Design and Implementation of Power Drive Controllers for LED Strings

by

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Abstract

In this thesis, design and implementation of power drives for light-emitting diode (LED) strings is investigated. We particularly focus on design methods for minimizing the size of output filter capacitor in flyback LED drivers. To this end, a novel constant power drive technique was developed to achieve better LED light regulation compared with the constant current technique. We present a filter capacitor minimization algorithm and applying it to an integrated buck-boost/flyback LED driver to achieve a long lasting LED driver.

Minimization of the filter capacitor in an ac-dc flyback converter is investigated by utilizing a descent algorithm. The algorithm was proposed using a relationship between input current harmonics and LED electrical and photometric characteristics. The performance of the proposed algorithm in terms of filter capacitor minimization was experimentally verified to achieve input power factor correction along with meeting light flicker requirements.

Furthermore, a primary-side constant power drive technique is proposed by utilizing a novel LED power estimation technique and an inner-outer-loop control structure. The proposed technique was implemented on an ac-dc flyback converter to attain simultaneous input power factor correction and LED light regulation. The enhanced performance of LED light regulation for the proposed technique is experimentally verified for different ambient temperatures and compared with the constant current drive method.

The above filter capacitor minimization algorithm was utilized and experimentally tested in an ac-dc integrated buck-boost/flyback converter. Utilizing this algorithm, the size of the required filter capacitors can be significantly reduced.

Keywords: Flyback LED driver, filter capacitor minimization, electrolytic capacitor-less LED driver, constant power, primary-side regulation, power factor correction.
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Chapter 1

Introduction

1.1 Motivation of the Research

Among all energy resources, electricity is the most convenient and accessible one ever known. There have been several sectors for electricity consumption, one of which is lighting which is responsible for 19% of the total electrical energy consumption [1]. Today, incandescent and fluorescent light bulbs are the most common light sources. About 40% of the electrical energy in lighting sector is used by incandescent light bulbs to produce 15% of the total light. The rest of the electrical energy in this sector is consumed by more efficient fluorescent light source to produce 85% of the total light [2]. The maximum luminous efficacy for incandescent and fluorescent light bulbs are 24 and 90 lum/W, respectively [3,4]. It shows that replacing incandescent light bulbs with more efficient light sources, has a significant effect on energy saving.

On the other hand, Solid State Lighting (SSL) is becoming more popular as the next generation of light source due to its higher luminous efficacy. Recently, high-brightness LEDs with high luminous efficacy of 294 lum/W was reported which is more than triple that of a fluorescent lamp (90 lum/W) [5]. Based on reported LED lighting applications, it can be concluded that incandescent and even fluorescent light sources can be successfully replaced with LEDs [6–9]. It shows that by using LED lamps, the potential for saving more energy in lighting sector is enormous.

In addition, LEDs enjoy relatively long lifetime up to 50 000 h [10]. Also, LED light bulbs are more reliable due to encapsulating with less glass. Moreover, mercury-free LEDs are environmentally friendly and can safely be disposed. The abovementioned advantages
Figure 1.1: Relative light output of red, blue and phosphor-converted white LEDs as a function of the junction temperature.

of LEDs compared with other light sources have made them highly promising light sources for future [11–13].

LED driver, as a media between power sources and LED, plays an important role in an LED lighting system. It is responsible for converting ac or dc power from source to appropriate dc power for the LED. However, the electrolytic capacitor is a necessary component in most LED drivers but its maximum lifetime is limited to typically 8000 h [14]. By comparing the lifetime of an electrolytic capacitor with an LED, it can be concluded that the lifetime of an LED system is limited to that of the electrolytic capacitor. Hence, designing an electrolytic capacitor-less LED driver can highly increase the life expectancy of an LED system.

Constant current LED driver is the most common driver in the market. It was proved that the LED light depends on the ambient temperature and its input power [15]. In constant current drive technique, the LED light reduces when the temperature increase as shown in Fig. 1.1 [16]. This dependency is an issue in applications exposed to temperature changes while regulation of light level in necessary. By reducing the sensitivity of the LED driver to ambient temperature, improved light regulation performance is expected.

The motivation of this research is the design and implementation of an electrolytic capacitor-less LED driver with long life expectancy. Meanwhile, improving the LED light regulation is another motivation of this work.
The emission mechanism of LEDs is electroluminescence. Captain Henry Joseph Round observed this phenomenon when a current flowed through a crystal of silicon carbide in 1907 [17]. In 1923 a Russian radio technician Oleg V. Lossev attempted to explain this phenomenon in a p-n junction by describing its current versus voltage characteristics [18, 19]. Since then LEDs have been commercially available for signaling, seven-segment displays, and remote control. The next breakthrough in LED technology was the introduction of blue LED by Shuji Nakamura at Nichia Corporation [20]. The blue LED helped the development of white LED by mixing red, green, and blue LEDs [21–23]. HB-LEDs was the next generation of light sources due to their long lifetimes, environmental friendliness, and high efficacy [24,25]. However, some fluorescent lamps showed higher luminous efficacy and lower heat generation compared with some LEDs [26] but, significant advancements in the manufacturing process of LEDs have made them highly promising light sources [27–31]. The advantages of LEDs versus incandescent and fluorescent lamps have initiated further research on their power drivers [28–37]. In the following subsections, different aspects of the research on LED drivers are briefly described.

1.2.1 Two-Stage vs. Single-Stage LED Drivers

An LED driver can be realized by two-stage or single-stage ac-dc converter. In a two-stage converter, a PFC stage is cascaded with a dc-dc converter. Each stage can be controlled individually to provide more flexibility, but the cost and complexity of the circuit will increase [38]. On the other hand, in single-stage LED drivers the two stages are integrated together to provide simpler power circuitry [39–45]. This type of LED driver is more desirable for low-power applications. One of the challenges of single-stage LED driver is the need for large filter capacitors. Eliminating this capacitor in LED drivers initiated further
research that will be described in the following subsection.

1.2.2 Electrolytic Capacitor Elimination

A single-stage LED driver requires an ac-dc converter with power factor correction (PFC). In a PFC converter, the input current is forced to be in phase with the input voltage, leading to a pulsating input power, whereas the output power is constant. Hence, a large filter capacitor is required to balance the instantaneous input and output powers. Large filter capacitor increases the size of the driver while reducing the power density. Usually electrolytic capacitors are used as the storage capacitors due to their large capacitance and low price. But they have short lifetime compared with other types of capacitors [14].

Several techniques have been proposed to reduce the size of filter capacitor [46–58]. In [46, 47], a new pulsating current LED driver is proposed in which the low frequency of the LED current is reduced without any electrolytic capacitor. In [48], a DCM power factor corrector and a CCM forward converter are combined into a single-switch topology. In this work, the energy storage capacitor is in the rectifier side with a multipurpose three-winding transformer. This transformer is used for isolation, power factor correction, and providing energy for the output. In [49, 50], a dc link capacitor and coupled inductors are used to provide energy for the LED. As a result, low ripple LED current is provided with small dc link capacitor.

In [51], the injection of third harmonic into the input current was studied for reducing the input power pulsation with an input power factor higher than 0.9 to comply with standards such as EN61000-3-2. The method of injecting odd harmonics to the input current was improved in [52]. In this work, the third and fifth harmonics were injected into the input current to reduce its peak-to-average ratio.

In another approach, the idea of replacing the filter capacitor with an active storage capacitor was studied. To this end, the concept of active power filter (APF) was studied in [53] by utilizing a bilateral converter. In [54], an H-bridge single-phase pulse width modulation (PWM) rectifier was studied. In this study, the feasibility of synthesizing active energy storage was proposed using a bidirectional buck and boost converter. Also, implementation of an active capacitor was proposed in [55] by utilizing an H-bridge configuration as an inverter. Bidirectional buck and boost converter was very common for active capacitor implementation [56–58]. It absorb the ac component of the LED driver’s input current using a small output filter capacitor. Consequently, thin film capacitors with longer lifetimes than
electrolytic ones were utilized in the implementation.

Reducing the filter capacitance improves the power density of the LED driver but it arises the issue of LED light flicker. This issue is described in more detail in the following subsection.

1.2.3 LED Light Flicker Considerations

Among different attributes of light sources, constant and flicker-free light is considered as an important factor on selecting the light source for lighting applications. This is not only depends on the light source but also on the driving circuitry. Flicker from light sources has been a concern for several decades. Whether visible or invisible, flicker can cause headache, migraines, fatigue, epilepsy, and other neurological responses [59]. It has been shown that flicker can degrade reading performance and cause distraction or annoyance for sensitive individuals [60]. According to the Illuminating Engineering Society (IES) lighting handbook [61], there are two measures for flicker that have been proposed by lighting designers, namely, Flicker Index and Percent Flicker which are given by

\[
\text{Flicker Index} = \frac{\text{Area 1}}{\text{Area 1+Area 2}},
\]

\[
\text{Percent Flicker} = \frac{E_{\text{max}} - E_{\text{min}}}{E_{\text{max}} + E_{\text{min}}} \times 100.
\]

Based on this handbook, the Flicker Index is preferred over the Percent Flicker, which is used to measure the relative cyclic variation of different light sources [62]. Such measures have to
be considered in the LED driver design when selecting factors such as the filter capacitance and control scheme. In this way, the output light signal was defined as a Fourier series of sinusoidal terms with different frequencies [63]. The low frequency terms of the Fourier series were considered in defining the flicker measures. However, the above work does not propose any relationship between the output light flicker and the size of filter capacitor. In fact, there has been no published work that has quantitatively studied the effect of filter capacitor on the output light flicker.

1.2.4 LED Drive Techniques

A common LED drive technique is current PWM that is widely used even in commercial LED drivers [32–34, 64]. In this technique, the dimming is achieved by changing the duty cycle but high junction temperature due to large ON-state currents reduces the output efficacy [11,13,32–34]. A dimmable LED driver with adaptive control was presented in [33]. In this work, the LED current is regulated through adjusting the duty cycle. Also, an improved PWM dimming technique was studied and implemented by simultaneous control over the amplitude and pulse width of the LED current. This techniques provides more flexibility for output current regulation but it increase the complexity of the driver. Another drive technique is direct current by which higher output luminous flux is achieved compared to PWM technique [36,37,56]. It also provides smooth and flicker-free LED light.

It was shown that, the LED’s junction temperature increase leads to its forward voltage drop and consequently its power drop [15,65–67]. As a result, the LED output light reduces due to its strong dependency to LED’s input power. This means, in constant current drive technique, the LED output luminous flux reduces as a result of any increase in its junction temperature [68–71]. In applications where temperature does not change considerably, constant current drive provides an almost constant output light. In these applications, the change of LED forward voltage versus temperature is small and can be neglected. When the temperature variations are large, the LED light variations is an issue as shown in Fig. 1.1.

1.2.5 Primary-Side Regulation Scheme

Isolated ac-dc converters are popular as LED drivers due to the safety issue. To regulate the LED current, secondary-side regulation (SSR) scheme is usually adopted as shown in
Figure 1.4: (a) Secondary-Side Regulation (SSR) scheme (b) Primary-Side Regulation (PSR).

Fig. 1.4a. In this scheme, the LED current is measured and regulated with a preset value in a control loop. In typical SSR systems, an optocoupler is used between the primary and the secondary to keep the isolation but it increases the complexity of the circuit [72, 73]. An integrated buck-boost/flyback converter is proposed in [42] with a buck-boost stage as a PFC. It converts the rectified input voltage to a dc voltage for the flyback stage. Then the flyback stage regulates the load voltage as a dc-dc converter. It measures the LED current on the secondary side using a series resistance and then regulates it with a preset value. This work was continued in [74] to focus on implementing a primary-side regulation (PSR) scheme. A primary-side LED current estimator is proposed using peak input current value and diode duty ratio. Figure 1.4b is an example of the PSR LED driver.

In PSR scheme, an auxiliary winding is usually used to measure the secondary-side information. This scheme enhances the efficiency of the circuit while reducing the size and cost [12, 45, 72, 73, 75–80]. In [45] the LED current is regulated using a primary side control scheme. In this work, the average current of the primary-side is measured and used for LED current estimation. In [72] a primary-side control scheme for TRIAC-dimmable LED driver is proposed. The proposed control scheme is suitable for discontinuous, continuous, and critical conduction modes. A similar work was done in [73] in which a new primary-side control scheme suitable for TRIAC dimmable LED drivers was proposed. In this work, the
LED power is estimated and controlled using the primary and auxiliary winding signals. In [75] a primary-side output current estimation and regulation circuit is proposed utilizing the auxiliary winding.

1.3 Summary of Contributions and Outline of Dissertation

Motivated by the issues mentioned in section 1.1 and utilizing previous works as summarized in section 1.2, the contributions of this thesis are briefly as follows:

1. Chapter 2: A novel algorithm is proposed to obtain the minimum required filter capacitor in an ac-dc flyback LED driver as shown in Fig. 1.5. To this end, two standards needs to be met, i.e., EN61000-3-2 class C for input current harmonics and ENERGY STAR for LED light flicker.

2. Chapter 3: A novel primary-side LED power regulation technique is proposed on the ac-dc flyback LED driver to alleviate the effect of ambient temperature on the LED light.

3. Chapter 4: The proposed filter capacitor minimization algorithm is extended on an ac-dc buck-boost/flyback LED driver as shown in Fig. 1.6 to achieve long lifetime
LED driver.

The contribution of each chapter is described in more detail in the following subsections.

1.3.1 Chapter 2: Filter Capacitor Minimization in a Flyback LED Driver Considering Input Current Harmonics and Light Flicker Characteristics

The contribution of this chapter is furnishing a quantitative relationship between the input current odd harmonics, the output light flicker, and the size of the filter capacitor in an ac-dc flyback LED driver. It is shown that, the input current odd harmonics and output light flicker are the only parameters that can affect the size of the required filter capacitor. Using this relationship, the required capacitor can be determined over the whole range of permissible amplitude for each harmonic, based on EN61000-3-2 standard, and the amplitude of output light flicker, based on ENERGY STAR standard. It is shown that, the input current’s third and fifth harmonics and the output light flicker are the most effective parameters that reduces the size of the filter capacitor. Thus, a minimum capacitance can be obtained when the maximum values of these parameters are selected based on the above mentioned standards.

The advantage of minimizing the filter capacitance is to increase the power density of the LED driver circuitry. Also, if the minimum required capacitor is in the range of long lifetime capacitors, e.g., ceramic or film ones, then long lifetime for the LED driver is expected. In addition, if any replacement circuitry, e.g., active capacitor, is aimed to be used instead of this capacitor, then a minimum capacitance needs to be provided by that circuitry. This leads to a lower current and voltage stress and higher lifetime for that circuit. It also helps to reduce the size, cost, and complexity of the circuit.

The results of this chapter was published in IEEE Transaction on Power Electronics journal as a paper entitled "Filter Capacitor Minimization in a Flyback LED Driver Considering Input Current Harmonics and Light Flicker Characteristics".
1.3.2 Chapter 3: A Double-Loop Primary-Side Control Structure for HB-LED Power Regulation

The contribution of this chapter is to improve the LED light regulation by developing a primary-side LED power control scheme. To this end, a novel primary-side LED power estimator and an inner-outer-loop controller is proposed. Utilizing the proposed method, power factor correction along with improved LED light regulation can be achieved. The proposed controller results in a relatively high power factor at different power levels. Meanwhile, a high accuracy primary-side LED power estimator is proposed in which no information from the secondary or auxiliary windings is needed.

The results of this technique was published in IEEE Transaction on Power Electronics journal entitled "A Double-Loop Primary-Side Control Structure for HB-LED Power Regulation". In addition, the results of primary-side non-auxiliary winding LED driver was submitted to 41st annual conference of the IEEE Industrial Electronics Society (IECON 2015) entitled "A Novel Primary-Side LED Power Regulation Without Auxiliary Winding".

1.3.3 Chapter 4: An Electrolytic Capacitor-less Buck-Boost/Flyback LED Driver Considering Light Flicker Characteristics

In this chapter, the method proposed in chapter 2 is used, modified, and extended for an ac-dc integrated buck-boost/flyback LED driver. In this chapter, a relationship between the input current, the output light flicker, and the size of the filter capacitor is obtained. Utilizing this relationship, an iterative algorithm is proposed by which the light flicker is determined for different values of filter capacitances. Then the capacitances by which the light flicker reaches its maximum value is considered as the minimum required filter capacitances. Experimental results are presented to verify the performance of the proposed method.

The result of this chapter was submitted to IEEE Transaction on Power Electronics journal entitled "An Electrolytic Capacitor-less Buck-Boost/Flyback LED Driver Considering Light Flicker Characteristics".
1.3.4 Chapter 5: Summary, Conclusions, and Suggestions for Future Work

The accomplishments of this research are summarized in this chapter. Based on theoretical studies and simulation and experimental results, a conclusion addressing the motivations of this research is presented. Finally, suggestions about potential research opportunities enlighten the path for future works.
Chapter 2

Filter Capacitor Minimization in a Flyback LED Driver Considering Input Current Harmonics and Light Flicker Characteristics

In this chapter, a comprehensive study is conducted on minimizing the size of output filter capacitor in an ac-dc flyback converter for driving HB-LED strings. To this end, a relationship between the input current harmonics, LED light flicker, and the magnitude of filter capacitor is obtained. It is shown that the size of the filter capacitor is mostly affected by the amplitude of the third and fifth harmonics of input current and the output light flicker. Considering the EN61000-3-2 standard for input current harmonics content and ENERGY STAR standard for flicker requirement, a procedure for obtaining the minimum value of filter capacitance is presented. Experimental studies are performed using the proposed method and tested on Cree XLamp XP-G and CR22-32L LED strings.

This chapter is organized as follows: The effect of input current odd harmonics and the output light flicker on the filter capacitance is studied in section 2.1. In this section, a quantitative relationship is derived, relating the size of filter capacitor, amplitudes of input current odd harmonics, and the output light Flicker Index or Percent Flicker measures. Section 2.2 illustrates how the above parameters can be utilized to reduce the size of the filter capacitance and obtain its minimum value. Section 2.3, presents the experimental
results to evaluate the validity of the proposed method.

2.1 Filter Capacitance Minimization

2.1.1 Effect of Input Current Odd Harmonics

Consider the ac-dc flyback LED driver shown in Fig. 2.1 with the input voltage as follows

\[ v_i(t) = V_m \sin \omega t \]  

(2.1)

where \( V_m \) and \( \omega \) are the amplitude and the angular frequency with period \( T \), respectively. For an input power factor of 1, the input current is given by

\[ i_{i1}(t) = I_1 \sin \omega t \]  

(2.2)

where \( I_1 \) is the amplitude of the first harmonic of the input current. From (2.1) and (2.2), the instantaneous input power is given by

\[ p_{i1}(t) = \frac{V_m I_1}{2} (1 - \cos 2\omega t). \]  

(2.3)

Equation (2.3) indicates that the input power to the converter consists of dc and ac components. Delivering dc power to the LED string requires a large capacitor across the LED to absorb the ac ripple components and provide power balance. By injecting some odd harmonics to the input current while meeting available standards for the input current harmonics content, the amplitude of the ac components of the input power is reduced. It help to use smaller filter capacitances to provide the power balance. Assume that the input current is a summation of the first \( h \) odd harmonics as follows

\[ i_{ih}(t) = I_1(\sin \omega t + \sum_{n=3}^{h} I_n^* \sin n\omega t) \]  

(2.4)
where $I_n^*$ is the normalized amplitude of $n^{th}$ harmonic of the input current with the amplitude of first harmonic given by $I_1$ and $n$ is an odd number.

Using Fig. 2.1 and assuming 100% efficiency for the converter, the diode current can be obtained by utilizing the converter’s power balance, i.e., $p_o(t) = p_i(t)$, as follows

$$i_D(t) = M(t) i_{irh}(t) \quad (2.5)$$

where $M(t) = v_{ir}(t)/v_C(t)$ and $i_{irh}(t)$ is the rectified of the input current given by (2.4). Since the capacitor voltage ripple has much smaller variations than the input voltage, $M(t)$ can be approximated using the average of capacitor voltage, i.e., $M(t) = v_{ir}(t)/v_{C_{avg}}$. Now, let us assume that the switch and diode forward voltages are negligible when compared with the input and capacitor voltages. Therefore, the diode current in (2.5) can be rewritten as follows

$$i_D(t) = \frac{v_{ir}(t) i_{irh}(t)}{v_{C_{avg}}} = \frac{p_{ih}(t)}{v_{C_{avg}}} \quad (2.6)$$

where $p_{ih}(t)$ is the input power when the input current contains the first $h$ odd harmonics. Using (2.1) and (2.4), the input power can be obtained as follows

$$p_{ih}(t) = \frac{V_m I_1}{2} \left(1 - \cos 2\omega t + \sum_{n=3}^h I_n^* \left(\cos(n-1)\omega t - \cos(n+1)\omega t \right) \right). \quad (2.7)$$

Finally, substituting (2.7) in (2.6) yields

$$i_D(t) = I_d \left(1 - \cos 2\omega t + \sum_{n=3}^h I_n^* \left(\cos(n-1)\omega t - \cos(n+1)\omega t \right) \right) \quad (2.8)$$

where

$$I_d = \frac{V_m I_1}{2v_{C_{avg}}} \quad (2.9)$$
A simple model of the ac-dc flyback LED driver shown in Fig. 2.1 is shown in Fig. 2.2. In this model, the flyback converter is modeled as a current source $i_D$ given by (2.8) and (2.9) and the LED is modeled as a dc voltage source $V_{\text{LED}}$ in series with a resistance $r_{\text{LED}}$ [81], i.e., LED small-signal model. To obtain the capacitor voltage $v_C$ in this model, the following differential equation needs to be solved

$$\frac{dv_C(t)}{dt} + \frac{1}{\tau} v_C(t) = \frac{r_{\text{LED}}}{\tau} i_D(t) + \frac{1}{\tau} V_{\text{LED}} \tag{2.10}$$

where $\tau = r_{\text{LED}} C$. Solving (2.10) results in

$$v_C(t) = V_{\text{LED}} + r_{\text{LED}} I_d - \frac{r_{\text{LED}} I_d (2\omega \tau \sin 2\omega t + \cos 2\omega t)}{1 + (2\omega \tau)^2}$$

$$+ \sum_{n=3}^{h} \frac{r_{\text{LED}} I_d I_n^* ((n-1)\omega \tau \sin (n-1)\omega t + \cos (n-1)\omega t)}{1 + ((n-1)\omega \tau)^2}$$

$$- \sum_{n=3}^{h} \frac{r_{\text{LED}} I_d I_n^* ((n+1)\omega \tau \sin (n+1)\omega t + \cos (n+1)\omega t)}{1 + ((n+1)\omega \tau)^2}. \tag{2.11}$$

Equation (2.11) represents the effect of input current odd harmonics on the capacitor voltage and will be used to describe the effect of these harmonics on the output light flicker as discussed in section 2.1.2. Utilizing (2.10), the time constant for the flyback circuit model in Fig. 2.2 is given by $\tau = r_{\text{LED}} C$, which is approximately 0.01 ms for $r_{\text{LED}}=1 \Omega$ and $C=100 \mu F$. Thus the exponential part of the capacitor voltage $v_C(t)$ decays quickly to zero in about 0.04 ms.
2.1.2 Effect on Output Light Flicker

Now, consider the LED model shown in Fig. 2.2. Hence, the LED current is given by

\[ i_{\text{LED}}(t) = \frac{v_{C}(t) - V_{\text{LED}}}{r_{\text{LED}}}. \]  

(2.12)

Therefore, the LED power can be obtained as follows

\[ p_{\text{LED}}(t) = v_{C}(t) \times i_{\text{LED}}(t) = \frac{v^{2}_{C}(t) - V_{\text{LED}}v_{C}(t)}{r_{\text{LED}}}. \]  

(2.13)

It has been shown that the output luminous flux of an LED has a second-order polynomial relationship with LED power [15]. Utilizing this result, we have

\[ \phi_{v}(p_{\text{LED}}(t)) = \alpha_{1}p_{\text{LED}}(t) - \alpha_{2}p_{\text{LED}}^{2}(t) \]  

(2.14)

where \( \phi_{v} \) is the output luminous flux and \( \alpha_{1} \) and \( \alpha_{2} \) are two positive coefficients that can be obtained using the thermal characteristics of the LED system. Finally, by substituting (2.13) in (2.14) the output luminous flux can be represented as a function of time as follows

\[ \phi_{v}(t) = -\frac{\alpha_{1}V_{\text{LED}}}{r_{\text{LED}}}v_{C}(t) + (\frac{\alpha_{1}}{r_{\text{LED}}} - \frac{\alpha_{2}V^{2}_{\text{LED}}}{r^{2}_{\text{LED}}})v^{2}_{C}(t) + \frac{2\alpha_{2}V_{\text{LED}}}{r^{2}_{\text{LED}}}v^{3}_{C}(t) - \frac{\alpha_{2}}{r^{2}_{\text{LED}}}v^{4}_{C}(t). \]  

(2.15)

In addition, the average capacitor voltage in (2.11) is given by

\[ v_{C_{\text{avg}}} = V_{\text{LED}} + r_{\text{LED}}I_{d} \]  

(2.16)
Figure 2.5: Measured data and exponential fit curve for LED current vs. voltage for Cree XLamp XP-G LED string.

which upon utilizing (2.9) yields

\[ v_{\text{avg}} = \frac{1}{2} \left( V_{\text{LED}} + \sqrt{V_{\text{LED}}^2 + 4P_\text{d}^\text{LED}} \right). \]  

(2.17)

Any analysis of photometric flicker requires accurate measurement of luminous flux modulation \( \phi_v \) emitted from a light source. A standard procedure for this measurement has not been proposed in the literature. Instead, the illuminance \( E_v \) can be measured using analog linear photosensors that are capable of measuring illuminance over a wide range, e.g., 10-10000 lx. The analog output of these types of photosensors can be monitored using a high resolution digital scope. The output illuminance and its average can be given by

\[ E_v(p_{\text{LED}}(t)) = k\phi_v(p_{\text{LED}}(t)), \]

(2.18)

\[ E_{\text{avg}} = k\phi_{\text{avg}} \]  

(2.19)

where \( k \) is a constant coefficient representing the portion of output luminous flux that illuminates the surface of the photosensor.

According to the IES Lighting flicker Handbook [61], there are two measures of flicker that have commonly been proposed by lighting designers, i.e., the Flicker Index and Percent Flicker. The Flicker Index is often used to measure the relative cyclic variations of different light sources. Referring to Fig. 2.3, this index is defined as the area above the line of average light \( E_{\text{avg}} \) divided by the total area of the light curve \( E_v(t) \) as follows

\[ \text{Flicker Index} = \frac{\text{Area 1}}{\text{Area 1+Area 2}}. \]  

(2.20)
Using Fig. 2.3, the numerator of (2.20) is given by

\[ \text{Area 1} = \int_{t_1}^{t_2} (E_v(t) - E_{v \text{avg}}) \, dt. \]  \hspace{1cm} (2.21)  

Also, the denominator of (2.20) is given by

\[ \text{Area 1} + \text{Area 2} = \int_0^{T/2} E_v(t) \, dt = \frac{T}{2} E_{v \text{avg}}. \]  \hspace{1cm} (2.22)  

Equation (2.20) can be rewritten by utilizing (2.21) and (2.22) in the following form

\[ \text{Flicker Index} = \frac{2f}{E_{v \text{avg}}} \int_{t_1}^{t_2} (E_v(t) - E_{v \text{avg}}) \, dt \]  \hspace{1cm} (2.23)  

where \( f = 1/T \) is the fundamental frequency. Finally, (2.23) can be rewritten by replacing illuminance with luminous flux by utilizing (2.18) and (2.19) as follows

\[ \text{Flicker Index} = \frac{2f}{\phi_{v \text{avg}}} \int_{t_1}^{t_2} (\phi_v(t) - \phi_{v \text{avg}}) \, dt \]  \hspace{1cm} (2.24)  

where \( \phi_v(t) \) is given by (2.15) and \( k \) has been canceled. Substituting (2.11) into (2.13) the LED power can be obtained as a function of input current harmonics. Utilizing the result in (2.14) and using (2.17) in (2.24), the relationship between amplitudes of the input current odd harmonics \( I_n^* \), capacitance \( C \), and the Flicker Index can be obtained. It should be noted that the filter capacitance \( C \) cannot be explicitly obtained, thus a numerical procedure will be used in its calculation.
Figure 2.7: Minimum required capacitor for Cree XLamp XP-G LED string with different third harmonic amplitudes when both criteria for Percent Flicker and Flicker Index are met when $P_d = 5, 10, 15, \text{ and } 20 \text{ W}$.

Furthermore, according to the IES Lighting Handbook and referring to Fig. 2.3, Percent Flicker is defined as follows

$$ \text{Percent Flicker} = \frac{E_{v_{\text{max}}} - E_{v_{\text{min}}}}{E_{v_{\text{max}}} + E_{v_{\text{min}}}} \times 100. \quad (2.25) $$

Again, utilizing (2.18) and (2.19) and replacing illuminance with luminous flux, (2.25) can be rewritten as follows

$$ \text{Percent Flicker} = \frac{\phi_{v_{\text{max}}} - \phi_{v_{\text{min}}}}{\phi_{v_{\text{max}}} + \phi_{v_{\text{min}}}} \times 100 \quad (2.26) $$

where $k$ has been canceled. According to the IES Lighting Handbook, the Flicker Index is preferred over Percent Flicker. Combining Flicker Index and Percent Flicker creates a measure for discussing flicker. Based on ENERGY STAR standard, most traditional light sources occupy an area enclosed by a maximum Flicker Index of 0.13 and a maximum Percent Flicker of 40%, referred to as the flicker frame of reference [63].

In the proposed control scheme, the input current follows a desired current which also determines the power factor. Thus capacitor degradation does not have any effect on the input power factor and only affects the LED light flicker. However, capacitance degradation would result in the LED light flicker to increase. Thus for the light flicker to meet the ENERGY STAR standard, the degradation effect has to be considered. The effect of capacitor degradation can be considered if the percentage of capacitor degradation during its lifetime is known and added to the calculated capacitance in the minimization scheme.
2.2 Experimental Study of the Proposed Method

An LED string consisting of 16 Cree XLamp XP-G LEDs in series was selected for experimental evaluation. To obtain the required capacitor based on different amplitudes of the input current harmonics and output light flicker (i.e., solving (2.24) and (2.26)), the electrical and photometric characteristics of the LED string given by $r_{LED}$, $V_{LED}$, $\alpha_1$ and $\alpha_2$ have to be obtained.

2.2.1 Electrical and Photometric Characteristics

Two experimental methods are described in the following subsections to obtain these parameters.

**Obtaining $r_{LED}$ and $V_{LED}$**

A typical current-voltage characteristic of an LED string is shown in Fig. 2.4. Using this figure and assuming that the range of operating power is from $P_{d1}$ to $P_{d2}$, it can be observed that the LED resistance $r_{LED}$ changes based on the operating point as follows: If $P_{d1} < P_{d2}$ then $r_{LED1} > r_{LED2}$. The LED currents for 16 Cree XLamp XP-G LEDs were measured and plotted in Fig. 2.5. For voltages lower than 35V the LED current is almost zero, whereas for higher voltages it follows an exponential function obtained by a regression analysis.
Therefore, the LED current-voltage characteristic is given by

\[ i_{\text{LED}} = \begin{cases} 0 & v_C \leq V_{\text{LED}} \\ ae^{b(v_C-V_{\text{LED}})} & v_C > V_{\text{LED}} \end{cases} \]  \hspace{1cm} (2.27)

where \( a = 0.012, \ b = 0.57, \) and \( V_{\text{LED}} \) is the dc source in Fig. 2.2 which was chosen as 35V.

Utilizing (2.27), LED resistance and power are given by

\[ r_{\text{LED}} = \left( \frac{di_{\text{LED}}}{dv_C} \right)^{-1} = \frac{1}{ab} e^{-b(v_C-V_{\text{LED}})}, \]  \hspace{1cm} (2.28)

\[ p_{\text{LED}} = av_C e^{b(v_C-V_{\text{LED}})}. \]  \hspace{1cm} (2.29)

It should be noted that the LED resistance is not only a function of the operating point, but also a function of the LED p-n junction temperature. When the LED turns ON, the junction temperature increases with time and the LED resistance reduces due to an increase in number of charge carriers in the semiconductor [82, 83]. Thus the minimum LED resistance can be obtained when the LED works under its maximum p-n junction temperature. To obtain the LED current-voltage characteristics shown in Fig. 2.5, the measurement at each point were acquired when the steady-state conditions for the p-n junction temperature were reached.

**Obtaining \( \alpha_1 \) and \( \alpha_2 \)**

By assuming the LED power as \( P_d \) that increases from zero based on (2.14), the output luminous flux \( \phi_v \) increases almost linearly because the second term is negligible when LED
power is small [15]. Thus for low LED power levels, the output luminous flux in (2.14) is given by \( \phi_v(P_d) \approx \alpha_1 P_d \). Hence, \( \alpha_1 = E \), where \( E \) is the efficacy of the LED. Using the Cree XLamp XP-G LED datasheet [84], \( \alpha_1 = 94 \) lum/W. For higher LED power levels, the second negative term in (2.14), which is proportional to the square of power, will significantly reduce \( \phi_v \). To obtain \( \alpha_2 \), the output illuminance was obtained in terms of the LED power up to 30 W and plotted in Fig. 2.6. Using this plot and regression analysis, the illuminance equation can be obtained in the following form

\[
E_v(P_d) = 478.9P_d - 9.4P_d^2. \tag{2.30}
\]

To obtain \( \alpha_2 \), let us substitute (2.30) in (2.18) as follows

\[
478.9P_d - 9.4P_d^2 = k\phi_v(P_d). \tag{2.31}
\]

Now, using (2.14) and \( \alpha_1 = 94 \) lum/W, (2.31) can be rewritten as follows

\[
478.9P_d - 9.4P_d^2 = 94kP_d - \alpha_2kP_d^2 \tag{2.32}
\]

where \( k \approx 5.1 \) and \( \alpha_2 \approx 1.84 \).

### 2.2.2 Effect of Input Current Harmonics and Output Light Flicker

Let us assume that the input current in (2.4) contains the first and third harmonics. Also, assume that the normalized amplitude of the third harmonic \( I_3^* \) and the filter capacitance \( C \) are unknown. Thus the capacitor voltage in (2.11) is a function of \( t \), normalized amplitude of the third harmonic \( I_3^* \), and the filter capacitance \( C \), i.e., \( v_C(t, I_3^*, C) \). Consequently, the output luminous flux in (2.15) and illuminance in (2.18) have the same independent
variables, i.e., $\phi_v(t, I_3^*, C)$ and $E_v(t, I_3^*, C)$, respectively. Equation (2.24) can be rewritten as follows

$$\text{Flicker Index} = \frac{2f}{\phi_{v,\text{avg}}} \int_{t_1}^{t_2} (\phi_v(t, I_3^*, C) - \phi_{v,\text{avg}}) \, dt. \quad (2.33)$$

If the input current contains the first and third harmonics then the input Power Factor is

$$PF = \frac{I_1}{\sqrt{I_1^2 + (I_1 I_3^*)^2}} = \frac{1}{\sqrt{1 + (I_3^*)^2}}. \quad (2.34)$$

From (2.34), PF as a function of $I_3^*$ is obtained. In order to comply with EN61000-3-2 standard, not only the input power factor has to be more than 0.9, but also there is a maximum permissible for each individual harmonic. Based on this standard, by considering the maximum permissible normalized third harmonic, i.e., 0.3, and using (2.34), the input power factor is 0.96 which complies with this standard. To study the effect of third harmonic on the size of the capacitor based on (2.26) and (2.33) for a desired power $P_d$, a numerical calculation can be used as follows:

1. Obtain $\phi_v(t, C)$ from (2.15) using a desired $I_3^*$, $f = 60$ Hz, $\alpha_1 = 94$ lum/W, $\alpha_2 = 1.84$ lum/W$^2$, $V_{LED} = 35$ V, and $r_{LED}$ from (2.28).
2. Obtain $\phi_v(t)$ by substituting a known capacitance.
3. Calculate $\phi_{v,\text{min}}, \phi_{v,\text{avg}},$ and $\phi_{v,\text{max}}$.
4. Calculate $t_1$ and $t_2$ by obtaining roots of $\phi_v(t) - \phi_{v,\text{avg}}$. 

Figure 2.11: Flicker measurement circuit using linear output ambient light sensor.
5. Calculate Percent Flicker and Flicker Index using (2.26) and (2.33), respectively.

6. If either the Percent Flicker or Flicker Index are higher than 40% or 0.13, respectively,
then increase the capacitance; otherwise, decrease the capacitance and repeat this
procedure from step 2 with the new capacitance.

The above algorithm has to be repeated to find $C$ with any desired resolution on the ca-
pacitance changes until both Percent Flicker and Flicker Index are below the maximum
permissible.

Using the above algorithm $I_3^*$ is swept from 0 to 0.3 and the minimum required capaci-
tance was obtained for four different powers as shown in Fig. 2.7. In this figure both criteria,
i.e., Percent Flicker $< 40\%$ and Flicker Index $< 0.13$ are met. This figure shows that when
$I_3^*$ assumed maximum value, the minimum filter capacitances of 110 $\mu$F, 200 $\mu$F, 265 $\mu$F,
and 310 $\mu$F were obtained for 5 W, 10 W, 15 W, and 20 W LED powers, respectively.

To study the effect of fifth harmonic, let us assume that the input current in (2.4)
contains the first, third, and fifth harmonics. Then the input Power Factor is

$$ PF = \frac{I_1}{\sqrt{I_1^2 + (I_1 I_3^*)^2 + (I_1 I_5^*)^2}} = \frac{1}{\sqrt{1 + (I_3^*)^2 + (I_5^*)^2}} \tag{2.35} $$

where $I_5^*$ is the normalized amplitude of fifth harmonic. Based on EN61000-3-2 standard,
maximum permissible for $I_5^*$ is 0.1. Considering $I_3^*$ and $I_5^*$ on their maximum the input power
factor given by (2.35) is 0.95 which complies with the abovementioned standard. Following
Figure 2.13: Experimental waveforms of input voltage, input current, LED voltage, and LED current with 120V input voltage and 20 W output power for Cree XLamp XP-G LED string.

the same procedure and using (2.24), a relationship between the normalized amplitude of third and fifth harmonics, filter capacitor, and Flicker Index is given by

$$\text{Flicker Index} = \frac{2f}{\phi_{\text{avg}}} \int_{t_1}^{t_2} (\phi_v(t, I_3^*, I_5^*, C) - \phi_{\text{avg}}) \, dt.$$  \hspace{1cm} (2.36)

Now, (2.26) and (2.36) can be used in the abovementioned algorithm. Figure 2.8 shows the minimum required capacitor with different normalized amplitudes of third and fifth harmonic into the input current. It shows that when $I_3^* = 0.3$ and $I_5^* = 0.1$, minimum 100 µF capacitor is required for 5 W LED power. This figure shows the effect of both third and fifth harmonics on the filter capacitance. If the fifth harmonic is zero the graph is the same as Fig. 2.7 in which the effect of only third harmonic was shown. By increasing the fifth harmonic up to its maximum permissible, i.e., 0.1, the required filter capacitance decreases and its minimum occurs when both third and fifth harmonics are on their maximum values. The algorithm was repeated for 10 W, 15 W, and 20 W LED powers and minimum 180 µF, 240 µF, and 270 µF capacitors were obtained. In these cases, injection of the fifth harmonic results in approximately 10% reduction in the minimum required capacitor. In order to comply with other standards such as IEEE STD 519 or JIS C 61000-3-2 class C, input current harmonic limits can be considered accordingly in (2.36) to obtain the minimum capacitor.

Solving (2.24) and (2.26) for harmonics higher than five revealed that the injection of those harmonics did not significantly reduce the filter capacitor. Thus it is concluded that
the most important factors involved in reducing the filter capacitor were the third and fifth harmonics of the input current.

For each operating point, the capacitor voltage $v_C(t)$ can be obtained using (2.29) which is used in (2.28) to obtain $r_{LED}$ for the minimization. Based on the results obtained in Figs. 2.7 and 2.8 for Cree XLamp XP-G LED, and Figs. 2.18 and 2.19 for Cree CR22-32L LED, a larger filter capacitance is required for a higher LED power. Therefore, by choosing the operating point corresponding to the highest power (worst case scenario) and obtaining its minimum capacitor, one can guarantee that for lower power levels the requirements for input current harmonics and output light flicker are both satisfied. As a result, the resistance at maximum power operating point is chosen in the minimization procedure.

Using the proposed method, the minimum required capacitor was greatly reduced. The idea of injecting odd harmonics to the input current to reduce the output capacitor was studied in literature. In [51], the injection of only third harmonic was studied with a 141 µF filter capacitor that is in the range of capacitors that we have obtained. In [52], a 8.75W LED driver was studied with the third and fifth harmonics injection to the input current and a very small capacitor less than 1 µF was used. But based on the provided results, the LED current contains a ripple with twice the input frequency. This ripple has a large peak-to-peak value leading to high Percent Flicker and Flicker Index. In non of these studies the output light flicker characteristics were considered. In addition, we have used 14W and 75W flyback LED drivers (from Power Integrations Inc.) in which 660µF and
4400 µF electrolytic capacitors were used, respectively, which are quite high when compared to the 100-270 µF capacitors obtained in this study. Furthermore, in the reference design RDR-195 (from Power Integrations Inc.) for driving a 14W LED load, a 660 µF capacitor is used, which in the proposed method for a 15W LED, a 175 µF capacitor is required, indicating an almost 73% savings in capacitor value.

2.2.3 Controller Implementation

Based on the result obtained in section 2.2.2 (discussion after (2.36)), if the input current contains the first, third, and fifth harmonics with amplitudes $I_1, I_3 = 0.3I_1$, and $I_5 = 0.1I_1$, respectively, then a minimum filter capacitance can be used. The control objective is thus to force the converter’s input current to follow a desired value as follows

$$i_{id}^* (t) = I_1 i_{id}^o (t) = I_1 (\sin \omega t + 0.3 \sin 3 \omega t + 0.1 \sin 5 \omega t).$$

Figure 2.9 illustrates the control system blocks. The input current after the rectifier is denoted by $i_{ir}$ (see also Fig. 2.10), which is controlled to follow the absolute value of the desired input current in (2.37). Therefore, the normalized rectified input voltage $v_{ir}^* (t)$ is used as the input to a lookup table. The relationship between the input and output of the lookup table is based on $i_{id}^* (t)$ in (2.37) in which the angular frequency $\omega t$ is obtained using the lookup table input, i.e., normalized rectified input voltage.

Using (2.7), the average input power to the driving circuit can be obtained by taking
the integral of the instantaneous input power over one period as follows

$$P_d = \frac{2}{T} \int_0^{T/2} p_{ih}(t) \, dt = \frac{V_m I_1}{2}. \quad (2.38)$$

Assuming 100% efficiency for the LED driver, the LED power is equal to the average input power $P_d$. Utilizing this relationship, the amplitude of the first harmonic $I_1$ in (2.37) can be obtained for any desired LED power as depicted in Fig. 2.9. By changing the desired power $P_d$, the amplitude of first harmonic $I_1$ would change, resulting in LED dimming capability without any changes to the input power factor.

Controller implementation can be achieved using both analog and digital methods. In this work, a TMS320F28027 microcontroller (MCU) from Texas Instruments Inc. was used. Using this MCU, a lookup table can be constructed to obtain the desired input current when the normalized rectified input voltage acts as its input.

### 2.3 Experimental Results

The schematic diagram of single-stage ac-dc flyback LED driver with proposed input current control capability is shown in Fig. 2.10. A PI input current controller was implemented and tested on Cree XLamp XP-G and CR22-32L LED strings.

The specifications of the prototype are as follows

1. Input voltage: $v_i = 120 \text{ V} / 60 \text{ Hz};$
2. Maximum output power: $P_d = 20\, \text{W}$;

3. Switching frequency: $f_s = 30\, \text{kHz}$.

The power devices and components of the flyback are as follows:

1. diode bridge $BR$: KBL10 (1000V, 4A);
2. peak detector diode $D_1$: 1N4007 (1000V, 1A);
3. switch $S$: IRFB11N50A N-channel MOSFET (500V, 11A);
4. voltage clamp Zener diode $Z$: 1.5KE200A (200V, 1500W);
5. voltage clamp diode $D_2$: UF5408 (1000V, 3A);
6. Transformer $T$: 1207 $\mu$H primary inductance, 15 $\mu$H primary leakage inductance, and turns are 50/14;
7. flyback diode $D$: MBR40250G Shottky diode (250V, 40A).

Figure 2.11 depicts the circuit used in this study to measure the output light flicker. In this circuit, a linear ambient light sensor (GA1A2S100) was used and its output current was converted to a voltage waveform using a $100\, \Omega$ resistor. This analog voltage was monitored using a high resolution (20 GS/s) digital scope to measure the light flicker. Using this voltage...
Figure 2.18: Minimum required capacitor for Cree CR22-32L LED string with different third harmonic amplitudes when both criteria for Percent Flicker and Flicker Index are met when $P_d = 5, 10, 15,$ and 20 W.

and the output characteristics of the photosensor, the output illuminance was obtained to measure both Flicker Index and Percent Flicker.

As it was discussed in Section 2.2.2, if the input current contains the first, third, and fifth harmonics, then the required filter capacitor for a constant output power is minimum. To this end, the amplitude of the third and fifth harmonics of the experimental setup are controlled to be 0.3 and 0.1 of the first harmonic, respectively as shown in Fig. 2.12. From this figure, it follows that the input current contains only the first, third, and fifth harmonics and higher harmonics are zero. Also, the amplitude of the third and fifth harmonics are in agreement with the results obtained in section 2.2.2. Figure 2.13 shows the input voltage and current of the LED driver driven by 120V input voltage source. As shown in this figure, the input current is in phase with input voltage results a high input power factor. Also, the LED voltage and current waveforms delivering 20 W of power to the LED string are illustrated in this figure.

2.3.1 Test on Cree XLamp XP-G LEDs

A 270\(\mu\)F capacitor, which is the minimum calculated capacitor for the maximum output power obtained in section 2.2.2, was used in the power driver circuitry to drive 16 Cree XLamp XP-G LEDs. In this way, the Flicker Index and Percent Flicker for 5W, 10W, 15W, and 20W output powers were measured every 5 minutes during 1 hour of operation.
of the circuit and plotted in Fig. 2.14 and Fig. 2.15, respectively. The results indicate that for the maximum output power of 20 W, the Flicker Index and Percent Flicker are below 0.13 and 40%, respectively, which is in good agreement with numerical results. For lower output power levels, lower Flicker Indices and Percent Flickers were measured, which is in agreement with the discussion in section 2.2.2.

2.3.2 Test on Cree CR22-32L LEDs

The proposed method was also investigated on a Cree CR22-32L LED string consisting of 20 LEDs. The LED current $i_{\text{LED}}$ were measured and plotted vs. LED voltage $v_C$ in Fig. 2.16 for voltages higher than 25 V. Using this plot and regression analysis, the exponential form of LED current given by (2.27) was obtained with $V_{\text{LED}} = 25 \text{ V}$ $a = 0.04$ and $b = 0.54$. In addition, the output illuminance was obtained in terms of the LED power up to 25 W and plotted in Fig. 2.17. Using regression analysis for this plot and CR22-32L datasheet [85], the output luminous flux in (2.14) can be obtained with $\alpha_1 = 100$ and $\alpha_2 = 1.44$.

Using the above information and the algorithm in section 2.2.2, the minimum required capacitance for different $I_3^*$ were obtained utilizing (2.26) and (2.33) and plotted in Fig. 2.18. This figure shows that when $I_3^*$ assumed its maximum value, the minimum capacitance as 220 $\mu$F, 375 $\mu$F, 485 $\mu$F, and 545 $\mu$F is needed for 5 W, 10 W, 15 W, and 20 W LED powers, respectively. To further reduce the required capacitance, the fifth harmonic was injected
and the minimum required capacitances were obtained using (2.26) and (2.36) as plotted in Fig. 2.19. Thus for $I_3^* = 0.3$ and $I_5^* = 0.1$, minimum value of 200 µF, 335 µF, 440 µF, and 500 µF capacitors are required for 5W, 10W, 15W, and 20 W LED powers, respectively. In these cases, injection of fifth harmonic results in approximately 8% reduction in the minimum required capacitor. Based on these results, a 500 µF capacitor was used in the power driver circuitry to drive CR22-32L LED string. The Flicker Index and Percent Flicker for different output powers were measured every 5 minutes during 1 hour of operation of the circuit and plotted in Fig. 2.20 and Fig. 2.21, respectively. The results indicate that for a maximum output power of 20 W, the Flicker Index and Percent Flicker are below 0.13 and 40%, respectively. Furthermore, for lower output power levels, lower Flicker Indices and Percent Flickers were measured, which is in agreement with the proposed method.

In [86], accelerated life tests were carried out on two different high brightness light emitting diodes. Based on experimental results, the authors concluded that the series resistance of the LED device can increase due to chip level degradation such as contact property worsening and contact detachment. Aging mechanisms in the optical and electrical characteristics of LEDs were presented in [87] which do not indicate major changes in the LED I-V curve for current ranges greater than 1mA. Thus, as a result of aging the series resistance can either increase or remain approximately constant, which implies that the time constant of the circuit shown in Fig. 2.2 can either increases or remain constant. Hence the overall effect would be to reduce output voltage ripple and light.
2.4 Summary and Conclusions

In this chapter, a comprehensive study was conducted on the parameters of the ac-dc flyback LED drive circuitry, from which a minimum required filter capacitor was obtained based on allowable light flicker measures and harmonic contents. Through analytic development, it was shown that the size of the filter capacitor would depend on the input current odd harmonics and the output light flicker. The study quantitatively indicated that injection of third and fifth harmonics to the input current can effectively help in reducing the size of the filter capacitor at the cost of reducing the input power factor. By setting the amplitude of these harmonics on their maximum value using EN61000-3-2 standard by which the input power factor of 0.9 is guaranteed, the size of filter capacitor was greatly reduced. It was further shown that higher odd harmonics content do not have a significant effect on reducing the size of this capacitor. On the other hand, the effect of output light flicker using the Flicker Index and Percent Flicker measures were studied. Reducing the filter capacitance leads to an increase in the output light flicker. Thus by setting the output light flicker to its maximum value based on the ENERGY STAR standard, a minimum value for filter capacitor was obtained. The proposed method was verified through experimental studies which validated analytic developments.

Figure 2.21: Percent Flicker vs. time obtained from experimental setup for Cree CR22-32L LED string when $C = 500 \mu\text{F}$ and $P_d = 5, 10, 15, \text{ and } 20 \text{ W.}$
Chapter 3

A Double-Loop Primary-Side Control Structure for HB-LED Power Regulation

In this chapter, a study of HB-LED strings under constant power control concept is conducted in which the LED’s input power is controlled rather than its current. In the conventional constant current drive technique, any changes in the LED string’s forward voltage results in output light variations due to a strong dependency between the LED output light to its input power. Under constant power drive control, the LED string’s forward voltage variations due to any effect such as ambient temperature changes and aging are compensated. To this end, a primary-side LED power estimator and controller are proposed to achieve simultaneous output power regulation and input power factor correction. The controller consists of an inner-outer-loop control structure implemented on an ac-dc flyback converter. Also, an LED power estimator is presented that requires no information from the secondary or auxiliary windings. An expression is obtained for the LED power in terms of its input current and ambient temperature that can provide qualitative and quantitative behavior in constant-current and constant-power drive regimes. Experimental results using a Cree CR22-32L LED string are presented and compared with the constant current drive technique for temperature variations in the range 25°C to 80°C. For the above temperature range and using similar hardware, the proposed controller results in an 8% reduction of the LED output light when compared to a 13% reduction using the constant current drive.
This chapter is organized as follows: In section 3.1, a relationship between the input current and LED power is derived, based on which a desired input current is constructed and utilized to control the converter’s input current. To achieve the desired input current, the converter is modeled and controlled using the averaging method. Section 3.2 presents the primary-side LED power estimator circuit and its control scheme. Both LED power estimator circuitry with and without auxiliary windings are presented in this section. In section 3.3, an expression between the LED input power and its current and ambient temperature is obtained to study variation of power versus temperature and input current. In section 3.4, experimental results are presented to highlight performance of the proposed constant power control concept compared with the constant current control method.

3.1 Input Current Control

3.1.1 Relationship between Input Current and LED Power

A schematic diagram of the ac-dc flyback LED driver utilized in this study is shown in Fig. 3.1. Let us denote the input voltage and current as $v_i(t)$ and $i_i(t)$ as given by (2.1) and (2.2), respectively. For now, assuming 100% efficiency for the LED driver, we have $p_o(t) = p_i(t)$. Assuming that the output capacitor in Fig. 3.1 is large enough and capable of absorbing
Figure 3.2: Simplified configuration of ac-dc flyback LED driver when (a) switch $S$ is ON and (b) switch $S$ is OFF.

The ac component of the output power, the average LED power is obtained as follows

$$P_{LED} = \frac{2}{T} \int_0^{T/2} p(t) \, dt = \frac{2}{T} \int_0^{T/2} v_i(t)i_i(t) \, dt = \frac{V_m I_1}{2}. \quad (3.1)$$

Let us denote the desired LED power by $P_d$. Utilizing (2.2) and (3.1), the desired instantaneous input current can be obtained as follows

$$i_{id}(t) = \frac{2P_d}{V_m} \sin \omega t = i_{idpeak} \sin \omega t. \quad (3.2)$$

From (3.2), for any desired LED power $P_d$, the amplitude of the desired input current $i_{idpeak}$ can be obtained. Hence, by controlling the amplitude of the first harmonic of the input current, the LED power can be kept under control.

### 3.1.2 Input Current Controller

A simplified configuration of the ac-dc flyback LED driver, when switch $S$ is ON/OFF, is shown in Fig. 3.2(a) and 3.2(b), respectively. In these figures, the voltage source $v_{ir}$ represents the voltage after the rectifier in Fig. 3.1. It is assumed that the switching frequency is much higher than the input voltage fundamental frequency. Therefore, the input voltage is almost constant during each switching cycle. Figure 3.3 shows the key waveforms of the ac-dc flyback LED driver in the discontinuous conduction mode. Consider the input current of the circuit as depicted in Fig. 3.3(d). In the time interval $kT_s < t \leq (k + d)T_s$, the input current is equal to the magnetizing current (see Fig. 3.3(c)), i.e., $i_{ir}(t) = i_{Lm}(t)$, with the rest of the switching cycle being zero as follows
Figure 3.3: Waveforms of ac-dc flyback LED driver for one cycle in discontinuous conduction mode. \( m_{ps} = \frac{N_p}{N_s} \), \( m_{as} = \frac{N_a}{N_s} \), and \( m_{ap} = \frac{N_a}{N_p} \) where \( N_p, N_s, \) and \( N_a \) are the primary, secondary, and auxiliary winding’s turn, respectively.

\[
\begin{align*}
\dot{i}_{ir}(t) &= \begin{cases} 
\frac{v_1}{L_m}(t - kT_s) & kT_s < t \leq (k + d)T_s, \\
0 & (k + d)T_s < t \leq (k + 1)T_s 
\end{cases} \\
&= \begin{cases} 
\frac{v_1}{L_m}(t - kT_s) & kT_s < t \leq (k + d)T_s, \\
0 & (k + d)T_s < t \leq (k + 1)T_s 
\end{cases}
\end{align*}
\]

(3.3)

where \( v_1 \) is the voltage across the magnetizing inductance when switch \( S \) is ON, \( L_m \) is the magnetizing inductance, \( k \) is the \( k^{th} \) switching cycle, \( d \) is the duty cycle of switch \( S \), and \( T_s \) is the switching period. Based on (3.3), the average input current in each switching cycle is given by

\[
< \dot{i}_{ir}(t) >_{T_s} = \frac{v_1 d^2}{2L_m f_s}
\]

(3.4)
Figure 3.4: ac-dc flyback LED driver with inner current and outer power control loops.

where \( f_s = 1/T_s \) is the switching frequency. Neglecting the drain-source voltage \( v_S \) of the switch \( S \) (see Fig. 3.2(a)) during the ON time, the voltage applied across the magnetizing inductance is equal to the rectified input voltage, i.e., \( v_1 \approx v_{ir} \). Considering (3.4), the peak amplitude of the rectified input current is given by

\[
I_1(t) = \frac{V_m}{2L_m f_s} d^2(t). \tag{3.5}
\]

From (3.5), it follows that the ac-dc flyback LED driver shown in Fig. 3.2 can be considered as a Single-Input Single-Output (SISO) system with duty cycle \( d(t) \) as the input and the peak amplitude of the rectified input current \( I_1(t) \) as the output. Thus the peak amplitude of the rectified input current can be controlled to follow the peak amplitude of the desired input current given by (3.2). Now, assuming a nominal model of the converter, the peak amplitude of the desired input current can be obtained if \( d(t) = D \). From (3.2) and (3.5) we have

\[
D = \sqrt{\frac{2L_m f_s i_{idpeak}}{V_m}} = \frac{2}{V_m} \sqrt{L_m f_s P_d}. \tag{3.6}
\]

To compensate parametric uncertainties and variations in \( i_{idpeak} \), let us take the duty cycle as follows

\[
d(t) = D + \hat{d}(t). \tag{3.7}
\]

By substituting (3.7) in (3.5) and assuming small variations for \( \hat{d}(t) \), the peak amplitude of the rectified input current can be linearized as follows

\[
I_1(t) \approx \frac{V_m D^2}{2L_m f_s} + \frac{V_m D}{L_m f_s} \hat{d}(t). \tag{3.8}
\]
Now, consider a PI controller given by

\[ \dot{e}(t) = K_P e_i(t) + K_I \int e_i(t) \, dt \] (3.9)

where \( K_P \) and \( K_I \) are the proportional and integral coefficients and \( e_i(t) = i_{id\text{peak}}(t) - I_1(t) \).

Substituting the error term \( e_i(t) \) and (3.9) in (3.8), and then utilizing (3.6), we have

\[ \left( K_P + \frac{L_m f_s}{V_m D} \right) e_i(t) + K_I \int e_i(t) \, dt = 0 \] (3.10)

from which

\[ e_i(t) = e_i(0) e^{-\frac{K_I}{K_P + \frac{L_m f_s}{V_m D}} t} \]. (3.11)

Since the coefficients of the PI controller are always positive, \( \frac{K_I}{K_P + \frac{L_m f_s}{V_m D}} \) is always positive, which guarantees that the error would exponentially converge to zero. It should also be noted that the above ratio specifies the convergence speed of the controller. By increasing this value reasonably, the controller is able to compensate the uncertainties caused by the nonlinear term given by (3.5). However, if the \( \frac{K_I}{K_P + \frac{L_m f_s}{V_m D}} \) ratio is increased too much it may have a destabilizing effect due to unmodeled dynamics.

### 3.2 LED Power Control

In this section, the LED power control concept is presented using the relationship between input current and LED power obtained in section 3.1.1. It should be noted that the LED power control loop has a lower sampling frequency than the input current control loop.
Figure 3.6: LED power as a function of LED current for different ambient temperatures for the following numerical values: $N = 10$, $E_0 = 100 \text{ lm/W}$, $V_{LEDO} = 3 \text{ V}$, $k_h = 0.85$, $k_v = -0.02$, $R_{jc} = 10 \degree \text{C/W}$, $R_{hs} = 6.3 \degree \text{C/W}$, and $T_0 = 20\degree \text{C}$.

3.2.1 LED Power Controller

Referring to (3.1), changing the peak amplitude of the input current $I_1$ will affect the average LED power $P_{LED}$. Considering (3.1), and assuming $P_{LED}$ and $I_1$ change slowly, we have

$$P_{LED}(t) = \frac{1}{2} V_m i_{idpeak}(t).$$  \hspace{1cm} (3.12)

The error between the LED power and its desired value is given by

$$e_p(t) = P_d - P_{LED}(t).$$  \hspace{1cm} (3.13)

By defining $i_{idpeak}(t)$ as follows

$$i_{idpeak}(t) = K_P e_p(t) + K_I \int e_p(t) \, dt$$  \hspace{1cm} (3.14)

where $K_P$ and $K_I$ are the proportional and integral coefficients of the PI controller in the power control loop, the LED power in (3.12) can be obtained as follows

$$P_{LED}(t) = \frac{V_m}{2} (K_P e_p(t) + K_I \int e_p(t) \, dt).$$  \hspace{1cm} (3.15)

By substituting $P_{LED}(t)$ from (3.13) into (3.15) and rearranging, we have

$$\left( K_P + \frac{2}{V_m} \right) e_p(t) + K_I \int e_p(t) \, dt = \frac{2P_d}{V_m}$$  \hspace{1cm} (3.16)
Figure 3.7: The single-stage ac-dc flyback LED driver prototype with proposed controller.

for which the error can be obtained as follows

$$e_p(t) = e_p(0)e^{-\frac{K_I}{K_P+\frac{1}{v_m}}t}.$$  \hfill (3.17)

Equation (3.17) reveals that by choosing $i_{idpeak}(t)$ based on (3.14), the error in (3.13) will exponentially converge to zero. The block diagram of the driver, current, and power controller are shown in Fig. 3.4.

### 3.2.2 LED Power Estimation

Consider the diode current depicted in Fig. 3.3(e). In the time interval $(k+d)T_s < t \leq (k+d+d_1)T_s$, where $d_1$ is the diode duty cycle, the diode current is equal to the magnetizing current mapped into the secondary side. Thus we have

$$i_D(t) = \begin{cases} 
    i_D((k+d)T_s) - m_{ps}^2 \frac{v_2}{L_m} (t-(k+d)T_s) & (k+d)T_s < t \leq (k+d+d_1)T_s, \\
    0 & kT_s < t \leq (k+d)T_s \text{ and } (k+d+d_1)T_s < t \leq (k+1)T_s \n\end{cases}$$ \hfill (3.18)

where $m_{ps} = \frac{N_p}{N_s}$ is the ratio between the primary and secondary windings of transformer $T$, and $v_2$ is the voltage across the magnetizing inductance mapped to the secondary side of transformer when diode $D$ is ON and assumed to be constant. The output power delivered to capacitor $C$ and the LED string, i.e., $p_o(t) = i_D(t)V_{LED}$ (see Figs. 3.2(b) and 3.3(f)), is
given by
\[ p_o(t) = \begin{cases} 
  p_o((k + d)T_s) - \frac{m^2_p s}{L_m} V_{LED} v_2 (t - (k + d)T_s) & (k + d)T_s < t \leq (k + d + d_1)T_s, \\
  0 & kT_s < t \leq (k + d)T_s \text{ and } (k + d + d_1)T_s < t \leq (k + 1)T_s
\end{cases} \]  

(3.19)

where \( p_o((k + d)T_s) = i_D((k + d)T_s)V_{LED} \), with \( V_{LED} \) denoting the LED voltage. Now, using (3.19) and \( p_o((k + d + d_1)T_s) = 0 \) as shown in Fig. 3.3(f), the output power at time \( t = (k + d)T_s \) is given by
\[ p_o((k + d)T_s) = \frac{m^2_p s V_{LED} v_2 d_1}{L_m f_s}. \]  

(3.20)

Now, consider \( p_o, d, \) and \( d_1 \) to be time-varying parameters. Hence, using (3.19), (3.20), and averaging \( p_o(t) \) over the switching period, the power delivered to the capacitor and LED is given by
\[ p_{oavg}(t) = < p_o(t) >_{T_S} = \frac{1}{2} p_o((k + d)T_s) d_1(t) \frac{m^2_p s}{2L_m f_s} (d_1(t)V_{LED})(d_1(t)v_2). \]  

(3.21)

Assuming that the storage capacitor \( C \) is large enough to absorb the ac component of the output power, the LED power \( P_{LED} \) is equal to the average output power in (3.21) at the fundamental frequency, i.e., \( 2f \), as follows
\[ P_{LED} = \frac{m^2_p s f}{L_m f_s} \int_{0}^{T/2} (d_1(t)V_{LED})(d_1(t)v_2) \, dt. \]  

(3.22)
To obtain the LED power using (3.22), $d_1(t)v_2$ and $d_1(t) V_{\text{LED}}$ are required. To this end, using the volt-second balance of the magnetizing inductance shown in Fig. 3.3(b), $d_1(t)v_2$ can be obtained as follows

$$d_1(t)v_2 = \frac{d(t)v_{ir}(t)}{m_{ps}}. \quad (3.23)$$

Neglecting the switch $S$ forward voltage when it is ON, $v_1(t) \approx v_{ir}(t)$. Referring to Fig. 3.2(b), (3.23) can be rewritten based on the LED voltage $V_{\text{LED}}$ as follows

$$d_1(t)(V_D + V_{\text{LED}}) = \frac{d(t)v_{ir}(t)}{m_{ps}} \quad (3.24)$$

where $V_D$ is the diode forward voltage when it is ON. Utilizing (3.24), we have

$$d_1(t)V_{\text{LED}} = \frac{d(t)v_{ir}(t)}{m_{ps}} - d_1(t)V_D. \quad (3.25)$$

From (3.25), it follows that by utilizing $d_1(t)$ from the auxiliary side, $v_{ir}(t)$ and $d(t)$ from the primary side, and $V_D$, the term $d_1(t)V_{\text{LED}}$ can be obtained. Note that $V_D$ is assumed to be constant during each switching cycle. Substituting (3.25) and (3.23) in (3.22), the LED power can be obtained.

The schematic diagram of the ac-dc flyback LED driver with the proposed input current and LED power controllers is shown in Fig. 3.5. The proposed LED power estimator circuit is also shown in this figure, consisting of the circuit with an auxiliary winding $N_a$ and firmware inside the MCU. Utilizing resistor $R_4$ and zener diode $Z_2$, the negative part of the
auxiliary winding voltage $v_a$ shown in Fig. 3.3(g) is clamped to zero and the positive part is scaled down to $V_Z$ (see voltage $v_+$ in Fig. 3.3(h)). Hence, to cancel the ripple in the time interval $(k + d + d_1)T_s < t < (k + 1)T_s$ shown in Fig. 3.3(h), the LM311 comparator is used with a constant and fixed $V_{th}$ applied to its negative pin. The output of the comparator is a PWM signal with the diode duty cycle representing the ON time of diode $D_3$. The comparator output is scaled down into a proper voltage range for the MCU using $R_8$ and $R_9$ resistors. The signal $v_{d1}$ is used to measure the ON time of diode $D_3$ using the built-in capture module (CAP) inside the MCU. Signals $v_{ir}$, $d$, and $d_1$ are required to estimate the LED power $P_{LED}$ as shown in Fig. 3.5.

### 3.2.3 Non-Auxiliary Winding LED Driver

Based on the results obtained in previous section, it is expected to measure the LED power only by the information from the primary-side. No information from the secondary or auxiliary winding is required. For high LED voltages where $V_D << V_{LED}$ (3.25) can be approximated as follows

$$d_1(t)V_{LED} \approx \frac{d(t)v_{ir}(t)}{m_{ps}}.$$  

(3.26)

Note that $V_D$ is assumed to be constant during each switching cycle. Substituting (3.26) and (3.23) in (3.22), the LED power can be obtained as follows
It shows that the average LED power can be obtained without using any information from the secondary or auxiliary sides that is the novelty of this method. It helps reducing the complexity, size, and cost of the controller and the whole LED driver. Figure 3.18 shows the measured LED power versus desired LED power and also the LED power error with the proposed method. As shown in this figure, the proposed LED power control and estimator are able to control the LED power with lower than 5% error for powers higher than 7 W.

### 3.3 Dependency of LED Power on Input Current and Ambient Temperature

In this section we obtain an expression for the LED power in terms of LED current and ambient temperature to discuss thermal characteristics of constant power LED control. To this end, let us consider the LED power as follows

\[ P_{\text{LED}} = V_{\text{LED}} I_{\text{LED}} \]  

(3.28)
where $I_{\text{LED}}$ is the LED current. From [71], the LED voltage in terms of its junction temperature approximated as follows

$$V_{\text{LED}} = V_{\text{LED0}}(1 + k_v(T_j - T_0)) \quad \text{for} \quad T_j \geq T_0 \quad \text{and} \quad V_{\text{LED}} \geq 0 \quad (3.29)$$

where $V_{\text{LED0}}$ is the rated LED voltage at the rated temperature $T_0$ and $k_v$ is the relative rate of reduction of LED voltage with increasing temperature. Furthermore, the junction temperature of an LED string is given by [15]

$$T_j = T_a + (R_{jc} + N R_{hs}) k_h P_{\text{LED}} \quad (3.30)$$

where $R_{jc}$ is the junction to case thermal resistance, $R_{hs}$ is the heat-sink thermal resistance, and $k_h < 1$ is a constant representing the portion of the LED power that turns into heat. Substituting (3.28) and (3.29) in (3.30) we have

$$T_j = T_a + (R_{jc} + N R_{hs}) k_h I_{\text{LED}} V_{\text{LED0}} (1 + k_v(T_j - T_0)). \quad (3.31)$$

Thus the junction temperature can be obtained as follows

$$T_j = \frac{T_a + (R_{jc} + N R_{hs}) k_h (1 - k_v T_0) V_{\text{LED0}} I_{\text{LED}}}{1 - (R_{jc} + N R_{hs}) k_h k_v V_{\text{LED0}} I_{\text{LED}}}. \quad (3.32)$$
Substituting (3.29) and (3.32) into (3.28), the LED power is given by

$$P_{LED} = \frac{(1 + k_v(T_a - T_0)) V_{LED0} I_{LED}}{1 - (R_{jc} + N R_{hs}) k_h k_v V_{LED0} I_{LED}}$$

(3.33)

which expresses the LED input power as a function of LED current and ambient temperature as illustrated in Fig. 3.6. Utilizing (3.33) we have (when $T_a > T_0$)

$$\frac{\partial P_{LED}}{\partial I_{LED}} = \frac{(1 + k_v(T_a - T_0)) V_{LED0}}{(1 - (R_{jc} + N R_{hs}) k_h k_v V_{LED0} I_{LED})^2} > 0,$$

$$\frac{\partial P_{LED}}{\partial T_a} = \frac{k_h V_{LED0} I_{LED}}{1 - (R_{jc} + N R_{hs}) k_h k_v V_{LED0} I_{LED}} < 0. \quad (3.34)$$

The above behavior can also be observed from Fig. 3.6. For instance, in response to an ambient temperature rise, the LED current increases for the case of a constant power controller. Thus, from (3.34) we have

$$\Delta P_{LED} \approx \frac{\partial P_{LED}}{\partial I_{LED}} \Delta I_{LED} + \frac{\partial P_{LED}}{\partial T_a} \Delta T_a. \quad (3.35)$$

Equations (3.34) and (3.35) can be used to analyze the change of LED current and power in constant power and constant current drives, respectively, in face of ambient temperature variations.
3.4 Experimental Verification

A prototype circuit was built as shown in Fig. 3.7 to drive Cree CR22-32L LED string consisting of 20 series LEDs. For the proposed primary-side LED power control scheme, digital implementation was done using TMS320F28027 microcontroller (MCU) from Texas Instruments.

The inductor discharge time, i.e., diode ON time, was obtained by using the auxiliary winding and its corresponding circuit. Figure 3.8 shows $v_+, V_{th}$, and $v_{d1}$ signals of the experimental circuit which indicates close agreement with Fig. 3.3(h) and 3.3(i). The duty cycle of signal $v_{d1}$ is measured using a built-in capture module CAP.

To achieve an accurate input current and LED power control, the frequency of analog to digital conversion was 5kHz to sample the 120Hz rectified input current. The maximum error occurs when the two samples are symmetrically located around $\frac{\pi}{2}$ radians. The percentage of the maximum error between the real peak current and the sampled value is about 0.07%.

Referring to Fig. 3.5, the rectified input voltage $v_{ir}$ was measured by using the $R_1 - R_2$ voltage divider. The peak value of the input voltage $V_m$ was provided to the controller using diode $D_1$ and capacitor $C_1$. The rectified input current $i_{ir}$ was obtained by using the $R_3 = 0.1\Omega$ and a low-pass filter consisting of $R_{10} = 1.8k\Omega$ and $C_2 = 100nF$. The cut-off frequency of this low pass filter is $f_c \approx 885Hz$. Note that the switching frequency is 30 kHz.
Figure 3.15: Effect of temperature on voltage, current, and illuminance of Cree CR22-32L LED string with proposed constant power control concept when $P_d = 14$ W. Illuminance reduction = $(9660 - 8930)/9660 \times 100 \approx 8\%$

Figures 3.9 and 3.10 present the LED voltage, LED current, input current, and input voltage of the ac-dc flyback LED driver when Cree CR22-32L LED string was connected to the output of the driver. As shown in these figures, the input current is in phase with the input voltage. Figure 3.9(a) and (b) show the input and output waveforms of the driver with 10W LED power and input voltage of 85 and 160\(V_{\text{rms}}\), respectively. Figure 3.10(a) and (b) show the same waveforms of the driver with a 25W LED power and input voltages of 85 and 160\(V_{\text{rms}}\), respectively. These figures show that the proposed controller is capable of delivering the desired power to the LED with different input voltage levels. We tested the proposed LED driver with other input voltage levels in the range 85-160\(V_{\text{rms}}\) and similar results were obtained.

Figures 3.11, 3.12, and 3.13 show the efficiency and power factor of the proposed LED driver with respect to the LED power with 85, 120, and 160 \(V_{\text{rms}}\) input voltage levels, respectively. For 85\(V_{\text{rms}}\) input voltage, a high input power factor and efficiency was achieved even for low LED powers as shown in Fig. 3.11. For higher powers, the input power factor is almost unity and the efficiency is greater than 80\%. However, by increasing the input voltage to 120\(V_{\text{rms}}\), the efficiency and power factor reduce as shown in Fig. 3.12. For LED
powers higher than 10W, these parameters are reasonably high. For a 160\textit{Vrms} input voltage, the efficiency and power factor reduce dramatically for low LED powers as shown in Fig. 3.13. This figure shows that for high input voltage levels the proposed LED driver is suitable for higher LED powers.

Figure 3.14 shows the LED power versus input voltages in the range of 85-160\textit{Vrms} which indicates the accuracy of the proposed controller in delivering the desired power with different input voltage levels for the whole input voltage range.

Figure 3.15 shows the effect of temperature on the LED voltage, current and illuminance in the range 25-80\textdegree C. This figure shows that the LED current increases about 24mA, the LED voltage reduces about 1.26V, and the LED light reduces by 8\% to keep the LED power constant on 14W. The temperature was changed using a Yamato Drying Oven DX402 and the LED illuminance was measured using the Extech Instruments Light Meter 401025. The same test was done using the 14W constant current LED driver RD195 from Power Integrations with the results shown in Fig. 3.16. As shown, the LED voltage reduces due to the temperature increase. Consequently, the LED power reduces when using the constant current technique. As a result, the output illuminance reduces about 13\% from its initial
value at 25°C. Comparing this test with the 14W test using the proposed constant power method (Fig. 3.15), it follows that the output illuminance in constant power drops 5% lower than the constant current case.

Referring to Fig. 3.15, the LED current increases about 24mA with the proposed constant power technique. Experimental results from [88] showed that the white LED system exhibited very little chromaticity shift (less than a 4-step MacAdam ellipse) when the light level was changed from 100% to 3% using both continuous current and PWM dimming schemes. It should be pointed out that the relative chromacity of certain LEDs shows very slight changes in the LED color versus LED current changes [84, 89–91]. As a result, with the LED current changes observed in Fig. 3.15, very slight change in the LED color will happen that is not perceivable by the human eyes.

Figure 3.17 shows that the proposed LED power control method is about 95% accurate for powers higher than 7W. Figure 3.19 shows the line current harmonics measured for 85, 120, and 160V_{rms} input voltage levels and 20W LED power. Thus the input current harmonics meet the class C minimum requirements with good margins.
3.5 Summary and Conclusions

In this chapter, a double-loop primary-side control structure was presented for a single-stage ac-dc flyback LED driver to regulate the LED power. The proposed controller was able to control the LED power on the primary-side using the proposed peak input current control method. Using this method, high input power factor along with LED power regulation with 95% accuracy for powers greater than 7 W was achieved. The proposed constant power control technique showed an 8% reduction of the LED output light when the ambient temperature increases from 25°C to 80°C. Experimental results validated the feasibility of the LED power control method.
Figure 3.19: Measured line current harmonics at $v_i = 120 \, V_{\text{rms}}$. 
Chapter 4

An Electrolytic Capacitor-less Flyback LED Driver Considering Light Flicker Characteristics

In this chapter, a comprehensive study is conducted on designing an electrolytic capacitor-less ac-dc buck-boost/flyback converter for driving High-Brightness LED strings. A relationship between the input current, LED light flicker, and the size of filter capacitors is obtained. Considering the ENERGY STAR standard for light flicker requirement, a procedure for obtaining the minimum value of filter capacitors is presented. It is concluded that relatively small filter capacitors, such as film or ceramic capacitors, can be chosen while meeting light flicker requirements. Thus a longer lifetime LED driver compared to a driver using electrolytic capacitors can be achieved. Experimental studies are presented for a 20W ac-dc buck-boost/flyback LED driver prototype on Cree CR22-32L and XLamp XP-G LED strings.

This chapter is organized as follows. The effect of input current and the output light flicker on the size of filter capacitors is studied in section 4.1. In this section, we present a quantitative relationship, relating the size of filter capacitors with the output light Flicker Index, or Percent Flicker measure. Section 4.2 illustrates how the above relationship can be utilized to reduce the size of the filter capacitors and obtain their minimum value. Section 4.3 presents experimental results for evaluating the validity of the proposed method.
4.1 Analytical Study of the Proposed Method

4.1.1 Effect of Input Current on Filter Capacitor Voltage

In this study, the ac-dc buck-boost/flyback topology shown in Fig. 4.1 is considered due to being able to provide a resistive input in the discontinuous conduction mode and a high input power factor. Denoting the input voltage and current as $v_i(t)$ and $i_i(t)$ as given by (2.1) and (2.2), respectively. Then, the instantaneous input power is given by

$$p_i(t) = \frac{1}{2} V_m I_m (1 - \cos 2\omega t).$$ (4.1)

Figure 4.2 shows the equivalent circuit of the LED driver shown in Fig. 4.1. In this figure, the voltage source $v_{ir}$ represents the voltage after rectifier in Fig. 4.1. The input power given by (4.1), is the power drawn from $v_{ir}$ and is delivered to the buck-boost equivalent resistance $R_B$ given by [36]

$$R_B = \frac{2L_B f_s}{D^2}$$ (4.2)

where $L_B$ is the buck-boost stage inductor, $f_s$ is the switching frequency, and $D$ is the quiescent value of duty cycle of switch $S$. In the equivalent circuit, the output of the buck-boost converter is modeled as a current source $i_B$ in parallel with its output filter capacitor $C_B$ [92]. Assuming that the efficiency of the buck-boost converter is 100%, the current source $i_B$ can be obtained by utilizing the converter’s power balance, i.e., $p_{oB}(t) = p_i(t)$, as follows

$$i_B(t) = M_B(t)i_{ir}(t)$$ (4.3)

where $M_B(t) = v_{ir}(t)/v_{CB}(t)$ and $i_{ir}(t)$ is the rectified input current. Since the ripple of voltage $v_{CB}(t)$ is much smaller than the input voltage variations then $M_B(t)$ can be

![Figure 4.1: Ac-dc integrated buck-boost/flyback LED driver.](image)
approximated using the average of this voltage, i.e., \( M_B(t) \approx v_{ir}(t)/V_{CB} \). Therefore, (4.3) can be rewritten as follows

\[
i_B(t) \approx \frac{v_{ir}(t)}{V_{CB}} i_{ir}(t) = \frac{p_i(t)}{V_{CB}} = I_b(1 - \cos 2\omega t) \tag{4.4}
\]

where \( I_b = \frac{V_m I_m}{2V_{CB}} \). Utilizing the equivalent circuit, the input of the flyback stage is modeled as a resistance given by

\[
R_F = \frac{2L_m f_s}{D^2} \tag{4.5}
\]

where \( L_m \) is the magnetizing inductance of the flyback transformer. To obtain the capacitor voltage \( v_{CB} \), the following differential equation needs to be solved

\[
\frac{dv_{CB}(t)}{dt} + \frac{1}{\tau_B} v_{CB}(t) = \frac{R_F}{\tau_B} i_B(t) \tag{4.6}
\]

where \( \tau_B = R_F C_B \). Solving (4.6) results in

\[
v_{CB}(t) = R_F I_b \left( 1 - \frac{2\omega \tau_B \sin 2\omega t + \cos 2\omega t}{1 + (2\omega \tau_B)^2} \right) + \text{exp}_{\text{decay}}(t) \tag{4.7}
\]

where \( \text{exp}_{\text{decay}}(t) \) is an exponentially decaying term. The time constant for the buck-boost stage is given by \( \tau_B \) which is approximately in the order of tens of milliseconds for \( L_m = 1 \) mH, \( f_s = 30 \) kHz, \( C_B = 10 \) \( \mu \)F, and small values of \( D \) to work in discontinuous conduction mode. Thus the exponential term in (4.7) decays quickly to zero. Furthermore, using the above values we have \( 2\omega \tau_B \gg 1 \), from which (4.7) can be simplified as follows

\[
v_{CB}(t) \approx R_F I_b \left( 1 - \frac{\sin 2\omega t}{2\omega \tau_B} \right) + \text{exp}_{\text{decay}}(t). \tag{4.8}
\]
Table 4.1: Flicker Index for Cree CR22-32L LED string with different $C$ and $C_B$ and $P_d = 10$ W.

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Using (4.8) and assuming that the steady-state is reached, i.e., the exponential term decay to zero, the input power to the flyback stage can be obtained as follows

$$p_iF(t) = \frac{v_{CB}^2(t)}{R_F} \approx R_F I_b^2 \left( 1 - \frac{\sin 2\omega t}{\omega \tau_B} \right)$$  \hspace{1cm} (4.9)

in which the second order term is canceled due to its small value. Using Fig. 4.2, the output of the flyback converter is modeled as a current source $i_D$ in parallel with the output filter capacitor $C$. Again, by assuming 100% efficiency for the flyback stage, the current source $i_D$ can be obtained by utilizing the converter’s power balance, i.e., $p_o(t) = p_iF(t)$, as follows

$$i_D(t) = M_F(t) i_iF(t)$$  \hspace{1cm} (4.10)

where $M_F(t) = v_{CB}(t)/v_C(t)$ and $i_iF(t)$ is the input current of the flyback stage. Since the ripple of voltage $v_C(t)$ is much smaller than the ripple of voltage $v_{CB}(t)$, $M_F(t)$ can be approximated using the average of voltage $v_C(t)$, i.e., $M_F(t) \approx v_{CB}(t)/V_C$. Therefore, the current source $i_D$ in (4.10) can be rewritten as follows

$$i_D(t) \approx \frac{v_{CB}(t)}{V_C} i_iF(t) = \frac{p_iF(t)}{V_C} = I_d \left( 1 - \frac{\sin 2\omega t}{\omega \tau_B} \right)$$  \hspace{1cm} (4.11)
Table 4.2: Percent Flicker (%) for Cree CR22-32L LED string with different $C_B$ and $P_d = 10$ W.

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<th>$C_B$ ($\mu$F)</th>
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where $I_d = \frac{R_F I_d^2}{V_C}$. In the equivalent circuit given by Fig. 4.2, the LED is modeled as a dc voltage source $V_{LED}$ in series with a resistance $r_{LED}$ [81]. To obtain the capacitor voltage $v_C(t)$, the following differential equation needs to be solved

$$\frac{dv_C(t)}{dt} + \frac{1}{\tau_F} v_C(t) = \frac{r_{LED}}{\tau_F} i_D(t) + \frac{1}{\tau_F} V_{LED}$$

(4.12)

where $\tau_F = r_{LED} C$. Solving (4.12) results in

$$v_C(t) = V_{LED} + r_{LED} I_d \left( 1 + \frac{2\omega \tau_F \cos 2\omega t - \sin 2\omega t}{\omega \tau_B (1 + (2\omega \tau_F)^2)} \right) + e^{\text{exp decay}}(t)$$

(4.13)

where $e^{\text{exp decay}}(t)$ is an exponentially decaying term. Equation (4.13) represents the effect of the first harmonic of the input current on the output capacitor voltage and will be used to describe the effect of this current on the output light flicker as will be discussed in the next section. From (4.12), the time constant $\tau_F \approx 0.01$ ms for $r_{LED} = 1\Omega$ and $C = 10\mu$F. Thus the exponential part of the capacitor voltage $v_C(t)$ decays quickly to zero.

4.1.2 Effect on Output Light Flicker

It is shown in chapter 2 that considering $v_C(t)$ across the LED string the output luminous flux is given by (2.15). Utilizing (2.24) and (2.26), the filter capacitance cannot be explicitly
Table 4.3: Flicker Index for Cree CR22-32L LED string with different $C$ and $C_B$ and $P_d = 20$ W.

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<th>$C_B$ ($\mu$F)</th>
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obtained, thus a numerical procedure is used in section 4.2. According to the IES Lighting Handbook, the Flicker Index is preferred over Percent Flicker. Combining Flicker Index and Percent Flicker creates a measure of light flicker. Based on ENERGY STAR standard, most traditional light sources occupy an area enclosed by a maximum Flicker Index of 0.13 and a maximum Percent Flicker of 40%, referred to as flicker frame of reference [63].

In the proposed control scheme, the input current follows the input voltage for achieving a unity power factor. Thus filter capacitor degradation does not have any effect on the input power factor and only affects the LED light flicker. However, capacitance degradation would result in the LED light flicker to increase. Hence, for the light flicker to meet the ENERGY STAR standard, the capacitor degradation effect has to be considered. The effect of capacitor degradation can be considered if the percentage of its degradation during its lifetime is known and added to the calculated capacitance in the minimization scheme.
Table 4.4: Percent Flicker (%) for Cree CR22-32L LED string with different $C$ and $C_B$ and $P_d = 20$ W.

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<th>$C_B$ ($\mu$F)</th>
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4.2 Filter Capacitance Minimization

4.2.1 Electrical and Photometric Characteristics of LED strings

In this chapter, Cree CR22-32L and XLamp XP-G LED strings were selected for experimental evaluation. To obtain the minimum required capacitors based on different LED powers and light flickers (i.e., solving (2.24) and (2.26)), the electrical and photometric characteristics of these LED strings given by $r_{\text{LED}}$, $V_{\text{LED}}$, $\alpha_1$ and $\alpha_2$ need to be obtained. The procedure for obtaining these parameters was described in chapter 2. Also, the photometric characteristics of these LED strings were obtained using LEDs’ datasheets that is given in the same chapter.

4.2.2 Filter Capacitance Minimization Algorithm

Assume that the filter capacitors $C$ and $C_B$ are unknown. Thus the output capacitor voltage in (4.13) is a function of $t$ and filter capacitors, i.e., $v_C(t, C, C_B)$. Consequently, the output luminous flux in (2.15) has the same independent variables, i.e., $\phi_v(t, C, C_B)$. As a result, (2.24) and (2.26) can be rewritten as follows
Figure 4.3: Schematic diagram of the ac-dc integrated buck-boost/flyback LED driver with input current and LED power controllers.

\[
\text{Flicker Index} = \frac{2f}{\phi_{\text{vavg}}} \int_{t_1}^{t_2} (\phi_v(t, C, C_B) - \phi_{\text{vavg}}) \, dt. \tag{4.14}
\]

For a given LED applied power, the effect of filter capacitors on the light flicker can be studied by utilizing (4.14) and (2.26) the following numerical procedure:

1. Obtain \( \phi_v(t, C, C_B) \) from (2.15) with known parameters, i.e., \( f, r_{\text{LED}}, V_{\text{LED}}, \alpha_1, \alpha_2, P_d, L_m, \) and \( L_B \).
2. Obtain \( \phi_v(t) \) by substituting a known \( C \) and \( C_B \).
3. Calculate \( \phi_{\text{vmin}}, \phi_{\text{vavg}}, \) and \( \phi_{\text{vmax}} \).
4. Calculate \( t_1 \) and \( t_2 \) by obtaining roots of \( \phi_v(t) - \phi_{\text{vavg}} \).
5. Calculate Percent Flicker and Flicker Index using (2.26) and (4.14), respectively.

The above algorithm was done for \( C \) and \( C_B = 10, 20, 30, 40 \) and \( 50 \mu \text{F} \) when \( P_d = 10\text{W} \) for Cree CR22-32L LED strings. The results are shown in Tables 4.1 and 4.2 for Flicker Index and Percent Flicker, respectively. For each power operating point, the capacitor voltage \( v_C(t) \) can be obtained using (2.29) which is then used in (2.28) to obtain \( r_{\text{LED}} \) for the minimization. In Table 4.1, the Flicker Index is below the maximum permissible, i.e., 0.13, for any capacitors in the table for both \( C \) and \( C_B \). Also, in Table 4.2, the Percent Flicker is below the maximum permissible, i.e., 40%, with the same capacitors and LED power. These results show that a larger filter capacitor leads to a lower light flicker as expected. The same procedure was followed for \( P_d = 20\text{W} \) with the results as shown in Tables 4.3 and 4.4 for Flicker Index and Percent Flicker, respectively. These results show that, with higher LED power, the light flicker is higher using the same filter capacitors. As a result, by
Table 4.5: Flicker Index for Cree XLamp XP-G LED string with different $C$ and $C_B$ and $P_d = 10$ W.

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<th>$C_B$ ($\mu$F)</th>
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choosing the operating point corresponding to the highest LED power (worst case scenario) and obtaining its minimum required capacitor, one can guarantee that for lower power levels the requirements for LED light flicker is satisfied. As shown in Table 4.3, by reducing the size of the filter capacitors the Flicker Index reaches above the maximum permissible value. In Table 4.4, the Percent Flicker is below the maximum permissible value for all capacitors specified in the table. To meet the light flicker standard, the filter capacitors by which both Flicker Index and Percent Flicker reach their maximum permissible values can be selected to be the minimum required capacitor. Based on these results, $C = 20 \mu$F and $C_B = 20 \mu$F are the smallest filter capacitors that are required to meet light flicker requirements. Thus, long lasting film or ceramic capacitors can be used.

The same calculation was done for $C$ and $C_B = 10, 20, 30, 40$ and $50 \mu$F when $P_d = 10$ and $20$ W for a Cree XLamp XP-G LED string. The results for $P_d = 10$W are shown in Tables 4.5 and 4.6 for Flicker Index and Percent Flicker, respectively. In these tables, both Flicker Index and Percent Flicker are below the maximum permissible values. For $P_d = 20$W, the results are shown in Tables 4.7 and 4.8 for Flicker Index and Percent Flicker, respectively. Thus the Flicker Index and Percent Flicker are below the maximum permissible values for all capacitors in the tables for both $C$ and $C_B$. These results show that for higher powers, both
Table 4.6: Percent Flicker (%) for Cree XLamp XP-G LED string with different $C$ and $C_B$ and $P_d = 10$ W.

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Flicker Index and Percent Flicker are higher. Therefore, choosing the minimum required capacitor, i.e., $10\mu$F, based on the higher power would guarantee that light flicker measures are satisfied for lower power levels.

4.3 Experimental Results

A schematic diagram of the ac-dc buck-boost/flyback LED driver utilized in this chapter is shown in Fig. 3.5. A 20W prototype of the power circuitry is as shown in Fig. 4.4. For the controller, LED power regulation using input current control method proposed in [93] was utilized in which the LED power is regulated in a flyback LED driver. Due to input behavior similarity between flyback and buck-boost/flyback converters, the same control concept is used in this study. The controller implementation was achieved using digital circuitry. A TMS320F28027 MCU from Texas Instruments Inc. was used for controller implementation.

Based on the results obtained in the section 4.2, two $20\mu$F capacitors were used for $C$ and $C_B$, in the power driver circuitry to drive Cree CR22-32L LED string. In this way, the Flicker Index and Percent Flicker for 10W and 20W were measured every 5 minutes during 1 hour of operation of the circuit and plotted in Figs. 4.5 and 4.6, respectively. Figure
Table 4.7: Flicker Index for Cree XLamp XP-G LED string with different $C$ and $C_B$ and $P_d = 20$ W.

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4.5 shows that for a power of 10W, the Flicker Index is about 0.11 and the Percent Flicker is about 27%, which are in good agreement with the results in section 4.2.2 and meet the ENERGY STAR standard. Also, Fig. 4.6 shows that for a power of 20W, the Flicker Index and Percent Flicker are approximately 0.12 and 30%, respectively, which are again in good agreement with numerical results in section 4.2.2.

The same tests were performed on a Cree XLamp XP-G LED string with two 10µF capacitors used for $C$ and $C_B$ in the power drive circuitry as discussed in section 4.2.2. The Flicker Index and Percent Flicker were measured for 10W and 20W and the results were updated every 5 minutes during 1 hour of operation of the circuit as shown in Figs. 4.7 and 4.8, respectively. These figures show that for a power of 20W, the Flicker Index and Percent Flicker are approximately 0.13 and 28%, respectively, which are in good agreement with the numerical results obtained in section 4.2.2. Also, the light flicker for 10W is lower than for 20W as shown in Fig. 4.7.

4.4 Conclusion

In this chapter, a comprehensive study was conducted on minimizing the size of the filter capacitors of an ac-dc buck-boost/flyback LED drive circuitry based on ENERGY STAR
Table 4.8: Percent Flicker (%) for Cree XLamp XP-G LED string with different $C$ and $C_B$ and $P_d = 20$ W.

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light flicker measures, i.e., Flicker Index and Percent Flicker. A relationship between filter capacitors and these measures was obtained utilizing small signal model of the converter. Furthermore, a numerical procedure was proposed to obtain the light flicker measure using circuit and LED parameters with different filter capacitors. It was found that the obtained capacitors were in the range of long lifetime capacitors such as film or ceramic ones. A 20W electrolytic capacitor-less LED driver was developed and its performance was verified through experimental studies which validated the analytic developments.
Figure 4.4: Prototype ac-dc integrated buck-boost/flyback LED driver with the proposed controller.

Figure 4.5: Flicker Index and Percent Flicker (%) vs. time obtained from experimental setup for Cree CR22-32L LED string when $C = 20 \ \mu F$, $C_B = 20 \ \mu F$ and $P_d = 10 \ \text{W}$. 
Figure 4.6: Flicker Index and Percent Flicker (%) vs. time obtained from experimental setup for Cree CR22-32L LED string when $C = 20 \mu F$, $C_B = 20 \mu F$ and $P_d = 20 W$.

Figure 4.7: Flicker Index and Percent Flicker (%) versus time obtained from experimental setup for Cree XLamp XPG LED string when $C = 10 \mu F$, $C_B = 10 \mu F$ and $P_d = 10 W$. 
Figure 4.8: Flicker Index and Percent Flicker (%) vs. time obtained from experimental setup for Cree XLamp XPG LED string when $C = 10 \, \mu F$, $C_B = 10 \, \mu F$ and $P_d = 20 \, W$. 
Chapter 5

Summary, Conclusions, and Suggestions for Future Work

5.1 Summary and Conclusions

In this work, the development of a novel electrolytic capacitor-less LED driver was studied and evaluated using experimental prototype. To this end, first a comprehensive study was conducted on relationship between electrical and photometric characteristics of an ac-dc flyback LED system. Then an iterative algorithm was proposed to obtain its minimum required filter capacitor. The performance of the proposed algorithm in terms of input power factor correction and LED light flicker was verified using simulation model and experimental setup. Second, constant power control technique was proposed for flyback LED driver to alleviate the effect of ambient temperature on LED light. In this way, a novel primary-side LED power estimator is presented that does not need to any information from the secondary or auxiliary windings. Then, a double-loop primary-side control structure was presented to achieve power factor correction and LED power regulation. The performance of the proposed primary-side LED power estimator and controller was experimentally verified. Using the proposed controller, LED power regulation with 95% accuracy for powers greater than 7 W was achieved. The proposed constant power control technique shows an 8% reduction of the LED light when the ambient temperature increases from 25°C to 80°C compare to 13% reduction in constant current control technique. Third, the proposed filter capacitor minimization algorithm was extended and applied to an ac-dc buck-boost/flyback LED driver.
Using this iterative algorithm, the minimum required filter capacitances was obtained. It was found that these capacitances were in the range of long lifetime capacitances, i.e., film or ceramic capacitors. Finally, a prototype of an electrolytic capacitor-less LED driver was implemented. The experimental results showing the LED light flicker, demonstrated the accuracy of the proposed theory.

5.2 Future Research

5.2.1 Implementing the constant power control concept for Continuous Conduction Mode (CCM)

The scope of this study was to design and implement an LED driver for low power applications. In this way, flyback converter operating in DCM is preferred by which high power factor can be achieved. The controller proposed in this study works in DCM and that the maximum deliverable power to the LED is limited. Therefore, for high power applications LED driver operating in CCM would be worthwhile studying in any future research.

5.2.2 Zero-voltage and zero-current switching

High switching frequency in power converters usually results in high switching losses. Soft switching techniques such as zero-voltage switching (ZVS) or zero-current switching (ZCS) are usually adopted to reduce these losses. In these techniques, the switching frequency is not constant, therefore, the proposed modeling and its controller does not hold true. Implementing the constant power control concept with soft switching techniques can be an opportunity for future work.

5.2.3 Implementing the constant power control concept for buck-boost/flyback LED driver

The constant power control concept proposed in chapter 3 can be applied to other LED driver topologies. Also, in chapter 4 an electrolytic capacitor-less LED driver is proposed. Combining these two concepts together to have an electrolytic capacitor-less constant power LED driver brings another opportunity for future work.
5.2.4 Constant light LED driver

The constant power control method proposed in this work alleviated the LED light reduction due to the ambient temperature increase. For applications in which tight LED light regulation is needed, extending the proposed method as a constant light LED driver is worth to be a subject of future work. In this way, measuring the LED light is a challenge.
Chapter 6

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