

Article

Design and Implementation of 150 W AC/DC LED Driver with Unity Power Factor, Low THD, and Dimming Capability

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Abstract: This paper presents the implementation of a two-stage light-emitting diode (LED) driver based on commercial integrated circuits (IC). The presented LED driver circuit topology, which is designed to drive a 150 W LED module, consists of two stages: AC-DC power factor correction (PFC) stage and DC/DC power converter stage. The implementation of the PFC stage uses IC NCP1608, which uses the critical conduction mode to guarantee a unity input power factor with a wide range of input voltages. The DC/DC power converter with soft-switching characteristics for the entire load range uses IC FLS2100XS. Furthermore, the design of an electromagnetic interference (EMI) filter for the LED driver and the dimming control circuit are discussed in detail. The hardware prototype, an LED lighting system, with a rated power of 150 W/32 V from a nominal 220 V/50 Hz AC voltage supply was tested to show the effectiveness of the design. The presented LED driver was tested for street lighting, and the experimental results show that the power factor (PF) was higher than 0.97, the total harmonics distortion (THD) was lower than 7%, and the efficiency was 91.7% at full load. The results prove that the performance of the presented LED driver complies with the standards: IEC61000-3-2 and CIRSP 15:2009.

Keywords: LED driver; LLC resonant converter; EMI filter; power factor correction

1. Introduction

As compared to conventional lighting solutions, the most important advantage of LEDs is their long lifespan. Furthermore, LEDs provide some more advantages such as energy efficiency, improved safety, and a great color rendering index. In addition, LEDs are gaining popularity as a green solution. Recently, with the rapid development of LED technology, LEDs are widely used in automotive electronics [1], street lighting [2], and LCD backlighting [3]. Recently, medium- and high-power LEDs have been used to replace traditional high-pressure sodium (HPS) lamps for street lighting applications. The use of LEDs for street lighting applications is a good solution to save electric energy, and the LED driver has become an attractive topic for much research [4]. In recent years, there have been some issues that researchers have concentrated on, to develop LED drivers for street lighting applications to improve input performance [5–7], reduce the size [8], and improve the system efficiency [9].

There are three types of LED driver systems: Single-stage [10,11], two-stage [12–14], and integrated LED drivers [15,16]. The single-stage LED driver is an AC/DC converter that provides a constant output current for the LED and a unity power factor. The two-stage system consists of two separate stages, an AC/DC power factor correction (PFC) converter and a DC/DC converter. Recently, some integrated topologies were introduced in order to reduce the size and cost. Single-stage and integrated solutions have some advantages such as low-cost designs and high efficiency due to only one energy conversion. However, the PFC of these topologies causes a high-output voltage ripple due to the absence of electrolytic capacitors. Hence, the low-frequency current ripple makes LEDs flicker. These single-stage and integrated solutions are not suitable LED street lighting systems and are instead used for LED-based replacement lamps where small-size converters are needed. Furthermore, the bridgeless topologies usually generate high electromagnetic interference (EMI) due to large dv/dt and di/dt . The two-stage topology is the most popular topology for LED drivers above 100 W [17]. In the two-stage LED driver, the constant output current regulation is more easily implemented due to the bus voltage. Additionally, an isolated DC-DC stage is needed in the LED driver in order to ensure constant current control.

Recently, there have been several commercial analog integrated circuits (ICs) available on the market that are recommended for a high power factor and low cost with a high efficiency. These commercial ICs fulfil the IEC 1000-3-2 Class C, but the total harmonics distortion (THD) is higher than 20% [18]. In reference [19], a resonant controller IC L6598 is used to control a single-stage AC-DC converter with the rated power of 100 W for a street lighting system. The experimental results in [19] show that the power factor (PF) is 0.95 and the THD is 20.5%.

In this paper, a 150 W two-stage LED driver solution and its complete design for street lighting applications using commercial ICs NCP1608 and FLS2100XS are presented. The unity power factor is achieved by a boost converter with PFC while galvanic isolation is provided by an isolated transformer in an LLC resonant DC-DC stage. This topology of the LED driver can be selected according to requirements, including PF, THD, galvanic isolation, efficiency, and dimmable control. Furthermore, in this paper, the design of the EMI filter is based on the measurement and analysis of the conducted EMI in the LED driver in the GHz frequency range. The high-frequency content of the current generated by the boost PFC converter and DC-DC converter may be beyond some EMC standards limits such as CISPR 15:2009 [20–22]. In order to validate the presented design, the performance of the driver is tested under full and light load conditions for a universal AC power supply.

The rest of this paper is organized as follows: The operating LED driver is presented briefly in Section 2. In Section 3, the design guideline of a two-stage converter and EMI filter are presented in detail. The LED dimming is implemented and presented in Section 3. Experimental results and some conclusions are provided in Sections 4 and 5, respectively.

2. Principle Operation of Two-Stage LED Driver

The presented converter for the high-power LED is based on two power stages, namely, a boost PFC AC/DC converter and an LLC resonant converter, depicted in Figure 1. The function of the LED driver is to provide a constant current to the LED load. As illustrated in Figure 1, the system consists of an EMI filter, a PFC circuit for improving the input PF, and a half-bridge DC/DC converter for building output voltage. There are five blocks in this LED driver topology.

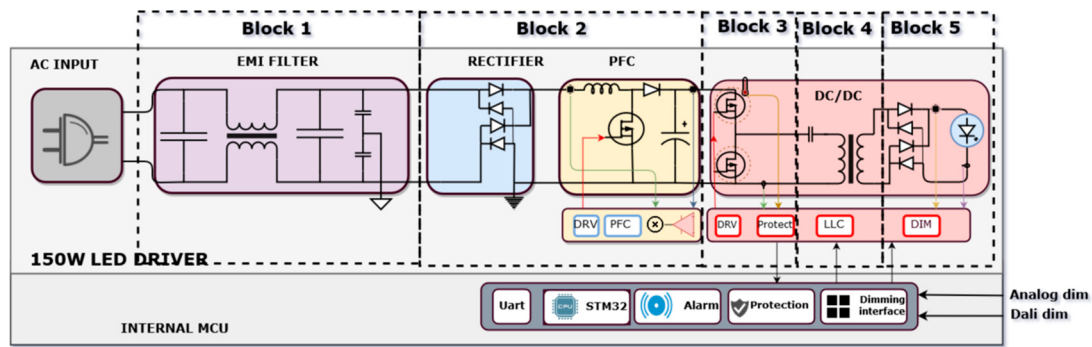


Figure 1. Two-stage LED driver topology with dimmable control.

Block 1: Any LED driver connected to the AC mains supply has to meet the requirement of EMI noise. Therefore, the EMI filter is needed at the input side of the LED driver.

Block 2: A PFC block is used to obtain a unity power factor with a wide input voltage range from 85 to 265 VAC. Besides, the PFC converter is mandatory in all AC-DC topologies connected to the line in order to comply with the IEC 61000-3-2 standard.

Block 3: A half-bridge converter block makes a square wave voltage from a 400 VDC input voltage and 50% duty ratio.

Block 4: A resonant tank block contains a capacitor and the magnetizing inductance of a transformer. The resonance tank operates with high frequency and creates a sinusoidal current from the square wave voltage from Block 3.

Block 5: A diode rectifier at the output side to generate the DC output voltage.

2.1. Principle Operation of Boost PFC Converter

Most DC power supplies use an uncontrolled rectifier and a storage capacitor to generate a DC voltage from an AC source. This causes a non-sinusoidal current consumption and increases the stress on the power delivery infrastructure. Active PFC circuits are the most popular method to eliminate input current harmonics. The system solution consists of a boost converter between the rectifier bridge and the bulk capacitor, as shown in Figure 2. The boost PFC stage comprises a diode bridge rectifier ($D_{R1}, D_{R2}, D_{R3}, D_{R4}$), a boost inductor (L_B), a boost diode (D_B), and a power switch (S_{PFC}). A bulk capacitor (C_{bulk}) is used to reduce DC-link voltage ripple. The functions of the boost PFC in the LED driver are as follows: Generate an output voltage equal to 400 VDC, which is higher than the input voltage, and force the average inductor current to track the reference current so that it has the same shape as the input voltage.

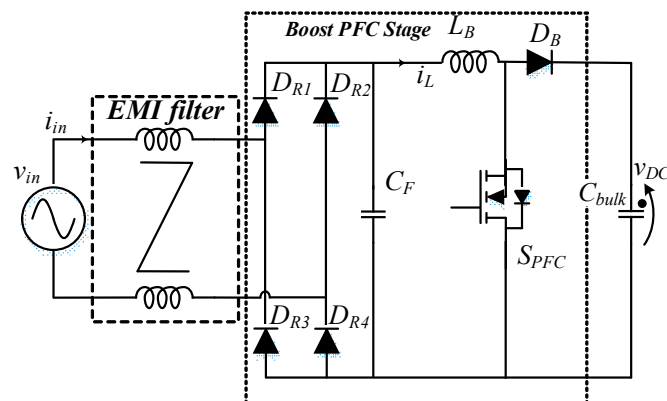


Figure 2. Boost power factor correction (PFC) stage.

Although the PFC can be achieved by several topologies, the boost converter, as shown in Figure 2, is the most popular topology, for the following reasons: The line voltage varies from zero to some peak value, typically 375 VDC (corresponding to maximum input voltage of 265 VAC); hence, a step-up converter is needed to generate a DC output voltage (V_{DC}) of 400 VDC. The boost converter has a filter inductor on the input side, which provides a smooth continuous input current waveform. A continuous input current is much easier to filter, which is a major advantage of this design due to the low cost. The boost converter can operate in three modes: Continuous conduction mode (CCM), discontinuous conduction mode (DCM), and critical conduction mode (CrCM). By comparing the DCM among the others, DCM operation seems simpler than that of the CrCM, as it may operate in constant frequency operation; however, the DCM has the disadvantage that it has the highest peak current compared to the CrCM and also to the CCM, without any performance advantage compared to the CrCM. For that reason, the CrCM is more a common practical design than the DCM.

Figure 3 shows the key waveforms of the boost PFC converter in the case of the CrCM. To obtain a unity power factor, the AC input voltage (v_{in}) is measured to provide a reference for inductor current. In the CrCM mode, the inductor current ripple ($I_{L(peak)}$) is twice the sinusoidal input current (i_{in}), and the switching frequency of the power switch is variable with constant on-time (T_{ON}), as shown in Figure 3.

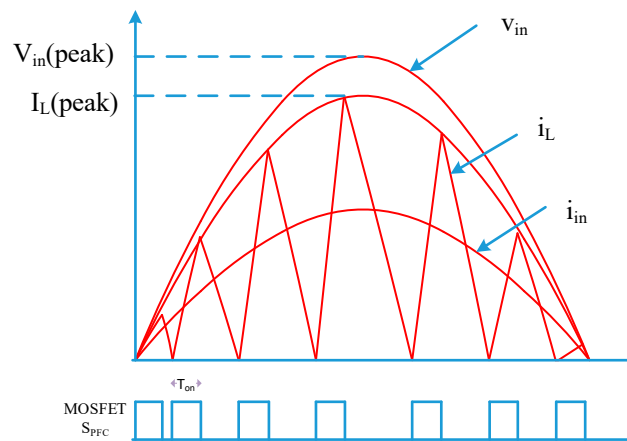


Figure 3. Key waveforms of the input voltage (v_{in}), input current (i_{in}), and inductor current (i_L) of the boost PFC.

2.2. Principle Operation of LLC Resonant Converter

Resonant DC/DC converters, which were first introduced in the 1980s [23], provide a low switching loss due to the enabling of resonant topologies to operate at high switching frequency. A half-bridge resonant converter has many advantages, such as high efficiency, high power density, and low EMI. In general, the LLC resonant converter consists of three blocks, which are a square-wave generator, resonant network, and rectifier network. The LLC resonant converter is shown in Figure 4, where L_M is the magnetizing inductance that acts as a shunt inductor, L_R is the series resonant inductor, and C_R is the resonant capacitor. The rectifier network includes a transformer and two diodes, D_1 and D_2 . The square-wave generator produces a square-wave voltage (V_{SQ}) from the DC power supply V_{DC} by driving switches S_1 and S_2 . The duty cycle of the two switches is 50%. In order to avoid the short circuit in the half-bridge converter, a dead-time is inserted in the switching state of S_1 and S_2 . In this circuit, only the fundamental component of the current is allowed to flow through the resonant network.

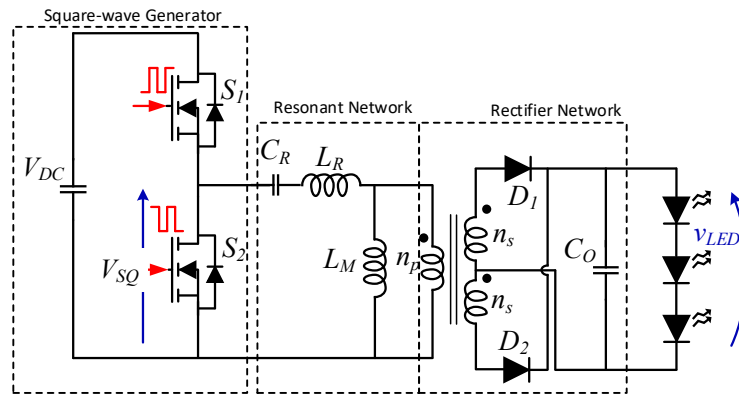


Figure 4. Half-bridge LLC resonant converter topology.

Figure 5 shows the equivalent circuit of the LLC half-bridge converter based on the first-harmonic approximation (FHA) method, where R_{AC} is the equivalent load resistance. There are two resonant frequencies (f_{r1}, f_{r2}), as shown in Equations (1) and (2), for the LLC half-bridge converter [24]:

$$f_{r1} = \frac{1}{2\pi \sqrt{L_R C_R}}, \quad (1)$$

$$f_{r2} = \frac{1}{2\pi \sqrt{(L_R + L_M) C_R}}. \quad (2)$$

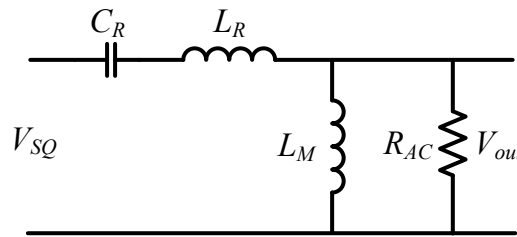


Figure 5. Simplified converter circuit of the LLC half-bridge converter.

The AC-equivalent load resistance (R_{AC}) can be expressed by Equation (3) [25]:

$$R_{AC} = \frac{8n^2}{\pi^2} R_{LED}, \quad (3)$$

where $n = \frac{n_p}{n_s}$ is the transformer ratio (n_p and n_s are the number of turns of the primary and secondary coil, respectively) and R_{LED} is the LED load.

From the equivalent model, as shown in Figure 5, the well-known voltage gain of the circuit is described by Equation (4) [25–27]:

$$M = \frac{V_{out}}{V_{SQ}} = \left| \frac{j\omega L_M \parallel R_{AC}}{j\omega L_M \parallel R_{AC} + j\omega L_R + \frac{1}{j\omega C_R}} \right| = \left| \frac{m \cdot f_n^2}{[(m+1) \cdot f_n^2 - 1] + j[(f_n^2 - 1) \cdot f_n \cdot Q \cdot m]} \right|, \quad (4)$$

where $m = \frac{L_M}{L_R}$ is the ratio of magnetizing inductance and leakage inductance, $Q = \frac{\sqrt{L_R/C_R}}{R_{AC}}$ is the quality factor, $f_n = \frac{f_{sw}}{f_0}$ is the ratio of the switching frequency and resonant frequency, and ω is the angular frequency.

The output voltage of the LLC half-bridge converter is achieved as follows:

$$V_{LED} = M \frac{n_s}{n_p} \frac{V_{DC}}{2}. \quad (5)$$

The behavior of voltage gain as a function of m and Q in Equation (4) can be plotted, as shown in Figure 6. The two resonant frequencies, as shown in Equations (1) and (2), divide the frequency range into three regions: Region 1 is the zero voltage switching (ZVS) region in the right half-plane of f_{r1} ; Region 2 is the ZVS region between f_{r1} and f_{r2} ; and (3) Region 3 is the zero current switching (ZCS) in the left half-plane of f_{r2} .

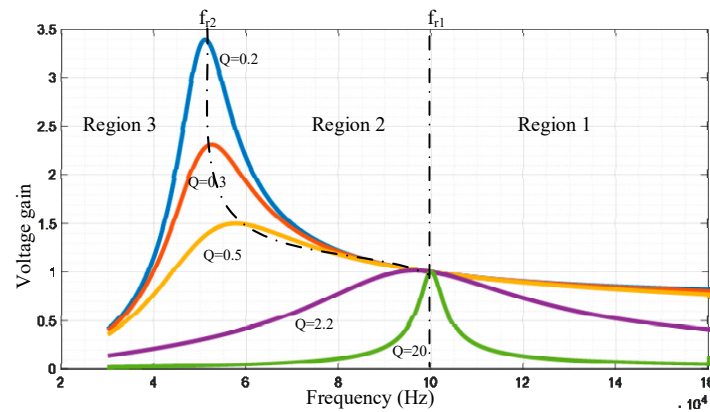


Figure 6. Family of voltage gain according to the quality factor Q .

The waveforms and equivalent circuits of the LLC resonant converter working in region 2 are presented in Figure 7a,b, and the operation principle is described as follows:

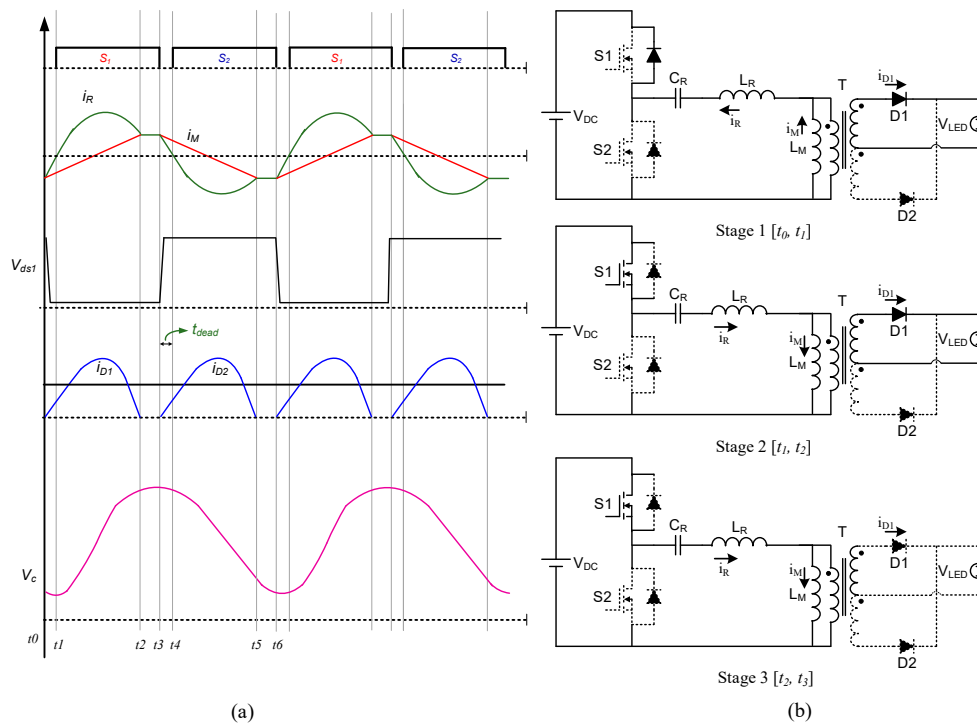


Figure 7. (a) Operating waveforms and (b) equivalent circuit.

Stage 1 [$t_0 - t_1$]: This stage begins when both the two switches S_1 and S_2 are in the off state at time t_0 , and the resonant current discharges the parasitic capacitor and then flows through the free-wheeling diode of S_1 . The input power does not transfer to the output side, and the output rectifier diode, D_1 , continues to transfer power to the load by the output capacitor. In this stage, C_R and L_R resonate because the voltage across L_M is clamped at the reflected output voltage.

Stage 2 [$t_1 - t_2$]: This stage begins when S_1 is turned on at t_1 , and S_2 is still in the off state. S_1 turns on with the ZVS condition. The current drops to 0 A before reversing and flowing through the switch, S_1 . The inductor L_M is clamped by the secondary voltage and does not engage resonance, and the magnetizing current i_M linearly increases. The load current is proportional to the difference between the resonant current, i_R and i_M .

Stage 3 [$t_2 - t_3$]: This mode begins when the resonant current i_R becomes positive. At t_2 , the load current drops to zero due to the resonant current, i_R , equaling i_M , and causes D_1 to turn off. As the switching period is bigger than the $L_R \cdot C_R$ resonant period, S_1 continues to conduct until t_3 .

3. Design Procedure of the Two-Stage LED Driver

In order to verify the correction of the previous analysis in Section 2, a two-stage LED driver is implemented by a combination between the boost PFC converter, as shown in Figure 2, and the LLC resonant half-bridge converter, as shown in Figure 4. The driver is designed for driving a 150 W LED with output voltage 32 VDC. Detailed calculation of the AC/DC and DC/DC stages, EMI filter, and dimming control module are presented as follows.

3.1. Boost-PFC Stage Design

In this paper, the boost PFC is implemented using a NCP1608BDR2G controller [28]. The advantages of this IC are a near-unity power factor, wide control range, for high-power application, and no input voltage sensing requirement. Furthermore, it works safely due to it providing over voltage, over current, and under voltage protection. This IC operates in the critical conduction mode to ensure a near-unity power factor across a wide range of input voltages and output power. It is also an active power factor correction controller specifically designed for a medium power converter. The design specification of the boost PFC converter is shown in Table 1.

Table 1. Parameters of boost PFC stage.

Boost PFC Design Parameters		
Input voltage	220 V	V_{in}
Frequency	50 Hz	f_{in}
Output voltage	400 V	V_{DC}
Input power	150 W	P
Minimum Efficiency	92%	η_{min}
Minimum switching frequency	40 kHz	$f_{sw(min)}$

The boost inductor L_B is designed to be operated in the CrCM. Here, V_{DC} is the DC-bus voltage, V_{in} is the input voltage, and f_{sw} is the switching period of the power switch S_{PFC} .

$$L_B = \frac{V_{in}^2 \cdot \left(\frac{V_{DC}}{\sqrt{2}} - V_{in} \right) \cdot \eta_{min}}{\sqrt{2} \cdot V_{DC} \cdot P \cdot f_{sw(min)}}. \quad (6)$$

The inductor value of the L_B at maximum input voltage (265 V) is calculated as

$$L_B = \frac{265^2 \cdot \left(\frac{400}{\sqrt{2}} - 265 \right) \cdot 0.92}{\sqrt{2} \cdot 400 \cdot 150 \cdot 40 \cdot 10^3} = 339 \mu\text{H}. \quad (7)$$

In order to guarantee that the switching frequency is higher than 40 kHz, the boost inductor is chosen at 250 μ H. Due to the tolerance of the inductor, the maximum inductance of the boost inductor is 300 μ H. The minimum frequency at full load is

$$f_{sw(\min)} = \frac{V_{in}^2 \cdot \eta_{\min}}{2 \cdot L_{B\max} \cdot P} \cdot \left(1 - \frac{\sqrt{2} \cdot V_{in}}{V_{out}}\right) = \frac{265^2 \cdot 0.92}{2 \cdot 300 \cdot 10^{-6} \cdot 150} \cdot \left(1 - \frac{\sqrt{2} \cdot 265}{400}\right) = 45.3 \text{ kHz}. \quad (8)$$

The IC NCP1608 has the over voltage function to prevent the output from exceeding a safe voltage. The threshold voltage for over voltage detection is 421 V. Therefore, the bulk capacitor is designed to ensure that the ripple of the DC-bus voltage is smaller than 42 V.

Hence,

$$C_{bulk} \geq \frac{P_{out}}{2 \cdot \pi \cdot V_{ripple} \cdot f_{in(\min)} \cdot V_{out}} = \frac{150}{2 \cdot \pi \cdot 42 \cdot 47 \cdot 400} = 30 \text{ } \mu\text{F}. \quad (9)$$

In this design, the value of the bulk capacitor is selected at 100 μ F.

Table 2 summarizes the device list of the boost PFC stage.

Table 2. Device list of boost PFC stage.

Boost PFC Design Specification	
Diode Rectifier ($D_{R1}, D_{R2}, D_{R3}, D_{R4}$)	TS15P05GC2 (Taiwan Semiconductor)
Boost Inductor L_B	300 μ (Core PQ2625)
Mosfet S_{PFC}	STF13NM60N (STMicroelectronics)
Boost Diode D_B	STTH15R06FP (STMicroelectronics)
C_{bulk}	100 μ F/450V (Nichion)
IC	NCP1608BDR2G (Onsemi)

3.2. LLC Resonant DC-DC Converter Design

In this section, the step-by-step consideration of the design of the LLC resonant converter will be discussed. The design of the LLC resonant converter implemented based on IC FLS2100XS [29] is described as follows:

Step 1: Define the converter design specification. The LLC converter electrical specifications are given in Table 3.

Table 3. Parameters of LLC resonant DC-DC converter.

LLC Resonant Half-Bridge Design Parameters		
Half-bridge input voltage	390–410 V	V_{DC}
Switching frequency	100–120 kHz	f_{sw}
Output voltage	32 V	V_{LED}
Output current	4.7 A	I_{LED}

Step 2: Determine the transformer turns ratio. The transformer turns ratio is determined by Equation (10):

$$M = \frac{2nV_{LED}}{V_{in}} \Rightarrow n = M \frac{V_{in}/2}{V_{LED}}. \quad (10)$$

To achieve a high operating efficiency, the switching frequency is chosen at near the resonant frequency. The voltage gain of the LLC converter is chosen from 1 to 1.4. From Equation (10), the turns ratio is $n = 6.3 \rightarrow 8.96$. Therefore, we choose the transformer turns ratio n as 8.75 (35:4:4), which is available in [30].

Step 3: Choose the transformer. In this research, the transformer core type is chosen as ETD-34 of Würth Elektronik. The primary inductance and the leakage inductance of this transformer are

$$L_P = 600 \mu\text{H}, L_R = 100 \mu\text{H}, L_M = 500 \mu\text{H}. \quad (11)$$

Step 4: Determine the equivalent load resistance:

$$R_{AC} = \frac{8n^2}{\pi^2} R_{LED} = 291.2 \Omega. \quad (12)$$

Step 5: Calculate the inductance ratio. From step 3, the inductance ratio m is determined as

$$m = \frac{L_P}{L_R} = 6. \quad (13)$$

Step 6: Calculate the resonant circuit parameters. As introduced in Section 2, the working frequency f_s must be lower than the resonant frequency in order to keep the converter working in region 2. In this design, the switching frequency is chosen from 100 to 150 kHz and the resonant frequency is set to be $f_r = 150$ kHz.

$$C_R = \frac{1}{2\pi Q f_r R_{AC}} = \frac{1}{4\pi^2 \cdot f_r^2 \cdot L_R} = 11.2 \text{ nF}. \quad (14)$$

Therefore, the resonant capacitor is selected as $C_R = 10$ nF.

$$Q = \frac{\sqrt{L_R/C_R}}{R_{AC}} = 0.32. \quad (15)$$

Step 7: Verify the resonant-circuit design.

From the above selected parameters, the resonant frequency is recalculated as

$$f_r = \frac{1}{2\pi \sqrt{L_R C_R}} = 159 \text{ kHz}. \quad (16)$$

Figure 8 shows the voltage gain characteristics of the LLC converter according to the different quality factor Q in the case of $m = 6$ and resonant frequency $f_r = 159$ kHz. It can be seen that, with the switching frequency of 100 kHz and the quality factor $Q = 0.32$, the voltage gain is 1.4. From Equation (10), the output voltage of the power converter is verified as Equation (17).

$$V_{LED} = \frac{MV_{in}}{2n} = \frac{1.4 \cdot 400}{2 \cdot 8.75} = 32 \text{ V}. \quad (17)$$

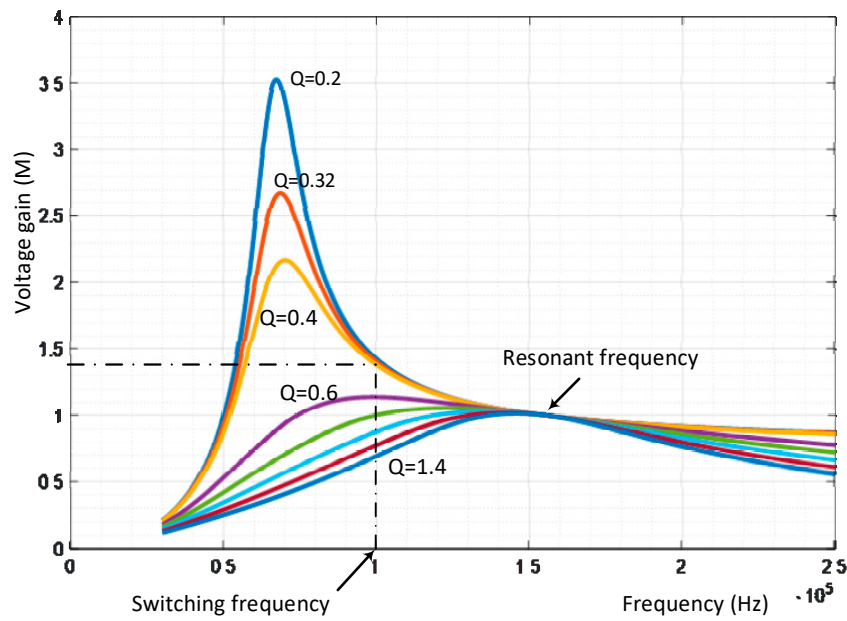


Figure 8. Voltage gain characteristics of the LLC resonant converter with the resonance frequency of 159 kHz, inductance ratio $m = 6$.

Table 4 shows the transformer information, switching frequency and the device lists which are used to implement DC/DC converter.

Table 4. Device list in the LLC resonant DC-DC converter.

LLC Resonant Half-Bridge Design Parameters	
Transformer	ETD 34 (35:4:4)
Switching frequency	100–120 kHz
Resonant Inductor L_R	100 μ H
Magnetizing Inductor L_M	500 μ H
Resonant Capacitor	10 nF
Diode D_1, D_2	STPS30H60CFP
Mosfet S_1, S_2 and Driver	IC FLS2100XS
Feedback Isolation Optocoupler	PC817

3.3. EMI Design

Due to the high-frequency switching noise current and voltage, EMI is a serious concern for the LED driver. Inserting the EMI filter between the AC power supply and the diode rectifier is necessary to comply with the EMC standards and to ensure a correct operation [31].

The LED power supply generally uses a two-stage EMI filter, including a common mode choke and a filter capacitor. Figure 9 is a conventional LED power filter. The differential-mode filter is built from an inductor L_{DM} and two capacitors C_{x1}, C_{x2} . When common-mode interference occurs, the common-mode choke L_{CM} and the Y-capacitors C_{y1}, C_{y2} are used for dampening. The additional capacitors $C_{y-f1}, C_{y-f2}, C_{y-f3}$ connect the ground to the line and neutral for high pulsating voltages. Based on the different current flows through the coupling capacitors C_{y-fi} , the windings of the current compensated choke L_{com} must also be somewhat unbalanced. The two-stage EMI filter shown in Figure 10 is suitable for the boost PFC, which operates in critical mode. In this paper, the EMI filters are designed based on the analysis of the conducted EMI and the use of a spectrum analyzer. According to the CISPR 16-1 standard, the peak and the average value of the EMI noise are measured. Figure 10 shows the peak and average value of conduction noise of the LED driver without using EMI

noise. From Figure 10, Table 5 shows the peak value, the limit of standard CISPR15:2009, and the distinguished limit.

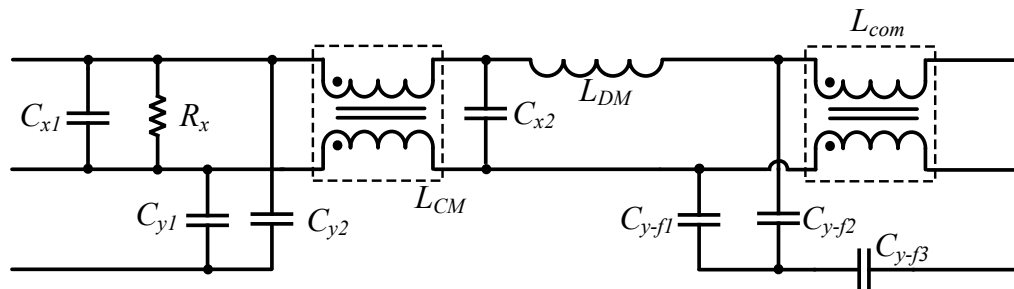


Figure 9. Two-stage electromagnetic interference (EMI) filter.

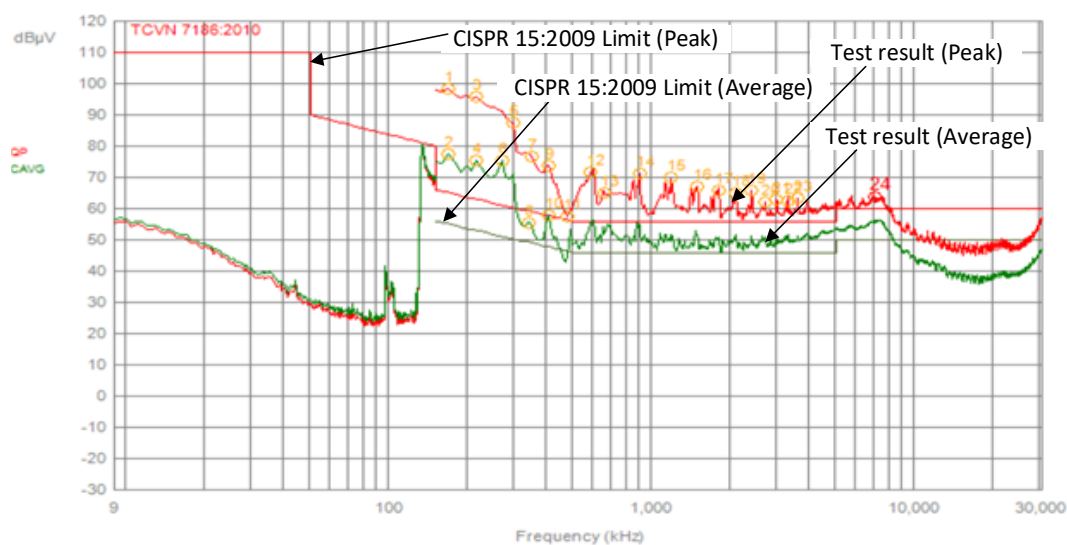


Figure 10. EMI frequency spectrum of LED driver without EMI filter and the limit of standard CISPR15:2009.

Table 5. Measurements of peak and average value of EMI frequency spectrum.

ID	Frequency	Average Value (dBμV)	Peak Value (dBμV)	Standard CISPR 15:2009 (dBμV)	Distinguished Limit (dBμV)
1	168 kHz	68.3	98.3	65.1	33.2
3	213 kHz	65.8	95.8	63.1	32.7
5	294 kHz	57.5	87.5	60.4	27.1
7	348 kHz	46.8	76.8	59.0	17.8
9	402 kHz	43.7	73.7	57.8	15.8
12	582 kHz	41.6	71.6	56.0	15.6
14	888 kHz	41.1	71.1	56.0	15.1
15	1176 kHz	39.8	69.8	56.0	13.8
16	1473 kHz	37.0	67.0	56.0	11.0
17	1779 kHz	35.8	65.8	56.0	9.8
19	2373 kHz	35.7	65.7	56.0	9.7

Based on Table 5, the EMI filter is designed as follows [32]:

Step 1: Measure information EMI noise without filter.

The maximum boundaries of the noise are shown in Table 5. In this case, the working conditions for the worst noise spectra are 68.3 dB with the frequency $f = 168$ kHz. The corner frequency is

calculated by using the frequency of the first highest interference amplitude point of CM interference. The corner frequency is determined by the following equation.

$$f_c = \frac{f_{sw}}{10^{\frac{\Delta V_{noise} - V_{limit}}{40}}}, \quad (18)$$

where f_c is the corner frequency (cut-off frequency) and $(\Delta V_{noise} - V_{limit})$ is the attenuation.

Step 2: Determine filter attenuation requirement.

For a proper margin, a limit under 6 dB is used in the design, and the attenuation requirement is calculated as

$$V_{req,dB} = V_{dB} - V_{limit,dB} + 6dB, \quad (19)$$

where V_{dB} is obtained from the EMI noise without a filter, $V_{limit,dB}$ is the conducted EMI limit specified by the CISPR 15:2009 standard, and +6 dB is reserved for safety.

Step 3: Determine filter corner frequency.

From Table 5, and Equations (18) and (19), the corner frequency is determined as

$$f_c = \frac{f_{sw}}{10^{\frac{V_{NOISE} - V_{LIMIT} + 6dB}{40}}} = \frac{168.10^3}{10^{\frac{33.2dB + 6dB}{40}}} = 17.6 \text{ kHz}. \quad (20)$$

Step 4: Determine filter component values.

Based on standard EN 60335-1, the maximum capacitance connected to the ground cannot exceed about 4700 pF on each phase for 250 VAC 50 Hz mains to meet the safety leakage current requirement. In this design, the Y-capacitors (C_{y1} , C_{y2}) are chosen at 1000 pF.

The L_{CM} and Y-capacitor should have a resonant frequency of f_c obtained in Step 3. Therefore, the inductor L_{CM} is calculated as follows:

$$L_{CM} = \frac{1}{(2\pi f_c)^2 \times 2C_y} = \frac{1}{(2\pi \times 17.6 \times 10^3)^2 \times 2 \times 10^{-9}} = 41 \text{ mH}. \quad (21)$$

For the differential-mode component, there is a freedom in choosing the differential mode inductor L_{DM} and X-capacitor capacitance. Hence, L_{DM} and C_{x1} , C_{x2} are decided according to Equation (22).

$$L_{DM} = \frac{1}{(2\pi f_c)^2 \times C_x}. \quad (22)$$

In this design, C_{x1} , C_{x2} are chosen at 0.47 μ F, which is commonly available. Thus, L_{DM} was obtained:

$$L_{DM} = \frac{1}{(2\pi f_c)^2 \times C_x} = \frac{1}{(2\pi \times 17.6 \times 10^3)^2 \times 470 \times 10^{-9}} = 174 \text{ } \mu\text{H}. \quad (23)$$

The parameters of the two-stage EMI filter are given in Table 6.

Table 6. Parameters of the input EMI filter.

Component	Values
C_{y1} , C_{y2}	1 nF/400 V
C_{y-f1} , C_{y-f2} , C_{y-f3}	1 nF/400 V
C_{x1} , C_{x2}	470 nF/240 V
L_{CM}	40 mH
L_{DM}	170 μ H
L_{com}	20 mH

For the filter, as shown in Figure 9, there are the two discrete components: L_{DM} and L_{CM} . As calculated in Equation (23), the inductance of the L_{DM} is small. Therefore, two discrete components L_{CM} and L_{DM} can be integrated into one inductive component L_C , as shown in Figure 11, where the leakage of the L_C is L_{DM} . The leakage inductance of the L_C is used as a series inductance, which represents a differential-mode filter component.

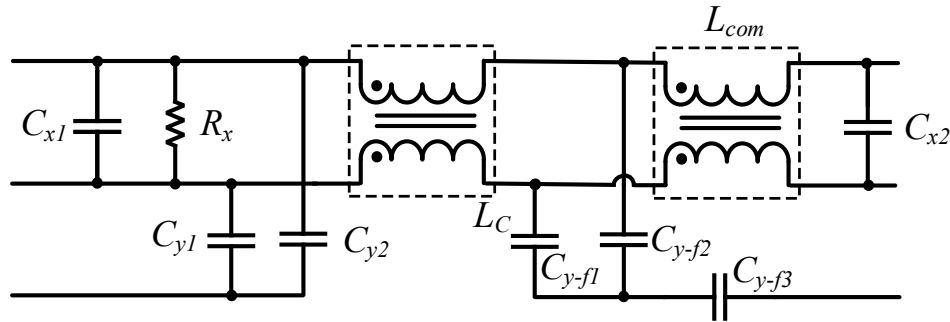


Figure 11. EMI filter with the combination L_{CM} and L_{DM} .

3.4. Dimming Control

LED lighting loads have very similar electrical characteristics to a diode, which is a constant-current source. The brightness of the LED light is dimmed by controlling the current. The load current is controlled by adjusting the switching frequency of the LLC resonant half-bridge converter. In this paper, the IC FLS2100XS is used for controlling the LLC resonant half-bridge converter. Figure 12 shows the typical circuit configuration for the RT pin of FLS2100XS, where an opto-coupler transistor is connected to the RT pin to control the switching frequency.

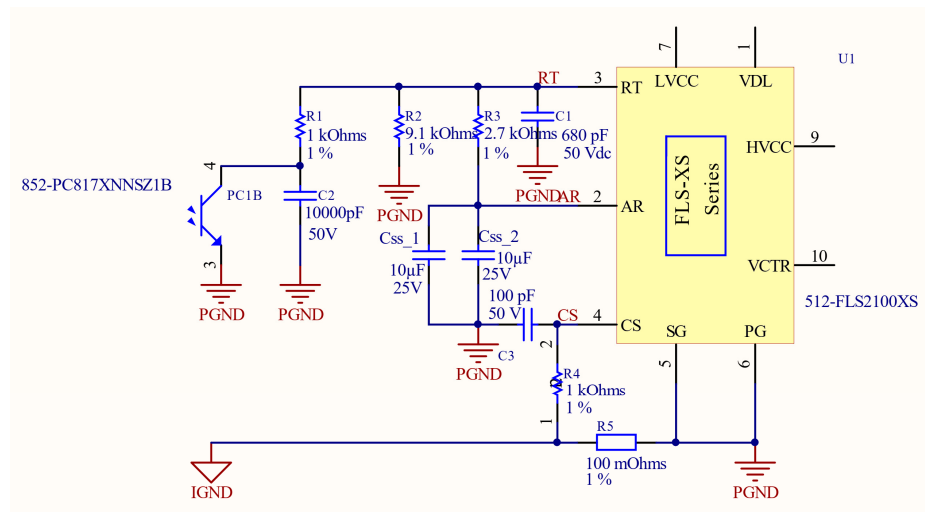


Figure 12. Typical circuit configuration for RT pin of IC FLS2100XS.

The LED dimming is implemented as shown in Figure 13, which is described as the error between the reference value and feedback value, and is the input of the analog proportional–integral (PI) controller. In addition, the output of the PI controller is used to feed the opto-coupler, which is connected to the RT pin of IC FLS2100XS to adjust the switching frequency.

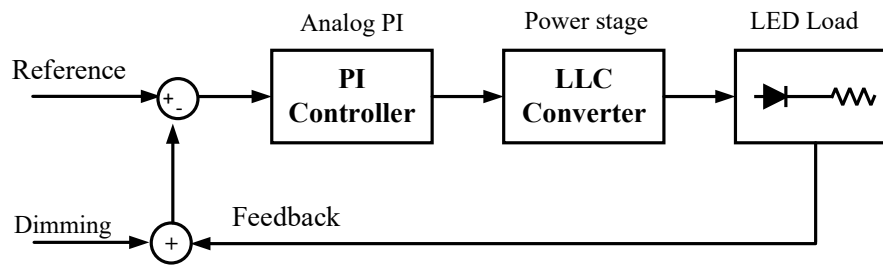


Figure 13. Principle of LED light dimming.

The schematic circuit diagram of the analog PI controller is presented in Figure 14. The analog PI controller is implemented by using op-amp LM358. First, the load current is measured by using the shunt resistor and feedback to the analog PI controller circuit through the variable resistor VR1. Second, the reference voltage is generated by using IC TL431, and it is set at 2.5 V. Hence, the 0–10 V or PWM dimmer is connected to R1 and the LC filter (R1, C1) generating the DC voltage. The feedback current and the dimming current signals are added and compared to the reference current signal. In addition, the error signal is fed into the analog PI controller, as shown in Figure 13.

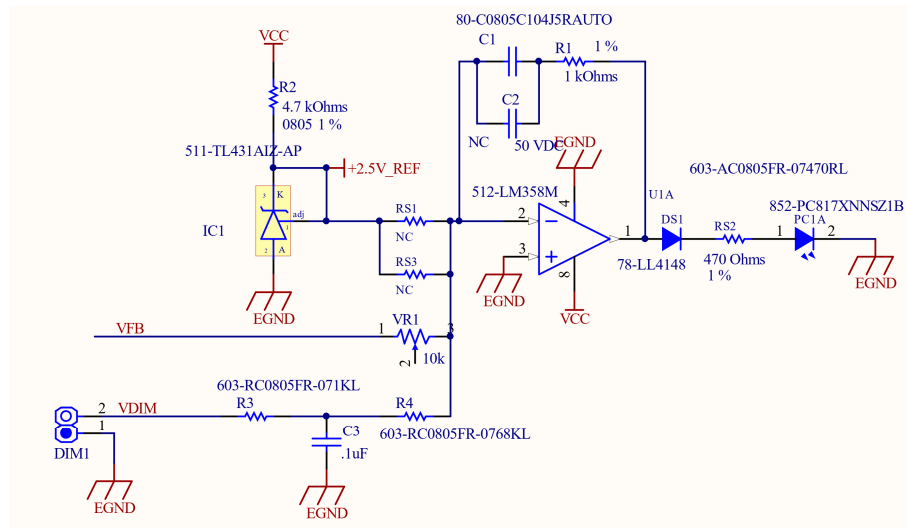


Figure 14. Schematic circuit diagram of the analog proportional–integral (PI) controller.

4. Hardware Prototype and Experimental Results

In this paper, a hardware implementation of the dimmable 150 W LED driver for street lighting application was carried out in laboratory, as shown in Figure 15. The specification and key components of the two-stage LED driver are presented in detail in Table 2, Table 4, and Table 6. In this research, the experimental results are measured by using a Fluke 437 Power Quality and Energy Analyzer, and the waveforms are captured by using the Tektronix TDS 2024B.



Figure 15. Experimental setup of 150 W LED driver.

Figure 16a,b show the prototype and the conducted EMI test results with the peak and average in the case of using L_{CM} and L_{DM} as two separated inductors L_{CM} and L_{DM} . Similarly, Figure 17a,b show the prototype and the conducted EMI test results with the peak and average value, in the case of using L_{CM} and L_{DM} in combination. It proves that the prototype can pass the CISPR 15:2009 standard.

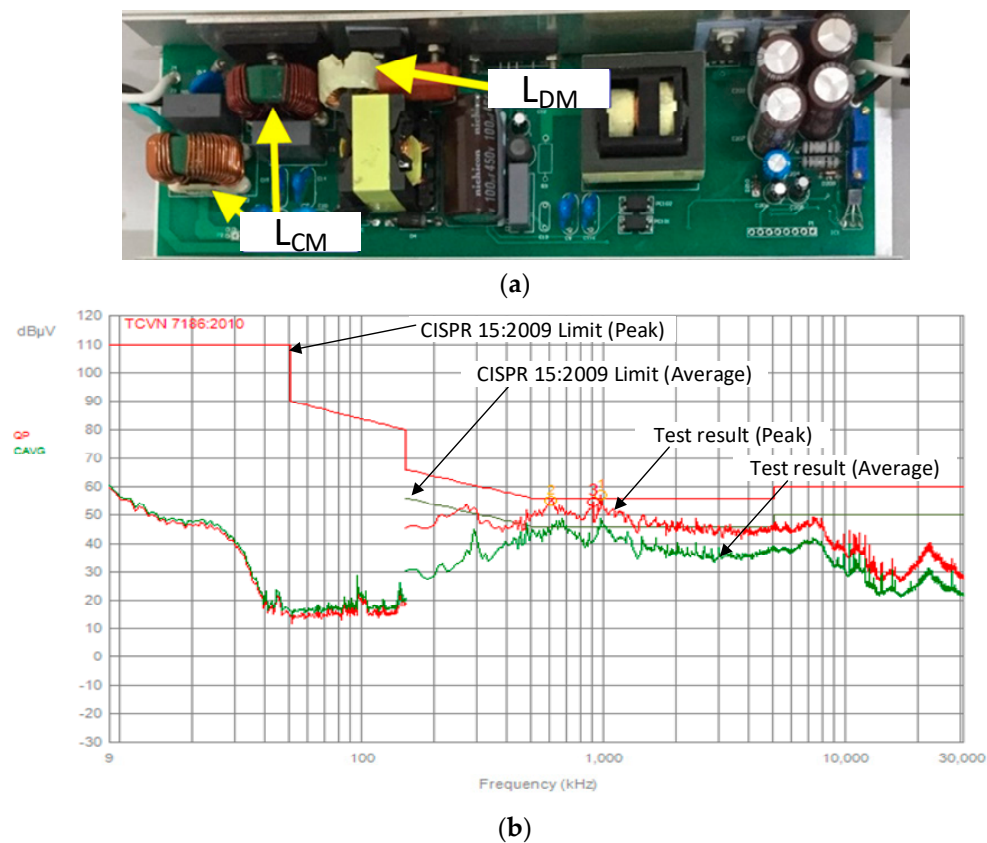


Figure 16. (a) The 150 W-based two-stage LED driver using EMI filter with discrete L_{CM} and L_{DM} . (b) EMI frequency spectrum.

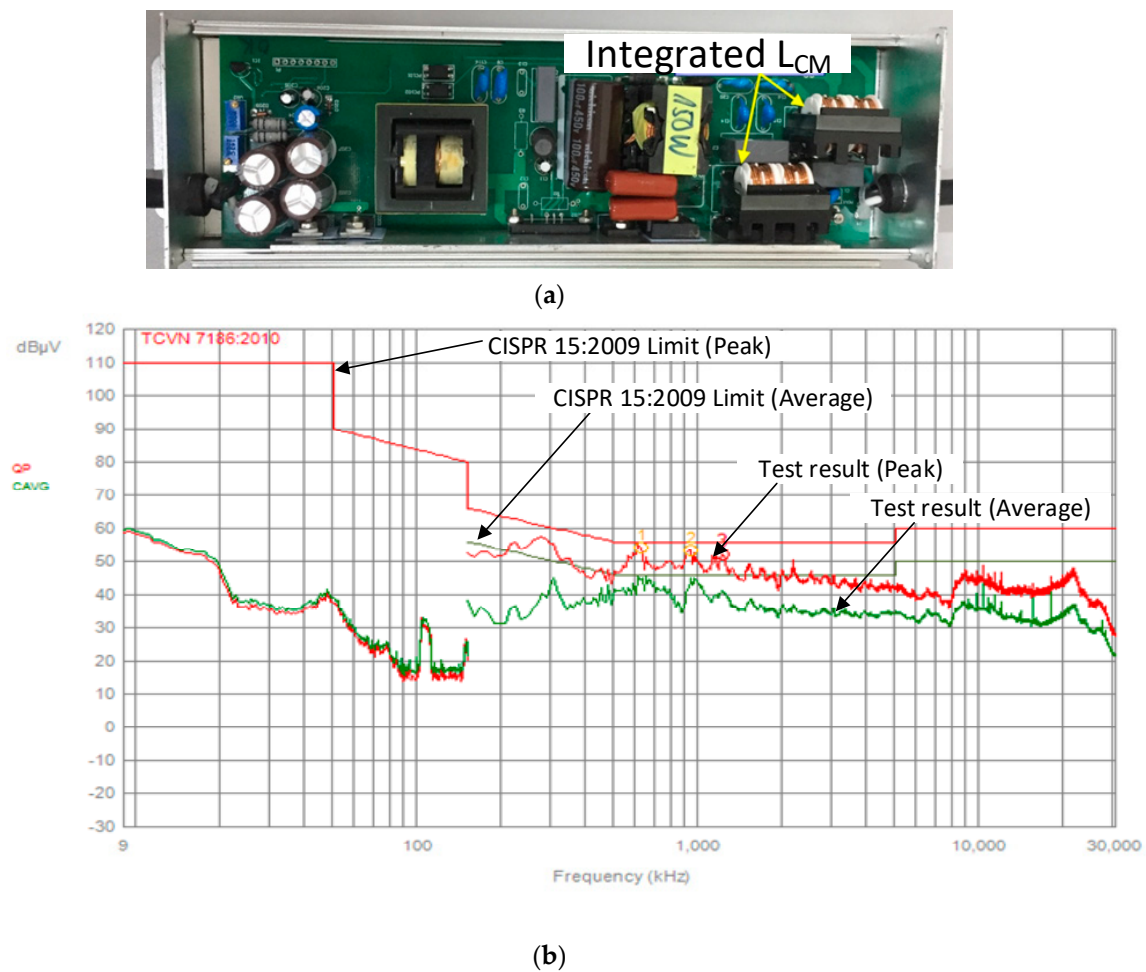


Figure 17. (a) The 150 W-based two-stage LED driver with combined L_{CM} and L_{DM} . (b) EMI frequency spectrum.

Figure 18a,b show the current/voltage waveforms in the PFC stage. In Figure 18a, experimental results of the input voltage v_{in} and input current i_{in} are shown, and it can be seen that the obtained power factor is nearly unity. The inductor current and the full-bridge rectifier voltage are shown in Figure 18b. It can be seen that the peak value of the inductor current is proportional to a full-wave rectified waveform. Figure 19 shows the waveforms of the input voltage and output voltage of the PFC stage. The data obtained from experimental results show that the average voltage is 405 V, which is higher than the nominal value (400 V), and its ripple is around 20 V, which corresponds to approximately 5%.

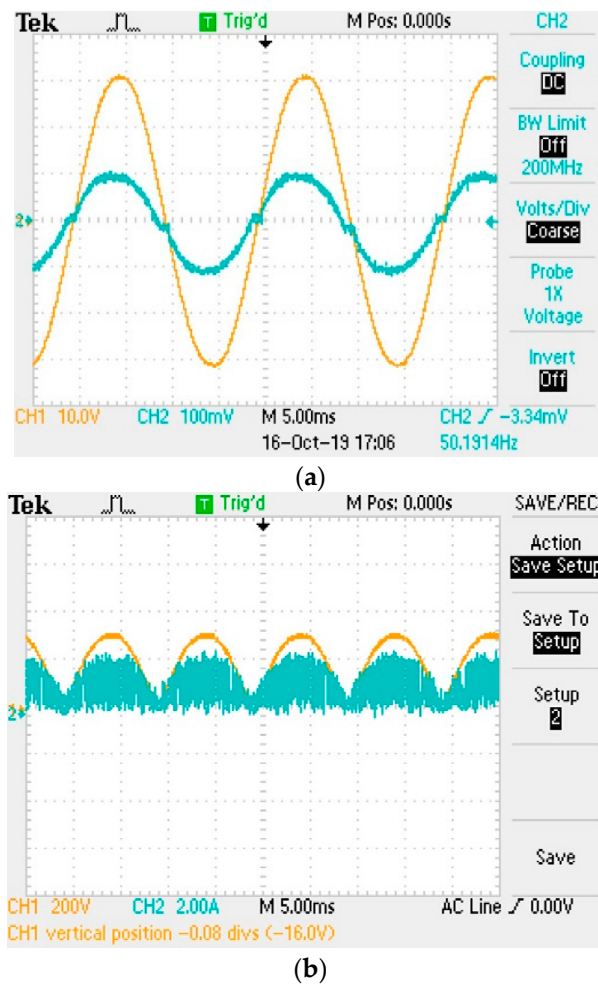


Figure 18. (a) Experimental results of input voltage/current, and (b) experimental results of full-bridge rectifier voltage and inductor current of boost PFC.

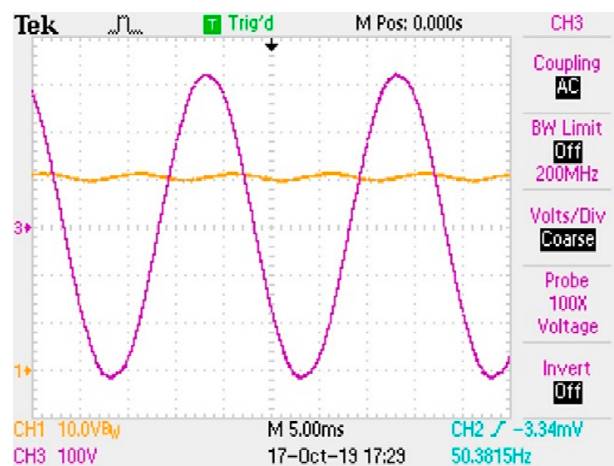


Figure 19. Experimental results of input voltage and DC-bus voltage.

Figure 20 shows experimental results of output voltage (V_{LED}) and output current (I_{LED}). The load voltage is 31.3 V, and the current is 4.9 A. This means that the power is 150 W. The voltage ripple is lower than 1 V, and the current ripple is about 150 mA. Table 7 presents the results of the input current, efficiency, PF, THD, with variation of the input voltage. With the nominal value, 220 Vrms, the PF is

higher than 0.98, the THD is lower than 7%, and the efficiency is higher than 91.6%. As expected, this converter is suitable for a wide range of input voltages, and it achieved a good performance for the THD and PF, as well as efficiency. Figure 21 depicts the input current spectrum, and the THD is 6.4%. The results show that all input harmonics are in accordance with the IEC 61000-3-2.

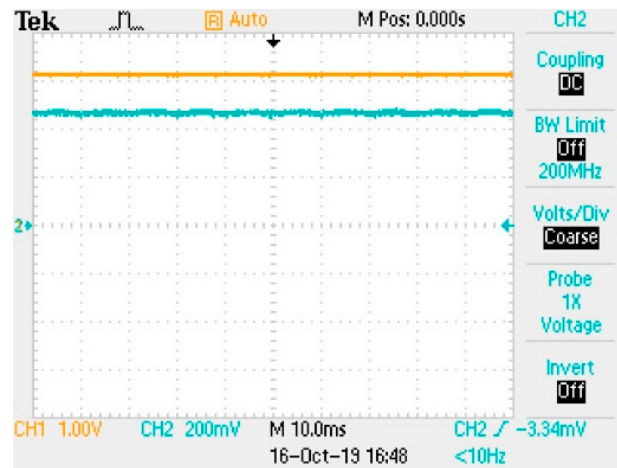


Figure 20. Experimental results of output voltage/current.

Table 7. Experimental results of input/output performance of 150 W two-stage LED driver.

Input Voltage (V)	Input Current (A)	Input Power (W)	Output Voltage (V)	Output Current (A)	Output Power	Efficiency %	Power Factor	THD %
260.7	0.676	170.1	31.33	4.995	156.5	92.00	0.966	9.20%
250.3	0.7	169.8	31.33	4.98	156.0	91.89	0.97	8.20%
240.4	0.731	171.1	31.38	5.003	157.0	91.76	0.975	7.10%
230.9	0.76	171.5	31.41	4.993	156.8	91.45	0.978	6.70%
220.8	0.786	170.3	31.32	4.985	156.1	91.68	0.981	6.40%
210.2	0.824	170.4	31.32	4.987	156.2	91.66	0.984	6.20%
200.3	0.866	171.1	31.33	5.003	156.7	91.61	0.987	6.10%
190.2	0.907	170.6	31.32	4.987	156.2	91.56	0.989	5.90%
180.6	0.955	171	31.31	4.995	156.4	91.46	0.991	5.50%
170.2	1.012	171.1	31.31	4.994	156.4	91.39	0.993	5.40%
160.2	1.075	171.2	31.3	4.988	156.1	91.19	0.995	4.90%
150.5	1.144	171.4	31.3	4.985	156.0	91.03	0.996	4.80%

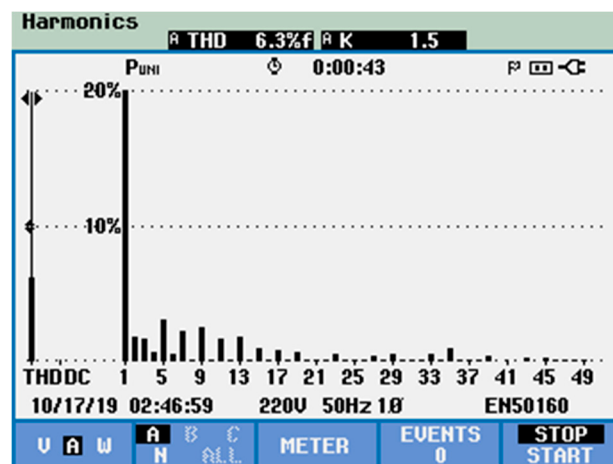
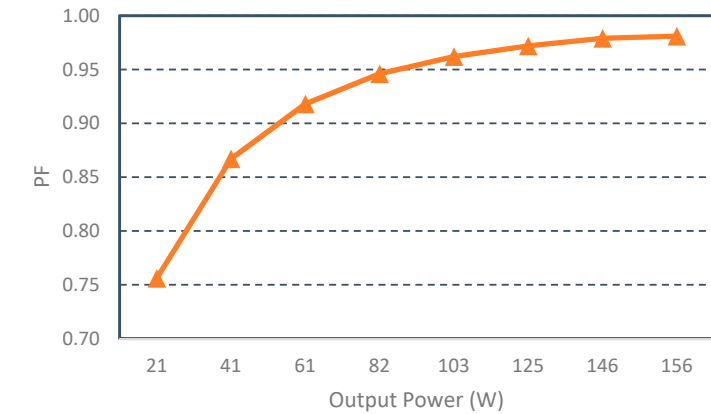
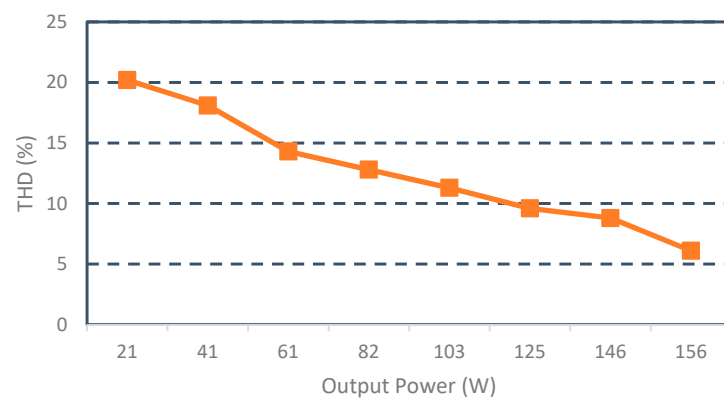


Figure 21. FFT analysis of the input current.

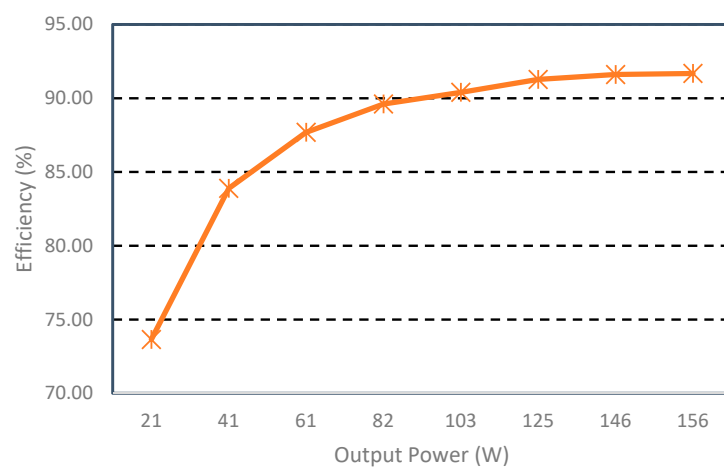
Figure 22a–c show the values of the input power factor, THD, and system efficiency according to the variation in load: From 10% to full load. The efficiency is always higher than 90% and the system power factor (PF) remains more than 0.97 when the load changes from 70% to 100%. Furthermore, the THD is smaller than 20% with 10% load and will decrease to 5.2% at full load.



(a)



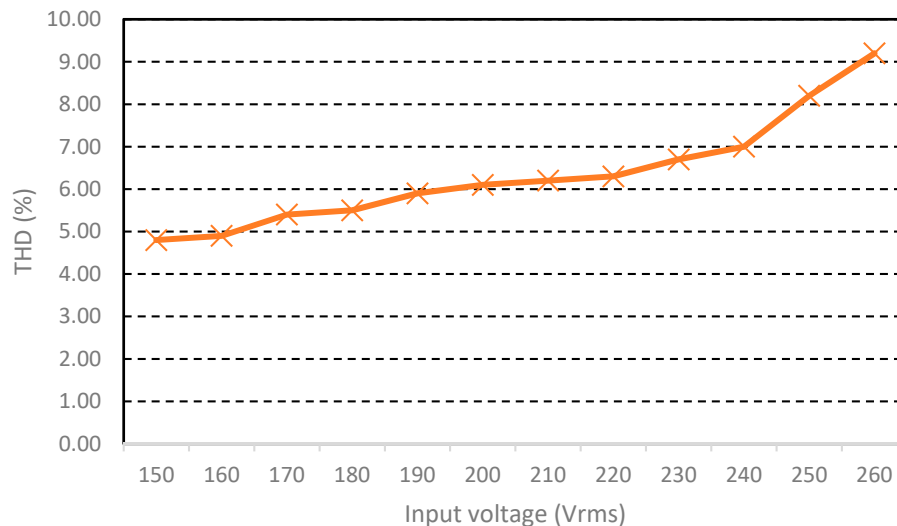
(b)



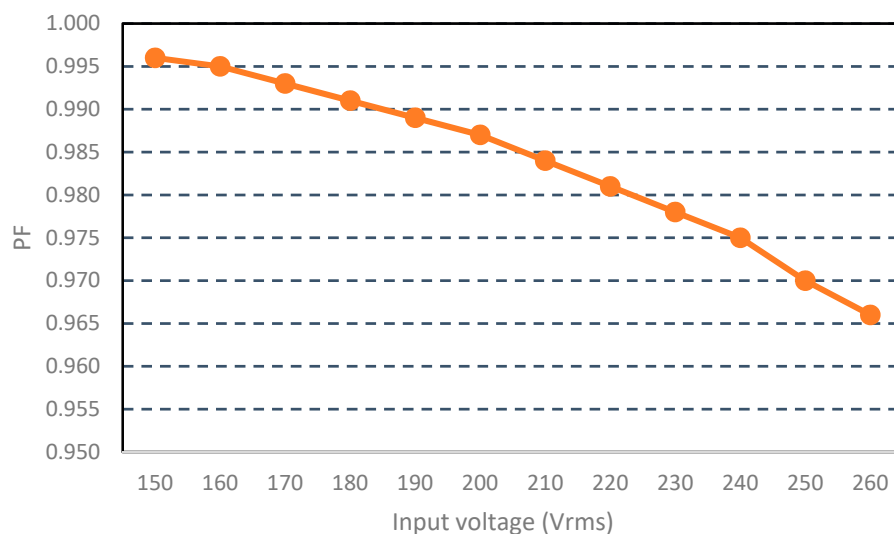
(c)

Figure 22. Experimental results of (a) PF and (b) total harmonics distortion (THD). (c) Efficiency with changing power load.

Figure 23a,b show the curves of the system PF and THD with changing input voltage. As shown in these figures, when the input voltage changes between 150 and 220 Vrms, the system THD is always lower than 10%, and the PF remains greater than 0.965. From Figures 22 and 23, the system THD always satisfies the IEC 61000-3-2 standard.



(a)



(b)

Figure 23. Experimental results of (a) PF and (b) THD with changing input voltage.

5. Conclusions

In this paper, the implementation of a 150 W LED driver, which was built from the combination between a boost PFC circuit and half-bridge-type LLC resonant circuit, was presented. An IC NCP1608 manufactured by ON Semiconductor was used to drive the switches in the boost converter, and an IC FLS2100XS was used in the LLC resonant half-bridge converter. The design of the AC/DC and DC/DC converter was analyzed and presented in detail. Furthermore, the optimization of the EMI filter has been presented, and discrete inductances were combined in order to make the filter smaller and cheaper. The output power has been tested from 10% to full load to show the effectiveness of the dimmable control. A prototype was built in the laboratory and some experimental results were

shown to demonstrate a PF as high as 0.981, THD lower than 6.5%, efficiency as high as 91.8%, and bus voltage as low as 400 V. The EMI and THD test results show that the driver complies with CIRSP 15:2009 and IEC 61000-3-2.

Author Contributions: L.M.P. and N.D.T. designed the methodology and wrote the manuscript. N.T.T. and N.M.H. conceived and designed the hardware. N.D.T. and N.H.P. implemented the experiments and collected data. N.C.C. provided validation and editing. L.M.P. provided supervision and revised the manuscript. All authors have read and agreed to the published version of the manuscript.

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