Abstract—A simple, efficient, and low-cost quasi-resonant LED driver is presented. The proposed driver relies on capacitive safety isolation barrier and eliminates the need for a bulky isolation transformer. Thus, cost and footprint area can be reduced. This paper describes the topology and presents the theory of operation. Theoretical predictions are verified by simulation and experimental results.

Index Terms—Capacitive power transfer, current source, dimming, resonant LED drive.

I. INTRODUCTION

The recent years have seen a major advancement of LED semiconductor technology, which resulted in improved brightness and color temperature, lower energy consumption, and longer life spans of LED light sources [1], [2]. LED lighting technology is ∼5 times more energy efficient than the incandescent lighting and can offer longer service life [1]–[4]. Hence, LEDs can reduce maintenance cost, and help energy conservation in stationary and automotive applications [5]. Since ∼20% of world’s generated power is consumed by lighting, LEDs can dramatically cut the CO2 polluting emissions from the power plants. In battery operated systems, the higher efficiency of LEDs and their drivers bear directly on the maximal operation time [6]. LED lamps are also flicker-free, emit directional light, can be dimed, and, for these reasons, are well suited for urban, office, commercial, industrial, and traffic environment. Typically, LEDs are powered by low-voltage source and, therefore, are safe to use in indoor and outdoor settings as well as in emergency applications.

To operate an LED lighting fixture, a specialized power converter, referred to as LED driver, is required. Offline LED lighting system presents several challenges. In these applications, the LED driver has to perform ac–dc conversion with large voltage step down, attain low distortion of the line current [7], and, for safety reasons, provide isolation [4], [8], [9]. High efficiency alleviates heat management, helps minimizing the driver’s size, and allows the lighting fixture to be fitted into congested space, where heat removal is difficult. Studies on the issue of characteristics matching between power sources and power converters were presented in several publications [10]–[12], and similar interfacing problematic applies to converter-load matching. Thus, although LED driver with regulated output voltage can do well, LEDs prefer constant current operating conditions. LED driver with current control can also provide precise dimming function. For these reasons, some LED drivers are rather complex and are a major contributor to the cost of LED lighting systems. High performance and small size at acceptable cost can promote widespread use of LED lighting and full realization of its potential benefits. However, lowering the cost usually requires compromises on performance [4], [13].

The space available for LED drivers is, in many cases, very limited (for example, in LED retrofit lamps), this being the motivation to miniaturization of the driver. Miniaturization of the driver circuit calls for high switching frequency and removal of the transformer, yet providing line isolation. In addition, transformers, involving manual procedures, do not comply well with manufacturability. Attaining high efficiency at high switching frequency requires resonant conversion methods to prevent switching losses [14]–[16]. High efficiency being of high importance not only from the perspective of energy saving but also due to the often limited heat removal and most importantly in order to prevent heating of the LEDs by their driver, which may be mounted in close proximity, such as in LED retrofits.

Transformer is a popular and widely used device for providing mains isolation. However, satisfactory safety feature can be attained by means of capacitive coupling [17]. Two capacitors (one on each bus line) small enough to block low-frequency current can be connected to provide mains isolation. These series capacitors are also part of the resonant tank [15] and conduct appropriate high-frequency current to power an LED string. The capacitive isolation has the merit of eliminating the need for an isolation transformer. Recently, a single stage, single switch, efficient, and low-cost multiresonant offline LED driver with capacitive safety isolation barrier and no transformer was briefly introduced [18]. In addition, [18] demonstrated zero-voltage-switching (ZVS) conditions at switching instances and thus lossless switching. This paper further investigates the idea of LED driver with capacitive isolation. The presented driver can be classified as a zero-current (ZC)-switched quasi-resonant type, whereas the earlier counterpart [18] is a multi-resonant ZV-switched topology. Although these circuits share the common idea of capacitive isolation, they differ in their operational principles. The concept of the quasi-resonant ZC-switched topology was introduced in [19].
In this paper, a detailed theoretical description of the proposed quasi-resonant LED driver (QRLEDD) for lighting applications is presented. Analytical relationships are verified by simulation and experimental results. Based on the theoretical analysis, a design approach is offered.

II. PROPOSED TOPOLOGY

Due to LEDs’ steep current–voltage \( i_D(v_D) \) characteristics, LEDs should be energized by a current source, whereas the input is a voltage source. This calls for driver circuits with gyrator characteristics (on average) [20]–[23]. The schematic of the proposed QRLEDD is shown in Fig. 1. The circuit uses a full-wave rectifier \( D_1-D_4 \), followed by a small high-frequency bypass capacitor, \( C_b \). A series input inductor, \( L_i \), is switched at high frequency by the switch, \( M \), to shape and control the magnitude of the input current. Fast input diode, \( D_i \), assures discontinuous input current mode. A pair of small series capacitors, \( 2C_s \), constitutes the capacitive isolation barrier. Diodes \( D_5 \) and \( D_6 \) are used to steer the capacitor current either into the small resonant inductor \( L_r \) or into the filter capacitance, \( C_{LED} \), across which a series string of LEDs is connected. In addition, \( D_5 \) prevents the switch parasitic capacitance oscillating with \( L_r \).

The QRLEDD possesses several desirable features, such as single active switch, ZVS turn-OFF, ZC turn-ON, which make high-frequency operation possible, resistive input port characteristics, ground referenced gate drive, inherent reset of blocking capacitors, and reduced circulating current.

III. MAIN WAVEFORMS

Principal waveforms of the proposed QRLEDD on the line frequency and on the switching frequency scale are shown in Fig. 2. The average line current waveform, \( i_{ac} \), shown in Fig. 2(a) appears as a near sinusoidal signal, which manifests that the QRLEDD possesses inherently high power factor and low harmonic content. The high power factor is not attained by a dedicated control (as in [23]) but rather due to the nature of the topology. This makes it a member of the loss-free-resistor family of converters [24], [25]. Examining the switch waveforms in Fig. 2(b) reveals favorable ZVS turn-OFF and ZC-switching (ZCS) turn-OFF lossless switching of the switch, which are a prerequisite to attain high efficiency at high frequency.

IV. ANALYSIS

A. Basic Assumptions

To facilitate the analysis approach, several simplifying assumptions are adopted. First, the rectified ac line voltage, \( V_{ac} \), is regarded as if frozen in time. This allows representing the line, the line rectifiers, and the bypass capacitor, \( C_b \), by an equivalent dc source, \( V_i \). Second, the LED string in parallel with the filter capacitance, \( C_{LED} \), can be modeled by an equivalent dc voltage source, \( V_{LED} \), determined by the number of diodes in the string. Finally, a pair of series connected isolating capacitors, \( 2C_s \), is represented by their equivalent value, \( C_s \). It is also assumed that all circuit components are ideal. The resulting model is shown in Fig. 3(a), whereas its detailed steady-state waveforms are presented in Fig. 3(b).

B. State Analysis

Inspection of the waveforms in Fig. 3(b) reveals that the switching cycle of QRLEDD is comprised of five topological states, the equivalent circuits of which are shown in Fig. 4 in the order of their appearance throughout the switching cycle.

State A: \( t_0-t_1 \) [see Fig. 4(a)] starts with ZCS of \( M_1 \). Here, \( L_i \) is charging linearly from the input source, \( V_i \), through \( D_i \). Meanwhile, \( C_s \) resonates with \( L_i \) through \( D_5 \) and discharges.

State B: \( t_1-t_2 \) [see Fig. 4(b)] starts as \( D_6 \) conducts and allows \( L_r \) to discharge linearly to the output, \( V_{LED} \), while \( C_s \) is clamped to the negative output voltage. During this interval, \( L_i \) continues its linear charging.

State C: \( t_2-t_3 \) [see Fig. 4(c)] starts when controller terminates switch conduction. ZVS turn-OFF condition is attained due to \( C_s \) snubbing the switch voltage, \( V_{ds} \).
Here, also $L_i$ and $C_s$ resonate charging $C_s$ to high voltage. The resonant current pulse flows to the output, $V_{LED}$, via $D_6$. Meanwhile, $L_r$ is clamped and keeps linearly discharging to the output. State C terminates as $D_4$ turns off at ZC.

**State D:** $t_3$–$t_4$ [Fig. 4(d)] commences upon ZCS of $D_4$. Here, $L_r$ keeps linear discharge to the output, while all the capacitors hold their voltage constant. State D terminates when $D_5$ and $D_6$ turn off under ZCS condition.

**State E:** $t_4$–$T_s$ [see Fig. 4(e)] is the idle state. Here, no current flows in any of the inductors, whereas the capacitors hold their voltage constant.

### C. Quantitative Analysis

1) **Output Power and Current:** Per each switching cycle, $T_s$, the average output current, $I_{LED}$, is generated by two charge components $q_r$ and $q_i$ supplied to the output filter capacitor, $C_{LED}$, by the output current $i_o(t)$ (see Fig. 5). The charge increment $q_i$ is contributed by the discharge current $i_i(t)$ of the input inductor, $L_i$, during State C, whereas the discharge current of the resonant inductor, $L_r$, which occurs during both States C and D, contributes the charge increment $q_r$.

During State A, the capacitor $C_s$ resonates with the resonant inductor $L_r$ [Fig. 4(a)]. Consequently, capacitor voltage falls from its peak value, $v_c(t_0) = V_m$, toward its minimum value $v_c(t_1) = -V_{LED}$ (Fig. 3). Hence, the energy, $E_r$, captured by the resonant inductor, $L_r$, is

$$E_r = \frac{1}{2} C_s (V_m^2 - V_{LED}^2).$$  \hspace{1cm} (1)

The energy, $E_r$, is then released by $L_r$ to the output during States C and D, depositing the corresponding charge increment, $q_r$, across $C_{LED}$, which is given by

$$q_r = \frac{E_r}{V_{LED}} = \frac{1}{2} C_s \left( \frac{V_m^2}{V_{LED}^2} - 1 \right). \hspace{1cm} (2)$$

During State C, the input inductor, $L_i$, resonates with the capacitor, $C_s$, the voltage of which rises from its lowest value $v_c(t_2) = -V_{LED}$ back to its maximum $v_c(t_3) = V_m$ (Fig. 3). Due to series connection [Fig. 4(c)], the charge, $q_i$, deposited on the output filter capacitor, $C_{LED}$, equals the
Fig. 5. Charge components of the output current, $i_o$.

charge increment across the capacitor, $C_s$

$$q_i = C_s(V_m + V_{LED}).$$  \hspace{1cm} (3)

The dc output current, $I_{LED}$, is the total charge supplied to the output $(q_r + q_i)$ averaged per switching cycle, $T_s$

$$I_{LED} = \frac{1}{T_s}(q_r + q_i).$$  \hspace{1cm} (4)

The output power can now be found plugging (2) and (3) into (4)

$$P_{LED} = I_{LED}V_{LED} = \frac{1}{T_s} C_s(V_m + V_{LED}) \left[ V_{LED} + \frac{V_m^2 - V_{LED}^2}{2} \right].$$  \hspace{1cm} (5)

Obtaining the explicit solution calls for finding the maximum voltage, $V_m$, across the capacitor $C_s$, as a function of circuit parameters and operating conditions (see the Appendix for details)

$$V_m = V_i \left( 1 - \frac{V_{LED}}{V_i} + 1 + \sqrt{1 + \left( \frac{2\pi D}{f_n} \right)^2} \right).$$  \hspace{1cm} (6)

Substitution of (6) into (5) and further manipulation yields the normalized output power of the QRLEDD

$$P_n = \frac{P_{LED}}{\frac{1}{2} V_i^2} = \left( \frac{f_n}{2\pi} \right) \left[ 1 + \sqrt{1 + \left( \frac{2\pi D}{f_n} \right)^2} \right]^2.$$  \hspace{1cm} (7)

The function $P_n$ is plotted in Fig. 7(a). Here, $f_s$ is the switching frequency, $T_s = (1/f_s)$ is the switching period, $T_{ON} = t_2 - t_0$ is the switch on time, $D = (T_{ON}/T_s)$ is the duty cycle, $f_n = (f_s/f_0)$ is the normalized switching frequency relative to the series resonant frequency $f_0 = (1/2\pi(L_iC_s)^{1/2})$, and $Z_{in} = (L_i/C_s)^{1/2}$ is the characteristic impedance relative to the input inductor $L_i$.

It is further assumed that LED string voltage can be represented by a simple linearized model

$$V_{LED} = n(V_o + R_{ON}I_{LED})$$  \hspace{1cm} (8)

where $n$ is the number of LEDs in the string, $V_o$ is the LED’s conduction threshold voltage, and $R_{ON}$ is the LED’s ON resistance (Fig. 6).

Hence, using (8), the power of the LED string as a function of string current, $I_{LED}$, and LED’s parameters can be written as

$$P_{LED} = V_{LED}I_{LED} = n(V_o + R_{ON}I_{LED})I_{LED}.$$  \hspace{1cm} (9)

Solving (9) yields an approximate solution for the current, $I_{LED}$

$$I_{LED} = \frac{V_o}{2R_{ON}} \left( \frac{V_o}{R_{ON}} \right)^2 + \frac{P_{LED}}{nR_{ON}}$$  \hspace{1cm} (10)

where $P_{LED}$ is given by (7).
2) **AC Operation:** Hitherto, QRLEDD analysis was conducted under assumption of dc input voltage $V_{in}$. However, typical operation condition of QRLEDD is from a rectified utility line, as shown in Fig. 1. Assuming that the amplitude of the rectified voltage is $V_i$ and its corresponding rms value is $V_{rms} = (1/\sqrt{2})V_i$; the ac power of QRLEDD at the peak of the ac line is about twice the average power, $P_{av}$, considering that the current waveform of the QRLEDD has some residual distortion and, its power factor is lower than unity, it was found that a correction factor 0.933 is appropriate. Therefore, (7) yields

$$P_{LED} = \left(\frac{V_{rms}^2}{Z_{oi}}\right) P_n = 0.933 \times 2 P_{av} = 1.866 P_{av}.$$  \hfill (11)

Fig. 2 reveals that the input port of the QRLEDD has a near-resistive input characteristic. The emulated resistance, $R_e$, seen by the line can be found from

$$R_e = \frac{V_{rms}^2}{P_{av}} = \frac{1.866 Z_{oi}}{P_n}. \hfill (12)$$

3) **Peak Switch Voltage:** Fig. 4(d) reveals that the peak switch voltage, $V_{ds,m}$, is the sum of the peak capacitor voltage, $V_m$, and the output voltage, $V_{LED}$: $V_{ds,m} = V_m + V_{LED}$. Hence, using (6), the normalized peak switch voltage can be obtained

$$V_{ds,mn} = \frac{V_{ds,m}}{V_i} = 1 + \sqrt{1 + \left(\frac{2\pi D}{f_n}\right)^2}. \hfill (13)$$

The normalized output power (2) and the normalized switch peak voltage (3) as a function of the normalized frequency, $f_n$, and duty cycle, $D$, are plotted in Fig. 7. The plot of the normalized power, shown in Fig. 7(a), reveals that duty cycle or frequency variations have somewhat limited the effect on QRLEDD’s power. Hence, burst control seems as preferred method of control. Such strategy involves feeding the LED string by a chain of discrete pulses, whereas the continuously variable off time can help avoiding quantization of the attainable average power. Since each pulse within the burst delivers same power to the LED, such method helps sustaining LED’s color.

**V. SIMULATION AND EXPERIMENTAL VERIFICATION**

Theoretical derivations of QRLEDD’s key relationships were verified by simulation. Fig. 8(a) shows the comparison of theoretically predicted versus simulated values of the peak switch voltage $V_{ds,m}$. The calculated power and the simulated power are compared in Fig. 8(b). Excellent agreement is found in both the cases.

To further verify the theoretical predictions, a 30 W, 110 V-input experimental QRLEDD prototype based on the schematic in Fig. 1 was built and tested. The circuit parameters were: 1) switching frequency $f_s = 250$ kHz; 2) input inductance $L_i = 30 \ \mu$H; 3) resonant inductor $L_r = 3 \ \mu$H; 4) capacitors $2C_t = 2 \ \mu$F/600 V; 5) $C_b = 0.5 \ \mu$F/400 V; and 6) $C_{LED} = 1 \ \mu$F/50 V.

The key waveforms of the experimental prototype are shown in Fig. 9(a) and the experimental results are shown in Fig. 9(b). Experiments confirm that ZCS at turn ON of the switch and ZVS at turn OFF of the switch were achieved.

Some ringing may be noticed in the $V_{ds}$ waveform due to the parasitic switch capacitance. Stray inductances of the wiring also result in current ringing that can be observed in $I_{ds}$ waveform. The line current of the QRLEDD, shown in Fig. 9(c), is seen to be nearly sinusoidal with a measured PF = 0.99.

The measured efficiency and output power of the experimental prototype as a function of the frequency are shown in Fig. 10.

The power is seen to be controllable through the switching frequency, indicating potential dimming capability. However, the efficiency is seen to be quite flat (constant) throughout the entire range of 110–190 kHz of switching frequency variation.

**VI. DESIGN CONSIDERATIONS**

A. **ZCS Boundary**

Examination of the waveform of the resonant inductor current, $i_r(t)$, shown in Fig. 11, reveals that during State $A_0 - t_1$: the inductor $L_r$ resonates with $C_t$. The maximum value, $I_{lm}$, of the resonant current can be derived from the energy
conservation considerations as

\[ I_{\text{rm}} = V_m \sqrt{\frac{C_s}{L_r}} \]  

(14)

Some of the energy is then returned to recharge \( C_s \) to the negative value of \((-V_\text{LED})\). Therefore (Fig. 5), at \( t = t_1 \), the output rectifier starts conducting with peak output current, \( I_{\text{om}} \), given by

\[ I_{\text{om}} = \sqrt{\frac{C_s}{L_r}} (V_m^2 - V_\text{LED}^2) \]  

(15)

The duration of State A can be approximated to a quarter of a resonant cycle

\[ \Delta t_1 = t_1 - t_0 = \frac{1}{4} T_{\text{ON}} = \frac{\pi}{2} \sqrt{C_s L_r} \]  

(16)

During States B, C, and D, the inductor \( L_r \) discharges toward the output voltage, \( V_\text{LED} \), assumed constant, and its current, \( i_{L_r}(t) \), falls linearly to reach zero after

\[ \Delta t_2 = t_4 - t_1 = \frac{L_r}{V_\text{LED}} I_{\text{om}} = \sqrt{C_s L_r} \sqrt{\left(\frac{V_{\text{dsm}}}{V_\text{LED}} - 1\right)^2 - 1} \]  

(17)

To assure ZC turn-OFF condition of the output rectifier, the switching frequency should be limited by a certain maximum value, \( f_{\text{sm}} \), given by

\[ f_{\text{sm}} = \frac{1}{\Delta t_1 + \Delta t_2} = \frac{1}{\sqrt{C_s L_r}} \frac{\pi}{2} \sqrt{\left(\frac{V_{\text{dsm}}}{V_\text{LED}} - 1\right)^2 - 1} \]  

(18)

Here, (16) and (17) were used. Normalizing (18) yields

\[ f_{\text{mn}} = \frac{f_{\text{sm}}}{f_0} = \frac{2\pi r}{\frac{\pi}{2} + \sqrt{\left(\frac{V_{\text{dsmn}}}{V_{\text{Ln}}} - 1\right)^2 - 1}} \]  

(19)
Here, \( V_{\text{dsmn}} \) is the normalized switch voltage, \( V_{\text{dsmn}} \), as a function of the highest normalized switching frequency, \( f_{\text{mn}} \) can be derived from (19)

\[
V_{\text{dsmn}} = V_{\text{Ln}} \left[ 1 + \frac{1}{\sqrt{1 + \pi^2 \left( \frac{2r}{f_{\text{mn}}} - \frac{1}{2} \right)^2}} \right].
\]

The solutions of (13) and (20) are plotted in Fig. 12 for selected values of the duty cycle, \( D \), and the inductor ratio, \( r \), for \( V_{\text{Ln}} = 30/155 = 0.19355 \). The crossover points mark the output rectifier ZCS mode boundary. The desired ZCS region lies to the left of the crossover points, where \( f_n < f_{\text{mn}} \).

B. Design Approach

QRLEDD can attain ZV turn-off and ZC turn-on conditions. Although ZV turn-off is lossless, the ZC turn-on losses are associated with the discharge of switch parasitic capacitance, \( C_{\text{oss}} \), which is highly nonlinear and strongly depends on the off-state voltage, \( V_{\text{ds}} \). The turn-on switching loss, \( P_{\text{ON}} \), of a power MOSFET can be estimated by [14]

\[
P_{\text{ON}} = \frac{2}{3} C_{\text{oss}} V_{\text{dsm}}^2 f_s.
\]  

The turn-on loss is frequency dependent, limits the efficiency of the ZCS converters and becomes dominant at low-power level. Hence, the highest switching frequency of QRLEDD can be established from (21) as a function of an acceptable switching power loss, \( P_{\text{ON}} \)

\[
f_{\text{sm}} = \frac{3 P_{\text{ON}}}{2 C_{\text{oss}} V_{\text{dsm}}^2}.
\]  

Some manufacturers provide plots of the energy, \( E_{\text{oss}}(V_{\text{ds}}) \), stored in switch parasitic capacitance \( C_{\text{oss}} \) [26]. This provides an alternative approach to obtain the highest switching frequency simply from

\[
f_{\text{sm}} = \frac{P_{\text{ON}}}{E_{\text{oss}}(V_{\text{dsm}})}. \quad (23)
\]

Either (22) or (23) can be used to find the resonant frequency, \( f_{\text{oi}} \)

\[
f_{\text{oi}} = \frac{f_{\text{sm}}}{f_{\text{mn}}}. \quad (24)
\]

According to the desired power level, \( P_{\text{av}} \), and line voltage, \( V_{\text{rms}} \), the characteristic impedance, \( Z_{\text{oi}} \), can be found using

\[
Z_{\text{oi}} = \frac{V_{\text{rms}}^2}{2 P_{\text{av}}}. \quad (25)
\]

After \( f_{\text{oi}} \) and \( Z_{\text{oi}} \) are established, the key parameters of QRLEDD can be calculated

\[
C_s = \frac{1}{2 \pi Z_{\text{oi}} f_{\text{oi}}}, \quad (26)
\]

\[
L_i = \frac{Z_{\text{oi}}}{2 \pi f_{\text{oi}}}, \quad (27)
\]

\[
L_r = \frac{L_i}{r^2}. \quad (28)
\]

C. QRLEDD Design Example

As an example, the proposed QRLEDD is designed to power a 30 V/30 W LED string from 110-Vrms line.

Since the QRLEDD is designed for \( P_{\text{av}} = 30 \) W, according to the peak LED power is \( P_{\text{LED}} = 56 \) W. To attain the lowest peak switch voltage, \( V_{\text{dsmn}} \), QRLEDD should be operated at the lowest possible duty cycle, \( D \), and the highest switching frequency, \( f_{\text{mn}} \) (see Fig. 7). Therefore, the first step in QRLEDD design is using Fig. 12 to choose the desired \( D \) and \( r \) and noting the corresponding normalized frequency \( f_{\text{mn}} \) at the borderline condition. For instance, for \( D = 0.15 \) and \( L_i/L_r = 4 \) (\( r = 2 \)), Fig. 12 reads \( f_{\text{mn}} = 0.95 \) and \( V_{\text{dsmn}} = 2.458 \). Using the established value of \( f_{\text{mn}} \), the corresponding values of \( P_n = 0.877 \) and \( Z_{\text{oi}} = 189 \) can be found applying (7), respectively.

The expected value of the peak switch voltage is \( V_{\text{dsm}} = 2.458 \times \sqrt{2} \times 110 = 382 \) [V]. Just as an example, SPA11N60C3 MOSFET, with \( E_{\text{oss}} (400 \text{ V}) = 3.9 \) [uJ] [12], was chosen to implement the switch. Assuming switching loss of \( P_{\text{ON}} = 1 \text{ W} \), the actual switching frequency \( f_s = 256 \text{ kHz} \) was estimated by (23), and the resonant frequency as \( f_{\text{oi}} = (f_s/f_{\text{mn}}) = 270 \text{ kHz} \). Then, the key circuit parameters: 1) \( C_s = 3.1 \text{ nF} \); 2) \( L_i = 112 \text{ uH} \); and 3) \( L_r = 28 \text{ uH} \) were found using (26)–(28).

To verify the design, the LT SPICE simulation program was run. The simulator used realistic model of the switch and took account of parasitic effects. The key simulated waveforms of the QRLEDD design example are shown in Fig. 13. As expected, the QRLEDD operates at the vicinity of the ZCS boundary of the output rectifier. The simulated peak switch voltage at the peak of the line is \( V_{\text{dsm}} = 376 \text{ V} \), matching the expected value of \( V_{\text{dsm}} = 382 \text{ V} \). The simulated average power throughout the line cycle was \( P_{\text{av}} = 29.97 \text{ W} \). The simulated average line current amplitude (Fig. 13) is \( I_{\text{ac}} = 0.374 \text{ A} \).
so that the resulting emulated resistance of the QRLEDD can be approximated $R_e = 155/0.374 = 414 \ \Omega$, as compared with $R_e = 402 \ \Omega$ predicted by (12) using the obtained above values of $P_n$ and $Z_{10}$. The simulation results closely match the design objectives and verify the theory and the proposed design approach.

VII. CONCLUSION

This paper described the key features and the theory of a single switch QRLEDD for lighting applications. Primary advantages of the proposed topology are capacitive isolation, ZV switching turn-off, ZC turn-on, high-frequency operation, inherently low input current distortion, high power factor, and high efficiency. Theoretical predictions were verified by simulation and experiment. Design guidelines were also presented. In a limited range, LED dimming can be attained by adjusting the switch duty cycle and the switching frequency. Better performance, however, can be attained energizing the LED at full power and applying burst control strategy with variable off time to create dimming effect.

The prototype described in the experiment was constructed to validate the proposed concept and theory. No optimization was attempted. Yet, it exhibited a typical efficiency of $\sim 85\%$. Several options exist to improve the QRLEDD efficiency. One possible approach is the application of a synchronous rectifier to replace the output rectifier diode, $D_6$. Improving the efficiency beyond 90% is the focus of ongoing research and will be reported elsewhere.

The topology presented in this paper is subject of patent application by GE and Tel Aviv University.

APPENDIX

The peak voltage across the series capacitor, $C_s$, can be found analyzing the equivalent circuit of State C shown in Fig. 4(c). By the state plane approach (see [27] for a brief review of the method), the normalized peak voltage, $V_{mn}$, can be calculated as

$$V_{mn} = \frac{V_n}{V_i} = M_T + R$$  \hspace{1cm} (29)

where the normalized voltage, $M_T$, applied to the series resonant tank $L_i - C_s$ is

$$M_T = \frac{V_i - V_{LED}}{V_i}$$ \hspace{1cm} (30)

and $R$ is defined by

$$R^2 = (m_2 - M_T)^2 + j^2 Z_{10}.$$ \hspace{1cm} (31)

According to the state analysis above at the beginning of State C, at $t = t_2$, the normalized initial voltage of the capacitor, $C_s$, is

$$m_2 = -\frac{V_{LED}}{V_i}.$$ \hspace{1cm} (32)

The input inductor $L_i$ is linearly charged during both States A and B [see Fig. 4(a) and (b)], so that at the end of State B, the input current is

$$I_2 = i_{i_1}(t_2) = \frac{V_i}{L_i} DT_s.$$ \hspace{1cm} (33)

From (33), the normalized initial current is

$$j_2 = \frac{I_2}{\frac{V_i}{Z_{10}}} = \left(\frac{V_i}{Z_{10}}\right) DT_s.$$ \hspace{1cm} (34)

where $Z_{10}$ is the characteristic impedance as defined above. Combining (29)–(34) yields the normalized peak capacitor voltage

$$V_{mn} = \frac{V_n}{V_i} = 1 - \frac{V_{LED}}{V_i} + \sqrt{1 + \left(\frac{2\pi D f_n}{f_0}\right)^2}.$$ \hspace{1cm} (35)

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