Two-Stage Power Conversion Architecture Suitable for Wide Range Input Voltage

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Two-Stage Power Conversion Architecture Suitable for Wide Range Input Voltage

Seungbum Lim, Student Member, IEEE, John Ranson, Member, IEEE, David M. Otten, Senior Member, IEEE, and David J. Perreault, Fellow, IEEE

Abstract—This paper presents a merged-two-stage circuit topology suitable for either wide-range dc input voltage or ac line voltage at low-to-moderate power levels (e.g., up to 30 W). This two-stage topology is based on a soft-charged switched-capacitor pre-regulator/transformation stage and a high-frequency magnetic regulator stage. Soft charging of the switched capacitor circuit, zero voltage switching of the high-frequency regulator circuit, and time-based indirect current control are used to maintain high efficiency, high power density, and high power factor. The proposed architecture is applied to an LED driver circuit, and two implementations are demonstrated: a wide input voltage range dc-dc converter and a line interfaced ac-dc converter. The dc-dc converter shows 88-96% efficiency at 30 W power across 25-200 V input voltage range, and the ac-dc converter achieves 88% efficiency with 0.93 power factor at 8.4 W average power. Contributions of this work include: 1) demonstration the value of a merged two-stage architecture to provide substantial design benefits in high input voltage, low-power step down conversion applications, including both wide-range-input dc-dc and line-input ac-dc systems; 2) introduction of a multi-mode soft-charged SC stage for the merged architecture that enables compression of an 8:1 input voltage range into a 2:1 intermediate range, along with its implementation, loss considerations and driving methods; and 3) merging of this topology with an resonant transition discontinuous-mode inverted buck stage and pseudo-current control to enable step-down power conversion (e.g., for LED lighting) operating at greatly increased frequencies and reduced magnetics size than with more conventional approaches.

Index Terms—dc-dc, ac-dc, high-frequency, switched capacitor, buck, soft-charging

I. INTRODUCTION

The performance and size of power converters are important for many applications. Achieving small size and high performance is particularly challenging in high-voltage and low-power applications (e.g., voltages up to a few hundred volts and powers up to several tens of watts). In this paper, we explore improved design in this voltage and power range, with a focus on LED driver circuits as an important application in this space. Light emitting diode devices promise unprecedented reductions in energy consumption in comparison to incandescent and fluorescent lights, but come with an as-yet unmet demand for high power density, high efficiency, and high-power-factor LED driver circuitry [1]. An examination of commercial LED drivers illustrates this: considering a group of commercial line-interfaced (120 V ac) LED drivers in the 3–12 W output power range, we have found efficiencies in the range of 64–83% and power factors of 0.73–0.93, with no systems achieving both high efficiency and high power factor as shown in Table I. The switching frequencies of these drivers were in the range of 57–104 kHz, with all having correspondingly low power densities below 5 W/in³. Recently published academic designs are harder to fully evaluate and compare because of large variation on LED output configurations (e.g., separation to multiple LED loads or diverse LED voltage specifications), but appear to provide generally similar performance with moderate improvements in individual aspects [2]–[5]. Overall, the volume of the converters was uniformly dominated by magnetic components, and in each case the driver circuit represented a major contribution to the LED system size. These examinations indicate that power electronics continues to be a significant limitation in solid-state lighting and that there is a need for major improvements in miniaturization and performance in this voltage and power range.

Miniaturization of power electronics requires reductions in magnetic components, which can be realized through increases in the switching frequency [6]. Recent work has shown the potential of high-frequency (or very-high-frequency) operation in systems operating at tens of watts and tens of volts (e.g., [6]–[10]), and there has been preliminary work at hundreds of watts and volts (e.g., [11]). Other switched-capacitor (SC) based design efforts has shown the advantages of merged operation of a switched capacitor (SC) transformation circuit and a high-frequency inductor based regulation circuit (e.g., [12]–[15]), but has only been studied for low-voltage and low power levels (i.e., under 2 watts and ten volts). Thus, achieving the necessary frequency increases at high voltages and modest powers (e.g., at up to hundreds of volts and tens of watts) and realizing the desired miniaturization and performance still remains a major challenge.

This paper introduces a merged two-stage architecture,
which operates at high voltages, low powers, and across a substantial miniaturization of the driver conversion circuit voltage range (e.g., 25–200 V input voltage). Lastly, we target applications we seek efficient operation across a wide input voltage and ac line voltage. Finally, Section IV of the paper presents experimental results demonstrating the proposed approach, including implementations for both wide-range dc voltage and ac line voltage. Finally, Section V concludes the paper.

II. SYSTEM REQUIREMENTS AND DESIGN CONSIDERATIONS

The system architecture we adopt targets LED driver circuits operating either from a wide-range dc voltage or from grid-scale ac voltages at low-to-moderate power levels. We introduce circuit implementations and control circuitries suitable for power levels of up to a few tens of watts, wide-range input voltages of up to 200 V (e.g., compatible with 120 Vrms ac operation), and LED string voltages in the range of 30–40 V. Furthermore, for ac-line applications we seek designs having high power factor (e.g., Power factor > 0.9), while for dc-input applications we seek efficient operation across a wide input voltage range (e.g., 25–200 V input voltage). Lastly, we target substantial miniaturization of the driver conversion circuit through a combination of system architecture, circuit topology, and adoption of greatly increased switching frequencies.

We begin with an overview of the design considerations in achieving miniaturization of a converter in this application, which operates at high voltages, low powers, and across a wide range of input voltage. To attain greater miniaturization, increases in switching frequency are necessary because the values of inductors and capacitors vary inversely with switching frequency. However, the sizes of passive components do not necessarily decrease monotonically with frequency, owing to magnetic-core loss, voltage breakdown, and heat transfer limits [6], [16]–[20]. Consequently, achieving substantial miniaturization through high frequency operation further relies upon appropriate passives design and careful selection of circuit topology to minimize the demands placed upon the passive components, especially the magnetic components.

A further consideration is that while passive component size may be made smaller with increased frequency, size reduces relatively slowly with frequency in practice. Moreover, some parasitic elements, such as semiconductor device parasitic capacitances and inductances, do not scale down at all with increased operating frequency. Consequently, the effects of parasitics become increasingly important as operating frequency increases. To achieve miniaturization through extreme high-frequency operation, an effective circuit topology should inherently absorb important parasitic components in its operation.

A third consideration specifically relates to operating characteristics at high voltages and low power levels. Converters operating at high voltages and low currents operate at high characteristic impedance levels, and consequently utilize relatively large inductors and small capacitors (e.g., characteristic impedance $Z_0 = \sqrt{L/C}$ scales as $V/I$ [6]). Furthermore, the values of inductors and capacitors scale down with increasing resonant frequency (e.g., $\omega_0 = 2\pi f = 1/\sqrt{LC}$). Thus, for a given topology operating at high voltage and low current,
increasing switching frequency beyond a certain point may lead to capacitance values that are too small to be practically achievable (e.g., given parasitic capacitance levels) placing a practical bound on frequency and miniaturization. For miniaturization of converters at high voltage and low power, it is preferable to select system architectures and circuit topologies that require relatively low characteristic impedance values (i.e., yielding smaller inductances and larger capacitances) to reduce constraints on scaling up in frequency.

Lastly, operating range and control scheme are important architectural considerations. We consider designs operating either from a wide range dc voltage or from an ac line voltage, such that the system sees a wide range of input voltages. Zero-voltage switching (ZVS) techniques are necessary to reduce capacitive discharge loss and to achieve high-frequency operation at high voltages. However, soft-switching is often difficult to maintain across a wide input voltage range. It is thus an important challenge to develop designs that maintain the desired soft-switched operation across wide voltage conversion ranges.

III. System Architecture and Circuit Topology

A. System Architecture

To address the above considerations, a merged-two-stage architecture is proposed, as shown in Fig. 1. The first stage is a variable-topology switched-capacitor (SC) circuit operating at moderate switching frequencies (e.g., tens to hundreds of kHz). The SC circuit can achieve high power density and efficiency at these frequencies because it employs only switches and capacitors and incorporates soft-charging operation [12], [13], [21]. However, the SC converter alone cannot efficiently provide the fine voltage regulation capability needed in this application [22]. Instead, this stage serves both to reduce the voltage range over which the second stage needs to operate, and to reduce the maximum voltage level (and hence impedance level) for which the second stage must be designed, in keeping with the design considerations described in the previous section.

The second stage is a magnetic-based stage that provides both additional voltage transformation and fine voltage regulation, and is operated at high frequency (e.g., HF, 3-30 MHz) in order to minimize magnetic component size. High-frequency operation is more readily achieved with high efficiency in the second stage because it operates at lower voltages and smaller voltage range with voltage transformation of the SC first stage. Furthermore, as described below, the topology of the second stage is selected such that it requires relatively small inductor values and inherently absorbs device parasitic capacitance (i.e., the drain-source capacitance of a switch transistor and the diode capacitance) as part of circuit operation, addressing some of the design considerations described in the previous section.

This architecture enables partitioning of the device requirements into “slow switching and high-voltage first stage” and “fast switching and low-voltage second stage” device categories. Moreover, as detailed below the two stages are designed to operate together (merged) in a manner that enables higher efficiency and higher power density than could be achieved in a conventional architecture having separate stages [12], [13], [21], [22]. We describe each of these stages in more detail below.

B. First Stage – Switched Capacitor Circuit

1) SC Power Stage:

The first stage is a variable-topology SC circuit that requires only two energy transfer capacitors and eight switches, and provides nearly continuous input and output currents. The fixed frequency 50% duty ratio gate signal is generated in the microcontroller, and the proper conversion mode is selected depending on the input voltage level.

![Fig. 4. The switched capacitor pre-regulator / transformation stage is comprised of two energy transfer capacitors and eight switches, and provides nearly continuous input and output currents. The fixed frequency 50% duty ratio gate signal is generated in the microcontroller, and the proper conversion mode is selected depending on the input voltage level.](image_url)

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<th>state B</th>
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<tbody>
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<td>M2, M4, M5, M7</td>
</tr>
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<td>M2, M4, M6, M7</td>
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<td>2:1</td>
<td>M1, M4, M5, M8</td>
<td>M2, M3, M6, M7</td>
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and eight switches. The switches turn on and off with 50% duty ratio with dead time. Table II shows the on-state switches in each state and conversion mode, and Table III presents the circuit configuration with this switch operation, where the current source \( i_{\text{int}} \) models the time-averaged currents drawn by the high-frequency regulation stage.

Additionally, our system is designed to “merge” operation of the two stages. A benefit of such a “merged two-stage” architecture [12], [13] is that the second high-frequency stage can “soft charge/discharge” the energy transfer capacitors in the SC stage, reducing loss and / or required capacitor size in the SC stage. In 1:2 step up conversion mode, the capacitors are soft discharged via the high-frequency second stage, and they are hard charged via the fixed input voltage source; thus in 1:2 conversion mode the SC circuit has a partial soft-charging characteristic. In 2:1 step down conversion mode, a portion (nominally half) of the second stage current \( (i_{\text{int}}) \) charges one capacitor and the remainder of the current discharges the other capacitor. The two capacitors charge and discharge such that the total voltage across them remains at the input voltage; thus in 2:1 conversion mode the SC circuit entirely operates with a soft-charging characteristic. Because the HF stage charges and discharges the SC stage as a current source (as opposed to high-current impulsive charging), the SC circuit can operate at low switching frequency (and have attendant low switching loss) while attaining low conduction losses. This loss of the SC stage is thus proportional to the frequency of the SC circuit. The other consideration is the intermediate voltage ripple over which the HF stage can operate well. Because of the soft-charging characteristic of the SC stage in our system architecture, the main dissipation comes from hard-switching of the switches. This loss of the SC stage is thus proportional to the frequency of the SC circuit. The other consideration is the intermediate voltage ripple that the HF regulation stage can tolerate. For a given energy transfer capacitor value in the SC circuit (i.e., \( C_1, C_2 \) of Fig. 4), the intermediate voltage ripple increases as the frequency is reduced. Therefore, the frequency should be properly chosen as a compromise between dissipation of the SC stage and intermediate voltage ripple. Our current HF regulation stage topology can tolerate substantial voltage ripple in the intermediate voltage; thus, the SC circuit can be operated at frequencies as low as 20-30 kHz with acceptable

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<td>1:2</td>
<td>+−</td>
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<tr>
<td>1:1</td>
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<td>+−</td>
</tr>
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<td>2:1</td>
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Fig. 5. The switched capacitor circuit comprises transistors, energy transfer capacitors, and bootstrap diodes and capacitors. The switches \( M_1 - M_9 \) are realized with back-to-back connections of two transistors to block bidirectional voltage. For the control circuitry, the logic power supply \( V_{\text{log}} (=5.5\, \text{V}) \) is applied to a set of bootstrap diodes and capacitors to generate appropriate floating supplies for driving the transistors. It should be noted that this bootstrap configuration applies for all conversion modes.

2) SC Stage Control and Driver Circuit:

The SC transformation stage is controlled by the microcontroller, independent of the HF regulation stage operation. The switches of SC stage are controlled to turn on and off by the microcontroller at a fixed frequency with 50% duty ratio.

To select the frequency of the SC circuit, two specifications should be considered: the dissipation of SC stage and the intermediate voltage ripple over which the HF stage can operate well. Because of the soft-charging characteristic of the SC stage in our system architecture, the main dissipation at the SC circuit comes from hard-switching of the switches. This loss of the SC stage is thus proportional to the frequency of the SC circuit. The other consideration is the intermediate voltage ripple that the HF regulation stage can tolerate. For a given energy transfer capacitor value in the SC circuit (i.e., \( C_1, C_2 \) of Fig. 4), the intermediate voltage ripple increases as the frequency is reduced. Therefore, the frequency should be properly chosen as a compromise between dissipation of the SC stage and intermediate voltage ripple. Our current HF regulation stage topology can tolerate substantial voltage ripple in the intermediate voltage; thus, the SC circuit can be operated at frequencies as low as 20-30 kHz with acceptable
capacitor size. An example loss calculation is illustrated with mathematical expressions in the appendix.

The switched capacitor circuit is realized with power transistors \((M_1 - M_8)\), bootstrap diodes \((D_{b1} - D_{b8})\), energy transfer capacitors \((C_1 - C_2)\), and bootstrap capacitors \((C_{b1} - C_{b8})\) as shown in Fig. 5. It can be seen that the switches M3, M4, M5, and M6 are implemented with a back-to-back connection of two transistors to block bidirectional voltage. For the control circuitry, the logic power supply \(V_{log} (= 5.5V)\) powers the switch drivers through a bootstrap driver circuit. Each driver circuit for every switch in the SC circuit was configured with a digital isolator IC, low-side gate driver, and bypass / bootstrap capacitors. The selected digital isolator IC (Adum5240) in our design can be used to generate the floating power supplies (using an internal dc-dc converter) as well as the logic drive signals. However, the dissipation of this converter is unduly high in this application. Thus, to increase the efficiency of the SC control circuit, a separate bootstrap power supply circuit is implemented as illustrated in Fig. 5. The internal power supply in the digital isolator is used to charge the bootstrap capacitors at start-up; after this, the lossy internal power supply capability is disabled and power is provided by the bootstrap circuit. During normal operation the digital isolator IC only operates to shift control signal levels to the flying switch control port.

It should be noted that even in 1:1 conversion mode the bootstrap technique can be achieved by alternatively selecting either the M3&M5 current path or the M4&M6 current path with 50% duty cycle instead of keeping all switches M3, M4, M5, and M6 on continuously. In the bootstrap 1:1 conversion mode, one turns on and off switches in the M3&M5 and M4&M6 paths along with switches M1, M2, M7, and M8 to provide the paths to charge the bootstrap capacitors. The bootstrap configuration shown in Fig. 5 is applicable to all three conversion modes when the switches are controlled as shown in Table II.

C. Second Stage – HF Regulation Circuit

1) HF Regulation Power Stage:

The second stage of our converter, shown in Fig. 6, is a resonant transition discontinuous-mode inverted buck converter operating at frequencies in the vicinity of 5-10 MHz, with ZVS soft-switching over part of its input voltage range, and near-ZVS soft-switching over the rest of the range. Over the portion of its operating range for which it is soft switched, the converter waveforms are similar to those of the “Quasi-Square-Wave” ZVS buck converter [23], though the control techniques are different. The buck converter is “inverted” in the sense that it is designed with “common positives” so that the active switch is referenced to a constant ground potential and the load has its positive terminal referenced to the positive terminal of the HF stage input voltage. The active switch having its control port referenced to a fixed potential greatly simplifies level shifting of drive signals and mitigates capacitive current injection at the switch gate that would otherwise occur owing to large, fast voltage transitions during switching. Referencing the switch in this way is of great practical importance in achieving extreme high-frequency operation [6], and becomes increasingly important at higher switching voltages where level shifting and capacitive injection affecting switching are major concerns.

The buck converter operation cycles through four phases in a HF switching cycle, as illustrated in Fig. 7. During Phase 1,
the switch is on, and the inductor current \( i_L \) ramps up linearly. In Phase 2, the switch turns off, and the switch drain to source voltage \( v_{ds} \) rings up to the intermediate voltage (i.e., \( V_{int} \), HF regulation stage input voltage). In Phase 3, the diode starts to conduct and \( i_L \) ramps down to zero. In Phase 4, both diode and switch are off, and the inductor rings with the net capacitance at the switch drain node \( (C_{sw,d}, \text{switch output capacitance plus diode output capacitance}) \). Inductor current \( i_L \) rings negative and \( v_{ds} \) rings down to zero or near zero volts (down to \( V_{int} - 2V_{out} \)). At this point, the switch is turned back on and the cycle repeats entering phase 1.

Equations (1)–(8) illustrate the inductor current \( i_L \) and switch drain source voltage \( v_{ds} \) in each phase, where \( Z_d = \sqrt{L/C_{sw,d}} \), \( w_d = 1/\sqrt{LC_{sw,d}} \), and \( i_L, pk = (V_{int} - V_{out})t_{on}/L, \) and the nonlinear capacitance variation (relative to voltage bias) is ignored. Equation (9) describes the switching period of the proposed high-frequency buck converter. It can be seen from equation (9) that for a given operating voltage \( (V_{int}, V_{out}) \) and parasitic capacitance level \( (C_{sw,d}) \), a converter with smaller inductance operates over a shorter time period (higher frequency).

Many converters operating from high voltages at low powers (and currents) suffer from large-valued (and consequently large-sized) inductors. A benefit of this topology is that it operates with high inductor current ripple, yielding a relatively small value of inductance and small magnetic energy storage. Moreover, as the system architecture enables a relatively low value of \( V_{int} \), the needed value of inductor \( L \) (at a particular switching frequency) is further reduced as per equation (9). Lastly, the soft-switched (or nearly soft-switched) nature of the topology enables relatively high switching frequency (small period) to be achieved at acceptable loss, further reducing the required inductance value. Thus, both the architecture selected and the second-stage topology help minimize the size of the required magnetics.

2) HF Regulation Stage Control and Driver Circuit:

The specific characteristic of the LED load is that the LED load voltage is nonlinear with respect to the load current and is almost determined by the pre-defined LED string forward drop voltage. Therefore, to control output power level, the current conducting in the LED string load should be adjusted; the fast closed-loop estimated-current control circuit introduced here modulates switch on-time of the HF second stage to adjust the current level. The ability of the controller to provide closed-loop control at the high operating frequency of the system is central to the effectiveness of the proposed approach.

The HF stage control circuit is designed to operate at high frequencies (to beyond 10 MHz) using fast comparators, RC integrators, and logic circuits as illustrated in Fig. 8. The control circuit drives the gate-source voltage \( v_{gs} \) of the switch (turning it on and off) as determined by the drain-source voltage \( v_{ds} \), zero-voltage-switching threshold voltage \( V_{zvs} \), cathode node voltage of the LED load (e.g., \( V_{int} - V_{out} \)), and peak-current threshold voltage \( V_{iref} \); each voltage signal in Fig. 8 is directly connected to one of the HF power conversion stage and microcontroller.

This HF control circuit regulates switching based on two thresholds using a pair of comparators: peak-current threshold voltage \( V_{iref} \) turn-off with comparator \( U_1 \) and ZVS threshold voltage \( V_{zvs} \) turn-on with comparator \( U_2 \). First, comparator \( U_1 \) determines the on-time duration of the switch and turns off the switch by comparing \( V_{iref} \) and the (approximate) integral of voltage \( V_{int} - V_{out} \). When the transistor begins its phase 1 on-state duration, \( V_{ds} \) is low, comparator \( U_2 \) output is low, and transistor \( U_3 \) is off such that the voltage on \( C_{c1} \) approximately integrates voltage \( V_{int} - V_{out} \). After the capacitor \( C_{c1} \) is charged for a short duration \( \Delta t \), \( V_{U_1} = (V_{int} - V_{out})\Delta t/(C_{c1}R_{c1}) \). The output of \( U_1 \) goes high and turns off the switch \( (v_{gs} \text{ low}) \) when \( V_{U_1} \) becomes larger than \( V_{iref} \). Because the proposed high-frequency regulation stage is always operating at the edge of discontinuous conduction mode, as shown in equation (1) the inductor current \( i_L \) during the switch is on-state (phase 1) is as follows:

\[
i_L = \frac{V_{int} - V_{out}}{L} t \quad (10)
\]

\[
i_L, pk = \frac{V_{int} - V_{out}}{L} t_{on} = \frac{V_{iref}C_{c1}R_{c1}}{L} \quad (11)
\]

Thus, the voltage \( V_{iref} \) directly modulates the switch turn-on time to regulate peak current of the inductor, regardless of
the high-frequency regulation stage input voltage \(V_{\text{int}}\) and output voltage \(V_{\text{out}}\).

The turn on of the switch (ending phase 4 and starting phase 1) is determined by comparator \(U_2\). After the switch turns off at the end of phase 1, the resistor-divided \(v_{\text{ds}}\) voltage goes above \(V_{\text{zvs}}\), and transistor \(U_3\) turns on. This discharges the voltage on \(C_{\text{c1}}\), resetting the past-integrated \(C_{\text{c1}}\) voltage for controlling on-time duration in the next cycle and making the output of \(U_1\) low. When the inductor current decreases down to zero at the end of phase 3 and the HF-stage diode turns off, phase 4 begins and the inductor starts to ring with the net capacitance at the switch drain node (That is, the switch drain-source voltage starts to ring down). When switch drain-source voltage \(v_{\text{ds}}\) rings down sufficiently, the resistor-divided \(v_{\text{ds}}\) voltage goes below \(V_{\text{zvs}}\), and the output of \(U_2\) goes low. The control circuit then turns on the switch again, and the cycle repeats. \(V_{\text{zvs}}\) is set such that — including logic delays — the switch is turned on at the correct point in the drain-voltage ringdown.

**D. DC configuration**

In a dc-dc configuration, the converter operates at wide input voltage range (i.e., 25–200 V). The switched capacitor stage changes the conversion mode as the input voltage varies and reduces the intermediate voltage range as shown in Fig. 2. The high frequency regulation stage then efficiently converts from the reduced-range intermediate voltage to the load voltage at a particular power level.

**E. AC configuration**

In an ac-dc configuration, the converter includes line frequency filter and full-bridge rectifier as shown in Fig. 9. The twice-line-frequency energy buffering for single-phase ac to dc conversion is realized with capacitance placed across the system output voltage. In an ac-dc converter high power factor is desired in addition to high efficiency and high power density. In commercial LED driver applications a power factor of at least 0.9 is desired, while in residential applications a power factor of 0.7 is acceptable. To attain high power factor, the input current waveform from the ac-line voltage should be proportional to the sinusoidal voltage waveform. In cases where the voltage waveform is sinusoidal and the current waveform is “clipped” to zero for low voltages (e.g., where the converter ceases to draw input current near ac-line zero voltage crossings), the current waveform providing the highest power factor is a “clipped” sine wave in phase with the voltage, as illustrated in Fig. 10. For a “cut in” voltage of 100 V (for a 170 V peak sinusoid), a power factor of \(> 0.95\) is realized with a clipped sinusoidal current waveform.

The proposed ac-dc converter is designed to approximately draw the current pattern of Fig. 10, with a “cut in” voltage of 100 V. Both the SC first stage and HF second stage are turned off below 100 V input voltage, and the intermediate voltage \(V_{\text{int}}\) remains at 50 V during this off duration.

When the rectified ac voltage is above 100 V, the merged two-stage circuit is controlled to yield an input current that approximates a clipped-sinusoidal waveform. The filtered input current is proportional to inductor peak current \(i_{\text{L,peak}}\) of the HF regulation stage, which can be directly changed by peak-current threshold voltage \(V_{\text{icref}}\) of the HF control circuit as described above. At low voltages, the converter is turned off by setting \(V_{\text{icref}}\) to 0, and when the rectified ac voltage goes above 100V, the \(V_{\text{icref}}\) peak-current threshold voltage is adjusted to be proportional to the rectified ac voltage by the microcontroller; the approximate clipped-sinusoidal input current then flows from the ac line voltage.

In the prototype converter, the ac-dc converter cuts in at 100 V (for a 170 V peak sinusoid) and ideally achieves 0.95 power factor. In LED driver applications, power factors exceeding 0.9 are not currently demanded in the market, so this operation mode is sufficient to provide the desired
TABLE IV
COMPONENTS OF THE SC STAGE AND HF STAGE

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<td>switch</td>
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<td>bootstrap diode</td>
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<td>bypass capacitor</td>
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<td>isolator / isoPower</td>
<td>ADum5240, Analog Devices</td>
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<td>low-side gate driver</td>
<td>FAN3111C, Fairchild</td>
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<tr>
<td>output capacitor</td>
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<td>SN74LVC2G06, Texas Instrument, Inverter with open-drain outputs</td>
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<td>resistor</td>
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<td>capacitor</td>
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<td>220 pF, Ceramic X1Y2 250 V</td>
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<tr>
<td>resistor</td>
<td>442 Ω</td>
</tr>
<tr>
<td>microcontroller</td>
<td>ATMEGA64M1, Atmel</td>
</tr>
</tbody>
</table>

performance. However, in applications for which higher power factor than 0.95 is required, the current conduction duration of the converter can be increased by cutting in at a lower voltage (e.g., 25 V for a 170 V peak sinusoid). To operate in this wide range, the SC stage then can change the conversion modes among 1:2, 1:1, and 2:1 ratios (like the variable-topology SC circuit in dc-dc converter as shown in Fig. 2) and sets the intermediate voltage between 50 V and 100 V. It should be noted that the same two-stage architecture can be applied for universal ac line voltage, but the appropriate voltage rate components and a higher step-down conversion ratio HF regulation stage are necessary.

IV. EXPERIMENTAL RESULTS

Two implementations of the proposed converter have been developed. One operates from a wide-range dc input voltage (25-200 V dc), while the second operates with a front-end filter and full-bridge diode rectifier from 120 Vrms ac line voltage. The components and parameters chosen for the dc-dc converter and ac-dc converter implementations are listed in Table IV. Some additional auxiliary bypass capacitors, resistors, and protection diodes are used to connect the converter circuit to the logic voltage and the microcontroller.

In the HF regulation stage, ZVS soft-switching condition is necessary to achieve high efficiency. Fig. 12 shows a experimental measurement of switch gate voltage $v_{ds}$ and switch drain voltage $v_{ds}$ of the HF stage in the merged two-stage converter, operating at a 60 V dc input voltage and a 35 V and 12.15 W load. The measured waveform illustrates high-frequency operation of the proposed HF regulation stage with ZVS soft-switching (i.e., the HF regulation stage operates at 7.8 MHz at this input voltage and power level. The switching frequency is changed and determined by the input voltage and power level, and the converter operates in high-frequency (3–30 MHz) region with ZVS (or near ZVS) condition).

Moreover, as previously discussed, soft-charging of capacitors is desired to boost the efficiency of the SC transformation stage. When the SC circuit operates in 2:1 step-down mode, it is fully soft-charged, and energy transfer capacitors charge and discharge based on current from the regulation stage. Fig. 13 shows the drain-source voltage ($v_{ds}$) of the HF regulation stage and the intermediate voltage ($v_{int}$) between SC transformation stage and HF regulation stage, when the converter operates from 165 V dc voltage. As can be seen, the intermediate voltage slowly discharges during each SC state, illustrating soft charging of the SC (first) stage by the HF (second) stage (merged two-stage operation). In the 1:2 mode, there is a mixture of soft- and hard-charging of the capacitors.

The microcontroller, SC stage and HF stage control circuit, and linear regulators are supplied by an external 5.5V power supply and dissipate about 0.6 W. The control circuit of the SC stage operates at 50 kHz frequency and uses about 0.3 W power to drive 8 level shifters and 8 low-side gate drivers. The HF control circuit operates in the 3-30 MHz frequency and
Fig. 12. Experimental $v_{gs}$ and $v_{ds}$ waveforms showing zero-voltage switching of the HF regulation stage ($x$: 40ns/div, $y$: $v_{gs}$: 2V/div, $y$: $v_{ds}$: 20V/div). The HF stage input voltage is 60 V and the output is a 35 V, 12.15 W load. The converter operates at 7.85 MHz under this condition.

Fig. 13. The implemented dc-dc converter is connected to 165 V dc voltage, where the SC circuit operates in 2:1 step down conversion mode. Ch1 (blue, $y$: 20V/div) shows the measured switch drain-source voltage $v_{ds}$ of the HF regulation stage, and Ch2 (red, $y$:20V/div) is the intermediate voltage ($V_{int}$) between SC transformation stage and HF regulation stage. It should be noted that the droop of intermediate voltage during each SC state shows the soft-charging characteristic of the SC (first) stage by the HF (second) stage (merged two-stage operation).

Fig. 14. The ac-dc implementation including a bridge rectifier and

sets the intermediate voltage to the desired operating range such that the HF regulation stage can operate more efficiently. Thus, as can be seen in Fig. 14, the efficiency makes discrete transitions when the input voltage crosses 50 V and 100 V, and owing to the mode changes the system maintains high efficiency across a wide input voltage range.

The ac-dc implementation including a bridge rectifier and
The experimental prototypes of both a dc-dc and an ac-dc converter using this architecture are implemented for LED driver applications and presented along with experimental results. Through appropriate control and merged operation between the SC transformation and HF regulation stages, soft-charging of the SC stage and ZVS switching of the HF stage are achieved. Additionally, this proposed architecture and topological approach are likewise expected to be applicable to not only this LED driver circuit but also a variety of electrical applications which have wide-range dc input voltage or ac line input voltage.

### APPENDIX

In this appendix we explore the dissipation and the intermediate voltage ($V_{int}$) fluctuation of the switched capacitor circuit, and illustrate how to choose a desired operating frequency of the switched capacitor circuit as a compromise among several considerations.

#### A. Dissipation of the Switched Capacitor Circuit

The dissipation of the switched capacitor circuit comes from two primary sources: energy transfer capacitor charge / discharge loss and switch output capacitor charge / discharge loss (we neglect gating loss, as this is a modest consideration here.)

The dissipation during charging and discharging the energy transfer capacitors can be calculated and estimated with a Thévenin model of the SC circuit [22]. In the Thévenin model, the output resistance represents the loss, and it changes with the operating frequency and capacitance level in the switched capacitor circuit. In the model of SC circuit with hard-charging operation, the output resistance can be plotted across the switching frequency with two asymptotes, the slow-switching-limit (SSL) and the fast-switching-limit (FSL), respectively. For instance, in our proposed 2:1 SC configuration, the effective SC stage Thévenin output resistance can be calculated as $1/(8C_{sw})$ in SSL and $3R_{sw, on}$ in FSL. Fig. 17 shows the output resistance of the hard-charging SC circuit across the frequency with different capacitance levels. To decrease this output resistance, the switched capacitor circuit can be operated with either larger capacitance or higher operating frequency, which consumes either more area or more loss from charging and discharging non-zero switch output capacitances (the second source of SC circuit dissipation described below).

Compared to hard-charging operation, soft charging operation adiabatically charges and discharges the energy transfer capacitors via the regulation stage, which allows it to operate at the FSL dissipation levels with small capacitance and low operating frequency. Fig. 17 also shows the output resistance of the proposed SC circuit with 2:1 conversion mode and soft-charging operation. With the output resistance of the Thévenin model, the first source of dissipated power in the 2:1 conversion mode of the SC circuit can be calculated as follows:

$$P_{SC-diss, 1} = i_{int}^2 R_{out} = \left(\frac{i_{in}}{2}\right)^2 3R_{sw, on} \quad (12)$$

### TABLE V

<table>
<thead>
<tr>
<th>TABLE V</th>
<th>The Specification of Implemented Prototype Converter</th>
</tr>
</thead>
<tbody>
<tr>
<td>output power</td>
<td>up to 30 W</td>
</tr>
<tr>
<td>SC stage switching frequency</td>
<td>50 kHz</td>
</tr>
<tr>
<td>HF stage switching frequency</td>
<td>varying in 3-30MHz (HF region)</td>
</tr>
<tr>
<td>Efficiency</td>
<td>up to 96%</td>
</tr>
<tr>
<td>Power Factor</td>
<td>0.93</td>
</tr>
<tr>
<td>Expected Power density (after opt.)</td>
<td>$&gt; 60 \text{ W/in}^3$</td>
</tr>
</tbody>
</table>
The second source of the loss in the SC circuit comes from the non-zero output capacitance across each switch. These output capacitances charge or discharge during every state change. For example, in 2:1 SC configuration, two switches \( M_7 \) and \( M_8 \) block \( V_{in} \) and the other switches (\( M_1-M_6 \)) block \( V_{in}/2 \), causing dissipation as follows:

\[
P_{SC-diss,2} = C_{oss} (2V_{in}^2 + 6(V_{in}/2)^2) f_{sw} \quad (13)
\]

\[
= 3.5 C_{oss} V_{in}^2 f_{sw} \quad (14)
\]

(In the equivalent model of the SC circuit, this loss could be captured by a shunt resistor across the converter input port of value \( R_{eq} = 1/(3.5C_{oss} f_{sw}) \).) It should be noted that while this loss can be minuscule in a low-voltage SC circuit, in a high voltage SC circuit this loss becomes significant.

Overall, the total dissipation can be calculated by adding the two sources of loss components. Fig. 18 shows the estimated total dissipation across the operating frequency when the proposed 2:1 conversion mode SC circuit operates at 150 V dc input voltage and 30 W power level. In this example, we supposed 0.1 \( \Omega \) switch on-resistance (\( R_{sw,on} \)) and 100 pF switch output capacitance (\( C_{oss} \)) referencing the EPC 2012 characteristic.

As can be seen in Fig. 18, soft-charging is always better for higher efficiency and higher power density compared to hard-charging operation. It should be noted that at this power, operating voltage, and soft-charging operation, the loss component from switch output capacitance is a dominant factor, and it is proportional to the SC operating frequency.

\[ V_{in} \] voltage fluctuation of the switched capacitor circuit

The essence of the soft-charging characteristic in SC circuit is to allow voltage fluctuation at the intermediate voltage, and adiabatically charge / discharge of energy transfer capacitors with a following regulation converter. The fluctuating intermediate voltage should reside in the operating range of the following HF regulation stage. The voltage fluctuation can be easily calculated at a particular power level, and here are example equations for the proposed 2:1 configuration case:

\[
\Delta Q_C = \frac{i_{int} t}{2} = \frac{i_{int}}{2} \frac{1}{2 f_{sw}} = \frac{i_{int}}{4 f_{sw}} = C \Delta V_{int} \quad (15)
\]

\[
\Delta V_{int} = \frac{i_{int}}{4C f_{sw}} = \frac{P_{out}}{4C f_{sw} V_{int}} \approx \frac{P_{out}}{2C f_{sw} V_{in}} \quad (16)
\]

The \( V_{int} \) voltage fluctuation is plotted across frequency as shown in Fig. 19, supposing 150 V input voltage and 30 W power level. As can be seen, the \( V_{int} \) voltage shows less voltage ripple at higher capacitance and higher operating frequency.

As illustrated above, the dissipation and intermediate voltage (\( V_{int} \)) fluctuation of the switched capacitor circuit are closely related to the switching frequency, capacitance level, and type of SC operation (hard charging or soft charging). It is thus essential to select a desired operating frequency compromising these trade-offs. Considering several characteristics at three different conversion modes simultaneously, the
The proposed SC circuit was designed with 1 μF energy transfer capacitors and operated with soft-charging characteristic at 50 kHz frequency. It should be noted that if wide output power range operation are desired, one might choose to also modulate the SC converter frequency with output power.

REFERENCES


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