Series-Connected High Frequency Converters in a DC Microgrid System for DC Light Rail Transit

Authors:

Bor-Ren Lin

Date Submitted: 2020-01-23

Keywords: light rail transit, half-bridge converter, zero-voltage switching, DC microgrid

Abstract:

This paper studies and presents a series-connected high frequency DC/DC converter connected to a DC microgrid system to provide auxiliary power for lighting, control and communication in a DC light rail vehicle. Three converters with low voltage and current stresses of power devices are series-connected with single transformers to convert a high voltage input to a low voltage output for a DC light rail vehicle. Thus, Metal-Oxide-Semiconductor Field-Effect Transistors (MOSFETs) with a low voltage rating and a turn-on resistance are adopted in the proposed circuit topology in order to decrease power losses on power switches and copper losses on transformer windings. A duty cycle control with an asymmetric pulse-width modulation is adopted to control the output voltage at the desired voltage level. It is also adopted to reduce switching losses on MOSFETs due to the resonant behavior from a leakage inductor of an isolated transformer and output capacitor of MOSFETs at the turn-on instant. The feasibility and effectiveness of the proposed circuit have been verified by a laboratory prototype with a 760 V input and a 24 V/60 A output.

Record Type: Published Article

Submitted To: LAPSE (Living Archive for Process Systems Engineering)

Citation (overall record, always the latest version): LAPSE:2020.0119
Citation (this specific file, latest version): LAPSE:2020.0119-1
Citation (this specific file, this version): LAPSE:2020.0119-1v1

DOI of Published Version: https://doi.org/10.3390/en11020266

License: Creative Commons Attribution 4.0 International (CC BY 4.0)
Article

Series-Connected High Frequency Converters in a DC Microgrid System for DC Light Rail Transit

Bor-Ren Lin

Department of Electrical Engineering, National Yunlin University of Science and Technology, Yunlin 640, Taiwan; linbr@yuntech.edu.tw; Tel.: +886-9-1231-2281

Received: 7 January 2018; Accepted: 18 January 2018; Published: 23 January 2018

Abstract: This paper studies and presents a series-connected high frequency DC/DC converter connected to a DC microgrid system to provide auxiliary power for lighting, control and communication in a DC light rail vehicle. Three converters with low voltage and current stresses of power devices are series-connected with single transformers to convert a high voltage input to a low voltage output for a DC light rail vehicle. Thus, Metal-Oxide-Semiconductor Field-Effect Transistors (MOSFETs) with a low voltage rating and a turn-on resistance are adopted in the proposed circuit topology in order to decrease power losses on power switches and copper losses on transformer windings. A duty cycle control with an asymmetric pulse-width modulation is adopted to control the output voltage at the desired voltage level. It is also adopted to reduce switching losses on MOSFETs due to the resonant behavior from a leakage inductor of an isolated transformer and output capacitor of MOSFETs at the turn-on instant. The feasibility and effectiveness of the proposed circuit have been verified by a laboratory prototype with a 760 V input and a 24 V/60 A output.

Keywords: DC microgrid; zero-voltage switching; half-bridge converter; light rail transit

1. Introduction

Direct current (DC) microgrids are studied to combine alternative current (AC) utility power, renewable energy sources, energy storage units and local DC and AC loads in order to reduce global warming and climbing temperature issues. The common DC voltage on DC microgrids can be 1500 V for traction vehicles; 760 V for light rail vehicles and industry applications; 380 V for residential and commercial buildings. For light rail transits, low frequency power transformers are used in conventional light rail vehicles to provide electrical isolation. The main drawback of line frequency transformers [1–3] is bulk volume. To avoid using bulky line frequency transformers, transformerless converter topologies [4–8] with primary-series and secondary-parallel connections have been proposed to lessen the blocking-voltage capability of active devices on the high voltage side and the current stress of power components on the low voltage side. Half-bridge (HB) or full-bridge (FB) converters have been widely adopted for medium power applications to provide a stable and low voltage output. Conventional FB converters and HB converters using high voltage rating devices such as insulated gate bipolar transistors (IGBTs) are widely adopted to convert a high voltage input to a low voltage output. However, the switching frequency of general IGBT devices is less than 50 kHz. To improve the low switching frequency problem of IGBT devices, multilevel converters [9–11] using Metal-Oxide-Semiconductor Field-Effect Transistor (MOSFET) devices have been developed in medium power converters to reduce converter volume. However, the high switching frequency will also increase core losses on magnetic components and switching losses on power MOSFETs. Zero-voltage switching converters [12–15] have been proposed to achieve a low switching loss. Asymmetric pulse-width modulation [16–18], frequency modulation [19–21] and a phase-shift pulse-width modulation (PWM) scheme [22–24] are general PWM control schemes used in DC/DC converters to regulate load voltage.
and also realize the mechanism of the zero-voltage turn-on switching. Therefore, the high efficiency converters can be obtained. The asymmetrical PWM scheme can reset the magnetizing flux in every switching cycle. Therefore, the voltage spike on drain-to-source of power MOSFET can decrease the safety operation region and the electromagnetic interference can be reduced.

This paper presents a series-connected soft switching converter using a single transformer for light rail vehicle applications. The adopted circuit topology includes three series-connected HB circuits on the primary side to reduce the blocking-voltage stress of power MOSFETs. Thus, low turn-on resistance and low voltage rating MOSFETs can be adopted to lessen conduction losses on power devices. The input split voltage balance is achieved by two flying capacitors. The asymmetric PWM scheme is employed to control power switches. The output capacitor of power MOSFETs and the leakage inductor of the transformer are resonant at the transition interval. Thus, the mechanism of the zero-voltage turn-on switching is realized and the electromagnetic interference can be reduced. Current doubler rectifier topology is used on the low voltage side to achieve a ripple current cancellation. Therefore, the resultant ripple current at the output capacitor is reduced. Three HB circuits share one transformer on the primary side so that the primary currents of the three HB circuits are reduced to one-third of the primary current in a conventional HB converter. Therefore, the conduction losses on active switches are reduced and the circuit efficiency is improved. The feasibility of the proposed circuit is verified with a 1.44 kW converter. This paper is organized as follows: in Section 2, the circuit configuration and operation principle of the developed converter used in the DC microgrid are presented and discussed in detail. In Section 3, the circuit characteristics of the developed circuit are presented. Experimental results based on a laboratory prototype are shown and discussed in Section 4. Finally, Section 5 presents the conclusions.

2. Proposed Converter

The basic circuit structure of a DC microgrid is given in Figure 1. The bidirectional AC/DC converters are used between the AC utility system and the DC microgrid to stabilize the DC bus voltage. Renewable energy sources from solar cell panels and wind power generators provide clear energy to the DC microgrid system through the unidirectional DC/DC converters and the AC/DC converters. Energy storage units are used to store/release energy from/to the DC microgrid through the bidirectional DC/DC converters in order to stabilize the DC bus voltage. In order to increase the voltage reliability, the bipolar voltage system (+\(V_{dc}\), −\(V_{dc}\) and neutral line) can be adopted on the DC microgrid. When a fault condition in one of the DC poles occurs, the DC power is still supplied to the local load by the other two wires and an auxiliary converter. It is clear that the reliability of the DC microgrid system is increased during fault condition. Therefore, three voltage levels (2\(V_{dc}\), +\(V_{dc}\) and −\(V_{dc}\)) can be used to provide industry and transportation with the 2\(V_{dc}\) voltage level and future residential and commercial buildings with the +\(V_{dc}\) or −\(V_{dc}\) voltage level. To keep two voltage levels (+\(V_{dc}\) and −\(V_{dc}\)) balanced during unbalanced loads, a voltage balancing circuit is normally needed to realize this goal. The DC/DC converter can be used to convert a high input voltage to a low output voltage and it does not relate to the bipolar voltage system. For a DC light rail vehicle, the input voltage is normally around 760 V. Power devices such as the 1500 V IGBT can be adopted in two-level power converters to supply auxiliary low voltage power for control units, telecommunication units and lighting systems in the light rail vehicle. The MOSFETs with a 600 V rating can be adopted in the three-level power converter to achieve the same goal but with a much higher switching frequency to reduce converter size. The studied converter is concentrated on power conversion for light rail vehicle applications. Figure 2a shows the basic block diagrams of power distribution in a conventional light rail vehicle. First, 750 \(V_{dc}\) is converted to the three-phase AC voltage 380 \(V_{ac}\) by the DC/AC inverter. Second, the AC/DC converter is adopted to convert 380 \(V_{ac}\) to a low DC voltage in an auxiliary power distribution system to supply a necessary amount of DC voltage for the battery bank, lighting equipment, door equipment, control power units and communication equipment in a light rail vehicle system. The AC/DC/AC converter is also adopted to convert 380 \(V_{ac}\) to a variable
AC voltage output for the AC motor drive. If the power of the DC light rail vehicle is supplied directly from the DC microgrid, then some power conversion stages can be saved to reduce costs and increase circuit efficiency. The basic power distribution diagrams in the proposed light rail vehicle are illustrated in Figure 2b. The auxiliary power in a light rail vehicle is directly converted from 760 V through a DC/DC converter. The AC motor drive and the air compressor are controlled by the DC/AC converters. Therefore, the AC/DC conversion can be saved to reduce costs and increase circuit efficiency and reliability.

The studied DC/DC converter is given in Figure 3. It directly converts a 760 V voltage input from the DC microgrid to a low voltage output by using low voltage rating power MOSFETs to provide auxiliary power in a light rail vehicle. The studied converter includes three HB circuits connected in a multilevel inverter to reduce voltage stress and balance input split voltages. Thus, two flying circuit efficiency. The basic power distribution diagrams in the proposed light rail vehicle are illustrated in Figure 2b. The auxiliary power in a light rail vehicle is directly converted from 760 V through a DC/DC converter. The AC motor drive and the air compressor are controlled by the DC/AC converters. Therefore, the AC/DC conversion can be saved to reduce costs and increase circuit efficiency and reliability.

The studied DC/DC converter is given in Figure 3. It directly converts a 760 V voltage input from the DC microgrid to a low voltage output by using low voltage rating power MOSFETs to provide auxiliary power in a light rail vehicle. The studied converter includes three HB circuits connected in a primary-series secondary-parallel with a single transformer. The primary sides of the HB circuits are connected in a series so that the voltage stress of the power switches $Q_1$~$Q_6$ is reduced to $V_{in}/3$ and...
the low turn-on resistance of power MOSFETs is adopted to reduce conduction losses and increase circuit efficiency. Each HB circuit provides $P_v/3$ to the output load. Flying capacitors are widely used in a multilevel inverter to reduce voltage stress and balance input split voltages. Thus, two flying capacitors $C_1$ and $C_2$ are used on the high voltage side to achieve an input voltage balance of $V_{C1} \sim V_{C3}$. The current doubler rectifier is used on the low voltage side to reduce the output ripple current. 

Figure 4 shows the timing sequence of switches $Q_1 \sim Q_6$ and the main PWM waveforms in the proposed converter. The operation principle and circuit analysis of the studied circuit are assumed under the following conditions: (1) the output capacitor $C_o$ is large enough to be treated as a constant voltage $V_o$, (2) power devices $Q_1 \sim Q_6$, $D_1$ and $D_2$ are ideal, (3) $C_f = C_r = C_f$, $C_{b1} = C_{b2} = C_{b3} = C_{b}$, $L_1 = L_2 = L_3 = L_r$, and $L_{m1} = L_{m2} = L_{m3} = L_{m}$, (4) $n_1 = n_2 = n_3 = n_p/3$, and (5) $V_{C1} = V_{C2} = V_{C3} = V_{C_b}$ and $V_{C1} = V_{C2} = V_{C3} = V_{C_f1} = V_{C_f2} = V_{m}/3$. The duty cycles of $Q_1$, $Q_3$ and $Q_5$ are $d$. Conversely, the duty cycles of $Q_2$, $Q_4$ and $Q_6$ are $1 - d$. The equivalent circuits of the operating steps in the proposed circuit are shown in Figure 5. There are eight operating stages for every switching period.

Figure 3. Circuit schematic of the studied converter.

Figure 4. Main waveforms during one switching period.
**Step 1 [t₀~t₁]:** Before step 1, power switches Q₁, Q₃ and Q₅ are conducting and the output current freewheels through D₁ and D₂. After time t₀, i₁p₁ is decreased to zero and D₁ becomes reverse biased. i₀ flows through D₂. Power is transferred from V_in to R₀ in this step. In step 1, V_C₁ = V₁, V_C₂ = V_C₃, V_C₄,d = V₁, V_C₅,d = V₃, V_C₆,d = V_C₄, V_Lₕ = V₁, V_Lₙ = V₃, V_Lₐ = V₁, V_L₂ = V₃, V_L₃ = V₁, V_L₄ = V₃, V_Lₕ₂ = V₁, V_Lₙ₂ = V₃, V_Lₐ₂ = V₁, V_L₂₂ = V₃, V_L₃₂ = V₁, V_L₄₂ = V₃, V_Lₕ₃ = V₁, V_Lₙ₃ = V₃, V_Lₐ₃ = V₁, V_L₂₃ = V₃, V_L₃₃ = V₁, V_L₄₃ = V₃. Since all circuit components in three HB circuits are identical and n₁ = n₂ = n₃ = n₄/n₅, it can obtain v_Lₕ₁ = v_Lₙ₁ = v_Lₐ₁ = v_L₂₁ = v_L₃₁ = v_L₄₁ = V₁, v_L₂₂ = v_L₃₂ = v_L₄₂ = V₃, v_Lₕ₃ = v_Lₙ₃ = v_Lₐ₃ = V₁. Therefore, V₁ = V₃ = 0. i₁p₁~i₃p₃ and i₁o₁ increase and i₁o₂ decreases.

**Step 2 [t₁~t₂]:** When Q₁, Q₃ and Q₅ are turned off at time t₁, the positive primary currents i₁p₁~i₃p₃ rapidly discharge C₁, C₄ and C₆, respectively. On the other hand, C₁, C₃ and C₅ are charged by i₁p₁~i₃p₃ respectively. The secondary winding voltage vₙ₅ is increased in this step.

**Step 3 [t₂~t₃]:** When C₁, C₃ and C₅ are charged to vₙ₅, v₁ and v₃ are discharged respectively, then the secondary winding voltage vₙ₃ = 0. Then the output current i₀ freewheels through D₁ and D₂. In this operation step, v_Lₕ₁ = v_Lₙ₁ = v_L₂₁ = v_L₃₁ = v_L₄₁ = 0, then the current i₀ freewheels through D₁ and D₂. The duty loss in step 3 is obtained as (1) shows:

\[
d_{loss,4} = \frac{\triangle t_{sw}}{T_{sw}} = \frac{L_m I_m f_{sw}}{3n_1 V_{in} C_{th}}
\]

**Step 5 [t₄~t₅]:** After t₄, i₁p₂ is decreased to zero so that D₂ is reverse biased. i₀ flows through D₁, L₁, and L₂. In step 5, power is transferred from V_in to R₀, V_C₁ = V₃, V_C₂ = V₃, V_C₃ = V₁, V_C₄,d = V₃, V_C₅,d = V₃, V_C₆,d = V₄, V_Lₕ = V₃, V_Lₙ = V₄, V_Lₐ = V₃, V_L₂ = V₄, V_L₃ = V₃, V_L₄ = V₄. Since all circuit components in three HB circuits are identical and n₁ = n₂ = n₃ = n₄/n₅, it can obtain v_Lₕ₁ = v_Lₙ₁ = v_Lₐ₁ = v_L₂₁ = v_L₃₁ = v_L₄₁ = V₃, v_L₂₂ = v_L₃₂ = v_L₄₂ = V₄. Thus, i₁p₁~i₃p₃ and i₁o₁ increase and i₁o₂ decreases.

**Step 6 [t₅~t₆]:** When Q₂, Q₄ and Q₆ turn off at time t₅, the negative primary currents i₁p₁~i₃p₃ rapidly charge C₂, C₄ and C₆, respectively. On the other hand, C₂, C₃ and C₅ are discharged by i₁p₁~i₃p₃ respectively. The secondary winding voltage vₙ₄ is increased in this step.

**Step 7 [t₆~t₇]:** When C₁, C₃ and C₅ are charged to vₙ₄, v₁ and v₃ are forward biased. Power switches Q₁, Q₃ and Q₅ can be turned on at this moment to realize zero-voltage turn-on. Since i₀ still freewheels through D₁ and D₂, it can obtain vₙ₆ = 0, v_Lₕ₁ = v_Lₙ₁ = v_L₂₁ = v_L₃₁ = v_L₄₁ = 0. Thus, i₁p₁~i₃p₃ and i₁o₁ increase and i₁o₂ decreases.

**Step 8 [t₇~t₈]:** When C₂, C₄ and C₆ are charged to vₙ₅, v₁ and v₃ are forward biased and are discharged through C₂, C₄ and C₆ respectively at t₈, it can obtain vₙ₆ = 0. i₀ flows through D₁ and D₂ in this step and v_Lₕ₁ = v_Lₙ₁ = v_L₂₁ = v_L₃₁ = 0. Thus, i₁p₁~i₃p₃ and i₁o₁ increase and i₁o₂ decreases. This step is ended when i₁o₂ = 0. During this freewheeling interval, i₄r₁~i₄r₃ increase from -i₄o₂/(3n₁) to i₄o₁/(3n₁) and the current variation on L₄r₁~L₄r₃ is about i₄/3n₁. The duty loss in step 8 is obtained as (2) shows:

\[
d_{loss,8} = \frac{L_m I_m f_{sw}}{3n_1(V_in/3 - V_{th})}
\]

At time t₈ + T_sw, i₁D₁ is decreased to zero and this switching period is completed. Since the duty cycle of all switches equals 0.5, the voltage balance of C₁, C₂ and C₃ is well achieved with two balance capacitors C₁ and C₂. If Q₁, Q₃ and Q₅ are in the on-state and Q₂, Q₄ and Q₆ are in the off-state, it obtains v_C₁ = V₃ and v_C₂ = V₄. If V₁ > V₃, then C₁ charges or discharges C₄ through Q₁ and Q₅. When Q₁, Q₃ and Q₅ are in the off-state and Q₂, Q₄ and Q₆ are in the on-state, it obtains v_C₁ = V₄ and v_C₂ = V₃. If V₁ > V₄ or V₃ < V₄, then C₁ charges or discharges C₄ through Q₂ and Q₄. In a similar way, C₂ can be used to balance V₃ and V₄. Therefore, V₁~V₃ are all controlled at V_in/3.
Energies 2018, 11, 266

**Figure 5.** Operation steps of the power converter during one switching cycle (a) step 1 (b) step 2 (c) step 3 (d) step 4 (e) step 5 (f) step 6 (g) step 7 (h) step 8.

3. Circuit Characteristics

The asymmetric PWM scheme is adopted to drive \(Q_1-Q_6\). Based on the flux balance of primary inductors such as \(L_{r1}\) and \(L_{m1}\), \(L_{r2}\) and \(L_{m2}\) and \(L_{r3}\) and \(L_{m3}\), the DC capacitor voltages \(V_{C1}, V_{C2}, V_{C3}\) in a steady state can be obtained as \(V_{C1} = V_{C2} = V_{C3} = \frac{dV_{in}}{3}\), where \(d\) is the duty cycle of \(Q_1, Q_3\) and \(Q_5\). In the same way, the output voltage can be derived from the flux balance of output inductors \(L_{o1}\) and \(L_{o2}\) in a steady state.

\[
V_o = \frac{V_{in}}{3n_1}(1 - d) - \frac{I_o L_f s_{SW}}{3n_1^2} - V_f
\]

where \(V_f\) is the voltage drop on \(D_1\) and \(D_2\).
Since the average winding currents $i_{p1} \sim i_{p3}$ equal zero and $I_{L01} + I_{L02} = I_o$, the average inductor currents $I_{L01}$ and $I_{L02}$ are derived as $I_{L01} = (1 - d)I_o$ and $I_{L02} = dI_o$. The ripple currents on $L_{o1}$ and $L_{o2}$ are given in (4) and (5).

$$\Delta i_{L01} = \frac{V_o (1 - d + d_{loss,3}) T_{sw}}{L_{o1}} = \frac{(1 - d) V_o T_{sw} + \frac{V_o L_o I_o}{n_1 (1 - d)V_{in}}}{L_{o1}}$$

(4)

$$\Delta i_{L02} = \frac{V_o (d + d_{loss,3}) T_{sw}}{L_{o2}} = \frac{dV_o T_{sw} + \frac{V_o L_o I_o}{n_1 n_2 V_{in}}}{L_{o2}}$$

(5)

The asymmetric PWM is adopted to regulate load voltage. From Figure 4, the turn-on time of $D_1$ is related to the duty cycle of Q2 and the turn-on time of $D_2$ is related to the duty cycle of Q1. Therefore, the average diode currents $I_{D1}$ and $I_{D2}$ are expressed as $I_{D1} = (1 - d)I_o$ and $I_{D2} = dI_o$. The voltage stress of diodes $D_1$ and $D_2$ is related to the secondary winding voltage. The secondary winding voltage $V_{sw}$ is dependent on the input voltage $V_{in}$ and clamped voltage $V_{Ch1}$. Thus, the voltage stress of $D_1$ and $D_2$ can be given in (6) and (7).

$$v_{D1} = (1 - d)V_{in} / (3n_1)$$

(6)

$$v_{D2} = dV_{in} / (3n_1)$$

(7)

The conduction losses on rectifier diodes $D_1$ and $D_2$ are approximately equal to $I_o V_f$. If the transformer is constructed, then the magnetizing inductances $L_{m1} \sim L_{m3}$ are given. Thus, the ripple currents $i_{Lm1} \sim i_{Lm3}$ are obtained in (8).

$$\Delta i_{Lm1} = \Delta i_{Lm2} = \Delta i_{Lm3} \approx \frac{(V_{in}/3 - V_{Ch1})(d - d_{loss,3}) T_{sw}}{L_{m1}} = \frac{d(1 - d)V_{in} T_{sw}}{3L_{m1}} - \frac{I_o L_r}{3n_1 L_{m1}}$$

(8)

Three HB circuits are connected in a series on the high voltage side to reduce the voltage stress of active switches. Therefore, it is able to obtain the voltage stress of each active switch clamped at $V_{in}/3$. If the ripple currents on magnetizing inductors and output inductors are neglected, the root-mean-square (rms) currents $i_{Q1,rms} \sim i_{Q3,rms}$ can be derived as (9) and (10) show.

$$i_{Q1,rms} = i_{Q3,rms} = i_{Q5,rms} \approx \frac{(1 - d) I_o \sqrt{d}}{3n_1}$$

(9)

$$i_{Q2,rms} = i_{Q4,rms} = i_{Q6,rms} \approx \frac{d I_o \sqrt{1 - d}}{3n_1}$$

(10)

The conduction losses on $Q_1 \sim Q_6$ are approximately equal to $d(1 - d) I_o^2 R_{sw} / (3n_1^2)$, where $R_{sw}$ is turn-on resistance of $Q_1 \sim Q_6$. The positive peak currents of $i_{p1} \sim i_{p3}$ at time $t_1$ are given in (11).

$$i_{p1}(t_1) = i_{p2}(t_1) = i_{p3}(t_1) \approx i_{Lm1,max} + \frac{i_{Lm1,max}}{3n_1} \approx \frac{d(1 - d)V_{in} T_{sw}}{6L_{m1}} - \frac{I_o L_r}{6n_1 L_{m1}} + \frac{(1 - d) I_o}{3n_1} \frac{(1 - d)V_{in} T_{sw} + V_o L_o}{6n_1 L_{m1}}$$

(11)

Likewise, the negative peak currents of $i_{p1} \sim i_{p3}$ at time $t_5$ are given in (12).

$$i_{p1}(t_5) = i_{p2}(t_5) = i_{p3}(t_5) \approx i_{Lm1,max} + \frac{i_{Lm1,max}}{3n_1} \approx \frac{d(1 - d)V_{in} T_{sw}}{6L_{m1}} + \frac{I_o L_r}{6n_1 L_{m1}} - \frac{d I_o}{3n_1} - \frac{dV_o T_{sw} + V_o L_o}{6n_1 L_{m2}}$$

(12)

The minimum primary current to realize zero-voltage turn-on switching of $Q_1$, $Q_3$ and $Q_5$ is given in (13).

$$i_{p1}(t_5) \geq \frac{V_{in}}{3} \sqrt{\frac{2C_Q}{L_{r1}}}$$

(13)
Similarly, the minimum primary current to realize zero-voltage turn-on switching of $Q_2$, $Q_4$ and $Q_6$ is given in (14).

$$i_{p1}(t_1) \geq \frac{V_{in}}{3} \sqrt{\frac{2C_Q}{L_{r1}}}$$ (14)

where $C_Q = C_{Q1} = C_{Q2} = C_{Q3} = C_{Q4} = C_{Q5} = C_{Q6}$.

4. Experimental Results

A laboratory prototype with 1.44 kW rated power was constructed and tested in order to verify the feasibility of the studied converter to supply the auxiliary power in a light rail vehicle from the DC microgrid system. The experimental circuit diagram of the developed converter is provided in Figure 6. The TL431 voltage regulator and photocoupler PC817 are used to regulate load voltage. The PWM UCC2893 is used to achieve asymmetric pulse-width modulation (APWM) generation. Pulse transformers are adopted to achieve electrical isolation and gate drive. The specifications of the experimental prototype are given in Table 1. The power rating of the magnetic transformer is the same as the transformer in the three-level converter and the two-level full-bridge converter. The turns-ratio of the transformer in the developed converter is one-third of the turns-ratio in the two-level full-bridge converter. The magnetizing voltage and primary turns of the transformer in the studied converter are only one-third of the magnetizing voltage and primary turns in the two-level full-bridge converter. The PC40 EER-42 magnetic core is used with 15 primary turns and 8 secondary turns to build the isolated transformer. The experimental results of the proposed converter while it supplies 1.44 kW to the output load under a 760 V input are shown in Figures 7–12. Based on the test results, the measured waveforms agree well with the theoretical waveforms as given in Figure 4. Figure 7 shows the gate voltages of $Q_1$–$Q_6$ at 20% and 100% loads. It is clear that $Q_1$, $Q_3$ and $Q_5$ have the same PWM waveforms. In the same manner, $Q_2$, $Q_4$ and $Q_6$ have the same PWM waveforms. The input split voltages and balance capacitor voltages at a full load are shown in Figure 8. From the experimental results, the input split voltages and two flying voltages are balanced well. The primary side currents of three HB circuits are illustrated in Figure 9. It is observed that the three primary currents are balanced well. Figure 10 gives the measured waveforms of three DC block capacitor voltages. It can be observed that the three voltages are balanced and the capacitor voltages are related to the duty cycle of $Q_1$, $Q_3$ and $Q_5$. Figure 11 shows the experimental waveforms of the secondary side currents. The ripple currents on $L_{o1}$ and $L_{o2}$ partially cancel each other so that the resultant ripple current on the load side is reduced. Figure 12 shows the test results of $Q_1$ and $Q_2$ at 20% and 100% loads. Before the switch is turned on, the switch current is negative to discharge the output capacitor $C_Q$ to zero voltage. Thus, the mechanism of the zero-voltage turn-on switching is clearly achieved for both $Q_1$ and $Q_2$ from a 20% load. It can also be observed that all drain voltages of $Q_1$ and $Q_2$ are clamped at $V_{in}/3$. Since the PWM signals of the three HB circuits are identical and the input split voltages are also balanced, $Q_3$–$Q_6$ can also be turned on at zero-voltage from a 20% load. Figure 13 shows the circuit efficiency of the studied converter under different load cases. The measured maximum efficiency is about 93.6%. The main advantage of the studied converter is the lower voltage rating of power switches compared to the conventional three-level converter with eight MOSFETs and the two-level full-bridge converter with four IGBTs. Considering the turns-ratio of the transformer, the root-mean-square current of the studied converter is the same as the three-level and two-level converters. However, the lower conduction resistance of MOSFETs with lower voltage rating is used in the developed converter. Therefore, the total conduction losses (six MOSFETs) in the studied converter can be reduced compared to the three-level converter with eight MOSFETs and the full-bridge converter with four IGBTs.
Table 1. Prototype Specifications.

| Items                        | Symbol | Parameter       |
|------------------------------|--------|-----------------|
| Input voltage                | \(V_{in}\) | 760 V          |
| Output voltage               | \(V_o\) | 24 V            |
| Output current               | \(I_o\) | 60 A           |
| Switching frequency          | \(f_{sw}\) | 100 kHz        |
| Power switches               | Q1-Q6  | IRFP460         |
| Rectifier diodes             | D1, D2 | MBR40100T       |
| Split capacitors             | C1, C2, C3 | 330 \(\mu\)F/400 V |
| Flying capacitors            | Cb1, Cb2, Cb3 | 750 nF/630 V |
| Block capacitors             | \(C_{bp}\), \(C_{bp}\) | 2.2 \(\mu\)F/630 V |
| Turns ratio of \(T\)         | \(n_p\), \(n_p\), \(n_p\), \(n_p\) | 15:15:15:8 |
| Primary inductances          | \(L_{r1}, L_{r2}, L_{r3}\) | 30 \(\mu\)H |
| Magnetizing inductances      | \(L_{m1}, L_{m2}, L_{m3}\) | 0.8 mH |
| Output filter inductances    | \(L_{o1}, L_{o2}\) | 32 \(\mu\)H |
| Output filter capacitance    | \(C_o\) | 4400 \(\mu\)F/50 V |

Figure 6. The experimental circuit diagram of the developed converter.

Figure 7. PWM signals of Q1-Q6 under (a) 20% load; (b) full load \([v_{Q1}~v_{Q6}: 10 \text{ V/div}; \text{time: } 2 \mu\text{s}]\).
Figure 8. Measured input capacitor voltages under full load [$V_{C1} - V_{C2}$: 200 V/div; time: 2 µs].

Figure 9. Test results of the primary side currents under (a) 20% load [$i_{p1} - i_{p3}$: 2 A/div; time: 2 µs]; (b) full load [$i_{p1} - i_{p3}$: 5 A/div; time: 2 µs].

Figure 10. Test results of the primary capacitor voltages under (a) 20% load [$V_{C1} - V_{C3}$: 100 V/div; time: 2 µs]; (b) full load [$V_{C1} - V_{C3}$: 100 V/div; time: 2 µs].
Figure 11. Measured secondary side currents at (a) 20% load \([i_{D1}, i_{D2}, i_{L1}, i_{L2}, i_{L1} + i_{L2}]\): 10 A/div; time: 2 µs; (b) 100% load \([i_{D1}, i_{D2}, i_{L1} + i_{L2}]\): 50 A/div; \(i_{L1}, i_{L2}\): 20 A/div; time: 2 µs.

Figure 12. Measured switch voltages and current (a) \(Q_1\) at 20% load \([v_{Q1,g}, v_{Q1,d}]\): 10 V/div; \(v_{Q1,p}\): 200 V/div; \(i_{Q1}\): 2 A/div; time: 1 µs; (b) \(Q_1\) at full load \([v_{Q1,g}, v_{Q1,d}]\): 10 V/div; \(v_{Q1,p}\): 200 V/div; \(i_{Q1}\): 10 A/div; time: 1 µs; (c) \(Q_2\) at 20% load \([v_{Q2,g}, v_{Q2,d}]\): 10 V/div; \(v_{Q2,p}\): 200 V/div; \(i_{Q2}\): 2 A/div; time: 1 µs; (d) \(Q_2\) at full load \([v_{Q2,g}, v_{Q2,d}]\): 10 V/div; \(v_{Q2,p}\): 200 V/div; \(i_{Q2}\): 10 A/div; time: 1 µs.

Figure 13. Test results of circuit efficiency.

5. Conclusions

In this paper, a series-connected HB converter with a single transformer is proposed for light rail transit applications. The proposed circuit’s main benefits are the low voltage rating of power semiconductors, low switching losses, high circuit efficiency and balance split voltages compared to conventional converter topologies used in light rail vehicles. The asymmetric PWM scheme is used to control power switches, regulate output voltage and achieve the mechanism of the zero-voltage

Conflicts of Interest:
The author declares no potential conflict of interest.

Chaudhary, P.; Samanta, S.; Sesarma, P. Input-series-output-parallel-connected buck rectifiers for high-voltage applications. IEEE Trans. Ind. Electron. 2015, 62, 193–202.
turn-on switching. The circuit analysis, operation principle and design example of the proposed circuit are presented and discussed in detail. Finally, the feasibility of the studied circuit is verified by experimental results with a 1.44 kW laboratory prototype. The current harmonics on the input side is dependent on the switching frequency and the load current. The large current harmonics will result in unstable voltage on the input DC voltage bus. Therefore, in a future study, the interleaved DC/DC converter of the studied circuit will be developed to reduce the input current harmonics.

Acknowledgments: This research is supported by the Ministry of Science and Technology, Taiwan, under contract MOST 105-2221-E-224-043-MY2. The author would like to thank Wei-Po Liu for his help in the experiment. The author would also like to thank the anonymous reviewers for their valuable comments and suggestions to improve the quality of the paper.

Conflicts of Interest: The author declares no potential conflict of interest.

References

1. Srwanson, J.D.; Thornes, C. Light-rail transit systems. IEEE Veh. Technol. Mag. 2010, 5, 22–27. [CrossRef]
2. Chaudhary, P.; Samanta, S.; Sesarma, P. Input-series-output-parallel-connected buck rectifiers for high-voltage applications. IEEE Trans. Ind. Electron. 2015, 62, 193–202. [CrossRef]
3. Dujic, D.; Kieferndorf, F.; Canales, F.; Drofenik, U. Power electronic traction transformer technology. In Proceedings of the Power Electronics and Motion Control Conference (IPEMC Conference), Harbin, China, 2–5 June 2012; pp. 636–642.
4. Chaudhary, P.; Sesarma, P. Front-end buck rectifier with reduced filter size and single-loop control. IEEE Trans. Ind. Electron. 2013, 60, 4359–4368. [CrossRef]
5. Akagi, H. Classification, terminology, and application of the modular multilevel cascade converter (MMCC). IEEE Trans. Power Electron. 2011, 26, 3119–3130. [CrossRef]
6. Winkelkemper, M.; Korn, A.; Steimer, P. A modular direct converter for transformerless rail interties. In Proceedings of the 2010 IEEE International Symposium on Industrial Electronics (IEEE ISIE Conference), Bari, Italy, 4–7 July 2010; pp. 562–567.
7. Coppola, M.; Del Pizzo, A.; Iannuzzi, D. A power traction converter based on modular multilevel architecture integrated with energy storage devices. In Proceedings of the Electrical systems for Aircraft, Railway and Ship Propulsion Conference, Bologna, Italy, 16–18 October 2012; pp. 1–7.
8. Alexandre, C.; Drazen, D. Galvanically isolated modular converter. IET Proc. Power Electron. 2016, 9, 2318–2328.
9. Nakanishi, T.; Itoh, J.I. Evaluation of isolation three-phase AC/DC converter using modular multilevel converter topology. In Proceedings of the 2014 International Power Electronics Conference (IEEE IPEC Conference), Hiroshima, Japan, 18–21 May 2014; pp. 1095–1111.
10. Mcneill, N.; Yuan, X.; Anthony, P. High-efficiency multilevel converter technology for electric vehicle applications using super-junction MOSFETs. In Proceedings of the IET Hybrid and Electric Vehicle Conference, London, UK, 5–6 November 2014; pp. 1–7.
11. Madhusoodhanan, S.; Tripathi, A.; Patel, D.; Mainali, K.; Kadavelugu, A.; Hazra, S.; Bhattacharya, S.; Hatua, K. Solid-state transformer and MV grid tie applications enabled by 15kV SiC IGBTs and 10kV SiC MOSFETs based multilevel converters. IEEE Trans. Ind. Appl. 2015, 51, 3343–3360. [CrossRef]
12. Satarupa, B.; Akshay, K.R.; Srinivasan, D. Naturally clamped snubberless soft-switching bidirectional current-fed three-phase push–pull DC/DC converter for DC microgrid application. IEEE Trans. Ind. Appl. 2016, 52, 1577–1587.
13. Zhang, J.; Wu, H.; Qing, X.; Xing, Y. PWM plus secondary-side phase-shift controlled soft-switching full-bridge three-port converter for renewable power systems. IEEE Trans. Ind. Electron. 2015, 62, 7061–7072. [CrossRef]
14. Gu, L.; Jin, K.; Liu, C. Current-tripler-rectifier pulse width modulation ZVS three-phase full-bridge DC/DC converter with Y–A connected transformer. IET Proc. Power Electron. 2015, 8, 1111–1118. [CrossRef]
15. Hiroyuki, H.; Kurokawa, F. Modulation method of a full-bridge three-level LLC resonant converter for battery charger of electrical vehicles. IEEE Trans. Power Electron. 2017, 32, 2498–2507.
16. Korotkov, S.; Meleshin, V.; Nemchinov, A.; Fraidlin, S. Small-signal modeling of soft-switched asymmetrical half-bridge DC/DC converter. In Proceedings of the APEC’95, Conference Proceedings 1995, Tenth Annual Applied Power Electronics Conference and Exposition, 1995 (IEEE APEC), Dallas, TX, USA, 5–9 March 1995; Volume 2, pp. 707–711.

17. Tschirhart, D.J.; Jain, P.K. Design procedure for high-frequency operation of the modified series-resonant APWM converter to reduce size and circulating current. *IEEE Trans. Power Electron.* 2012, 27, 4181–4191. [CrossRef]

18. Kim, B.C.; Park, K.B.; Moon, G.W. Asymmetric PWM control scheme during hold-up time for LLC resonant converter. *IEEE Trans. Ind. Electron.* 2012, 59, 2992–2997. [CrossRef]

19. Haga, H.; Kurokawa, F. A novel modulation method of the full bridge three-level LLC resonant converter for battery charger of electrical vehicles. In Proceedings of the 2015 IEEE Energy Conversion Congress and Exposition (IEEE ECCE Conference), Montreal, QC, Canada, 20–24 September 2015; pp. 5498–5504.

20. Johnson, S.; Erickson, R. Steady-state analysis and design of the parallel resonant converter. *IEEE Trans. Power Electron.* 1988, 3, 93–104. [CrossRef]

21. Gu, Y.; Lu, Z.; Hang, L.; Qian, Z.; Huang, G. Three-level LLC series resonant DC/DC converter. *IEEE Trans. Power Electron.* 2005, 20, 781–789. [CrossRef]

22. Mishima, T.; Akamatsu, K.; Nakaoka, M. A high frequency-link secondary-side phase-shifted full-bridge soft-switching PWM DC-DC converter with ZCS active rectifier for EV battery charged. *IEEE Trans. Power Electron.* 2013, 28, 5758–5773. [CrossRef]

23. Lin, B.R.; Shiau, T.Y. Zero-voltage switching full-bridge DC/DC converter with parallel-connected output and without output inductor. *IET Proc. Power Electron.* 2013, 6, 505–515. [CrossRef]

24. Safaee, A.; Jain, P.; Bakhshai, A. A ZVS pulsewidth modulation full-bridge converter with a low rms current resonant auxiliary circuit. *IEEE Trans. Power Electron.* 2016, 31, 4031–4047. [CrossRef]

© 2018 by the author. Licensee MDPI, Basel, Switzerland. This article is an open access article distributed under the terms and conditions of the Creative Commons Attribution (CC BY) license (http://creativecommons.org/licenses/by/4.0/).