Single-stage ZETA-SEPIC-based multifunctional integrated converter for plug-in electric vehicles

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Abstract: A single-stage-based integrated power electronic converter has been proposed for plug-in electric vehicles (PEVs). The proposed converter achieves all modes of vehicle operation, i.e. plug-in charging, propulsion and regenerative braking modes with wide voltage conversion ratio \( M \) [\( M < 1 \) as well as \( M > 1 \)] in each mode. Therefore, a wide variation of battery voltage can be charged from the universal input voltage (90–260 V) and allowing more flexible control for capturing regenerative braking energy and dc-link voltage. The proposed converter has least components compared to those existing converters which have stepping up and stepping down capability in all modes. Moreover, a single switch operates in pulse width modulation in each mode of converter operation hence control system design becomes simpler and easy to implement. To correctly select the power stage switches, a loss analysis of the proposed converter has been investigated in ac/dc and dc/dc stages. Both simulation and experimental results are presented to validate the operation of the converter.

1 Introduction

The electric vehicles or plug-in electric vehicles (PEVs) are now a promising solution to curb the air pollution that uses pollution-free battery power to produce clean energy for the vehicle [1]. The PEVs are combination of on-board charger, battery, and the inverter-drive system [2–5]. In majority of PEVs, a bidirectional dc/dc converter is interfaced between the battery and dc-link of machine inverter [6–8] for power flow during propulsion and regenerative braking operation. Therefore, an individual ac/dc converter is used to charge the battery from the grid side. In this conventional structure, two separate power electronic converters are needed for two independent operations (charging and discharging of the battery). The bidirectional dc/dc converter in conventional structure can be integrated with the on-board charger, to have one power electronics interface for complete operation of PEVs. The overall block diagram of an integrated charger with the inverter drive is shown in Fig. 1a. This integration reduces the number of components because some of the switches and inductors are utilised both in ac/dc and dc/dc stages. Therefore, reduced number of switches and inductors lead to higher power density, compact size and lower cost. In this regard, this paper proposes, a new ZETA-SEPIC-based integrated converter for PEVs, as shown in Fig. 1b which has buck/boost capability in each mode of operation. In addition, buck/boost operation in each mode allows selection of wide range of the battery voltage, efficient control of dc-link voltage and capturing the regenerative braking with a wide variation of the motor speed.

A comparison of existing integrated converters and other competitive converters with respect to the proposed converter is described in the following paragraph.

An integrated converter in [9] utilises a number of semiconductor devices to achieve each mode; therefore, it may not be an efficiency optimised and cost-effective solution. In addition, the presence of a large number of devices, this converter requires a complex control strategy to turn on the switches. An integrated converter in [10] has only boost charging capability; thus, the selection of wide range of battery voltages is compromised. In [11], an integrated converter does not have buck/boost operation in any mode; thus, selection of the dc-link and battery voltage range is sacrificed. A three-level quasi two-stage converter in [12] with two inductors has buck/boost operation only in charging mode as a result, aforementioned advantages of buck/boost operation in each mode is sacrificed. In [13, 14], an SEPIC-based converter has been proposed for the battery charging using three inductors and at least one extra inductor is also required for propulsion and regenerative braking modes. Thus, the increase of magnetic components has a negative effect on weight, cost and volume of the charger. Authors in [15, 16] have proposed a CuK converter based on-board battery charger, which operates only in charging mode, does not include propulsion and regenerative braking modes. A single-stage converter in [17] operates only battery charging mode using four switches, eight diodes and two inductors. However, to achieve other modes of the vehicle, some more components will be employed. Therefore, this converter will utilise a large number of active and passive components, which will have an adverse effect...
on cost and compactness of the charger. Authors in [18] have proposed front-end power factor correction (PFC) converter for EV battery charger, which is a bridgeless type converter that uses four inductors and at least one additional inductor requires to achieve other modes of the vehicle. A single-stage-based inductive charger has been proposed in [19] that provides a wide range voltage for battery charging, but this converter uses a large number of passive components and semiconductor devices; therefore, the floor area of the charger will increase and less suitable for on-board application of PEVs.

Motivation of the work: The universal voltage range of single phase is around 90–260 V and a majority of commercially available battery voltage range are between 200 and 450 V [20–22]. Therefore, the buck/boost operation of converter is needed in plug-in charging mode for universal voltage supply. Moreover, in propulsion mode, usually, the battery voltage is stepped up to the dc-link voltage (inverter dc-link voltage) to propel the motor drive system. In a case of high state of charge (SOC) of the battery, the battery discharges by supplying its stored energy to the capacitor and voltage across capacitor gradually increases, which is shown in Fig. 2c. While inductor \( L_2 \) supplies energy to the output stage (capacitor and battery) shown in Fig. 2b and current through \( L_2 \) decreases linearly, as shown in Fig. 2d. The capacitor \( C_b \) is charged to \( V_{L_{b\ max}} \) through the body diode of \( S_1 \) in very short duration then after it retains this value forever in this mode. If the duty ratio of the converter is \( d_t \) then voltage-second balance either of inductor \( L_1 \) or \( L_2 \) for one switching period, \( T_s \), can be given as

\[
V_{\text{max}} |\sin(\omega t)| * d(t) = V_b * (1 - d(t)) * T_s
\]

(1)

From (1), the voltage conversion ratio \( M_t \) as

\[
M_t = \frac{V_b}{V_{\text{max}} |\sin(\omega t)|} = \frac{d(t)}{1 - d(t)}
\]

(2)

2.2 Propulsion mode

When this mode begins, battery started supplying power to the dc-link of the inverter and vehicle comes in running mode. During motion of the vehicle, the SOC of the battery continuously decreases. In this mode, switches \( S_1 \) and \( S_2 \) are kept in OFF state using mode selector logic, and switch \( S_3 \) is gated through PWM signal. When switch \( S_2 \) is turned OFF, inductor \( L_1 \) stores energy through the path \([v_d] - L_1 - S_2 - V_b - [v_g]\) and inductor \( L_2 \) stores energy through the path \([v_d] - L_4 - S_3 - C - L_2 - V_b - [v_g]\), as shown in Fig. 2a. When switch \( S_3 \) is turned OFF, inductor \( L_1 \) discharges by supplying its stored energy to the capacitor \( C \) and voltage across capacitor gradually increases, which is shown in Fig. 2d, and this capacitor is charged to the battery voltage \( V_b \). While inductor \( L_2 \) supplies energy to the output stage (capacitor and battery) shown in Fig. 2b and current through \( L_2 \) decreases linearly, as shown in Fig. 2d. The capacitor \( C_b \) is charged to \( V_{L_{b\ max}} \) through the body diode of \( S_1 \) in very short duration then after it retains this value forever in this mode. If the duty ratio of the converter is \( d_t \) then voltage-second balance either of inductor \( L_1 \) or \( L_2 \) for one switching period, \( T_s \), can be given as

\[
V_{\text{max}} |\sin(\omega t)| * d(t) = V_b * (1 - d(t)) * T_s
\]

(1)

\[
M_t = \frac{V_b}{V_{\text{max}} |\sin(\omega t)|} = \frac{d(t)}{1 - d(t)}
\]

(2)

2 Operation of the proposed converter

The proposed integrated converter operates in three modes: battery charging from the grid (plug-in charging), propulsion, and regenerative braking of charging. In the following section, operation of converter is discussed in detailed manner.

Fig. 2 Equivalent operating circuits and theoretical waveforms of plug-in charging mode

(a) Battery charging from the grid, when switch \( S_1 \) is turned ON and, (b) when switch \( S_3 \) is turned OFF, (c) Waveforms during one-line cycle, (d) One switching cycle
2.3 Regenerative braking mode

Operation of regenerative braking mode is similar to the grid mode operation, when switch \( S_3 \) is turned OFF, inductor \( L_3 \) transfers its stored energy to the capacitor \( C \) through the path \( V_{bS3L3}C \), as shown in Fig. 3b, and current through \( L_3 \) gradually decreases, which is shown in Fig. 4a. If the duty ratio of the converter is \( d_2 \) and applying voltage-second balance either in inductor \( L_3 \) or \( L_2 \) for one switching period then one can obtain:

\[
V_b \times d_2 \times T_s = V_{bS} \times (1 - d_2) \times T_s
\]  

The voltage conversion ratio \( M_2 \) from (3) can be expressed as

\[
M_2 = \frac{V_{bS}}{V_b} = \frac{d_2}{1 - d_2}
\]  

3 Component selection

3.1 Selection of semiconductor devices

Usually, power rating of each mode of PEVs is not same due to that the selection of switch ratings, i.e. voltage/current are decided by particular mode of operation. Therefore, voltage/current ratings of a particular switch will be selected as peak voltage/current developed either in switch or its body diode conduction. Table 1 shows peak voltage and current stresses of switch in all modes. From Table 1, switch ratings are selected as follows:

- Peak voltage rating of \( S_1 \) and \( S_2 \) is \( V_{max} \)
- Peak current rating of \( S_3 \) is \( V_{max} \)
- The peak current rating of \( S_3 \) is \( V_{max} \)

While voltage and current ratings of \( S_3 \) are chosen same as \( S_3 \) because body diode of \( S_3 \) conducts in propulsion mode and power rating of propulsion mode in PEVs usually highest among all modes.

3.2 Selection of inductors \( L_3 \) and \( L_2 \)

In the PEVs, the power level is usually high. At high-power-level continuous conduction mode (CCM) is preferred compared to discontinuous mode (DCM) since current stress in CCM are much smaller than DCM. During time interval \( [0, d_1(t)T_s] \), the inductor current ripple of \( L_3 \) can be written as

\[
\Delta i_L = \frac{V_{bS}}{L_3}d_1(t)T_s
\]  

Boundary condition for inductor current ripple happens when current ripple is the twice of the average input current. Thus,

\[
\Delta i_L = \frac{V_{bS}}{L_3}d_1(t)T_s = 2I_b
\]  

Fig. 3 Equivalent operating circuits during propulsion and regenerative braking modes

(a) Propulsion mode of operation, when switch \( S_3 \) is ON, (b) When switch \( S_3 \) is OFF, (c) Operation of regenerative braking, when switch \( S_3 \) is ON, (d) When switch \( S_3 \) is OFF

Fig. 4 Switching waveforms

(a) In propulsion mode, (b) In regenerative braking mode

...


Table 1 Voltage and current ratings of switches in all modes

<table>
<thead>
<tr>
<th>Modes</th>
<th>Voltage</th>
<th>Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>plug-in charging</td>
<td>( V_{b,\text{max}} + V_h )</td>
<td>( I_{b,\text{max}} + I_h )</td>
</tr>
<tr>
<td>propulsion</td>
<td>( V_h + V_h )</td>
<td>NC</td>
</tr>
<tr>
<td>regenerative braking</td>
<td>( V_h + V_h )</td>
<td>NC</td>
</tr>
</tbody>
</table>

Table 2 Simulation circuit parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>grid voltage ((V_g))</td>
<td>220/70.7 V</td>
</tr>
<tr>
<td>DC-link voltage ((V_{\text{dc}}))</td>
<td>400/100 V</td>
</tr>
<tr>
<td>line frequency ((f_L))</td>
<td>50/50 Hz</td>
</tr>
<tr>
<td>battery nominal voltage ((V_b))</td>
<td>300/60 V</td>
</tr>
<tr>
<td>nominal charging power ((P_b))</td>
<td>1 kW/230 W</td>
</tr>
<tr>
<td>(L/v_L)</td>
<td>2/2 mH</td>
</tr>
<tr>
<td>switching frequency ((f_s))</td>
<td>20/20 kHz</td>
</tr>
<tr>
<td>(C_{N} / C_{B} )</td>
<td>SET1 = SET2 = 550/10/2200 µF</td>
</tr>
</tbody>
</table>

Assuming unity power factor, the ratio of input voltage and input current can be expressed by an effective resistor, \( R_e \):

\[
R_e = \frac{V_g}{I_g} = \frac{V_g}{I_b} \tag{9}
\]

Thus, (CCM) condition for inductor current \((i_L)\) can be expressed as

\[
I_L > \max\left(\frac{R_e i(t)}{2f_s}\right) > \frac{V_g}{2} \frac{1}{P_b} \frac{V_b}{2f_s V_{g,\text{max}} + V_b} \tag{10}
\]

From (10), larger the minimum input power \(P_{g,\text{min}}\) and switching frequency \(f_s\) is, the easier the converter can enter in CCM. Similarly, the boundary condition for \(I_v\) can be derived. The CCM condition for inductor current \(i_L\) can be calculated as

\[
I_L > \left(\frac{R_e i(t)T_s}{2}\right) > \frac{R_{g,\text{min}}}{2} \frac{1}{V_{g,\text{min}} + V_b 2f_s} \tag{11}
\]

According (11), higher the switching frequency and smaller load resistance, \(R_{g,\text{min}}\) is, the easier the converter can enter in CCM mode. Using specification given in Table 2, inductors \(L_e\) and \(L_s\) are selected as 2 mH in the simulation as well as in hardware.

3.3 Selection of capacitor \(C\)

The coupling capacitor \(C\) is an important element in the prototype converter since its value greatly influence the quality of input current. Therefore, this capacitor will be designed under the following constraints: The resonant frequency of \(L_{in}\), \(L_s\) and \(C\) during CCM operation must be greater than line frequency \(f_{L}\) to avoid input current oscillations at every half line cycle and lower than switching frequency \(f_s\) to assure constant voltage in a switching period, i.e.

\[
f_L < f_s < f_s \tag{12}
\]

where

\[
f_s = \frac{1}{2\pi \sqrt{(L_s + L_c)C}} \tag{13}
\]

In this work, switching frequency \((f_s)\) is set as 20 kHz, and \(f_{L}\) is set as 1 kHz. The capacitor \(C\) is selected as 10 µF both in simulation and hardware. The voltage rating of this capacitor is selected based on battery voltage range because both in ZETA mode or SEPIC modes, the voltage across coupling capacitor follows battery voltage with ripple.

3.4 Selection of \(L_e\) and \(C_b\)

The design of input filter is essential for maintaining low harmonic distortion in grid current [23, 24]. The maximum value of filter capacitance is expressed as

\[
C_{b,\text{max}} = \frac{I_{g,\text{max}}\tan \theta}{2\pi f_L V_{g,\text{max}}} \tag{14}
\]

where \(V_{g,\text{max}}\) and \(I_{g,\text{max}}\) are peak input voltage, and \(\theta\) is considered below 5° for maintaining high input power factor. The capacitance \(C_{b,\text{max}}\) is calculated as 1.14 µF and selected value in the simulation and hardware is 1 µF.

The filter inductor to maintain low ripple is calculated as

\[
L_e = \frac{1}{4\pi f_L C_{b,\text{min}}} \tag{15}
\]

where \(f_L\) is cut off frequency, which is selected such that it should more than the grid frequency \(f_1 = 50\) Hz and less than switching frequency \(f_s = 20\) kHz. Therefore, it is chosen as 4 kHz. \(L_e\) from (15) is calculated as 1.58 mH, and the selected value in the simulation and hardware is 1.5 mH.

3.5 Selection of capacitor \(C_b\)

With regards to the parallel capacitor across the battery terminal, the switching frequency voltage ripple is negligible as this capacitor is typically very large. Twice of the line frequency voltage ripple is more critical since it directly affects the charging voltage. The low-frequency voltage ripple on the battery capacitor is given [25] as

\[
C_{b,\text{min}} > \frac{P_b}{4f_L \Delta V_{b} V_b} \tag{16}
\]

where \(f_{L}\) is the grid frequency and \(\Delta V_{b}\) is the capacitor ripple. To reduce the low-frequency voltage ripple, a large value of capacitor is preferred. However, this will make the electrolytic capacitor bulky. Therefore, the trade off between output voltage ripple and capacitor size must be taken into consideration for designing the converter. From above expression, the dc-link capacitance is selected as 2.2 mF for \(\Delta V_{b} < 5\%\) of the battery voltage.

4 Loss analysis

Due to high-power application of the proposed converter, it is necessary to estimate the semiconductor power losses of the converter in each mode to correctly select the power stage components. The conduction and switching losses for two semiconductor devices, i.e. 1200 V/100 A and 1700 V/72 A in ac/dc (plug-in charging) and dc/dc (propulsion) stages are analytically calculated using manufacturers date sheets. Moreover, the passive components that come in main current path are also presented in this analysis to accurately find efficiency of the converter in operation modes. The conduction loss for an MOSFET switch can be calculated as follows:

\[
L_e = \frac{1}{4\pi f_L C_{b,\text{min}}} \tag{15}
\]

where \(f_L\) is the grid frequency and \(\Delta V_{b}\) is the capacitor ripple. To reduce the low-frequency voltage ripple, a large value of capacitor is preferred. However, this will make the electrolytic capacitor bulky. Therefore, the trade off between output voltage ripple and capacitor size must be taken into consideration for designing the converter. From above expression, the dc-link capacitance is selected as 2.2 mF for \(\Delta V_{b} < 5\%\) of the battery voltage.
The total semiconductor losses are modelled as

\[ P_{\text{sw}} = \frac{1}{2}(V_{\text{s},\text{max}} - I_{s,\text{rms}}) \ast (t_i + t_f) \ast f_s \]  

(19)

where \( V_{\text{s},\text{max}} \) and \( I_{s,\text{rms}} \) are peak values of voltage and current and \( t_i \) and \( t_f \) are the rising and falling times of a switch, respectively. Power loss of diode is the sum of diode conduction loss (product of average current \( I_{d,\text{avg}} \) and knee voltage \( V_{d,\text{knee}} \)) and reverse recovery loss, which is modelled as

\[ P_D = V_F \ast I_D + P_{Q_R} \]  

(20)

where \( P_{Q_R} \) is the reverse recovery loss of a diode. The \( P_{Q_R} \) can be neglected for fast recovery diode and lower switching frequency. The total semiconductor losses are modelled as

\[ P_S = P_{\text{cond}} + P_{sw} + P_D \]  

(21)

The passive components that come in main current path are \( L_1, L_2, L_j \) and capacitor \( C \), then inductors copper losses are modelled as

\[ P_L = I_{L,\text{rms}} \ast r_f + I_{L,\text{rms}} \ast r_1 + I_{L,\text{rms}} \ast r_2 \]  

(22)

where \( r_f, r_1 \) and \( r_2 \) are series resistances of inductors \( L_1, L_2 \) and \( L_3 \), respectively. 

The ohmic loss in equivalent series resistance \( r_c \) of the capacitor \( C \) is expressed as

\[ P_c = I_{C,\text{rms}} \ast r_c \]  

(23)

Then total passive component losses \( P_p \) is written as

\[ P_p = P_{L_1} + P_{L_2} + P_{L_3} + P_c \]  

(24)

where \( P_{L_1}, P_{L_2} \) and \( P_{L_3} \) are the copper loss in \( L_1, L_2 \) and \( L_3 \), respectively.

With the help of these loss equations, Table 3 shows total losses of semiconductor devices \( (P_S) \) and passive components \( (P_p) \) in charging \((ac/dc \) stage\) and propulsion \((dc/dc \) stage\) modes, where \( P_{dc} \) and \( P_{dc,\text{dc}} \) denote power rating of charging and propulsion modes, respectively. It is noticed that from Table 3 in plug-in charging mode \((ac/dc)\), semiconductor losses with grid voltage of 240 V are around 1.5–2.2% of the rated powers for mentioned switching devices. While with 120 V grid supply, total losses in charging are found between 2 and 4% of the rated power. However, the total losses in \( dc/dc \) stage are around 2.5–3.1% of the rated powers \((P_{dc,\text{dc}} \leq 10 \text{kW})\).

The passive components losses \( P_p \) in \( ac/dc \) stage with 120 V grid supply is 2.5% of the rated power while with 240 V grid voltage is 1.5–2.78% of the rated powers. In \( dc/dc \) stage, \( P_p \) is around 1–2% of the rated powers. The percentage of passive component losses in \( dc/dc \) stage is slightly lower than the \( ac/dc \) stage, it is due to fact that inductor \( L_3 \) copper loss is not present in \( dc/dc \) stage. Moreover, the total losses (semiconductors and passive components) in \( ac/dc \) and \( dc/dc \) stages are between 5–5% and 2–4% of the rated powers with both devices, respectively. A loss breakdown of the converter system at full load for 1200 V/100 A switch is given in Table 4.

### Table 3: Semiconductors and passive components losses of the proposed converter in \( ac/dc \) and \( dc/dc \) stages with \( V_b = 300 \text{ V} \) and \( V_{\text{grid}} = 400 \text{ V} \)

<table>
<thead>
<tr>
<th>( P_{\text{ch}}, \text{kW} )</th>
<th>( P_{dc,\text{dc}}, \text{kW} )</th>
<th>( V_{\text{grid}}, \text{V} )</th>
<th>Switch rating</th>
<th>Total semiconductor losses ( (P_S) ), W</th>
<th>Total passive components losses ( (P_p) ), W</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.8</td>
<td>5</td>
<td>120</td>
<td>1200 V/100 A [( V_b = 2.55 \text{ V}, t_i = 65, t_f = 80 )]</td>
<td>59</td>
<td>44.88</td>
</tr>
<tr>
<td>3.2</td>
<td>5</td>
<td>240</td>
<td></td>
<td>68.93</td>
<td>48.69</td>
</tr>
<tr>
<td>1.8</td>
<td>10</td>
<td>120</td>
<td>1200 V/100 A [( V_b = 2.55 \text{ V}, t_i = 65, t_f = 80 )]</td>
<td>59</td>
<td>40.13</td>
</tr>
<tr>
<td>6.6</td>
<td>10</td>
<td>240</td>
<td></td>
<td>154.11</td>
<td>185.445</td>
</tr>
<tr>
<td>1.8</td>
<td>5</td>
<td>120</td>
<td>1700 V/72 A [( V_b = 3.6 \text{ V}, t_i = 20, t_f = 18 )]</td>
<td>33.44</td>
<td>44.88</td>
</tr>
<tr>
<td>3.2</td>
<td>5</td>
<td>240</td>
<td></td>
<td>46.18</td>
<td>48.69</td>
</tr>
<tr>
<td>1.8</td>
<td>5</td>
<td>120</td>
<td>1700 V/72 A [( V_b = 3.6 \text{ V}, t_i = 20, t_f = 18 )]</td>
<td>33.44</td>
<td>44.88</td>
</tr>
<tr>
<td>6.6</td>
<td>10</td>
<td>240</td>
<td></td>
<td>126.45</td>
<td>185.445</td>
</tr>
</tbody>
</table>

### Table 4: Loss breakdown of the system in \( ac/dc \) (plug-in charging) and \( dc/dc \) (propulsion) stages with \( 400 \text{ V} \) dc-link and \( 300 \text{ V} \) battery

<table>
<thead>
<tr>
<th>Stage</th>
<th>( P_{\text{ch}} ), kW</th>
<th>( P_{dc,\text{dc}} ), kW</th>
<th>( V_{\text{grid}} ), V</th>
<th>( P_{\text{kon}}, W )</th>
<th>( P_{\text{sw}}, W )</th>
<th>( P_{\text{D}}, W )</th>
<th>( P_{L_1}, W )</th>
<th>( P_{L_2}, W )</th>
<th>( P_{L_3}, W )</th>
<th>( P_{p}, W )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( ac/dc ) stage</td>
<td>[120 V, 1.8 kW]</td>
<td>33.55</td>
<td>18</td>
<td>7.45</td>
<td>13.5</td>
<td>24.75</td>
<td>5.63</td>
<td>1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( dc/dc ) stage</td>
<td>[240 V, 6.6 kW]</td>
<td>65</td>
<td>56.62</td>
<td>31.32</td>
<td>45.375</td>
<td>73.18</td>
<td>65.81</td>
<td>1.08</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( 5 \text{ kW} )</td>
<td>[5 kW]</td>
<td>39.75</td>
<td>29.49</td>
<td>31.7</td>
<td>—</td>
<td>17.18</td>
<td>28.98</td>
<td>1.094</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( 10 \text{ kW} )</td>
<td>[10 kW]</td>
<td>79.4</td>
<td>58.99</td>
<td>69</td>
<td>—</td>
<td>68.75</td>
<td>115.95</td>
<td>4.379</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

5 Comparative analysis of the proposed converter with single-stage converters

The conventional single-stage battery charger topologies, namely the boost PFC converter, inverting buck/boost PFC converter, SEPIC PFC converter, and CuK PFC converter is shown in Fig. 5. In order to have a fair comparison of the proposed converter with these single-stage converters, the \( dc/dc \) converter connected between the battery and \( dc/dc \) link in Fig. 5, is assumed to be a four-quadrant bidirectional converter. The boost PFC converter can only charge the battery when battery voltage is more than the peak grid voltage \((V_b > V_{\text{grid}})\). While inverting buck/boost and CuK converter both have negative output voltage with respect to the input. The same polarity between the input and the output has an advantage of solving electromagnetic interference/electromagnetic compatibility problems, and in designing filters easily, because the internal ground of a vehicle, the ground of the on-board battery charger (OBC) and the cathode of the battery can have the same
potential [17]. However, in CuK and SEPIC PFC-based converter topologies have one additional inductor compared to the proposed integrated converter. In addition to these conventional chargers, other existing integrated chargers are also included in comparative analysis. It can be noticed from Table 5, the proposed converter has least components compared to those converters which have buck/boost operation in each mode. However, the integrated converter [9] has lower voltage and current stresses across the devices (either input or output voltage and current) in each mode as well as buck/boost operation in each mode, but the efficiency of this converter sacrifices because three to four semiconductor devices come in the main current path. The integrated converter [10] has low stresses (voltage and current) like converter [9], but the major limitation of this converter is only to have a boost charging capability hence selection of the wide range of battery voltage is not possible. The efficiency plots of the proposed converter and integrated converters of [9, 10] using 1200 V/100 A device and 220 V grid voltage in each mode are shown in Fig. 6. The integrated converter [11] has bridge less nature in plug-in charging mode as well as low-voltage and current stresses in propulsion and regenerative modes and one–two devices come in current path; therefore, this converter will have higher efficiency than proposed converter in each mode. However, the major limitation of this converter is none of the modes have buck/boost operation.

From Table 5, the proposed converter has two additional passive components (one inductor and one intermediate capacitor C) compared to other existing integrated converters. The voltage across capacitor C follows battery voltage both in SEPIC and ZETA modes, the peak voltage across the capacitor $C = V_b + \Delta V_C$, where $\Delta V_C$ is voltage ripple of $V_C$ then voltage rating of this capacitor is selected according to slightly higher than the battery voltage. Usually, the capacitance value of $C$ is between 0.5 and 10 $\mu$F for 20 kHz switching frequency range [23, 26]; therefore, the size of this capacitor is smaller compared to battery terminal capacitor $C_b$ (designed based on twice of the grid frequency). Nevertheless, the additional increase of two passive components as well as higher losses in the proposed converter compared to existing converters [10, 11] results increase in a volume of the proposed converter. However, converter [9] has a number of semiconductor devices (switches are twice of the proposed converter) and switches with driver circuit occupy a considerable floor area as well as higher losses, result in the overall size of the proposed converter may be lower or comparable to this converter. Moreover, compared to conventional single-stage converters which have buck/boost operation in each mode, the proposed converter has lower volume due to lower passive and semiconductor components.

### 6 Control system design

The control structure during different modes of converter operation [10] is shown in Fig. 7. Each mode is implemented by mode selector logic, which requires external input such as torque ($T$),...
the mode selection is done manually.

During battery charging from the grid, the reference charging power is divided by instantaneous battery voltage, which is input to the outer proportional-integral (PI) controller. The output of the outer PI controller is a reference dc signal for inner controller $G_{ib}(z)$, and multiplied by a unit rectified sinusoidal wave to generate the final reference input to inner current PI controller. The inner current controller is used for the correction of power factor at the grid side.

The output controller $G_{ib}(z)$ is expressed as

$$G_{ib}(z) = K_p + \frac{K_i T_s}{z - 1}$$  \hspace{1cm} (25)

where $K_p$ is the proportional gain to adjust the control bandwidth and $K_i$ is the integral gain to achieve zero steady-state error.

The inner PI controller is given by the following equation as

$$G_{ic}(z) = K_{pc} + \frac{K_{ic} T_s}{z - 1}$$  \hspace{1cm} (26)

where $K_{pc}$ is the proportional gain and $K_{ic}$ is the integral gain. These two coefficients should be tuned such that the bandwidth of the controller is kept between one-sixth to one-tenth of the low-frequency battery ripple current and compactness of the charger. The inductance value of this filter is compromised between allowable low-frequency battery ripple current and high-frequency voltage ripples are present in $C$.

In propulsion and regenerative braking mode of operation, average current mode controller is used. Depending upon operating modes appropriate switches are turned ON. In propulsion mode, the dc-link voltage is regulated through two-loop controller. The outer PI controller $G_{ib}(z)$ regulates the dc-link voltage by generating reference battery current for the inner current controller $G_{ic}(z)$, which is provided by the average current mode controller.

In regenerative braking modes, the reference quantity is usually the grid side.

The simulation of the proposed converter has been verified in MATLAB/SIMULINK environment with two sets of circuit parameters, as shown in Table 2.

Set-I: In charging mode, the grid voltage is fixed at 220 V, charging power is 1 kW and the nominal battery voltage is 300 V with 20% of SOC. Fig. 8a shows simulated waveforms of grid voltage ($v_g$), grid current ($i_g$), battery voltage ($V_b$), battery current ($I_b$) and grid filter voltage and current. In this mode, the proposed converter operates as ZETA PFC converter and it is seen that the grid voltage ($v_g$) and grid current ($i_g$) are in the same phase with sinusoidal shape while measured total harmonic distortion (THD) of the grid current from fast Fourier transform (FFT) tool of the Matlab is 3.77% and calculated power factor (PF) is 0.99. The battery voltage and current are also shown in Fig. 8a. From this figure, it is noticed that the battery current has low-frequency pulsation, which is inherent in single-phase system. This pulsation of battery current $i_b$ depends on filter inductor connected in series with the battery. The inductance value of this filter is compromised between allowable low-frequency battery ripple current and compactness of the charger. The grid side filter capacitor voltage ($V_f$) and filter inductor current ($I_L$) are also shown in Fig. 8a. The voltage across coupling capacitor $C$ follows battery voltage and both switching and low-frequency voltage ripples are present in $C$.

The relevant waveforms in propulsion mode are shown in Fig. 8b with 400 V dc-link and 1 kW output load power. The control objective of this mode is to keep the dc-link constant for satisfactory operation of the inverter drive system. Therefore, dynamic performance of this mode is verified by step load variations from 1 to 2 kW at $t = 1.2$ s and from 2 to 1 kW at $t = 1.8$ s as shown in Fig. 8c. Despite the step change in loads at these instants, the dc-link voltage is well regulated at 400 V.

The simulation waveforms of regenerative mode are shown in Fig. 8d. To verify this mode, the dc-link voltage is varied manually using bridge rectifier and autotransformer, which is equivalent to the speed change of motor in real-time operation of the vehicle. The variations in the dc-link voltage are between 290 and 350 V and the battery is charged with constant current of 3.5 A. The corresponding variations in duty signal with the dc-link voltage are also shown in Fig. 8d. The duty signal is <0.5 whenever the dc-link voltage is more than the battery voltage, and >0.5 whenever, the dc-link voltage is lower than the battery voltage.

SET-2: In this Set of simulation, the proposed converter is verified at lower voltage and power levels to provide consistency between simulation and hardware results (Section 7.2). The parameters used in this Set are given in Table 2.

In plug-in charging mode, the peak grid voltage is fixed at 100 V, charging power is 230 W and the nominal battery voltage is 60 V with initial battery SOC of 20%. Fig. 9 shows simulated waveforms of this mode. It is seen in Fig. 9a, the grid voltage ($v_g$)
and grid current \(i_g\) are in the same phase with sinusoidal shape. The measured THD of the grid current and PF are around 3.53% and 0.999, respectively, which demonstrates the effectiveness of the converter operation at lower charging power, grid and battery voltages. The capacitor filter voltage \(V_{cf}\) and filter inductor current \(I_{lf}\) are also shown in Fig. 9a. In Fig. 9b, the voltage across capacitor \(C\), i.e. \(V_C\) follows the battery voltage and voltage ripple of this capacitor depends on the battery voltage ripple as well as the value of capacitor \(C\). However, size of capacitor \(C\) is compromised between allowable voltage ripple and overall charger size. Moreover, a simulation with 90 V of grid voltage (minimum universal voltage) is also performed, which is shown in Fig. 9c.

7.2 Experimental results

The parameters and specifications of prototype model of the proposed converter are mentioned in Table 6. A field-programmable gate array (FPGA)-based dSPACE (1104) controller is used for the development of the proposed converter. The isolation between controller and gate drivers of semiconductor switches is developed using the optocoupler TLP-250, which can be operated up to 25 kHz. The slow speed of switch body diode is replaced with the fast recovery diode MUR-1560 to reduce the reverse recovery loss and removes unwanted transients of the
Fig. 10 Simulation waveforms during propulsion and regenerative braking with 60 V battery and 100 V dc link
(a) Waveforms in propulsion mode with 400 W load, (b) Dynamic operation of propulsion mode with load variations, (c) Closed loop verification of regenerative braking mode by varying the dc-link voltage

Table 6 Experimental circuit parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Grid voltage ( (V_g) )</td>
<td>42.42/70.7 V</td>
</tr>
<tr>
<td>DC-link voltage ( (V_{hv}) )</td>
<td>60 V</td>
</tr>
<tr>
<td>Line frequency ( (f_L) )</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Battery voltage ( (V_{b}) )</td>
<td>36 V</td>
</tr>
<tr>
<td>Battery type</td>
<td>Exide power safe plus sealed lead acid</td>
</tr>
<tr>
<td>Capacity</td>
<td>26 Ah</td>
</tr>
<tr>
<td>( L_1/L_2 )</td>
<td>2 mH</td>
</tr>
<tr>
<td>( C_{hv}/C_{L}/C_b )</td>
<td>550/10/2200 ( \mu )F</td>
</tr>
<tr>
<td>Power switch</td>
<td>MOSFET IRFP-460</td>
</tr>
<tr>
<td>Current sensor</td>
<td>TELCON HTP-25</td>
</tr>
<tr>
<td>Voltage sensor</td>
<td>AD202JN</td>
</tr>
</tbody>
</table>

respectively. The dc-link voltage is also measured using voltage sensor. The grid current is measured through an ac current probe of FLUK power analyser. All waveforms are recorded in Agilent Technology, DSO1014A.

Fig. 12 shows waveforms during plug-in charging mode. As we know that in this mode, the battery is charged from the grid supply (60 V peak and 85 W charging power); therefore, converter must operate in PFC mode. In Fig. 12a, the grid current at CH1 has nearly sinusoidal shape as well as in phase with the grid voltage at CH2, which shows converter is operating near unity power factor condition. The battery current and voltage are shown at CH3 and CH4, respectively. A low-frequency oscillation in the battery current is seen at CH3 of Fig. 12a, if this low-frequency oscillation is high (without output filter), the reverse recovery loss of diode \( D_s \) will increase; therefore, the filter inductor at battery terminal suppresses the oscillation of battery current as well as reduces the reverse recovery loss of the diode \( D_s \) but addition of this filter may depend on manufactures and type of battery (capability to take low-frequency ripple of the battery current). In PFC mode of ZETA converter, it is mandatory to connect an LC filter after rectifier bridge diode to filter out discontinuous input current. The waveforms of filter capacitor voltage and filter inductor current are shown in Fig. 12b. Also in Zeta converter voltage across coupling capacitor is same as output voltage, i.e. battery voltage \( V_b \), which is shown in Fig. 12c. Fig. 12d shows the measured power factor is 0.99, and total harmonic distortion (THD) is 7.1%. The high-power factor operation of the proposed converter relives the burden on grid system and reduces the cost of electricity usages. Furthermore, with 100 V (peak) grid voltage and 85 W charging power (Fig. 12e), the measured PF and THD of the grid current are 0.99 and 7.2%, respectively, as shown in Fig. 13b. However, the measured input power \( P_g \) is 89.5 W at 85 W charging power (\( P_g \)); therefore, calculated efficiency is 94.9%. However, with 210 W charging power and 100 V (peak) grid voltage, the measured input power is 226 W (Fig. 13c) and calculated efficiency is 92.9%, which is lower compared to 85 W charging power, it is due to fact that conduction losses in MOSFET and inductors significantly increase at higher power. An experimental efficiency curve in charging mode with 60 and 100 V (peak) grid voltages is shown in Fig. 13d. The maximum efficiency for 60 V is found to be 94.76%. For 100 V, the maximum efficiency is 95.1% at 115 W but it may be more than this value at some other power level.

In propulsion mode, the battery voltage either stepped up or stepped down based on the battery SOC and dc-link voltage range. The objective of this mode is to keep the dc-link voltage constant irrespective of load or supply variations. The screenshot of oscilloscope in Fig. 14a, load has been shown in Fig. 14b. The dynamic performances of this mode is tested with step change in load variations from 125 to 215 W and corresponding change in battery and load current is shown in Fig. 14b at CH3 and CH4, respectively.

In regenerative mode, the battery is charged with regenerative braking energy of motor, the generated voltage is rectified with feedback diodes of the inverter, then depending upon speed of

Fig. 11 Gate driver circuit of switch with voltage protection
circuit. For voltage protection of the switches, an RC snubber circuit is used along with driver circuit shown in Fig. 11.

During plug-in charging mode, the grid voltage and output battery current are measured using voltage and current sensor.
motor the dc-link voltage varies. However, in this work, this mode is tested by rectifying ac voltage, which is applied through an auto transformer. The relevant waveforms of this mode are shown in Fig. 14. In Fig. 14c, the dc-link voltage has been fixed at 70 V, and the battery is charged with 2.78 A of current. In Fig. 14d, the dc-link voltage varies from $V_{hv_{\text{min}}}$ (45 V) to $V_{hv_{\text{max}}}$ (70 V) and during this variation of the dc-link voltage, the battery current is maintained at 2.78 A through the closed loop operation.

8 Conclusion

In this work, a ZETA-SEPIC-based single-stage power electronics interface has been proposed for PEVs. The proposed converter operates in three modes, i.e. plug-in charging (PFC mode), propulsion and regenerative modes. In PFC and regenerative braking modes, the proposed converter operates as ZETA converter, while in propulsion mode, it operates as SEPIC converter. It means the proposed converter has buck/boost operation in each mode of converter operation without voltage reversal which allows selection of a wide range of the battery voltage, efficient control of dc-link voltage and capturing the regenerative braking energy with wide variations of the motor speed. In comparison with existing single-stage converters, the proposed converter has the least component to those converters which have buck/boost operation in each mode. The functionality and performance of the proposed integrated converter have been verified through both in simulation and hardware. The performance of control algorithm is tested with the step load variations in propulsion mode and dc-link voltage variations in regenerative braking mode. An extensive loss analysis of the proposed converter is investigated to correctly select the power stage switches. The maximum theoretical efficiency of the converter in plug-in operation is 96%.
charging, propulsion and regenerative braking modes is found 95.9%, 97.1%, and 96.7%, respectively. While in hardware, the peak efficiency is found 94.76% at 85 W and 60 V peak grid voltage.

9 References

Fig. 14 Experimental waveforms during propulsion and regenerative modes
(a) Screenshot of propulsion mode (CH1: dc-link voltage [30 V/div], CH2: battery voltage [20 V/div], CH3: battery current [5 A/div]), CH4: dc-link current [3 A/div], time scale = 1 s/div.
(b) Dynamic performance of propulsion mode (CH1: dc-link voltage [30 V/div], CH2: battery voltage [20 V/div], CH3: battery current [5 A/div], CH4: dc-link current [3 A/div], time scale = 1 s/div).
(c) Screenshot of regenerative mode (CH1: dc-link voltage [30 V/div], CH2: battery voltage [20 V/div], CH3: battery current [5 A/div], CH4: battery voltage [20 V/div]), (d) Closed loop verification of regenerative mode (CH1: dc-link voltage [30 V/div], CH2: battery voltage [20 V/div], CH3: battery current [5 A/div], time scale = 1 s/div.)