Abstract—A novel transformerless high step-up buck–boost dc–dc converter is proposed in this paper. The output voltages of some sources such as fuel cell, photovoltaic are not regulated. Therefore voltage regulation is required to fix the DC-link voltage. Hence, a buck boost DC/DC converter is suitable to regulate the output voltage of these sources. The presented converter voltage gain is higher than that of the conventional boost, buck boost, CUK, SEPIC and ZETA converters and high voltage gain can be obtained with a suitable duty cycle. In this converter, only one power switch is utilized. The voltage stress across the power switch and diodes is low. Hence, the low on-state resistance of the power switch can be selected to decrease conduction loss of the switch and improve efficiency. The low voltage stress across the diodes allows the prevent the reverse-recovery current problem. The proposed converter can be operated in the continuous conduction mode (CCM) and the discontinuous conduction mode (DCM). The presented converter has simple structure; therefore the control of the proposed converter will be easy. The principle of operation and the mathematical analyses of the proposed converter are explained. The validity of the proposed converter is verified by the experimental results.

Index Terms— Buck boost dc–dc converter; voltage gain; duty cycle, voltage stress.

I. INTRODUCTION

In recent years, environmental troubles, such as climate change and global warming by increased emissions of carbon dioxide, are very important. With increasing attention to environmental problems, energy achieved from the fuel cell systems is focused on the low environmental effects and clean energy. Fuel cells are an effective alternative to replace fuels in emergency power systems and vehicles. Fuel cells can be used as clean energy by users with low emissions of carbon dioxide. Due to steady operation with renewable fuel supply and high effectiveness and efficiency, the fuel cell has been recognized increasingly as a suitable alternative source. There are some problems of this fuel such as high costs, but they have brilliant features such as high efficiency and small size. Due to this explanation, the fuel cell is appropriate as power supplies for telecom back-up facilities and hybrid electric vehicles. The output voltage of the fuel cell unit cell is low and is not steady and it cannot be directly connected to the load. For applications that need a steady DC voltage, buck-boost dc–dc converter is required [1-5]. However, the traditional buck-boost converter is not suitable for fuel cells sources. The traditional buck-boost converter efficiency is expected high; however, it is low and is

Nomenclature

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>( r_{DS} )</td>
<td>Switch on-state resistance</td>
</tr>
<tr>
<td>( R_{F1} ) and ( R_{F2} )</td>
<td>Diodes ( D_1 ) and ( D_2 ) forward resistances</td>
</tr>
<tr>
<td>( V_{F1} ) and ( V_{F2} )</td>
<td>Diodes ( D_1 ) and ( D_2 ) threshold voltages</td>
</tr>
<tr>
<td>( R_{L1} ) and ( R_{L2} )</td>
<td>ESR of inductors ( L_1 ) and ( L_2 )</td>
</tr>
<tr>
<td>( r_{C1} ), ( r_{C2} ) and ( r_{C3} )</td>
<td>ESR of capacitors ( C_1 ), ( C_2 ) and ( C_3 )</td>
</tr>
<tr>
<td>( \Delta V_{C1} )</td>
<td>Voltage ripple of the capacitor ( C_1 )</td>
</tr>
<tr>
<td>( \Delta V_{C1,ESR} )</td>
<td>Voltage ripple on capacitor ( C_1 ) created from the current that flows through the equivalent series resistance</td>
</tr>
<tr>
<td>( \tan \delta_1 )</td>
<td>Voltage ripple of the charging and discharging Dissipation factor of capacitor ( C_1 )</td>
</tr>
<tr>
<td>( \Delta V_{C2} )</td>
<td>Voltage ripple on capacitors ( C_2 ) created from the current that flows through the equivalent series resistance</td>
</tr>
<tr>
<td>( \Delta V_{C2,ESR} )</td>
<td>Voltage ripple of the capacitors ( C_2 ) created from the charging and discharging Dissipation factor of capacitors ( C_2 )</td>
</tr>
<tr>
<td>( \tan \delta_2 )</td>
<td>Voltage ripple of the capacitors ( C_3 )</td>
</tr>
<tr>
<td>( \Delta V_{C3} )</td>
<td>Voltage ripple on capacitors ( C_3 ) created from the current that flows through the equivalent series resistance</td>
</tr>
<tr>
<td>( \Delta V_{C3,ESR} )</td>
<td>Voltage ripple of the capacitors ( C_3 ) created from the charging and discharging Dissipation factor of capacitor ( C_3 )</td>
</tr>
<tr>
<td>( \tan \delta_{C3} )</td>
<td></td>
</tr>
</tbody>
</table>

Although the flyback converter can obtain the high step-up voltage gain, the power switches suffer a voltage spike across the switches and the converter efficiency is not high because of the reverse-recovery problems and leakage inductor [9-11]. In [12-13] high voltage gain dc–dc converters with a coupled-inductor is proposed. The leakage inductance of the coupled inductor is so important that it cause high voltage spikes and adds the voltage stress. In [14] the switched capacitor method is used to obtain high step up voltage gain. In [15] a high step-up bidirectional dc–dc converter with low voltage stress on the
switch is proposed. In [16-17] transformer less high step-up dc-dc converters are proposed. In [18-19] a transformer less buck-boost dc-dc converter is proposed. The voltage gain for this converter is twice as large as the conventional buck boost converter. In [20] a transformer less interleaved high step-down converter is proposed, but, in the presented converter, two power switches have been utilized and the capacitors of the converter are suddenly charged. In [21] a buck-boost converter based on KY converter is proposed. In this converter, two main switches are used and the voltage gain of the presented converter is 2D. In [22] a multi-output buck boost DC/DC converter is proposed. This converter has several output voltages, but, in the presented converter, many power switches have been used. In [23] a two-stage inverting buck-boost converter is proposed. The presented converter is constructed of two parallel conventional buck-boost converters. In [24] a two-stage buck-boost converter for power factor correction is presented and this converter does not require additional power switch. In [25] a multi-input DC-DC converter is proposed to connect two power sources with a DC bus or load. The presented converter has high efficiency due to obtaining turn-on zero voltage switching (ZVS) of power switches. However, it requires a bidirectional port. In [26] a multi-port bidirectional DC-DC converter is proposed for DC micro grid. An integrated three phase transformer is utilized to enhance the voltage level and isolate low voltage side and high voltage side; therefore, the number of power switches is high hence, two input sources require to share a common ground. In [27] a non-inverting buck-boost converter for fuel-cell system using three power switches is proposed with a voltage gain equal to the proposed converter in this paper. In [28] the high step-up converters, based on coupled inductors to extend voltage gains by adjusting the turns ratios of coupled inductors, are presented. However, the use of coupled inductors would lead to some problems in leakage inductance and pulsating input current. In [29] the transformer less converters are presented. Although the converters do not use coupled inductors, there are too many components, which result in complexity. In [30] an interleaved high step-up converter is presented. The presented converter combines an interleaved boost converter, a voltage multiplier and a cascade three-level boost converter. In this converter, the interleaved boost converter is used to reduce the input current ripple, and the voltage multiplier and the three-level boost converter are used not only to upgrade the voltage gain but also to make the voltage stresses of semiconductors lower than the output voltage. In [31] the interleaved converters with coupled inductors are presented. However, additional active clamp and passive clamp circuits are adopted to reduce the spike voltages across the active switches, and this would result in complicated structure. In [32] a transformerless high voltage gain dc-dc converter, is proposed, the presented converter is consisted of two inductors and two active switches which share the same operation signal and the topology of the converter is very simple. In this paper, a novel single switch transformerless buck boost dc-dc converter with high step-up voltage gain and low voltage stress on the power switch is proposed. The voltage transfer gain of the proposed converter is higher than the classic buck-boost converter, SEPIC, CUK and ZETA converters. The structure of the proposed converter is simple, hence the control of the converter will be easy. In this topology, only one power switch is used which makes the control scheme simple as well as reducing the switching power losses. Moreover, CCM operational region is broadened in the proposed converter in comparison with the converter the voltage stress across the active power switch is reduced which enables the use of MOSFETs with a lower voltage rating and low $R_{DS\text{ON}}$. Thus, the conversion efficiency can be improved due to the reduced conduction and switching losses. Similarly, the reduced voltage stress across all the diodes in the circuit allows the use of Schottky rectifiers for alleviating the reverse-recovery current problem, leading to a further reduction in the switching losses. The proposed buck-boost converter is utilized in many applications like fuel-cell systems, car electronic devices, LED drivers and gadgets such as mobile phones and notebooks. In this paper, the mathematical analyses of the proposed converter are explained. Besides, to verify the feasibility of the converter, experimental results are provided.

II. OPERATING PRINCIPLE OF THE PROPOSED CONVERTER

The proposed converter is shown in Fig. 1(a). This converter consists one power switch $S$, two diodes $D_1$ and $D_2$, two inductors $L_1$ and $L_2$, three capacitors $C_1$, $C_2$ and $C_3$ and load $R$. In this paper, duty cycle is denoted by $D$, the switching time frequency are denoted by $F_S$ and $T_S$, respectively, the input voltage is signified by $V_i$, the output voltage is represented by $V_o$, the voltages across the capacitors $C_1$, $C_2$ and $C_3$ are indicated by $V_{C1}$, $V_{C2}$ and $V_{C3}$, respectively, the voltage across the inductors $L_1$ and $L_2$ are signified by $V_{L1}$ and $V_{L2}$ respectively and the output current is denoted by $I_o$.

For simplicity of the analysis of the operating principles, the following assumptions are considered:

1) The capacitors of the presented converter are large enough, hence the voltage across capacitors are assumed to be constant.

2) The main switch of the proposed converter is treated as ideal and the parasitic capacitor of the main switch is neglected.

The presented converter can be operated in both the continuous conduction mode (CCM) and the discontinuous conduction mode (DCM). The continuous conduction mode can be divided into two operation modes. The analysis of the presented converter in one switching period under continuous condition mode (CCM) is explained in detail as follows:

1) First mode $[0 \leq t \leq DT_1]$: During this time interval as shown in Fig. 1(b), the switch $S$ is turned on and the diodes $D_1$ and $D_2$ are turned off. The inductors $L_1$ and $L_2$ are magnetized linearly and energy storage in capacitor $C_3$ is discharged to capacitor $C_1$ and the capacitor $C_2$ is discharged. Thus, the relevant equations are found to be:

$$V_{L1} = V_i \quad \text{(1)}$$

$$V_{L2} = V_i + V_{C3} - V_{C1} \quad \text{(2)}$$
Second mode [DT_1 \leq t \leq T_2] : The equivalent circuit is shown in Fig. 1(c). During this time interval, the switch S is turned off and the diodes D_1 and D_2 are turned on. The inductors L_1 and L_2 are demagnetized linearly. The capacitor C_2 is charged by the inductor L_1. The capacitor C_3 is charged by the inductor L_2 and the capacitor C_1 is discharged. The associated equations are found to be:

\[ V_{L1} = -V_{C3} \quad (3) \]
\[ V_{L2} = -V_{C1} = -V_{C2} \quad (4) \]

By applying volt-sec balance principle on the inductor L_i and using (1) and (3), we have:

\[ \int_0^{T_2} V_i \, dt + \int_0^{T_2} (-V_{C_i}) \, dt = 0 \quad (5) \]

By simplify (5), the voltage across capacitor C_3 (V_{C3}) can be achieved as follows:

\[ V_{C3} = \frac{DV_i}{1-D} \quad (6) \]

By applying volt-sec balance principle on the inductor L_1 and using (2) and (4), we have:

\[ \int_0^{T_2} (V_i + V_{C3} - V_{C1}) \, dt + \int_0^{T_2} (-V_{C2}) \, dt = 0 \quad (7) \]

By simplify (7), the voltage across capacitor C_2 (V_{C2}) can be obtained as follows:

\[ V_{C2} = \frac{DV_i}{1-D} \quad (8) \]

Using (6) and (8), the voltage transfer gain (M_{CCM}) can be achieved as follows:

\[ M_{CCM} = \frac{V_o}{V_i} = \frac{V_{C2} + V_{C3}}{V_i} = \frac{2D}{1-D} \quad (9) \]

According to (9), the voltage gain of the proposed converter is higher than that of the conventional boost, buck boost, CUK, SEPIC and ZETA converters and is twice as large as the voltage gain of the conventional buck boost converter.

Fig. 2. Shows some typical key waveforms of the proposed converter in continuous conduction mode (CCM). The voltage gain curves for the proposed converter and conventional boost and buck boost converter are shown in Fig. 3. It is seen that the voltage gain of the proposed converter is higher than the other converters.
B. Calculation of the currents

The current that flows through the capacitor $C_2$ during switch on period ($I_{C2,on}$) can be achieved as follows:

$$I_{C2,on} = -I_o$$

(10)

The current that flows through the capacitor $C_1$ during switch on period ($I_{C1,on}$) can be earned as follows:

$$I_{C1,on} = I_{L2}$$

(11)

The current that flows through the capacitor $C_1$ during switch off period ($I_{C1,off}$) can be achieved as follows:

$$I_{C1,off} = I_{L2} - I_{C2,off} - I_o$$

(12)

Where, $I_{C2,off}$ is the current that flows through the capacitor $C_2$ during switch off period.

By applying current-see balance principle on capacitor $C_1$, the following equation can be earned as follows:

$$\int_0^{D_{t}} I_{C1,on} dt + \int_{D_{t}}^{T} I_{C1,off} dt = 0$$

(13)

By substituting (10), (11) and (12) into (13), the currents that flow through the capacitor $C_1$ and the inductor $L_2$ ($I_{L2}$) can be obtained as follows:

$$I_{C1,on} = I_{L2} = I_o$$

(14)

From Fig. 1(b), the current that flows through the capacitor $C_2$ during switch on cycle ($I_{C3,on}$) can be achieved as follows:

$$I_{C3,on} = I_{C2,on} = -I_o$$

(15)

According to Fig. 1(c), the current that flows through the inductor $L_1$ ($I_{L1}$) can be earned as follows:

$$I_{L1} = (I_{C3} + I_{L2} - I_{C1} - I_{C2,off}) = \frac{1+D}{1-D} I_o$$

(16)

The average of input current ($I_i$) can be achieved as follows:

$$I_i = \frac{1}{T_s} \int_0^{DT_s} (I_{L1} - I_{C1,off}) dt = \frac{2D}{1-D} I_o$$

(17)

According to Fig. 1(b), the current that flows through the switch $S$ ($I_S$) can be obtained as follows:

$$I_S = I_{L1} + I_{C1,on} = \frac{2}{1-D} I_o$$

(18)

The current that flows through the diodes $D_1$ and $D_2$ ($I_{D1}$ and $I_{D2}$) can be achieved as follows:

$$I_{D1} = I_{L2} + I_{C1,off} = \frac{1}{1-D} I_o$$

(19)

$$I_{D2} = I_{L1} - I_{C1,off} = \frac{1}{1-D} I_o$$

(20)

C. Discontinuous conduction mode

The operation modes in discontinuous conduction mode (DCM) can be divided into three modes. The first mode in (DCM) is the same as the first mode in (CCM). In the second mode, the diodes currents are decreasing and in the third mode the diodes $D_1$ and $D_2$ currents will be zero and the diodes and switch will turn off. The equivalent circuit and the typical waveform in third mode are shown in Figs. 4 and 5. In this mode, the inductors $L_1$ and $L_2$ currents will be constant; therefore, the voltage of the inductors $L_1$ and $L_2$ will be zero. According to (19) and (20), the sum of the diodes $D_1$ and $D_2$ currents can be obtained as follows:

$$I_{D1} + I_{D2} = I_{L1} + I_{L2}$$

(21)

Using (19) and (20), the average currents of diodes $D_1$ and $D_2$ ($I_{D1,av}$ and $I_{D2,av}$) can be achieved as follows:

$$I_{D1,av} = I_{D2,av} = \frac{V_o}{R}$$

(22)

According to Fig. 5, the sum of the average current of diodes $D_1$ and $D_2$ during switching off period can be obtained as follows:
\[ I_{D_{1,av}} + I_{D_{2,av}} = \frac{1}{2} \times D_{m2} \times I_{D-PK} \]  \hspace{1cm} (23)

Where, \( D_{m2} \) is duty cycle in first mode at discontinuous conduction mode (DCM) and \( I_{D-PK} \) is sum of the peak currents of diodes \( D_1 \) and \( D_2 \) \( \left(I_{D1-pk} \text{ and } I_{D2-pk}\right) \).

\[ I_{D-pk} = I_{D1-pk} + I_{D2-pk} = \frac{V_I DT_s}{L_s} \]  \hspace{1cm} (24)

Where,

\[ \frac{1}{L_c} = \frac{1}{L_1} + \frac{1}{L_2} \]  \hspace{1cm} (25)

By applying volt-sec balance principle on inductors \( L_1 \) and \( L_2 \), duty cycle in first mode at discontinuous conduction mode \( (D_{m1}) \) can be achieved as follows:

\[ D_{m2} = \frac{2DV_I}{V_o} \]  \hspace{1cm} (26)

Using (22)-(26), the voltage transfer gain in DCM \( (M_{DCM}) \) can be earned as follows:

\[ M_{DCM} = \frac{D}{\sqrt{\tau_L}} \]  \hspace{1cm} (27)

Where, the parameter \( \tau_L \) is defined as follows:

\[ \tau_L = \frac{2L_c}{RT_s} \]  \hspace{1cm} (28)

**D. Boundary condition mode**

In this mode, the voltage transfer gain of the CCM is equal to the voltage transfer gain of the DCM. Using (9) and (27), the boundary normalized inductor time constant \( (\tau_b) \) can be achieved as follows:

\[ \tau_b = \frac{(1-D)^2}{4} \]  \hspace{1cm} (29)

The relationship between the boundary normalized inductor time constant \( (\tau_b) \) with different duty cycle is shown in Fig. 6. If \( \tau_L \) is larger than \( \tau_b \), the proposed converter operates in continuous condition mode (CCM).

**E. Efficiency analysis**

For efficiency analysis of the proposed converter, parasitic resistance, the voltage ripple across the capacitors and the inductors is ignored.

The rms current of power switch \( S \) \( (I_{S,rms}) \) can be achieved as follows:

\[ I_{S,rms} = \sqrt{\frac{1}{T} \int_0^T \left(I_{L1} + I_{C1,av}\right)^2 dt} = \frac{2\sqrt{D}}{(1-D)} I_o \]  \hspace{1cm} (30)
The power loss of switch $S$ ($P_{iDS}$) can be calculated as follows:

$$P_{iDS} = r_{iDS} I_{s,ms}^2 = \frac{4D}{(1-D)^2} I_o^2$$  \hspace{1cm} (31)

The rms current of diodes $D_1$ and $D_2$ ($I_{D1,ms}$ and $I_{D2,ms}$) can be obtained as follows:

$$I_{D1,ms} = I_{D2,ms} = \frac{1}{\sqrt{T_s}} \int_0^T \frac{I_o}{1-D} dt = \frac{1}{\sqrt{1-D}} I_o$$  \hspace{1cm} (32)

Forward resistance loss of diode $D_1$ ($P_{RF1}$) can be obtained as follows:

$$P_{RF1} = R_F I_{D1,ms}^2 = R_F I_1 \frac{1}{1-D} I_o^2$$  \hspace{1cm} (33)

Forward voltage loss of diode $D_1$ ($P_{VF1}$) can be earned as follows:

$$P_{VF1} = V_F I_{D1,av} = V_F I_o$$  \hspace{1cm} (34)

Forward resistance loss of diode $D_2$ ($P_{RF2}$) can be achieved as follows:

$$P_{RF2} = R_F I_{D2,ms}^2 = R_F \frac{1}{1-D} I_o^2$$  \hspace{1cm} (35)

Forward voltage loss of diode $D_2$ ($P_{VF2}$) can be obtained as follows:

$$P_{VF2} = V_F I_{D2,av} = V_F I_o$$  \hspace{1cm} (36)

The rms current of capacitors $C_1$ and $C_2$ ($I_{C1,ms}$ and $I_{C2,ms}$) can be obtained as follows:

$$I_{C1,ms} = I_{C2,ms} = \frac{1}{\sqrt{T_s}} \int_0^T \left[ \left( \frac{D}{1-D} I_o \right)^2 dt \right]$$  \hspace{1cm} (37)

The power loss of capacitor $C_1$ ($P_{RC1}$) due to ESR, can be achieved as follows:

$$P_{RC1} = r_{C1} I_{C1,ms}^2 = r_{C1} \frac{D}{1-D} I_o^2$$  \hspace{1cm} (38)

The power loss of capacitor $C_2$ ($P_{RC2}$) due to ESR, can be calculated as follows:

$$P_{RC2} = r_{C2} I_{C2,ms}^2 = r_{C2} \frac{D}{1-D} I_o^2$$  \hspace{1cm} (39)

The power loss of capacitors $C_3$ ($P_{RC3}$) can be expressed as follows:

$$P_{RC3} = r_{C3} I_{C3,ms}^2 = r_{C3} \frac{4D}{1-D} I_o^2$$  \hspace{1cm} (41)

The rms current of inductor $L_1$ ($I_{L1,ms}$) can be obtained as follows:

$$I_{L1,ms} = \frac{1+D}{1-D} I_o$$  \hspace{1cm} (42)

The conduction loss of inductor $L_1$ ($P_{iL1}$) can be obtained as follows:

$$P_{iL1} = R_L I_{L1,ms}^2 = R_L I_o^2$$  \hspace{1cm} (44)

The conduction loss of inductor $L_2$ ($P_{iL2}$) can be expressed as follows:

$$P_{iL2} = R_L I_{L2,ms}^2 = R_L I_o^2$$  \hspace{1cm} (45)

Finally, the total power loss and efficiency of the proposed converter ($P_{loss}$ and $\eta$) can be obtained as follows:

$$P_{loss} = P_{iDS} + \sum_{u=1}^2 (P_{RFu})_{Duo} + \sum_{u=1}^2 (P_{VFu})_{Du} + \sum_{u=1}^2 P_{RCu} + P_{iL1} + P_{iL2}$$  \hspace{1cm} (46)

$$\eta = \frac{P_o}{P_o + P_{loss}} = \frac{1}{1 + \frac{P_{loss}}{P_o}}$$  \hspace{1cm} (47)

The voltage stress of the diodes $D_1$ and $D_2$ ($V_{D1}$ and $V_{D2}$) can be achieved as follows:

$$V_{D1} = V_{D2} = \frac{V_i}{1-D}$$  \hspace{1cm} (48)

The voltage stress of the power switch $S$ ($V_S$) can be obtained as follows:
The relationship between the normalized voltage stress across power switch of the proposed converter and other converters is depicted in Fig. 7. According to Fig. 7, the normalized voltage stress of the switch in the proposed converter is lesser than that in other converters.

E. Calculation of the voltage ripple of the capacitors

From Fig. 8, \( \Delta V_{C1} \) can be achieved as follows:

\[
\Delta V_{C1} = \Delta V_{C1,ESR} + \Delta V_{C1,cap}
\]  
(50)

Also, \( \Delta V_{C1,ESR} \) can be calculated as follows:

\[
\Delta V_{C1,ESR} = ESR_{C1} \Delta I_{C1} = ESR_{C1} (I_{C1,off} - I_{C1,on}) = \frac{ESR_{C1} I_0}{1 - D}
\]  
(51)

Where,

\[
ESR_{C1} = \frac{\tan \delta_{C1}}{2 \pi f_s}
\]  
(52)

\( \Delta V_{C1,cap} \) can be achieved as follows:

\[
\Delta V_{C1,cap} = \frac{I_{C1,off} DT_s}{C_1} = \frac{DT_s V_i}{RC_1}
\]  
(53)

Also, \( \Delta V_{C2,ESR} \) can be expressed as follows:

\[
\Delta V_{C2,ESR} = ESR_{C2} \Delta I_{C2} = ESR_{C2} (I_{C2,off} - I_{C2,on}) = \frac{ESR_{C2} I_0}{1 - D}
\]  
(55)

Where,

\[
ESR_{C2} = \frac{\tan \delta_{C2}}{2 \pi f_s}
\]  
(56)

\( \Delta V_{C2,cap} \) can be achieved as follows:

\[
\Delta V_{C2,cap} = \frac{I_{C2,off} (1 - D) T_s}{C_2} = \frac{DT_s V_i}{RC_2}
\]  
(57)

From Fig. 10, \( \Delta V_{C3} \) can be obtained as follows:

\[
\Delta V_{C3} = \frac{\Delta V_{C3,cap}}{V_{C2}}
\]  
(58)

Also, \( \Delta V_{C3,ESR} \) can be expressed as follows:

\[
\Delta V_{C3,ESR} = ESR_{C3} \Delta I_{C3} = \frac{2 ESR_{C3} I_0}{1 - D}
\]  
(59)

Where,

\[
ESR_{C3} = \frac{\tan \delta_{C3}}{2 \pi f_s}
\]  
(60)

\( \Delta V_{C3,cap} \) can be achieved as follows:

\[
\Delta V_{C3,cap} = \frac{I_{C3,off} (1 - D) T_s}{C_3} = \frac{2 DT_s V_i}{RC_3}
\]  
(61)

From Fig. 9, \( \Delta V_{C2} \) can be obtained as follows:

\[
\Delta V_{C2} = \Delta V_{C2,ESR} + \Delta V_{C2,cap}
\]  
(54)
The component normalized voltage and rms current stresses for the presented converter in CCM are shown in Table 1.

**TABLE I**

<table>
<thead>
<tr>
<th>Circuit parameter</th>
<th>Normalized voltage</th>
<th>Normalized rms current</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switch $S$</td>
<td>$\frac{M_{CCM}+1}{2M_{CCM}}$</td>
<td>$\frac{M_{CCM}+2}{\sqrt{M_{CCM}}}$</td>
</tr>
<tr>
<td>Capacitors $C_1$ and $C_2$</td>
<td>$\frac{1}{2}$</td>
<td>$\frac{1}{\sqrt{2M_{CCM}}}$</td>
</tr>
<tr>
<td>Capacitor $C_3$</td>
<td>$\frac{1}{2}$</td>
<td>$\frac{1}{\sqrt{2M_{CCM}}}$</td>
</tr>
<tr>
<td>Diodes $D_1$ and $D_2$</td>
<td>$\frac{M_{CCM}+1}{2M_{CCM}}$</td>
<td>$\frac{M_{CCM}+2}{\sqrt{2M_{CCM}}}$</td>
</tr>
<tr>
<td>Inductor $L_4$</td>
<td>-</td>
<td>$\frac{M_{CCM}+1}{M_{CCM}}$</td>
</tr>
<tr>
<td>Inductors $L_2$</td>
<td>-</td>
<td>$\frac{1}{M_{CCM}}$</td>
</tr>
</tbody>
</table>

In order to show the total device number and voltage gain of the proposed converter, conventional buck-boost and converters in ref 18-19, a comparison is made between the proposed topology and other converters. The device number and voltage gain of the structures are given in Table 2. As shown in Table 2, the proposed structure uses lower number of elements. and the total device of the other converters is higher comparing to their gains and voltage stresses. Based on the low voltage stress of the proposed converter, the efficiency of the proposed converter is higher compared to its gain.

**TABLE II**

<table>
<thead>
<tr>
<th>Quantities of switches</th>
<th>Proposed converter</th>
<th>Converter in 18</th>
<th>Converter in 19</th>
</tr>
</thead>
<tbody>
<tr>
<td>Quantities of diodes</td>
<td>2</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>Quantities of capacitors</td>
<td>3</td>
<td>4</td>
<td>1</td>
</tr>
<tr>
<td>Quantities of inductors</td>
<td>2</td>
<td>3</td>
<td>2</td>
</tr>
<tr>
<td>Total device count</td>
<td>8</td>
<td>10</td>
<td>9</td>
</tr>
</tbody>
</table>

To verify the operation of the proposed converter, the experimental results are provided. The specifications are as follows:

1. input voltage : 23 V
2. switching frequency: 25 kHz
3. switch: IRFP460A
4. switch on-state resistances: 0.03 ohm

**E. Capacitor and inductor design**

To confirm that the inductors $L_1$ and $L_2$ always work in CCM, the needed inductance values can be achieved as follows:

$$I_o \geq \frac{\Delta I_s}{2}$$ (62)

The minimum of inductors $L_1$, $L_2$ and $L_3$ to work in ccm operation mode can be achieved as follows:

$$\frac{1+D}{1-D}I_o \geq \frac{(1-D)V_o}{4L_o f_s}$$ (63)

$$L_1 \geq \frac{V_o(1-D)^2}{4(1+D)L_o f_s} = \frac{42 \times (1-0.48)^2}{4 \times 2.6 \times (1+0.48) \times 25 \times 10^3}$$

$$= 29 \mu H$$ (64)

$$I_o \geq \frac{(1-D)V_o}{4L_{3,3} f_s}$$ (65)

$$L_2 \geq \frac{V_o(1-D)}{4f_s} = \frac{42 \times (1-0.48)}{4 \times 2.6 \times 25 \times 10^3}$$

$$= 77 \mu H$$ (66)

The minimum of the capacitors $C_1$, $C_2$ and $C_3$ can be obtained as follows:

$$C_{1,2} \geq \frac{D T V_o}{R A V_{C1}} = \frac{D I_o}{0.01 \times V_o \times f_s}$$

$$= \frac{0.48 \times 2.6}{0.02 \times 42 \times 25000} = 59 \mu F$$ (67)

$$C_3 \geq \frac{L_{C2,3} f_s}{\Delta V_{C2,3}} = \frac{2D T V_o}{R A V_{C2,3}} = \frac{2D I_o}{0.01 \times V_o \times f_s}$$

$$= \frac{2 \times 0.48 \times 2.6}{0.01 \times 42 \times 25000} = 118 \mu F$$ (68)

**III. EXPERIMENTAL RESULTS**

To verify the operation of the proposed converter, the experimental results are provided. The specifications are as follows:

1) input voltage : 23 V
2) switching frequency: 25 kHz
3) switch: IRFP460A
4) switch on-state resistances: 0.03 ohm
5) diodes $D_1$ and $D_2$: MUR860
6) diodes $D_1$ and $D_2$ forward resistances: 0.02 ohm
7) diodes $D_1$ and $D_2$ threshold voltages: 0.7 V
8) inductors $L_1$: 83 μH
9) inductors $L_2$: 245 μH
10) the equivalent series resistances (ESR) of inductor $L_1$: 0.012 ohm
11) the equivalent series resistances (ESR) of inductors $L_2$: 0.020 ohm
12) capacitor $C_1$ and $C_2$: 100 μF
13) capacitor $C_3$: 470 μF
14) the equivalent series resistances (ESR) of capacitors $C_1$ and $C_2$: 0.013 ohm
15) the equivalent series resistances (ESR) of capacitors $C_1$ and $C_2$: 0.022 ohm

The proposed converter is operated in CCM operation mode. The output voltage waveform is shown in Fig. 11(a). The output voltage is equal to 42 V and the output power is 110 W. The inductors $L_1$ and $L_2$ currents waveforms are shown in Figs. 11(b), 11(c) respectively. According to (14) and (16), the average values of inductors $L_1$ and $L_2$ currents are equal to 7.4 and 2.6 A, respectively, which closely agree with the experimental results. The voltages of diodes $D_1$ and $D_2$ waveforms is shown in Fig. 11(d). According to (48), the voltage across diodes $D_1$ and $D_2$ is equal to 44 V, which is equal to experimental results. The voltages of switch waveform is shown in Fig 11(e). According to (49), the voltage across switch $S$ is equal to 44 V, that is verified by the obtained value from experimental results. The current that flows through the diodes $D_1$ and $D_2$ is shown in Fig. 11(f). According to (19) and (20), the average value of diodes $D_1$ and $D_2$ currents is equal to 6.5 A, which closely agree with experimental results. The switch $S$ current waveform is shown in Fig. 11(g). According to (18), the average value of switch $S$ current is equal to 13 A, which closely agree with the experimental results. From Fig. 12, the measured efficiency of the proposed converter is 92.6% at the full-load condition and the maximum efficiency is 95.1%.
IV. CONCLUSION

In this paper, a novel transformer less buck boost dc-dc converter is presented. The output voltage of some sources such as fuel cell, photovoltaic and battery based systems are not regulated. Therefore voltage regulation is required to fix the DC-link voltage. Hence, a buck boost DC/DC converter is suitable to regulate the output voltage of these sources. The structure of the presented buck-boost converter is simple. In the proposed converter, only one main switch is utilized, which decreases the conduction loss of power switch and improves efficiency. The voltage stress across the power switch is low and switch with low on-state resistance can be selected. The step-up voltage gain of the proposed buck-boost converter is higher than that of the classic boost, buck-boost, CUK, SEPIC and ZETA converters. The proposed converter has simple structure; therefore, the control of the presented converter will be easy. The buck-boost converters is utilized in many applications like gadgets such as mobile phones and notebooks, fuel-cell systems, car electronic devices and LED drivers. Finally, the experimental results are provided to verify the feasibility of the proposed converter.

REFERENCES


