I. INTRODUCTION

In recent years, the microgrids have been continuously growing, especially hybrid systems that use many elements in direct current [1]. However, it is worth noting that energy storage systems with lithium-ion batteries are the main technology used for the new storage capacity [2], [3]. A DC microgrid can be described as a grid subsystem that allows the interconnection of distributed generators, power generation elements, energy storage, and local loads within a restricted region. Hybrid microgrids are composed of AC and DC subgrids, as shown in Figure 1, which are linked by one or more interface converter [4]. Microgrid must provide a secure and economical operation at all times for the system, while simultaneously providing a high level of reliability in the bus voltage for local loads. In the event of disconnection or AC-bus failures or any fluctuation in the photovoltaic generation, as discussed in [5], the dc-dc converter is used as a strategy to provide effective regulation of the DC-bus voltage [6]. The stability of DC-bus voltage is very important and when the microgrid is connected to the utility grid, the bus voltage is regulated by the AC-DC converter. However, when the microgrid works in an islanded operation, DC-bus voltage must be regulated by renewable energy generations and storage systems [7].

II. PROPOSED COUPLED INDUCTOR BIDIRECTIONAL DC-DC CONVERTER

In this section, the theoretical analysis and the operating stages for CCM are presented. Some simplifications are adopted and the components are considered ideals. The
converter operates in a steady-state and the variation of the energy stored in the inductors and capacitors is zero in a switching period $T_s$ and, the volt-second balance is applied.

Figure 2 shows the topology of the bidirectional DC-DC converter with coupled inductor (also known as the bidirectional coupled inductor non-isolated DC-DC converter) which was proposed in [13]. The circuit is composed by three power switches $S_1$, $S_2$ and $S_3$, one coupled inductor ($L_1$ and $L_2$), two filter capacitors $C_1$ and $C_2$, and dc sources $E_1$ and $E_2$. The topology has fewer components as compared to the bidirectional cascade, Buck-Boost, Cuk and Interleaved bidirectional dc-dc converters [14]–[17], it allows bidirectional power flow between dc sources $E_1$ and $E_2$. In general, the proposed converter operation can be divided into two operating modes: forward operating mode and the backwards operating mode. In forward operating mode, the power flow from $E_1$ to $E_2$ is obtained by modulating $S_2$, keeping $S_1$ turned-on and $S_3$ turned-off. In backwards operating mode, the power flow from $E_2$ to $E_1$ is obtained by modulating $S_3$, keeping the switches $S_2$ and $S_1$ turned-off.

![Fig. 2. Topology of the bidirectional coupled inductor DC-DC converter.](image)

The turns ratio of the coupled inductor is defined by (1), where the $n_1$ is the turns number of winding $L_1$, and $n_2$ is the turns number of winding $L_2$.

$$n = \frac{n_2}{n_1} = \sqrt{\frac{L_2}{L_1}} \quad (1)$$

Where:
- $L_1$ - The inductance value of the winding $n_1$;
- $L_2$ - The inductance value of the winding $n_2$.

A. Forward Operating Mode ($E_1$ to $E_2$)

In CCM, the inductor currents $i_{L_1}$ and $i_{L_2}$ will not reach zero. Based on this, the circuit operation in a switching period $T_s$ can be divided into two stages, as shown in Figures 3.a and 3.b. Figure 4 adequately demonstrates the main idealized waveforms of the proposed topology in forward operating mode. The operational modes are described in the following.

![Fig. 3. Topological stages in forward operating mode for CCM: a) First stage, and b) Second stage.](image)

In this operation mode, the power switch $S_1$ remains conducting on all operating time, therefore the waveforms of $S_1$ are unshown in Figure 4.

1° stage [$0 - DT_s$]: This stage starts at instant $t_0$, when the switches $S_1$ and $S_2$ are turned-on. The input current $i_{E1}(t)$ is flowing through the switches $S_1$, $S_2$ and inductor $L_1$, the current in $L_2$ is zero and current $i_{L_1}$ increases linearly. The voltages across the inductors $L_1$ and $L_2$ are $E_1$ and $n \cdot E_1$, respectively. The load $R_2$ receives energy from the output capacitor $C_2$. The equivalent circuit of first stage is illustrated in Figure 3.a.

2° stage [$DT_s - T_s$]: Starts when switch $S_2$ is turned-off and the internal diode of $S_3$ is forward biased. The currents $i_{L_1}$ and $i_{L_2}$ decrease linearly. The voltages across $L_1$ and $L_2$ decrease linearly. The voltages across $L_1$ and $L_2$ are $E_1 - E_2$ and $(E_1 - E_2) \cdot n / (1 + n)$, respectively. The energy stored in the coupled inductor is transferred to the load $R_2$ through $L_2$ and internal diode of $S_3$. Furthermore, the output current $i_{E2}$ is equal $i_{L_2} - i_{C_2}$. The equivalent circuit of second stage can be seen in the Figure 3.b.

1) DC Voltage Gain for Forward Mode: The expression for the DC voltage gain for CCM, can be given from the analysis of the average voltage across the inductor winding $L_1$.

$$V_{LL_{AVG}} = \frac{1}{T_s} \left[ \int_0^{DT_s} E_1 \, dt + \int_{DT_s}^{T_s} \frac{E_1 - E_2}{1 + n} \, dt \right] = 0. \quad (2)$$

Rearranging (2), the DC voltage gain in forward operating mode for CCM is presented by (3).

$$G_{forw} = \frac{E_2}{E_1} = \frac{1 + n \cdot D}{1 - D}. \quad (3)$$

The DC voltage gain of the proposed converter in forward operating mode ($G_{forw}$) is exposed graphically in Figure 5 for different $n$ condition and, compared to the conventional Boost converter, where $G_{Boost} = 1 / (1 - D)$.

2) Inductor Current Ripple for Forward Mode: The current ripple $\Delta i_{L_1}$ in the inductor $L_1$ can be obtained from the first stage, which is important to determine the inductance value of...
the coupled inductor.

\[ v_{L1} = L_1 \cdot \frac{\Delta i_{L1}}{dt} ; \quad \Delta i_{L1} = \frac{E_1 \cdot D}{L_1 \cdot f_s}. \]  \hfill (4)

Equation (4) can be used to calculate the inductance value of \( L_1 \) and \( L_2 \), as shown in the following expressions:

\[ L_1 = \frac{E_1 \cdot D}{\Delta i_{L1} \cdot f_s} ; \quad L_2 = L_1 \cdot n^2. \]  \hfill (5)

3) Output Capacitor Voltage Ripple: Voltage ripple of the output capacitors \( \Delta V_{C2} \) against the available charge \( \Delta Q \), is expressed as follows:

\[ \Delta V_{C2} = \Delta V_{C2} = \frac{\Delta Q}{C_2} = \frac{i_{C2} \cdot \Delta t}{C_2}. \]  \hfill (6)

During the first stage, the capacitor current \( i_{C2} \) is equal to the output current \( i_2 \). The interval time is equal to \( D \cdot T_s \), after replacing that in the (6), the output voltage ripple is given by (7).

\[ \Delta E_2 = \frac{i_2 \cdot D}{C_2 \cdot f_s}. \]  \hfill (7)

Through expression (7), the capacitance value can then be determined.

4) Components’ Stress Calculation for Forward Mode: The RMS value of capacitor current \( C_2 \) can be calculated by (8).

\[ i_{C2} = \left[ \frac{1}{T_s} \int_0^{D} (i_2)^2 \, dt \right]^{1/2} + \int_{D}^{T_s} \left( \frac{i_2}{1-D} - i_2 \right)^2 \, dt. \]  \hfill (8)

Through expression (8), the RMS current in the capacitor \( C_2 \) is calculated by (9).

\[ i_{C2} = i_2 \cdot \sqrt{\frac{D}{1-D}}. \]  \hfill (9)

The average and RMS current values through \( L_1 \) can be expressed by (10).

\[ I_{L1(AVG)} = I_1 ; \quad I_{L1(RMS)} = \frac{i_2}{1-\frac{D}{(2+n) \cdot n \cdot D + 1}}. \]  \hfill (10)

The average and RMS current values through \( L_2 \) can be expressed by (11).

\[ I_{L2(AVG)} = I_2 ; \quad I_{L2(RMS)} = \frac{i_2}{\sqrt{1-D}}. \]  \hfill (11)

The current stresses of switch \( S_1 \) can be expressed by (12). As the switch is constantly conducting, its drain-source voltage is equal to zero.

\[ I_{S1(AVG)} = I_1 ; \quad I_{S1(RMS)} = \frac{i_2}{1-D} \cdot \sqrt{\frac{D}{2+n} \cdot n \cdot D + 1}. \]  \hfill (12)

The current and voltage stresses on switch \( S_2 \) are given by (13) and (14), respectively.

\[ I_{S2(AVG)} = I_1 - I_2 ; \quad I_{S2(RMS)} = \frac{i_2}{1-D}. \]  \hfill (13)

\[ V_{S2} = \frac{E_1 \cdot n + E_2}{1+n}. \]  \hfill (14)

The current and voltage stresses at switch \( S_3 \) are given by (15) and (16).

\[ I_{S3(AVG)} = I_2 ; \quad I_{S3(RMS)} = \frac{i_2}{\sqrt{1-D}}. \]  \hfill (15)

\[ V_{S3} = E_1 + E_2 \cdot n. \]  \hfill (16)

B. Backward Operating Mode (\( E_2 \) to \( E_1 \))

In backward operating mode, the proposed topology presents two topological stages for CCM in each switching period \( T_s \), according to Figure 6. Following the description of the operating principle and idealized waveforms are presented.

1st stage \([0 - DT_s] \): In \( t = 0 \), the switch \( S_3 \) is turned on, and the coupled inductor windings store energy. The input current \( i_2 \) is flowing through the internal diode of \( S_1 \), \( S_2 \), \( L_2 \), and \( L_1 \). The voltages across \( L_2 \) and \( L_1 \) are equal to \( (E_2 - E_1) \cdot n/(1+n) \) and \( (E_2 - E_1)/(1+n) \), respectively. The load \( R \) and output capacitor \( C_1 \) receive energy from the inductor \( L_1 \) through the internal diode of \( S_1 \). The equivalent circuit of the first stage is illustrated in Figure 6.a.

2nd stage \([DT_s - T_s] \): This stage starts at instant \( t = D \cdot T_s \), the switch \( S_3 \) is turned off, and the internal diode of \( S_2 \) is forward biased. The current \( i_{L2} \) is zero, and \( i_{L1} \) decreases linearly. The voltage inductor across \( L_2 \) and \( L_1 \) are equal \( n \cdot E_1 \) and \( E_1 \), respectively. The energy stored in the coupled inductor is transferred to the load \( R \) through inductor \( L_1 \) and internal diode of \( S_1 \) and \( S_2 \). Additionally, the output current is \( i_{L1} - i_{L1} \) and the input current is equal to zero. The equivalent circuit of the second stage is illustrated in Figure 6.b.

Figure 7 illustrates the main idealized waveforms of proposed topology in backward operating mode.
1) DC Voltage Gain for Backward Mode: Considering the proposed topology is operating in steady state, CCM and backward operating mode, the average voltage across inductor winding \( L_2 \) should be zero for a switching period. Then the DC voltage gain can be given from (17).

\[
V_{L2,AVG} = \frac{1}{T_s} \left[ \int_0^{D T_s} \frac{(E_2 - E_1) \cdot n}{1 + n} \cdot D \cdot t \cdot d t \right] .
\]  

Manipulating the average voltage of the inductor \( V_{L2} = 0 \), the DC voltage gain in the backward operating mode is calculated by (18).

\[
G_{\text{back}} = \frac{E_1}{E_2} = \frac{D}{(1 + n) \cdot n \cdot D} .
\]  

The DC voltage gain of the proposed topology in backward operating mode is shown graphically in Figure 8, and compared to the conventional Buck converter.

![Fig. 7. Idealized waveforms in backward operating mode for CCM.](image)

2) Inductor Current Ripple for Backward Mode: The inductor current ripple in CCM can be calculated during the second stage. From Figure 7, the inductor current is obtained, which is given by (19).

\[
v_{L2} = \frac{L_2}{D T_s} \frac{\Delta I_{L2}}{d t} ; \quad \Delta I_{L2} = (E_2 - E_1) \cdot n \cdot D \cdot \frac{1 + n \cdot D}{1}.
\]  

From (19), the inductance values of \( L_2 \) and \( L_1 \) are calculated by (20).

\[
L_2 = \frac{(E_2 - E_1) \cdot n \cdot D}{1 + n \cdot D} ; \quad L_1 = \frac{L_2}{n^2} .
\]  

3) Output Capacitor Voltage Ripple: In the previous analysis, the output capacitor is assumed to be very large to keep their voltages constant. In practice, the voltages cannot be kept constant with a finite capacitance value. Thus, there is inevitably a voltage ripple over it. The relationship of the voltage ripple and the capacitance is given by (21).

\[
\Delta V_{C1} = \Delta E_1 = \frac{\Delta Q}{C_1} .
\]  

The output capacitor \( C_1 \) is discharged by the load current \( I_1 \) during the second stage. The total time is equal to \( (1 - D) \cdot T_s \). Therefore, the voltage ripple can be obtained by (22).

\[
\Delta E_1 = \frac{I_1 \cdot D - I_2}{C_1 \cdot f_s} .
\]  

Through expression (22), the capacitance value can then be determined.

4) Components’ Stress Calculation for Backward Mode: The mathematical expression that describing the RMS value of capacitor current can be given by:

\[
I_{C1} = \sqrt{\frac{1}{T_s} \left[ \int_0^{D T_s} \left( I_1 - \frac{I_2}{D} \right)^2 \cdot d t \right] + \int_0^{T_s} \left( \frac{I_1 - I_2}{1 - D} - \frac{I_2}{D} \right)^2 \cdot d t} .
\]  

Through expression (23), the RMS current is calculated by (24).

\[
I_{C1} = (I_1 \cdot D - I_2) \cdot \frac{1}{(1 - D) \cdot D} .
\]  

The average and RMS values of the current through \( L_2 \) can be expressed by (25).

\[
I_{L2(AVG)} = I_2 ; \quad I_{L2(RMS)} = \frac{I_2}{\sqrt{D}} .
\]  

The average and RMS values of the current through \( L_1 \) can be expressed by (26).

\[
I_{L1(AVG)} = I_1 ; \quad I_{L1(RMS)} = \sqrt{I_2^2 + \left( I_1 - I_2 \right)^2 \cdot \left( 1 - D \right)} .
\]  

The average and RMS values of the current through switch \( S_1 \) are defined by (27). During this operation mode, the switch will remain permanently “ON”, so the drain-source voltage is
The DC voltage gain in the backward operating mode is

\[ AV = \frac{E_2}{E_1} \]

Fig. 7. Idealized waveforms in backward operating mode for CCM.

From Figure 7, the inductor current ripple is

\[ i_{L2}(t) = i_{L1}(t) + \Delta I \]

The winding backward operating mode, the average voltage across inductor

\[ V_L = \frac{1}{2} (E_2 - E_1) \]

The voltage ripple over it. The relationship of the

\[ \Delta V_L = \frac{E_2 - E_1}{2-n} \]

Considering the

\[ E = \frac{1}{2} \cdot DT_s \]

of capacitor current can be given by:

\[ i_{L2}(t) = \frac{1}{2} \cdot DT_s \]

Through expression (23), the RMS current is calculated by

\[ I_{RMS} = \frac{1}{\sqrt{2}} \cdot I_{DC} \]

The current and voltage stresses of switch \( S_2 \) are given by

\[ I_{S2} = I_2 \quad V_{S2} = E_2 + (E_1 - E_2) \cdot n \]

Based on the aforementioned operation analyses, the current and voltage stress of switches \( S_3 \) can be obtained directly as follows:

\[ I_{S3} = I_2 \quad V_{S3} = E_2 + E_1 \cdot n \]

Table I shows the comparison between the calculated values with the numerical simulation of the circuit for evaluating all the mathematical expressions. By which, it is verified that the presented equations are able to correctly represent the operation of the converter.

In Table II the proposed converter is compared with similar topologies that have a single coupled inductor and bidirectional power flow, presented in [18]–[26]. The comparison is made in terms of voltage gain, number of components, voltage stress and soft switching performance. It should be noted that the voltage is compared for same \( E_1 \) and \( E_2 \) and the leakage inductance of coupled-inductors based converters is ignored. The proposed converter offers the following improvement over those reported; low device count, good voltage gain ratio, less device voltage stress and substantial increase in utility rate of magnetic core.

### III. EXPERIMENTAL RESULTS

Initial investigation of the steady-state operation analysis is provided, in order to verify the operation and adequately evaluate the performance of the proposed converter, a 600 W prototype was designed and verified experimentally. The converter specifications are exposed in Table III. A photograph of the implemented prototype is exhibited in Figure 9. The PWM modulation was generated on LAUNCHXL-F28377S C2000 LaunchPad.

The experimental waveforms are obtained from the prototype operating in both modes of operation. For the suppression and damping of voltage spikes produced by the leakage inductance of the coupled inductor, a RCD-type circuit (snubber) was connected in parallel to switch \( S_2 \) and another RC series circuit was connected in parallel with \( S_3 \) [27].

Fig. 9. Experimental prototype of proposed converter.

Figure 10 shows the experimental results when the converter operates in forward mode.

Fig. 10. Currents and voltages in the load and power supply for forward operation \( (E_1 = 100 V, \ E_2 = 256 V) \).

Figure 11 shows the waveforms in the coupled inductor for forward operating mode. It is confirmed that converter presents two operating stages in CCM. Moreover, there is a great similarity between experimental and theoretical waveforms. The waveforms were obtained for a duty cycle of \( D = 0.45 \). In this case, the converter is operating in the step-up voltage mode.

Fig. 11. Voltage and current waveforms in the coupled inductor for forward operation; \( Ch1 \) - current \( i_{L1}(t) \), \( Ch2 \) - current \( i_{L2}(t) \), \( Ch3 \) - voltage \( V_{L1}(t) \) and, \( Ch4 \) - voltage \( V_{L2}(t) \).

The voltage and current of switch \( S_2 \) and diode \( D_3 \) are shown in Figures 12 and 13, respectively.
TABLE II
Comparison of the Proposed Topology with the Existing Coupled Inductor Converters

| Topology          | [18] | [19] | [20] | [21] | [22] | [23] | [24] | [25] | [26] | proposed |
|-------------------|------|------|------|------|------|------|------|------|------|----------|
| No. of components | 3/3  | 4/4  | 3/4  | 4/4  | 4/4  | 4/4  | 4/5  | 4/4  | 4/4  | 3/3/2    |
| Switch / Diode / Capacitor | 2      | 2    | 2    | 2    | 2    | 2    | 2    | 2    | 2    | 2        |
| Voltage gain      | 2(n+1) - D | n+1 | n+1 | n+1 | n+1 | n+1 | n+1 | n+1 | n+1 | n+1     |
| Voltage gain      | 2(n+1) - D | 2(n+1) | 2(n+1) | 2(n+1) | 2(n+1) | 2(n+1) | 2(n+1) | 2(n+1) | 2(n+1) | 2(n+1) |
| Voltage stress on switches | No-present | nD | nD | nD | nD | nD | nD | nD | nD | nD |
| Turn ratio (n/E1) | 1.55, 17 turns/11 turns - 38AWG Litz wire |
| Current ripple (ΔI1) | 20A/Hz |
| Inductance (L2) | 288 µH | 691 µH |
| Capacitance (C2) | 120 µF | 450 V | 13.6 µF | 630 V | 28 V |
| Switches (S1, S2, S3) | S1C1008KE by ROHM |
| Snubber (RC1 + RC2) | 68kΩ, 220nF, FR207 + 120Ω, 0.56 nF |

TABLE III
Main Prototype Specifications and Parameters

| Parameters                  | Value          |
|----------------------------|----------------|
| Rated power (P)            | 600 W          |
| Low voltage (E1)           | 100 V          |
| High voltage (E2)          | 300 V          |
| Switching frequency (fS)   | 20kHz          |
| Duty cycle (D)             | forward mode (0.45) / backward mode (0.58) |

Fig. 12. Voltage and current waveforms in the switch S3 for forward operation: Ch2 - current iS2(t) and, Ch4 - voltage VS2(t).

Fig. 13. Voltage and current waveforms in the diode D1 for forward operation: Ch1 - current iD1(t) and, Ch3 - voltage VL1(t).

In Figure 14 is presented an experimental result that shows in detail the voltage and the current in the capacitor C2. In this result, it is verified that the voltage applied to the load is stabilized and with low ripple.

Figures 15–18 show the experimental results that correspond to the operation in the backward mode.

In Figure 15, the voltages and the windings currents of the coupled inductor are shown for backward operating mode (E2 = 300 V, E1 = 100 V). It is observed that the converter also operates in CCM. In this case, it is operating in the step-down voltage mode with a duty cycle of D = 0.58 .

Fig. 16. Voltage and current waveforms in the diode D2 for backward operation: Ch2 - current iD2(t) and, Ch4 - voltage VS2(t).

Fig. 17. Voltage and current waveforms in the coupled inductor for backward operation; Ch1 - current iL1(t), Ch2 - current iL2(t), Ch3 - voltage VL1(t) and, Ch4 - voltage VL2(t).

Fig. 18 shows the capacitor voltage VC1(t) and its current iC1(t). The capacitor voltage is same as E1 and is continuous with low ripple.

The results confirmed that the proposed dc-dc converter can be used as a step-up and step-down converter. In addition, it
allows a wide range of applications that require bidirectional power flow and a wide range of regulation of the dc-bus.

IV. CONCLUSIONS

This paper presented a dc-dc bidirectional converter with coupled inductor for dc-microgrid applications where dc voltage of source varies in a wide range. The paper discussed operating principle, equivalent circuits, theoretical waveforms and mathematical analysis of the proposed converter. From the experiment result, it was verified that the proposed dc-dc converter is bidirectional, it allows a wide output voltage range and greater dc voltage gain is compared with the traditional Buck and Boost converters.

The topology merges the main characteristics of Boost and Buck dc-dc converters, resulting in a converter with small volume, low weight and reduced number of components. It has advantages in terms of dc voltage gain without any isolation and can better handle the dc bus voltage regulation. Moreover, the converter is competitive and qualified for energy storage elements, such as batteries, and ultracapacitors with the high voltage dc bus in electric vehicles. The experimental results with resistive loads, in forward and backward modes, verified the operating of the proposed converter, when bidirectional power flow is required and a good voltage regulation of the dc bus is essential.

ACKNOWLEDGEMENTS

The authors thank to Universidade do Estado de Santa Catarina - UDESC, the Coordenação de Aperfeiçoamento de Pessoal de Nível Superior - Brasil (CAPES), the MCTIC and CTI Renato Archer for providing institutional resources to this research and technical support. This research registered a patent number BR 10 2018 010019-0 of intellectual property under the domain of the Fundação Universidade do Estado de Santa Catarina - UDESC.

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