Novel Bidirectional DC-DC Converters Based on The Three-State Switching Cell

Abstract—It is well known that there is an increasing demand for bidirectional dc-dc converters for applications that range from renewable energy sources to electric vehicles. Within this context, this work proposes novel dc-dc converter topologies that use the three-state switching cell (3SSC), whose well-known advantages over conventional existing structures are ability to operate at high current levels, while current sharing is maintained by a high frequency transformer; reduction of cost and dimensions of magnetics; improved distribution of losses, with consequent increase of global efficiency; and reduction of cost associated to the need of semiconductors with lower current ratings. Three distinct topologies can be derived from the 3SSC: one dc-dc converter with reversible current characteristic able to operate in the first and second quadrants; one dc-dc converter with reversible voltage characteristic able to operate in the first and third quadrants; and one dc-dc converter with reversible current and voltage characteristics able to operate in four quadrants. Only the topology with bidirectional current characteristic is analyzed in detail in terms of the operating stages in both nonoverlapping and overlapping modes, while the design procedure of the power stage elements is obtained. In order to validate the theoretical assumptions, an experimental prototype is also implemented, so that relevant issues can be properly discussed.

Index Terms—adjustable speed drives, choppers, dc-dc converters, dc machines.

I. INTRODUCTION

The widespread use of dc-dc converters include distinct applications e.g. standalone and grid-connected renewable energy systems [1] [2], uninterruptible power supplies [3], aerospace and defense [4], motor drives [5], battery chargers [6], although many other examples can be found in literature. Among the existing topologies, bidirectional dc-dc converters have been the scope of recent works due to the increasing interest in clean-energy vehicle applications [7].

For instance, the proposed structure in [8] employs a coupled inductor with the same number of turns in the primary and secondary sides. Even though the converter employs only one additional switch compared with the conventional buck/boost converter (also known as synchronous buck converter or class C chopper [9]), the voltage stress across the main switches is significantly increased and the topology does not seem to be adequate for high current applications. Another bidirectional dc-dc topology is introduced in [10], where four active switches are used. In this case, the voltage stresses are reduced to half of the high voltage side. However, high component count and increased conduction losses are expected in this case. A topology with high voltage gain characteristic is proposed in [11] in order to achieve bidirectional power flow. Cost and complexity involving many active and passive components represent a serious drawback in this case. A similar circuit is described in [12], but the use of five active switches brings significant complexity associated to appreciable conduction losses.

Although there are numerous approaches in literature, some desirable and mandatory characteristics for EV applications are reduced dimensions associated to high power density, high efficiency and reduced conduction and commutation losses in the semiconductor elements, and, of course, bidirectional power flow [13]. With the introduction of the three-state switching cell (3SSC), many novel converter topologies have been proposed over the last few years [14]. For instance, a 3SSC-based buck converter is studied in [15], whose performance is superior than that of the conventional buck converter for high power, high-current applications. Besides, it has also been demonstrated in [16] that 3SSC-based dc-dc converter topologies present good
performance over the entire range of the duty cycle i.e. for narrow or wide voltage conversion ratios. In terms of high voltage step-up, dc-dc boost-type converters using the 3SSC have also been proposed in [17] and [18], which are interesting choices for EV applications [19]. High power factor rectifiers have also been analyzed in [20]–[22], thus validating the 3SSC as an interesting solution to achieve high power density with improved efficiency.

This paper proposes novel bidirectional dc-dc converters using the 3SSC to obtain class C, class D, and class E choppers that can be used in EV applications. Initially, a review on the aforementioned basic circuits is presented, so that the 3SSC-based circuits can be properly derived. Although three structures are introduced, the study is focused on a converter with reversible current characteristic, whose qualitative and quantitative analyses are presented in detail so that an experimental prototype can be implemented. Finally, some results on a dc motor drive are presented and properly discussed to validate the proposal.

II. BRIEF REVIEW OF BASIC DC CHOPPERS AND PROPOSAL OF NOVEL TOPOLOGIES USING THE 3SSC

The PWM (pulse width modulation) switch model proposed in [23] shows that the semiconductors in the classic dc-dc converters are arranged in a three-terminal approach i.e. active, passive, and common. The aforementioned circuit model is also known as two-state switching cell (2SSC), from which bidirectional converters can be derived by associating unidirectional topologies, as shown in Fig. 1, Fig. 2, and Fig. 3. For instance, the class C chopper in Fig. 1 consists in the integration of buck and boost converters [23].

When the operation with bidirectional current and/or voltage characteristic is desired, the so-called class C, class D, and class E choppers in Fig. 1, Fig. 2, and Fig. 3 can be used. These are simple topologies which can be used to drive a dc motor, thus allowing two-quadrant or four-quadrant operation. Typically, the simplified representation of a dc motor is given by a resistance \( R \) connected in series with an inductance \( L \) and a dc voltage source \( E \). In the class C chopper shown in Fig. 1, the current through the motor can be either positive (if switch \( S_1 \) is turned on) or negative (if switch \( S_2 \) is turned on), while the voltage is always positive. In the class D chopper shown in Fig. 2, the voltage across the motor is either positive (if switches \( S_1 \) and \( S_2 \) are both turned on) or negative (when switches \( S_1 \) and \( S_2 \) are both turned off and the energy stored in the inductor is delivered to the source through the diodes), while the current is always positive. On the hand, both current and voltage can be positive or negative in Fig. 3, where four-quadrant operation is possible.

The topologies represented in Fig. 1, Fig. 2, and Fig. 3 are adequate for bidirectional applications e.g. renewable energy sources and EVs [24]–[25]. However, as the power levels increase, traditional structures may become inadequate considering that the series or parallel association of converter may be necessary in this case. It is then reasonable to assume that the 3SSC concept can be applied in the conception of novel choppers class C, class D, and class E, which are depicted in Fig. 4, Fig. 5, and Fig. 6. The load is represented by a motor \( M \) and the 3SSC is evidenced by the high-frequency autotransformer composed by windings \( T_1 \) and \( T_2 \).

At a first glance, the proposed structures seem more complex than the conventional approaches presented in Fig. 1, Fig. 2, and Fig. 3 due to high component count. However, they are adequate for high power, high current applications considering that the load current is equally shared between two branches due to the presence of the autotransformer. It is also worth to mention that this work is dedicated to the detailed study of the chopper shown in Fig. 4 only, while the remaining converters will not be further analyzed. Of course, the analysis carried out in the paper can be easily extended to the topologies in Fig. 5 and Fig. 6.

The converter in Fig. 4 is composed by one 3SSC-based boost converter [14] and one 3SSC-based buck converter [15] in an arrangement that is similar to the topology shown in Fig. 1. Initially, the qualitative analysis is presented so that it is possible to define the design procedure for the topology. Then it is possible to implement an experimental prototype so that the aforementioned theoretical assumptions are validated and relevant issues can be discussed.
Fig. 1. Class C chopper based on the 2SSC.

Fig. 2. Class D chopper based on the 2SSC.

Fig. 3. Class E chopper based on the 2SSC.

Fig. 4. Class C chopper based on the 3SSC.

Fig. 5. Class D chopper based on the 3SSC.
Fig. 6. Class E chopper based on the 3SSC.

III. QUALITATIVE STUDY

By definition, two modes regarding the states of the active switches can be obtained for every topology based on the 3SSC [14]. For value of the duty cycle $D$ higher than 0.5, the semiconductors operate in overlapping mode (OM) considering that two switches remain turned on simultaneously. For $D<0.5$, the converter is supposed to operate in nonoverlapping mode (NOM), as only one switch is on in a given stage.

Analogously to the converter in Fig. 1, the proposed topology in Fig. 4 is able to operate:

- as a buck converter when switches $S_1$ and $S_2$ are driven, while switches $S_3$ and $S_4$ must remain off and the current flows from the input voltage source to the load i.e. both voltage and current are positive;
- as a boost converter when switches $S_3$ and $S_4$ are driven, while switches $S_1$ and $S_2$ must remain off and the current flows from the load to the input voltage source i.e. the voltage is positive and the current is negative.

The analysis carried out in this paper considers that the current through the inductor $L$ never becomes null as operation occurs in continuous conduction mode (CCM). Four distinct operating modes will result then and the following conditions are assumed:

- all semiconductors are ideal;
- the converter operates in steady-state condition;
- the drive signals of the main switches are phase-shifted by 180°;
- a unity turns ratio autotransformer is used;
- the magnetizing current is negligible if compared with the load current.

The following parameters can also be defined:

$V_{gs1}, V_{gs2}, V_{gs3}, V_{gs4}$ - gating signals applied to switches $S_1, S_2, S_3, \text{ and } S_4$, respectively;

$I_L$ - current through inductor;

$I_{s1}, I_{s2}, I_{s3}, I_{s4}$ - currents through switches $S_1, S_2, S_3, \text{ and } S_4$;

$I_{d1}, I_{d2}, I_{d3}, I_{d4}$ - currents through diodes $D_1, D_2, D_3, \text{ and } D_4$;

$V_{s1}, V_{s2}, V_{s3}, V_{s4}$ - voltages across switches $S_1, S_2, S_3, \text{ and } S_4$;

$V_{d1}, V_{d2}, V_{d3}, V_{d4}$ - voltages across diodes $D_1, D_2, D_3, \text{ and } D_4$.

A. Operation as a Buck Converter in NOM and CCM

The operating stages that define this mode are shown in Fig. 7 and the relevant theoretical waveforms are presented in Fig. 8.

**First stage** [$t_0, t_1$] (Fig. 7 (a)): Switch $S_1$ is turned on, while $S_2$ remains off. Besides, switches $S_3$ and $S_4$ are not turned on in this operation mode. According to Kirchhoff’s laws, it can be stated that the voltages across the autotransformer windings are equal to $V/2$, where $V$ is the input voltage. By analyzing the loop that comprehends $V_i, S_1, T_2, L$ and $V_o$, the following expression results:

$$\frac{V}{2} - L \frac{dI_L}{dt} - V_o = 0$$

(1)

where $V_o$ is the voltage across the load.

**Second stage** [$t_1, t_2$] (Fig. 7 (b)): Switch $S_1$ is turned off, while $S_2$ still remains off. The load current flows through diodes $D_2$ and $D_4$, which are forward biased in this case. The voltage across the autotransformer windings is null due to the existing magnetic coupling. The energy stored in the inductor is delivered to the load, but it is not fully discharged, thus characterizing the operation in CCM. In this case, expression (2) is valid.
\[-L \frac{dI_L}{dt} + V_o = 0\] (2)

Third stage \([t_2, t_3]\) (Fig. 7 (c)): Switch \(S_1\) remains off while switch \(S_2\) is turned on. This stage is analogous to the first one and expression (1) is also valid in this case.

Fourth stage \([t_3, t_4]\) (Fig. 7 (d)): Both switches \(S_1\) and \(S_2\) are turned off. This stage is analogous to the second one and expression (2) also represents such condition.

The static gain of the converter is defined as:

\[G_V = \frac{V_o}{V_i}\] (3)

Considering that the current ripple through the inductor defined as \(\Delta I_L\) is the same for time intervals \((t_1-t_0)\) and \((t_2-t_1)\), one can obtain:

\[\Delta I_L (t_1-t_0) = \Delta I_L (t_2-t_1)\] (4)

From (1), it is possible to write:

\[\Delta I_L = \frac{dI_L}{dt} = \frac{-V_o}{L} + \frac{V_i}{2L}\] (5)

Analogously, the following expression can be obtained from (2):

\[\Delta I_L = \frac{dI_L}{dt} = \frac{V_0}{L}\] (6)

The aforementioned time intervals can be determined from Fig. 8 as:

Fig. 7. Operating stages as a buck converter in NOM-CCM.
\((t_1 - t_0) = DT_s\) 

\((t_2 - t_1) = \frac{T_s}{2} (1 - 2D)\) 

where \(T_s\) is the switching period.

Substituting expressions (5) to (8) in (4) gives:

\[-\frac{V_o}{L} + \frac{V_i}{2L} (DT) = \frac{V_i}{L} \left[ \frac{T_s}{2} (1 - 2D) \right] \]

Finally, expression (9) can be rearranged as:

\[\frac{V_o}{V_i} = G_V = D\] 

Therefore the static gain given in (10) is the same as that valid for the classical buck converter in CCM.

**B. Operation as A Buck Converter in OM and CCM**

The operating stages that correspond to this mode are shown in Fig. 9 and the relevant theoretical waveforms are presented in Fig. 10.

**First stage** \([t_0, t_1]\) (Fig. 9 (a)): Both switches \(S_1\) and \(S_2\) are turned on simultaneously, while \(S_3\) and \(S_4\) remain turned off when the converter operates in buck mode. In this case, the current through the inductor corresponds to the sum of the currents through windings \(T_1\) and \(T_2\). Considering that current sharing is maintained by the autotransformer and also the polarity given in the circuit, it can be stated that the voltage across the windings is null.

Then the following expression can be obtained:

\[L \frac{di_L}{dt} + V_o - V_i = 0\]
Second stage $[t_1, t_2]$ (Fig. 9 (b)): Switch $S_1$ remains on, but switch $S_2$ is turned off. The energy stored in the inductor is delivered to the load and the current decreases linearly. The load current corresponds to the currents through windings $T_1$ and $T_2$, which are equal because unity turns ratio is considered.

By analyzing the circuit, the following expressions can be obtained:

$$V_{T1} = V_{T2} = \frac{V_i}{2}$$

and

$$-L \frac{dI_L}{dt} + V_o - \frac{V_i}{2} = 0$$

where $V_{T1}$ and $V_{T2}$ are the voltages across $T_1$ and $T_2$, respectively.

Third stage $[t_2, t_3]$ (Fig. 9 (c)): Both switches $S_1$ and $S_2$ are on. Besides, it can be seen that the switching states of all semiconductor elements are the same as those in the first stage. Of course, expression (11) is also valid in this stage.

Fourth stage $[t_3, t_4]$ (Fig. 9 (d)): Switch $S_1$ is turned off, but switch $S_2$ remains turned on. This stage is analogous to the second one and the analysis of the circuit also leads to expressions (12) and (13).

![Diagrams of operating stages](image-url)
Once again, let us obtain the static gain of the converter considering the waveforms shown in Fig. 10 and also that expression (4) is valid. From (11), it is possible to write:

\[ \Delta I_L = \frac{dI_L}{dt} = \frac{V_i}{L} - \frac{V_o}{L} \] (14)

Analogously, the following expression can be obtained from (13):

\[ \Delta I_L = \frac{dI_L}{dt} = \frac{V_o}{L} - \frac{V_i}{2L} \] (15)

Besides, the analysis of Fig. 10 gives:

\[ (t_1 - t_0) = \frac{T_s}{2} (2D - 1) \] (16)
\[ (t_2 - t_1) = T_s (1 - D) \] (17)

The static gain can then be obtained by substituting (14) to (17) in (4):

\[ \left( \frac{V_i}{L} - \frac{V_o}{L} \right) \left[ \frac{T_s}{2} (2D - 1) \right] = \left( \frac{V_o}{L} - \frac{V_i}{2L} \right) \left[ T_s (1 - D) \right] \]

\[ \frac{V_o}{V_i} = G_v = D \] (19)

Comparing (10) and (19), it is possible to state that the static gain of the proposed topology operating as a buck converter (where the current is positive) is the same in either NOM or OM. Besides, it behaves as a classical buck converter in both aforementioned modes.

### C. Operation as A Boost Converter in NOM and CCM

The operating stages that define this mode are shown in Fig. 11 and the relevant theoretical waveforms are presented in Fig. 12.
First stage [$t_0$, $t_1$] (Fig. 11 (a)): Switch $S_3$ is turned on, while switch $S_4$ is turned off. Both switches $S_1$ and $S_2$ are never turned on during the operation in boost mode. The voltage across the transformer windings is $V/2$ according to Kirchhoff’s laws. By analyzing the loop that comprehends $V_i$, $S_3$, $T_2$, $L$, and $V_o$ and considering that the current flows from the load to the source, the following expression results:

$$\frac{V}{2} + L \frac{dI_L}{dt} - V_o = 0$$  \hspace{1cm} (20)

Second stage [$t_1$, $t_2$] (Fig. 11 (b)): Switch $S_3$ is turned off and switch $S_4$ is still off. The current flows from the load to the source though the inductor and diodes $D_1$ and $D_2$. The voltage across the transformer windings is null due to the existing magnetic coupling between $T_1$ and $T_2$. By analyzing the equivalent operating stage, the following expression results:

$$V_o + L \frac{dI_L}{dt} - V_i = 0$$  \hspace{1cm} (21)

Third stage [$t_2$, $t_3$] (Fig. 11 (c)): Switch $S_4$ is turned on and switch $S_3$ remains off. This condition is analogous to the first stage and expression (20) is also valid.

Fourth stage [$t_3$, $t_4$ (Fig. 11 (d)): Switch $S_3$ remains off, while $S_4$ is turned off. This condition is analogous to the second stage and expression (21) is also valid.

In order to obtain the static gain, expressions (3) and (4) must be considered. From the first stage, the following expression can be derived:

$$\Delta I_L = \frac{dI_L}{dt} = \frac{V_i}{L} - \frac{V_i}{2L}$$  \hspace{1cm} (22)

From the second stage, it is also possible to write:

$$\Delta I_L = \frac{dI_L}{dt} = \frac{V_o}{L} - \frac{V_i}{L}$$  \hspace{1cm} (23)

Fig. 11. Operating stages as a boost converter in NOM-CCM.
According to Fig. 12, the time intervals represented in (4) are given by:

\[(t_1 - t_0) = DT\]  \hspace{1cm} (24)

\[(t_2 - t_1) = \frac{T}{2} (1 - 2D)\]  \hspace{1cm} (25)

Substituting expressions (22) to (25) in (4) gives:

\[
\frac{V_o}{L} - \frac{V_i}{2L} (DT) = \left( \frac{V_i - V_o}{L} \right) \left[ \frac{T}{2} (1 - 2D) \right]
\]  \hspace{1cm} (26)

\[
\frac{V_i}{V_o} = G_V = \frac{1}{1 - D}
\]  \hspace{1cm} (27)

According to (27), the static gain is the same as that valid for the classical boost converter in CCM.

D. Operation as A Boost Converter in OM and CCM

The operating stages that correspond to this mode are shown in Fig. 13 and the relevant theoretical waveforms are presented in Fig. 14.

First stage \([t_0, t_1]\) (Fig. 13 (a)): Switches \(S_3\) and \(S_4\) are turned on simultaneously. The current is equally shared between \(T_1-S_4\) and \(T_2-S_3\) due to the unity turns ratio of the autotransformer. According to the given polarity, the voltage across the windings is null. There is no energy transfer from the load to the source and the inductor is charged by the load itself. Besides, the following expression can be obtained:

\[
L \frac{dI_i}{dt} - V_o = 0
\]  \hspace{1cm} (28)

Second stage \([t_1, t_2]\) (Fig. 13 (b)): Switch \(S_4\) is turned off and switch \(S_3\) is still on. The inductor and the load are responsible for
delivering energy to the source through \(T_1-D_2\) and \(T_2-S_3\) and current sharing is maintained. The following expression represents this stage:

\[
V_o + L \frac{dI_L}{dt} - \frac{V_i}{2} = 0
\]  
(29)

**Third stage** \([t_2, t_3]\) (Fig. 13 (c)): This stage is identical to the first one and the current flows through \(S_1\) and \(S_2\).

**Fourth stage** \([t_3, t_4]\) (Fig. 13 (d)): This stage is analogous to the second one, although the current flows through \(S_4\) while \(S_2\) is off. However, expression (29) is still valid.

Finally, it is possible to define the static gain of the converter in the operating mode represented by the waveforms shown in Fig. 10, from which the following time intervals can be obtained:

\[
(t_1 - t_0) = \frac{T_s (2D - 1)}{2}
\]  
(30)

\[
(t_2 - t_1) = T_s (1 - D)
\]  
(31)

Substituting (28) to (31) in (4) gives:

\[
\frac{V_o}{L} \left[ \frac{T_s (2D - 1)}{2} \right] = \left( \frac{V_i}{2L} - \frac{V_o}{L} \right) [T_s (1 - D)]
\]  
(32)

\[
\frac{V_i}{V_o} = G_v = \frac{1}{(1 - D)}
\]  
(33)

Expression (33) demonstrates that the static gain of the proposed topology in NOM and OM is the same as that for the classical boost converter in CCM.
This section presents the expressions that allow the design of the 3SSC-based dc-dc converter. Considering the operating modes described in Session III, the accurate design of the converter considers the highest values assumed by the current and voltage stresses involving the power stage elements.

A. Inductor

According to the study developed in Session III, it has been demonstrated that the time intervals that define each operating stage are not the same considering the operation in NOM and OM. Besides, the static gain is not the same for the operation as a buck or a boost converter. Therefore, the current ripple through the inductor can be defined by the expressions shown in Fig. 15. It is also possible to normalize them as a function of the current ripple, inductance, switching period, and input voltage, as parameters \( \alpha_1 \) and \( \alpha_2 \) can be obtained for NOM and OM, respectively and plotted in Fig. 16.

\[
\alpha_1 = \frac{2L\Delta I_L}{T_s V_i} = (1-2D)D \\
\alpha_2 = \frac{2L\Delta I_L}{TV_i} = (D-1)(1-2D)
\]

The normalized curves in Fig. 16 are valid for the operation in either buck or boost mode and show that the maximum values of the current ripple are \( \alpha_1=0.125 \) at \( D=0.25 \) and \( \alpha_2=0.125 \) at \( D=0.75 \). Substituting \( D=0.25 \) and \( D=0.75 \) in the expressions given in Fig. 15, it is possible to define the maximum ripple current corresponding to the worst possible case. Therefore the inductance can be obtained by:
The rms current through the inductor in NOM can be calculated from the analysis of the relevant operating stages described in Session III i.e.

$$I_{L_{(rms)}} = \sqrt{\frac{2}{T_s} \int_0^{D_T} \left( I_m + \frac{V_i (1-2D)t}{2L} \right)^2 dt + \frac{2}{T_s} \int_0^{T_s (2D-1)} \left( I_m - \frac{V_i D t}{L} \right)^2 dt}$$

(37)

Analogously, the rms current through the inductor in OM is given by:

$$I_{L_{(rms)}} = \sqrt{\frac{2}{T_s} \int_0^{\frac{T_s (1-D)}{2}} \left( I_m + \frac{V_i (1-D)t}{L} \right)^2 dt + \frac{2}{T_s} \int_0^{\frac{T_s (1-D)}{2}} \left( I_m - \frac{V_i (2D-1)t}{2L} \right)^2 dt}$$

(38)

The inductor core can be designed according to the following expression:

$$A_e \cdot A_w = \frac{L \cdot I_{L_{(max)}} \cdot I_{L_{(rms)}}}{K_w \cdot J_{(max)} \cdot B_{(max)}} \cdot 10^4 \quad [\text{cm}^4]$$

(39)

where $A_e$ is the effective core area; $A_w$ is the window area; $I_{L_{(max)}}$ is the maximum value of the current through inductor $L$; $I_{L_{(rms)}}$ is...
the rms current through inductor \( L \); \( K_u \) is the utilization factor of the window; \( J_{\text{max}} \) is the maximum current density; and \( B_{\text{max}} \) is the maximum variation of the magnetic flux.

The number of turns \( N_L \) for the inductor is:

\[
N_L = \frac{LI_L(\text{max})}{B(\text{max})A_c} \cdot 10^4
\]

(40)

The core loss in the inductor can be obtained from:

\[
P_{\text{L(coi)}} = \Delta B^{2.4} \left( K_H f_L + K_E f_L^2 \right) V_c
\]

(41)

where \( \Delta B \) is the magnetic flux variation; \( K_H = 4 \cdot 10^{-5} \) is the hysteresis loss coefficient; \( f_L = 2 f_s \) is the operating frequency of the inductor i.e. twice the switching frequency \( f_s \); \( K_E = 4 \cdot 10^{-10} \) is the eddy-current loss coefficient; and \( V_c \) is the effective volume of the ferrite core.

The copper loss in the inductor is:

\[
P_{\text{L(copper)}} = \rho l_L N_L I_{L(\text{rms})}^2
\]

(42)

where \( \rho \) is the copper resistivity measured at 70 °C; \( l_L \) is the average length of one turn; \( N_L \) is the number of turns of the inductor; \( n_L \) is the number of parallel wires; \( S_f \) is the cross section area of copper wire.

B. Autotransformer

The maximum voltage across the autotransformer windings is:

\[
V_{T(\text{max})} = \frac{V_i}{2}
\]

(43)

The rms current through the autotransformer is equal to half of the rms current through the inductor in any operation mode (NOM/OM and/or buck/boost) i.e.

\[
V_{T(\text{max})} = \frac{V_i}{2}
\]

(44)

The transformer can be designed according to the following expression [15]:

\[
A_c \cdot A_w = \frac{P_o}{2 f_s K_t K_p K_u J_{\text{max}} B_{\text{max}}} \cdot 10^4
\]

(45)

where \( f_s \) is the switching frequency, \( K_t \) is the topology factor, \( K_p \) is the utilization factor of the primary winding, and \( P_o \) is the output power.

The core loss in the autotransformer is given by the following expression:

\[
P_{\text{T(core)}} = \Delta B^{2.4} \left( K_H f_T + K_E f_T^2 \right) V_c
\]

(46)

where \( \Delta B \) is the magnetic flux variation and \( f_T = 2 f_s \) is the operating frequency of the transformer.

The copper loss in the autotransformer windings is:

\[
P_{\text{T(copper)}} = \frac{2 \rho l_T N_T I_{T(\text{rms})}^2}{n_T S_f}
\]

(47)

where \( l_T \) is the average length of a single turn; \( N_T \) is the number of turns of the 3SSC autotransformer; and \( n_T \) is the number of parallel wires.
C. Diodes

The maximum voltage across the diodes is:

$$V_{D_{\text{max}}} = V_i$$  \hspace{1cm} (48)

The average and rms currents through the diodes in NOM can be calculated from the analysis of the relevant operating stages described in Session III i.e.

$$I_{D_{\text{avg}}} = \frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} + \frac{V_i(1-2D)t}{4L} \right) dt + \frac{2}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} - \frac{V_iD_t}{2L} \right) dt$$  \hspace{1cm} (49)

$$I_{D_{\text{rms}}} = \sqrt{\frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} + \frac{V_i(1-2D)t}{4L} \right)^2 dt + \frac{2}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} - \frac{V_iD_t}{2L} \right)^2 dt}$$  \hspace{1cm} (50)

Analogously, the average and rms currents through the diodes in OM are given by:

$$I_{D_{\text{avg}}} = \frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} - \frac{V_i(2D-1)t}{4L} \right) dt$$  \hspace{1cm} (51)

$$I_{D_{\text{rms}}} = \sqrt{\frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} - \frac{V_i(2D-1)t}{4L} \right)^2 dt}$$  \hspace{1cm} (52)

D. Main Switches

The maximum voltage across the switches is:

$$V_{S_{\text{max}}} = V_i$$  \hspace{1cm} (53)

The average and rms currents through the switches in NOM are:

$$I_{S_{\text{avg}}} = \frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} + \frac{V_i(1-2D)t}{4L} \right) dt$$  \hspace{1cm} (54)

$$I_{S_{\text{rms}}} = \sqrt{\frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} + \frac{V_i(1-2D)t}{4L} \right)^2 dt}$$  \hspace{1cm} (55)

The average and rms currents through the switches in OM are:

$$I_{S_{\text{avg}}} = \frac{2}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} + \frac{V_i(1-D)t}{2L} \right) dt + \frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} - \frac{V_i(2D-1)t}{4L} \right) dt$$  \hspace{1cm} (56)

$$I_{S_{\text{rms}}} = \sqrt{\frac{2}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} + \frac{V_i(1-D)t}{2L} \right)^2 dt + \frac{1}{T_s} \int_0^{\frac{T_s}{2}} \left( \frac{I_m}{2} - \frac{V_i(2D-1)t}{4L} \right)^2 dt}$$  \hspace{1cm} (57)

Finally, the maximum current through the switches corresponds to half of the maximum current through the inductor i.e.

$$I_{S_{\text{max}}} = \frac{I_m}{2} = \frac{I_o}{2} + \frac{DT_sV_i(1-2D)}{8L}$$  \hspace{1cm} (58)

where $I_o$ is the total load current.
V. EXPERIMENTAL RESULTS

In order to validate the theoretical assumptions, the experimental prototype shown in Fig. 17, whose ratings are given in Table I, was designed and implemented according to the procedure developed in Session IV. A dc motor is also used as load in the proposed topology. Some relevant waveforms are obtained and discussed as follows.

During the operation of the dc motor, the converter may assume NOM or OM depending on value of the duty cycle. Fig. 18 presents the gating signals applied to switches $S_1$, $S_2$, $S_3$, and $S_4$. The drive signals applied to $S_1$ and $S_3$ are complementary, as well as those regarding $S_2$ and $S_4$. In this case, switches $S_1$ and $S_2$ operate in NOM, while $S_3$ and $S_4$ are in OM. The opposite occurs in Fig. 19, where $S_1$ and $S_2$ are in OM and $S_3$ and $S_4$ are in NOM.

Fig. 20 shows the currents through switches $S_1$ and $S_2$, which remain turned on simultaneously, although the waveforms are phase-displaced by 180°. Besides, it is possible to see that the output voltage is positive, while the converter operates in buck mode.

**TABLE I**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>$V_i = 24$ V</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>$f_s = 30$ kHz</td>
</tr>
<tr>
<td>Ripple current through inductor $L$</td>
<td>$\Delta I_L = 15% I_o$</td>
</tr>
</tbody>
</table>

**Designed Elements**

- **Inductor** $I = 10$ μH, core NEE-42/21/20-IP12R by Thornton, $N_L = 6$ turns – 30×AWG 23
- **Transformer** Core NEE-42/21/20-IP12R by Thornton, $N_{T1} = N_{T2} = 2$ turns – 2×AWG 23
- **Main switches** MOSFET IRFP260P by International Rectifier
- **Diodes $D_1$...$D_n$** Ultrafast diode RHRP3060 by Fairchild Rectifier

![Fig. 17. Experimental prototype.](image-url)
Fig. 18. Gating signals applied to switches $S_1-S_2$ (NOM) and $(S_3-S_4)$ (OM) (CH1: $V_{GS(S1)}$, CH2: $V_{GS(S2)}$, CH3: $V_{GS(S3)}$, CH4: $V_{GS(S4)} - 10 \text{ V/div.}, 10 \mu\text{s/div.}$).

Fig. 19. Gating signals applied to switches $S_1-S_2$ (OM) and $(S_3-S_4)$ (NOM) (CH1: $V_{GS(S1)}$, CH2: $V_{GS(S2)}$, CH3: $V_{GS(S3)}$, CH4: $V_{GS(S4)} - 10 \text{ V/div.}, 10 \mu\text{s/div.}$).

Fig. 20. Currents through switches $S_1$ (CH3: $I_{S1} - 10 \text{ A/div.}, 10 \mu\text{s}$) and $S_2$ (CH2: $I_{S2} - 10 \text{ A/div.}, 10 \mu\text{s}$) in OM and output voltage (CH1: $V_o - 20 \text{ V/div.}, 10 \mu\text{s}$).

VI. CONCLUSION

Three dc-dc converter topologies employing the 3SSC have been presented in this paper. By employing the 3SSC, the current is equally shared among the semiconductor elements, as the proposed approach is adequate for high-current applications where the efficiency is supposed to be high. Besides, current sharing is maintained considering that the autotransformer windings present nearly the same impedance.
The quantitative and qualitative analyses of a converter with bidirectional current characteristic have been derived in order to obtain a detailed and accurate design procedure of the power stage elements, from which an experimental prototype could be implemented and evaluated. The behavior is similar to that expected in theory in terms of the current and voltage waveforms and stresses regarding the semiconductor elements.

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REFERENCES


