Three-Phase Flyback Push–Pull DC–DC Converter: Analysis, Design, and Experimentation

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Abstract—In this paper a step-up/step-down isolated dc–dc converter referred to as a three-phase flyback push–pull dc–dc converter is presented. The power circuit is constituted by a pair of coupled inductors, a three-phase transformer, a capacitor, three switching transistors and three power diodes. The proposed converter offers the advantages of compact passive devices, low conduction power losses, full duty cycle range (0–100%), and inherent protection against transformer saturation. Furthermore, filter sizes are minimized to duty cycles of around 1/3 and 2/3. These characteristics make this converter suitable for many applications, especially in low-voltage high-power applications such as telecommunications power supply, battery chargers, and renewable power systems. The operating principle and the idealized mathematical analysis in continuous conduction mode are presented. Experimental data were obtained from a laboratory prototype with an input voltage of 125 V, output voltage of 100 V, load power of 1000 W, and switching frequency of 42 kHz. The measured prototype efficiency was 94% for full load and 96% for 400 W.

Index Terms—DC–DC converter, flyback converter, push–pull converter, three-phase dc–dc converter.

I. INTRODUCTION

The voltage-fed and current-fed push–pull converters are classical isolated dc–dc converters used widely in low voltage applications due to their simplicity and good transformer core utilization since with these converters two quadrants of the BH curve are used [1]–[7]. However, only the current-fed configuration is able to drive large current rate without causing transformer saturation due to an unequal volt-second product. Also, these converters are not able to operate in the full duty cycle range, because this would lead to a short-circuit in the voltage-fed mode and an open-circuit in the current-fed mode.

The Weinberg and flyback-current-fed push–pull converters are also push–pull topologies and are able to operate in the full duty cycle range (0 to 100%) [8]–[16]. This characteristic is achieved by the use of a pair of coupled inductors. However, only the flyback-current-fed push–pull converter shown in Fig. 1 uses effectively the coupled inductors, since its windings drive the energy for the whole duty cycle range, while the secondary winding of other such converters drives the energy only when the duty cycle is less than 50%. Moreover, this converter does not increase the number of semiconductor devices required. In addition, both of these converters provide continuous constant current shape at the input and output when operated with a constant duty cycle of 1/2. Consequently, operation around this point provides smaller filter sizes.

This paper proposes the three-phase flyback push–pull converter, which is a three-phase version of the flyback-current-fed push–pull dc–dc converter [13] as are shown in Fig. 1. The proposed converter is constituted by three switching transistors (S1, S2, S3), three diodes (D1, D2, D3), a capacitor C, a pair of coupled inductors Lf, and a high-frequency three-phase transformer T in a wye-wye connection. The resulting characteristics of the three-phase version are as follows:

1) operation in full duty cycle range;
2) absence of the transformer saturation problem, since the high-source impedance provided by the coupled inductors avoids transformer saturation due to the unequal volt-second product applied to its windings;
3) continuous constant current shapes at the input and the output for set duty cycles of 1/3 and 2/3 with consequently smaller filter sizes;
4) the transistors are connected to the same reference point; thus, these devices can be driven by nonisolated circuit drivers;
5) low conduction losses since there is only a single semiconductor device in series with the input and output sources;
6) small filter sizes, the three-phase system increases the operating frequency of filters to three times the switching frequency;
7) a good transformer copper and core utilization since the transformer windings are placed in a single common core as is shown in Fig. 2;
8) high-power processing, the sharing of the input and output current improves the heat transfer of the power devices allowing high current rates.

The characteristics listed previously show that the proposed converter merges the main push–pull converter characteristics with those of the three-phase converters [17]–[26]. The result is a converter with smaller filter sizes able to drive high current rates at low voltage rates.

This study presents an idealized theoretical analysis and experimental results for the proposed converter operating in continuous conduction mode (CCM). Also, a comparison with the conventional and push–pull types of flyback converters highlights the advantages provided by the proposed converter.

II. PRINCIPLE OF OPERATION

The following assumptions are made to simplify the theoretical description of the operating principle:
1) the converter is operating in steady state;
2) the magnetic flux of the coupled inductors is continuous, this means that the converter is operating in CCM;
3) all switching devices and the three-phase transformer are ideals, so these power devices do not cause power losses and the transformer magnetizing inductance is infinite;
4) the capacitance of C is large enough to consider a constant output voltage;
5) the windings of the coupled inductors are tightly coupled and have negligible resistance;
6) the magnetizing inductance $L_m$ of the coupled inductors $L_f$ is referred to its primary side and has a finite value since this device stores energy;
7) the power transistors are symmetrically driven using the three-phase gating signals as is shown in Fig. 3;
8) the turns ratios of the coupled inductors and the three-phase transformer are considered equal ($N_L = N_f$). This equality is assumed since it provides a single static gain in CCM for all duty cycle range.

The output voltage $V_o$ is regulated employing pulse-width modulation (PWM) to control the duty cycle $D$, which is defined as the ratio between the transistor on time $t_{on}$ and the switching period $T_s$. The transistors can be driven using a duty cycle range from 0% to 100%, that is, the proposed converter admits overlapping and nonoverlapping conduction of transistors. Thus, the converter operation is divided into three regions listed in Table I. In region R1 the transistors are not overlapped, while in regions R2 and R3 overlapping of the conduction of two and three transistors, respectively, occurs.

### A. Operating Principle in R1

In this region, the energy supplied by the input source is discontinuous and it occurs during the transistor conduction periods. On the other hand, the capacitor and load arrangement is continuously supplied with energy from the circuit. Therefore, the input and output currents have discontinuous and continuous shapes, respectively. The main waveforms in CCM for region R1 are shown in Fig. 3.

The first stage starts at instant $t_0$ when $S_1$ is turned ON, causing $D_1$ to be reverse biased. Diodes $D_2$ and $D_3$ remain conducting. The equivalent circuit for this stage is shown in Fig. 4(a). Ideally, the magnetomotive forces (mmf) of the three-phase transformer are equal; thus, the currents flowing through $S_1$, $D_2$, and $D_3$ have the relation $N_f i_{s_1} = i_{d_2} = i_{d_3}$. Therefore, one-third of the magnetizing current of the coupled inductors flows through the primary winding ($i_{l_p} = i_{m}/3$), while two-thirds are conducted through the secondary winding ($i_{l_s} = 2N_f i_{m}/3$). The voltages across the active primary windings of transformer and coupled inductors are functions of the input and output sources and are given as $v_{T_1} = 2(E_i + N_f V_o)/3$ and $v_{L_p} = (E_i - 2N_f V_o)/3$, respectively. In this stage, the coupled
inductors store part of the energy supplied by the input source and, consequently, the input voltage is greater than $2N_TV_o$. The voltage applied across transistors $S2$ and $S3$ is $3/2$ times the active primary transformer winding ($V_{S2} = V_{S3} = E_i + N_TV_o$). This voltage referred to the secondary side is applied across diode $D1$ ($V_{D1} = E_i/N_T + V_o$).

The second stage starts at instant $t_1$ when $S1$ is turned OFF. All diodes are forward biased; therefore, the voltages across the transformer windings are zero. Thus, the stored energy of the coupled inductors is directly transferred to the output through its secondary winding ($V_{Ls} = -V_o$) as shown in Fig. 4(b). The magnetizing current referred to the secondary side is evenly shared among the diodes ($i_{D1} = i_{D2} = i_{D3} = N_Ti_m/3$). The voltage across any transistor is $E_i + N_TV_o$.

In the third and fifth stages, the input source again supplies energy to the circuit, part of this energy is stored by the coupled inductors, which is transferred to the output in the fourth and sixth stages, respectively. Fig. 3 shows that the capacitor behavior is the opposite to that of the coupled inductors, that is, it charges in the even stages and discharges in the odd stages.

### B. Operating Principle in R2

In this region, the energy flowing through the input and output are continuous. Therefore, the input and output currents are continuous but both currents have pulsating shapes. The main waveforms in CCM for region R2 are shown in Fig. 5.

The first stage starts at instant $t_0$ when $S1$ is turned ON while $S3$ is conducting. Only diode $D2$ is forward biased. The equivalent circuit for this stage is shown in Fig. 6. The relation between the currents flowing through active power devices is $N_Ti_{S1} = N_Ti_{S3} = i_{D2}$. Thus, two-thirds of the magnetizing current of coupled inductors flow through the primary winding ($i_{Lp} = 2i_m/3$) and one-third is conducted by the secondary windings ($i_{Ls} = N_Ti_m/3$). The voltages across the active primary windings of the transformer and coupled inductors are $v_{Tp1} = (E_i + N_TV_o)/3$ and $v_{Lp} = (2E_i - N_TV_o)/3$, respectively. In this stage, the input voltage is higher than $N_TV_o/2$, but it subsequently becomes lower than $2N_TV_o$. Therefore, the coupled inductors store part of the energy supplied by the input source. The voltage applied across transistor $S2$ is three times the voltage across active primary transformer windings ($V_{S2} = E_i + N_TV_o$). This voltage referred to the secondary side is applied through diodes $D1$ and $D3$ ($V_{D1} = V_{D3} = E_i/N_T + V_o$).

The second, fourth, and sixth stages are similar to the even stages of region R1 with the difference that now the coupled inductors transfer their stored energy to the output since the input voltage is lower than $2N_TV_o$. In the third and fifth stages, the coupled inductors store again part of the energy supplied by the input source. Fig. 5 shows that the capacitor has the same behavior as that of R1, that is, it charges in the even stages and discharges in the odd stages.
C. Operating Principle in R3

In this region, the input source supplies energy to the circuit throughout the switching cycle, while the output receives energy from the circuit if only two transistors are turned ON. Therefore, the input current is continuous while the output current is discontinuous. The main waveforms for a switching period in R3 are shown in Fig. 7.

The first stage starts at instant $t_0$ when $S1$ is turned ON while $S2$ and $S3$ are conducting. All diodes are reverse biased. Therefore, the voltages across the transformer windings are zero as shown in Fig. 8. Thus, the input source transfers energy directly to the coupled inductors through its primary winding ($v_{LP} = E_i$). The voltage across any diode is $E_i/N_T + V_o$. Whole magnetizing current of coupled inductors flows through the primary side and is evenly distributed through $S1$, $S2$, and $S3$ ($i_{S1} = i_{S2} = i_{S3} = i_i/3$). This storage stage occurs again in the third and fifth stages.

The second, fourth, and sixth stages are similar to the even stages of region R3 with the difference that now the coupled inductors transfer their stored energy since the voltage $N_T V_o/2$ becomes higher than the input voltage. Fig. 7 shows that in the even stages, the energy consumed by the load is supplied by the capacitor that is recharged in the odd stages.
III. MATHEMATICAL ANALYSIS

In this section, the main equations for CCM that allow the current and voltage stresses of the proposed three-phase flyback push–pull dc–dc converter to be computed are developed. The quantitative information contained in Figs. 3, 5, and 7 is used for this purpose.

The input and output currents of the proposed converter such as the flyback converter have pulsating shapes. Consequently, the rms currents in the power devices as well as the output voltage ripple in CCM are largely dependent on the average value of the magnetizing current. Thus, the expressions to compute these electrical values are developed assuming that the magnetizing inductance of coupled inductors is large enough for its current ripple to be neglected.

A. Static Gain

If the proposed converter is operating in steady state then the average voltage across primary winding of coupled inductors should be zero for a switching cycle. Therefore, considering the symmetrical operation, this is also true for one-third of a switching period

\[ V_{Lp} = \frac{1}{T_s} \int_{t_o}^{t_2+T_s} v_{Lp} dt = \frac{3}{T_s} \int_{t_o}^{t_2} v_{Lp} dt = 0. \]  

(1)

The substitution of voltages and times that correspond to regions R1, R2, and R3 yields

\[ (E_i - 2N_T V_o) t_{on} - N_T V_o \left( \frac{T_s}{3} - t_{on} \right) = 0 \]  

\[ (2E_i - N_T V_o) \left( t_{on} - \frac{T_s}{3} \right) \]  

\[ + (E_i - 2N_T V_o) \left( \frac{2T_s}{3} - t_{on} \right) = 0 \]  

\[ E_i \left( t_{on} - \frac{2T_s}{3} \right) + (2E_i - N_T V_o) \left( T_s - t_{on} \right) = 0. \]  

(2)

(3)

(4)

Solving these equations, (5) is obtained which represents the static gain of the proposed converter in CCM. The result shows that for the converter, there is a single expression for the whole duty cycle range and that it is the same as that of the conventional flyback converter

\[ \frac{V_o}{E_i} = \frac{D}{N_T (1 - D)}. \]  

(5)

B. Magnetizing Current Ripple of Coupled Inductors

The magnetizing current ripple can be determined in the stored stage or transferred stage. Considering the former, we have the following:

\[ \Delta I_m = \frac{1}{L_m} \int_{t_o}^{t_1} v_{Lp} dt. \]  

(6)

Substituting the voltages and times for each region results in (7), (8), and (9), which correspond to regions R1, R2, and R3, respectively

\[ \Delta I_m = \frac{(E_i - 2N_T V_o) t_{on}}{3L_m} \]  

(7)

\[ \Delta I_m = \frac{(2E_i - N_T V_o) \left( t_{on} - \frac{T_s}{3} \right)}{3L_m} \]  

(8)

\[ \Delta I_m = \frac{E_i \left( t_{on} - \frac{2T_s}{3} \right)}{3L_m}. \]  

(9)

Using (5) in the aforementioned equations results in the per-unit magnetizing current ripple given by

\[ \Delta I_m = \frac{\Delta I_m}{V_o T_s} = \begin{cases} \frac{1}{3} & \text{for R1} \\ \frac{(2 - 3D)(3D - 1)}{9D} & \text{for R2} \\ \frac{(1 - D)(3D - 2)}{3D} & \text{for R3}. \end{cases} \]  

(10)

C. RMS Currents of Coupled Inductors

The primary and secondary rms currents of coupled inductors are computed through

\[ I_{Lp,prms} = \left( \frac{3}{T_s} \int_{t_o}^{t_2} \left( i_{Lp} \right)^2 dt \right)^{1/2} \]  

(11)

\[ I_{Ls,prms} = \left( \frac{3}{T_s} \int_{t_o}^{t_2} \left( i_{Ls} \right)^2 dt \right)^{1/2}. \]  

(12)

The average values for the magnetizing current and the output current of any region are related as follows:

\[ I_m = \frac{I_o}{N_T (1 - D)}. \]  

(13)

Neglecting the magnetizing current ripple of the current shapes shown in Figs. 3, 5, and 7, and using the relationship given by (13) the per-unit primary and secondary rms currents of the coupled inductors are obtained as (14) and (15)

\[ I_{Lp,rms} = \frac{I_{Lp,prms}}{I_o} = \begin{cases} \frac{1}{(1 - D)N_T \sqrt{\frac{D}{3}}} & \text{for R1} \\ \frac{1}{3(1 - D)N_T \sqrt{\frac{5D - 2}{3}}} & \text{for R2} \\ \frac{1}{(1 - D)N_T \sqrt{\frac{3 - 5D}{3}}} & \text{for R3} \end{cases} \]  

(14)

\[ I_{Ls,rms} = \frac{I_{Ls,prms}}{I_o} = \begin{cases} \frac{1}{1 - D \sqrt{\frac{3 - 5D}{3}}} & \text{for R1} \\ \frac{1}{3(1 - D) \sqrt{\frac{7 - 9D}{3}}} & \text{for R2} \\ \frac{1}{1 - D \sqrt{\frac{1 - D}{3}}} & \text{for R3}. \end{cases} \]  

(15)

D. RMS Currents of Three-Phase Transformer

The sharing of the input and output currents by the proposed converter provides constant current amplitudes through the primary and secondary transformer windings for whole duty cycle range as shown in Figs. 3, 5, and 7, which have instantaneous
TABLE II

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed</th>
<th>Push-pull</th>
<th>Conventional</th>
</tr>
</thead>
<tbody>
<tr>
<td>$G_{mec}$</td>
<td>$\frac{D}{N_T(1-D)}$</td>
<td>$\frac{D}{N_T(1-D)}$</td>
<td>$\frac{D}{N_T(1-D)}$</td>
</tr>
<tr>
<td>$\Delta I_m$</td>
<td>$\begin{cases} N_T \frac{1-3D}{3} \ N_T \frac{2-3D}{3} \ N_T \frac{9D}{3D-2} \end{cases}$ for R1</td>
<td>$\begin{cases} N_T \frac{1-2D}{2} \ N_T \frac{2(2D-1)}{2D} \end{cases}$ for R2</td>
<td>$N_T(1-D)$</td>
</tr>
<tr>
<td>$\Delta V_o$</td>
<td>$\begin{cases} \frac{D(1-3D)}{3(1-D)} \ \frac{3(1-D)(2-3D)}{9(1-D)} \ \frac{3D-2}{3} \end{cases}$ for R1</td>
<td>$\begin{cases} \frac{D(1-2D)}{2(D-1)} \ \frac{2(D-1)}{2(D-1)} \end{cases}$ for R2</td>
<td>$D$</td>
</tr>
<tr>
<td>$I_{Crms}$</td>
<td>$\begin{cases} \frac{\sqrt{3}D}{\sqrt{9D-2}} \ \frac{\sqrt{3}(3-5D)}{3(1-D)} \ \frac{\sqrt{3}(1-D)}{3(1-D)} \end{cases}$ for R1</td>
<td>$\begin{cases} \frac{\sqrt{2}D}{2N_T(1-D)} \ \frac{\sqrt{2}(2-3D)}{2(1-D)} \end{cases}$ for R2</td>
<td>$\frac{\sqrt{D}}{N_T(1-D)}$</td>
</tr>
<tr>
<td>$I_{Lprms}$</td>
<td>$\begin{cases} \frac{3N_T(1-D)}{\sqrt{9D-2}} \ \frac{3N_T(1-D)}{\sqrt{3(5D-2)}} \ \frac{3N_T(1-D)}{3(1-D)} \end{cases}$ for R1</td>
<td>$\begin{cases} \frac{\sqrt{2}D}{2N_T(1-D)} \ \frac{\sqrt{2}(2-3D)}{2(1-D)} \end{cases}$ for R2</td>
<td>$\frac{\sqrt{D}}{N_T(1-D)}$</td>
</tr>
<tr>
<td>$I_{Lrms}$</td>
<td>$\begin{cases} \frac{\sqrt{3}(3-5D)}{3(1-D)} \ \frac{\sqrt{3}(1-D)}{3(1-D)} \end{cases}$ for R1</td>
<td>$\begin{cases} \frac{\sqrt{2}(2-3D)}{2(1-D)} \end{cases}$ for R2</td>
<td>$\frac{\sqrt{1-D}}{1-D}$</td>
</tr>
<tr>
<td>$I_{Tprms}$</td>
<td>$\begin{cases} I_m \sqrt{D} \ I_m(1-D) \sqrt{1-D} \end{cases}$</td>
<td>$\begin{cases} 2N_T(1-D) \sqrt{1-D} \end{cases}$</td>
<td>$N_T(1-D)$</td>
</tr>
<tr>
<td>$I_{Trms}$</td>
<td>$\begin{cases} I_m \sqrt{D} \ I_m(1-D) \sqrt{1-D} \end{cases}$</td>
<td>$\begin{cases} 2N_T(1-D) \sqrt{1-D} \end{cases}$</td>
<td>$N_T(1-D)$</td>
</tr>
<tr>
<td>$V_{Smx}$</td>
<td>$E_i + N_T V_o$</td>
<td>$E_i + N_T V_o$</td>
<td>$E_i + N_T V_o$</td>
</tr>
<tr>
<td>$V_{Dmx}$</td>
<td>$E_i/N_T + V_o$</td>
<td>$E_i/N_T + V_o$</td>
<td>$E_i/N_T + V_o$</td>
</tr>
</tbody>
</table>

values of $i_m/3$ and $N_T i_m/3$, respectively. Therefore, the rms values of these currents, neglecting the magnetizing current ripple, are a function of the conduction time of the semiconductor devices in series with these windings. Thus, the primary and secondary rms currents of the three-phase transformer are computed through

\[
I_{Tprms} = \frac{I_{Lprms}}{I_o} = \frac{\sqrt{D}}{3(1-D)N_T} \quad (18)
\]

\[
I_{Trms} = \frac{I_{Tprms}}{I_o} = \frac{\sqrt{1-D}}{3(1-D)} \quad (19)
\]

E. Output Voltage Ripple

The output voltage ripple neglecting the magnetizing current ripple is given by

\[
\Delta V_o = \frac{1}{C} \int_{t_1}^{t_2} (i_{Ls} - I_o) dt. \quad (20)
\]
From Figs. 3, 5, and 7, the voltage ripple for regions R1, R2, and R3 yields

\[ \Delta V_o = \frac{1}{C} \left( I_m N_T - I_o \right) \left( \frac{T_s}{3} - t_{on} \right) \]  
(21)

\[ \Delta V_o = \frac{1}{C} \left( 2I_m N_T - I_o \right) \left( \frac{2T_s}{3} - t_{on} \right) \]  
(22)

\[ \Delta V_o = \frac{1}{C} \left( \frac{I_m N_T}{3} - I_o \right) \left( T_s - t_{on} \right). \]  
(23)

The substitution of (5) into (21), (22), and (23) yields (24), which is the per-unit output voltage ripple in CCM

\[ \Delta V_o = \frac{\Delta V_o C}{I_o T_s} = \begin{cases} \frac{D(1 - 3D)}{3(1 - D)} & \text{for R1} \\ \frac{(3D - 1)(2 - 3D)}{9(1 - D)} & \text{for R2} \\ \frac{3D - 2}{3} & \text{for R3}. \end{cases} \]  
(24)

The substitution of (5) into (21), (22), and (23) yields (24), which is the per-unit output voltage ripple in CCM

\[ \Delta V_o = \frac{\Delta V_o C}{I_o T_s} = \begin{cases} \frac{D(1 - 3D)}{3(1 - D)} & \text{for R1} \\ \frac{(3D - 1)(2 - 3D)}{9(1 - D)} & \text{for R2} \\ \frac{3D - 2}{3} & \text{for R3}. \end{cases} \]  
(24)

**TABLE III**  
SPECIFICATIONS OF PROTOTYPE

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage ( (E_i) )</td>
<td>100 V±25%</td>
</tr>
<tr>
<td>Output voltage ( (V_o) )</td>
<td>100 V</td>
</tr>
<tr>
<td>Output power ( (P_o) )</td>
<td>1000 W</td>
</tr>
<tr>
<td>Switching frequency ( (f_s) )</td>
<td>42 kHz</td>
</tr>
</tbody>
</table>

**F. RMS Capacitor Current**

If negligible output voltage ripple is assumed, then the rms current of the capacitor can be calculated as follows:

\[ I_{C_{\text{rms}}}^2 = I_{L_{\text{rms}}}^2 - I_o^2. \]  
(25)

Substituting (15) in (25), the per-unit capacitor rms current is given by the following:

\[ I_{C_{\text{rms}}} = \frac{I_{C_{\text{rms}}}}{I_o} = \begin{cases} \sqrt{\frac{D(1 - 3D)}{3(1 - D)^2}} & \text{for R1} \\ \sqrt{\frac{(3D - 1)(2 - 3D)}{9(1 - D)^2}} & \text{for R2} \\ \frac{3D - 2}{3(1 - D)} & \text{for R3}. \end{cases} \]  
(26)
G. Voltage Stresses Across Semiconductor Devices

The voltages across turned-off transistor and reverse-biased diode for any region such as in the case of the conventional flyback converter is the sum of the output and input voltages referred to the primary or secondary side, respectively

\[ V_S = E_i + N_T V_o \]  
\[ (27) \]
\[ V_D = \frac{E_i}{N_T} + V_o. \]  
\[ (28) \]

In CCM, these voltages can be expressed as

\[ V_S = E_i \frac{D}{1 - D} \]  
\[ (29) \]
\[ V_D = \frac{E_i}{N_T} \frac{D}{1 - D}. \]  
\[ (30) \]

IV. COMPARISON OF CONVERTERS

In this section, the proposed converter is compared with the flyback-current-fed push–pull and conventional flyback converters, as shown in Table II. It can be seen that the static gain and semiconductor voltage stresses are the same for all converters, while the current and voltage ripples, and the rms currents differ for each converter. The comparison reveals that the current and voltage ripples of the proposed converter are 1/3 lower than those of the conventional converter, and 3/2 lower than the values required for the push–pull type as shown in Fig. 9. Moreover, the rms currents through the capacitor and the secondary inductor windings of the proposed converter are lower than those values for the other ones as can be seen in Fig. 10. Consequently, the proposed converter allows a greater reduction in filter sizes than the other converters.

The comparison of the rms transformer currents, which are the same rms currents across the switching transistors and diodes, shows that the rms currents of the proposed converter are 1/3 lower than those of the conventional converter and 3/2 lower than those of the push–pull type. Therefore, the proposed converter is able to drive high-current rates with lower current stresses across semiconductor devices.

In addition, the proposed converter presents two particular points, one-third and two-thirds of the duty cycle, where the current ripple as well as the voltage ripple and rms current of the capacitor fall to zero. Therefore, for these points, the coupled inductors and output capacitor sizes are minimized.

V. EXPERIMENTAL RESULT

The validation of theoretical analysis was carried out through an experimental prototype designed with the specifications shown in Table III. The schematic circuit of the laboratory prototype is shown in Fig. 11, the power components of which are listed in Table IV. The gate drive signals were generated by Kit DSP TMS320LF2407. Nonisolated drive circuits were employed to drive the switching transistors. The photo in Fig. 12 shows the laboratory prototype, where the passive power devices and main parts of the circuit can be identified. It can be observed that the windings of the three-phase high-frequency transformer were placed in a single common core.

The experimental current waveforms obtained from the experimental prototype operating in regions R1, R2, and R3 are...
shown in Figs. 13, 14, and 15, respectively. The results show that the input and output currents have pulsating shapes in any of the regions. The currents at the input and output sides differ for each region, in region R1 the input current is discontinuous and the output current is continuous, in region R2 both of currents are continuous, while in region R3 the input current becomes continuous and output current assumes a discontinuous flow. In all regions, the input and output currents are evenly shared through the transistors and diodes, and their current amplitudes across these devices have constant values.

The converter efficiency curves shown in Fig. 16 were obtained for regions R1 and R2 with input voltages of 125 and 85 V, respectively, and in both cases the output voltage was maintained at a constant value of 100 V. The efficiency curve for R3 is not shown since the laboratory prototype is not designed to operate in this region, so the excessive power losses at low input voltage result in a poor efficiency. The efficiency for an output power of 980 W was 94% in region R1 and 93% in region R2. The maximum efficiency in both regions was approximately 96% which occurred at an output power of 400 W.
VI. CONCLUSION

In this paper, a three-phase dc–dc isolated converter, referred to as a three-phase flyback push–pull converter, is introduced. Theoretical analysis of the converter is presented and experimental results are used for its validation. The proposed converter merges the main characteristics of push–pull converters with those of three-phase dc–dc converters, resulting in a converter with low power losses, small filter sizes, compact transformer, and high-power processing.

In addition, the operation of the proposed converter in region R2 provides continuous power flow from the input source to the output load that allows filters sizes to be reduced. The prototype efficiencies obtained for regions R1 and R2 show that the proposed converter has good efficiency, considering the power losses in dissipative overvoltage clamp circuits. Therefore, the proposed converter is suitable for many applications, particularly for low-voltage high-power applications such as telecommunications power supply, battery chargers, and renewable power systems.

REFERENCES


