Control and Modeling of Push-Pull Forward Three-Level Converter for Microgrid

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Abstract—Renewable energy sources are widely used in microgrid. Output voltage of them is often low and varies widely. Because diodes in three-level legs in traditional three-level (TL) converter are substituted by MOSFETs, the push-pull forward (PPF) TL converter is very suitable for wide and low input-voltage applications, It can operate at two-level and three-level modes. Modeling of the proposed converter is established. Control of the proposed converter is designed, which contains output voltage loop and current limit loop. In addition, input voltage feedforward is often added to the control system in order to eliminate influences of input voltage on output voltage and inductor current can be eliminated. The divider in the input voltage feedforward is simplified by two piecwise linear functions, which can be easily implemented by operational amplifiers. Therefore, cost can be reduced. Simulation results of a 1-kW PPF TL converter verify the theoretical analysis.

Keywords—three-level converter; DC-DC converter; three-level mode; two-level mode; input voltage feedforward

I. INTRODUCTION

Microgrids support a flexible and efficient electric grid, powered by renewable energy sources, such as fuel cell and PV cell [1]. However, output voltage of renewable energy sources is often low and wide range, such as 20-50 V [2]. The three-level (TL) converter is very suitable for wide input-voltage range application, which can reduce the filter inductor current ripple and solve the problem of high voltage stress of the output rectifier diodes in conventional two-level voltage-source converter [3]-[4]. Nevertheless, conduction loss of the diodes in three-level legs in the traditional TL converter is high in low input-voltage application.

A push-pull forward (PPF) TL converter for low and wide input-voltage range application was proposed in [5]. The freewheeling diodes connected with three-level legs are substituted for the MOSFETs to reduce the conduction loss, so one more control degree of freedom is added compared with the conventional TL converter. Only operating principle is illustrated in [5].

The TL converter for wide input-voltage range application can operate at three-level and two-level modes, two control degrees of freedom exist [6]-[7]. However, modeling of the TL converter is not given. Moreover, three control degrees of freedom exist in the PPF TL converter, so it is necessary to establish modeling of the PPF TL converter and simplify the control at three-level and two-level modes.

In addition, input voltage feedforward is often added to the control system in order to eliminate influences of input voltage on output voltage and output filter inductor current in wide input-voltage range application [8]-[10]. Since the function of the input voltage feedforward is inversely proportional to the input voltage [9], it is complicated to implement. For example, a divider is often needed [10]. Therefore, cost is increased. A linear function of the input voltage is used in [9]. However, the error between the ideal feedforward signal and the linear function is large.

In order to solve the aforementioned problems, control and modeling of the PPF TL converter is proposed. The operating principle is illustrated in Section II. Only two independent control variables are needed, which are the same as the traditional TL converter. Then modeling of the PPF TL converter is established in Section III. Small-signal model of the PPF TL converter is deduced. Section IV gives the control design and example for the PPF TL converter. Output voltage controller, current limit regulator, and input voltage feedforward are designed in detail. Simulation results of a 1-kW PPF TL converter verify the theoretical analysis in Section V. Finally, the concluding remarks are given in Section VI.

II. OPERATING PRINCIPLE

Fig. 1 shows the PPF TL converter, where $R_{ESR}$ is the equivalent series resistance of filter capacitor $(C_f)$, $i_L$ is the current through output filter inductor $(L_f)$, $R_L$ is the load, and $D_{S1} \sim D_{S6}$ are the body diodes of switches $S_1 \sim S_6$, respectively. Fig. 2 gives key waveforms of the proposed converter, where $\tau_{on1}$ is turn-on time of $S_1$ or $S_2$, $\tau_{off}$ is turn-off time of $S_2$ or $S_3$, $T_s$ is the switching period, and $n$ is the windings ratio of secondary and primary windings. There are two methods to control $S_3$ and $S_6$. On the one hand, $S_3$ and $S_6$ can be complementary switched; $S_4$ and $S_5$ can be turned complementary, as shown in Fig. 2. $S_1$ and $S_2$ only sustain input voltage $(U_{in})$ in two-level mode. On the other hand, the drive signal of $S_4$ is that of $S_2$ ANDing with NOT that of $S_1$. Likewise, the drive signal of $S_5$ is that of $S_3$ ANDing with NOT that of $S_4$. $S_2$ and $S_3$ sustain one and one-half of $U_{in}$ in two-level mode. Therefore, the former control method is
selected to reduce the voltage stresses of S₂ and S₃. Detailed operating principle can be referred to [5].

![Fig. 1. Main circuit of the PPF TL converter.](image)

In three-level mode, S₁ and S₄ are PWM controlled. S₂ and S₃ are switched out of phase with a small dead time, so \( t_{on2} \) is equal to the maximum turn-on time (\( t_{on\max} \)), which is constant. The secondary rectified voltage (\( U_{AB} \)) is a three-level waveform. In two-level mode, S₂ and S₃ are PWM controlled. S₁ and S₄ are turned off all the time, whereas S₅ and S₆ are always turned on in this mode. Voltage \( U_{AB} \) is a two-level waveform.

From above analysis and Fig. 2, there are only two independent control variables in the proposed converter, which are the same as that in traditional TL converters.

### III. Modeling of the PPF TL Converter

From Figs. 1 and 2, there are three types of levels of \( U_{AB} \), i.e., 0 level, 1/2 level, and 1 level, shown in Fig. 3. The state-space equations of 0 level, 1/2 level, and 1 level are given in (1), (2), and (3), respectively.

![Fig. 3. Different types of equivalent circuits.](image)

The average state-space equation can be deduced from (1)–(3), as shown in (4), where \( u_f \) is the voltage across \( C_f \), \( d_1 \) is the duty cycle of S₁, i.e., \( t_{on1}/T_s \), and \( d_2 \) is the duty cycle of S₂, i.e., \( t_{on2}/T_s \).

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Relationship between \( u_{cf} \) and output voltage \( u_o \) can be obtained from Fig. 1.

\[
\hat{u}_{cf} = \frac{-1}{C_f R_{ESR}} u_{cf} + \frac{1}{C_f R_{ESR}} u_o .
\]  

From (4) and (5), small-signal model can be deduced, shown in (6) at the top of this page, where \( N(s) = L / C_f (R_{ESR} + R_L) s^2 + (R_L C_f R_{ESR} + L_f) s + R_L . \)

\[
\begin{bmatrix}
    \frac{n[1 + C_f (R_L + R_{ESR}) s]}{N(s)} & \frac{n[1 + C_f (R_L + R_{ESR}) s]}{N(s)} & \frac{n[1 + C_f (R_L + R_{ESR}) s]}{N(s)} \\
    \frac{nR_L (1 + R_{ESR} C_f) s}{N(s)} & \frac{nR_L (1 + R_{ESR} C_f) s}{N(s)} & \frac{nR_L (1 + R_{ESR} C_f) s}{N(s)} \\
    \frac{nR_L (1 + R_{ESR} C_f) s}{N(s)} & \frac{nR_L (1 + R_{ESR} C_f) s}{N(s)} & \frac{nR_L (1 + R_{ESR} C_f) s}{N(s)} \\
\end{bmatrix}
\begin{bmatrix}
    \hat{u}_m \\
    \frac{n}{d_t} \hat{d}_t \\
    \frac{n}{d_t} \hat{d}_2 \\
\end{bmatrix}
\]

IV. CONTROL DESIGN AND EXAMPLE

The whole control system is given in Fig. 4, where only one control chip SG3525 is used.

A. Specifications

The parameters used in the control design are shown in Table I.

<table>
<thead>
<tr>
<th>Specifications</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage ( (U_i) )</td>
<td>20-50 V</td>
</tr>
<tr>
<td>Output voltage ( (U_o) )</td>
<td>380 V</td>
</tr>
<tr>
<td>Output power ( (P_o) )</td>
<td>1 kW</td>
</tr>
<tr>
<td>Switching frequency ( (f_s) )</td>
<td>100 kHz</td>
</tr>
<tr>
<td>Gain of the modulator ( (F_{v1} ) and ( F_{v2} )</td>
<td>1/2.4</td>
</tr>
<tr>
<td>Feedback coefficient of output voltage ( (K_v) )</td>
<td>5/380</td>
</tr>
<tr>
<td>Feedback coefficient of input voltage ( (K_i) )</td>
<td>0.1</td>
</tr>
<tr>
<td>Secondary to primary turns ratio of the transformer ( (m) )</td>
<td>20.5</td>
</tr>
<tr>
<td>Output filter capacitor ( (C_f) )</td>
<td>470 ( \mu )F</td>
</tr>
<tr>
<td>ESR of ( C_f ) and ( R_{ESR} )</td>
<td>138 m( \Omega )</td>
</tr>
<tr>
<td>Output filter inductor ( (L_f) )</td>
<td>640 ( \mu )H</td>
</tr>
</tbody>
</table>

From (6) and (5), small-signal model can be deduced, shown in (6) at the top of this page, where \( N(s) = L / C_f (R_{ESR} + R_L) s^2 + (R_L C_f R_{ESR} + L_f) s + R_L . \)

\[
G_{out}(s) = \frac{\hat{u}_m}{d} = \frac{nR_L (1 + R_{ESR} C_f) s}{L / C_f (R_{ESR} + R_L) s^2 + (R_L C_f R_{ESR} + L_f) s + R_L} .
\]
If \( F_{m1} \) and \( F_{m2} \) are set to the same, i.e., \( F_{m} \), by adjusting the amplitude of the carrier waveforms, control block diagram in both modes can be simplified to Fig. 6. Therefore, performance of the system at both three-level and two-level modes is the same by using the same controller. In addition, control design of the system can be simplified.

From Fig. 6 and (7), the open loop transfer function under voltage controlled mode can be obtained as

\[
G_v(s) = G_{vd}(s) \times F_m \times K_v
\]

\[
= \frac{nR_c(1+R_{ESR}C, s)U_{in}}{L_c/C_f(R_{ESR}+R_c)s^2+(R_c/C_f+L_p)s+R_p} \times \frac{1}{2.4 \times 380}
\]  

(8)

The crossover frequency is set to 5 kHz at \( U_{in} = 50 \text{ V} \). From (8), about a zero of 2.5 kHz is generated by \( C_f \) and \( R_{ESR} \), which is close to the crossover frequency during the whole input-voltage range. Therefore, a type III compensation network is adopted to eliminate the zero, as shown in Fig. 7, where \( U_r \) is the output of voltage regulator. The transfer function of the controller is given in (9).

\[
G_{vd}(s) = \frac{(1+R_cC_f)[1+s(R_c+R_p)C_f]}{sR_c(C_f+C_p)(1+s+R_c/C_f)(1+sR_c/C_f)}
\]

(9)

From (9), the following two zeros and three poles can be deduced.

\[
f_{z1} = \frac{1}{2\pi R_cC_f}
\]

(10)

\[
f_{z2} = \frac{1}{2\pi(R_c+R_p)C_f}
\]

(11)

\[
f_{p1} = 0
\]

(12)

\[
f_{p2} = \frac{1}{2\pi R_cC_f}
\]

(13)

The zeros and poles are set as follows:

1) the zeros \( f_{z1} \) and \( f_{z2} \) are set as 250 Hz to maintain the phase margin being larger than 45°;

2) the pole \( f_{p2} \) is equal to the zero of 2.5 kHz generated by \( C_f \) and \( R_{ESR} \);

3) the pole \( f_{p3} \) is set as half of the switching frequency to attenuate high frequency noise, i.e., 50 kHz.

Therefore, parameters \( C_f \), \( C_p \), \( R_f \), \( R_p \), and \( R_c \) can be calculated as 64 nF, 320 pF, 152 nF, 1.88 kΩ, 10 kΩ, 427 Ω, and 1.88 kΩ. In addition, the bode diagram of output voltage loop can be obtained, given in Fig. 8. From Fig. 8, the crossover frequency is between 2 kHz and 5 kHz under the input voltage range of 20-50 V. Moreover, the phase margin is between 74.47° and 78.94°.

C. Current Limit Controller

In order to limit the output filter inductor current during startup or overload, a current limit loop should be added limiting the output filter inductor current to three times the rated load current.

Since the transfer functions of both duty ratios to inductor current are also the same, i.e., \( G_{cd}(s) = G_{vd}(s) = G_{al}(s) \) from (6), the simplified control block diagram of the current limit loop can be obtained like output voltage loop, as shown in Fig. 9, where \( G_{cd}(s) \) is the transfer function of current regulator.

\[
G_{al}(s) = \frac{i_{Ls}}{d} = \frac{n[1+s(C_fR_{ESR}+R_c)s^2+(R_cC_fR_{ESR}+L_p)s+R_p]}{L_c/C_f(R_{ESR}+R_c)s^2+(R_cC_fR_{ESR}+L_p)s+R_p}
\]

(15)

From Fig. 9 and (15), the open loop transfer function under current limit controlled mode can be expressed as
The crossover frequency is set to 20 kHz at $U_{in} = 50$ V. From (16), about a zero of 2.5 Hz is generated by $C_f$, $R_L$, and $E_{SR}$, which can improve the phase margin of the system. Thus, a type II compensation network is used, as shown in Fig. 10, where $U_r$ is the output of current limit regulator. The transfer function of the controller is given in (17).

$$G_c(s) = G_d(s) \times F_{in} \times K_c$$

$$= \frac{n[1+C_f(R_{ESR}+R_i)s]U_{in}}{L_fC_f(R_{ESR}+R_i)s^2+(R_iC_fR_{ESR}+L_f)s+R_i} \times \frac{0.6}{2.4}$$

(16)

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$$G_c(s) = \frac{1+R_iC_f s}{sR_i(C_1+C_f)(1+sR_iC_f)}$$

(17)

From (17), the following one zero and two poles can be obtained.

$$f_{z1} = \frac{1}{2\pi R_iC_f}$$

(18)

$$f_{p1} = 0$$

(19)

$$f_{p2} = \frac{1}{2\pi R_iC_f}$$

(20)

The zero and pole are set as follows:

1) the zero $f_{z1}$ is set as 1 kHz to maintain the phase margin being larger than 45°;

2) the pole $f_{p2}$ is the same as $f_{p1}$ to attenuate high frequency noise.

Therefore, parameters $C_f$, $C_i$, $R_f$, $R_{ESR}$, and $R_i$ can be calculated as 64 nF, 922 pF, 10 kΩ, 3.45 kΩ, and 10 kΩ. In addition, the bode diagram of current limit loop can be obtained, given in Fig. 11. From Fig. 11, the crossover frequency is between 8.6 kHz and 20 kHz under the input voltage range of 20-50 V. Moreover, the phase margin is between 65.4° and 73.8°.

D. Input Voltage Feedforward

From (6), the input-to-output voltages and input-voltage-to-inductor current transfer functions can be obtained as

$$G_1(s) = \frac{\hat{d}_i}{U_{in}} = \frac{nR_i(1+R_{ESR}C_f)s(d_i+d_j)}{L_fC_f(R_{ESR}+R_i)s^2+(R_iC_fR_{ESR}+L_f)s+R_i}$$

(21)

$$G_2(s) = \frac{\hat{i}_{L}}{U_{in}} = \frac{n[1+C_f(R_i+R_{ESR})s](d_i+d_j)}{L_fC_f(R_{ESR}+R_i)s^2+(R_iC_fR_{ESR}+L_f)s+R_i}$$

(22)

As the change of input voltage affects the output voltage and inductor current from (21) and (22), respectively, input voltage feedforward should be added into the control system. The quiescent operation point of reference voltage ($U_{ff_th}$) should be added into the common point of the output voltage and inductor current regulators, given in (23).

$$U_{ff_th} = 5.7 - \frac{94.655}{U_{in}}.$$  

(23)

From (23), the feedforward voltage $U_{ff_th}$ is inversely proportional to the input voltage, so a divider should be used and cost will be increased. In order to reduce the cost and improve the accuracy, the approximate feedforward voltage ($U_{ff_ap}$) can be realized by two piecewise linear functions.

$$U_{ff_ap} = \min \left\{ \frac{0.166U_{in}-2.308}{0.0585U_{in}+0.87} \right\}.$$  

(24)

From (24), the voltage $U_{ff_ap}$ is the smaller value in the two piecewise linear functions, which can be easily implemented by analog devices. Fig. 12 gives the implementation of $U_{ff_ap}$, where $U_{r}$ is the reference output voltage of SG3525, i.e., 5.1 V and $U_{inf}$ is the feedback input voltage. Equations (23) and (24) can be plotted in Fig. 13. The error between (23) and (24) related to (23), i.e., $e\%$, is given in (25) and plotted in Fig. 13.

$$e\% = \frac{U_{ff_ap} - U_{ff_th}}{U_{ff_th}} \times 100\%.$$  

(25)

From Fig. 13, the error percentage $e\%$ is less than 5%, so the approximation accuracy is high.
A 1-kW PPF TL converter is simulated to verify the theoretical analysis using the SABER software with the parameters given in Section IV. An ideal DC voltage source and an ideal resistive load are adopted in the simulation.

Fig. 14 shows the simulation results of input voltage feedforward. From Fig. 14, voltage $U_{\text{ff}}$ very approximates theoretical value $U_{\text{ff}}$. The error percentage is less than 5%. Therefore, simulation results verify the theoretical analysis.

Simulation results under startup is given in Fig. 15, where $U_r$ is the reference voltage used to compare with the carrier voltage. As can be seen from Fig. 15, the proposed converter operates at current-limit control mode during startup, and the inductor current is limited to three times the rated output current. Therefore, the output voltage increases linearly. After the output voltage reaches the rated value, i.e., 380 V, the proposed converter operates at output-voltage control mode. After that, the system transfers to steady state gradually.

Fig. 16 shows the simulation results under load and input voltage variations without input voltage feedforward. When the input voltage changes suddenly, current $i_L$ has large inrush current and the output voltage is largely affected, shown in the left oval dashed line in Fig. 16. However, when the load changes from full to half loads and from half to full loads, the output voltage maintains constant, seen from the right oval dashed line in Fig. 16. When the input voltage changes from...
30 V to 50 V, the maximum of $U_{AB}$ is reduced. It means that three-level mode transfers to two-level mode. Therefore, the voltage stress of the rectifier diodes is reduced compared with the traditional two-level converter.

Fig. 17 gives the simulation results under load and input voltage variations with input voltage feedforward. As can be seen from Fig. 17, the feedforward voltage $U_{ff_{ap}}$ changes with the input voltage, and current $i_L$ is less affected and the output voltage keeps constant when input voltage changes, shown in the left oval dashed line in Fig. 17. Therefore, influences of input voltage on the output voltage and inductor current can be eliminated. When load current changes, the output voltage keeps constant, seen from the right oval dashed line in Fig. 17. Therefore, simulation results verify the theoretical analysis.

Fig. 17. Simulation results under load and input voltage variations with input voltage feedforward.

VI. CONCLUSIONS

This paper has proposed control and modeling of the PPF TL converter. The proposed converter can operate at both two-level and three-level modes, so it is suitable for wide input-voltage range application. Control degrees of freedom are the same as that in the traditional TL converter by adopting the proposed control method. Small-signal model of the proposed converter is established. Control block diagrams in both three-level and two-level modes are the same by maintaining the same gain of the modulator at both modes, so the control can be simplified, which is composed of output voltage loop, current limit loop, and input voltage feedforward. The output voltage loop is adopted to maintain the output voltage constant. The current limit loop is used to limit the output filter inductor current during startup or overload, while the input voltage feedforward is adopted to cancel the influences of input voltage on the output voltage and inductor current. The feedforward input voltage is approximated by two piecewise linear functions, which limit the error percent within 5%. Therefore, cost can be reduced since the divider is not needed. The concept also can be applied to other linearization approximation of the nonlinear function.

REFERENCES