Grid-friendly Control Strategy with Dual Primary Side Series-connected Winding Transformers

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Abstract

High-power three-level voltage source converters are widely used in high performance ac drive systems. In some ultra-power occasions, it is very important to reduce the harmonics in grid side by using multiple rectifying. A combined harmonic elimination method, which includes the dual primary side series-connected winding transformer and SHEPWM modulation, is proposed to eliminate low-order current harmonics on the primary side and the secondary side of transformers. Based on the analysis of the harmonic influence caused by dead time and dc magnetic-bias, a synthetic compensation control strategy is presented to minimize the grid side harmonics in dual primary side series-connected winding transformer application. Both simulation and experimental results demonstrate that the proposed control strategy can significantly reduce the converter input current harmonics, and eliminates the dc magnetic-bias on the transformer.

Key words: Dead time, Harmonic, Magnetic-bias, Series-connected, SHEPWM modulation, Transformer

I. INTRODUCTION

Active rectifiers which can feedback the breaking energy to the grid are commonly used in high power drive systems such as metallurgical rolling mill, mine, etc.. In medium voltage applications, active rectifiers often adopt three-level neutral-point-clamped topology with the merit of high voltage and low harmonics. However, in this case, the converter must be controlled with low switching frequency considering the switching losses. The reduction of current harmonics with low switching frequency is an important issue for the design of high power converter systems.

The modulation methods and control strategies [1-8] of three-level converters have long been a hot spot for researching. In order to reduce the grid side harmonics as much as possible, many scholars have made a lot of fruitful work. Most of them focus on modulation strategy and system design, and make a comprehensive optimization. The specific harmonics eliminating modulation strategy (SHEPWM) is a preferable modulation method at low switching frequency with lower grid current THD. A kind of three-level SHEPWM modulation strategy has been presented in [9], which can realize the balance of the neutral voltage. A certain minimum harmonic modulation strategy has been proposed in [10], which can achieve the most optimal harmonic modulation ratio in the whole scope. The above papers are focus on the theory analysis of single aspect of SHEPWM modulation strategy under ideal conditions.

In practical applications, for the design of grid side system, the structure of the transformer also directly affects the grid side harmonics. For example, the split type pulse transformer, which is widely used in urban rail transit power system, can effectively eliminate the primary side current harmonics of the transformer. [11-12] proposed a three-level harmonic optimal modulation strategy based on a 12 pulse rectifier transformer cooperated with a dual main circuit structure. It can realize low harmonics in transformer primary side with low switching frequency. A multiple tandem transformer winding structure has been described in [13-14], which is mainly used for HVDC power system as well as the STATCOM. However, the working principle, advantages and output current harmonic characteristics on the system of grid side have no in-depth analysis.

The dead time of the high power device will influence the control precision and voltage harmonics distribution in converter output valve side fundamental wave. Besides some non-ideal factors, such as the influence of grid voltage harmonic, the difference of the device switch characteristic and control system error and so on, the magnetic-bias in
transformer windings, which is caused by the DC component of the PWM converter output voltage [15-18], will lead to the grid side current distortion, and also take an effect on grid side current harmonic. [19-21] discussed the influence of dead time and compensation strategies where the SVPWM modulation was used in detail. [22-24] put forward some control strategies for the inhibition of magnetic-bias, but they were focused on the carrier modulation and square wave modulation strategies. In low switching frequency applications, the dead time effect of SHEPWM for three-level converters and the corresponding magnetic-bias compensation methodology have not been released in literatures yet.

At first, a combined harmonic elimination method, which includes the dual primary side series-connected winding transformer and SHEPWM modulation, is introduced to eliminate low-order harmonics of primary side and secondary side current of transformer in this paper. Furthermore, influences of dead time and magnetic-bias are analyzed in principle. Then, a synthetic compensation control strategy (SHEPWM-SC) is proposed to minimize the grid side harmonics in dual primary side series-connected winding transformer application. Finally, experiments are conducted in the dual primary series-connected winding transformers based three-level IGCT converters. Both simulation and experimental results show that the proposed control strategy can eliminate $12k \pm 1 (k = \frac{r-1}{2})$ order harmonics and below; and further reduce grid harmonics by dead time and magnetic-bias compensation.

II. COMBINED HARMONIC ELIMINATION

As an important system component, the structure and design parameters of transformer will directly decide the converter performance level in the system design. Multiple primary side series-connected winding transformer as a special transformer structure is mainly used in high power systems [25-27]. Compared with multiple split winding transformer, a lower short-circuit impedance of the multiple primary side series-connected winding transformer can be designed but having the same current THD. So the transformer size can be decreased. The secondary side current of spilt winding transformer contains abundant harmonics when connected to high power converter. However, the primary current harmonic can be in low level by carrier phase-shifting or transformer wind-shifting. The structure of primary side series-connected winding transformer determines that the current harmonic level is the same in both winding, only the amplitude is different. After using a certain control strategy, the current harmonic in both sides can be in low level. Besides, the waveform factor will be optimized and it can take full advantage of the capacity.

The schematic diagram of dual series-connected winding transformer is shown in Fig.1. It can be seen that the primary and secondary windings of dual transformers are connected in star (Y) and delta (Δ) forms. The dual secondary windings are connected separately to the as sides of PWM converters. The transformer leakage inductors are used as the input inductors. The voltages/currents in the dual secondary windings have 30 degree phase shift.

In Fig.1, a.b.c are defined as the three input phases of the converter respectively. $V_{1ab}$, $V_{1bc}$ is line voltage in grid side. $V_{2ab}$, $V_{2bc}$ is the line voltage of secondary Y-connected winding, $V_{3ab}$ $V_{3bc}$ is the line voltage of secondary Δ-connected winding. $U_{1ab}$, $U_{1bc}$ is the primary side line voltage inducted from secondary winding. $i_{1a}$, $i_{1b}$, $i_{1c}$ is Phase A, B, C primary side current, $i_{2a}$ is Phase A current of secondary Y-connected winding, $i_{3a}$ is Phase A current of secondary Δ-connected winding. $L_{f}$ and $r_{f}$ are the equivalent impedance of transformer winding.

Let turn ratios of the primary winding and the secondary winding are $k_{1}$ and $k_{2}$, to guarantee the consistency of the two secondary voltages, the turn ratios of the two windings satisfy the relationship shown as below

$$k_{1} = \frac{1}{\sqrt{3}} = k_{2} = f.$$  (1)

Because of the special structure of transformer, there is a strong coupling relationship within the windings. In order to keep current-sharing and voltage-sharing within the windings, the output voltages of PWM rectifiers connected to the dual secondary windings have the same magnitude but 30 degree phase shift. Analyzing the dual converters output line voltages, the inverter output voltage can be written as follow

$$V_{2ab} = \sum_{n=1,5,7,11,13,\ldots}^{\infty} V_{n} \sin(n(\alpha + \varphi_{n}) + \varphi_{n}),$$

$$V_{3ab} = \sum_{n=1,5,7,11,13,\ldots}^{\infty} V_{n} \sin(n(\alpha + \varphi_{n}) + \varphi_{n}).$$

where, $V_{n}$ is the amplitude of each harmonic in line voltage, $\varphi_{n}$ is the initial phase of each harmonic in Y-connected winding line voltage.

So, it can be got

$$U_{ab} = 2\sqrt{3}f[V_{1} \sin(\alpha + \varphi_{1}) + \sum_{n=1,2,\ldots}^{r_{f} \geq 1} V_{n} \sin(n(\alpha + \varphi_{n}))],$$  (2)

where, $k=1,2,3\ldots$, $U_{ab}$ is the primary side Line AB voltage inducted from secondary winding.
From the Eqn.2, the primary side line voltage $U_{ab}$ inducted from the secondary PWM voltage of transformer only contains $12k \pm 1 \ (k=1,2,3\cdots)$ order harmonic components. When using the SHEPWM, it can be concluded that

$$I \sum_{k=1}^{N} (-1)^{k+1} \cos \alpha_k = M$$

$$I \sum_{k=1}^{N} (-1)^{k+1} \cos \alpha_k = 0 \quad n = 11, 13, 23, 25 \cdots$$

where, $M$ is the modulation ratio, $\alpha_k$ is the switching angle in a quarter of switching cycle.

If the switching frequency is $f$ (T is odd) times of power frequency (50Hz), the $12k \pm 1 \ (k=1,2,3\cdots\frac{r-1}{2})$ harmonics can be eliminated in Line AB voltage, so

$$U_{ab} = \frac{2\sqrt{3} f}{\pi} \left[ 2 \sum_{n=1}^{\infty} \hat{V}_n \sin(n+1) \left( \frac{\varphi_1}{2} \right) + \sum_{n=12k \pm 1}^{\infty} \hat{V}_n \sin(n+1) \left( \frac{\varphi_2}{2} \right) \right] \ (4)$$

where, $k = \frac{T+1}{2}, \frac{T+1}{2} + 1, \frac{T+1}{2} + 2 \cdots \cdots, U_{dc}$ is the DC link voltage.

Line BC voltage $U_{bc}$ inducted from secondary winding is the same as Eqn.4.

For the primary current, equation can be written as

\[
\begin{align*}
\frac{di_{1a}}{dt} &= -\frac{r_i}{L_1} i_{1a} + \frac{2}{3L_4} (V_{1ab} - U_{ab}) + \frac{1}{3L_1} (V_{1bc} - U_{bc}) \\
\frac{di_{1b}}{dt} &= -\frac{r_i}{L_1} i_{1b} - \frac{1}{3L_4} (V_{1ab} - U_{ab}) + \frac{1}{3L_1} (V_{1bc} - U_{bc})
\end{align*}
\]

where, $i_{1a}$ is phase A primary side current, $i_{1b}$ is phase B primary side current, $V_{1ab}$ is Line AB voltage in grid side, $V_{1bc}$ is Line BC voltage in grid side.

If the grid voltage is ideal, the primary side phase current $i_{1i}$ from equations above, only contains $12k \pm 1 \ (k = \frac{T+1}{2}, \frac{T+1}{2} + 1 \cdots \cdots)$ harmonics, and the secondary phase current can be written as

\[
\begin{align*}
i_{2a} &= k_1 i_{1a} \\
i_{3a} &= k_2 i_{1a}
\end{align*}
\]

Therefore, the secondary winding phase current $i_{2a}$ and $i_{3a}$ of transformer contain $12k \pm 1 \ (k = \frac{T+1}{2}, \frac{T+1}{2} + 1 \cdots \cdots)$ order harmonics, from which the current harmonics in each side of transformer can be eliminated. It improves the waveform coefficient of converter input current which is more-friendly to the grid.

III. DEAD TIME AND MAGNETIC-BIAS

A. Influence of Dead Time

In order to prevent shoot-through phenomenon, a dead time $T$ must be inserted in the complementary pulse trigger signal to guarantee that the same phase switch can reliably turn-off. Due to the characteristic of turn-on and turn-off snubber circuit, high-power semiconductor devices (such as IGCT) need to set up longer dead time. Therefore, the effect of dead time is still severe though switching frequency is low.

High-power electric drive converters always work in unit power factor. When the motor operates as motor mode, the current and voltage have a same phase in the grid side. When the motor operates as generator mode, the current and voltage have an opposite phase. When using SHEPWM, dead time effects on the phase voltage of the converter are shown in Fig.2, respectively. As is shown in the Fig.2 (a), the system works as a rectifier, the dead time will lead to the pulse width of the ideal phase voltage level wider. From Fig.2 (b), the system works as an inverter, and the dead time will lead to the pulse width of the ideal phase voltage level narrow.

Fig.2. The effect of dead time on the phase voltage

The actual output voltage of the converter is shown as follow

\[
U'_{po} = \begin{cases} U_{po} + U_{gain} & \text{in the same phase} \\ U_{po} + U_{lose} & \text{in the reverse phase} \end{cases}
\]

where, $U'_{po}$ is actual phase voltage, $U_{po}$ is ideal phase voltage, $U_{gain}$ is voltage gain caused by dead time, $U_{lose}$ is voltage loss caused by dead time, $i$ is the input current in grid side.

Under the condition of the rectifier mode, taking the impact of dead time as an example, after Fourier transform for the periodic function, there are

\[
U_{gain}(t) = \frac{a_0}{2} + \sum_{n=1}^{\infty} (a_n \cos n\varphi + b_n \sin n\varphi)
\]

where,
Due to $U_{gain}(t)$ is an odd function, there is no even-order harmonic components on it. For SHEPWM, substituting Eqn.3 into Eqn.5, harmonic amplitudes can be calculated as

$$
\begin{align*}
    a_n &= -\frac{u_{dc}}{\pi n} [\sin n\pi \sum_{k=1}^{N} (-1)^{k+1} \cos n\alpha_k - (1 - \cos n\pi) \sum_{k=1}^{N} \sin n\alpha_k] \\
    b_n &= -\frac{u_{dc}}{n} \frac{(1 - \cos n\pi) \sum_{k=1}^{N} (-1)^{k+1} \cos n\alpha_k - \sin n\pi \sum_{k=1}^{N} \sin n\alpha_k}{\sin n\pi} 
\end{align*}
$$

(5)

Harmonic phase is given by

$$
\arctan \frac{a_n}{b_n} = \arctan \frac{1 - \cos n\pi}{\sin n\pi}
$$

If the switch frequency is 350Hz, it works in normal modulation scope ($0.675<M<0.934$). Let $n=11, 13, 23, 25, 35, 37$ in Eqn.5, the converter is working in rectifier state, the dead time effects on each harmonic are shown in Fig.3.

![Fig.3](image1)

(a) p/M change curve in 60us dead time

![Fig.3](image2)

(b) 37th harmonic/M change curve in different dead time

Fig.3. The dead time effects

In Fig.3 (a), $p$ is the proportion of harmonic amplitude and half voltage, $M$ is the modulation ratio. As the dead time is fixed, the dead time’s impacts of each harmonic’s amplitude of phase voltage are not identical. In the normal modulation ratio, the 23rd harmonic which should has been eliminated is up to 6% of a half bridge voltage amplitude, 37th harmonic reaches 4%, and the other all increase in different degrees.

By Eqn.6, along with the rising of the dead time, harmonic amplitudes in every phase voltage have been increased. Taking the 37th harmonic for example, as shown in Fig.3 (b), harmonic amplitude increases with the dead time. If the dead time reaches 90us, the 37th harmonic’s amplitude can reach 6.5% of half bridge voltage, which will result in the increasing of current harmonic on grid side. B. Influence of Magnetic-bias

For high-power three-level converters, the input inductor of grid side is generally equivalent by the leakage inductor of transformers. A single-phase model of transformer is shown in Fig.4. The transformer secondary side usually directly connects to the converter valve side, so PWM voltage directly imposes on the transformer secondary side. Due to sampling bias of control system, delay characteristics, power grid voltage harmonics and semiconductor voltage drop, PWM voltage in valve side will contain dc component which will make the transformer core saturated over a long period, so as to make the current distortion on transformer secondary side.

![Fig.4](image3)

Fig.4. The single phase model of transformer

There, $L_m$ and $r_m$ are the magnetizing impedance of transformer.

Magnetic induction intensity $B(t)$ can be written as

$$
B(t) = \int \frac{U_{1}}{N_{S}} dt,
$$

where, $U_1$ is the voltage in valve side of converter, $N_S$ is the number of turns in the secondary side of transformer, $S$ is the effective cross-sectional area of transformer core.

Let

$$
U_1 = U_{AC} + U_{DC},
$$

where, $U_{AC}$ represents the AC component of converter voltage in valve side, $U_{DC}$ is the DC component.

Hence

$$
B(t) = \int \frac{U_{AC}}{N_{S}} dt + \int \frac{U_{DC}}{N_{S}} dt
$$

(7)

From Eqn.7, when $U_{DC}=0$, the V-S area is equal to the forward and reverse pulse as well as the maximum operating magnetic induction intensity, as shown in Fig.5. The magnetic core operating point moves along the hysteresis loop symmetrically and without biasing. In some cases, if $U_{DC}>0$, the V-S area of positive pulse is larger than the reverse pulse as well as the maximum magnetic induction intensity, the hysteresis loop in the whole pulse cycle transfers to the first quadrant which results in biasing.
In the next cycle, if the time difference no longer increases, the bias does not increase either. But it cannot be eliminated automatically. However, if the magnetic-bias increased sequentially, that will lead to magnetic core saturate, the nonlinearity of the magnetic curve increase. The magnetizing current grows rapidly and leads to transformer saturated eventually, which causes a shape rise in current through the transformer. This part of shape increased current are superimposed on the secondary side of transformer current which will cause the current distortion and deteriorate the grid side harmonic. The analysis of dc component is the same, and no more details are given.

In three phase transformer, any dc component existing in line voltage will lead to magnetic-bias corresponding to the two phase of transformer and cause serious current distortion. It shows the waveform of grid side input current when the magnetic-bias saturation occurs in Fig.6 (a). The converter three-phase current distortion occurs periodically at the wave crest and trough, where a sharp current slew rate di/dt will result in overvoltage and damage the devices severely in the extreme situation.

![Fig.5 Transformer dc bias magnetic saturated](image)

IV. PROPOSED CONTROL STRATEGY

Because of the special structure of transformer, there is a strong coupling relationship within windings. It must keep voltage-sharing and current-sharing within windings. Especially in the high-voltage case, the dead time and magnetic-bias will have more significant effects on the system. Therefore, the traditional multiple current loop and inverter control will not be suit for this system. In this section, a synthetic compensation control strategy (SHEPWM-SC), which includes the combined harmonic elimination, dead time and magnetic-bias compensation, is proposed to minimize the harmonics with dual Series-connected Winding Transformers. The structure of proposed control strategy is shown in Fig.7.

In an ideal transformer model, the secondary side phase current is proportional to the primary side phase current, any one current change will affect the rest of two currents. Unlike the traditional multiple split transformer winding which is decoupling, secondary windings of dual primary side series-connected winding transformer have a strong coupling relationship. It must ensure that the dual winding inducted to the primary voltage and current are strict equilibrium. Otherwise, it will cause the primary winding voltage unbalanced. So the traditional utility voltage outer loop and independent current inner loop control are no longer feasible. Under an ideal condition, the secondary side current $i_2$ and $i_3$ are the same controlled quantity, taking dual current into three-phase average $dq$ transformation, and make a feedback control with instantaneous average current, the strategy transform inner current closed-loop output quantity into three-phase modulation wave, after the phase shift, dual grid side converter module control pulse is generated, it can ensure strict equalizing between dual windings. In order to reduce the grid side of harmonics in voltage and current feedback adverse influence on control performance, sampling signal of voltage and current on the grid side must get through a filter. Through the control diagram, the dc voltage is controlled by the active current $I_d$ and the power factor is controlled by the reactive current $I_w$ when setting the reactive current is 0, the system will work in unit power factor.

![Fig.6 The transformer magnetic-bias saturated](image)
As mentioned before, the primary side and secondary side currents only contain $12k \pm 1 (k = \frac{n+1}{2}, \frac{n+1}{2} + 1, \ldots)$ order harmonics by using the combined harmonic elimination in this application. As for dead time and magnetic-bias compensations, because of the limitation of length, there are no more details here.

V. SIMULATION RESULTS

A simulation model is given to validate the proposed control strategy in Matlab. Transformer parameters are presented as follows: capacity is 13.6MVA, dual primary side series-connected winding, valve side windings are Y-connected and D-connected with their short-circuit impedances 12%, and valve-side windings output voltage is 3.16kV. The dc link voltage is 4840V. The switching frequency is 350Hz and dead-time is 60us (depended on the characteristics of IGCT and snubber circuit).

The three phases output current waveforms of Y-connected winding with SPWM, SHEPWM-DC (SHEPWM with magnetic-bias compensation), SHEPWM-SC are shown in Fig. 8. Accordingly, their $12n \pm 1 (n=1,2,3)$ harmonics value are listed in Table I. It shows that SPWM technique has resulted in 11th and 13rd output current harmonic values are far greater than SHEPWM-DC and SHEPWM-SC. The amplitude of each harmonic component of SHEPWM-SC is smaller than others.

**TABLE I**

<table>
<thead>
<tr>
<th>OUTPUT CURRENT HARMONICS VALUE OF Y WINDING</th>
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<tbody>
<tr>
<td>(FUNDAMENTAL WAVEFORM VALUE 2600A) WITH SPWM, SHEPWM-DC, SHEPWM-SC</td>
</tr>
</tbody>
</table>
As known, the control system sampling errors, data transmission delay, device characteristics and other factors will lead to a dc offset in the converter output voltage. To verify the influence of the dc offset on the transformer magnetic-bias hardware in the loop simulations based on dSPACE are conducted.

Fig.9 shows the diagram of three-level medium voltage transmission system hardware-in-loop setup. It has been built up in crate, and DSP and FPGA have been used for achieving control algorithm. Based on dSPACE, the transformer, electric motor, converter, power grid, and complete set of transmission system mathematical models can be built up.

<table>
<thead>
<tr>
<th>order</th>
<th>harmonic amplitude(A)</th>
<th>harmonic amplitude(A)</th>
<th>harmonic amplitude(A)</th>
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<td>11</td>
<td>102.73</td>
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<td>3.23</td>
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<td>13</td>
<td>83.56</td>
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<td>23</td>
<td>22.15</td>
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<td>1.53</td>
</tr>
<tr>
<td>37</td>
<td>5.08</td>
<td>3.94</td>
<td>2.31</td>
</tr>
</tbody>
</table>

Fig.9 Three-level medium voltage transmission system hardware-in-loop setup

In Fig.10, magnetic flux-exciting current of transformer characteristic curve is shown. When magnetic flux is located in the rated area, exciting current of transformer is far below the rated current. Once magnetic flux is saturated, exciting current of transformer would increase sharply which generates output current distortion.

Fig.10 Magnetic flux-exciting current of transformer

Fig.11 (a) shows Y-connected winding exciting current waveforms using SHEPWM without compensation technique. The winding exciting current mutates because of flux saturation, and its peak value is up to 300A. While the output current is distorted due to the exciting current effect as shown in Fig. 11 (b).

With SHEPWM-SC technique, Y-connected winding exciting current peak value is 2.5A, 0.1% of the rated value without mutation in Fig.12 (a). And its output current has no distortion in Fig. 12 (b). The transformer operates in regular scope.

Contrastive analysis of Fig.11 and Fig.12, it obtains the conclusion that the dc component of PWM voltage has been restrained by using SHEPWM-SC scheme. The transformer magnetic flux will not be saturated and output current has no distortion.
VI. EXPERIMENTAL RESULTS

To validate the control strategy, experiments have been carried out in a practical engineering application (Fig. 13). Two three-level IGCT converters with a common dc bus drive two synchronous motors respectively. The motor rated power is 5MW and 200% overload.

Main technical parameters of the application as followers
1) transformer output voltage: 3.16kV
2) switching frequency: 350Hz
3) dc link voltage: 4840V
4) dead-time: 60us
5) short circuit impedance of transformer: 7.5%

Fig. 14 phase A output current of secondary windings with SPWM

Where, \( i_{a1} \) is Y-connected winding Phase A output current, \( i_{a2} \) is \( \Delta \)-connected winding Phase A output current. Fig. 14 shows output current of transformer secondary windings with SPWM in heavy load case. Phase A current of \( \Delta \)-connected winding is 30 degrees lead than Phase A current of Y-connected winding. Current peak value is 2000A. According to Y-connected winding current \( i_{a1}, \) 400A ripple in current peak will affect the capacity utilization of the converter in heavy load condition.

Fig. 15 shows the transformer output current is badly mutated in no-load and heavy load by using SHEPWM without compensation. And ripple peak value is up to 350A. In view of three-phase equilibrium, other two phases are affected as well. \( i_b \) and \( i_c \) in Fig. 15 (a) mutate simultaneously, which
indicates the dc component of output line voltage $U_{dc}$ has generated magnetic-bias in Phase B and C. Therefore, the exciting current increases sharply. In Fig. 15(b), Phase C current mutation emerges out at the fundamental wave peak point. The convert over-current protection would be triggered probably is adverse factor for convert reliable and safety operation in heavy load condition.

Output current of Y-connected winding with SHEPWM-DC control scheme is shown in Fig.15 (c). The peak value of input current of converter reaches 2500A. There is not mutation in heavy load case. Compared with Fig.15 (b), the wave coefficient has been improved, and current ripple has been decreased as well.

Output current of Y-connected winding with SHEPWM-SC control scheme is shown in Fig.10 (d). $12\pi \pm 1(n=1, 2, 3)$ order current harmonic amplitudes with SPWM, SHEPWM-DC, SHEPWM-SC are shown in Table II. With SPWM, the convert input current $11th$, $13th$, $23rd$ harmonics are greater than SHEPWM-DC’s and SHEPWM-SC’s. Its $11th$ harmonic amplitude is 126.55A which is 10.1 times of SHEPWM-DC and 18.6 times of SHEPWM-SC. The size of each harmonic component with SHEPWM-SC is smallest than others. The similar results have been shown in simulation.

<table>
<thead>
<tr>
<th>Harmonics</th>
<th>SPWM</th>
<th>SHEPWM-DC</th>
<th>SHEPWM-SC</th>
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<tr>
<td>11</td>
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<td>12.53</td>
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<td>37</td>
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<td>4.196</td>
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</table>

VII. CONCLUSIONS

In medium voltage high power converter, grid side harmonics are affected by some key factors. Based on analysis of the factors, a grid-friendly control strategy has been proposed in this paper.

1) A combined harmonic elimination method, which includes the dual primary side series-connected winding transformer and SHEPWM scheme, is introduced to eliminate $12k \pm 1(k = \frac{T-1}{2})$ order harmonics and below of the transformer primary side and secondary side current.

2) A synthetic compensation control strategy for three-level medium-voltage high-power converters is proposed to eliminate the grid side current distortion resulted from dead time and dc magnetic-bias in the low switching frequency application.

Simulation and experimental results have validated the control strategy. Compared with SPWM method, using the proposed SHEPWM-SC method, the $11th$, $13th$, and $23rd$ current harmonics are reduced to 5.38%, 2.85%, 20.31%. The secondary side line current THD is improved from 8.62% to 2.21% at rated operating condition.

REFERENCES


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