

一种用于推挽式电压型逆变器的 低损耗无源吸收电路

伍家驹¹, 杉本英彦²

(1. 南昌航空工业学院无损检测技术教育部重点实验室, 江西省 南昌市 330034; 2. 福井大学, 日本 福井)

Low Loss Passive Snubber in Push-pull Voltage Source Inverter

WU Jia-ju¹, HIDEHIKO Sugimoto²

(1. Nondestructive Test Key Laboratory of Ministry Education, Nanchang Institute of Aeronautical Technology, Nanchang 330034, Jiangxi Province, China; 2. Fukui University, Fukui 910-8507, Japan)

ABSTRACT: A low loss passive snubber is developed for the PWM push-pull Voltage Source Inverter (VSI). In order to avoid the superposition of the high frequency discharge energy on the output voltage, an independent discharge circuit is introduced to feed back part of the energy in the capacitor to DC source. Due to the voltage clamp on the capacitor, there is, for the doubled voltage of the push-pull circuit, no additional loss in the snubber. Details of design and a 200V DC input, 30kW VSI application example are also presented. Experimental results show that the circuit topology is a good choice in low loss snubber design.

KEY WORDS: electric power engineering; push-pull voltage source inverter; surge voltage; switching loss; snubber circuit

摘要: 提出并实现了一种推挽式 PWM 电压型逆变器用无源吸收电路; 设有独立的放电回路让缓冲电容器上的能量部分地回馈到直流电源中, 高频放电能量不会叠加在输出电压上; 缓冲电容器的电压被钳位, 使得推挽电路特有的倍压不会在吸收电路上造成额外的损耗。给出吸收电路参数的设计准则和计算实例, 在 200VDC 和 30kW 级的中压中功率逆变装置中发挥了推挽拓扑的优点。理论分析和实验验证了所提方法的有效性。

关键词: 电力工程; 推挽式电压源逆变器; 浪涌电压; 开关损耗; 吸收回路

0 INTRODUCTION

Owing to the advantages of less switching devices, good electrical isolation and better output

adjustability, push-pull topology VSI are often used in DC/DC and DC/AC converters in the cases such as aeroplane[1], railway[2], etc. High voltage across the switching devices is its disadvantage[1-4]. IGBT brings more application choices to push-pull topology VSI. It is evident that switching condition plays a very important role in the development of push-pull topology VSI. To meet the increasing power density, so as to obtain higher reliability, following are the main topics in the studies of snubber/soft switching technology: ①Combination of active and passive modes. ②No more switching devices needed. ③No more sensors and control circuits. ④As less devices as possible are used. ⑤Free from parasitic oscillation. ⑥ Currents in the switching devices remain unchanged. ⑦Independent discharge loop[5-7]. In the snubber/soft switching technology of bridge topology VSI, studies are mainly on multilevel[8], multi target optimization[9], low loss[10], no loss[7-8] and free from snubber[11]. Though there are disadvantages of higher loop current and strict quality demands to the voltage clamp capacitor, the push-pull forward topology is good at suppressing the turn-off voltage on the switching devices. Studies on resonant ZCS and ZCT were also seen[12]. Compared with the studies mentioned above, the progress of snubber/soft switching technology in push-pull topology VSI is relatively slow. In reference [2], the passive snubber consumes energy because the switching devices are

基金项目: 国家自然科学基金项目(50467003)。

Project Supported by National Natural Science Foundation of China(50467003).

part of the discharge loop. There is not only the energy dissipation stored in related inductors, but also the additional loss of the doubled voltage due to the push-pull topology. Analyses on total harmonic distortion and distortion factor of PWM push-pull and two-armed push-pull VSIs were presented in references [3] and [13], but the snubber discharge passes through the transformer. It works well in DC/DC conversion, but in DC/AC conversion, there are significant high frequency components in the VSI output. The main reasons affect the study on push-pull VSI are: ①Parameters such as leakage inductance should be taken into consideration in the snubber design. But the relativity of the leakage flux between the 3 windings of the transformer, as well as the relativity of corresponding leakage inductances, still remain to be solved both theoretically and practically[14]. ②More energy is absorbed and dissipated by the snubber. ③The two switching devices are connected by their emmitters. Compared with bridge VSI and push-pull forward VSI, it is very difficult either to find corresponding circuit, or to learn from other snubber/soft switching technology.

In the development of a 30 kW instantaneous voltage compensator[4,15], push-pull topology was employed due to the consideration of voltage superposition. To suppress the turn-off voltage overshooting on the switching devices, a low loss passive snubber was developed and introduced into the voltage compensator. Experiments show that the snubber well meets the 7 topics[5-7]given before. The turn-off voltage overshooting on the switching devices was successfully suppressed. Fine AC output was also obtained.

1 CIRCUIT ANALYSIS

1.1 Main circuit and snubber

(1) Main circuit.

Fig.1 shows the main circuit of push-pull topology VSI^[2-3,13], where N_1 and N_2 are the primaries, N_3 the secondary, i_1 , i_2 and i_3 are corresponding currents, l_1 and l_2 are the effective mutual inductive leakages which affect the snubber design directly. Driven by U_{ge1} and U_{ge2} , the input current of the

transformer i_i consists of i_T and i_D , where i_T is the current passing through the switching devices T_1/T_2 , i_D the current passing through the sustaining diodes D_2/D_1 , respectively. Equ.(1) shows the transformer input current i_i . Fig.2 shows the waveform of i_i .

$$i_i(t) = i_1(t) + i_2(t) = i_{T1}(t) + i_{D2}(t) + i_{T2}(t) + i_{D1}(t) \quad (1)$$

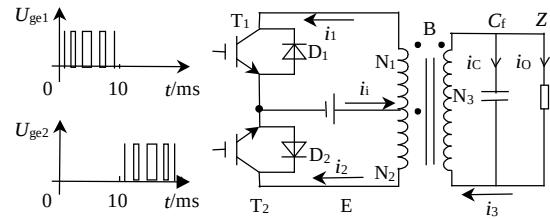


图1 一种基本的推挽式电压源逆变器

Fig.1 A basic pushi-pull voltage source inverter

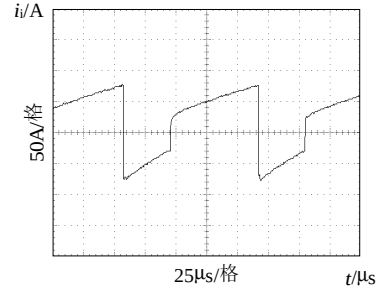


图2 变压器的输入电流

Fig.2 Input current of transformer

Based on Fig.1 and Fig.2, following conclusions can be drawn: ① i_T is of the same polarity as E_0 , energy is passed to load Z as T is turned on. On the other hand, i_T is of the opposite polarity as E_0 , the energy stored in transformer B and load Z will be fed back to E_0 as D works in sustaining state. ②In Fig.2, the positive part in the waveform shows i_T , the negative part shows i_D . ③Values of $|di_T/dt|$ and $|di_D/dt|$ are very large in the switching transient moment. As there is no snubber, the turn-off voltage overshooting on the switching devices will reach 1200V, as shown in Fig.3.

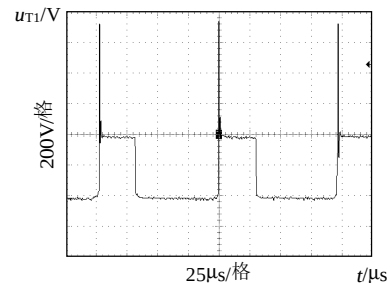


图3 主管关断电压

Fig.3 Turn-off voltage overshooting on switching device

The high voltage overshooting not only brings significant switching loss, but also is very close to the maximum inverse voltage limit of the switching device. It is necessary to suppress the turn-off voltage overshooting to a safe level.

(2) The low loss passive snubber in push-pull VSI.

To suppress the turn-off voltage overshooting on the switching devices, a low loss passive snubber is introduced into the push-pull VSI, as shown in Fig.4.

1.2 Turn-off analysis on switching devices (u_o and i_o of same polarity)

(1) (t_0-t_1), T_1 turned on.

In this case, $U_{ge2} < 0$, T_2 is turned off. Diodes D_1 , D_2 , D_3 and D_4 are also turned off. Fig.5 shows the equivalent circuit in Fig.4.

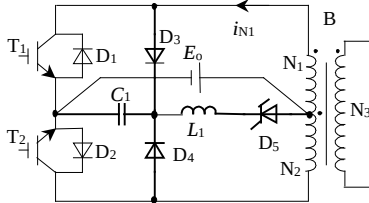


图 4 低损耗吸收电路

Fig.4 The low loss snubber circuit

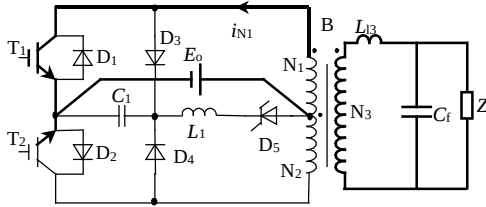


图 5 T_1 导通时的相关回路

Fig.5 The relevant circuit during the T_1 on-period

$i_1(t)$ keeps increasing within (t_0-t_1). The energy is passed from E_0 to Z by the transformer, when the leakage inductance between N_1 and N_2 is ignored[14].

Fig.6 shows the currents in N_1 and C_1 .

(2) (t_0-t_2), T_1 turned off.

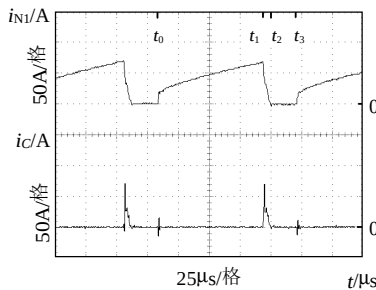


图 6 i_1 和 i_C

Fig.6 Current in N_1 and C_1

As T_1 is turned off, i_{T1} decreases very sharply, $L_1 di_1/dt$ is of the same polarity as E_0 , where L_{11} is the effective leakage inductance between N_1 and N_2 [15], D_3 is turned on. In this case, the magnetic energy stored in L_1 becomes electric energy, together with the energy from E_0 , starts charging C_1 . Note that the L_1 branch is ignored in the analysis because the current in L_1 is not allowed to change suddenly. According to $i_1 = i_{T1} + i_{C1} = I_0$, and i_{C1} can change immediately, there is $i_1 > 0$. The IGBT turn-off time $t_{off} \approx 200\text{ns}$. The turn-off time of $i_1(t)$ is accordingly adjusted to $3.5\mu\text{s}$. Furthermore, there is $i_1 = i_{T1}$ for $t = 0$. As a result, $|di_1/dt| \ll |di_{T1}/dt|$. It brings significant $|di/dt|$ decrease to the transformer leakage inductance, as well as to the related stray inductances. Finally, u_{ce1} , the turn-off voltage overshooting on the switching device T_1 , is lowered significantly. Fig.7 shows the charging circuit of C_1 .

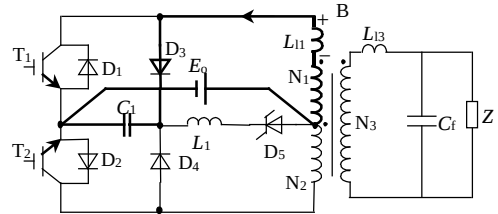


图 7 C_1 的充电回路

Fig.7 Charging circuit of C_1

As the winding resistance is ignored, the induced voltage of N_2 can be expressed by $u_2 = -L_{21} di_1/dt + L_{22} di_2/dt - L_{23} di_3/dt$, where L_{ij} stands for the inductance between winding i and winding j . When D_2 starts current sustaining by u_2 , the transformer input current i_1 changes from i_1 to i_2 . Fig.8 shows the superposition of i_2 with i_1 . It should be pointed out that because $i_{D2} + i_{C1} = I_0$, the transformer input current i_i will remain unchanged within the 200ns IGBT turn-off time, and puts no effects on the secondary.

(3) (t_2-t_3), the current sustaining period.

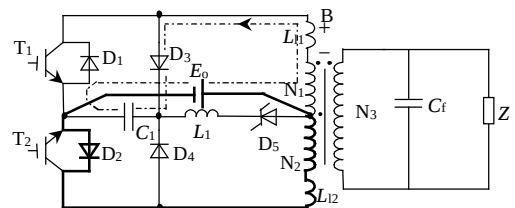


图 8 叠流过程

Fig.8 Superposition of i_1 and i_2

Fig.9 and Fig.10 are the current sustaining circuit and the sustained currents, respectively. In this period, i_i moves from N_1 to N_2 , the energy stored in transformer B feeds back to E_0 , and partly passed to Z by L_{13} [14].

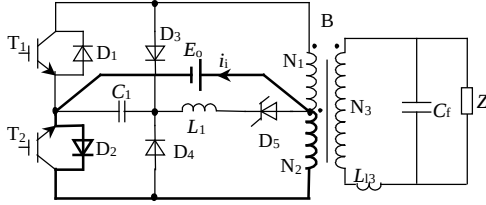


图9 续流过程

Fig.9 Current of sustaining circuit

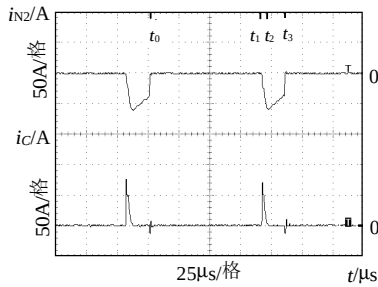


图10 i_2 和 i_C

Fig.10 Waveforms of sustained current

There are two things need to be noted in the analysis: ① $|i_i| < I_0$. ② $|i_2/dt|$ decreases due to the existence of $i_C(\tau)$.

(4) Discharge of C_1 . Fig.11 and Fig.12 show the C_1 discharge circuit and its equivalent circuit, respectively.

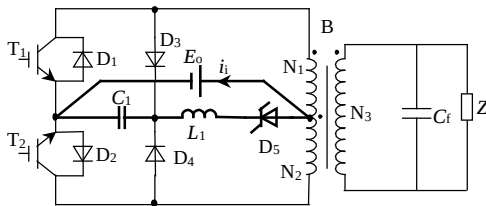


图11 电容器放电回路

Fig.11 C_1 Discharge circuit

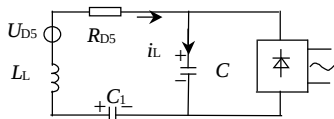


图12 放电等价回路

Fig.12 Equivalent circuit of Fig.11

In Fig.12, R_{D5} is the equivalent resistance of Zener diode D_5 , U_{D5} is the discharge voltage of D_5 , L_L is the limiting inductor, C is the decoupling capacitor of DC source. These components form an independent discharge circuit for capacitor C . Because $C \gg C_1$, in

the turn-off period, there is only a tiny DC voltage increase on capacitor C . It is worth knowing that the discharge circuit is completely independent from load Z , and of no need for a special-designed transformer.

(5) Waveforms in the snubber.

Fig.13 and Fig.14 are the voltage waveforms for the switching devices, the sustaining diodes and the capacitor, respectively.

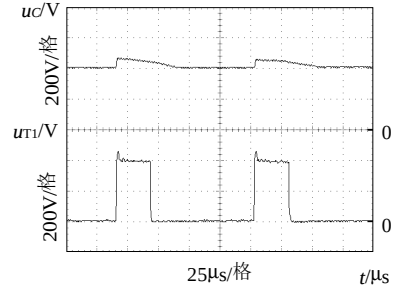


图13 u_o 与 i_o 同向时 T_1 的电压和 C_1 的电压

Fig.13 Voltages waveforms of T_1 and C_1 (u_o and i_o of same polarity)

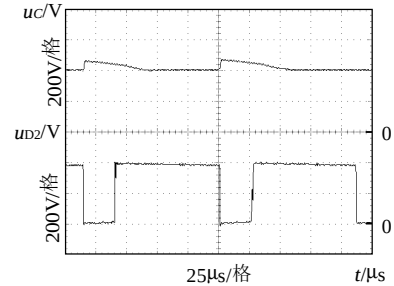


图14 u_o 与 i_o 同向时 D_2 的电压和 C_1 的电压

Fig.14 Voltages waveforms of D_2 and C_1 (u_o and i_o of same polarity)

It is evident by the waveforms that ① The turn-off voltage overshooting has been lowered significantly. ② Due to D_5 , there is a room for the selection of the minimum residual value after the discharge. It keeps $u_{C1} = 2E_0 + \Delta u$ throughout whole discharge period. Therefore, the doubled voltage in the push-pull VSI brings no additional loss to the snubber. ③ The voltage increase on capacitor C is caused by the IGBT turn-off operation. Two switching devices share the capacitor C , as well as the discharge circuit in the snubber. ④ The discharge current peak value is suppressed due to L_1 , with time interval $25\mu s$. ⑤ There are no harmful oscillations as the sustaining diodes are turned-on, or turned-off.

1.3 Turn-off analysis on switching devices (u_o and i_o of opposite polarities)

(1) The transformer input current i_i .

With load power factor $\cos a < 1$, as the filter loss omitted, fundamental of i_i can be expressed by Equ.(2). Fig.15 shows the current waveform in the case of load impedance angle $a \approx -60^\circ$.

$$\begin{cases} E_0 i_i = \sqrt{2} u_o \sin wt \cdot \sqrt{2} i_o \sin(wt - a) \\ i_i = \frac{u_o i_o}{E_0} \cos a - \frac{u_o i_o}{E_0} \cos(2wt - a) \end{cases} \quad (2)$$

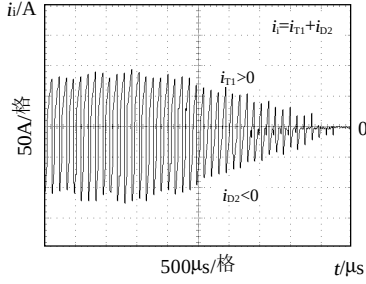


图 15 a 滞后 60° 时的 i_i

Fig.15 Current waveform of $a \approx -60^\circ$

There are following conclusions:

1) The i_i fundamental frequency is as twice as that of the power frequency, and which is consistent with Equ.(1).

2) The right part in the waveform corresponds to the case that u_o and i_o are of opposite polarities. It shows that the sustaining time of $i_2(i_{D2})$ is longer than that of $i_1(i_{T1})$.

3) Taking the whole waveform into consideration, there is the equation for u_o and i_o of opposite polarities:

$$\left| \int_0^t i_{T1}(t) dt \right| < \left| \int_0^t i_{D2}(t) dt \right|, \left| \int_0^t i_{T2}(t) dt \right| < \left| \int_0^t i_{D1}(t) dt \right|$$

With $E_0 > 0$, the integral result will be negative. It means the DC source absorbs energy.

4) The left part in the waveform reflects the case that u_o and i_o are of same polarity. It shows that the sustaining time of $i_2(i_{D2})$ is less than the sustaining time of $i_1(i_{T1})$, as shown in Fig.13 and Fig.14.

5) Taking the whole waveform into consideration, there is the equation for u_o and i_o of same polarity:

$$\left| \int_0^t i_{T1}(t) dt \right| > \left| \int_0^t i_{D2}(t) dt \right|, \left| \int_0^t i_{T2}(t) dt \right| > \left| \int_0^t i_{D1}(t) dt \right|$$

With $E_0 > 0$, the integral result will be positive. It means the DC source outputs energy.

(2) Voltage stress on D_2 .

As T_1 is turned on, D_2 will be turned off by the inverse voltage in transformer B, as shown in Fig.16.

Corresponding voltage stress is shown in Fig.17.

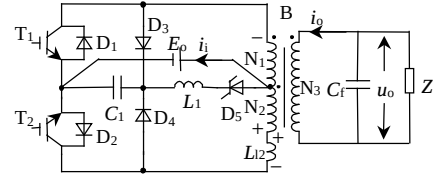


图 16 续流结束时的漏感反电势

Fig.16 Inverse induced voltage of leakage inductance at the end of current sustaining

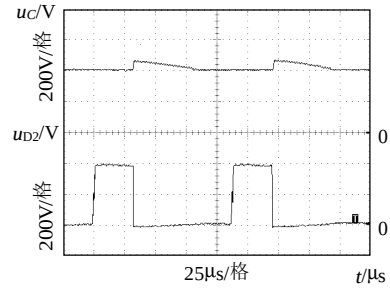


图 17 u_o 与 i_o 反向时 D_2 的电压和 C_1 的电压

Fig.17 Voltages D_2 and C_1 when u_o and i_o are of opposite polarity

It can be seen that ①As D_2 is turned off, the induced voltage polarity of L_{l2} is opposite to E_0 . ② There is no voltage overshooting along with the increase of u_{D2} . ③The sustaining time of $i_2(i_{D2})$ is longer than the sustaining time of $i_1(i_{T1})$. ④There are also no harmful oscillations as the sustaining diodes are turned on, or turned off.

(3) Voltage stress on T_1 .

When u_o and i_o are of opposite polarities, the T_1 turn-on and turn-off voltage waveforms are shown in Fig.18.

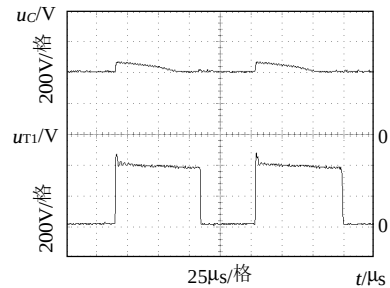


图 18 u_o 与 i_o 反向时 T_1 的电压和 C_1 的电压

Fig.18 Voltages T_1 and C_1 when u_o and i_o are of opposite polarity

It can be seen that u_{T1} stays a longer time at the high level than that at the low level, and the turn-off voltage overshooting on the switching devices has been suppressed significantly.

2 DESIGN EXAMPLE

2.1 Parameters of Snubber Components

(1) Diodes D_3 and D_4 .

D_3 and D_4 are the switching diodes as used in ordinary snubbers. Switching diode FR157 is selected in the design example.

(2) Capacitor C_1 .

Capacitance of C_1 is obtained by

$$C_1 = \frac{L_L \cdot i_{\max}^2}{u_{C1}(0) - u_{C1}(T)} \quad (3)$$

where $u_{C1}(0)$ and $u_{C1}(T)$ are the voltages on C_1 before and after its discharge, respectively; $i_{\max}$ is the maximum current; L_{11} is the transformer leakage inductance, L_{11s} is the stray inductance of the main circuit. Besides, the current falling time of the switching devices, as well as the overall power loss in the switching devices and in the snubber, should also be taken into C_1 computation. A general rule $u_{C1}(0)/u_{C1}(T) \approx 1.1$ is suggested. In the design example, $L_{11s} \approx L_{12s} \approx 2\mu\text{H}$, $L_{11} \approx L_{12} \approx 11\mu\text{H}$, $u_{C1}(0) \approx 460\text{V}$, $u_{C1}(T) = 420\text{V}$. Finally, there is $C_1 \approx 4\mu\text{F}$.

(3) Zener Diode D_5 .

1) D_5 is aimed to keep the C_1 voltage much than twice the DC voltage: $u_{C1} = u_{D5} + E_0 > 2E_0$, or $u_{D5} > E_0$. D_3 is turned on and D_4 only at the transient moment as the switching device is turned off, so as to avoid the additional snubber loss caused by the doubled voltage in push-pull circuit. $u_{D5} \approx 1.05E_0 \approx 210\text{V}$ is selected in the example.

2) To speed up the discharge, the dynamic resistance R_Z of D_5 should be about 10Ω . 100W210 Zener diodes are used.

3) According to Fig.13, Fig.14, Fig.17 and Fig.18, the discharge time is less than 1/2 cycle, or the value of effective current is low.

(4) Inductance L_L .

Following things should be taken into the computation of L_L .

1) The discharge current peak value must be lower than the maximum current limit of D_5 .

2) Free from parasitic oscillation.

3) The discharge should finish within one cycle, or $T = 1/f_c = 100\mu\text{s}$. Theoretically, L_L can be obtained by

the circuit equations based on Fig.9. But due to the nonlinearity of D_5 , it is difficult to find the analytical solution. Engineering solutions are experimentally available. In the design example, $L_L = 5\mu\text{H}$.

2.2 Energy analyses in snubber

(1) Discharge circuit analysis

Fig.13, Fig.14, Fig.17 and Fig.18 show that: ① There is no oscillation in the discharge circuit. ② The current peak is limited by L_L only at the very beginning of the discharge. To simplify the computation, the nonlinearity of R_Z is ignored, or the discharge analysis is taken as a constant $R_Z C$ circuit. Consequently, the discharge of C_1 can be shown by Equ.(4)

$$u_{C1} = E_0 + \{u_{C1}(0) - E_0\} e^{-\frac{t}{R_Z C}} \quad (4)$$

In Fig.4, let $u_{C1}(0) = 2.3E_0$, $u_{C1}(T) = 2.1E_0$, there is Equ.(5).

$$e^{-\frac{T}{R_Z C}} = \frac{2.1E_0 - E_0}{2.3E_0 - E_0} = 0.85 \quad (5)$$

(2) Analyses on energy feedback and dissipation.

Insert Equations (4) and (5) into Equation (6),

$$w_1 = \int_0^T R_Z i^2 dt = \int_0^T R_Z \left[-C \frac{du_{C1}}{dt} \right]^2 dt = \frac{\{u_{C1}(0) - E_0\}^2}{R} \int_0^T e^{-\frac{2t}{R_Z C}} dt = 0.18CE_0^2 \quad (6)$$

there is the energy dissipation of D_5 within one cycle T : $w_1 = 0.18CE_0^2$.

The energy difference in the discharge of C_1 , or the energy transferred from the transformer leakage inductance and the stray inductances to capacitor C_1 is

$$w_2 = 0.5C\{u_{C1}^2(0) - u_{C1}^2(T)\} = 0.44CE_0^2 \quad (7)$$

Energy fed back to DC source is shown by Equ.(8)

$$w_3 = w_2 - w_1 = 0.26E_0^2 \quad (8)$$

It is obvious that $w_3/w_2 = 0.59$, or 59% of the energy is fed back to DC source.

(3) More details of the snubber.

Oscilloscopes TDS-3032B and TDS-340A were used in the snubber development, with A622 current probe. The IGBT SKM100GB-123D was selected as switching device.

Fig.19 is the picture of the snubber.

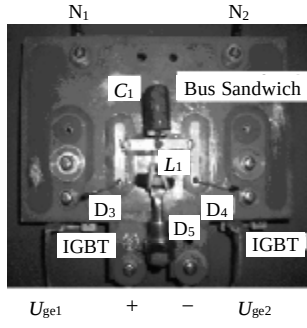


图 19 吸收电路照片

Fig.19 Picture of the snubber

2.3 Comparison analysis

(1) Comparison analysis with energy-consuming snubber.

Switching device loss and snubber loss are the two parts of the energy loss in the energy-consuming snubber. To be convenient, only the energy transmission in the capacitor is taken into consideration. Fig.20 shows the typical circuit of energyconsuming snubber.

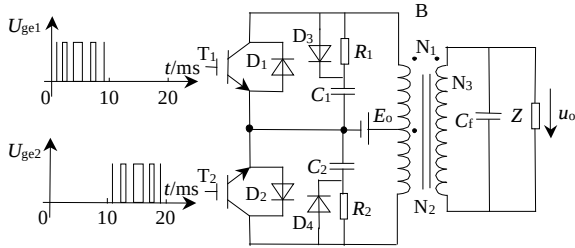


图 20 现行耗能型吸收电路图

Fig.20 Circuit of traditional energy-consuming snubber

Within time interval $t \in (0 \sim 10 \text{ ms})$, the discharge circuit of C consists of: ① $C_1 \rightarrow R_1 \rightarrow$ the turned-on T_1 . ② $C_2 \rightarrow R_2 \rightarrow$ the current sustaining D_2 . Not only the current burden on switching devices, but also the switching loss, is significantly increased. Equ. (9) shows the energy loss within one cycle.

$$w_4 = 0.5Cu_T^2(0) + 0.5Cu_D^2(0) = 4.65CE_0^2 \quad (9)$$

$u_T(0)$ is the capacitor charge voltage in the snubber as the IGBT is turned off, here $u_T(0) = u_{C1}(0) = 2.3E_0$. $u_D(0)$ is another capacitor charge voltage in the snubber, which is caused by the doubled voltage of the push-pull VSI as the IGBT is turned on, here $u_D(0) = 2E_0$. The energy consumption ratio between the design example and the energy-consuming snubber is:

$$w_1/w_4 = 0.26CE_0^2/4.645CE_0^2 = 0.06$$

(2) Comparison analysis with energy-feedback-to-load snubber[1,12].

In the low loss passive snubber presented by the author, the energy is fed back to DC source, the discharge circuit is free from load Z . As a result, they make no contributions to the output, or better sine output is available. Fig.21 shows the output waveform obtained in the design example, with 220V AC output, load current 45A, and $\cos\alpha = 0.99$.

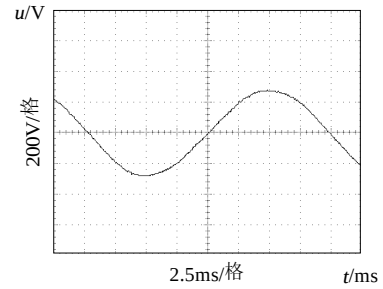


图 21 输出电压

Fig.21 Output waveform in the design example

(3) Comparison analysis with bridge power inverter.

The turn-off voltage overshooting in bridge power inverter is mainly due to the DC stray inductance (usually $\times 10^{-9} \text{H}$ [9]) and the stray inductance caused by the snubber leads. In push-pull VSI, despite the stray inductances mentioned above, the effective inductance (usually $\times 10^{-6} \text{H}$ [14]) between the transformer primary and secondary is also in relation to the overshooting. Therefore, more power is absorbed and fed back in the snubber used in push-pull VSI.

(4) Discussion on snubbers used in other push-pull VSIs.

The VSI in reference [14] is of push-pull topology. It is DC/DC converter. DC/DC converter of push-pull topology is unnecessary to consider the sine characteristics of the output. In reference [12], the VSI belongs to the push-pull forward topology and differs from the VSI topology presented here. Therefore, no comparison is available.

3 CONCLUSIONS AND DISCUSSIONS

(1) As of same power capacity, the turn-off voltage overshooting in push-pull VSI is higher than that in bridge power inverter.

(2) The low loss passive snubber is also available in PWM push-pull VSI, as in bridge power inverter. The energy loss in the low loss snubber is only 1/5 of that in energy-consuming snubber. Besides, better sine output is also available by the low loss snubber.

(3) The voltage higher than $2E_0$ on the capacitor avoids the additional loss in the snubber due to the doubled voltage in push-pull VSI. The energy is fed back to DC source. The discharge circuit is independent from the load. As a result, neither current burden increase, nor voltage stress increase is added on the switching devices.

(4) Only Zener diode and limiting inductor are used in the energy feedback circuit. It is no need to introduce mutual inductor to solve the problems such as flux restoration and diode maximum voltage limit in the secondary.

ACKNOWLEDGMENT

The authors wish to thank Prof. Li Zhou from Guangxi University (China) and Prof. Chen Xuan at NIAT, for his work and support. We would also like to thank Samvel Guglloetto, Visiting professor at Nanchang Institute of Aeronautical technology, for his editorial assistance.

REFERENCES

- [1] 陈道炼, 李磊, 胡育文. 双极性移相控制电压源高频交流环节 AC/AC 变换器研究[J]. 中国电机工程学报, 2002, 22(9): 69-73.
Chen Daolian, Li Lei, Hu Yuwen. Research on the bi-polarity phase-shifted controlled voltage mode AC/AC converters with high frequency AC link[J]. Proceedings of the CSEE, 2002, 22(9): 69-73(in Chinese).
- [2] 李铃铃. 一种 DC/DC 大功率变换器的设计[J]. 电气传动, 1997, 27(3): 51-55.
Li Lingling. The design of a large power DC-DC converter[J]. Electric Drive, 1997, 27(3): 51-55(in Chinese).
- [3] 谷春艳, 刘超英. 双臂推挽式单相电压源逆变器及其调制方案[J]. 中国电机工程学报, 2004, 24(9): 99-103.
Gu Chunyan, Liu Chaoying. Two-armed push-pull single-phase voltage source inverters and their modulation scheme[J]. Proceedings of the CSEE, 2004, 24(9): 99-103(in Chinese).
- [4] 伍家驹, 刘桂英, 王牛保. 瞬间电压降补偿器的研制[J]. 电力电子技术, 1999, 33(6): 7-10.
Wu Jiaju, Liu Guiying, Wang Niubao. Development of the instantaneous voltage drop compensator[J]. Power Electronics, 1999, 33(6): 7-10(in Chinese).
- [5] 林渭勋. 现代电力电子电路[M]. 杭州: 浙江大学出版社, 2002.
- [6] 邓焰, 何湘宁, 钱照明. 无源吸收技术的原理与典型电路[C]. 第十三届全国电源技术年会论文集, 中国, 深圳, 2000: 358-362.
- [7] 邓焰, 吴燮华, 何湘宁, 等. 一种新型高功率逆变桥臂无源无损耗吸收电路[J]. 中国电机工程学报, 2000, 20(3): 5-9.
Deng Yan, Wu Xiehua, He Xiangning, et al. A novel passive lossless snubber for power inverter bridge legs[J]. Proceedings of the CSEE, 2000, 20(3): 5-9(in Chinese).
- [8] 吴洪洋, 邓焰, 何湘宁. 一种新颖的多电平逆变器无源无损耗吸收电路[J]. 中国电机工程学报, 2002, 22(10): 71-76.
Wu Hongyang, Deng Yan, He Xiangning. A novel passive lossless snubber for multilevel inverters[J]. Proceedings of the CSEE, 2002, 22(10): 71-76(in Chinese).
- [9] 唐欣, 罗安, 李刚. 大功率绝缘栅双极晶体管模块缓冲电路的多目标优化设计[J]. 中国电机工程学报, 2004, 24(11): 144-147.
Tang Xin, Luo An, Li Gang. Multi-objective optimal design of snubber circuits for high power IGBT modules[J]. Proceedings of the CSEE, 2004, 24(11): 144-147(in Chinese).
- [10] 金宏信. モータ端子サージ電圧抑制用低損失スナバ[J]. 电气学会论文志 D, 2003, 123(8): 918-925.
Hironobu Kon. Low loss snubber for reducing motor terminal surge voltage[J]. IEE Japan, 2003, 123(8): 918-925(in Chinese).
- [11] 加藤修治, 上田茂太, 酒井洋满, 等. スナバレス直列接続に適したアクティブゲートコントロール[J]. 电气学会论文志 D, 2002, 122(8): 816-822.
Shuji Katoh, Shigeta Ueda, Hiromitsu Sakai, et al. Active gate control for Snubberless IGBTs connected in Series[J]. T. IEE Japan, 2002, 122(8): 816-822.
- [12] 王慧珍, 戴卫力. 推挽正激软开关电路谐振网络的分析与设计[J]. 电工技术学报, 2004, 19(8): 78-81.
Wang Huizhen, Dai Weili. Analysis and design considerations for the resonant network of push-pull forward soft-switching converter[J]. Transactions of China Electrotechnical Society, 2004, 19(8): 78-81(in Chinese).
- [13] 刘超英, 成红英. 一种适用于推挽式逆变器的低损耗缓冲吸收器及其对功率电路的影响[J]. 电工电能新技术, 2002, 21(2): 46-48, 60.
Liu Chaoying, Cheng Hongying. A low-loss snubber for pushu-pull inverters and its effect to power circuits[J]. Advanced Technology of Electrical Engineering and Energy, 2002, 21(2): 46-48, 60(in Chinese).
- [14] 伍家驹, 杉本英彦, 余达祥, 等. 推挽变压器漏感的一种间接测算方法[J]. 中国电机工程学报, 2005, 25(23): 129-137.
Wu Jiaju, Hidehiko Sugimoto, Yu Daxiang, et al. Determining the leakage inductance of push-pull transformers with an indirect method[J]. Proceedings of the CSEE, 2005, 25(23): 129-137(in Chinese).
- [15] 伍家驹, 杉本英彦, 任吉林. 一种综合提高用电质量的方法和装置[P]. 批准专利权号: ZL 99.1 17667.7.

APPENDIX

Power frequency magnetization analysis on push-pull transformer

Charge and discharge are the two operations in the passive snubber which is aimed to suppress the turn-off voltage overshooting on the switching devices. As the switching device

is turned off, At the transient moment (approximately 200ns) energy stored in leakage inductance and stray inductance becomes electric energy, then stored in the capacitor by charging it. It brings no additional effects on the magnetization of the push-pull transformer. The discharge circuit is completely independent from the main circuit, and brings no effects on the magnetization of the push-pull transformer.

To obtain higher power density, the magnetizing curve of the transformer is often positioned at the turning section in front of the B-H loop saturation area. There should be low frequency components in the magnetizing current, so as to make the flux fundamental a sine wave. In the modulation that used in references [2-3], and the low loss passive VSI development, the magnetizing current is PWM controlled by a closed loop. The transformer input current i_i is shown by

Equ.(1) and Fig.15. i_3 is a rippled sine wave with power frequency, and shown in Fig.1. Acted by i_i and i_3 , The push-pull power transformer will be magnetized both bilaterally and symmetrically within a cycle of power frequency[14].

收稿日期：2006-01-09。

作者简介：

伍家驹(1951—)，男，教授，主要从事电力电子技术的科研与教学工作，主要研究方向为功率变换器及其控制策略，wujiaj2003@yahoo.com.cn；

杉本英彦(1943—)，男，博士生导师，主要从事电力电子技术的科研与教学工作，主要研究方向为交流电机的高性能控制，太阳能发电，超导电机及直线电机；

(责任编辑 韩 蕾)