Direct Power Control without Current Sensors for Nine-Switch Inverters

Lei Pan†, Junru Zhang*, Kai Wang*, Beibei Wang*, Yi Pang*, and Lin Zhu*

†,*School of Control and Mechanical Engineering, Tianjin Chengjian University, Tianjin, China

Abstract

Recently, the nine-switch inverter has been proposed as a dual output inverter. To date, studies on the control strategies for NSIs have been mostly combined with their application. However, in this paper, a mathematical model and control strategy for nine-switch inverters has been proposed in view of the topology. A switching function model and equivalent circuit model of a nine-switch inverter have been built in αβ coordinates. Then, a novel current observer with an improved integrator is proposed based on the switching function model, and a direct power control strategy is proposed. No current sensors are used in the proposed strategy, and only two voltage sensors are employed. The performance of the proposed control method is verified by simulation and experimental results.

Key words: Current observer, Direct power control, Nine-switch converter, Switching function

I. INTRODUCTION

Recently, a novel three-phase three-leg inverter, which is referred to as a nine-switch inverter (NSI), with nine IGBTs has been proposed in the literature [1]. This inverter, shown in Fig. 1, is composed of two three-phase inverter units. These units are named upper and lower units, and they share the same dc-link voltage.

In the literature, different modulation methods [2]-[12] and applications [13]-[25] have been widely reported.

There are two main kinds of modulation methods. These methods are the carrier-based pulse-width modulation (PWM) method [2]-[8] and the space-vector modulation (SVM) method [9]-[12]. PWM can be used for the different frequency (DF) operation mode and the constant frequency (CF) operation mode [2], and the research content includes the carrier-based pulse width modulation mechanism [3], the constraint relationship between the dc link voltage, phase difference and voltages of two three-phase ac terminals [4], the carrier-based pulse width modulation with dead-time elimination [5], and the switching and conduction losses [7]. SVM can also be used for the DF mode and the CF mode [9], [10]. In some literatures on SVM, the space vector modulation mechanism [11] and the spatial distribution of the voltage vector [12] are developed in-depth.

The applications of NSIs have been proposed in many fields, such as hybrid electric vehicles [13], power conditioners [14], [15], torque control [16], wind power systems [17], [18], uninterruptible power supplies (UPS) [19], photovoltaic systems [20], [21], and electrical machines [22].

At present, studies on the control strategies for NSIs are mostly combined with applications [15]-[25]. Considering that the switches in the NSC are shared by two three-phase ports, a new adapted control method is proposed to flexibly control and fully use of the current capacity of the switches for DFIG wind power systems [17]. In the DVCA strategies for DFIGs [23], the rotor-side gets more voltage to suppress overcurrent when the symmetrical grid voltage dips with dynamic voltage assignment. In addition, the grid-side gets more current capacity for reactive current compensation with dynamic current assignment. In the integrated motor drive and battery charger systems for EVs, grid voltage-oriented control, which is obtained through a phase-locked loop (PLL) algorithm with resonance, is adopted to obtain a highly satisfactory performance [24]. The steady-state control strategy and transient-management scheme for system dynamics and grid faults in a NSI are developed to ensure proper performance in the steady-state, dynamic operation and enhanced fault ride-through capability of the FSIG-WT [25].

© 2018 KIPE
However, in this paper, a mathematical model of a NSI is established in view of the topology. Then, the direct power control method for a NSI is proposed based on $\alpha\beta$ coordinates. In this control method, only two voltage sensors are used and there are no current sensors. In order to improve the control effect, a new type of current observer, where an improved integrator with saturated feedback (IISF) is proposed. Simulation and experimental studies have been carried out to verify the effectiveness of the proposed scheme.

II. SWITCHING FUNCTION MODEL FOR A NSI

The topology of a NSI is shown in Fig. 1. Considering that there are three switches in each leg of the NSI, the semiconductors in each leg can have eight different ON-OFF positions. However to avoid DC bus short circuit, all three switches cannot be ON at same time. On other hand to avoid floating of the loads, at least two switches should be ON. Therefore, only three ON-OFF positions are possible. The constraint condition between the three switches in each leg of a NSI is shown in (1). In (1), A, B or C refers to leg A, B or C, and U, M or L refers to the upper, mid or lower semiconductor.

\[
\begin{align*}
S_{AA} + S_{AM} + S_{AL} &= 2 \\
S_{BH} + S_{BL} + S_{SL} &= 2 \\
S_{CH} + S_{CM} + S_{CL} &= 2
\end{align*}
\]

(1)

where $S_{JX}=1'$, when the switch $S_{JX}$ is ON; and $S_{JX}=0'$, when the switch $S_{JX}$ is OFF ($J=A, B, C; X=U, M, L$).

From Fig. 1, it is possible to obtain the voltage equation of a NSI with the switching variable shown in (2).

\[
\begin{align*}
u_{A1} &= u_{c1} S_{AA} - u_{c2} S_{AM} S_{AL} \\
u_{B1} &= u_{c2} S_{BH} - u_{c2} S_{BM} S_{BL} \\
u_{C1} &= u_{c3} S_{CH} - u_{c2} S_{CM} S_{CL} \\
u_{A2} &= u_{c1} S_{AA} - u_{c2} S_{AM} S_{AL} \\
u_{B2} &= u_{c1} S_{BH} S_{BM} - u_{c2} S_{BL} \\
u_{C2} &= u_{c1} S_{CH} S_{CM} - u_{c2} S_{CL}
\end{align*}
\]

(2)

where $u_{JY}$ is the voltage between $JY$ ($J=A, B, C; Y=1, 2$) and point o.

From Fig. 1, it is also possible to obtain the differential equation of a NSI, as shown in (3).

\[
\begin{align*}
L_1 \frac{di_{A1}}{dt} &= -R_i i_{A1} + u_{A1} - u_{nA,0} \\
L_1 \frac{di_{B1}}{dt} &= -R_i i_{B1} + u_{B1} - u_{nB,0} \\
L_1 \frac{di_{C1}}{dt} &= -R_i i_{C1} + u_{C1} - u_{nC,0} \\
L_2 \frac{di_{A2}}{dt} &= -R_i i_{A2} + u_{A2} - u_{nA,2} \\
L_2 \frac{di_{B2}}{dt} &= -R_i i_{B2} + u_{B2} - u_{nB,2} \\
L_2 \frac{di_{C2}}{dt} &= -R_i i_{C2} + u_{C2} - u_{nC,2}
\end{align*}
\]

(3)

where $u_{nY,0}$ is the voltage between nY ($Y=1, 2$) and point o.

In three-phase, three-wire systems, it can be known that:

\[
\begin{align*}
i_{A1} + i_{B1} + i_{C1} &= 0 \\
i_{A2} + i_{B2} + i_{C2} &= 0
\end{align*}
\]

(4)

From (2), (3) and (4), it can be concluded that:

\[
\begin{align*}
u_{nA,0} &= (u_{A1} + u_{B1} + u_{C1})/3 \\
u_{nB,0} &= (u_{A2} + u_{B2} + u_{C2})/3 \\
u_{nC,0} &= (u_{A3} S_{AA} + S_{BB} + S_{CC}) - u_{c2} (S_{AM} S_{AL} + S_{BM} S_{BL} + S_{CM} S_{CL})
\end{align*}
\]

(5)

From (2), (3) and (5), it is possible to obtain the switching function model for a NSI as shown in (6).

\[
\begin{align*}
L_1 \frac{di_{A1}}{dt} &= -R_i i_{A1} + \frac{1}{3} u_{c1} (2S_{AA} S_{AH} - S_{BH} - S_{CH}) \\
&\quad - \frac{1}{3} u_{c2} (2S_{AM} S_{AL} - S_{BM} S_{BL} - S_{CM} S_{CL}) \\
L_2 \frac{di_{A2}}{dt} &= -R_i i_{A2} + \frac{1}{3} u_{c1} (2S_{CH} - S_{AH} - S_{BH}) \\
&\quad - \frac{1}{3} u_{c2} (2S_{CM} S_{CL} - S_{AM} S_{AL} - S_{BM} S_{BL}) \\
L_1 \frac{di_{C1}}{dt} &= -R_i i_{C1} + \frac{1}{3} u_{c2} (2S_{AA} S_{AH} - S_{BH} - S_{CH}) \\
&\quad - \frac{1}{3} u_{c3} (2S_{AM} S_{AL} - S_{BM} S_{BL} - S_{CM} S_{CL}) \\
L_2 \frac{di_{C2}}{dt} &= -R_i i_{C2} + \frac{1}{3} u_{c3} (2S_{CH} - S_{AH} - S_{BH}) \\
&\quad - \frac{1}{3} u_{c3} (2S_{CM} S_{CL} - S_{AM} S_{AL} - S_{BM} S_{BL}) \\
L_1 \frac{di_{B1}}{dt} &= -R_i i_{B1} + \frac{1}{3} u_{c1} (2S_{BB} - S_{AH} - S_{BH}) \\
&\quad - \frac{1}{3} u_{c3} (2S_{BH} S_{BL} - S_{AH} S_{AL} - S_{CM} S_{CL}) \\
L_2 \frac{di_{B2}}{dt} &= -R_i i_{B2} + \frac{1}{3} u_{c2} (2S_{CH} - S_{AH} - S_{BH}) \\
&\quad - \frac{1}{3} u_{c3} (2S_{CM} S_{CL} - S_{AM} S_{AL} - S_{BM} S_{BL}) \\
L_1 \frac{di_{B1}}{dt} &= -R_i i_{B1} + \frac{1}{3} u_{c3} (2S_{CH} - S_{AH} - S_{BH}) \\
&\quad - \frac{1}{3} u_{c2} (2S_{CM} S_{CL} - S_{AM} S_{AL} - S_{BM} S_{BL})
\end{align*}
\]

(6)
Suppose a balanced three-phase voltage for two AC terminals, then:

\[ \begin{align*}
    u_{a1} + u_{b1} + u_{c1} & = 0 \\
    u_{a2} + u_{b2} + u_{c2} & = 0
\end{align*} \]  

(7)

Substituting (7) into (6) yields:

\[ \begin{align*}
    u_{a1}[S_{at}(1 + S_{at})] + S_{bt}(1 + S_{bt}) + S_{ct}(1 + S_{ct})] - u_{a2}[(S_{at} + 1)S_{bt} + (S_{bt} + 1)S_{ct} + (S_{ct} + 1)S_{ct}] & = 0 \\
\end{align*} \]  

(8)

(8) describes the constrained relationship between \( u_{a1} \) and \( u_{a2} \). From (9), it can be seen that the relationship between \( u_{a1} \) and \( u_{a2} \) depends on \( S_{at}(X=\text{ABC}) \) and \( S_{at}(X=\text{ABC}) \). Namely, \( u_{a1} = u_{a2} \) when \( S_{at} + S_{bt} + S_{ct} = S_{at} + S_{bt} + S_{ct} \).

III. DIRECT POWER CONTROL FOR A NSI

A \( \alpha \beta \) transformation or Clarke transformation can map the three-phase instantaneous line current, \( i_{a1}, i_{b1}, \) and \( i_{c1} \) into an instantaneous current on the \( \alpha \beta \)-axes \( i_{a2} \) and \( i_{b2} \). Therefore, the switching function model of a NSI in the A, B, and C coordinates can be transformed into the \( \alpha \beta \) coordinates with a Clarke transformation, and the transformed switching function model of the NSI is shown in (9).

\[ \begin{align*}
    [L_1 \frac{di_{a1}}{dt} + L_2 \frac{di_{a2}}{dt}] & = \begin{bmatrix} -R_s & 0 & 0 & 0 \\
                        0 & -R_s & 0 & 0 \\
                        0 & 0 & -R_s & 0 \\
                        0 & 0 & 0 & -R_s \end{bmatrix} \begin{bmatrix} i_{a1} \\
                                i_{a2} \\
                                i_{b1} \\
                                i_{b2} \end{bmatrix} + \begin{bmatrix} 0 \\
                                      0 \\
                                      0 \\
                                      1 \end{bmatrix} i_c \\
\end{align*} \]  

(9)

According to (9), it is possible to obtain an equivalent circuit model of a NSI in the \( \alpha \beta \) coordinates as shown in Fig. 2, when the three-phase voltage is balanced.

Where:

\[ \begin{align*}
    A & = \frac{3}{8}[S_{at} + S_{bt} + S_{ct} + S_{at}S_{bt} + S_{at}S_{ct} + S_{bt}S_{ct} + S_{at}S_{bt}S_{ct}] \\
    B & = \frac{3}{8}[S_{at}S_{bt} + S_{at}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct}] \\
    C & = \frac{3}{8}[S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct}] \\
    D & = \frac{3}{8}[S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct}] \\
    E & = \frac{3}{8}[S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct}] \\
    F & = \frac{3}{8}[S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct} + S_{at}S_{bt}S_{ct}] \end{align*} \]

Accordingly, the integral observer from Fig. 2 can be obtained as shown in Fig. 3.

In Fig. 3, there is an integrator in each of the current observers. This integrator introduces an integral initial value error and a dc offset error, which eventually leads to integrator saturation. The current estimation error curves are provided in Fig. 4, and the simulation condition is described in section IV.

Therefore, an improved integral method is needed to overcome this problem.

In Fig. 5, an improved current observer for the upper AC terminal is adopted. In this observer, an IISF is introduced,
and the effects of IISF are as follows:

When the system has just started, the current is small or close to zero, and the IISF is equivalent to a low-pass filter. When the system is in normal operation, the IISF is equivalent to an integrator. When the current is beyond the limiting value, the IISF is also equivalent to a low-pass filter.

The IISF can have the advantages of a pure integrator and a low-pass filter. It can eliminate the amplitude attenuation of a low pass filter, and inhibit the current of the dc component. The threshold value C is determined by the rated current of the AC terminal, and a certain margin is needed.

After improvement, current estimation error curves are shown in Fig. 6. With a comparison between Fig. 4 and Fig. 6, it can be seen that the response speed, overshoot and steady-state error of Fig. 6 are better than those of Fig. 4. That is to say, the performance of the current observer with an IISF is significantly better than the current observer with an integrator.

In addition, the IISF can be also applied to the lower AC terminal of a NSI, and the effects are the same as those of the upper AC terminal of a NSI.

Fig. 5. Improved current observer for the upper AC terminal of a NSI.

In addition, it is possible to conclude the \( u_{a1}, u_{b1}, u_{a2}, \) and \( u_{b2} \) from (2).

\[
\begin{align*}
    u_{a1} &= u_{c1}S_{a1} - u_{c2}S_{aM}S_{aL} \\
    u_{b1} &= u_{c1}S_{b1} + \frac{1}{2}u_{c2}(S_{aM}S_{bL} + S_{aL}S_{bM}) \\
    u_{a2} &= u_{c1}S_{a1}S_{aM} - u_{c2}S_{aL} \\
    u_{b2} &= -\frac{1}{2}u_{c2}(S_{a1}S_{bM} + S_{aM}S_{bL}) - u_{c2}S_{bL}
\end{align*}
\]

(10)

Therefore, a direct power control system can be built for a NSI, and the system diagram is shown in Fig. 7.

IV. SIMULATION AND EXPERIMENTAL RESULTS

Based on Fig. 7, a simulation system is built using Matlab/ Simulink for a NSI, and SVPWM is adopted [11]. The simulation conditions are the input voltage 500VDC, \( R_s = 50\Omega \), and \( L_{s1} = L_{s2} = 100\text{mH} \); and the output line voltages are 220Vrms with 50Hz for the upper load, and 110Vrms with 60Hz for the lower load. Fig. 8-13 show simulation results for the proposed direct power control system of a NSI.

From Fig. 8, it can be seen that the peak values of the upper and lower line voltages are about 310V and 155V, respectively. It can also be seen that the frequencies of the upper and lower line voltages are 50Hz and 60Hz, respectively. In Fig. 9, the peak values of the line current for
the upper and lower loads are about 3A and 1.4A, respectively. The frequencies of the upper and lower line current are 50Hz and 60Hz, respectively.

From Fig. 10 and Fig. 11, for the upper load, it can be seen that the active and reactive powers are about 750W and 435var; the maximum overshoots of the active and reactive power are about 21W and 20var; and the transition processes are about 0.01s for the active power and 0.02s for the reactive power. In addition, the steady-state errors are about 1W for the active power and 1var for the reactive power.

For the lower load, in Fig. 12 and Fig. 13, it can be seen that the active and reactive powers are about 155W and 115var; the maximum overshoots of the active and reactive power are about 8.5W and 7var; the transition processes are about 0.015s for the active power and 0.02s for the reactive power. In addition, the steady-state errors are about 0.5W for the active power and 0.5var for the reactive power.

To confirm the viability of the proposed the direct power control scheme, as shown in Fig. 7, a laboratory prototype for a NSI has been developed based on a TMS320F2812 DSP as shown in Fig. 14. The experimental conditions are the input voltage 500VDC, \( R_{s1} = R_{s2} = 50 \Omega \) and \( L_{s1} = L_{s2} = 100 \text{mH} \).

Firstly, in the CF mode, the frequencies of the output line voltages are 50HZ; and the upper and lower load output line
voltage are 110Vrms and 220Vrms, respectively. Fig. 15-21 show experimental results for the proposed direct power control system for a NSI.

Fig. 15 and Fig. 16 show that the peak values of the upper and lower line voltage are about 155V and 310V; the peak values of the phase voltage are about 90V and 180V; and the frequencies of the line voltage and phase voltage for the upper and lower load are 50Hz, respectively. In Fig. 17, the peak line current of the upper load is about 1.5A, and the peak line current of the lower load is about 3A. The frequencies of the upper and lower line current are also 50Hz. In Fig. 18, the THDs of the upper and lower line current are 3.69% and 3.32%, respectively. From Fig. 19 and Fig. 20, it can be seen that the active powers are about 175W and 700W for the upper and lower load, and that the reactive powers are about 110var and 435var for the upper and lower load, respectively.

From Fig. 21, the maximum errors of the active power are about 5W and 13W for the upper and lower load; the steady-state errors of the active power are smaller than 1W and 1W for the upper and lower load; and the transition processes of the active power are about 30ms for the upper power and 70ms for the upper and lower load, respectively. From Fig. 22, the maximum errors of the reactive power are about 4.5var and 13var for the upper and lower load; the steady-state errors of the reactive power are smaller than 0.5var and 1var for the upper and lower load; and the transition processes of the reactive power are about 20ms and 70ms for the upper and lower load, respectively.
Fig. 19. Active power curves for a NSI.

Fig. 20. Reactive power curves for a NSI.

Fig. 21. Active power error curves for a NSI.

Secondly, in the DF mode, the frequencies of the upper and lower load output line voltages are 60HZ and 50HZ; and the upper and lower load output line voltages are 110Vrms and 220Vrms, respectively. Fig. 23-30 show experimental results.

Fig. 22. Reactive power error curves for a NSI.

Fig. 23. Line voltage curves for a NSI.

Fig. 24. Phase voltage curves for a NSI.

Fig. 23 and Fig. 24 show that the peak values of the upper and lower line voltage are about 155V and 310V; the peak values of the phase voltage are about 90V and 180V; and the frequencies of the upper and lower line voltage are 60Hz and 50Hz, respectively.
Fig. 25 shows that the peak line current of the upper load is about 1.4A, and that the peak line current of the lower load is about 2.9A. The frequencies of the upper and lower line current are 60Hz and 50Hz, respectively. In Fig. 26, the THDs of the upper and lower line current are 4.09% and 4.04%, respectively.

From Fig. 27 and Fig. 28, it can be seen that the active powers are about 155W and 700W for the upper and lower load, and that the reactive powers are about 115var and 435var for the upper and lower load, respectively.

From Fig. 29, it can be seen that the maximum errors of the active power are about 5W and 20W for the upper and lower load; the steady-state errors of the active power are smaller than 0.5W and 4W for the upper and lower load; and the transition processes of the active power are about 25ms and 60ms for upper and lower load, respectively.
V. CONCLUSION

This paper proposes a novel direct power control method for a NSI. In the proposed method, a switching function model and an equivalent circuit model for a NSI have been built in αβ coordinates. Then, a novel current observer is proposed, and a novel direct power control method for a NSI has been proposed with only two voltage sensors. Simulation and experimental results show the effectiveness of the proposed method.

In further research, a NSI will be used in a power quality conditioner, and the proposed equivalent circuit model and the method will be used in a unified power quality conditioner to improve the power quality of a power grid.

ACKNOWLEDGMENT

This work was supported by the Universities Science and Technology Fund Planning Project of Tianjin: [Grant Number 20130419]; Tianjin Research Program of Application Foundation and Advanced Technology: [Grant Numbers 15JCQNJC04500, 16JCQNJC04200 and 16JCTPJJC49600]; Tianjin Science and Technology Support Project: [Grant Numbers 15zczdsf00080].

REFERENCES

AC/AC converter,” in Power Electronics, Drive Systems and Technologies Conference, pp. 5-9, 2011.


Lei Pan received his Ph.D. degree from the Hebei University of Technology, Tianjin, China, in 2014. He is presently working as an Associate Professor at Tianjin Chengzjian University, Tianjin, China. His current research interests include power converters and motor drives.

Jiuru Zhang received her B.S. degree from Tianjin Chengjian University, Tianjin, China, in 2016, where she is presently working towards her M.S. degree. Her current research interests include power converters and motor drives.

Kai Wang received the B.S. degree from Tianjin Chengjian University, Tianjin, China, in 2017. He is currently a master degree candidate of Tianjin Chengjian University. His research interests are power converters and motor drives.

Beibei Wang received his B.S. and M.S. degrees in Electrical Engineering and Power Electronics from Liaoning Technical University, Liaoning, China, in 2006 and 2009, respectively. He is presently working as a Lecturer at Tianjin Chengjian University, Tianjin, China. His current research interests include power converter systems and electrical motor drives.

Yi Pang received his Ph.D. degree from Nankai University, Tianjin, China, in 2015. He is presently working as a Lecturer at Tianjin Chengjian University, Tianjin, China. His current research interests include power converter systems and electrical motor drives.

Lin Zhu received her B.S. degree from the Tianjin University of Commerce, Tianjin, China, in 2008; and her M.S. and Ph.D. degrees from Tianjin University, Tianjin, China, in 2010 and 2015, respectively. She is presently working as a Lecturer at Tianjin Chengjian University, Tianjin, China. Her current research interests include neural-network-based methods for microwave device modeling and circuit design, and the development of a neural-network-based circuit simulator.