An Effective Control Technique for Medium-Voltage High Power Induction Motor Fed by Cascaded Neutral Point Clamped Inverter

Ge Baoming, School of Electrical Engineering, Beijing Jiaotong University, Beijing 100044, China
Email: gebaoming@tsinghua.org.cn
Fang Zheng Peng, Department of Electrical & Computer Engineering, Michigan State University, East Lansing, MI 48824, USA

Abstract - An effective control solution for speed sensorless vector control of induction motor fed by a cascaded neutral point clamped (NPC) inverter is proposed, where the novel SPWM pulse rotation control approach provides a simple way to implement vector control for induction motor when the cascaded NPC inverter is employed. The single carrier is used in the novel pulse control, instead of the multi-carriers, which makes the voltage modulation simpler and the hardware burden decrease. It is of great benefits to the cascaded NPC inverter, i.e., the output voltages and powers of all inverter modules, and two series-capacitor dc voltages of each inverter module are perfectly balanced, and a low switch frequency of all inverter modules provides a synthesized high frequency PWM phase voltage. A rotor flux-oriented vector control is combined with back EMF-based MRAS speed estimation, which results in a speed closed-loop control. The voltage sensors together with the filters of changeable parameters ensure the precision of speed estimation for the whole frequency range. The proposed scheme is applied to control an 800kW 4160V induction motor fed by the 1MV A 6000V 17-level cascaded NPC inverter, where the controller mainly consists of one DSP and one CPLD. The experimental results verify the proposed scheme.

Keywords: cascaded multilevel inverters, vector control, speed estimation, neutral point clamped inverter, pulse rotation.

1 Introduction

Medium voltage adjustable speed drive (MV-ASD) systems offer significant advantages in a wide range of industrial applications such as fan, pump, and many improved process control systems with higher efficiencies combined with energy savings. The induction motor (IM) drives fed by a cascaded multilevel inverter (CMI) presents an attractive solution, because the induction motors are cheapest and rugged, also the CMI utilizes small inverter bridges with relatively low voltage to synthesize and reach high voltage, thus is more suitable for high-voltage, high power applications [1-8]. At present, the cascaded H-bridge inverter is the best-selling product in the MV-ASD market worldwide. However, the need of an excessive number of transformer windings presents its main drawback. The cascaded neutral point clamped (NPC) inverter was proposed to alleviate this problem, which is based on the NPC 5-level inverters [9]. But it is crucial in practical uses to control the power and voltage balance among all modules, and the voltage balance between the series-capacitors in each NPC bridge module.

Vector control may ensure the induction motor drive of a high performance, but the required speed sensor often experiences failure. It is always desirable to implement a speed sensorless control in terms of cost reduction and reliability improvement. Several methods of speed estimators have been proposed to replace the speed sensor in the recent years, such as the rotor flux-based MRAS and back EMF-based MRAS speed estimators, slip frequency-based estimator, observer-based estimator, and so on [10-12]. Among the approaches above, the MRAS speed estimators are most attractive approaches, due to their design simplicities. The back EMF-based MRAS speed estimators can cover a very low speed range, where the back EMF observer is used in the reference model to avoid a pure integration. It is important to measure accurate terminal voltages of the IM for the precision of speed estimation. However, the voltage measurement with a constant voltage ratio can not satisfy both requirements in very low frequency, and also very high frequency. Definitely it will be an ideal method to use different voltage ratios for the low and high frequencies, respectively.

This paper presents a MV-ASD system, where a MV induction motor is fed by a cascaded NPC inverter. The back EMF-based MRAS speed estimator is employed in the vector control to fulfill a speed sensorless vector control, where the different voltage ratio is designed in the voltage sensing unit for the purpose of ensuring the precision of voltage sensing in the low and high frequencies, moreover the
low pass filter is used to eliminate harmonics of PWM voltages from the cascaded inverter. An effective SPWM method for the cascaded NPC inverter utilizes the single carrier modulation, pulse encode, and pulse rotation, which simplifies the implementation of whole control system and guarantees the power and voltage balance among all NPC bridge modules, especially the voltage balance between the series-capacitors in each NPC module. The proposed control scheme is applied to control a developed 1MVA 6000V 17-level cascaded NPC inverter to drive an 800kW 4160V induction motor, where the controller mainly consists of one DSP and one CPLD. The experimental results verify the proposed scheme of speed sensorless vector control for IM fed by cascaded NPC inverter.

2 Cascaded NPC Inverter-based IM Drive System

Fig. 1 (a) shows the system configuration of the 1MVA 6000V cascaded multilevel inverter-based IM drive system, in which an IM is fed by a 3-phase cascaded inverter in the Y connection. Each phase has 4 cascaded NPC inverter modules, and each inverter module shown in Fig. 1 (b) is capable of producing 5 different voltage levels, and therefore generating 17 levels in each phase to neutral and 33 levels in phase to phase voltage. For each NPC module, we define three control signals named as K1, K2, and K3, and the 8 switching signals named as Sa1, Sa2, Sb1, Sb2, Sa1’, Sa2’, Sb1’, and Sb2’, then Table I shows the output voltages and their switching states, along with which capacitor is used when applicable, where the high level 1 is assigned as OFF, and the low level 0 as ON. Total 12 control signals control 32 switches each phase. Two series-capacitor voltages should be balanced during operation, namely the average voltages across the two capacitors should be maintained the same, i.e. \( V_{C1}=V_{C2} \). In addition, the voltage and power of all NPC modules must be kept balance as well.

3 SPWM Technique for Cascaded NPC Inverter

3.1 SPWM Technique

Since each phase has 4 NPC modules and each NPC module has two dc capacitors in the current 17-level cascaded inverter, we can use 16 shifted carriers to produce SPWM signals for the entire 17 levels, i.e., so-called multi-carriers method.

However, the multi-carriers method causes quite a hardware burden due to generating 16 shifted carriers. Instead of that, another method is to use a single carrier to produce the needed SPWM signals by chopping the reference signal into the single carrier range according to the magnitude. To illustrate the basic principle of the SPWM and to make it simpler and less busy in the figure, Fig. 2 shows an example to use 8 shifted carriers and to use a single carrier to generate a 9-level SPWM signal, respectively. However, the obtained PWM pulse sequence should be decoded into different levels and distributed among the NPC modules of the cascaded inverter. For this purpose, five bits of binary variables \( a, b, c, d, \) and \( e \) are defined, where \( a=0 \) presents the positive half cycle of the reference signal, \( a=1 \) as the negative half; \( b=0 \) denotes the positive slope of the reference signal, \( b=1 \) as the negative slope; \( c, d, \) and \( e \) represent level number from 000 to 111.

Fig. 3 is used as a general case to calculate the duty cycle, where \( M \) denotes modulation in-
dex, $T_c$ is the period of the carrier, $t_1$ and $t_2$ are two adjacent time instants corresponding to the maximum and minimum values of the carrier, respectively, and $Q$ denotes the $Q$th level, where $1 \leq Q \leq 8$. Then the duty cycle, $t_{on}$, is calculated as

$$t_{on} = 4T_c \left[ |M \sin \omega_1| + |M \sin \omega_2| - \frac{Q-1}{8} \right]$$

(1)

![Fig. 2 9-level modulation.](image)

### Table I  Switching States, Output Voltages, and Capacitors Used

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<th>K1</th>
<th>K2</th>
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<th>Sb2</th>
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<th>Sa2'</th>
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3.2 Pulse Rotation

Pulse rotation is needed for balancing the capacitor voltages and module powers. We define the binary variable $x$ as a pulse sequence, namely, $x=1$ denotes the high level and $x=0$ as the low level, together with other 5 variables $a, b, c, d,$ and $e$ result in total 64 switching states, which will provide sufficient information for the pulse rotation.

A true value table of the 64 switching states can be obtained, where the voltage balance of series-capacitors in each NPC module is taken into account. From the true value table, the deduced logical functions are summarized as

$$A_{ij} = \overline{a}c + \overline{a}\overline{b}\overline{c}e + \overline{a}b\overline{c}d + \overline{a}c(\overline{b}d + b\overline{d})e$$

$$+ b\overline{c}d\overline{e} + \overline{b}c\overline{d}(a\overline{e}x + \overline{a}ex)$$

$$A_{i2} = \overline{b}\overline{c}d\overline{e} + b\overline{c}d\overline{e}$$

$$A_{i1} = ad + ac\overline{d} + \overline{a}c\overline{d}(e\overline{x} + \overline{e}x) + a\overline{c}d\overline{x}$$

$$A_{i2} = \overline{a}c + \overline{a}\overline{d} + \overline{a}\overline{c}\overline{d}(\overline{b}\overline{e} + be) + a\overline{c}d\overline{e}$$

$$+ \overline{a}\overline{b}\overline{c}d\overline{e} + \overline{a}b\overline{c}d\overline{e}$$

$$A_{i2} = \overline{c}\overline{d} + \overline{b}\overline{c}d\overline{x} + b\overline{c}d\overline{e}$$

$$A_{i3} = ac + \overline{c}d(e\overline{x} + ae) + \overline{a}\overline{c}d\overline{e}$$

$$A_{i4} = \overline{c} + \overline{a}\overline{d} + \overline{a}\overline{c}\overline{d}(\overline{b}\overline{e} + be) + a\overline{c}d\overline{e}$$

$$+ \overline{a}\overline{c}\overline{d}(\overline{b}e + be)$$

$$A_{i5} = \overline{c} + \overline{c}\overline{d} + \overline{b}\overline{c}d\overline{e} + b\overline{c}d\overline{e}$$

where $A_{mn}$ denotes the control signal corresponding to $K_i$ ($i=1, 2,$ and 3) in Table I, the subscript $m$ denotes the $m$th module ($m=1, 2, 3,$ and 4), the subscript $n$ denotes the $n$th control signal ($n=1, 2,$ and 3). The symbol “+” denotes a logical relationship OR.

The outputting pulse is rotated one time each pulse, the modules A1, A2, A3, and A4 will
be rotated at the same time. Also each series-capacitor used in each module will be rotated, so that it can alternately use different capacitor of the same module when only one capacitor is used, which ensures the series capacitor voltage balance.

The following steps should be finished for the pulse rotation:

1) Every new pulse coming will incur variable b reverse.
2) Four sets ($A_{1n}, A_{2n}, A_{3n}$, and $A_{4n}$) of control signals calculated from the logical functions above should be rotated each pulse.

### 4 Speed Sensorless Vector Control for MV IM Drive

In the stationary frame the motor model can be written by

$$\mathbf{v}_s = \mathbf{i}_s R_s + L_m^2 p \mathbf{i}_m / L_r + L_s \delta \mathbf{i}_s$$  \hspace{1cm} (2)

$$0 = \frac{1}{T_s} (\dot{\mathbf{i}}_m - \mathbf{i}_s) - j \omega_\delta \mathbf{i}_m + p \mathbf{i}_m$$  \hspace{1cm} (3)

where $L_m$ is the mutual inductance; $R_s$ is the stator resistance; $L_s$ and $L_r$ are respective stator and rotor self-inductances; and $\delta = 1 - L_s^2 / L_r L_r$. $\mathbf{v}_s$, $\mathbf{i}_s$, and $\mathbf{i}_m$ are stator voltage, current, and magnetizing current, respectively. Moreover, there is

$$\mathbf{v}_s = \mathbf{v}_{sd} + j \mathbf{v}_{sq}$$

where the subscripts $d$ and $q$ denote the respective $d$-axis and $q$-axis components. $\mathbf{i}_s$ and $\mathbf{i}_m$ have the similar expression with $\mathbf{v}_s$. $\omega_\delta$ is the rotor speed; $T_s$ is the rotor circuit time constant; $p$ is differential operator.

Define

$$L_m' = L_m^2 / L_r$$

Then the back EMF vector can be obtained by

$$\mathbf{e}_m = \mathbf{v}_s - \mathbf{i}_s R_s - L_s \delta \mathbf{i}_s$$  \hspace{1cm} (4)

$$\mathbf{e}_m = -\frac{L_m'}{T_r} (\dot{\mathbf{i}}_m - \mathbf{i}_s) + j \omega_\delta L_m' \mathbf{i}_m$$  \hspace{1cm} (5)

where

$$\mathbf{e}_m = L_m' p \mathbf{i}_m$$

Rotor speed is unknown in (5), so it can be rewritten as

$$\dot{\mathbf{e}}_m = -\frac{L_m'}{T_r} (\dot{\mathbf{i}}_m - \mathbf{i}_s) + j \omega_\delta L_m' \mathbf{i}_m$$  \hspace{1cm} (6)

Thus, $\dot{\mathbf{e}}_m$ will equal to $\mathbf{e}_m$ when the estimated rotor speed equals to the actual rotor speed. So, the adaptive method to identify rotor speed can be

$$\dot{\omega}_r = \left( k_p + \frac{k_i}{p} \right) (\mathbf{e}_m \otimes \mathbf{e}_m)$$  \hspace{1cm} (7)

where $k_p$ and $k_i$ are proportional and integral constants, respectively [10].

Referring to the rotor flux frame, the dynamic model of the IMs becomes

$$\mathbf{v}_{r'} = (R_s + \delta L_s p + j \alpha \delta L_s) \mathbf{i}_{r'} + \frac{L_m'}{L_r} (j \alpha + p) \lambda_{r'}$$

$$p \lambda_{r'} = R_r \frac{L_m'}{L_r} \dot{\mathbf{i}}_{r'} + \left[ -j \alpha - \frac{R_r}{L_r} \right] \mathbf{\lambda}_{r'}$$

where $\alpha$ and $\alpha'$ are respective synchronous and slip angular frequencies; $\lambda_r$ is the rotor flux linkage, and the superscript $r$ denotes the rotor flux frame; $R_r$ is the rotor resistance.

According to the rotor flux-oriented control, we have

$$\lambda_{dr} = \lambda_{dr}' + j \lambda_{qdr}' = \lambda_{dr}'$$

The slip frequency is

$$\omega_s = \omega_s / \left[ T_r, \lambda_{dr}' / L_m \right]$$

The $d$-axis component of stator current is

$$i_{ds} = (1 + T_r p) \lambda_{dr}' / L_m$$

The magnetizing current can be estimated as

$$i_{mr}' = \lambda_{dr}' / L_m = i_{ds} / (1 + T_r p)$$

The torque is

$$T_e = \frac{3 L_m}{4 L_r} p \lambda_{dr}' i_{qdr}$$

where $P$ is the number of poles.

When the current closed-loop control method is employed in 3-phase system to force the actual currents track the desired currents, Fig. 4 shows its rotor flux-oriented vector control for IM drives, where $G_c$ is a speed regulator with the proportional and integral type, and its output is

$$T_e = k_p e_m + k_i \int_0^t e_m dt$$

where $e_m$ is the speed tracking error. $k_p$ and $k_i$ are the proportional and integral gains, respectively.

Therefore, the speed closed-loop transfer function from the reference speed to the actual speed is

$$\frac{\omega_s (s)}{\omega_s' (s)} = \frac{k_p s + k_i}{J s^2 + k_p s + k_i}$$

where $J$ is the moment of inertia. A desired speed response can be achieved by designing the control parameters $k_p$ and $k_i$. 

5 Experimental Results

An 800kW 4160V IM is fed by the developed cascaded NPC inverter shown in Fig. 5 (a) and the CPLD XC95288 cooperates with the DSP TMS320LF2407 to implement the proposed control technique, as shown in Fig. 5 (b). The flux-oriented vector control, the current closed-loop control, speed estimation, the speed tracking, and the SPWM pulses are produced in the DSP, and the pulse encode and pulse rotation are fulfilled in the CPLD. The line voltage contains plenty of harmonics due to PWM inverter, which will influence the precision of speed identification, therefore a first-order low-pass filter is used to detect the fundamental components of stator voltages. Moreover, the stator voltage is extremely small at low speed, and extremely large at high speed. A voltage sensing method with constant voltage ratio is not suitable for both sides, and the adjustable output range is expectant. For this purpose, the optical signal from DSP-based board controls three switches’ states such as turn-on and turn-off in the designed voltage sensing circuit, which will change the voltage sensing circuit for different frequency ranges. In particular, three switches will turn on when the voltage frequency is over 6 Hz. On contrary, the switches will turn off when below 6 Hz. The input voltage of voltage sensor is restricted under 250 V. Fig. 5 (c) gives the prototype of voltage sensing circuit.

Fig. 6 (a) and (b) show the 9-level and 3-level voltages, respectively, which are from the cascaded NPC inverter that operates in the proposed pulse rotation control. The pulse voltage $v_{c1}$ is from one of the four modules in phase “c”. The phase-neutral voltage waveform is composed of the sum of the pulses from all four modules. While the effective switching frequency of the phase-neutral voltage is 2 kHz, the switching frequency of each module is limited to 500 Hz.
Fig. 6  Pulse rotation in the cascaded NPC inverter.

Fig. 7  Output voltages and currents at 60 Hz.

Fig. 8  Speed sensorless induction motor drive runs below 6 Hz.

The voltage sensing circuit change will adjust the range of measured voltage, which will improve the voltage sensing precision. At the same time, the control parameters should be changed due to the change of the detected voltage. Fig. 9 shows the starting dynamics in output of voltage sensor, the estimated speed, the actual speed, and phase current when the desired motor speed is 394 r/min (13.1 Hz), where the voltage sensing circuit will be switched at 6 Hz. The output voltage of voltage sensor will change lot when its circuit is switched, but the estimated speed still can match the actual speed very well. Fig. 10 presents the experimental results when the speed increases. At the beginning, the motor drive runs below 6 Hz, then the desired speed increases to 490 r/min (16.3 Hz). Fig. 11 presents the drive’s responses when its speed decreases to 85 r/min (2.84 Hz), from a frequency more than 6 Hz.

It is obvious that the estimated speeds always are identical to the actual speed when the voltage sensor circuit is switched.

Fig. 12 shows the experimental results when the desired speed is a trapezoid waveform,
where the motor drive starts from standstill, and passes through 180 r/min (6 Hz), and operates in 403 r/min (13.4 Hz) for 25 s, and decelerates, and passes through 180 r/min (6 Hz), and then stops. Fig. 13 presents the steady responses in output of voltage sensor, the estimated speed, the actual speed, and phase current when the desired motor speed is 348 r/min (11.57 Hz). We find that the drive works very well without speed sensor and the estimated speed presents perfect precision during whole operation.

6 Conclusions

This paper presented a speed sensorless vector control technique for MV high power IM drives fed by a cascaded NPC inverter. For this purpose, a novel SPWM technique was proposed, where the single-carrier together with pulse code and pulse rotation provided an effective pulse control method for the cascaded NPC inverter. It simplified the voltage modulation and made the hardware burden decrease, also the great benefits included not only that the output voltages and powers of all inverter modules, and two series-capacitor dc voltages of each inverter module are perfectly balanced, moreover a low switch frequency of all inverter modules provided a synthesized high frequency PWM phase voltage. On the basis of the novel SPWM technique, a rotor flux-oriented vector control was combined with back EMF-based MRAS speed estimation, and a speed closed-loop control was implemented to control the rotor speed. A current closed-loop control overcome the internal voltage drop of the inverter. The experimental results verified the proposed speed sensorless
vector control technique for MV high power IM drives fed by a cascaded NPC inverter.

References