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Three-vector Model Predictive Control without Weight Coefficient based on Three-phase Neutral Point Clamped Inverter

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Abstract: Three-phase three-level neutral point clamped inverters are widely used in new energy fields such as photovoltaic power generation and wind power generation by virtue of their multilevel topology. The traditional three-vector model predictive control can solve the problems of current harmonics and unfixed switching frequency that exist in the single-vector model predictive control, but the traditional three-vector model predictive control still has the problems of large computation and difficult design of weighting factors. Based on this, this paper proposes a three-vector model predictive current control without weighting factors. The method firstly selects the voltage vectors by dividing the sectors to eliminate the weighting factors, secondly adopts the principle of non-differential beat control to obtain the reference voltage and the optimal sectors, and finally calculates the action time of each vector and designs the switching sequences by means of the cost function. The experimental results show that the proposed control strategy can quickly regulate the midpoint potential balance, and the current harmonic content is reduced to 3.5%.

Keywords : Three-phase NPC inverter; Model predictive current control; Neutral point potential balance; Weight factor; Delay compensation;

1. Introduction

Compared with two-level inverters, three-level Neutral Point Clamped (NPC) inverters have the advantages of low output harmonics and low voltage stress. These inverters are widely used in the fields of high-voltage frequency control and grid-connected generation of renewable energy, and have become a research hotspot in domestic and international academic circles[1-3].

Conventional control methods mainly include proportional-integral control[4], space vector pulse width modulation control[5] and model predictive control[6]. Among them, model predictive control, as an advanced control strategy with the advantages of multivariable control, prediction of future outputs, and rolling optimization, has become a hot research topic in the field of inverter control in recent years[7-9].

Model predictive control in power electronics can be divided into Continuous Control Set Model Predictive Control[10] (CCS-MPC) and Finite Control Set Model Predictive Control[11] (FCS-MPC). FCS-MPC is able to achieve multi-objective optimization by traversing all the voltage vectors of the inverter and using the cost function to determine the optimal output. This control method does not require a modulator and determines the optimal voltage vectors for the output directly through the prediction model and the cost function[12]. However, only one voltage vector is output in each cycle, leading to issues such as variable switching frequency and large current ripple.

To solve the above problems, researchers have proposed the Three Vector Model Predictive Current Control (TV-MPCC), which applies two non-zero voltage vectors and one zero voltage vector in one control cycle. This approach effectively reduce the current ripple and improve the system steady state performance^[13]. However, its high switching frequency and substantial computational requirements make it more demanding on the processing chip^[14]. At the same time, in order to realize multiple control objectives such as output

current control and neutral point potential balance control, it is also necessary to design reasonable weighting factors, and there is no mature weighting factor design theory^[15]. Therefore, how to reduce the computational amount and design the weighting factor under the premise of ensuring the steady state performance of the system has become a research hotspot of three-vector model predictive current control.

In practice, system delays and time delays are common. At the present time, it is common to use Lagrangian extrapolation for delay compensation to reduce the error caused by compensating part of the delay, but this method has a large peak when there is a step change. Based on this, this paper introduces a phase angle compensator, which is simpler compared to the Lagrangian extrapolation method and can effectively mitigates the issue of large peaks at step changes.

In summary, this paper proposes a TV-MPCC method without weight factor based on a three-phase NPC inverter. By dividing the sectors to select the corresponding voltage vectors and designing the switching sequence, the midpoint potential balance control can be realized while eliminating the weighting factor. By converting the reference current to the reference voltage, the optimal sector and the optimal voltage vector can be obtained without the need of optimization search, which can reduce the computation amount. Aiming at the phase lag caused by the delay of digital control, a phase angle compensator is introduced to reduce the error caused by the delay. Finally, the experiments compare and analyze the differences between FCS-MPCC, traditional TV-MPCC and the control strategy proposed in this paper in terms of steady state performance, dynamic performance and midpoint potential balancing ability, which verifies the feasibility of the proposed method.

2. Mathematical modeling of three-phase NPC inverters

The three-phase NPC inverter topology is shown in **Fig1**. As shown, the inverter consists of a DC side, NPC topology and an AC side. The DC side consists of a DC input power supply U_{dc} and two capacitors C_1, C_2 . The AC side consists of an LC filter and a resistive load. The NPC topology consists of four switching devices $S_{x1}, S_{x2}, S_{x3}, S_{x4}$ and two clamping diodes D_{x1}, D_{x2} ($x=a,b,c$).

Define the switch function $S_x (x=a,b,c)$ as follows

$$S_x = \begin{cases} 1, & S_{x1} = S_{x2} = 1, S_{x3} = S_{x4} = 0 \\ 0, & S_{x2} = S_{x3} = 1, S_{x1} = S_{x4} = 0 \\ -1, & S_{x3} = S_{x4} = 1, S_{x1} = S_{x2} = 0 \end{cases} \quad (1)$$

where: $S_{xi} (i = 1, 2, 3, 4)$ denotes the switching state, $S_{xi}=1$ denotes turn-on and $S_{xi}=0$ denotes turn-off. The inverter can generate a total of 27 switching states, corresponding to 27 voltage vectors. The voltage vectors can be categorized into zero, small, medium, and large vectors. The voltage vector distribution is shown in **Fig2**, which can be divided into six large sectors (I-VI), and each large sector can be divided into four small sectors (1-4).

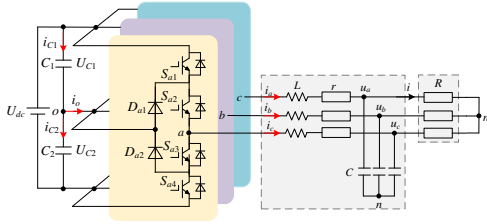


Fig1. Three-phase NPC inverter topology

For the three-phase NPC inverter, the output current of the AC side is the main objective of the study, and the mathematical model needs to be established first. From Kirchhoff's law, the mathematical model of three-phase NPC inverter can be obtained as

$$L \frac{di_x}{dt} = V_x - r i_x - U_{xo} \quad (x=a,b,c) \quad (2)$$

where: L is the filter inductor, r is the filter inductor equivalent resistance, V_x is the output voltage vector, i_x is the current of filter inductor, U_{xo} is the output voltage ($x=a,b,c$).

Since the model in the two-phase stationary coordinate system is simple and easy to control, the above mathematical model is converted to the coordinate system by Clark transformation as follows

$$L \frac{di_{\alpha\beta}}{dt} = V_{\alpha\beta} - r i_{\alpha\beta} - U_{\alpha\beta o} \quad (3)$$

where: $i_{\alpha\beta} = [i_\alpha \ i_\beta]^T$ is the inductor current in the $\alpha\beta$ coordinate system, $V_{\alpha\beta} = [V_\alpha \ V_\beta]^T$ is the output voltage vector in the $\alpha\beta$ coordinate system, and $U_{\alpha\beta o} = [U_{\alpha o} \ U_{\beta o}]^T$ is the load voltage in the $\alpha\beta$ coordinate system.

Assuming that the sampling time of the controller is T_s , discretizing and organizing the dynamic equations using the forward Euler method yields

$$i_{\alpha\beta}(k+1) = \left(1 - \frac{rT_s}{L}\right) i_{\alpha\beta}(k) + \frac{T_s}{L} (V_{\alpha\beta}(k) - U_{\alpha\beta o}(k)) \quad (4)$$

where: $V_{\alpha\beta}(k)$ is the voltage vector corresponding to the 27 switching states.

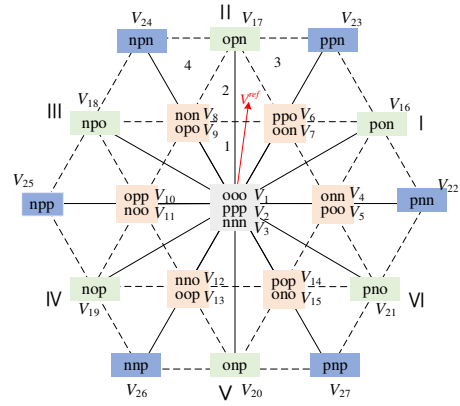


Fig2. Voltage vector

The mathematical model of the current flowing through the DC side capacitor is as follows

$$i_{Cx} = C_{dc} \frac{dU_{Cx}}{dt} \quad (x=1,2), C_1 = C_2 \quad (5)$$

Using the forward Euler method to discretize and simplify the above equation can be obtained as follows

$$U_{C1}(k+1) = U_{C1}(k) + \frac{T_s}{C_1} i_{C1}(k) \quad (6)$$

$$U_{C2}(k+1) = U_{C2}(k) + \frac{T_s}{C_2} i_{C2}(k)$$

According to (6), the voltage value of the DC side capacitor at the moment $k+1$ can be predicted from the current flowing through the capacitor at the moment k as well as the voltage across the DC side.

The neutral point potential offset at the moment $k+1$ is:

$$U_o(k+1) = U_{C1}(k+1) - U_{C2}(k+1) \quad (7)$$

According to Kirchhoff's current law, the neutral point current is

$$i_o(k) = i_{C1}(k) - i_{C2}(k) \quad (8)$$

From the topology in **Fig1**, different operating states of the inverter correspond to different neutral point currents. The neutral point current can be expressed as

$$i_o(k) = \sum_{x=a,b,c} (1 - |S_x(k)|) i_x(k) \quad (9)$$

where: $S_x(k)$ is the switching state at moment k and $i_x(k)$ is the output current at moment k .

Associate(6)(7)(8)(9):

$$U_o(k+1) = U_o(k) + \frac{T_s}{C} \sum_{x=a,b,c} (1 - |S_x(k)|) i_x(k) \quad (10)$$

where: $U_o(k)$ is the neutral point potential at moment k , and $U_o(k+1)$ is the neutral point potential at moment $k+1$.

In order to realize the balance control of the neutral point potential, it is necessary to keep $U_o(k+1)$ near zero.

3. Traditional finite set model predictive current control

For a three-phase NPC inverter, it is usually required to realize the tracking of the AC side current and the equalization control of the DC side capacitor voltage. The conventional FCS-MPCC uses the cost function as follows:

$$g_i(k) = |i_{\alpha}^{\text{ref}}(k+1) - i_{\alpha}(k+1)| + |i_{\beta}^{\text{ref}}(k+1) - i_{\beta}(k+1)| + \lambda |U_o^{\text{ref}}(k+1) - U_o(k+1)| \quad (11)$$

where: $i_{\alpha}^{\text{ref}}(k+1)$ and $i_{\beta}^{\text{ref}}(k+1)$ are the current reference value, $U_o^{\text{ref}}(k+1)$ is the reference value of the neutral point voltage. In order to realize the balance control of the neutral point potential, $U_o^{\text{ref}}(k+1)$ is usually zero, λ is the weighting coefficient of the neutral point potential control.

The traditional FCS-MPCC requires traverse all basic voltage vectors and calculate the cost function, and output the switching state corresponding to the smallest cost function. Each sampling needs to traverse all the voltage vectors, resulting in significant computational load and slow system dynamic response.

Without considering the delay, FCS-MPCC outputs only one optimal switching state. However, the disordered nature of the conventional FCS-MPC may cause the same switching state to be repeatedly output as the optimal state in multiple cycles, resulting in a non-constant switching frequency. This non-constant switching frequency triggers the

dispersion of the output signal spectrum of the NPC three-level inverter, the aggravation of harmonic distortion, and the degradation of control accuracy. This not only increases the current THD, but also increases the complexity of the AC side filter parameter design.

In order to solve the above problems, researchers have proposed the TV-MPCC. The traditional TV-MPCC can reduce the current ripple, but it still needs to design the weighting factor, and the computation volume is large.

4. Three-vector model predictive control without weighting factors and considering delay compensation

4.1. Selection of voltage vector

Associate(9)(10), it can be seen that only both the mid vector and the small vector have an effect on the fluctuation of the neutral point potential. Small vector can be classified into positive and negative small vectors according to the different effects of the small vectors on the mid-point potential. The effects of switching states ppo and oon on the neutral point potential of the DC side of the inverter are shown in **Fig3**.

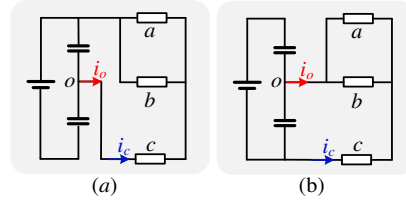


Fig3. Simplified diagram of ppo and oon

From **Fig3**, it can be seen that when the positive small vector ppo acts, the neutral-point current $i_o = i_c$, the upper capacitor charges, the lower capacitor discharges, and the neutral-point potential decreases, while when the negative small vector oon acts, the neutral-point current $i_o = -i_c$, and the neutral point potential increases.

According to the effect of different switching states on the neutral point potential, it can be seen that the positive and negative small vectors have opposite effects on the neutral point potential. Therefore, in one switching cycle, the small vector can be selected according to the state of the neutral point potential to control the balance of the neutral point potential thus eliminating the neutral point potential control weight coefficient in the cost function.

Table 1 Voltage vector selection principle

small sector	vector type	Vector Selection
1	positive	$V_1-V_6-V_9$
	negative	$V_1-V_7-V_8$
2	positive	$V_6-V_9-V_{17}$
	negative	$V_7-V_8-V_{17}$
3	positive	$V_6-V_{17}-V_{23}$
	negative	$V_7-V_{17}-V_{23}$
4	positive	$V_9-V_{17}-V_{24}$
	negative	$V_8-V_{17}-V_{24}$

Taking sector II as an example, Table 1 shows the principle of selecting the voltage vector. The principle of vector selection for other sectors is the same.

4.2. Designing the switching sequence

The traditional FCS-MPC acts on only one voltage vector in one control cycle, and the relationship between the voltage vectors is not fixed, resulting in frequent changes in the switching frequency. In this paper, the switching frequency is fixed by rationally designing the switching sequence.

In order to avoid vector switching, it is necessary to ensure that the first and last voltage vectors are the same when switching states in the sector; at the same time, in order to reduce the switching noise, when switching states, this paper adopts the single-phase bridge-arm switching strategy, i.e., only one-phase bridge-arm switching tube is allowed to act in each control cycle, which makes the average switching frequency fixed.

Table 2 Switching sequence design

small sector	neutral point voltage	switching sequence
1	≥ 0	ooo-opo-ppo-opo-ooo
	< 0	non-onon-ooo-onon-non
2	≥ 0	opn-opo-ppo-opo-opn
	< 0	non-onon-opn-onon-non
3	≥ 0	opn-ppn-ppo-ppn-opn
	< 0	oon-opn-ppn-opn-oon
4	≥ 0	nnp-opn-opo-opn-nnp
	< 0	non-nnp-opn-nnp-non

As can be seen from Table 2, there is no redundant vector in each switching cycle, so the three candidate vectors form a five-segment hairpin, and this five-segment allocation reduces the number of vector switching times and the cumulative number of single-bridge-arm switching-tube actions to four times in one control cycle, and reduces the switching

frequency while fixing the switching frequency, reduces switching losses, reduces switching noise impact on the control system, and enhances the system reliability.

4.3. Time calculations

The FCS-MPC and the traditional TV-MPCC calculates the cost function to find the optimum. In order to reduce the computation, this paper calculates the reference voltage value $V_{\alpha\beta}^{ref}$, and then determines the region where the reference voltage is located. the three voltage vectors of this sector are the optimal voltage vectors.

$$V_{\alpha\beta}^{ref}(k) = \frac{L}{T_s} (i_{\alpha\beta}^{ref}(k+1) - i_{\alpha\beta}(k)) + r i_{\alpha\beta}(k) - u_{\alpha\beta o}(k) \quad (12)$$

Rewriting the traditional current cost function as a voltage cost function

$$g_i = |V_{\alpha}^{ref}(k) - V_{\alpha_i}| + |V_{\beta}^{ref}(k) - V_{\beta_i}| \quad (13)$$

Assuming that the reference voltage vector falls in the sector 2 of the sector II of **Fig2**.

The voltage vectors are ppo, oon, non, opo, opn, according to the principle that the vector action time is inversely proportional to its value function, it can be obtained that the vector action time is:

$$\begin{cases} t_{ppo} = t_{oon} = \frac{g_2 g_3}{g_1 g_2 + g_1 g_3 + g_2 g_3} T_s \\ t_{non} = t_{opo} = \frac{g_1 g_3}{g_1 g_2 + g_1 g_3 + g_2 g_3} T_s \\ t_{opn} = \frac{g_1 g_2}{g_1 g_2 + g_1 g_3 + g_2 g_3} T \end{cases} \quad (14)$$

where: $g_1 = g_{ppo} = g_{oon}$, $g_2 = g_{non} = g_{opo}$, $g_3 = g_{opn}$. They are the value of cost functions under the switching states of ppo, oon, non, opo, and opn, which can be calculated by(13).

This method does not need to perform the value function optimization search. It is only obtains the three optimal voltage vectors after sector judgment, calculates the value function and action time corresponding to the voltage vectors. This method simplifies the calculation process, significantly reduces the calculation volume and improves the overall efficiency of the system while guaranteeing the good steady state performance and dynamic performance.

4.4. Delay compensation

The above analysis is carried out in an ideal state, however, there exists a system delay in a digital control system. If the delay is not taken into account, it will lead to the deviation between the actual output and the predicted output, which will reduce the dynamic and steady state performance of the system. At high switching frequency, the delay will lead to the increase of the current harmonic content. Therefore, it is necessary to compensate for the delay of the system.

At the moment t_k , the current at the moment t_{k+1} is estimated using equation(4). This estimated current value is used as a starting point to store the optimal switching state S_{k+1} , which is predicted at the moment t_{k+1} , with the predicted value of

$$i_{\alpha\beta}(k+2) = (1 - \frac{rT_s}{L})i_{\alpha\beta}(k+1) + \frac{T_s}{L}(V_{\alpha\beta}(k+1) - u_{\alpha\beta o}(k+1)) \quad (15)$$

From the principle of non-differential beat control, the voltage reference value can be obtained as:

$$V_{\alpha\beta}^{ref}(k+1) = \frac{L}{T_s}(i_{\alpha\beta}^{ref}(k+2) - i_{\alpha\beta}(k+1)) + ri_{\alpha\beta}(k+1) - u_{\alpha\beta o}(k+1) \quad (16)$$

Rewriting the cost function:

$$g_i = |V_{\alpha}^{ref}(k+1) - V_{\alpha_i}| + |V_{\beta}^{ref}(k+1) - V_{\beta_i}| \quad (17)$$

In a digital control system, the DSP computation delay and the PWM output delay dominate the system delay. The digital control delay does not interfere with the amplitude-frequency characteristics of the system, but rather causes phase lag, as shown by the red dashed line in **Fig4**.

In order to compensate for the delay error caused by the phase lag, this paper adopts a simple but effective design method, i.e., an additional phase angle compensator is introduced into the synchronization process, as shown by the blue dashed line in **Fig4**.

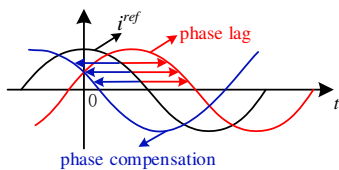


Fig4. Phase angle compensation diagram

The phase angle compensator is expressed as

$$\varphi = NT_s w_0 \quad (18)$$

where: $N=T_d/T_s$, where T_s is the sampling period and T_d is the delay period, $w_0 = 2\pi f$, $f=50\text{Hz}$.

As known from the previous text, $T_d = 2T_s$, thus $N=2$. The reference current is expressed as

$$i_{\alpha\beta}^{ref}(k+2) = i_{\alpha\beta}^{ref}(k)e^{j\omega 2T_s \alpha_0} \quad (19)$$

With this phase angle compensation scheme, the error due to delay can be effectively reduced. Fig5 shows the block diagram of the proposed TV-MPCC control.

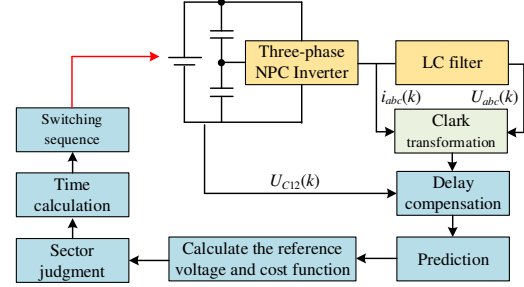


Fig5. Block diagram of TV-MPCC control

5. Experimental verification

This paper establishes a comparative experimental analysis of Method 1(FCS-MPCC), Method 2(traditional TV-MPCC) and Method 3(TV-MPCC proposed). To verify the feasibility of the proposed algorithm, this paper builds a platform with TMS320F28335D-type DSP as the core. The three-phase NPC inverter experimental platform as shown in **Fig6**.

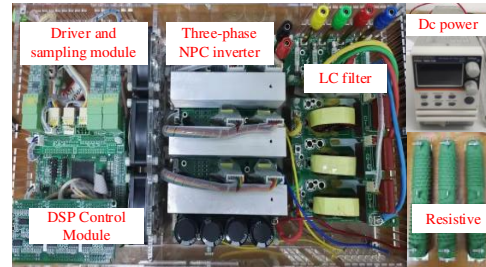


Fig6. Experimental platform

The parameters of the main circuit are: $C_1=C_2=2040 \mu\text{F}$, filter inductor $L=1\text{mH}$, filter inductor equivalent resistance $r=0.001\Omega$, filter capacitance $C=4.7 \mu\text{F}$, the DC input voltage is $U_{dc}=40\text{V}$, the output resistance load $R=4\Omega$. The fundamental frequency is 50Hz, and the sampling frequency of the controller is 20kHz.

5.1. Steady-state experiment

Under the three control methods, the

inverter output current and the THD are shown in **Fig7**, and the output voltage is shown in **Fig8**. The reference current is given as $3\sqrt{2}\cos(\omega t)$ A.

From **Fig7(a)**, the root mean square (RMS) of the output current of the Method 1 is 3.21 A, and the THD is 14.0%. The current harmonic distribution is large. From **Fig7(b)**, the RMS of the output current of the Method 2 is 3.08A and THD is 8.4%. From **Fig7(c)**, the RMS of the output current of the Method 3 is 3.05 A, and the THD is 3.5%. The Method 3 has the best tracking effect and effectively reduce current harmonics. As shown in **Fig8**, the output voltage of the Method 3 is more stable, indicating that the proposed control strategy has a better improvement effect on the system steady state performance.

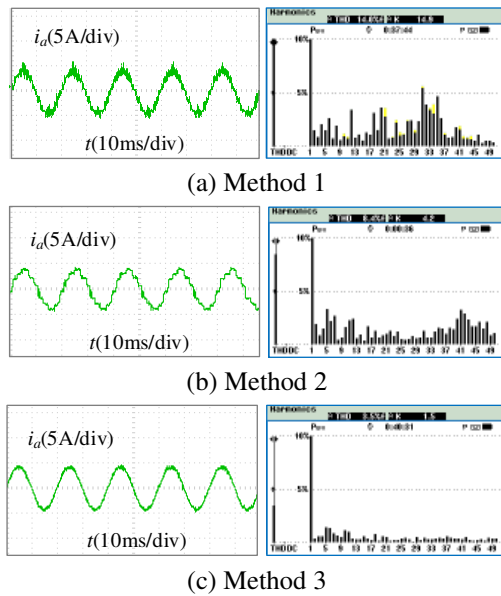


Fig7.Output current

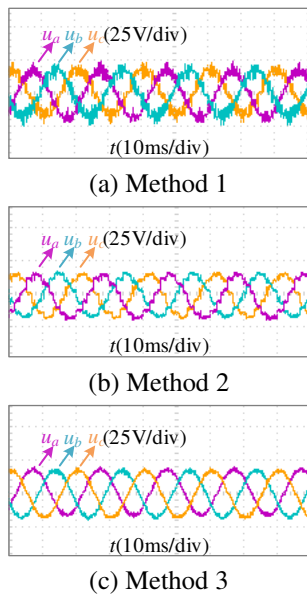


Fig8.Output Voltage

Fig9 shows the output line voltage U_{ab} waveform of the inverter before filtering. Because the proposed TV-MPCC rationally designs the switching sequence based on the voltage vector selection principle. From **Fig9**, it can be seen that the output three-level has no higher level jumps.

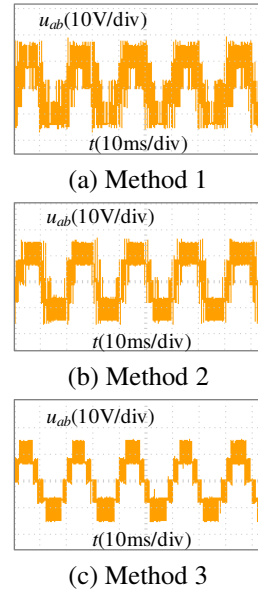
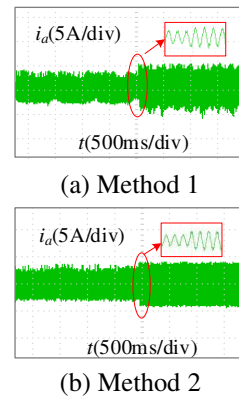
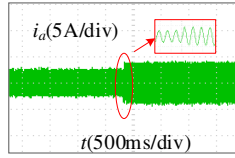


Fig9.Unfiltered line voltage

5.2 Dynamic experiment

In order to verify and compare the dynamic response performance of the three control methods, the given reference current is abruptly added from $2\sqrt{2}\cos(\omega t)$ A to $3\sqrt{2}\cos(\omega t)$ A. **Fig10** shows the current waveform of the inverter with the sudden change of the reference current. As shown in **Fig.10**, three methods can realize fast current tracking, but the Method 3 has smaller current harmonics, faster response speed.





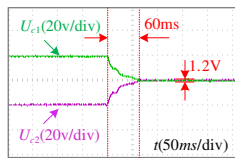
(c) Method 3

Fig10. Output Current

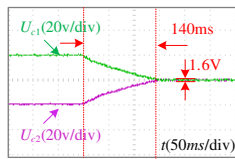
5.3. Neutral point Potential Equilibrium Effect

The initial capacitor voltages are set to 40V and 0V, to simulate the neutral point potential imbalance. **Figure 11** presents the neutral point voltage dynamic waveform under the three control algorithms.

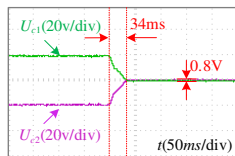
As shown in **Fig. 11**, three methods can realize the capacitor voltage equalization control. From **Fig11(a)**, under the Method 1, the capacitor voltage is discharged 20 V after 60 ms. The maximum deviation after equalization is about 1.2 V. From **Fig11(b)**, the capacitor voltage is discharged 20 V after 140 ms. The maximum deviation after equalization is about 1.6 V. From **Fig11(c)**, the capacitor voltage is discharged 20 V after 34 ms. The maximum deviation after equalization is about 0.8 V. The neutral point potential equalization rate is faster and the equalization effect is stronger under the Method 3.



(a) Method 1



(b) Method 2



(c) Method 3

Fig. 11 Neutral Point Voltage

The experimental results under the three control strategies are shown in Table 3. From Table 3, the Method 3 has the best tracking effect, the smallest current error, the smallest

THD, the fastest neutral point potential equalization and the best balancing effect.

Table 3 Comparison of experimental results

	Steady-state experiment		Neutral point potential	
	current RMS	current THD	time	deviation
Method 1	3.21A	14.0%	60ms	1.2V
Method 2	3.08A	8.4%	140ms	1.6V
Method 3	3.05A	3.5%	34ms	0.8V

6. Conclusion

This paper proposes an optimized three-vector multi-objective current prediction control method for three-phase NPC inverters. The method aims to address the difficulties in selecting the weighting factors, the large computation volume and the high harmonic content in the FCS-MPCC and the traditional TV-MPCC. The experimental results show that the proposed TV-MPCC has the following superiorities:

- (1) Elimination of the weighting factor in the cost function and realization of fast neutral point potential equilibrium.
- (2) It reduces the computation of the algorithm by calculating the reference voltage to select the optimal sector and the optimal voltage vector, avoiding the calculation of rolling optimization search.
- (3) It has fast dynamic response speed and certain anti-interference ability, and can quickly adjust the output with good dynamic performance and parameter robustness;
- (4) Effectively reduces the total harmonic distortion rate, specifically, the THD is reduced from 14.0% to 3.5% in the traditional algorithm.

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