Retraction

Retraction: The improvement of model predictive control based on three-level three phase inverter (J. Phys.: Conf. Ser. 1601 042032)

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This article has been retracted at the request of the authors, who have identified errors in the paper which affect the scientific validity of the work. The authors specifically state "the coefficient of i\textsuperscript{T1} should be (1-2k) instead of (2k-1) in the paper. This error is likely to cause conclusions to violate the actual situation."

The authors apologise to the readers, conference organisers and IOP Publishing for these errors.

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The improvement of model predictive control based on three-level three phase inverter

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Abstract. To improve the dynamic response of the inverter, the conventional proportional-integral controller and the modulation part are substituted by the finite-control step model predictive control (FCS-MPC) strategy. Unlike the two-level inverter, the drifted neutral point potential is the issue that cannot be neglected. To resolve this issue, the cost function is modified in this paper. Besides, as all of the voltage vectors are employed during optimal research progress in FCS-MPC, the switching frequency is variable, which burdens the filter. The complex calculation progress also burdens the microcontroller. According to the conventional modulation algorithm, the voltage vectors are divided into several groups and the optimal calculations are simplified with them. Based on that modification, the switching frequency is fixed. Finally, the simulation and the experimental results are demonstrated, they reveal the validity and the merits of the proposed algorithm.

1. Introduction

With the increasing awareness of environmental protection and the reduction of fossil fuel reserves, distributed generation (DG) such as photovoltaic and wind power has been vigorously developed [1, 2]. In order to supply the energy collected by the micro source to the local power source and the power grid in the form of electrical energy, the concept of the micro grid has been proposed, and it has been extensively studied by relevant staff and experts. As the interface of the distributed generation system, grid-connected inverters play an important role in power conversion and maintenance of power quality of the power grid [3,4].

Among the three-phase three-level inverters, the NPC three-level topology is currently the most widely used topology. In this topology, the midpoint voltage will shift due to load and modulation mode during operation. Therefore, the adjustment of three-level midpoint clamp type midpoint voltage balance has always been a research hotspot of scholars at home and abroad, and a large number of control strategies for adjusting midpoint voltage have been proposed [5-7]. Literature [5] proposes that additional hardware circuits can be used to adjust the midpoint potential. However, this method increases the size of the system and has limited adjustment capability. The current main midpoint potential adjustment method is a software adjustment strategy. The literature [6] pointed out that the midpoint current generated by the mutually redundant small voltage vectors has the opposite effect on the midpoint potential. Using this characteristic, the midpoint potential is adjusted by a small vector. The literature [7] analyses the modulation function and midpoint potential adjustment on the basis of carrier modulation. Literature [8] proposed to control the midpoint potential in the case of a single modulated wave by adjusting the duty cycle. However, the single modulation wave modulation method has the problem of midpoint fluctuation. Literature [9, 10] proposed a dual modulation wave...
modulation method to adjust the midpoint potential and keep the midpoint potential constant. However, this dual-modulated wave approach is computationally complex and has no unique solution. To overcome this problem, it is necessary to artificially add constraints.

In order to ensure the balance of the midpoint potential, the accuracy of adjusting the balance of the midpoint potential needs to be improved. Reference [4] pointed out that the control link needs to go through AD sampling, digital calculation and other processes, and the adjustment of the duty cycle needs to be delayed by one switching cycle before it can be executed. Therefore, the imbalance state of the midpoint voltage cannot be adjusted in time, and the offset of the midpoint voltage is enlarged. In order to achieve the adjustment of the midpoint potential, the literature [11,12] applies model predictive control to the rectifier, and proposes to realize the control of the converter in a two-phase rotating coordinate system. Reference [13] realizes the control of the converter by using the switching time as the control quantity in the model predictive control, and realizes continuous adjustment. Reference [14] takes discrete voltage vectors as control elements and uses finite set model predictive control to control the three-phase inverter. In the above literature, in the process of model predictive control, all voltage vectors are used as variables for optimal control, and tracking of a given voltage is achieved by this method. Unlike the two-level converter system, the three-level inverter voltage vector increases, which increases the burden on the controller.

2. Three-phase three-level midpoint potential adjustment

The topology of the three-phase three-level midpoint clamped inverter is shown in Figure 1. In order to simplify the analysis process, the following assumptions can be made: (1) the two capacitors on the DC side are equal, namely; (2) the switch is in an ideal state. Taking a bridge arm in the figure as an example, the four switching devices on the bridge arm are respectively.

When the two capacitor voltages on the DC side are half of the DC voltage, when S1 and S2 are turned on, S3 and S4 is turned off and the output phase voltage \( V_{ao} = 0.5V_{dc} \), where \( x = A,B,C \) is recorded as P; when the S1, S2 is turned on and S3, S4 is turned off, the output phase voltage \( V_{ao} = 0 \) which is recorded as the state O; when S3, S4 is turned on, S1 and S2 is turned off, the output phase voltage \( V_{ao} = -0.5V_{dc} \). It can be seen that the output voltage vector can be described by the switching state, and the spatial voltage vector distribution is shown in Figure 2.

Reference [14] divides the voltage vector into small vector, medium vector and large vector according to the magnitude of the vector. Among them, the small vector and the middle vector can affect the DC midpoint voltage. The two small vectors that are redundant with each other have opposite effects on the midpoint potential, and it is concluded that the midpoint potential can be adjusted through the redundant small vectors.
Reference [15] pointed out that when the output voltage vector is a small vector or a medium vector, the midpoint is connected to the load and a midpoint current is generated. Then the relationship between midpoint voltage and midpoint current is shown in Figure 3.

From Figure 3, the relationship between the midpoint current and the midpoint potential can be expressed as:

\[
\begin{align*}
  i_{c1} &= C \frac{du_{c1}}{dt} \\
  i_{c2} &= C \frac{du_{c2}}{dt} \\
  i_{NP} &= C \frac{d(u_{c1} - u_{c2})}{dt}
\end{align*}
\]  

(1)

From (1), the midpoint current can be adjusted by the midpoint current. Taking Figure 2 as an example, assume that the target voltage vector falls within a vector triangle enclosed by voltage vectors (PPO, OON), PON, and PNN. Suppose the action time of the three voltage vectors are, and. The duty cycle of the final output of the modulation link is shown in Figure 4.

It can be seen from Fig. 4 that when the voltage vector is ONN, the midpoint current is \( i_a \); when the voltage vector is PON, the midpoint current is \( i_b \); when the voltage vector is POO, the midpoint current is \( -i_a \). Taking the situation in Figure 4 as an example, suppose that the voltage difference between the two capacitors on the DC side is \( \Delta u_{dc} \), the small vector action time is \( T_1 \), the average midpoint current is \( i_1 \), the voltage vector action time is \( T_2 \), the midpoint current average is \( i_2 \), and the switching period is \( T_s \). The adjustment coefficient of the redundant small vector is \( k \), then the adjustment coefficient can be obtained from (2),

\[
\Delta u_{dc} = \frac{1}{C} \int (2k - 1)i_1T_1 + i_2T_2 dt
\]

(2)

3. Mathematical model of three-phase inverter model predictive control

3.1 Mathematical model of three-phase inverter

The topology of the three-phase inverter is shown in Figure 1. The output terminal is connected to an LC filter. The voltage of the output phase is \( u_u \), the current flowing through the inductor \( L \) is \( i_x \), and the voltage on the filter capacitor is \( u_c \). The parameters are assumed exactly the same, the capacitance is \( C \), the inductance is the value \( L \), and the inductance is the resistance \( R \). It can be obtained by Clarke transformation and Park transformation as (3) and (4) shown:

\[
\begin{align*}
  L_f \frac{du_{\alpha}}{dt} &= u_{\alpha} - u_{\alpha} - R_i i_{\alpha} \\
  L_f \frac{du_{\beta}}{dt} &= u_{\beta} - u_{\beta} - R_i i_{\beta}
\end{align*}
\]

(3)


\[
\begin{align*}
L_f \frac{di_d}{dt} &= u_{rd} - u_{qd} - R_i j_d + \omega L_i j_q \\
L_f \frac{di_q}{dt} &= u_{rq} - u_{qg} - R_i j_q - \omega L_i j_d
\end{align*}
\]  

(4)

Variables on the same coordinate axis are marked with the same subscript. It can be seen from (4) that the coupling amount is introduced into the expressions of the d-axis and the q-axis. In order to eliminate the influence between them, given the need to add feed-forward to eliminate the coupling effect, reference [16] gives the control equations of the voltage outer loop and current inner loop:

\[
\begin{align*}
\begin{cases}
i_d = (PI)_{\text{loop}} (u_{rd}^{\text{ref}} - u_{rd}) - \omega L_i j_d + i_{d}^{\text{ref}} \\
i_q = (PI)_{\text{loop}} (u_{rq}^{\text{ref}} - u_{rq}) + \omega L_i j_q + i_{q}^{\text{ref}} \\
u_{rd} = (PI)_{\text{loop}} (i_d^{\text{ref}} - i_d) - \omega L_i j_d + u_{rd}^{\text{ref}} \\
u_{rq} = (PI)_{\text{loop}} (i_q^{\text{ref}} - i_q) + \omega L_i j_q + u_{rq}^{\text{ref}}
\end{cases}
\end{align*}
\]  

(5)

Among them, $u_{rd}^{\text{ref}}$ and $u_{rq}^{\text{ref}}$ are the given voltage reference value of the voltage outer loop respectively, $(PI)_{\text{loop}}$ is the voltage outer loop controller; $i_d^{\text{ref}}$ and $i_q^{\text{ref}}$ are the given voltage reference value of the current inner loop, respectively, $(PI)_{\text{loop}}$ is the current inner loop controller.

It can be seen from the mathematical model that the control of the traditional three-phase inverter system is to enter the modulation link through the reference voltage signal obtained by the current inner loop to control the output signal. This approach can utilize PI control while decoupling the axis variables. Because PI can only track the deviation of the DC quantity, Park transformation is required, which increases the amount of calculation.

### 3.2 Model predictive control

Because the traditional control method uses a PI controller, and in order to achieve the error-free tracking of a given amount, the input of the controller is DC, so Park transformation is required. Model predictive control is different. It is a control method that predicts its future output based on the mathematical model of the controlled object. For predictive control of the finite set model, the optimal vector is selected by evaluating the future state of the system corresponding to each discrete voltage vector [17]. Considering the discrete characteristics of the numerical controller used, delays are introduced in the sample-hold and calculation processes. When the sampling frequency is much greater than the measured frequency, the differential component in the system can be approximated by the first-order forward differential component, that is

\[
\frac{dx}{dt} = \frac{x(k+1) - x(k)}{T_s}
\]  

(7)

Where and represent the sampling time, respectively, the controlled quantity, and the sampling period. Through (7), the original three-phase inverter mathematical model (3) can be discretized,

\[
\begin{align*}
\begin{cases}
i_{\alpha}(k+1) = \frac{T_s}{L_f} (u_{\alpha\alpha}(k) - u_{\alpha\alpha}(k)) + \left(1 - \frac{T_s R_e}{L_f}\right) i_{\alpha}(k) \\
i_{\alpha}(k+1) = \frac{T_s}{L_f} (u_{\alpha\beta}(k) - u_{\alpha\beta}(k)) + \left(1 - \frac{T_s R_e}{L_f}\right) i_{\beta}(k)
\end{cases}
\end{align*}
\]  

(8)

The predicted value of current can be obtained from (8). Where $u_{\alpha\alpha}(k)$ and $u_{\alpha\beta}(k)$ in the finite set prediction model, respectively, all the separate voltage vectors are projected on the two-phase stationary coordinate system, namely

\[
\vec{V}_s = V_{\alpha\alpha} + j V_{\alpha\beta}
\]  

(9)
Where \( \bar{v}_s \) is the selected voltage vector. Finally, the effect of each voltage vector is evaluated by the evaluation function, and the required voltage vector is finally obtained

\[
J_1(n) = (I_{ref\alpha}^n - I_{ref\alpha}(n))^2 + (I_{ref\beta}^n - I_{ref\beta}(n))^2
\]

(10)

Where \( I_{ref\alpha}^n \) and \( I_{ref\beta}^n \) are the projections of the reference current value on the \( \alpha \) axis and the \( \beta \) axis, respectively, and \( I_{ref\alpha}(n) \) and \( I_{ref\beta}(n) \) are the predicted current at time \( n \). For a three-phase three-level system, on the basis of ensuring the output voltage waveform, the midpoint current needs to be constrained. Assuming that the midpoint voltage offset is \( \Delta u_{dc} \), the second evaluation function is

\[
J_2(n) = \lambda \left( C \frac{\Delta u_{dc}}{I_0(n)} - I_0(n) \right)^2
\]

(11)

Where \( I_0 \) is the midpoint current generated by the voltage vector \( \bar{v}_s \) in a cycle, and \( \lambda \) is the weighting factor for midpoint current adjustment. Combining (8), (10) and (11), firstly through (8), (10) can realize the control of three-phase three-level inverter.

It can be seen from Figure 2 that the three-phase three-level inverter has 27 voltage vectors. In the traditional finite set control model predictive control, all voltage vectors need to be online for each prediction process (8) and (11) calculation, large amount of calculation. At the same time, each voltage vector needs to continue to act for one switching cycle, and only the voltage vector of the next cycle is considered from the aspect of voltage synthesis. This voltage synthesis method can ensure that the output follows the target vector.

In practice, not all voltage vectors need to participate in online calculations. Assuming that the target voltage vector is in sector III, both the discrete voltage vector and the target voltage vector are projected onto the coordinate axis of the two-phase stationary coordinate system. It can be seen that the components of some voltage vectors on the two coordinate axes are opposite to the desired synthesized voltage vector. It can be simplified here, as long as the voltage vector in the same large sector as the target voltage vector is involved in the synthesis. After adding this constraint, the voltage vectors involved in the selection were reduced from 27 to 10, reducing the amount of calculation. At the same time, considering the demand for the switching frequency, each switching cycle needs to have an effective vector and a zero vector. In a cycle, it can only change in the adjacent voltage state, that is, P-O, O-N, and cannot directly change between P and N states.

4. Simulation and experimental results

In order to verify the correctness of the proposed algorithm, this paper uses simulation and experiment to verify. The simulation software adopts MATLAB 2013b, and the experiment is completed on the three-phase inverter prototype built in the laboratory. Among them, the experimental platform adopts the switching device as the FS3L30R07W2H3F_B11 three-phase three-level module produced by Infineon, the controller uses the 32-bit floating point DSP of TI company: TMS320C6748, and the coprocessor is the FPGA of Xilinx company: Spartan6E, switch The tube drive is MAST5-6C-U12.

Table 1. System parameters.

| Items        | DC side | AC side filter |
|--------------|---------|----------------|
| Voltage      | 30V     | 470 \( \mu F \) |
| Capacitance  |         | 4 \( mH \)     |
| Inductance   |         | 3.3 \( \mu F \) |
| Capacitor    |         | 10 \( \Omega \) |
| Load         |         | 5 \( kHz \)     |
| On-off level |         | 0.8             |

4.1 Simulation results

Set the parameters in the simulation software according to Table 1. Comparing the simulation results of the two methods, the output current waveform is shown in Figure 5 at steady state.
Figure 5. The simulation of current (a) Traditional FCS-MPC (b) Improved algorithm

Figure 5 (a) is the current waveform when using traditional FCS-MPC control, and Figure 5 (b) is the current waveform when using the improved algorithm. Figure 6 shows the results of two methods for controlling the midpoint potential in the simulation platform.

Figure 6. The simulation of current (a) Traditional FCS-MPC (b) Improved algorithm

4.2 Experimental results

To further verify the effectiveness of the improved algorithm, a comparative experiment was conducted on the experimental platform. Except for the selection of 1.4mH for the filter inductance during the test, the remaining parameters are communicated with the values given in Table 1. Figure 6 shows the difference waveform of the midpoint voltage between the two methods.

Figure 7. The experimental differences of neutral point potential (a) Midpoint voltage of traditional method (b) Midpoint voltage of improved algorithm

In Figure 7, both the original method and the improved algorithm can maintain a constant midpoint voltage. Take the flip of a certain pin level of DSP as the mark, and the high level is the time required for calculation. The time required for the two algorithms to complete the control algorithm is shown in Figure 8. It can be seen from the comparison of Figure 8 that the traditional FCS-MPC algorithm needs 140 to complete the control, and the improved algorithm needs 54 to complete the control. This shows that the use of improved algorithms reduces the amount of calculation.
5. Conclusion

This paper proposes a fixed frequency model predictive control algorithm for a three-phase three-level inverter system. Based on the original algorithm model predictive control, the evaluation function is improved, the midpoint potential control of the three-level system is realized, the control process is simplified, the calculation time required for the control is shortened, and it is possible to further improve the system performance.

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