Model-Free Predictive Current Control for Voltage-Source PWM Rectifiers (Postprint)

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Abstract

Model predictive control has been extensively studied for voltage-source PWM rectifiers in recent years, featuring advantages such as simple principle, fast dynamic response, and ease of implementation; however, its shortcomings include heavy computational burden and the requirement for accurate system models and parameters. Through analysis of the PWM rectifier model combined with predictive control algorithms and fast vector selection, a direct current control method is obtained. This method can identify, via a single prediction calculation and comparison, the vector that yields optimal control performance at the current instant. Simplification leads to model-free predictive current control, which is characterized by requiring only one sampling of the grid current per control period and independence from rectifier model and parameters. This method demonstrates strong robustness and favorable dynamic and steady-state performance. This paper presents a comparative analysis of direct current control and model-free predictive current control in terms of steady-state performance, dynamic performance, and robustness, with experimental results verifying the correctness and effectiveness of both methods.

Full Text

Model-Free Predictive Current Control of the Voltage Source PWM Rectifier

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Abstract

Model predictive control has been extensively studied for voltage source PWM rectifiers in recent years, offering advantages such as simple principle, fast dynamic response, and easy implementation. However, its drawbacks include high computational burden and the requirement for accurate system models and parameters. By analyzing the PWM rectifier model and combining predictive control algorithms with fast vector selection, a direct current control method is obtained. This method can identify the optimal vector for the current moment through a single prediction calculation and comparison, which simplifies into model-free predictive current control. The key feature is that only one grid current sample is required per control cycle, independent of the rectifier model and parameters. This method demonstrates strong robustness with good dynamic and steady-state performance. This paper compares direct current control and model-free predictive current control in terms of steady-state performance, dynamic response, and robustness, with experimental results verifying the correctness and effectiveness of both methods.

Keywords: PWM rectifier, direct current control, model-free predictive current control, robustness

1 Introduction

Three-phase voltage source PWM rectifiers exhibit excellent performance characteristics including bidirectional energy flow, constant DC output voltage, low grid-side current harmonics, and controllable power factor, making them widely applicable in motor drives, renewable energy grid integration, power factor correction, and other fields [1]. Control methods for PWM rectifiers primarily include Voltage Oriented Control (VOC) [2], Model Predictive Control (MPC) [3], and Direct Power Control (DPC) [4]. Among these, model predictive control has recently become a research hotspot for PWM rectifiers due to its simple principle, fast dynamic response, and ease of implementation.

MPC can be categorized based on controlled variables into Predictive Power Control (PPC) [5-6] and Predictive Current Control (PCC) [7-8]. Since PWM rectifier grid currents are measurable, current predictive control is more direct than power predictive control, making PCC an important research direction. Traditional PCC calculates the next moment' s current based on the PWM rectifier discrete model and measured grid current values, then enumerates voltage vectors to minimize the cost function, thereby selecting the optimal voltage vector. This approach suffers from excessive computational load, and implementing one-step delay compensation [9] further increases the algorithmic burden. Additionally, this method' s limitation lies in its implementation only in the three-phase stationary coordinate system, where current difference data occupies substantial memory space.

For PWM rectifiers, this paper proposes an improved Model-Free Predictive Current Control (MFPCC) based on [10]. This method requires only one current sample per control cycle, reducing computational load. In fact, due to high sampling frequency, a single current sample can achieve current difference calculation, thereby simplifying the algorithm. Unlike [10], the proposed MFPCC operates in the two-phase stationary coordinate system, requiring only a small array to store current difference values, thus reducing computational effort and facilitating digital implementation. Experiments conducted on a two-level PWM rectifier experimental platform based on TMS320F28335 DSP compare the dynamic and steady-state performance and parameter robustness of DCC and MFPCC, with results demonstrating the effectiveness of both methods.

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2 Direct Current Control

2.1 Predictive Current Calculation

The three-phase voltage source PWM rectifier circuit is shown in Figure 1 [Figure 1: see original paper], where the three-phase AC voltages on the grid side are e_a, e_b, and e_c; L and R represent the equivalent inductance and resistance of the AC-side reactor.

Using Equation (1) to transform the three-phase stationary coordinate system to the two-phase stationary coordinate system:

$$x = \frac{2(x_a + e^{j2\pi/3}x_b + e^{-j2\pi/3}x_c)}{3}$$

where x is the complex vector in the two-phase coordinate system.

Transforming the rectifier model quantities to the two-phase coordinate system using Equation (1), the three-phase PWM rectifier model can be expressed as:

$$e = Ri + L\frac{di}{dt} + v$$

where e is the grid voltage complex vector, i is the grid-side current complex vector, and v is the rectifier AC-side output voltage complex vector.

The term di/dt in Equation (2) can be discretized using first-order Euler discretization:

$$\frac{di}{dt} \approx \frac{i_{k+1} - i_k}{T_s}$$

Substituting Equation (3) into Equation (2) yields the discretized form:

$$i_{k+1} = \left(1 - \frac{RT_s}{L}\right) i_k + \frac{T_s}{L}(e_k - v_k)$$

According to Equation (4), using the grid voltage e_k, grid current i_k, and rectifier AC-side voltage v_k sampled at time k, the predicted grid current value i_{k+1} at time $k+1$ can be calculated.

Using the sampled grid voltage e and grid current i, the grid complex power S can be calculated [11]:

$$S = P + jQ = 1.5(i^* \cdot e)$$

where "*" denotes the complex conjugate.

From Equation (5), the current reference value i_{ref} is:

$$i_{ref} = \frac{S_{ref}}{1.5e^*}$$

2.2 Optimal Vector Selection

A two-level voltage source PWM rectifier can output eight voltage space vectors, corresponding to eight switching states. These include six non-zero active vectors $(u_1, u_2, \dots, u_6)$ and two zero vectors (u_0, u_7). Traditional model predictive current methods require eight enumerations, resulting in heavy computational load. This paper proposes a fast vector selection method.

From Equation (4), the grid current prediction can be rewritten as:

$$i_{k+1} = i_{0,k+1} - \frac{T_s}{L} v_k$$

where $i_{0,k+1}$ is the predicted current under zero vector action.

Equation (7) shows that the predicted current i_{k+1} can be expressed using $i_{0,k+1}$ and v_k. The error vector between reference current and predicted current can be expressed as:

$$\varepsilon_{k+1} = i_{ref,k+1} - i_{k+1} = \varepsilon_{0,k+1} + \frac{T_s}{L} v_k$$

where the error vector generated by the zero vector is:

$$\varepsilon_{0,k+1} = i_{ref,k+1} - i_{0,k+1}$$

As shown in Figure 2 [Figure 2: see original paper], if the error vector $-\varepsilon_{0,k+1}$ lies in the shaded region, the vector u_1 yields the smallest error vector magnitude, making u_1 the optimal non-zero vector for the current moment. Substituting this vector into Equation (7) yields the current error vector under u_1 action: $\varepsilon_{k+1} = i_{ref,k+1} - i_{k+1} = i_{ref,k+1} - i_{0,k+1} + \frac{T_s}{L}u_1 = \varepsilon_{0,k+1} + \frac{T_s}{L}u_1$.

After calculating $|\varepsilon_{k+1}|$ and $|\varepsilon_{0,k+1}|$, if $|\varepsilon_{k+1}| > |\varepsilon_{0,k+1}|$, the optimal voltage vector is the non-zero vector u_1. If $|\varepsilon_{k+1}| < |\varepsilon_{0,k+1}|$, the optimal voltage vector is the zero vector. If the final optimal vector is a zero vector, the principle of minimizing switching actions at the current moment should be followed, considering the vector applied in the previous moment and selecting the zero vector that results in fewer switching actions.

2.3 One-Beat Delay Compensation

After sampling the current i_k at time k, digital signal processing requires a certain period, and the selected voltage vector u_k can only be applied at time $k+1$. This leads to additional power ripple. To achieve precise current control, a one-step ahead prediction method is needed to solve the control delay problem. The specific process involves predicting the grid current i_{k+1} at time $k+1$ using the predicted current value combined with fast vector selection to choose the optimal voltage vector to be applied at time $k+1$.

With one-beat delay compensation, Equation (10) is rewritten as:

$$\varepsilon_{k+2} = i_{ref,k+2} - i_{k+2}$$

where:

$$i_{k+2} = \left(1 - \frac{RT_s}{L}\right) i_{k+1} + \frac{T_s}{L}(e_{k+1} - v_{k+1})$$

3 Model-Free Predictive Current Control

Since only one voltage vector is selected for application in each control cycle, model-free predictive current control can theoretically be implemented. It predicts the next moment' s current based on the currently sampled current and current differences. Due to the short system control cycle, grid current is considered to change linearly within each control cycle. Based on this consideration, current differences in each cycle can be accurately calculated. In practical systems, voltage jumps and dead time affect sampling accuracy. To obtain more accurate current differences, the model-free predictive current control proposed in [10] samples current twice within one control cycle and then predicts the current difference value. This approach increases hardware sampling frequency and software implementation difficulty, while also increasing computational burden and potentially detecting current spikes. To avoid this situation, the common practice is to delay the second current sampling for a period after voltage vector application. However, the delay time requires empirical tuning, which hinders the method' s generalization.

This paper employs a single current sampling method, sampling current only once per control cycle. Since the sample at the beginning of each control cycle is also the sample at the end of the previous cycle, the same effect can be achieved. Because the impact of one-beat delay must be considered, the grid current at time $k+2$ should be predicted at the current time k. As shown in Figure 3 [Figure 3: see original paper], the current at time $k+2$ can be obtained as:

$$i_{k+2} = i_k + \Delta i_k(u_k) + \Delta i_{k+1}(u_x)$$

where $\Delta i_k(u_k)$ is the current difference vector under u_k action between times k and $k+1$, and $\Delta i_{k+1}(u_x)$ is the current difference vector under u_x action between times $k+1$ and $k+2$. It can be seen that the predicted current at time $k+2$ requires no rectifier parameters.

From Equation (13), the values of Δi_k and Δi_{k+1} cannot be calculated at time k because they are computed at times $k+1$ and $k+2$, respectively. The fundamental idea of predictive control is to predict future information based on present and past system information. If the sampling interval is sufficiently short, present and past current differences can be directly used to predict the next moment' s current. In other words, Δi_k and Δi_{k+1} can be approximated using previously stored current difference vectors:

$$\Delta i_k \approx \Delta i_{old}|_{S_i=S_k}, \quad i = 0, \dots, 7$$

$$\Delta i_{k+1} \approx \Delta i_{old}|_{S_j=S_{k+1}}, \quad j = 0, \dots, 7$$

where $\Delta i_{old}|_{S_i=S_k}$ and $\Delta i_{old}|_{S_j=S_{k+1}}$ are the current differences under voltage vectors S_i and S_j applied in past moments. Substituting Equations (16) and (17) into Equation (13) yields the current at time $k+2$:

$$i_{k+2} = i_k + \Delta i_{old}|_{S_i=S_k} + \Delta i_{old}|_{S_j=S_{k+1}}$$

Equation (18) shows that the predicted current at time $k+2$ depends only on the current moment' s current and current differences. To more accurately approximate Δi_k and Δi_{k+1}, previously stored current differences should be updated with new calculated values in each sampling cycle. Different voltage vectors have different update frequencies, which may lead to inaccurate current predictions under certain voltage vectors. However, due to the high sampling frequency, the impact of different update frequencies is minimal.

The cost function is defined as:

$$g = |i_{ref} - i_{k+2}|$$

All seven voltage vectors are substituted into Equations (18) and (19), and the voltage vector corresponding to the minimum value calculated by the cost function in Equation (19) is selected as the final vector.

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4 Experimental Results

4.1 Steady-State Performance

A two-level PWM rectifier prototype was built to experimentally verify the DCC and MFPCC control strategies. The experimental platform architecture is shown in Figure 4 [Figure 4: see original paper]. The control board uses a DSP TMS320F28335 controller, with four-channel DA added to facilitate observation of internal variables. During experiments, current probes directly measure current, while remaining electrical quantities are output through 12-bit DA. System parameters include: AC-side line voltage RMS 150V, frequency 50Hz, AC-side inductance 10mH, DC bus capacitance 840 F, and sampling frequency 30kHz.

To compare the steady-state performance of both methods, analysis was conducted with DC bus voltage in open-loop and only power closed-loop. Figures 5a [Figure 5: see original paper] and 5b show the steady-state waveforms for DCC and MFPCC with active power reference of 1000W and reactive power reference of 0var. The oscilloscope’ s four channels from top to bottom are active power reference, actual active and reactive power, and AC-side phase A current.

Figure 5 indicates that DCC has inferior steady-state performance compared to MFPCC, with larger active and reactive power ripples. Further quantitative analysis by calculating the RMS values of active and reactive power ripple over 0.1s shows that DCC has active power ripple of 33.4985W, reactive power ripple of 31.6836var, and phase A current THD of 2.2008%. MFPCC exhibits active power ripple of 20.9632W, reactive power ripple of 22.0266var, and phase A current THD of 1.9334%, demonstrating that MFPCC steady-state performance is superior to DCC.

4.2 Dynamic Performance

Figures 6a [Figure 6: see original paper] and 6b show the experimental waveforms when PWM rectifier active power steps from 600W to 1000W under DCC and MFPCC, respectively. Both methods exhibit similar dynamic performance, with actual active power quickly tracking the reference without overshoot, proving that DCC and MFPCC have similarly excellent dynamic performance.

4.3 Robustness

Figures 7a [Figure 7: see original paper] and 7b show the waveforms when model inductance changes from 10mH to 5mH during steady-state operation with active power reference of 1000W and reactive power reference of 0var. The analysis above indicates that both methods have similar steady-state performance under ideal parameters. However, when inductance varies, DCC' s active and reactive power ripples both increase, while MFPCC' s power ripples remain constant, demonstrating strong parameter robustness because MFPCC itself does not depend on system parameters.

Figure 8 [Figure 8: see original paper] compares the active power ripple, reactive power ripple, and THD for DCC and MFPCC when model inductance equals 10mH and 5mH, respectively. Figures 7 and 8 demonstrate that MFPCC has stronger parameter robustness compared to DCC.

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5 Conclusion

This paper first proposes a direct current control method that obtains the final selected vector through predictive calculation of current vectors and comparison of their error vectors. Since the entire process requires only one prediction and one comparison, computational load and complexity are reduced compared to traditional direct current control. Subsequently, a model-free predictive current control method is proposed, which samples grid current once per cycle and obtains the optimal voltage vector without requiring rectifier parameters, offering strong robustness and easier implementation than the proposed direct current control method. Experimental results from a two-level PWM rectifier prototype show that under ideal parameters, DCC has larger active power ripple, reactive power ripple, and current THD than MFPCC in steady-state, indicating MFPCC' s superior steady-state performance. During dynamic conditions, both methods exhibit very similar performance. Considering MFPCC' s strong parameter robustness, it has greater practical value in real systems.

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Note: Figure translations are in progress. See original paper for figures.