Model Predictive Direct Power Control for Single Phase Three-Level Rectifier at Low Switching Frequency

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Abstract—This paper presents a new model predictive direct power control (MP-DPC) to overcome the drawbacks of model predictive control (MPC) for single phase three-level rectifiers in the railway traction drive system, including huge online calculation, poor power control precision at the low switching frequency and variable switching frequency. To do so, an exact analytical solution of instantaneous power estimation is adopted to predict active and reactive powers in next duty cycle updating interval for achieving the deadbeat control and reducing the predictive error at the low switching frequency (below 1kHz). The optimal $d$-axis and $q$-axis components of input voltage within next duty cycle updating interval of the adopted rectifier in rotating coordinate system are directly calculated by minimizing the cost function. And the optimal drive pulses are generated by pulse width modulation stage in the proposed MP-DPC, other than evaluating cost function for each voltage vector in traditional MP-DPC. Finally, the influence of inductance mismatch on control system is analyzed, and an inductance estimation method is shown to improve the control precision. An experimental comparison with other five different DPC schemes has verified the effectiveness of the proposed MP-DPC scheme.

Index Terms—Direct power control, low switching frequency, model predictive control, single phase three-level rectifiers, inductance estimation.

I. INTRODUCTION

Compared with single phase two-level pulse width modulation (PWM) converters, single phase three-level converters present more advantages of low current harmonics, high voltage level, and high power density. Therefore, single phase three-level converters are applied widely as railway locomotive traction rectifiers, grid-tied inverter, uninterruptible power supply etc. Various current control methods have been proposed for the adopted rectifiers, which include hysteresis current control [1], $dq$-axis current control [2], proportional-resonant (PR) current control [3] etc. However, due to the continual power fluctuation and the low switching frequency in railway locomotive application, these traditional current control schemes cannot achieve the desired requirement.

Being different from traditional current control, another advanced approach is direct power control (DPC) [4], [5], which directly controls active and reactive powers to realize constant DC voltage and unit power factor, respectively. Traditional DPC [6], [7] originated from direct torque control (DTC) in motor drives, can achieve power decoupling control by power switching table lookup. Converters can achieve high dynamic and robustness in table-based DPC, but variable switching frequency limit its application. The developed DPC schemes [8]-[11] combining with pulse width modulation (PWM) are proposed to realize constant switching frequency. In [8], a simple linear power prediction method in the $dq$ rotating coordinate system is proposed to achieve the power predictive control with constant switching frequency. But it is difficult for this method to solve the optimal control output within next duty cycle updating interval. In [10], the power prediction method based on two-steps prediction is proposed to compensate the time delay caused by digital control. In [11], a proportional-integral (PI) control scheme with power feedforward is proposed to achieve power decoupling control. However, the control response depends on the parameters of PI controllers. And it is not easy to tune these parameters. Therefore, many advanced DPC schemes, such as advanced table-based DPC [12], [13], DPC based on fuzzy sliding mode [14], model predictive DPCs (MP-DPC) [15]-[26], are proposed to further enhance control performance. And MP-DPC schemes, due to the advanced dynamic and steady precision, have been widely studied in recent years.

In MP-DPC, a pre-defined cost function is used to evaluate and choose the appropriate switching state in next control interval. And by this way, traditional MP-DPC [19], [20] chooses an optimal vector from these possible vectors to minimize power error and decouple active and reactive powers, other than switching table-lookup in the traditional DPC. Furthermore, the optimal duty cycle method [21] by combining effective vectors with zero vector are proposed to minimize cost function of the traditional MP-DPC. And the current harmonics are observably reduced by this method. In addition, the cost function is further advanced by choosing a pair of effective vectors other than zero vector [22]. Although the power fluctuation of these MP-DPCs is reduced, the switching frequency still is variable with the operation condition. Therefore, the optimal MP-DPC based on multi-vectors is proposed to realize constant switching frequency [23]. In above all MP-DPCs, the cost function value needs to be calculated and evaluated repeatedly to select the appropriate combination of these vectors. Hence, the huge
computing resource is consumed and the control frequency is limited, even if some schemes are proposed to reduce the calculation complexity [24]-[26]. In [26], a virtual-flux-based predictive DPC of AC/DC converters is proposed for three phase rectifiers, the switching table combined with multi-vectors optimization can reach constant switching frequency and reduce the calculation complexity. In addition, the MP-DPC combining with PWM is proposed to reduce the calculation complexity in [27]-[29], which calculate the optimally modulated wave and generate drive pulses by PWM. However, since the modulated wave is the ac value, the predicted modulation signal is optimal in the sampling instant, but is not always the best within the entire switching interval, especially at the low switching frequency. And the control performance will deteriorate with the decrease of the switching frequency. Moreover, all of the model predictive control (MPC) schemes are sensitive to circuit parameters. Thus, parameter estimation [20], [26], [29]-[31] schemes are reported to improve the control accuracy. In [20], the least square method is applied to accurately estimate the inductance value, but the response of this estimator will become slow with an increase of the program runtime. Thus, [26] proposed a low pass filter (LPF) to replace least square method. But the response of this estimator is limited by LPF. In order to improve the response of this estimator, some estimators based on the controller characteristics, such as Luenberger observer and Newton- Raphson algorithm, are proposed in [29]-[31].

But the complexity of these schemes restricts control frequency.

In low switching frequency condition, a high precision predictive model to predict the future value, during one duty cycle updating interval, is only significant for the MP-DPC to obtain the optimal control output within entire switching period. Some predictive control methods have been proposed [32]-[37]. In [33], an MPC method with switching frequency reduction is achieved by forming a new cost function with two-step prediction, which can reduce the average switching frequency. In [34], MPC with the fixed switching frequency is realized by the predetermined sequences that are based on the behavior of the regular sampling PWM. In order to obtain the higher control precision, a very high sampling frequency (400kHz) is applied in this control method. Meanwhile, in order to reduce calculation complexity, the coefficient table needs to be predefined, and the size of this lookup table will increase with the predictive horizon and the switching states. In [35], [36], the weighting of the switching frequency is introduced into the cost function for reducing the switching frequency, but the switching frequency varies with operation condition, which is similar to [33]. Almost all above control schemes are applied in the three-phase system, and MP-DPC applied in single phase converters should be discussed and studied in detail according to its own characteristics.

Currently, in low switching frequency condition (below 1kHz) condition, above MPC schemes have some drawbacks which are summarized as follows: 1) The power prediction schemes reported in existing literatures are not accurate at low switching frequency; 2) The sector selection and vectors evaluation for minimizing cost function lead to the high calculation complexity, 3) the sensibility of MPC to circuit parameter will enlarge power errors and deteriorate control performance; 4) The control precision is positively correlated with control frequency, which is usually set as a high value to obtain a better control performance, leading to the relatively high switching frequency. Especially, in the single phase system, the instantaneous power estimation is more complicated than that in the three-phase system, which limits the existing MP-DPC schemes in the single phase system.

To solve the above problems existing in MP-DPC scheme at low switching frequency, in this paper, combining PWM scheme reported in [38], a new MP-DPC scheme for single-phase three-level rectifiers in railway traction drive system is proposed to simplify calculation, increase power control precision, and improve power decoupled ability at low switching frequency. The rest of this paper is organized as follows: Section II describes the general power mathematical model and instantaneous power estimation method based on fictive-axis emulation (FAE) [39]. The proposed MP-DPC with the boundary optimization and its compensation scheme are discussed in Section III. The influence of inductance mismatch on the control system and inductance on-line estimation are presented in Section IV. In Section V, a comparison of five traditional control methods and the proposed control method is carried out in the scaled-down experimental platform. Section VI concludes this paper.

II. SYSTEM DESCRIPTION AND MATHEMATICAL MODEL

A. System description and d-q mathematical model

Fig. 1 shows the topology of a single phase three-level neutral-point-clamped (NPC) converter, where \( u \) and \( i \) are the main voltage and the line current, respectively. \( L \) and \( R \) are the equivalent inductance and resistance of the traction transformer, respectively; \( C_1 \) and \( C_2 \) are symbols for the two capacitors in the DC-link; \( u_{ab} \) is the input voltage of the H-bridge converter; \( u_{d1} \) and \( u_{d2} \) are the voltage of capacitors \( C_1 \) and \( C_2 \), respectively. \( R_L \) is the symbol for the inverter-motor side equivalent resistance; \( S_{a1} - S_{a4} \) are IGBT modules with anti-parallel diodes of a phase. \( S_{b1} - S_{b4} \) are IGBT modules with anti-parallel diodes of b phase.

![Fig. 1. Single phase three-level NPC converter](image)

The main voltage \( u \), the line current \( i \) and the input voltage \( u_{ab} \) of the adopted converter are defined as

\[
u = u_x \sin(\alpha t) + u_y \cos(\alpha t)\]  

\[i = i_x \sin(\alpha t) + i_y \cos(\alpha t)\]  

\[u_{ab} = u_{d1} \sin(\alpha t) + u_{d2} \cos(\alpha t)\]

Where \( u_{d1} \) and \( u_{d2} \) are the \( d \)-axis components in \( dq \) rotating frame of the voltage and current vectors \( u \), \( i \), \( u_{ab} \), respectively;
\[ u_{\alpha}, i_{\alpha}, \text{and } u_{abq} \text{ are the } q\text{-axis components in } d-q \text{ rotating frame of the voltage and current vectors } u, i, u_{ab}, \text{respectively.} \]

As shown in Fig. 1, Kirchhoff voltage law (KVL) is adopted to analyze voltage across inductor \( L \), and the voltage equation is shown as

\[ L \frac{di}{dt} = u - Ri - u_{ab} \]  \hspace{1cm} (4)

Substituting (1)–(3) into (4), \( d-q \) mathematical model of the adopted converter is deduced as

\[
\begin{align*}
L \frac{di}{dt} &= u_{d} - R_{d}i_{d} + \omega L_{d}i_{q} - u_{abd} \\
L \frac{di}{dt} &= u_{q} - R_{q}i_{q} - \omega L_{q}i_{d} - u_{abq}
\end{align*}
\]  \hspace{1cm} (5)

B. Power mathematical model

On the basis of the instantaneous power theory, the instantaneous active power \( P \) and reactive power \( Q \) are defined as

\[ \left[ \begin{array}{c} P \\ Q \end{array} \right] = \frac{1}{2} \left[ \begin{array}{cc} u_{d} & u_{q} \\ -u_{q} & u_{d} \end{array} \right] \left[ \begin{array}{c} i_{d} \\ i_{q} \end{array} \right] \]  \hspace{1cm} (6)

Where \( u_{d}, u_{q}, i_{d}, i_{q} \) are \( \alpha \text{-axis and } \beta\text{-axis components of the main voltage and the line current in the two-phase stationary coordinate system, respectively.} \)

Assuming there is no fluctuation in the main voltage, which is supplied by railway power supply system. And voltage components \( u_{d} \) and \( u_{q} \) in \( d-q \) rotating frame are constant values in normal condition. According to (6), the differential values of active and reactive powers are deduced as

\[ \frac{dP}{dt} = \frac{1}{2} \left[ u_{d} \frac{di_{d}}{dt} + u_{q} \frac{di_{q}}{dt} \right] \]  \hspace{1cm} (7)

Substituting (7) into (5), the components \( u_{abd} \) and \( u_{abq} \) of the input voltage \( u_{ab} \) in \( d-q \) rotating frame can be expressed as

\[
\begin{bmatrix}
 u_{abd} \\
 u_{abq}
\end{bmatrix} = \frac{1}{2} \begin{bmatrix}
 u_{d} - 2L \frac{di_{d}}{dt} + u_{q} \frac{di_{q}}{dt} \\
 u_{q} + u_{d} - 2L \frac{di_{d}}{dt} - u_{q} \frac{di_{q}}{dt}
\end{bmatrix}
\]  \hspace{1cm} (8)

It is assumed the voltage vector is oriented to \( d\text{-axis in rotating coordinate system, and } d\text{-axis and } q\text{-axis components of } u_{ab} \text{ can be further simplified as}

\[
\begin{bmatrix}
 u_{abd} \\
 u_{abq}
\end{bmatrix} = \begin{bmatrix}
 u_{m} \\
 0
\end{bmatrix} - \frac{2L}{u_{m}} \begin{bmatrix}
 \frac{dP}{dt} \\
 \frac{dQ}{dt}
\end{bmatrix} + \frac{2}{u_{m}} \begin{bmatrix}
 -\omega L Q \\
 \omega R P + \omega L Q
\end{bmatrix}
\]  \hspace{1cm} (9)

Where \( u_{m} \) is the amplitude of the main voltage \( u \).

Traditionally, proportional-integral (PI) controllers are applied to replace the differential elements of active and reactive powers in (9), constituting power feed-forward decoupling controller [11].

C. Instantaneous power estimation

According to (6), in order to estimate active and reactive powers, \( \beta\text{-axis components of the main voltage vector } u \) and the line current vector \( i \) in the stationary coordinate system need to be estimated at first. \( \beta\text{-axis component of the main voltage vector can be estimated by the second-order generalized integral (SOGI) [40]. } \)

\[
L \frac{di_{\beta}}{dt} = u_{\beta} - R_{\beta}i_{\beta} - u_{ab\beta}
\]  \hspace{1cm} (10)

Where \( u_{ab\beta} \) is the \( \beta\text{-axis component of the input voltage of the adopted converter.} \)

Therefore, \( i_{\beta} \) can be estimated by the fictive-axis emulation (FAE) [39]. And the block diagram of FAE is shown in Fig. 2.

\[ \text{Fig. 2 The estimation of } i_{\beta} \text{based on FAE} \]

Where \( u_{ab\beta} \) is gotten from rotating coordinate transformation, which transforms \( u_{ab} \) to \( u_{ab\beta} \) and \( u_{ab\beta}, i_{\beta} \) are estimated by SOGI. And then active and reactive powers can be estimated by (6).

III. MODEL PREDICTIVE DIRECT POWER CONTROL

A. Analytical solution of instantaneous power

The amplitude of the main voltage \( u_{m}, u_{abd} \) and \( u_{abq} \) in (9) can be considered as constant in one switching interval. Therefore, (9) can be transformed into linear non-homogeneous ordinary differential equations with constant coefficients, shown as

\[
\begin{align*}
\frac{dP}{dt} + \frac{R}{L} P + \frac{\omega Q}{L} - u_{m} u_{abd} &= 0 \\
\frac{dQ}{dt} + \frac{R}{L} Q - \frac{\omega P}{L} + u_{m} u_{abq} &= 0
\end{align*}
\]  \hspace{1cm} (11)

Therefore, according to analytical method of the differential equations [41], the exact analytical solution of active and reactive powers is solved as

\[
\begin{align*}
P &= e^{-\frac{R}{L} t} \left( A_{1} \cos \omega_{c} t + A_{2} \sin \omega_{c} t \right) - \frac{u_{m} u_{abq} - \frac{\omega}{L} \left( u_{m}^{2} - u_{m} u_{abd} \right)}{2\omega L + \frac{\omega}{L} R} \\
Q &= e^{-\frac{R}{L} t} \left( A_{1} \sin \omega_{c} t - A_{2} \cos \omega_{c} t \right) + \frac{\frac{\omega}{L} u_{m} u_{abq} + \left( u_{m}^{2} - u_{m} u_{abd} \right)}{2\omega L + \frac{\omega}{L} R}
\end{align*}
\]  \hspace{1cm} (12)

Where \( A_{1} \) and \( A_{2} \) are determined by the initial power value at the beginning of each switching interval. \( \omega_{c} \) is the system characteristic frequency, which is defined as

\[ \omega_{c} = R/L \]  \hspace{1cm} (13)

It is clear from (12) that damped oscillation will occur while \( u_{abd} \) and \( u_{abq} \) are constants unless both \( A_{1} \) and \( A_{2} \) are zero.
Hence, $A_1$ and $A_2$ will be zero in steady state. And active and reactive powers $P$, and $Q$, in steady state can be expressed as

$$
P_s = \frac{u_n u_{abq} - \omega L (u_n^2 - u_n u_{abq})}{2\omega L + 2\omega L R} = \frac{\omega L u_n u_{abq} + (u_n^2 - u_n u_{abq})}{2\omega L + 2\omega L R}
$$

\[(14)\]

The relative small parasitic resistance $R$ of the traction transformer can be ignored in practice, the active power $P$ and reactive power $Q$ in (12) can be simplified as

$$
P = A_1 \cos \omega t + A_2 \sin \omega t - \frac{u_n u_{abq}}{2\omega L}
$$

$$
Q = A_1 \sin \omega t - A_2 \cos \omega t + \frac{u_n^2}{2\omega L} - \frac{u_n u_{abq}}{2\omega L}
$$

\[(15)\]

B. Instantaneous power prediction

The regular sampling method is widely applied to generate pulse signals with the desired widths in actual digital controllers. The duty cycle will be updated once or twice in each switching interval according to the different configuration of the digital controller. And the adopted updating process of the duty cycle in proposed control system is shown in Fig. 3.

![Fig. 3 The control system timing sequence](image)

Where duty cycle is updated at the top and bottom of the triangular carrier in TMS320F28335. $u_{abq}$ is the $\alpha$-axis component of $u_{ab}$ in the stationary reference frame, and it is held within each half switching interval. $u_{abq*}$ represents the discrete value of $u_{abq}$ in the actual digital controller. The control and sampling frequency $f_s$ are set to the integral multiple of switching frequency $f_s$ to ensure the synchronization of the switching and controlling sequence. The integer $a$ in Fig. 3 is defined as

$$
a = \frac{T_s}{T_x}
$$

\[(16)\]

Where $T_s$ and $T_x$ represent the switching interval and the control interval; $u_{abq*}$ is the $\alpha$-axis component of $u_{ab}$ in the the stationary reference frame, $u_{abq*}$ represents the the discrete value of $u_{abq}$ in the the actual digital controller, which is updated in every half switching interval, $n$ represents the $n$th hal switching interval.

It is assumed that both $u_{ab}$ and $u_{abq}$ are constant within each half switching period, and according to (15), the active power and reactive power within the $n$th half switching interval can be expressed as

$$
P(n\frac{T}{2} + t) = A_1(n\frac{T}{2}) \cos \omega t + A_2(n\frac{T}{2}) \sin \omega t
$$

$$
-\frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2}) (0 \leq t \leq \frac{T}{2})
$$

\[(17)\]

$$
Q(n\frac{T}{2} + t) = A_1(n\frac{T}{2}) \sin \omega t - A_2(n\frac{T}{2}) \cos \omega t
$$

$$
+ \frac{u_n^2}{2\omega L} - \frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2})
$$

Where $P(n\frac{T}{2})$ and $Q(n\frac{T}{2})$ are instantaneous powers at the beginning of the $n$th half switching interval. $u_{abq}(n\frac{T}{2})$ and $u_{abq}(n\frac{T}{2})$ are $d$-axis and $q$-axis components during the $n$th half switching interval, respectively, $A_1(n\frac{T}{2})$ and $A_2(n\frac{T}{2})$ in (17) are determined by initial powers at the beginning of the $n$th half switching interval.

In order to achieve deadbeat model predictive power control, the predictive value of active and reactive powers need to be estimated from (17). Firstly, while the control frequency is much higher than switching frequency, $P(nTs)$ and $Q(nTs)$ are approximate to the estimated powers $P(kTs)$ and $Q(kTs)$, and expressed as

$$
P(kTs) \approx P(n\frac{T}{2})
$$

$$
Q(kTs) \approx Q(n\frac{T}{2})
$$

\[(18)\]

Where $k$ represents $k$th control interval being closest to the $n$th half switching interval.

According to (17) at $t=0$ and (18), $A_1(nTs/2)$ and $A_2(nTs/2)$ should satisfy

$$
P(kTs) \approx P(n\frac{T}{2}) - \frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2})
$$

$$
Q(kTs) \approx Q(n\frac{T}{2}) + \frac{u_n^2}{2\omega L} - \frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2})
$$

\[(19)\]

Therefore, $A_1(nTs/2)$ and $A_2(nTs/2)$ can be expressed as

$$
A_1(n\frac{T}{2}) = \frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2}) + P(kTs)
$$

$$
A_2(n\frac{T}{2}) = \frac{u_n^2}{2\omega L} - \frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2}) - Q(kTs)
$$

\[(20)\]

According to (17), The predictive active and reactive powers during the next half switching interval are expressed as

$$
P(n\frac{T}{2} + T / 2) = A_1(n\frac{T}{2}) \cos \omega T \frac{T}{2} + A_2(n\frac{T}{2}) \sin \omega T \frac{T}{2}
$$

$$
-\frac{u_n}{2\omega L} u_{abq} (n\frac{T}{2})
$$

\[(21)\]
Therefore, the active power and reactive power during the next half switching interval can be predicted from (20) and (21), and the deadbeat control can be achieved by this way.

C. The optimal solution of \( u_{abd} \) and \( u_{abq} \)

The reference of active power \( P_{ref} \) is usually obtained by PI controller of DC-link voltage, and the reference of reactive power \( Q_{ref} \) is usually set as zero to realize unit power factor. Therefore, in steady state, \( P_{ref} \) and \( Q_{ref} \) can be regarded as constant in a half switching interval, and expressed as

\[
P_{ref}(n \frac{T}{2} + \frac{T}{2}) = P_{ref}(n \frac{T}{2}) = P_{ref}(kT_s) \]

\[
Q_{ref}(n \frac{T}{2} + \frac{T}{2}) = Q_{ref}(n \frac{T}{2}) = Q_{ref}(kT_s) \]

(22)

A simple cost function [42] can be used to weigh power error after half switching interval shown as

\[
J = \left[ P_{ref}(kT_s) - P(n \frac{T}{2} + \frac{T}{2}) \right]^2 + \left[ Q_{ref}(kT_s) - Q(n \frac{T}{2} + \frac{T}{2}) \right]^2
\]

(23)

The cost function \( J \) in (23) is relative to \( u_{abd} \) and \( u_{abq} \). And the relation curve of the normalized \( J \) with respect to \( u_{abd} \) and \( u_{abq} \) is drawn as Fig. 4. From this figure, there is only one minimum value point in the entire operation range.

![Fig. 4 The relation curve of the normalized cost function with respect to \( u_{abd} \) and \( u_{abq} \)](image)

Where normalized \( J, u_{abd} \) and \( u_{abq} \) are defined as

\[
J = \frac{J}{P_{ref}^2 + Q_{ref}^2} \]

(24)

\[
\begin{aligned}
J_{u_{abd}} &= \frac{u_{abd}}{u_{dc\_ref}} \\
J_{u_{abq}} &= \frac{u_{abq}}{u_{dc\_ref}}
\end{aligned}
\]

(25)

Where \( u_{dc\_ref} \) represents dc-link reference voltage. Therefore, in order to minimize cost function, \( u_{abd} \) and \( u_{abq} \) need to satisfy

\[
\frac{\partial J}{\partial u_{abd}} = 0, \quad \frac{\partial J}{\partial u_{abq}} = 0
\]

(26)

Solving the simultaneous equations (20)-(26), the optimal solution of \( u_{abd} \) and \( u_{abq} \) can be obtained as

\[
\begin{aligned}
u_{abd}(n \frac{T}{2}) &= \frac{L}{u_{m}} \left[ Q_{ref}(kT_s) + Q(kT_s) \right] - \frac{aOL\sin(\omega T_s/2)}{u_{m}[1 - \cos(\omega T_s/2)]} \left[ P_{ref}(kT_s) - P(kT_s) \right] \\
u_{abq}(n \frac{T}{2}) &= \frac{L}{u_{m}} \left[ P_{ref}(kT_s) + P(kT_s) \right] + \frac{aOL\sin(\omega T_s/2)}{u_{m}[1 - \cos(\omega T_s/2)]} \left[ Q_{ref}(kT_s) - Q(kT_s) \right]
\end{aligned}
\]

(27)

D. Volt-second Compensation and boundary optimization

Traditionally, \( u_{abd} \) and \( u_{abq} \) are transformed into \( u_{ab\dagger} \) and \( u_{ab\ddagger} \) by the rotating coordinate transformation matrix, which can be expressed as

\[
\begin{bmatrix}
u_{ab\dagger} \\
u_{ab\ddagger}
\end{bmatrix} = \begin{bmatrix}
\sin(\omega T_s) & -\cos(\omega T_s) \\
-\cos(\omega T_s) & \sin(\omega T_s)
\end{bmatrix} \begin{bmatrix}
u_{abd} \\
u_{abq}
\end{bmatrix}
\]

(28)

After that, PWM scheme [38] will be applied to generate drive pulse, which should satisfy volt-second balance. However, the volt-second balance can hardly hold due to the pulse generation process in the actual digital controller shown in Fig. 3. Therefore, volt-second need be compensated before updating duty cycle.

The discrete \( u_{ab\dagger} \) in Fig. 3 should ensure volt-second balance within \( nth \) half switching interval. According to (28), \( u_{ab\dagger} \) satisfies

\[
u_{ab\dagger} = \frac{1}{2} \int \frac{T}{2} \frac{T}{2} \sin(\omega T_s) dt + \frac{1}{2} \int \frac{T}{2} \frac{T}{2} \cos(\omega T_s) dt
\]

(29)

Simplifying (29), \( u_{ab\dagger} \) can be expressed as

\[
u_{ab\dagger} = \frac{2}{\omega T_s} u_{ab\dagger}[\cos(n\frac{\omega T_s}{2}) - \cos(n\frac{\omega T_s}{2} + \frac{\omega T_s}{2})] + \frac{2}{\omega T_s} u_{ab\dagger}[\sin(n\frac{\omega T_s}{2}) + \frac{\omega T_s}{2}] - \sin(n\frac{\omega T_s}{2})]
\]

(30)

Similarly, the fictitious voltage \( u_{ab\ddagger} \) is shown as

\[
u_{ab\ddagger} = \frac{2}{\omega T_s} u_{ab\ddagger}[\sin(n\frac{\omega T_s}{2}) + \frac{\omega T_s}{2}] - \sin(n\frac{\omega T_s}{2})]
\]

(31)

\( u_{ab\dagger} \) in (30) should replace \( u_{ab\dagger} \) in (28) to guarantee volt-second balance, especially at low switching frequency, and \( u_{ab\ddagger} \) is used to calculate instantaneous power in FAE shown in Fig. 2.

The modulated wave \( u_{ab\dagger} \) is limited by the dc-link voltage of the adopted converter, which signifies that \( |u_{ab\dagger}| \) should be lower than \( u_{dc} \) in normal condition. If the \( |u_{ab\dagger}| \) is larger than \( u_{dc} \), the optimal solution of the cost function in (23) is out of the linear operation range. Actually, in this condition, the input voltage value \( u_{ab\dagger} \) of the adopted converter retains \( u_{dc} \) or \( -u_{dc} \). And the fictitious voltage \( u_{ab\ddagger} \) keeps zero within one switching interval. Therefore, \( d \)-axis and \( q \)-axis components of \( u_{ab\dagger} \) can be rewritten as

\[
\begin{bmatrix}
u_{ab\dagger} \\
u_{ab\ddagger}
\end{bmatrix} = \begin{bmatrix}
\sin(\omega T_s) & -\cos(\omega T_s) \\
-\cos(\omega T_s) & \sin(\omega T_s)
\end{bmatrix} \begin{bmatrix}
\sgn(u_{ab\dagger}) u_{dc} \sin(\omega T_s) \\
\sgn(u_{ab\dagger}) u_{dc} \cos(\omega T_s)
\end{bmatrix}
\]

(32)

Where, \( u_{ab\dagger} \) and \( u_{ab\ddagger} \) are adopted to replace \( d \)-axis and \( q \)-axis components solved by (27) in this condition, respectively. \( \sgn(u_{ab\dagger}) \) represents the polarity of \( u_{ab\dagger} \). The reference voltages \( u_{ab\dagger} \) and \( u_{ab\ddagger} \) should be recalculated from (30) and (31).

The flow diagram of the proposed MP-DPC is drawn as Fig. 5. The first step is to sample the required signals and estimate active power and reactive power based on FAE shown in Fig.
2 and (6). The second step is to estimate the phase angle and frequency of the main voltage using phase lock loop (PLL) [43] for rotational coordinate transformation and obtain the active and reactive reference powers. Then, the third step is to calculate the optimal values \( u_{abdl} \) and \( u_{abdq} \) in (27) for obtaining the desired modulation wave \( u_{abl}^* \) and \( u_{abq}^* \) from (30) and (31), respectively. If the \( u_{abl}^* \) is out of the linear operation range, \( u_{abdl} \) and \( u_{abdq} \) will be recalculated by (32) for the aim of boundary optimization.

IV. ANALYSIS OF INDUCTANCE MISMATCH

A. The analysis of inductance mismatch

Almost all MPC schemes are sensitive to system parameters, and the proposed MPC scheme is also no an exception. Usually, the parasitic resistance \( R \) is much smaller than inductive reactance in the adopted converter. Therefore, the control precision is mainly affected by inductance mismatch.

It is assumed that \( L_n \) is used in the proposed method, which is not equal to the actual inductance \( L \). \( u_{abdl} \) and \( u_{abdq} \) in (27) can be rewritten as

\[
\begin{align*}
\left\{ \begin{array}{l}
\left( \frac{T}{2} \right) = u_m - \frac{\omega L_n (kT_e)}{u_m} [Q_{ref} (kT_e) + P(kT_e)]
\left( \frac{T}{2} \right) = \frac{\omega L_n (kT_e)}{u_m} [P_{ref} (kT_e) - P(kT_e)]
\left( \frac{T}{2} \right) = \frac{\omega L_n (kT_e)}{u_m} [Q_{ref} (kT_e) - Q(kT_e)]
\end{array} \right.
\end{align*}
\]

The active and reactive powers will remain constant in the steady state, which can be described as

\[
\begin{align*}
\left\{ \begin{array}{l}
P(n \frac{T}{2} + \frac{T}{2}) = P(n \frac{T}{2}) = P(kT_e)
Q(n \frac{T}{2} + \frac{T}{2}) = Q(n \frac{T}{2}) = Q(kT_e)
\end{array} \right.
\]

Substituting (33) into (20) and (21) to solve the power predictive values, which are substituted into (34), and then the active and reactive powers in inductance mismatch condition can be expressed as

\[
\begin{align*}
P(kT_e) &= \frac{\eta [P_{ref} (kT_e) \cos(\omega \frac{T}{2}) + Q_{ref} (kT_e) \sin(\omega \frac{T}{2})]}{1 + 2\eta [1 - \cos(\omega \frac{T}{2})]}
+ \frac{(\eta + 1)P_{ref} (kT_e)}{1 + 2\eta [1 - \cos(\omega \frac{T}{2})]}
\frac{Q(kT_e)}{1 + 2\eta [1 - \cos(\omega \frac{T}{2})]}
\end{align*}
\]

Where the deviation rate of inductance \( \eta \) is defined as

\[
\eta = \frac{L_n (kT_e) - L_n (kT_e)}{L_n (kT_e)}
\]

The reference reactive power \( Q_{ref} \) is usually zero. Therefore, the normalized active and reactive powers are defined as

\[
\begin{align*}
P(n \frac{T}{2} + \frac{T}{2}) &= P(n \frac{T}{2}) = \frac{1 + \eta - \cos(\omega \frac{T}{2})}{1 + 2\eta [1 - \cos(\omega \frac{T}{2})]}
+ \frac{(\eta + 1)P_{ref} (kT_e)}{1 + 2\eta [1 - \cos(\omega \frac{T}{2})]}
\frac{Q(kT_e)}{1 + 2\eta [1 - \cos(\omega \frac{T}{2})]}
\end{align*}
\]

According to (38), the pattern of \( P \) and \( Q \) with respect to \( \eta \) and \( T_s \) are shown as in Fig.6(a) and (b), respectively.
increasing of inductance error in constant switching interval, and \( Q \) is just opposite.

B. Inductance estimation

From (38), the result of dividing \( P \) by \( Q \) can be expressed as

\[
\frac{Q(kT_e)}{P(kT_e)} = \frac{\eta \sin \left( \frac{\omega T_e}{2} \right)}{1 + \eta - \eta \cos \left( \frac{\omega T_e}{2} \right)}
\]  

\[ (39) \]

According to (39) and (37), the deviation of inductance \( L_e \) can be solved as

\[
L_e(kT_e) = L(kT_e) - L_e(kT_e) = \frac{Q(nT_e)L_n(kT_e)}{P(nT_e)\sin \left( \frac{\omega T_e}{2} \right) + Q(nT_e)\cos \left( \frac{\omega T_e}{2} \right) - Q(nT_e)}
\]

\[ (40) \]

Therefore, the real inductance \( L \) can be estimated as

\[
L(kT_e) = L_n(kT_e) + L_e(kT_e)
\]

\[ (41) \]

There are switching noise and sampling errors in the active and reactive powers in the actual system. In order to suppress system noise, coefficient \( \lambda \) is introduced into inductance estimation, the compensated inductance \( L_{\text{comp}} \) can be expressed as

\[
L_{\text{comp}}(kT_e) = L_n(kT_e) + \lambda L_e(kT_e) \quad (0 < \lambda \leq 1)
\]

\[ (42) \]

V. EXPERIMENTAL TEST

A. steady state

A scaled-down experimental platform is developed to verify the performance of the proposed scheme. The system consists of a digital signal processor (DSP) TMS320F28335, and the main circuit with IGBT module F3130R06W1E3_B11. The main parameters of this platform are listed in Table I. The proposed algorithm implemented in this experimental platform has been described as shown in Fig. 5. And the other control strategies are implemented at 10- to 30-kHz sampling frequency for the aim of comparison during the experimental test.

<table>
<thead>
<tr>
<th>TABLE I</th>
<th>EXPERIMENTAL PARAMETER</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameter</td>
<td>Value</td>
</tr>
<tr>
<td>The rated main voltage (RMS)</td>
<td>65V</td>
</tr>
<tr>
<td>The rated load</td>
<td>30Ω</td>
</tr>
<tr>
<td>The rated DC voltage</td>
<td>120V</td>
</tr>
<tr>
<td>The equivalent inductance</td>
<td>5.6mH</td>
</tr>
<tr>
<td>The DC-link capacitor</td>
<td>3300μF</td>
</tr>
</tbody>
</table>

The traditional control methods including MPC based on single vector [19], optimal duty cycle [42], optimal duty cycle with constant switching frequency [26], PI-based DPC [11] and MPC with the optimal modulated wave [29], which are respectively named as MPC-I, MPC-II, MPC-III, PI-DPC and MPC-IV in this paper, are compared with the proposed MPC method during the experimental tests. The experimental waveforms and total harmonic distortion (THD) in the steady state at 20kHz sampling frequency are shown in Fig. 7, where the switching frequency in MPC-III, PI-DPC, MPC-IV and the proposed method are same(2kHz). At the same sampling frequency of 20kHz, the MPC-IV and the proposed MPC present much better performance than the MPC-1, MPC-II, MPC-III, and PI-DPC in terms of less current harmonics. Furthermore, there is a wide harmonic spectrum for MPC-I and MPC-II, which is not easy to be filtered. On the contrary, the dominated harmonics of the proposed MPC and traditional MPCSs with constant switching frequency (MPC-III, PI-DPC, and MPC-IV) distribute around the double switching frequency (4kHz), which is favorable for filter design, and the current control precision of the proposed method and MPC-IV are much better than that of MPC-III and PI-DPC from the THD of the line current in these methods.

Fig. 7 Experimental waveforms and FFT results of the line current in steady state in four MPC methods at 2kHz switching frequency: (a) MPC-I, (b) MPC-II, (c) MPC-III, (d) PI-DPC, (e) MPC-IV, and (f) Proposed MPC (u: 50V/div, \( u_{d1} \) and \( u_{d2} \): 20V/div, i: 10A/div, Time: 10ms/div)
A quantitative comparison of these six schemes with control frequency from 10- to 30-kHz in rated power is shown in Table II. It is clear from Table II that the switching frequency of both MPC-I and MPC-II increases with the sampling frequency, and the former spreads from 4.1kHz to 13.1kHz when the sampling frequency increases from 10kHz to 30kHz. The later always keeps and locates below the sampling frequency. Therefore, both MPC-I and MPC-II are not suitable for the condition of the low switching frequency due to the high sampling frequency demand in these traditional MPC. It is different from MPC-I and MPC-II based on finite-control-set, MPC-III, PI-DPC, MPC-IV and the proposed MPC by using PWM stage can keep constant switching frequency with the different sampling frequency, and present less power ripple and lower THD than MPC-I and MPC-II. The proposed MPC has a slight advantage in the line current THD and power fluctuation over MPC-IV at 2kHz switching frequency. The power ripples of MPC-III and PI-DPC are obviously reduced with the increase of control frequency. On the contrary, power ripples in PI-DPC and the proposed MPC always retain the small value with the changing of control frequency, which indirectly verifies that high accurate power prediction and deadbeat power control is achieved by PI-DPC and the proposed method.

<table>
<thead>
<tr>
<th>$f_s$ Methods</th>
<th>$f_{sw}(kHz)$</th>
<th>$P_{sw}(W)$</th>
<th>$Q_{sw}(Var)$</th>
<th>THD(%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10kHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MPC-I</td>
<td>4.1</td>
<td>162</td>
<td>150</td>
<td>9.97</td>
</tr>
<tr>
<td>MPC-II</td>
<td>9.5</td>
<td>140</td>
<td>126</td>
<td>9.33</td>
</tr>
<tr>
<td>MPC-III</td>
<td>2.0</td>
<td>85.6</td>
<td>83.7</td>
<td>7.07</td>
</tr>
<tr>
<td>PI-DPC</td>
<td>2.0</td>
<td>78.1</td>
<td>73.7</td>
<td>6.75</td>
</tr>
<tr>
<td>MPC-IV</td>
<td>2.0</td>
<td>45.1</td>
<td>42.3</td>
<td>3.07</td>
</tr>
<tr>
<td>Proposed-MPC</td>
<td>2.0</td>
<td>43.3</td>
<td>40.8</td>
<td>3.07</td>
</tr>
<tr>
<td>20kHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MPC-I</td>
<td>7.3</td>
<td>121</td>
<td>108</td>
<td>6.93</td>
</tr>
<tr>
<td>MPC-II</td>
<td>18.5</td>
<td>105.3</td>
<td>86.5</td>
<td>6.01</td>
</tr>
<tr>
<td>MPC-III</td>
<td>2.0</td>
<td>75.7</td>
<td>72.7</td>
<td>5.98</td>
</tr>
<tr>
<td>PI-DPC</td>
<td>2.0</td>
<td>75.1</td>
<td>73.1</td>
<td>5.34</td>
</tr>
<tr>
<td>MPC-IV</td>
<td>2.0</td>
<td>45.5</td>
<td>42.1</td>
<td>3.15</td>
</tr>
<tr>
<td>Proposed-MPC</td>
<td>2.0</td>
<td>43.0</td>
<td>40.6</td>
<td>2.97</td>
</tr>
<tr>
<td>30kHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MPC-I</td>
<td>13.1</td>
<td>104</td>
<td>101</td>
<td>6.28</td>
</tr>
<tr>
<td>MPC-II</td>
<td>27.0</td>
<td>72.5</td>
<td>71.2</td>
<td>5.61</td>
</tr>
<tr>
<td>MPC-III</td>
<td>2.0</td>
<td>70.2</td>
<td>69.5</td>
<td>5.60</td>
</tr>
<tr>
<td>PI-DPC</td>
<td>2.0</td>
<td>68.3</td>
<td>67.5</td>
<td>4.91</td>
</tr>
<tr>
<td>MPC-IV</td>
<td>2.0</td>
<td>45.1</td>
<td>42.3</td>
<td>3.07</td>
</tr>
<tr>
<td>Proposed-MPC</td>
<td>2.0</td>
<td>43.2</td>
<td>40.3</td>
<td>3.03</td>
</tr>
</tbody>
</table>

Where the average switching frequency $f_{sw}$ is defined as commutation number of every switching device between on and off states within one second.

In order to further compare the performance of these MPC methods with constant switching frequency at low switching frequency, experimental results of MPC-III, PI-DPC, MPC-IV and the proposed MPC in 0.8kHz are shown in Fig. 8. From this figure, it is clear that the proposed MPC can achieve lowest THD of the line current. However, the control precision of MPC-III, PI-DPC and MPC-IV will worsen rapidly with the decrease of switching frequency. There exists phase difference between the line current and the main voltage in low switching frequency condition. Experimental results show that the control performance of the proposed MPC is much better than that of traditional constant switching control methods at low switching frequency.

Fig. 9 shows the line current THD curves of MPC-III, PI-DPC, MPC-IV and the proposed MPC methods at low switching frequency from 500 to 2000kHz. It is clear that current THD value in traditional methods is much higher than that in the proposed MPC at the low switching frequency.
B. Dynamic response

Experimental results of these six schemes at 20kHz sampling frequency under active power step-change condition are shown in Fig. 10, where the switching frequency is set to 2kHz. The active power reference steps from 300W to 480W and then back to 300W again. The reactive power is set to zero to realize unit power factor. It is clear that the dynamic performances are very excellent, but active and reactive powers in traditional methods are not decoupled completely due to the limitation of power estimation method in single phase system. Comparing with the traditional schemes, since the advanced FAE is applied to calculate instantaneous power by using the optimal $u_{abd}$ and $u_{abq}$ in the proposed MPC, power estimation precision is improved obviously. And the power decoupling ability in the proposed MPC is more excellent than that in other five schemes.

![Fig. 10 Experimental results in active power step change condition at 2kHz switching frequency: (a) MPC-I, (b) MPC-II, (c) MPC-III, (d) PI-DPC, (e) MPC-IV, and (f) Proposed MPC (P: 120W/div, Q: 120var/div, and $i$: 10A/div, Time: 50ms/div)](image)

Experimental results of MPC-III, PI-DPC, MPC-IV and the proposed MPC at low switching frequency (0.8kHz) are further shown in Fig. 11. It is seen that the transient response of the proposed MPC is faster than that of MPC-III, PI-DPC, MPC-IV. The obvious steady state power errors appear in MPC-III and MPC-IV under this condition. It means that the power prediction precision and the control performance of MPC-III, PI-DPC, MPC-IV are unsatisfactory in this case, even if the switching frequency keeps constant by PWM. Comparatively, from Fig. 11(d), the active and reactive powers track their references without steady state error, and are decoupled independently in the proposed MPC.

![Fig. 11 Experimental results in active power step change condition at 2kHz switching frequency: (a) MPC-I, (b) MPC-II, (c) MPC-III, (d) PI-DPC, (e) MPC-IV, and (f) Proposed MPC (P: 120W/div, Q: 120var/div, and $i$: 10A/div, Time: 50ms/div)](image)
Dynamic response of the proposed method in external load disturbance is shown in Fig. 12, where load steps up from 300W to 480W and then steps down to 300W (load resistance steps from 48Ω to 30Ω). It is seen that the reactive power is always kept to zero, and DC-link voltage recovers constant value rapidly after the load step change. The experimental results verify its strong robustness against load disturbance.

C. Inductance parameter mismatch

The experimental results of the proposed MPC method under the inductance mismatch condition at switching frequency 2kHz and 4kHz are shown in Fig. 13(a) and (b), respectively. Where $L_n$ is set to 0.8mH in programming. From this figure, it is clear that the active power is lower than its reference, and there is a positive reactive power in this condition. Comparing Fig. 13(a) and (b), the power errors will reduce with an increase of switching frequency, which is consistent with the description in Fig. 6. It verifies the correctness of theoretical analysis. Therefore, power errors caused by the inductance mismatch don't require to be compensated if the proposed method is applied in high switching frequency.

From the above experimental results and theoretical analysis, the performance comparison of four methods applied in single phase converters can be demonstrated as Table III, which shows the excellent performance of the proposed method in these aspects: algorithm complexity, steady state and dynamic characteristics.

<table>
<thead>
<tr>
<th>Performance</th>
<th>Scheme</th>
<th>Performance</th>
<th>Scheme</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switching frequency</td>
<td>variable</td>
<td>variable</td>
<td>constant</td>
</tr>
<tr>
<td>Operation time of the algorithm</td>
<td>18.5μs</td>
<td>19.3μs</td>
<td>20.1μs</td>
</tr>
<tr>
<td>THD of line current</td>
<td>high</td>
<td>high</td>
<td>medium</td>
</tr>
<tr>
<td>Dynamic response</td>
<td>fast</td>
<td>fast</td>
<td>normal</td>
</tr>
<tr>
<td>Power decoupling</td>
<td>medium</td>
<td>medium</td>
<td>medium</td>
</tr>
<tr>
<td>Steady state power error</td>
<td>no</td>
<td>no</td>
<td>yes</td>
</tr>
</tbody>
</table>

TABLE III

PERFORMANCE COMPARISON OF FOUR MPC METHODS
VI. CONCLUSION

In this paper, an MP-DPC combining with PWM stage is proposed to achieve high control accuracy for the single phase three-level converters in low switching frequency application. In the proposed method, the optimal d-axis and q-axis components of the input voltage in the adopted converter can be deduced from the simple equations based on cost function minimization and the accurate power predictive model. And then the desired modulated wave is obtained by dq-dq axis transformation, which is set as the input reference modulated wave in PWM to generate pulse driving signals for each switching device. The effect of inductance mismatch on the active and reactive powers is discussed, and an inductance on-line estimation method is discussed for the MP-MPC. By comparing the experimental results of the proposed MP-DPC scheme with those of the traditional control schemes including MPC based on single vector [19], optimal duty cycle [42], optimal duty cycle with constant switching frequency [26], PI-based DPC [11] and optimal modulated wave [29], it can be concluded that the salient features of the proposed MP-DPC scheme with inductance parameter mismatch compensation are as follows:

1) Compared to traditional finite-control-set MP-DPC [19], [42], the proposed method can immensely simplify calculation and achieve constant switching frequency by adopting PWM.

2) Contrasting with traditional DPC combining with PWM [11], [26], [29], the proposed method can gain higher control precision at the low switching frequency, due to applying an exact instantaneous power predictive model in dq coordinate system to predict the active and reactive power within next half switching interval other than one switching interval. And the deadbeat power control to compensate the digital delay is achieved by this way. Meanwhile, the active and reactive power can be decoupled more thoroughly in the proposed MPC than the traditional methods.

3) The proposed inductance estimation method can improve the robustness and precision of the proposed method, and is almost not affected by power fluctuation.

REFERENCES


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