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Performance Improvement of Model Predictive Control Using Control Error Compensation for Power Electronic Converters Based on the Lyapunov Function

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Abstract

This paper proposes a model predictive control based on the discrete Lyapunov function to improve the performance of power electronic converters. The proposed control technique, based on finite control set model predictive control (FCS-MPC), defines a cost function about control law which is determined under the Lyapunov stability theorem with a control error compensation. The steady state and dynamic performance of the proposed control strategy has been tested under the single phase AC/DC voltage source rectifier (S-VSR). The experimental results demonstrate that the proposed control strategy not only offers global stability and good robustness but also leads to a high quality sinusoidal current with a reasonably low total harmonic distortion (THD) and a fast dynamic response under linear loads.

Key words: Power electronic converters, Discrete Lyapunov function, Model predictive control, Lyapunov stability theorem, Robustness

I. ITRODUCTION

Power electronic converters are widely used in the national economy and in people's livelihoods. The control strategies of converters have gradually become a research beacon concerning the requirement of power quality [1-3].

With the development of high speed and powerful digital signal processors (DSPs) and microprocessors, growing attention and interest have been attracted to the use of model predictive control (MPC) in power electronics. Generally, the MPC techniques applied to power electronics have been classified into two main categories: continuous control set MPC (CCS-MPC) and finite control set MPC (FCS-MPC) [4-5]. In CCS-MPC, a modulator using sinusoidal pulse width modulation (SPWM) or space vector pulse width modulation (SVPWM) generates the switching states starting from the continuous output of the predictive controller [6-7]. On the other hand, the FCS-MPC takes advantage of the discrete nature of power converters for solving the optimization problem [8-9]. Without the modulation stage, FCS-MPC applies the direct control action to the converter.

The conventional FCS-MPC employs one voltage vector during one sampling period, and needs a high sampling frequency to achieve a better performance. In one sampling period, the FCS-MPC consists of two main steps: prediction of the behavior for the next sampling instant for all possible voltage vectors and evaluation of the cost function for each prediction, and to find the optimal voltage vector based on the traversal algorithm. The fact increases the computation burden [10-12]. Furthermore, due to the limited number of voltage vectors in the converter, the performance improvement caused by the conventional FCS-MPC is limited, and the THD of the controlled variable is higher than the conventional control based on modulator [13-14].

The Lyapunov function based control strategy is powerful for considering global stability and robustness, several works have been published in the literature, such as [15-21]. Using discrete energy function to achieve the superior performance and global asymptotic stability for a boost PFC converter used in electric vehicles in [15-17]. In [18-19], the Lyapunov function based control approach was applied for single phase inverter with LCL filter and single phase inverter with LC filter, respectively. In [20-21], the three phase AC-DC voltage source rectifier achieved a fast dynamic performance by adopting the Lyapunov function based control strategy, especially, the proposed control approach was modified with model predictive control in [21]. In this paper, a model predictive control based on discrete Lyapunov function is proposed to improve the control performance by adding the error term of the controlled variable and the reference variable to the control law. The discrete model of the S-VSR and the principle of the conventional FCS-MPC are elaborately described in Section II. The control law is calculated using the Lyapunov stability theorem based on the discrete Lyapunov function and the proposed control strategy is given in Section III. In Section IV, the control coefficient,α, of the error term in the control law is selected by analyzing its influence on the steady state and the dynamic performance regarding the stability and the robustness. In Section V, the performance of the proposed method for the S-VSR is investigated with an experimental system, and the experimental results are presented and compared with the conventional FCS-MPC. Finally, conclusions are drawn in Section VI.

II. CONVENTIONAL FCS-MPC

Fig. 1 shows an S-VSR. The equation describing the operation of the converter can be written as

$$L_s \frac{di}{dt} = e - Ri - V_r \tag{1}$$

where:

e the grid voltage;

V_r the rectifier voltage;

i the grid current;

R the equivalent series resistance;

L_s the inductance of the line filter.

The discrete model of the converter is obtained to approximate the derivative di/dt in (1) by

$$\frac{di}{dt} = \frac{i(k+1) - i(k)}{T} \tag{2}$$

where:

T the sampling time.

By substituting (2) into (1), the following expression is obtained for the future current at (k+1)th instant, from (1), the equivalent eddy currents are straight- forwardly derived as follows:

$$i(k+1) = (1 - \frac{RT}{L_s})i(k) + \frac{T}{L_s}\left[e(k) - V_r(k+1)\right] \tag{3}$$

where:

$V_r(k+1)$ the future rectifier voltage of S-VSR, and it is a continuous vector.

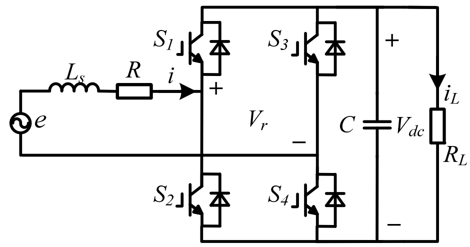


Fig. 1. Single phase AC/DC voltage source rectifier

There are three voltage vectors, which can predict three future current values. The conventional FCS-MPC method is based on this property.

$$i^p(k+1) = (1 - \frac{RT}{L_s})i(k) + \frac{T}{L_s}\left[e(k) - V(k+1)\right] \tag{4}$$

where:

$i^p(k+1)$ the future predicted current;

$V(k+1)$ the discrete voltage vector of S-VSR, and is selected from the three voltage vectors 0,$-V_{dc}$ and V_{dc}.

Using this property, we can use the following cost function to select the optimal switching state at the next step as

$$J = \left| i^p(k+1) - i^*(k+1) \right| \tag{5}$$

where:

$i^*(k+1)$ the future reference current value.

The optimal future switching state selected from the cost function J can force the future current value to approach the reference current value at the next step. Finally, the selected voltage state which can minimize the current error is applied to the rectifier at the next sampling instant.

III. PROPOSED CONTROL STRATEGY BASED ON DISCRETE LYAPUNOV FUNCTION

An effective control algorithm is essential for S-VSR so that the current, $i(k)$,tracks the reference value, $i^*(k)$. Therefore, it is necessary to find a control function such that the current tracking error, $\Delta i(k)$, asymptotically converges to zero. The Lyapunov direct method is used for the specific application.

And the error $\Delta i(k)$ is taken as

$$\Delta i(k) = i(k) - i^*(k) \tag{6}$$

According to the Lyapunov stability theorem, the discrete Lyapunov function, $L(x(k))$, satisfies the following properties:

1）$L(0)=0$；

2）$L(x(k))>0$ for all $x(k)\neq 0$；

3）$L(x(k)) \rightarrow \infty$ as $\|x(k)\| \rightarrow \infty$；

4）$\Delta L(x(k))<0$ for all $x(k)\neq 0$.

Thus, the discrete Lyapunov function $L(\Delta i(k))$ of the S-VSR could be taken as

$$L(\Delta i(k)) = \frac{1}{2}\Delta i^2(k) \tag{7}$$

From (6) and (7), the rate of change of the Lyapunov function, $L(\Delta i(k))$, can be expressed for the rectifier mode as

$$\begin{aligned}\Delta L(\Delta i(k)) &= L(\Delta i(k+1)) - L(\Delta i(k)) \\ &= \frac{1}{2}\left[i(k+1) - i^*(k+1)\right]^2 \\ &\quad - \frac{1}{2}\left[i(k) - i^*(k)\right]^2\end{aligned} \tag{8}$$

To satisfy the Lyapunov stability theorem we define the following expression

$$i(k+1) - i^*(k+1) = \alpha\left[i(k) - i^*(k)\right] \tag{9}$$

where α is a control coefficient with constant value.

Substituting (9) into (8), and obtain the following expression for $\Delta L(\Delta i(k))$.

$$\begin{aligned}\Delta L(\Delta i(k)) &= L(\Delta i(k+1)) - L(\Delta i(k)) \\ &= \frac{1}{2}(\alpha^2 - 1)\left[i(k) - i^*(k)\right]^2\end{aligned} \tag{10}$$

It is apparent that $\Delta L(\Delta i(k)) < 0$, if α is chosen as

$$0 < \alpha < 1 \tag{11}$$

The future control law $V_r(k+1)$ at the $(k+1)$th instant of the proposed control strategy can be determined with (3) and (9) and can be written as

$$\begin{aligned}V_r(k+1) &= e(k) + (\frac{L_s}{T} - R)i(k) - \frac{L_s}{T}i^*(k+1) \\ &\quad - \alpha\frac{L_s}{T}\left[i(k) - i^*(k)\right]\end{aligned} \tag{12}$$

It is clearly shown that (12) is not only related to the controlled variable $i(k)$ at the kth instant and the reference variable $i^*(k+1)$ at the $(k+1)$th instant but also related to the error term of the controlled variable and the reference variable at the kth instant. Therefore, the proposed control law has a feed forward and feedback structure which is the same as model predictive control. And when α=0, solving (12), the control law can be expressed as

$$\hat{V}_r(k+1) = e(k) + (\frac{L_s}{T} - R)i(k) - \frac{L_s}{T}i^*(k+1) \tag{13}$$

which is the same as the control law for the deadbeat control.

In the proposed strategy, the control law in (12) is used as the continuous future reference voltage vector, in order to choose one of three future voltage vectors of the S-VSR in a finite set. If the future voltage vector of the S-VSR closest to the future reference voltage vector obtained from (12) is applied to the S-VSR, the current at the next sampling instant could track the future reference current. Since the S-VSR only generates the three voltage vectors in its finite set in contrast to the continuous reference voltage vector in (12), the cost function defined as (14) enables one proper future voltage vector to be selected among three possible vectors.

$$G = \left|V_r(k+1) - V(k+1)\right| \tag{14}$$

IV. SELECTION AND INFLUENCE OF CONTROL COEFFICIENT

As shown in (12), the key point of the proposed control strategy is to increase the error term with the control coefficient, α, in the control law of the deadbeat control, and the selection of α is closely related to the control performance.

A. Influence of the stability

The discrete voltage vector applied to the rectifier in the next sampling period is regarded as the sum of the continuous future reference voltage vector and the quantization error vector as

$$V(k+1) = V_r(k+1) + \Delta v(k+1) \tag{15}$$

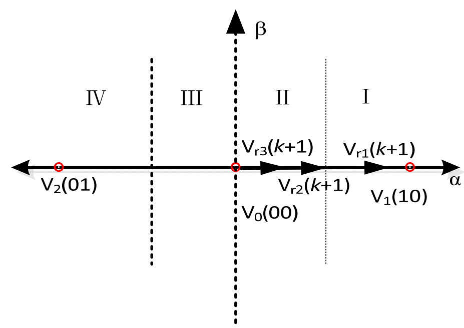


Fig. 2. Optimal voltage vector Selection

Fig.2 shows the selection principle of the optimal voltage vector based on (15) in the positive period of $e(k)$. According to the cost function (14), the optimal voltage vector $V(k+1)$ minimizes G. From Fig.2 we know that there are 3 cases for the future reference voltage vector $V_r(k+1)$ in the positive period of $e(k)$:

1): $V_r(k+1)$ is situated in Sector I as $V_{r1}(k+1)$. To minimize G, V_1 is selected. The quantization error vector satisfies $0 \le \Delta v(k+1) \le 0.5V_{dc}$.

2): $V_r(k+1)$ is situated in juncture of Sector I and Sector II as $V_{r2}(k+1)$. To minimize G, V_1 or V_0 is selected. The quantization error vector satisfies $|\Delta v(k+1)| = 0.5V_{dc}$.

3): $V_r(k+1)$ is situated in Sector II as $V_{r3}(k+1)$. To minimize G, V_0 is selected. The quantization error vector satisfies $-0.5V_{dc} \le \Delta v(k+1) \le 0$.

The selection principle of the optimal voltage vector in negative period of $e(k)$ is similar to that in positive period of $e(k)$.According to the analysis, the quantization error vector is bounded in

$$\|\Delta v(k+1)\| \le 0.5V_{dc} \tag{16}$$

The direct Lyapunov method gives the following stability criteria for a function $L(\Delta i(k))$ is uniformly and ultimately bounded [22], i.e.,

$$\begin{aligned} &L(\Delta i(k)) \ge c_1 \left|\Delta i(k)\right|^l, \quad \forall \Delta i(k) \in G \\ &L(\Delta i(k)) \le c_2 \left|\Delta i(k)\right|^l, \quad \forall \Delta i(k) \in \Gamma \\ &L(\Delta i(k+1)) - L(\Delta i(k)) < -c_3 \left|\Delta i(k)\right|^l + c_4 \end{aligned} \tag{17}$$

where c_1, c_2, c_3, and c_4 are positive constants, $l \ge 1$, $G \subseteq R_n$ is a positive control invariant set, and $\Gamma \subset G$ is a compact set.

By applying the value of future voltage vector (15) ,$V(k+1)$, for the rectifier, the rate of change of the Lyapunov function, $\Delta L^p(\Delta i(k))$, can be written as

$$\begin{aligned} \Delta L^p(\Delta i(k)) = &\frac{1}{2}[i^p(k+1) - i^*(k+1)]^2 \\ &-\frac{1}{2}[i(k) - i^*(k)]^2 \end{aligned} \tag{18}$$

Substitute (4) and (12) into (18), $\Delta L^p(\Delta i(k))$ can be written as

$$\begin{aligned} \Delta L^p(\Delta i(k)) = &\frac{1}{2}(\alpha^2 - 1)\Delta i^2(k) \\ &+\frac{1}{2}\left[\frac{T}{L_s}\Delta v(k+1)\right]^2 \\ &-\alpha\frac{T}{L_s}\Delta v(k+1)\Delta i(k) \end{aligned} \tag{19}$$

Solving (19), it can be expressed as

$$\begin{aligned} \Delta L^p(\Delta i(k)) = &(\frac{1}{2} - b)(\alpha^2 - 1)\Delta i^2(k) \\ &+ p(\Delta i(k)), \\ p(\Delta i(k)) = &b(\alpha^2 - 1)\Delta i^2(k) \\ &-\alpha\frac{T}{L_s}\Delta v(k+1)\Delta i(k) \\ &+\frac{1}{2}\left[\frac{T}{L_s}\Delta v(k+1)\right]^2 \end{aligned} \tag{20}$$

where b is a positive constant and within $0<b<1/2$.

It is clear that $p(\Delta i(k))$ has a maximum shown as (21) based on (11)

$$p(\Delta i(k))_{\max} = \left[\frac{T}{L_s}\Delta v(k+1)\right]^2 \frac{(1-2b)\alpha^2 + 2b}{4b(1-\alpha^2)} \tag{21}$$

As a result, by considering (16), the rate of change of the Lyapunov function in (19) is

$$\begin{aligned} &\Delta L^p(\Delta i(k)) \\ &\le (\frac{1}{2} - b)(\alpha^2 - 1)\Delta i^2(k) + p(\Delta i(k))_{\max} \\ &\le (\frac{1}{2} - b)(\alpha^2 - 1)\Delta i^2(k) \\ &+\frac{1}{4}(\frac{T}{L_s}V_{dc})^2 \frac{(1-2b)\alpha^2 + 2b}{4b(1-\alpha^2)} \end{aligned} \tag{22}$$

Therefore, the stability condition (17) is satisfied by the constant values as

$$\begin{aligned} &c_1 = c_2 = 1;\ c_3 = (\frac{1}{2} - b)(1-\alpha^2); \\ &c_4 = \frac{1}{4}(\frac{T}{L_s}V_{dc})^2 \frac{(1-2b)\alpha^2 + 2b}{4b(1-\alpha^2)} \end{aligned} \tag{23}$$

And (22) can be expressed as

$$\Delta L^p(\Delta i(k)) \le -2c_3 L^p(\Delta i(k)) + c_4 \tag{24}$$

This inequality implies that, as time increases, the current control error converge to the compact set as

$$A = \left\{ \Delta i(k) \middle| \|\Delta i(k)\| \le \sqrt{\frac{c_4}{c_3}} \right\} \tag{25}$$

It is clear that when α^2 is larger, the convergence domain is larger and the convergence domain is closely to the robustness.

B. Influence of the convergence speed

From(8), the relationship between the Lyapunov function of the $(k+1)$th instant and the Lyapunov function of the kth instant can be described as

$$L(\Delta i(k+1)) = \alpha^2 L(\Delta i(k)) \tag{26}$$

The convergence speed of the Lyapunov function can be studied using ρ which is given by

$$\rho = \frac{L(k)}{L(k+1)} = 1/\alpha^2 \tag{27}$$

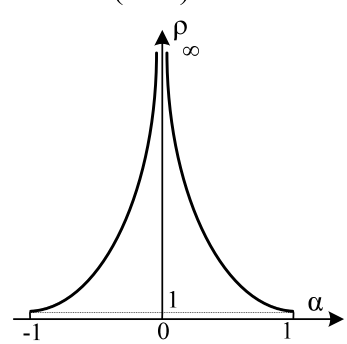


Fig. 3. The relationship between ρ and α

Fig. 3 shows ρ as a function of α. It is clear that when α^2 is gradually increased from -1 to 1, ρ is decreased and the convergence speed is also decreased. As we know, the convergence speed is closely related to the dynamic performance.

C. Influence of the steady-state performance

To study the effect of α on the steady state performance, we neglect the equivalent series resistance R and suppose that the current reference value does not change considerably in one sampling interval. The future current value can be expressed by:

$$i(k+1)=i(k)+\frac{T}{L_s}[e(k)-V_r(k+1)] \quad (28)$$

And the continuous future reference voltage vector can be expressed by:

$$V_r(k+1)=e(k)+(1-\alpha)\frac{L_s}{T}\left[i(k)-i^*(k)\right] \quad (29)$$

In the positive period of $e(k)$,we hope the $i(k+1)$ could be close to $i^*(k+1)$. When $i(k)>i^*(k)$ at the kth instant, the optimal voltage vector should be V_1. To increase the possibility of V_1 by increasing the $V_r(k+1)$, so the better range of α is $-1<\alpha<0$.When $i(k)<i^*(k)$ at the kth instant, the optimal voltage vector should be V_0. To increase the possibility of V_0 by decreasing the $V_r(k+1)$, so the better range of α is also $-1<\alpha<0$.

So, the better value range of α is given as

$$-1<\alpha<0 \quad (30)$$

D. Selection of α

According to the analyses above, when α^2 is selected to be larger within the value range, the steady-state performance and the robustness are better, but the convergence speed and the dynamic performance are worse, so we should make a compromise during the selection of α. A typical range of α for the rectifier in this study is found to be

$$-0.6<\alpha<0.2 \quad (31)$$

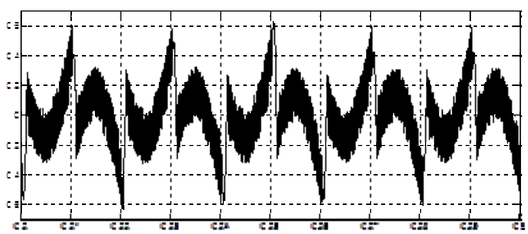
V. SIMULATION AND EXPERIMENTAL

The proposed control strategy has been verified by simulation and experimental results. The Simulations were carried out by MATLAB/Simulink. The experimental test was performed using a single phase PWM AC/DC voltage source rectifier prototype in a fully DSP system based on a TMS320F28069.The system parameters are given in Table I.

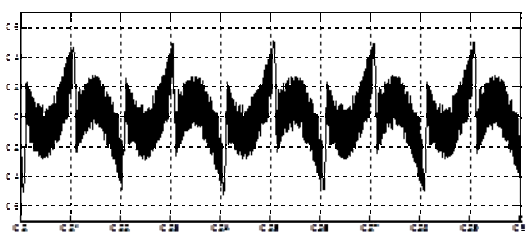
TABLE I
SYSTEM PARAMETERS

System parameters	Symbol	Value
AC voltage(RMS)	e	$50V/50Hz$
Filter inductance	L_s	$6mH$
Equivalent series resistance	R	0.3Ω
DC side capacitor	C	$1000\mu F$
DC side voltage	V_{dc}	$100V$
Sampling frequency	T	5e-5s

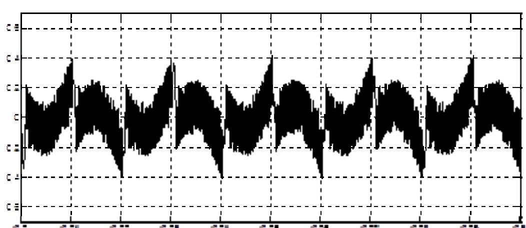
A. Simulation results



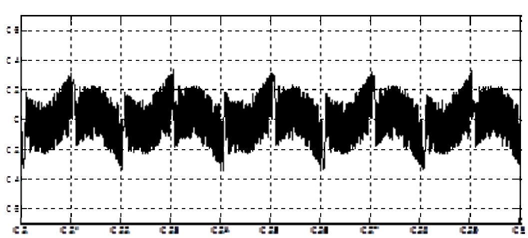
(a) α=0.2



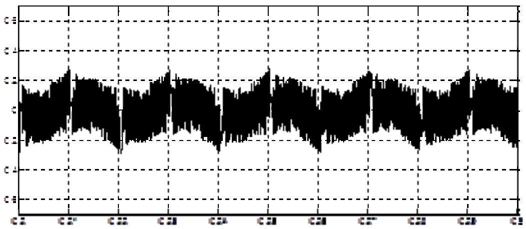
(b) α=0



(c) α=－0.2



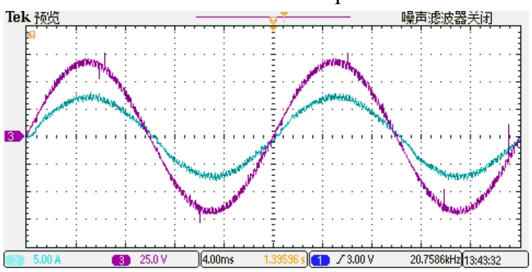
(d) α=－0.5



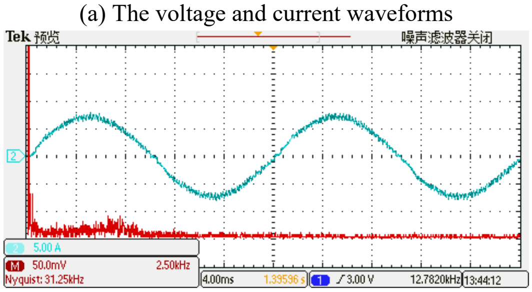
(e) α=−0.8

Fig.4. Simulated error $\Delta i(k)$ in steady state based on different values of α (0.2A/div).

Fig.4 shows the simulation results of the steady-state error of $i(k)$ based on different values of α, it is clear that the smaller the α, the better the steady-state performance. In this paper, α is selected as −0.45 in the experimental test.

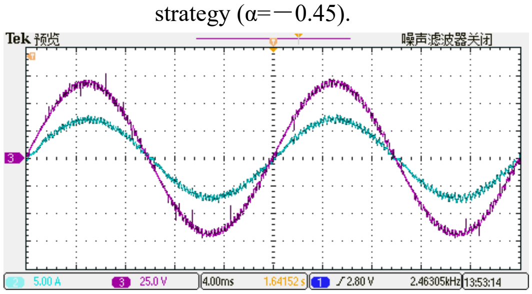


(a) The voltage and current waveforms

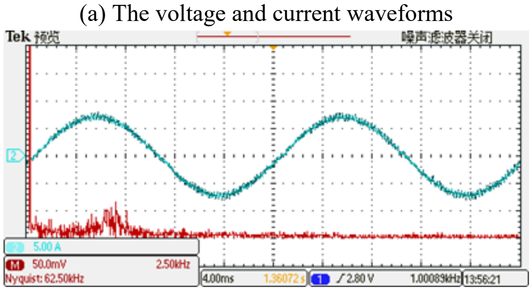


(b) Harmonic spectrum of current

Fig.5. Steady-state test result of the proposed control strategy (α=−0.45).

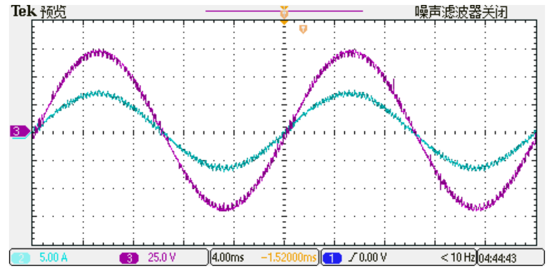


(a) The voltage and current waveforms

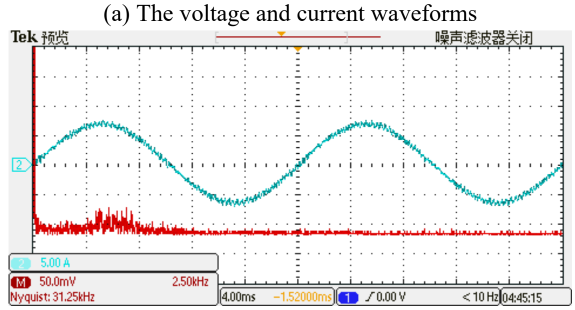


(b) Harmonic spectrum of current

Fig.6. Steady state test result of the conventional FCS-MPC.



(a) The voltage and current waveforms



(b) Harmonic spectrum of current

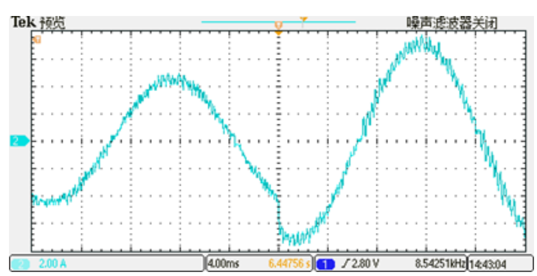
Fig.7. Steady-state test result of the online parameter estimation control strategy in [23]

B. Steady-State Response Tests

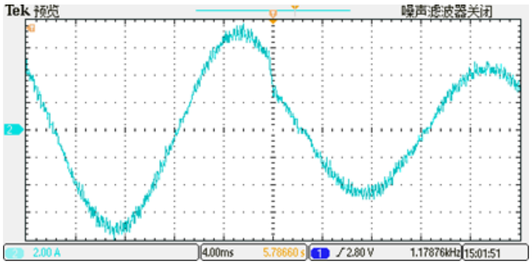
Fig. 5 shows the steady-state test result of the proposed control strategy when the reference current peaks at 6.8 A. Fig.5(a)depicts the input current and voltage waveforms, where current follows the voltage to achieve the unit power factor. Fig.5(b) is the harmonic spectrum of current. The test results demonstrate the improved steady-state response of the proposed control strategy. The converter input current is highly sinusoidal with a measured total harmonic distortion (THD) of 2.16%.

Fig. 6 shows the test result of the conventional FCS-MPC. As observed from the current waveform, the fluctuation range of the current is larger, and the THD is 3.38%.

Fig.7 shows the test result of the online parameter estimation control method referred in[23]. The THD of the current waveform is 2.51%. From the comparison above, we can find that the proposed control method has the best steady-state response, then the online parameter estimation FCS-MPC control method. The Steady-state test result of the conventional FCS-MPC is worst.

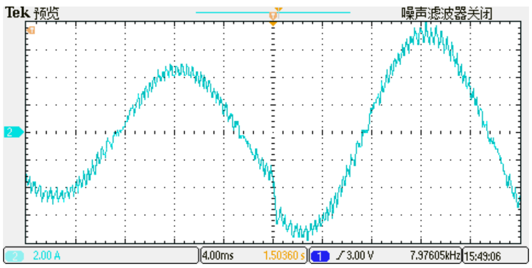


(a) Step change from 4- to 6.8- A peak

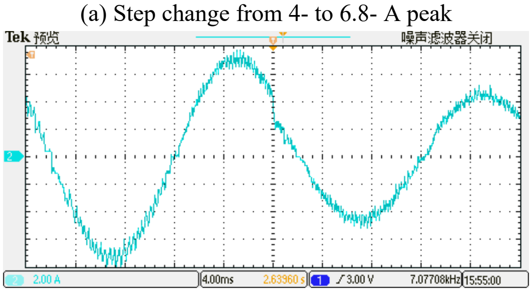


(b) Step change from 6.8- to 4- A peak

Fig.8. Current behavior during reference step with the proposed control strategy (α=−0.45).



(a) Step change from 4- to 6.8- A peak

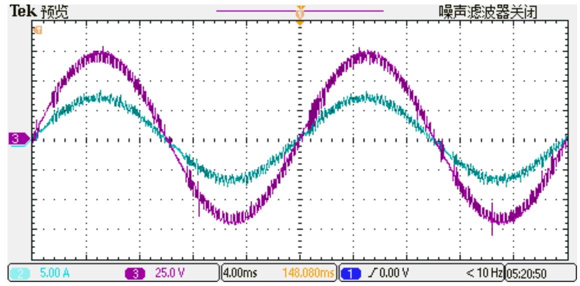


(b) Step change from 6.8- to 4- A peak

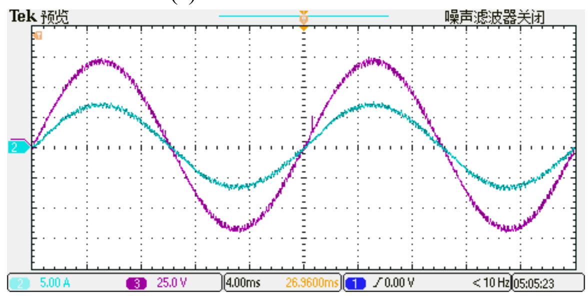
Fig.9. Current behavior during reference step with the conventional FCS-MPC.

C. *Dynamic Response Tests*

An important aspect of any control system is the dynamic response to changes in the reference. Fig. 8 depicts the current behavior with the proposed control strategy (α=−0.45) when the reference step change from 4- to 6.8- A peak and vice versa; the current reached a steady-state level in Fig. 8(a) within 126μs and requires 268μs to reach a steady state in Fig. 8(b). Compared with the proposed control strategy, the conventional FCS-MPC has a faster dynamic response as the analysis in section IV. Fig.9 shows the current behavior with conventional FCS-MPC when the reference step change from 4- to 6.8- A peak and 6.8- to 4- A peak. The dynamic response time in Fig. 9(a) is 107μs and that in Fig. 9(b) is 231μs. The experiment result is consistent with the analysis in Section IV.



(a) *L* mismatch with -25%



(b) *L* mismatch with +25%

Fig.10. Steady-state test result of the proposed control strategy when the inductance *L* had a mismatch.

D. *Robustness Tests*

The system parameters, such as the inductance and the equivalent resistance, vary with temperature, core saturation, and other environmental conditions, and the parameter errors influence the whole control performance. The robustness of the proposed control strategy is tested when the actual inductance is mismatch with -25% and +25%.

Fig. 10 shows the steady-state test result of the proposed control strategy when the inductance has a mismatch with -25% and +25%.we can find that when the inductance has a mismatch with -25%, there are higher input current ripples. But inductance mismatch does not influence the system stability in the proposed control strategy.

VI. Conclusions

A model predictive control based on discrete Lyapunov function with a control error compensation of power electronic converters is proposed in this paper. The criterion for selecting the control coefficient,α, is described. Furthermore, the influence of changingαis also studied.

The proposed control strategy based on the discrete direct Lyapunov method leads to a globally asymptotically stable system, shows improved steady-state performance and has a fast dynamic response just a little slower than the conventional FCS-MPC. The results associated in this investigation are very encouraging and will continue to play a strategic role in the improvement of modern digital control systems.

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