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Robust Deadbeat Current Control Method for Three-Phase Voltage-Source Active Power Filter

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ABSTRACT

This paper is concerned with a deadbeat current control implementation of shunt-type three-phase active power filter (APF). Although the one-dimensional deadbeat control method can attain time-optimal response of APF compensating current, one sampling period is actually required for its settling time. This delay is a serious drawback for this control technique. To cancel such a delay and one more delay caused by DSP execution time, the desired APF compensating current has to be predicted two sampling periods ahead. Therefore an adaptive predictor is adopted for the purpose of both predicting the control error of two sampling periods ahead and bringing the robustness to the deadbeat current control system. By adding the adaptive predictor output as an adjustment term to the reference value of half a source voltage period before, settling time is made short in a transient state. On the other hand, in a steady state, THD (total harmonic distortion) of the utility grid side AC source current can be reduced as much as possible, compared to the case that ideal identification of controlled system could be made.

Keywords : Active Power Filter, One-Dimensional Deadbeat Control Scheme, Adaptive Predictor, Robust Control

1. Introduction

Deadbeat control technique is often proposed as an advanced current control method of active power filter (APF)[1],[2], because the superiority of the digital control scheme to analog control one can be effectively utilized by adopting the deadbeat-control strategy. In this paper, the advanced deadbeat control method adopting an adaptive predictor for a three-phase voltage-source APF is introduced.

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The time-optimal response of the APF compensating current is based on a one-dimensional finite time settling algorithm, and it is understood that one sampling period is required for a settling time[3]. To cancel both the delay and the one in DSP control strategy, the desired APF compensating current of two sampling periods ahead has to be predicted.

For the prediction value, the reference value of half a source voltage period before is employed as a main term, because the compensation current is in the form of the periodical wave. However, the prediction in a transient state can't be made only by examining the value before half a period. Furthermore, there is a problem that the mathematical and dynamic circuit model of the controlled APF system is not always perfectly exact in most cases.

So due to parameter uncertainties, control error occurs in deadbeat current control system without robustness [4].

Therefore, the adaptive predictor[5] is adopted for the purpose of both predicting the control error of two sampling periods ahead and bringing the robustness to the deadbeat current control system. And, by adding the adaptive predictor output as an adjustment term to the main reference term, settling time is made short in a transient state. On the other hand, in a steady state, THD of source side utility AC power current can be reduced as much as possible, compared to the case that ideal identification of controlled system could be made.

Finally, the newly proposed shunt voltage source active power filter is applied to the nonlinear load represented by diode bridge rectifiers. It is proven, from an experimental point of view, that the compensating ability of the active power filter with a proposed adaptive predictor is actually extremely high in accuracy.

2. THREE-PHASE VOLTAGE SOURCE ACTIVE POWER FILTER

2.1 System Configuration

Figure.1 shows the schematic configuration of a three-phase voltage-source APF, in addition to a utility AC power side harmonic-producing load and a current control implementation scheme. The harmonic-producing load system consists of a typical full bridge diode rectifier with a serial smoothing reactor or a parallel smoothing capacitor on the DC side. In the proposed control system, both load current and APF compensating current have to be detected by the hall effect type current sensor device.

The APF compensating current is controlled by a deadbeat current controller, such that the current drawn by the APF from the utility is equal to the current harmonics and reactive current required for the nonlinear load. And two sets of adaptive predictor are to be introduced to bring

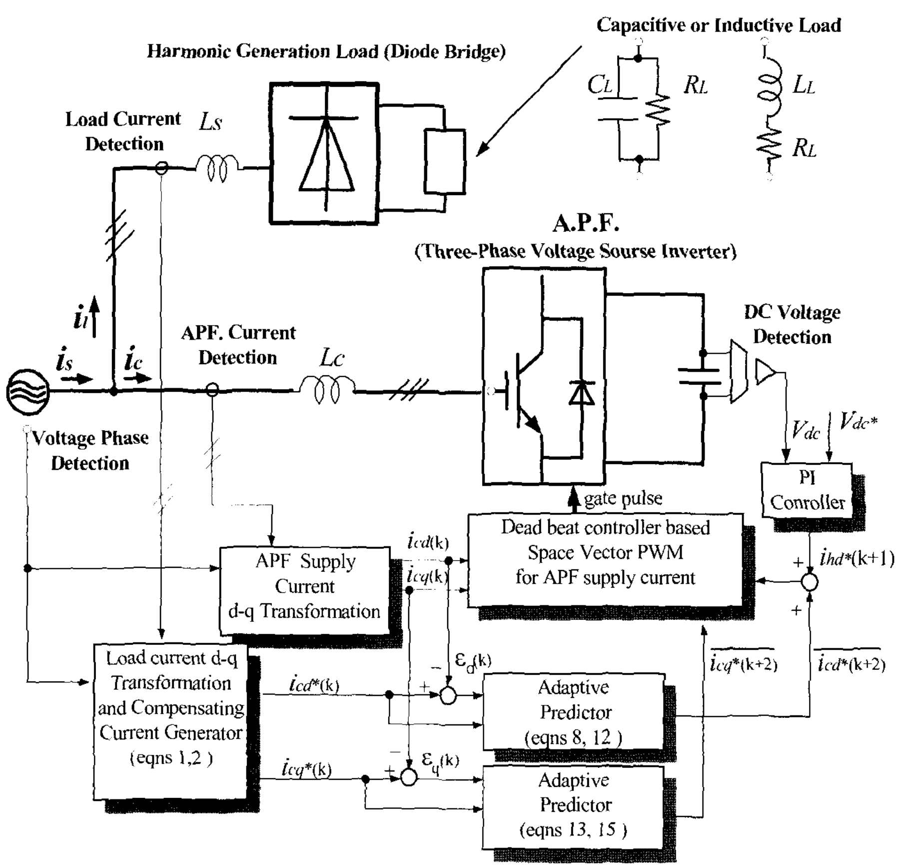


Fig. 1. A schematic three-phase voltage source APF, load system and control system.

robustness to the deadbeat current control system. And more, the PI controller is necessary to make the capacitor voltage on the DC side of APF constant.

2.2 Implementation of Instantaneous Compensating Current Reference

Figure. 2 shows each current vector on d-q synchronous rotating coordinate frame. When source voltage to each phase is expressed as,

$$e_u(t)=\sqrt{\frac{2}{3}}V_l\sin(\omega t+\frac{\pi}{2}),\ e_v(t)=\sqrt{\frac{2}{3}}V_l\sin(\omega t-\frac{\pi}{6})$$
$$e_w(t)=\sqrt{\frac{2}{3}}V_l\sin(\omega t-\frac{5\pi}{6})$$

Where, ω : source voltage angular frequency, V_l : line voltage rms value.

Instantaneous space vector of source voltage, $\boldsymbol{e}$ can be expressed on α−β stationary frame as follows:

$$\boldsymbol{e}=\sqrt{\frac{2}{3}}\ [e_u(t)\ +e_v(t)e^{+j2/3\pi}+e_w(t)e^{-j2/3\pi}]=V_l\,e^{j\omega t}$$

$\boldsymbol{e}$ is also expressed on d-q synchronous rotating frame as,

$$\boldsymbol{e}=\ V_l\,e^{j\omega t}e^{-j\omega t}=V_l$$

d and q components of load current $\boldsymbol{i}_l$ are defined on the basis of detection currents i_{lu} and i_{lv} as follows:

$$\begin{bmatrix} i_{ld} \\ i_{lq} \end{bmatrix}=\sqrt{2}\begin{bmatrix} -\sin(\theta-2\pi/3) & \sin\theta \\ -\cos(\theta-2\pi/3) & \cos\theta \end{bmatrix}\begin{bmatrix} i_{lu} \\ i_{lv} \end{bmatrix} \quad (1)$$

where, θ: the direction of instantaneous space vector of source voltage $\boldsymbol{e}$, θ =ωt

The d and q components of current reference $\boldsymbol{i}_c$* are respectively calculated at a sampling point k as follows:

$$i_{cd}^*(k) = \text{moving average}(\ i_{ld}) - i_{ld}(k)$$

$$i_{cq}^*(k) = -\ i_{lq}(k) \quad (2)$$

In this case, the moving average signal processing value is estimated using the 64 sampling data during the last half of a source voltage period, where the sampling period is 360/128 degree in the electrical angle. By setting such a moving average processing, it is possible to compensate an unbalance state in three-phase voltage source type load currents. On the other hand, by setting the reactive current reference i_{cq}* as indicated in eq.(2), the total power factor of the utility AC power side can be implemented so as to be equal to unity.

Thus current reference vector $\boldsymbol{i}_c$* for APF output $\boldsymbol{i}_c$ is able to be calculated for not each phase but each d-q component, respectively. And the instantaneous space vector of inverter output voltage $\boldsymbol{v}$ is to be directly adjusted for APF as described in the next section.

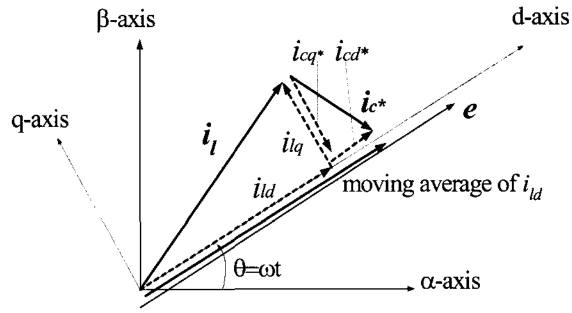


Fig. 2. Instantaneous space vectors i_l and i_c*.

2.3 Current Control in Three-Phase Voltage-Source Active Power Filter

The following current control algorithm for voltage -source three-phase active power filter can attain one -dimensional finite time settling control response. Fig. 3 represents the main schematic circuit of the three-phase voltage-source active power filter using IGBT power modules.

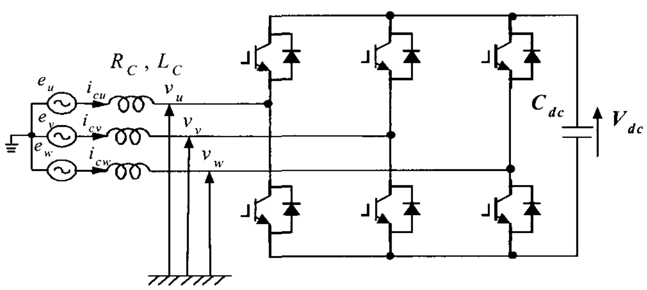


Fig. 3. three-phase active power filter.

The circuit equations in AC power side of this inverter are written as follows:

$$\begin{bmatrix} e_u \\ e_v \\ e_w \end{bmatrix} = \left(R_C + L_C \frac{d}{dt} \right) \begin{bmatrix} i_{cu} \\ i_{cv} \\ i_{cw} \end{bmatrix} + \begin{bmatrix} v_u \\ v_v \\ v_w \end{bmatrix} \tag{3}$$

where, R_C and L_C are the resistive component and inductive component of the AC side filter reactor.

The three-phase variables in eq.(3) are transformed into instantaneous space vectors, $\boldsymbol{e}$: source voltage vector, $\boldsymbol{i}_c$: compensating current vector and $\boldsymbol{v}$: inverter output voltage vector respectively on α–β stationary coordinate frame by using a transformation matrix; $\sqrt{2/3}$ $[1\ e^{+j2/3\pi}\ e^{-j2/3\pi}]$.

$$\sqrt{2/3}\,(e_u + e_v e^{+j2/3\pi} + e_w e^{-j2/3\pi})$$
$$= \sqrt{2/3}\,(R_C + L_C \frac{d}{dt})\,\{(i_{cu} + i_{cv} e^{+j2/3\pi} + i_{cw} e^{-j2/3\pi})\}$$
$$+\sqrt{2/3}\,(v_u + v_v e^{+j2/3\pi} + v_w e^{-j2/3\pi})$$

And circuit vector equation in α–β stationary coor -dinate frame is derived as follows,

$$\boldsymbol{e} = (R_C + L_C \frac{d}{dt})\,\boldsymbol{i}_c + \boldsymbol{v} \tag{4}$$

To solve the differential vector equation (4), $\boldsymbol{e}$ and $\boldsymbol{v}$ are assumed to stand still during a sampling period Ts. In this case, the system equations using the discretized value representation is derived as ,

$$\boldsymbol{i}_c(k+1) = p \cdot \boldsymbol{i}_c(k) + \frac{q}{R_c}\{\,\boldsymbol{e}(k) - \boldsymbol{v}(k)\,\} \tag{5}$$

where, $p = \exp(-R_c / L_c \cdot T_S)$, $q=1-p$ and T_s is the sampling period.

Control target is one-dimensional finite time settling response concerning instantaneous space vector $\boldsymbol{i}_c$. The inverter output voltage $\boldsymbol{v}$ has to be calculated by deadbeat current controller and regulated by space vector modulation, so as to make the phase current vector $\boldsymbol{i}_c$ reach its reference by the end of the following modulation period.

In order to compensate current harmonics with high accuracy by digital deadbeat technique, it is vital to predict the reference value of two sampling periods ahead $\boldsymbol{i}_c$*(k+2) and the control variable of one sampling period ahead $\boldsymbol{i}_c$ (k+1). The prediction of $\boldsymbol{i}_c$ (k+1) can be easily made by using eq.(5). On the other hand, the prediction of $\boldsymbol{i}_c$ *(k+2) is difficult especially in a transient state. When this prediction is made successfully by some means, the inverter output voltage in the next period $\boldsymbol{v}$(k+1) can be calculated at a sampling point k as follows.

$$\boldsymbol{v}(k+1) = \boldsymbol{e}(k+1) - \frac{R_c}{q}\{\,\boldsymbol{i}_c^*(k+2) + i_{hd}(k+1)e^{j\omega(k+2)T_s} - \boldsymbol{i}_c(k+1)\,\} \tag{6}$$

In above equation, i_{hd}(k+1) is the active component for making the capacitor voltage V_{dc} on DC side of APF constant. In order to carry out the deadbeat control scheme easily, it is effective to implement the control algorithm on not α–β stationary coordinate frame but d-q synchronous rotating frame. For the latter frame, equation (6) can be rewritten in the form of sum of the steady state value $\boldsymbol{v}^*$(k+1) and the deviation Δ$\boldsymbol{v}$(k+1) as follows.

$$\boldsymbol{v}(k+1) = \boldsymbol{v}^*(k+1) + \Delta\boldsymbol{v}(k+1)$$
$$= \boldsymbol{e}(k+1) - \frac{R_c}{q} \cdot \{\overline{\boldsymbol{i}_c^*(k+2)} + i_{hd}(k+1)\} \cdot \left(e^{j\theta_0} - p\right)$$
$$- \frac{R_c}{q} \cdot p \cdot \Delta \boldsymbol{i}_c(k+1) \tag{7}$$

where, $\theta_0 = \omega T_s$ and

$$\Delta\,\boldsymbol{i}_c(k+1) = \overline{\boldsymbol{i}_c^*(k+2)} + i_{hd}(k+1) - \overline{\boldsymbol{i}_c(k+1)}$$

By the way, the active and reactive components of $\boldsymbol{i}_c$*(k+2) are respectively predicted by using two sets of adaptive predictor. Two sets of adaptive predictor have another function of bringing robustness to the deadbeat control system about not only the circuit parameter uncertainties of the APF model but also the disturbance of higher harmonic components contained in source voltage.

The details about the adaptive predictor for voltage source active power filters are to be discussed in the next section.

2.4 Prediction of Compensating Current Reference

In many harmonics generation load, each waveform of active and reactive compensating current reference i_{cp}*

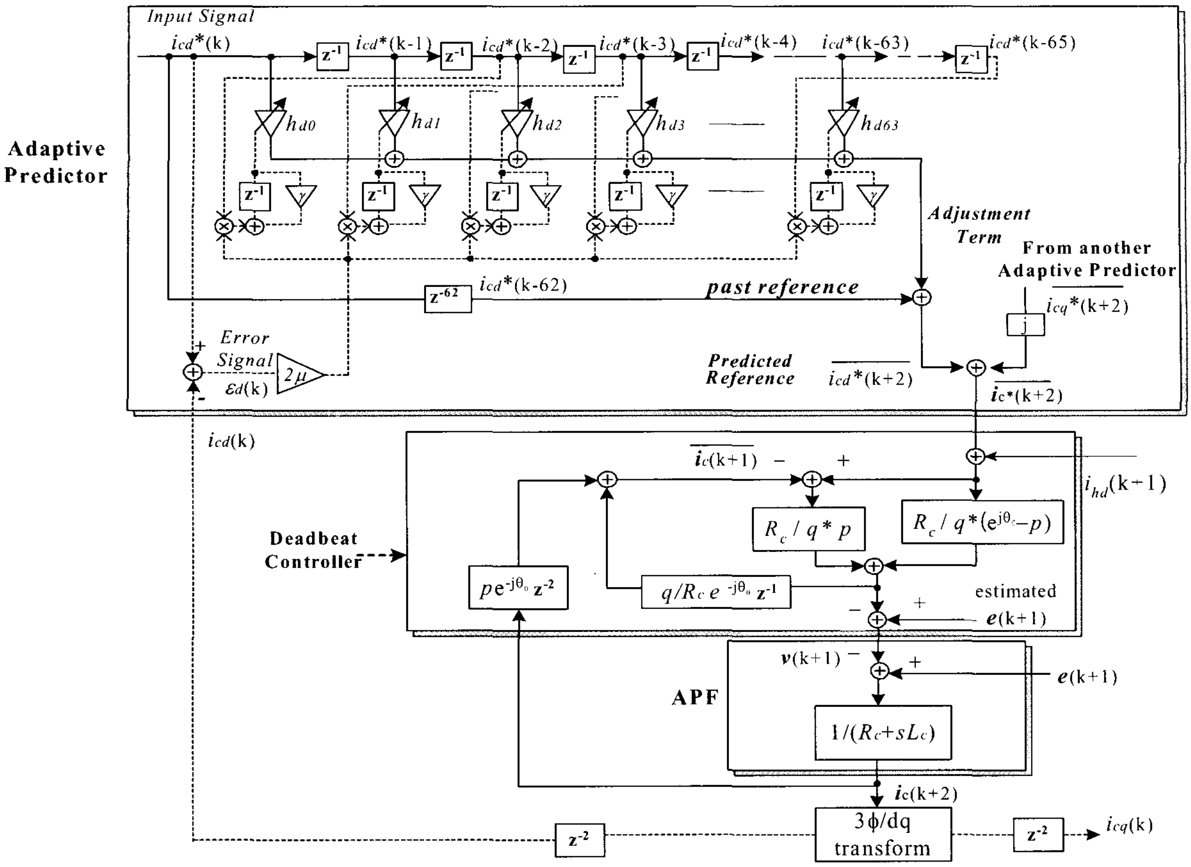


Fig. 4. A deadbeat control scheme with adaptive predictor.

and $i_{cq}*$ is periodical as shown in the latter experimental results. And the period is half a source voltage period (1/2T), where T : source voltage period. The predicted reference value of two sampling periods ahead can be inherently determined by summing the past reference value which has been detected 1/2T before and the adjustment term which is the output of the adaptive predictor.

2.5 Adaptive Predictor

In this APF system, two sets of adaptive predictor are introduced to predict the each component of the control error vector of two sampling periods ahead. Figure.4 shows the proposed deadbeat control scheme. The constitution of the proposed adaptive predictor for the reference of active component is represented in this figure. That for the reference of reactive component has a same constitution. And the accuracy and responsiveness of the deadbeat current control can be improved by adding the output of the each adaptive predictor as an adjusting term to the each component of current reference vector of 1/2T before. In other words, these adaptive predictors act as integrating controllers in order to bring robustness to the deadbeat current control implementation.

2.5.1 Adaptation of Filter Coefficient for Active Component

The adaptation process of digital filter coefficients is illustrated with dash lines in Fig.4. It can be seen that the adaptation of the FIR filter is concretely implemented by using the following Leaky Least Mean Square (LMS) algorithm as follows,

$$\boldsymbol{H}_d(\mathrm{k}+1)=\boldsymbol{H}_d(\mathrm{k})+2\mu_d\varepsilon_d(\mathrm{k})\,\boldsymbol{i}_{cd}*(\mathrm{k}-\mathrm{M}) \qquad (8)$$

where, $\boldsymbol{H}_d$ (k) is the filter coefficient vector: $\boldsymbol{H}_d$(k)

$=[h_{d0}(k), h_{d1}(k), h_{d2}(k), \ldots, h_{d63}(k)]$. As regards the order of the filter, the larger the order is, the higher the control accuracy can be made. It is also necessary to set the order of the filter low so that the operation may be made to end in the sampling period. In the proposed adaptive predictor, the order is set at 64. And data sampling frequency is 128 times as many as source voltage frequency in order to make data vector in the filter cover the complete one cycle of the compensating current reference $\boldsymbol{i}_c^*$. M is the delay number, set at 2 in order to train the adaptive predictor to predict the control error of two sampling periods ahead. $\boldsymbol{i}_{cd}^*(k-M)$ is the data vector for renewing coefficients:

$$\boldsymbol{i}_{cd}^*(k-M)= \boldsymbol{i}_{cd}^*(k-2)= [\, i_{cd}^*(k-2), i_{cd}^*(k-2-1), i_{cd}^*(k-2-2), \ \ldots \ , i_{cd}^*(k-2-63)\,] \qquad (9)$$

$\varepsilon_d(k)$ is the error signal. It is the function of an adaptive FIR filter that the filter coefficient vector is renewed in order to decrease the square expectation value of the error signal as much as possible. The purpose of APF (to make the AC utility grid side current sinusoidal without higher harmonics and reactive component) is thus achieved if the error signal is defined as shown in thus following equation.

$$\varepsilon_d(k) = i_{cd}^*(k) \ - \ i_{cd}(k) \qquad (10)$$

In the adaptive predictor, the distinctive feature is that delayed signal is employed as data vector in eq.(8). And equation (8) is cross-correlation function computed form between an error signal ε_d and a desired value i_{cd}^*. h_{d0} (t) is the cross-correlation value between them when phase difference is set to 5.625° (=360/128*2) in fundamental frequency. Similarly, $h_{d1}(t)$ is the cross-correlation value when phase difference is set to 8.4375° (=360/128*3).

And $h_{d63}(t)$ is the cross-correlation value when phase difference is set to 182.8125° (=360/128*65). For example, $h_{d0}(k)$ can be extended as follows:

$$h_{d0}(k)= 2\mu_d\{\, \varepsilon_d(k)\, i_{cd}^*(k-2) + \gamma\, \varepsilon_d(k-1)\, i_{cd}^*(k-3) + \gamma^2\, \varepsilon_d(k-2)\, i_{cd}^*(k-4) + \gamma^3\, \varepsilon_d(k-3)\, i_{cd}^*(k-5) + \ldots + \gamma^n\, \varepsilon_d(k-n)\, i_{cd}^*(k-n-2) + \gamma^{n+1}\, \varepsilon_d(k-n-1)\, i_{cd}^*(k-n-3) + \ldots \} \qquad (11)$$

If the error signal ε_d (k) contained the same frequency component that exists in reference signal $i_{cd}^*(k)$, filter coefficient $\boldsymbol{H}_d(k)$ would expand. ε_d (k) must become zero as long as the coefficient $\boldsymbol{H}_d(k)$ does not expand. This is the principle that the proposed adaptive predictor can reduce the control error as much as possible. In the actual application of the adaptive predictor, the adjustment of two parameters is vital as follows.

$\gamma = 0.999$ $(=1-2^{-10})$. This constant value γ, which is just less than one, is introduced to prevent the accumulation of error, which often occurs in 16-bit fixed decimal point DSP.

μ_d is the step size. This is fixed to a small enough value to guarantee the stability of the LMS-algorithm. It is necessary to set the value of μ_d from zero within $1/\sum_{j=0}^{63} E\{i_{cd}{}^{*2}[k-j]\}$, in order to settle the error signal ε_d.

This value is experimentally found. The concrete values are 5 % for a the bridge rectifier with a capacitive load, where $1/\sum_{j=0}^{63} E\{i_{cd}{}^{*2}[k-j]\}$ is 100%.

2.5.2 Prediction Equation for Active Component

The prediction equation for the compensating current reference of two sampling periods ahead at sampling point k is illustrated with solid lines in Fig. 4. It can be seen that the prediction equation is as follows:

$$\overline{i_{cd}^*(k+2)} = i_{cd}^*(k-62) + \ \boldsymbol{H}_d(k+1)\, \boldsymbol{i}_{cd}^{*T}(k) \qquad (12)$$

where, $\boldsymbol{i}_{cd}^*(k)=[i_{cd}^*(k), i_{cd}^*(k-1), i_{cd}^*(k-2), \ldots, i_{cd}^*(k-63)]$

The first part $i_{cd}^*(k-62)$ is the reference of half a source voltage period before, which is necessary not to eliminate higher harmonics to be contained in compensating current reference. On the other hand, the second part $\boldsymbol{H}_d$ (k+1) $\boldsymbol{i}_{cd}^{*T}(k)$ is an adjustment term for improving control accuracy and responsiveness. As a meaningful cross-correlation value between i_{cd}^* and ε_d can be obtained in frequency range from 0 Hz to within half of the sampling frequency, as shown by the sampling principle, the cut-off frequency of the proposed adaptive predictor is a half of the sampling frequency; 3.84 kHz in this case.

2.5.3 Adaptation of Filter Coefficient and Prediction Equation for Reactive Component

The similar equations about q-component have to be derived as follows. Renewal of the filter coefficient as

$$\boldsymbol{H}_q(\mathrm{k+1}) = \gamma\, \boldsymbol{H}_q(\mathrm{k}) + 2\mu_q \varepsilon_q(\mathrm{k})\boldsymbol{i}_{cq}{}^*(\mathrm{k-2}) \tag{13}$$

Where, $\boldsymbol{H}_q(\mathrm{k}) = [h_{q0}(\mathrm{k}), h_{q1}(\mathrm{k}), h_{q2}(\mathrm{k}), \; \ldots \;, h_{q63}(\mathrm{k})]$

$$\boldsymbol{i}_{cq}{}^*(\mathrm{k-2}) = [i_{cq}{}^*(\mathrm{k-2}), i_{cq}{}^*(\mathrm{k-3}), i_{cq}{}^*(\mathrm{k-4}), .., i_{cq}{}^*(\mathrm{k-65})\,]$$

The error signal must be defined as follows, in order to achieve power factor correction (PFC):

$$\varepsilon_q(\mathrm{k}) \;=\; i_{cq}{}^*(\mathrm{k}) - i_{cq}(\mathrm{k}) \tag{14}$$

μ_q : step size, μ_q is set at 10% , where

$$1/\sum_{j=0}^{63} E\{i_{cq}{}^{*2}[k-j]\} \text{ is } 100\% .$$

The prediction equation for reactive component is as follows:

$$\overline{i_{cq}{}^*(\mathrm{k+2})} = i_{cq}{}^*(\mathrm{k-62}) + \boldsymbol{H}_q(\mathrm{k+1})\, \boldsymbol{i}_{cq}{}^{*\mathrm{T}}(\mathrm{k}) \tag{15}$$

3. Experimental Results and Their Discussions

3.1 Experimental Condition

In this section, the current control performance of the proposed three-phase voltage source APF is verified from an experimental point of view.

Experimental conditions are described as follows; the line voltage 50V in RMS value, the source frequency is 60Hz, the three-phase sinusoidal voltages are balanced.

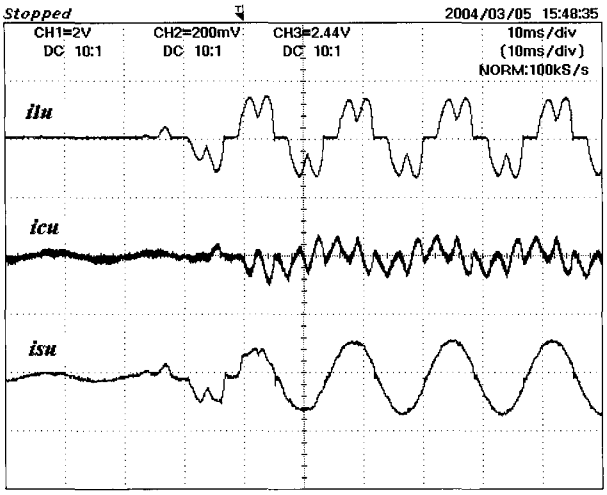
The circuit constants designed for a filter reactor is given as inductance L_c=4mH and internal resistance R_c=0.1Ω. The sampling period Ts is set to 0.130 ms. Circuit constant on the DC side of the APF is designed for capacitance C_{dc}=3300μF.

On the other hand, the load constants of the diode bridge rectifier are R_L=27.8Ω , the DC smoothing capacitor C_L=3300μF. In addition, series inductance L_s is necessary to suppress the expansion of higher harmonic currents including in load current. The series inductance is designed for L_s = 2mH.

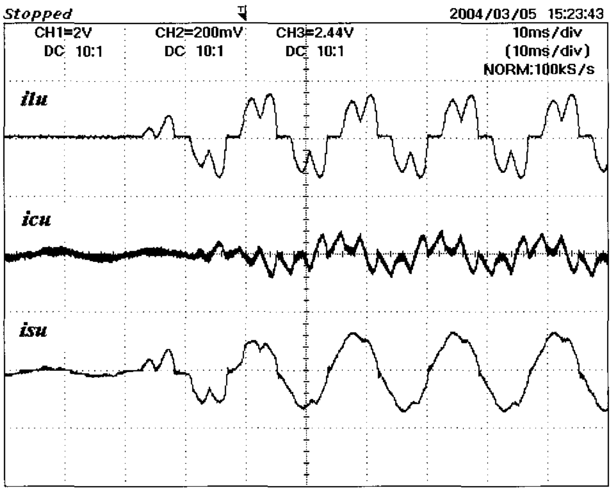
3.2 Waveforms of the Three-phase Variables

The overall operational waveforms of the experimental APF system are shown in Fig. 5. The function of the APF is to make the source current sinusoidal without distortion, by adding the appropriate compensation current to the distorted load current. Fig. 5(a) shows the result as the adaptive predictor is operating. Fig. 5(b) shows the result when the adaptive predictor is stopped by setting both ε_d and ε_q at zero. Fig. 5 shows the transient response to a step on load from 0% to 100 %. Although it is common to the step in which the reference value is calculated by using eqn.(1) and (2), it is different to predict the reference of two sampling periods ahead. When the adaptive predictor is not operating, the past reference value (detected 1/2T before) is taken as the prediction value of two sampling periods ahead.

When, Figs. 5(a) and (b) are compared, it can be seen that the THD of the source current in Fig. 5.(a) can be reduced owing to the adaptive predictor. The reason why the THD of a source current in Fig. 5(b) cannot be reduced as well as in Fig. 5.(a) is explained as follows. Although the filter inductance decreases with harmonic order and its internal resistance increases with harmonic order, the consideration about the parameter changes are never made in the control algorithm. On the other hand, in Fig. 5.(a), the robustness about the parameter uncertainties is brought by the two sets of the adaptive predictor. The harmonic spectra of the load and source currents in Fig. 5(a) are shown in Fig. 6 as a percentage of the fundamental current component. The THD is 37.6% in the load current and it is 4.0% in the source current after the harmonic compensation which is a significant improvement. The results based on THD measurement in the source current after harmonic compensation by APF show that the proposed method is effective for improving performance of the deadbeat current control. In the following, the effectiveness of the two sets of the adaptive predictor is confirmed respectively.



(a) adaptive predictor is operating



(b) adaptive predictor is not operating

Fig. 5. Transient response of u-phase current to step on load from 0% to 100 %.

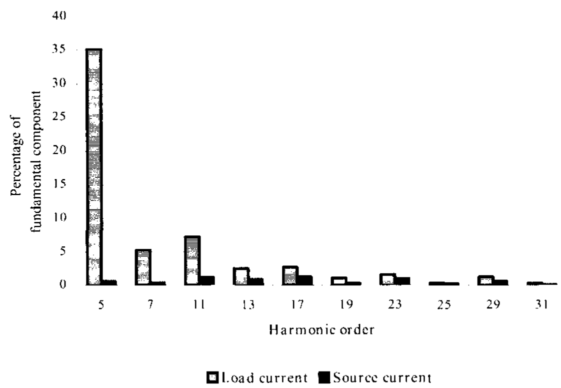


Fig. 6. Harmonic spectra of the load and source current.

3.3 Compensation of Higher Harmonics

Figure. 7 shows the transient response of active component of a source current i_{sd}., when the adaptive predictor is operating. In addition to the load current i_{ld} and compensating current i_{cd} are illustrated respectively in the same figure. It's proven that due to the adaptive predictor, the source current is almost fixedly kept after the settling.

In Fig. 8, it is proven that the term of $\boldsymbol{H}_d$ (k+1) $\boldsymbol{i}_{cd}{}^{*T}$(k), which is used as an adjustment term of the current reference in Fig. 5(a), is almost equal to the control error in Fig. 5(b). That is to say, the high compensating performance in Fig. 5(a) can be attained by adding the adjustment term as large as the control error to the compensating current reference.

3.4 Compensation of Reactive Component

Figure. 9 shows the transient response of reactive component of a source current i_{sq}. In addition, the load current i_{lq} and the compensating current i_{cq} are illustrated, respectively. It's proven that due to the adaptive predictor, the reactive component of a source current i_{sq} is almost kept zero in both a steady and a transient state, as the adaptive predictor brings robustness to the deadbeat controller.

3.5 Compensation to a Diode Rectifier with an Inductive load

The proposed APF can be applied to compensate current harmonics produced by a diode bridge rectifier with an inductive load. The overall operational waveforms of the experimental APF system are shown in Figure 10.

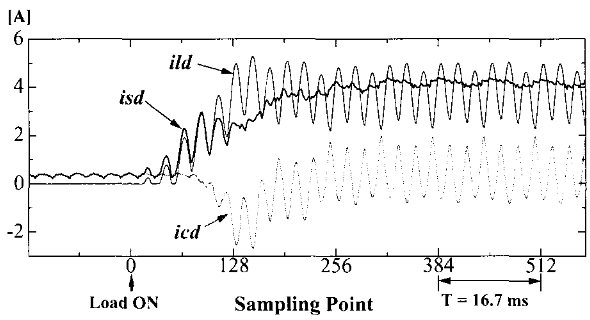


Fig. 7. Transient response of active component. i_{sd}(source current)= i_{ld} (load)+i_{cd}(compensating).

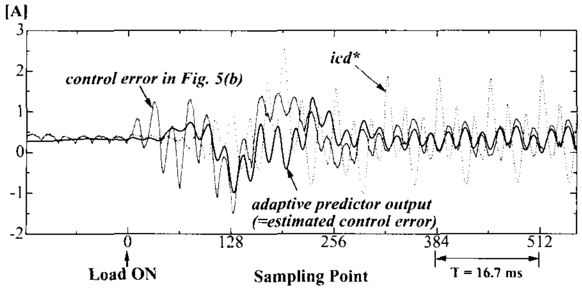


Fig. 8. Control error and adaptive predictor output; $H_d(k+1)i_{cd}^{*T}(k)$.

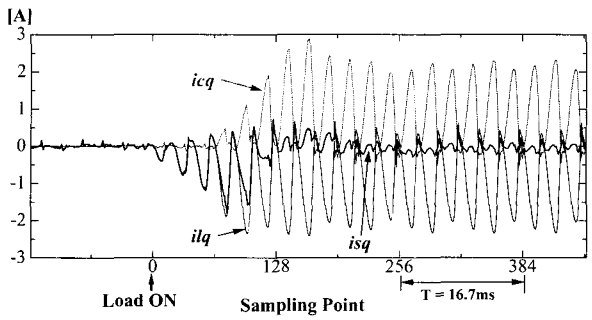


Fig. 9. Transient response of reactive component i_{sq} (source current)= i_{lq} (load)+i_{cq} (compensating).

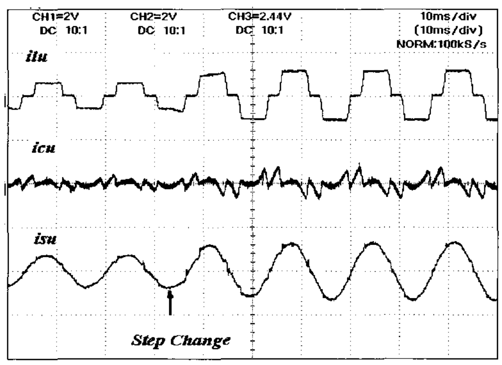


Fig. 10. Transient response of u-phase current to a step on the load resistance R_L: from 55.6 to 27.8Ω in an inductive load (L_L=90mH), (step size is set at μ_d =4%, μ_q =40%).

4. Conclusions

In this paper, the authors have actually derived the effective method by which it is possible to apply three-phase voltage-source PWM inverter with a deadbeat controller to APF.

In this APF, the compensation of current harmonics is to be attained on the basis of setting the control scheme for the purpose of compensating the instantaneous reactive current component in load current as well as reducing the ripple factor in the instantaneous active current component.

The adaptive predictor was originally applied to predict the compensating current reference and to bring robustness to a deadbeat controller. And the 64 load current data during the last half of a voltage source period were used in the adaptive filter.

The feasible validity of the proposed current control algorithm developed for deadbeat based active power filter was substantially confirmed from an experimental point of view.

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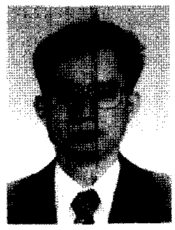

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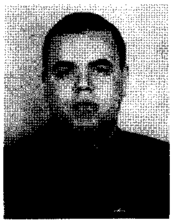

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