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전력설비의 안정한 운용을 위한 3상 능동전력필터의 강인한 내부모델제어

Robust Internal Model Control of Three-Phase Active Power Filter for Stable Operation in Electric Power Equipment

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Abstract - A new simple control method for active power filter, which can realize the complete compensation of harmonics is proposed. In the proposed scheme, a model-based digital current control strategy is presented. The proposed control system is designed and implemented in a form referred to as internal model control structure. This method provides a convenient way for parameterizing the controller in term of the nominal system model, including time-delays. As a result, the resulting controller parameters are directly set based on the power circuit parameters, which make tuning of the controllers straightforward task. In the proposed control algorithm, overshoots and oscillations due to the computation time delay is prevented by explicit incorporating of the delay in the controller transfer function. In addition, a new compensating current reference generator employing resonance model implemented by a DSP(Digital Signal Processor) is introduced. Resonance model has an infinite gain at resonant frequency, and it exhibits a band-pass filter. Consequently, the difference between the instantaneous load current and the output of this model is the current reference signal for the harmonic compensation.

Key Words : Active power filter, Internal model control, Deadbeat response, Resonance model

1. Introduction

In recent years, power converter that can result in harmonics have been widely used in industrial applications. These harmonics can be occurred deterioration in reliability and stability of computers, factory automation systems, electronic equipments and communication systems. Therefore, active power filters have been widely investigated for the compensating of harmonics in electric power system[1]. The control of active power filter is composed of two parts; the calculation of harmonic compensating current reference and the control of mains or active power filter current. In active power filters, the compensating currents with very complicated waveform have to be generated. To achieve this, a high speed and accurate current control capability is required. One of current control methods is the hysteresis control, in which a hysteresis is used to determine the permitted deviation of the actual phase current from the reference value. This method has a very fast response and is easy to implement. However, It has the drawback of high and non-constant switching frequency[2]. Another is the PI synchronous reference frame control. In this method, the PI controller output is taken as the modulation signal, and a sine-triangular PWM(Pulse Width Modulation) scheme is applied. This method, though simple and giving fixed switching frequency, has difficulty in adjusting the controller parameters[3]. The third method is the predictive control using a model of the converter-grid system and past samples of the currents. The principle of this method is the error between the reference and compensating current at the next sampling instant can be predicted. Finally, a deadbeat control is as known guarantees the best possible dynamic performances among the fully digital solutions[4]. In this method, the switching pulse width is adopted so that active power filter current is been exactly equal to its reference at the next sampling instant. An important advantage of deadbeat control is that it does not required line voltage measurement in order to generate the current reference. On the other hand, the inherent delay due to the calculation time is indeed a serious drawback in active power filter application.

In this paper, a new simple control method for active power, which can realize the complete compensation of harmonics is proposed. The current control system consists of two parts; an internal model controller and a

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modeling error feedback loop which incorporates an explicit representation of the active power filter system nominal model[5]. The output of the nominal model is determined and is subtracted from the actual output, to yield a feedback signal. The difference the feedback signal and the harmonic compensating current reference is supplied to the internal model controller, which determines the control signal. The main features of this method are as follows[6]. The control system has a wide bandwidth while being very stable. This method provides an essentially overshoot free reference-to-output response with a minimum possible rise time. This internal model structure presents a convenient way to parameterize the controller in terms of the system model and to monitor, analyze and compensate for the modeling errors. The control system is implemented without a dq coordinate transformation. In active power filter system, the currents are not single-frequency sinusoids, therefore a dq transformation does not provide a particular advantage. In the proposed scheme, a new compensating current reference generator employing resonance model implemented by a DSP is introduced[7, 8]. The resonance model has an infinite gain at resonant frequency. The gain of resonance model is small except at the resonant frequency. Thus it does not cause the stability problem. In this case, the resonant frequency is set to the fundamental frequency of the main voltage. The detected load current is applied to the resonance model without any coordinate transformation. The output signal of it is used as the fundamental component of the source current and is adjusted automatically to a value, which makes the fundamental frequency component of the main voltage. The difference between the instantaneous load current and this fundamental component is the current reference signal for the harmonic compensation. Comparing with the convenient calculation method of harmonic compensating current reference, this method does not require the dq coordinate transformation. Therefore, it can be applied also to single-phase active power filter[9-12]. From the results of the experiment, there are demonstrated that the complete compensation of the harmonic components in the source current can be achieved.

2. Current Control System

2.1 System Modeling

Fig. 1 shows a configuration of the proposed active power filter system. The voltage equation of actual plant model using space vector to represent three-phase variables in Fig. 1 is given by

$$L_f \frac{di_c}{dt} + R_f i_c = v_c - v_s \tag{1}$$

where, i_c and v_c are output current and voltage vector of active power filter, and v_s is source voltage vector, respectively. Parameters R_f and L_f represent respectively the equivalent per-phase resistance and inductance of the path between the active power filter and the point of common coupling, including the wiring, the filter inductor, and the transformer.

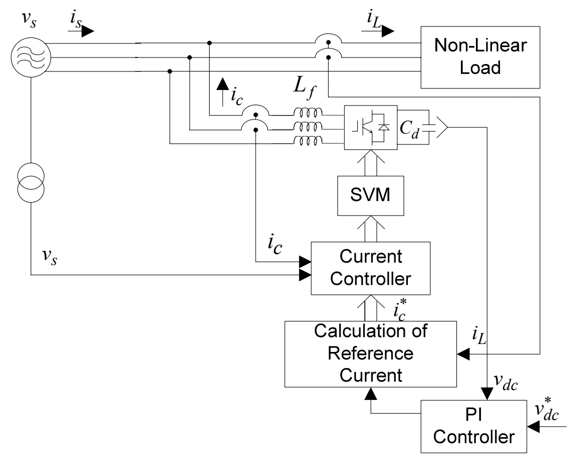


그림 1 제안된 시스템 구성도

Fig. 1 Configuration of the proposed system

The voltage equation of nominal plant model is defined by

$$\tilde{L}_f \frac{di_c}{dt} + \tilde{R}_f i_c = v_c - v_s \tag{2}$$

where, parameters $\tilde{L}_f$ and specially $\tilde{R}_f$ are known with a limited accuracy, and are dependent on the frequency and operating conditions in practice. Therefore, a distinction is made between the actual parameter values and the nominal ones, by using "~" to denote the nominal values.

In the s-domain, Eq. (1) can be written as,

$$I_c(s) = G_p(s)[V_c(s) - V_s(s)] \tag{3}$$

$$G_p(s) = \frac{1}{L_f s + R_f} \tag{4}$$

The transfer function of nominal system is the same as the transfer function of actual system $G_p(s)$, except that the actual parameter values are replaced by the nominal ones.

$$\tilde{G}_p(s) = \frac{1}{\tilde{L}_f s + \tilde{R}_f} \tag{5}$$

As long as the control signal changes only at sampling times and stays constant between consecutive sampling points in digitally controlled systems, the zero-order-hold equivalent form of the transfer function $G_p(s)$ is given by

$$G_p(z) = \frac{b}{z-a} \tag{6}$$

where, $a = e^{-R_f T_s / L_f}$, $b = \frac{1}{R_f}(1-a)$ and T_s is the sampling period.

Similarly, the z-domain transfer function of nominal system is represented by

$$\widetilde{G}_p(z) = \frac{\tilde{b}}{z-\tilde{a}} \tag{7}$$

where, $\tilde{a} = e^{-\widetilde{R}_f T_s / \widetilde{L}_f}$, $\tilde{b} = \frac{1}{\widetilde{R}_f}(1-\tilde{a})$

The characteristics of digitally controlled system depend on the sampling time of the digital controller and the time delay caused by the computational time of the microprocessor. A basic condition to improve the performance is to make the time delay and a sampling period shorter as far as possible. Hence, the essential requirements for digital control methods are compensation of the time delay and a simple control algorithm. To realize the above mentioned requirements, the calculation time delay is modeled to a unit lag transfer function z^{-1} as if it is a part of the system model.

2.2 Internal Model Control

The block diagram of the proposed current control system is shown in Fig. 2. It consists of two parts; an internal model controller and a modeling error feedback loop which incorporates an explicit representation of the active power filter system nominal model. The output of the nominal plant model $I_{cn}(z)$ is determined and is subtracted from the actual output $I_c(z)$, to yield a feedback signal $I_{cf}(z)$. The difference the feedback signal and the harmonic compensating current reference $I_c^*(z)$ is supplied to the internal model controller, which determines the control signal $V_c(z)$. If the actual and nominal model is exact, then the actual output $I_c(z)$ and the nominal output $I_{cn}(z)$ are the same and the feedback signal $I_{cf}(z)$ is zero. Therefore, the current control system operates open-loop system, when there is no parameter uncertainty.

The reference-to-output transfer function including the calculation time delay is given from Fig. 2

$$\left[\frac{I_c(z)}{I_c^*(z)}\right]_{V_s(z)=0} = \frac{G_I(z) z^{-1} G_p(z)}{1 + G_I(z) z^{-1} [G_p(z) - \widetilde{G}_p(z)]} \tag{8}$$

Based on assumption that the actual and nominal model is exact, that is $G_p(z) = \widetilde{G}_p(z)$, the actual output and the nominal output are the same and $I_{cf}(z) = 0$. In order to achieve zero steady state error, Eq. (8) must be

$$\left[\frac{I_c(z)}{I_c^*(z)}\right]_{V_s(z)=0} = G_I(z) z^{-1} \widetilde{G}_p(z) = 1 \tag{9}$$

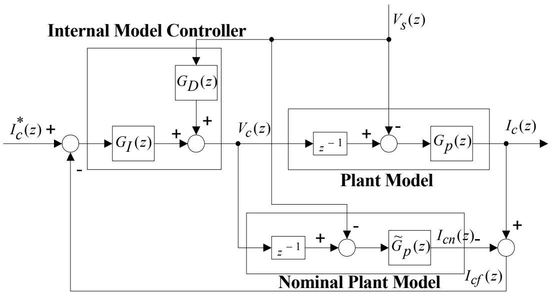


그림 2 제안된 전류제어시스템의 구성도

Fig. 2 Block diagram of the proposed current control system

The transfer function $G_I(z)$ is derived as following from the ideal condition in Eq. (9)

$$G_I(z) = \frac{1}{z^{-1}\widetilde{G}_p(z)} = \frac{z(z-\tilde{a})}{\tilde{b}} \tag{10}$$

However, the above transfer function $G_I(z)$ is not proper because its numerator has a greater order than the respective denominator. That is to say, this internal model controller is impossible to realize considering calculation time delay. Therefore, the transfer function $G_I(z)$ is adopted to a second order deadbeat reference-to-output response which means that its response reaches the reference in two sampling time as following

$$G_I(z) = \frac{1}{z^2}\frac{1}{z^{-1}\widetilde{G}_p(z)} = \frac{(z-\tilde{a})}{\tilde{b}z} \tag{11}$$

Under this condition, the reference-to-output transfer function including the calculation time delay in Eq. (9) is obtained as

$$\left[\frac{I_c(z)}{I_c^*(z)}\right]_{V_s(z)=0} = \frac{1}{z^2} \tag{12}$$

That is, the average current of active power filter is the exact replica of the reference current with a time-lag of two sampling intervals. The frequency response of the reference-to-output transfer function is

$$\left[\frac{I_c(e^{j\omega T_s})}{I_c^*(e^{j\omega T_s})}\right]_{V_s(e^{j\omega T_s})=0} = e^{-2j\omega T_s} \tag{13}$$

which has unity gain and a phase lag of $2\omega T_s$. The phase lag in correspondence of the two-sampling-period delay produces appreciable phase error between the reference and actual currents especially at harmonic frequencies. To overcome this problem, an equal and opposite phase shift is added to the harmonic current reference in the compensating current reference generator.

Similarly, the disturbance-to-output transfer function including the calculation time delay is given from Fig. 2

$$\left[\frac{I_c(z)}{V_s(z)}\right]_{I_c^*(z)=0} = \frac{G_D(z)z^{-1}G_p(z) - G_p(z)}{1+G_D(z)z^{-1}[G_p(z)-\widetilde{G}_p(z)]} \tag{14}$$

The transfer function $G_D(z)$ is chosen such that the source voltage, which intervenes as a disturbance in the control loop, is eliminated by feedforward under the same condition of Eq. (9).

$$\left[\frac{I_c(z)}{V_s(z)}\right]_{I_c^*(z)=0} = \widetilde{G}_p(z)[G_D(z)z^{-1}-1]=0 \tag{15}$$

The following transfer function $G_D(z)$ is sufficient from the ideal condition in Eq. (15)

$$G_D(z) = z \tag{16}$$

However, this transfer function $G_D(z)$ is not also proper in case of $G_I(z)$. Hence, the value of $V_s(z)$ in the next sampling instant is predicted from its past and present values by extrapolation. A linear extrapolation used for such a purpose yields

$$v_s(k+1) \approx 2v_s(k) - v_s(k-1) \tag{17}$$

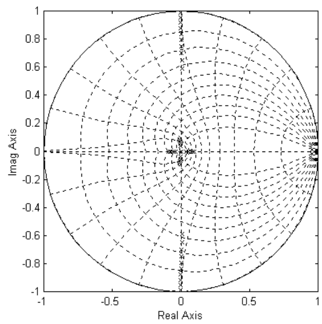
The feasible form of transfer function $G_D(z)$ is given by from Eq. (17)

$$G_D(z) = 2 - z^{-1} \tag{18}$$

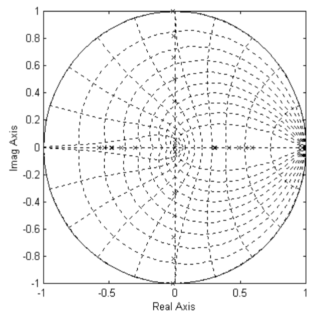
Consequently, the internal model controller has the following form

$$V_s(z) = G_I(z)[I_c^*(z) - I_{cf}(z)] + G_D(z)V_s(z) \tag{19}$$

To evaluate the robustness of the internal model controller against parameter variations, we investigate the root-loci of its closed-loop transfer function by varying R_f and L_f. Fig. 3 shows that R_f and L_f vary within ±50[%], respectively. From Fig. 3(a), we observe that internal model controller is a little oscillated as R_f is decreased, and the response of internal model controller is late as R_f is increased. On the contrary, we observe that internal model controller is a little oscillated as L_f is increased, and the response of internal model controller is late as L_f is decreased. From a stability point of view, internal model controller is always stable, regardless of parameter variations, since its roots are located inside unit circle in the z plane. This result demonstrates that the internal model controller with a second-order deadbeat response is robust against parameter variations.



(a) When R_f vary within ±50[%]



(b) When L_f vary within ±50[%]

그림 3 파라미터 변동에 대한 내부모델제어기의 근궤적

Fig. 3 Root-loci of internal model controller with respect to parameter variation

3. Current Reference Generator

In active power filters, the reference signal of the current components for the harmonic compensation is required. In the proposed scheme, a new compensating current reference generator employing resonance model implemented by a DSP is introduced. Fig. 4 shows an equivalent circuit of resonance model.

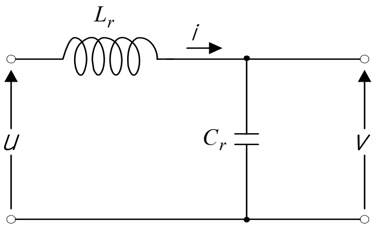


그림 4 공진모델의 등가회로

Fig. 4 Equivalent circuit of resonance model

We assume $\omega_r = 1/L_r = 1/C_r$ for the convenience in the implementation of the calculation. The state equation of this circuit is given by in continuous time domain

$$\begin{bmatrix} \dot{v} \\ \dot{i} \end{bmatrix} = \begin{bmatrix} 0 & \omega_r \\ -\omega_r & 0 \end{bmatrix} \begin{bmatrix} v \\ i \end{bmatrix} + \begin{bmatrix} 0 \\ \omega_r \end{bmatrix} u \tag{20}$$

The output equation is given by

$$y = k_r \begin{bmatrix} -\sin\theta_r & \cos\theta_r \end{bmatrix} \begin{bmatrix} v \\ i \end{bmatrix} \tag{21}$$

From Eq. (20) and (21), the s-domain transfer function of resonance model is given by

$$G_r(s) = \frac{k_r(\cos\theta_r \omega_r s - \sin\theta_r \omega_r^2)}{s^2 + \omega_r^2} \tag{22}$$

where, k_r and ω_r are a control gain and resonant frequency of resonance model, respectively. θ_r is a parameter for the adjustment of the phase angle of the resonance model at the resonant frequency to compensate a phase error. Note that the value of $G_r(s)$ goes to infinity when $s = j\omega_r$. In this case, the resonant frequency is set to the fundamental frequency of the main voltage. Thus it exhibits a 60[Hz] band-pass filter. Although the proposed resonance model acts as a LC filter, it has the following properties, which can eliminate the shortcomings of the convenient LC filters. The parameters of the resonance model, such as the resonant frequency, quality factor and characteristic impedance can be changed even during the operation condition. There are no physical restrictions in the selection of the circuit parameters of the resonance model component, such as the inductance, capacitance and resistance.

From Eq. (22), we can obtain the state equation in discrete time domain as follow

$$\begin{bmatrix} v(k+1) \\ i(k+1) \end{bmatrix} = \begin{bmatrix} \cos\omega_r T_s & \sin\omega_r T_s \\ -\sin\omega_r T_s & \cos\omega_r T_s \end{bmatrix} \begin{bmatrix} v(k) \\ i(k) \end{bmatrix} + \begin{bmatrix} 1-\cos\omega_r T_s \\ \sin\omega_r T_s \end{bmatrix} u(k) \tag{23}$$

where, $v(k)$, $i(k)$ and $u(k)$ are k^{th} sampled values of v, i and u, respectively. Using Eq. (23), the state variables at the $(k+1)^{th}$ sampling instant can be calculated consecutively using the state variables and input at the k^{th} sampling instant. In this way, resonance model represented by Eq. (22) can be implemented by a DSP.

The detected load current is applied to resonance model without any coordinate transformation. The output of this model is used as the fundamental component of source current, and the difference between the instantaneous load current and this fundamental component is the current reference signal of harmonic compensation.

For the investigation of the effect of k_r, Bode plots of resonance model when $\theta_r = 0$ are shown in Fig. 5. From this figure, the control gain k_r is the bandwidth of resonance model. If k_r is large, the current reference generator is achieved a faster transient response, but the resulting system is unstable. Therefore, the selection of k_r is a trade-off between transient response and stability of current reference generator. To meet this requirement, k_r is set to 0.4 in simulation and experiment.

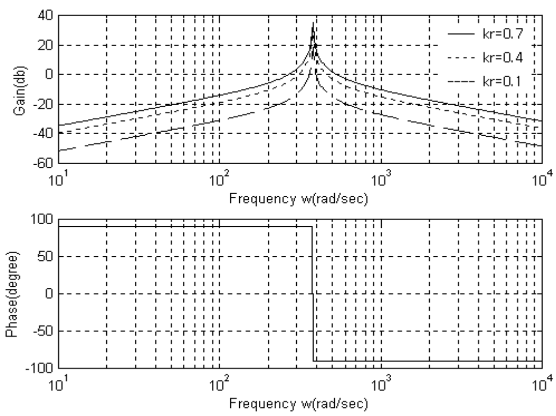


그림 5 $\theta_r = 0$ 에서 공진모델의 보드선도

Fig. 5 Bode plots of resonance model when $\theta_r = 0$

For the convenience in the investigation of θ_r, we will consider resonance model with a finite quality factor. When the quality factor is Q, the transfer function of resonance model is given by

$$G_r'(s) = \frac{k_r(\cos\theta_r \omega_r s - \sin\theta_r \omega_r^2)}{s^2 + \omega_r s / Q + \omega_r^2} \tag{24}$$

Substituting $s = j\omega_r$ in Eq. (24), we can obtain the characteristics of resonance model at the resonant frequency.

$$G_r'(j\omega_r) = k_r Q(\cos\theta_r + j\sin\theta_r) \quad (25)$$

From this result, we can see that the phase angle of the transfer function given by Eq. (24) is θ_r at the resonant frequency ω_r. Resonance model whose transfer function in Eq. (22) is a special case in which the quality factor Q is infinity.

The phase lag of the reference-to-output transfer function in Eq. (13) produces appreciable phase error between the reference and actual currents especially at harmonic frequencies. Thus the phase angle of resonance model is set to an equal and opposite phase shift, the phase error of the reference-to-output transfer function is compensated automatically.

4. Experimental Result

To verify the effectiveness of the proposed system, experiment was performed by 5[kVA] laboratory test system. The system parameters used for experiment are listed in Table 1. The control algorithm is implemented by a DSP(TMS320C32). The calculation interval is 92.6 [μs]. In this experiment, a diode rectifier is used as the harmonic current source.

표 1 시스템 파라미터

Table 1 System parameters

v_s	220[V], 60[Hz]
$\widetilde{L}_f$	2[mH]
$\widetilde{R}_f$	1.7[Ω]
C_d	2200[μF]
R_{load}	30[Ω]
f_{sw}	5.4[kHz]
v_{dc}^*	700[V]

The steady state response of the proposed system is shown in Fig. 6. Fig. 6(a), 6(b) and 6(c) are the experimental results of load current, source current and compensating current, respectively. Fig. 7 shows the frequency spectra of load and source current from experiment results in Fig. 6. The lower order harmonic component of source current has been reduced significantly, and the THD(Total Harmonic Distortion) has been reduced from 23.37[%] to 4.71[%] with the proposed control system. From these figures, the proposed system can be achieved the complete compensation of the harmonic components in the source current. Fig. 8 shows the transient response of the proposed system when there is 50[%] step change of load resistance. From this figure, the proposed control system can be achieved a faster transient response.

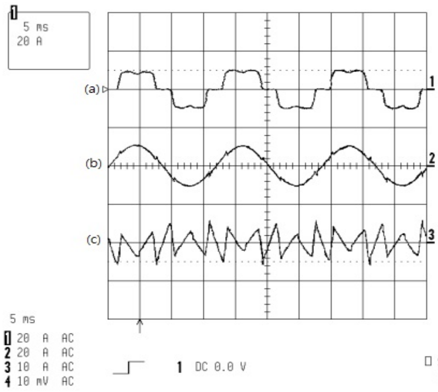
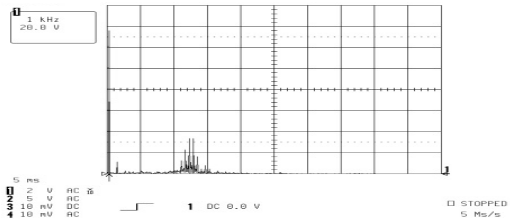
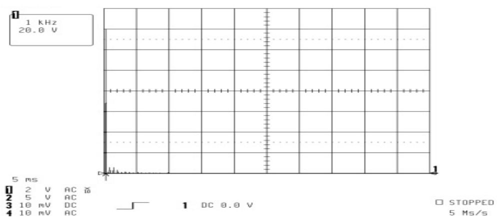


그림 6 정상상태 실험결과: (a) 부하전류; (b) 전원전류; (c) 보상전류

Fig. 6 Experimental results of the steady state: (a) load current; (b) source current; (c) compensating current



(a) load current



(b) source current

그림 7 그림 6의 THD

Fig. 7 Frequency spectra of Fig. 6

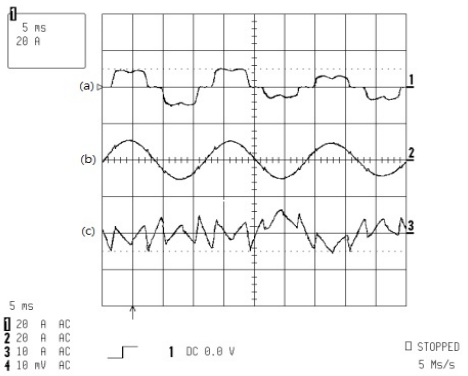


그림 8 과도상태 실험결과: (a) 부하전류; (b) 전원전류; (c) 보상전류

Fig. 8 Experimental results of the transient response : (a) load current; (b) source current; (c) compensating current

5. Conclusion

In this paper, a new simple control method for active power, which can realize the complete compensation of harmonics is proposed. The current control system consists of two parts; an internal model controller and a modeling error feedback loop which incorporates an explicit representation of the active power filter system nominal model. The output of the nominal model is determined and is subtracted from the actual output, to yield a feedback signal. The difference of the feedback signal and the harmonic compensating current reference is supplied to the internal model controller, which determines the control signal. In the proposed scheme, a new compensating current reference generator employing resonance model implemented by a DSP is introduced. The resonance model has an infinite gain at resonant frequency, and it exhibits the band-pass filter. In this case, the resonant frequency is set to the fundamental frequency of the main voltage. The detected load current is applied to the resonance model without any coordinate transformation. The output signal of it is used as the fundamental component of the source current and is adjusted automatically to a value, which makes the fundamental frequency component of the main voltage. The difference between the instantaneous load current and this fundamental component is the current reference signal for the harmonic compensation. From the results of the simulation and experiment, there are demonstrated that the complete compensation of the harmonic components in the source current can be achieved.

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