Periodic errors elimination in CVCF PWM DC/AC converter systems: Repetitive control approach

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Abstract: A plug-in digital repetitive learning (RC) controller is proposed to eliminate periodic tracking errors in constant-voltage constant-frequency (CVCF) pulse-width modulated (PWM) DC/AC converter systems. The design of the RC controller is systematically developed and the stability analysis of the overall system is discussed. The periodic errors are forced toward zero asymptotically and the total harmonics distortion (THD) of the output voltage is substantially reduced under parameter uncertainties and load disturbances. Simulation and experimental results are provided to illustrate the validity of the proposed scheme.

1 Introduction

Constant-voltage constant-frequency pulse-width modulated (CVCF PWM) DC/AC converters are widely employed in various AC power-conditioning systems, such as automatic voltage regulators and uninterruptible power supply systems. Output voltage THD is one important index to evaluate the performance of the converters, associated with communication interference, excessive heating in capacitors and transformers *etc.* Nonlinear loads, causing periodic distortion, are major sources of THD in AC power systems. To minimise THD, several high precision control schemes are proposed for the CVCF PWM DC/AC converters. A deadbeat (or OSAP) controller has been proposed [1–3]. Sliding mode controller (SMC) [4, 5] and hysteresis controller (HC) [6] can overcome parameter uncertainties and load disturbance. However, the deadbeat control is highly dependent on the accuracy of the parameters; random switching pattern of SMC or HC will impose excessive stress on power devices and cause difficulty in lowpass filtering.

Repetitive learning control (RC) law is closely similar to iterative learning control (ILC) [7–11]. Although an ILC system only updates the control input once each cycle and resets the plant at the beginning of each iteration, RC continuously adjusts its control input and needs no reset. The RC method [12], based on the internal model (IM) principle [13], has proposed [14, 15] to achieve high accuracy in the presence of uncertainties for servomechanism. Applications of RC [16] include robots [17], disc drives [18], steel casting process [19], satellites [20]. Without a complete design method and stability analysis of RC system, it has been applied to DC/AC converters [21, 22] with preliminary results.

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In this paper, the design of discrete time RC controller is presented systematically. A plug-in RC controller is proposed and developed for the OSAP controlled CVCF PWM DC/AC converters. The stability of overall system is discussed. To show the validity of proposed method, simulation and experimental results are illustrated.

2 Plug-in discrete time repetitive controller

Fig. 1*a* shows a periodic signal generator. Consider the discrete time RC system shown in Fig. 1*b*, where $y_d(z)$ is the reference input signal, $y(z)$ is the output signal, $d(z)$ is the disturbance signal, $e(z)$ is the tracking error signal, $G_s(z)$ is the transfer function of the plant, $G_r(z)$ is the repetitive signal generator; plug-in RC controller $G_{rc}(z)$ is the feedforward compensator, and $G_c(z)$ is the conventional feedback controller. $G_c(z)$ is chosen so that the following closed-loop transfer function is asymptotically stable.

$$H(z) = \frac{G_c(z)G_s(z)}{1 + G_c(z)G_s(z)} = \frac{z^{-d}B(z^{-1})}{A(z^{-1})} = \frac{z^{-d}B^{+}(z^{-1})B^{-}(z^{-1})}{A(z^{-1})} \tag{1}$$

where d is the known number of pure time step delays; $B^{-}(z^{-1})$ is the uncancellable portion of $B(z^{-1})$;$B^{+}(z^{-1})$ is the cancellable portion of $B(z^{-1})$.

Based on the internal model (IM) principle [13], the zero-error tracking of any reference input in the steady state can be achieved if a generator of the reference input is included in the stable closed-loop system. Therefore, for a periodic reference input, the RC controller $G_{rc}(z)$ is plugged into the system shown in Fig. 1 as follows [16]:

$$G_{rc}(z) = G_r(z)G_f(z) = \frac{k_r z^{-N_1}}{1 - z^{-N}} G_f(z) \tag{2}$$

where the repetitive signal generator $G_r(z)$ and the filter $G_f(z)$ are chosen as follows:

$$G_r(z) = \frac{k_r z^{-N_1}}{1 - z^{-N}} = \frac{k_r z^{N_2}}{z^N - 1} \tag{3}$$

$$G_f(z) = \frac{z^{-n_u}A(z^{-1})B^{-}(z)}{B^{+}(z^{-1})b} \tag{4}$$

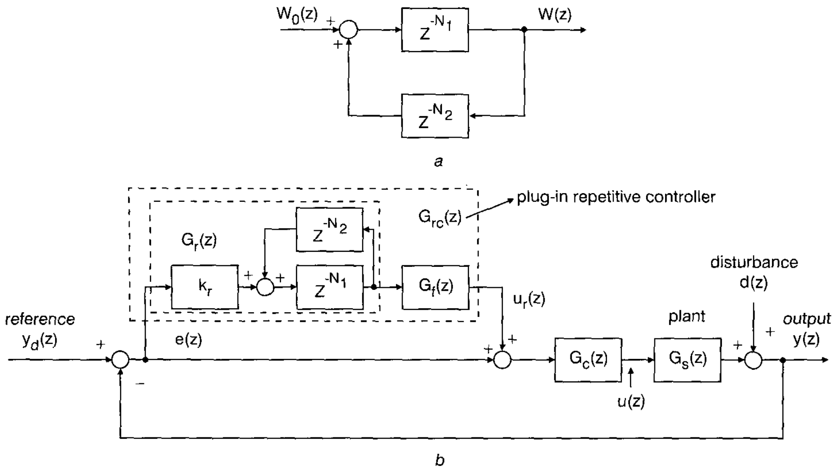


Fig. 1 *Repetitive control system*
a Periodic signal generator
b Plug-in repetitive control system

where k_r is the *repetitive control gain*; $N = N_1 + N_2 = f/f_c$ with f being the reference signal frequency and f_c the sampling frequency; $N_2 = n_u + d$; $B^-(z)$ is obtained from $B^-(z^{-1})$ with z^{-1} replaced by z; b is a scalar chosen so that $b \geq [B^-(1)]^2$; n_u is the order of $B^-(z^{-1})$, and z^{-n_u} makes the filter realizable. $G_f(z)$ in (4) is an implementation of the zero phase error tracking controller (ZPETC) design [23].

From Fig. 1, the transfer functions from $y_d(z)$ and $d(z)$ to $y(z)$ in the overall closed loop system are respectively derived as

$$\begin{aligned}\frac{y(z)}{y_d(z)} &= \frac{(1 + G_r(z)G_f(z))G_c(z)G_s(z)}{1 + (1 + G_r(z)G_f(z))G_c(z)G_s(z)} \\ &= \frac{(1 - z^{-N} + k_r z^{-N_1} G_f(z))H(z)}{1 - z^{-N}(1 - k_r z^{N_2} G_f(z)H(z))}\end{aligned} \tag{5}$$

$$\frac{y(z)}{d(z)} = \frac{1}{1 + G_c(z)G_s(z)}(1 - z^{-N}) \frac{1}{1 - z^{-N}(1 - k_r z^{N_2} G_f(z)H(z))} \tag{6}$$

'rom eqns. 5 and 6, it can be concluded that the overall :losed-loop system is stable if the following two conditions ıold: (i) The roots of $1 + G_c(z)G_s(z) = 0$ are inside the unit :ircle; and (ii):

$$\|1 - k_r z^{N_2} G_f(z)H(z)\| < 1, \qquad \text{for all } z = e^{j\omega}, 0 < \omega < \frac{\pi}{T} \tag{7}$$

Iowever, in practice, it is impossible to implement the :PETC filter $G_f(z)$ exactly because of model uncertainty. .et the unmodelled dynamics $\Delta(z)$ be represented in the orm of a multiplicative modelling error. Then the relation ıetween the actual system transfer function $H_t(z)$ and the ominal system transfer function $H(z)$ can be written as

$$H_t(z) = H(z)(1 + \Delta(z)) \tag{8}$$

/here all poles of $H_t(z)$ are assumed to be inside the unit ircle.

Then, using eqns. 1 and 4, the following product can be expressed as

$$G_f(z)H_t(z) = z^{-N_2}\frac{B^-(z)B^-(z^{-1})}{b} + \Delta(z) \tag{9}$$

And it is assumed that there exists a constant ϵ such that $\|\Delta(z)\| \leq \epsilon$. Therefore, eqn. 7 leads to a conservative stable range for k_r as follows

$$0 < k_r < \frac{2}{\max \|B^-(z)B^-(z^{-1})/b + \Delta(z)\|} \leq \frac{2}{1+\epsilon} \tag{10}$$

From Fig. 1, the error transfer function for the overall system can be derived as

$$G_e(z) = \frac{e(z)}{y_d(z) - d(z)} = \frac{1}{1 + G_c(z)G_s(z)}(1 - z^{-N}) \frac{1}{1 - z^{-N}(1 - k_r z^{N_2} G_f(z)H_t(z))} \tag{11}$$

Obviously, if the overall closed-loop system shown in Fig. 1 is asymptotically stable and the angular frequency ω of the reference input $y_d(t)$ and the disturbance $d(t)$ approaches $\omega_m = 2\pi mf$, $m = 0, 1, 2, \ldots, M$ ($M = N/2$ for even N and $M = (N-1)/2$ for odd N), then $z^{-N} \to 1$, $lim_{\omega \to \omega_m} \|G_e(j\omega)\| = 0$, and thus

$$\lim_{\omega \to \omega_m} \|e(j\omega)\| = 0 \tag{12}$$

eqn. 12 indicates that zero steady-state tracking error is obtained with the RC controller for any periodic reference input whose frequency is less than half of the sampling frequency. To enhance the robustness of the system, a low-pass filter $Q(z, z^{-1})$ is used in RC controller as follows [17]:

$$G_r(z) = \frac{k_r Q(z, z^{-1}) z^{-N_1}}{1 - Q(z, z^{-1}) z^{-N}} \tag{13}$$

where

$$Q(z, z^{-1}) = \frac{\sum_{i=0}^{m} \alpha_i z^i + \sum_{i=1}^{m} \alpha_i z^{-i}}{2\sum_{i=1}^{m} \alpha_i + \alpha_0} \tag{14}$$

where $\alpha_i (i = 0, 1, \ldots, m;\ m = 0, 1, 2, \cdots)$ are coefficients to be designed.

Notice that $Q(z, z^{-1})$ is a moving average filter that has zero phase shift and brings all open-loop poles inside the unit circle except the one at +1. On the other hand, high-frequency periodic disturbances are not perfectly cancelled by this controller. In this case, there is a trade-off between tracking precision and system robustness [24]. And correspondingly, eqn. 7 is modified as follows [18]:

$$\|1 - k_r z^{N_2} G_f(z) H_t(z)\| < \| \frac{1}{Q(z, z^{-1})} \| \qquad (15)$$

3 RC controller for CVCF PWM DC/AC converters

3.1 Modelling CVCF PWM DC/AC converters

The dynamics of the CVCF PWM DC/AC converter (as shown in Fig. 2(a)) can be described as follows [1]:

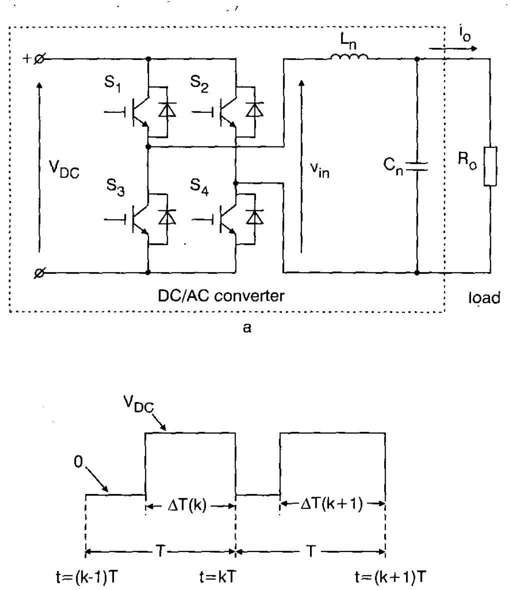
$$\begin{bmatrix} \dot{v}_c \\ \ddot{v}_c \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ -\frac{1}{L_n C_n} & -\frac{1}{C_n R_n} \end{bmatrix} \begin{bmatrix} v_c \\ \dot{v}_c \end{bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{L_n C_n} \end{bmatrix} v_{in} \qquad (16)$$

where v_c is the output voltage; i_o is the output current; v_{dc} is the DC bus voltage; L_n, C_n, and R_n are the nominal values of the inductor, capacitor and load, respectively; as shown in Fig. 2b, the control input v_{in} is a PWM voltage pulse of magnitude v_{dc} (or $-v_{dc}$) with width ΔT centred in the sampling interval T.

For a linear system $\dot{\boldsymbol{x}} = A\boldsymbol{x} + B\boldsymbol{u}$, its sampled-data equation can be expressed as $\boldsymbol{x}(k+1) = e^{AT}\boldsymbol{x}(k) + \int_0^T e^{A(T-\tau)} B\boldsymbol{u}(\tau) d\tau$. Therefore, a sampled-data form for eqn. 16 can be derived as follows:

$$\begin{bmatrix} v_c(k+1) \\ \dot{v}_c(k+1) \end{bmatrix} = \begin{bmatrix} \varphi_{11} & \varphi_{12} \\ \varphi_{21} & \varphi_{22} \end{bmatrix} \begin{bmatrix} v_c(k) \\ \dot{v}_c(k) \end{bmatrix} \pm \begin{bmatrix} g_1 \\ g_2 \end{bmatrix} \Delta T(k) \qquad (17)$$

where coefficients $\varphi_{11} = 1 - T^2/2L_nC_n$, $\varphi_{21} = -T/L_nC_n + T^2/2L_nC_n^2R_n$, $\varphi_{12} = T - T^2/2C_nR_n$, $\varphi_{22} = 1 - T/C_nR_n - T^2/2L_nC_n + T^2/2C_n^2R_n^2$, $g_1 = ET/2L_nC_n$, $g_2 = E/L_nC_n(1 - (T/2C_nR_n))$.



b

Fig. 2 *CVCF PWM DC/AC converter and PWM input waveform*
a CVCF PWM DC/AC converter
b PWM waveform for v_{in}

3.2 Problem formulation

Consider the DC/AC converter described by eqn. 17 and its output equation

$$y(k) = v_c(k) \qquad (18)$$

The objective of the controller is to force the tracking error between $y(k)$ and its sinusoidal reference $y_d(k)$ with the period of $N*T$ to approach zero asymptotically.

3.3 Controller design

According to the theory in Section 2, the controller for CVCF PWM DC/AC converter comprises a conventional feedback controller and a plug-in RC controller.

3.3.1 Conventional feedback controller: The ARMA equation for the dynamics (eqns. 17, 18) can be obtained as follows:

$$y(k+1) = -p_1 y(k) - p_2 y(k-1) + m_1 u(k) + m_2 u(k-1) \qquad (19)$$

where $u(k) = \pm\Delta T(k)$; $p_1 = -(\varphi_{11} + \varphi_{22})$, $p_2 = \varphi_{11}\varphi_{22} - \varphi_{21}\varphi_{12}$, $m_1 = g_1$, $m_2 = g_2\varphi_{12} - g_1\varphi_{22}$. If the control law for the plant (eqn. 19) is chosen as

$$u(k) = \frac{1}{m_1}[y_d(k) - m_2 u(k-1) + p_1 y(k) + p_2 y(k-1)] \qquad (20)$$

then $y(k+1) = y_d(k)$. It yields a deadbeat responsc $H(z) = z^{-1}$. eqn. 20 describes a one sampling aheac preview (OSAP) controller [1].

3.3.2 Plug-in repetitive controller

As well as a sampling time tracking delay, the OSAI controller depends on the model having accurate L_n, C and R_n. In practice, parameter uncertainties ΔL, ΔC an load disturbance ΔR yield large tracking errors. Therefore a RC controller is proposed to overcome the periodi disturbance and parameters variation. According t design theory mentioned in Section 2, $G_f(z) = 1/H(z) = z$ and the RC controller $G_{rc}(z)$ is proposed

$$G_{rc}(z) = G_r(z)G_f(z) = \frac{k_r z^{-N+1} Q(z, z^{-1})}{1 - Q(z, z^{-1})z^{-N}} \qquad (21$$

For simplicity of design and analysis, $Q(z, z^{-1}) = 1$ an $N_2 = 0$ in this case. In sampled-data form, the RC controller can be expressed as follows

$$u_r(k) = u_r(k-N) + k_r e(k-N+1) \qquad (22$$

In fact, eqn. 22 is the same as anticipatory learning contrc law [11].

3.4 Robustness analysis

In view of the uncertainties ΔL, ΔC and ΔR, the ARM equation for the actual plant becomes

$$y(k+1) = -a_1 y(k) - a_2 y(k-1) + b_1 u(k) + b_2 u(k-1) \qquad (23$$

where $a_1 = p_1 + \Delta p_1$, $a_2 = p_2 + \Delta p_2$, $b_1 = m_1 + \Delta m_1$ an $b_2 = m_2 + \Delta m_2$ are calculated on the basis of the practic parameters $L = L_n + \Delta L$, $C = C_n + \Delta C$, $R = R_n$ $\Delta R \in (R_{min}, \infty)$. When an OSAP controller (eqn. 20) i

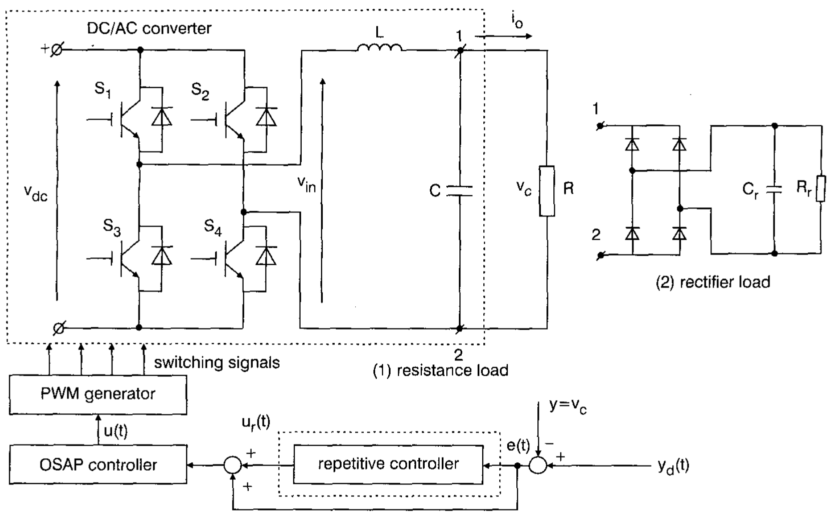


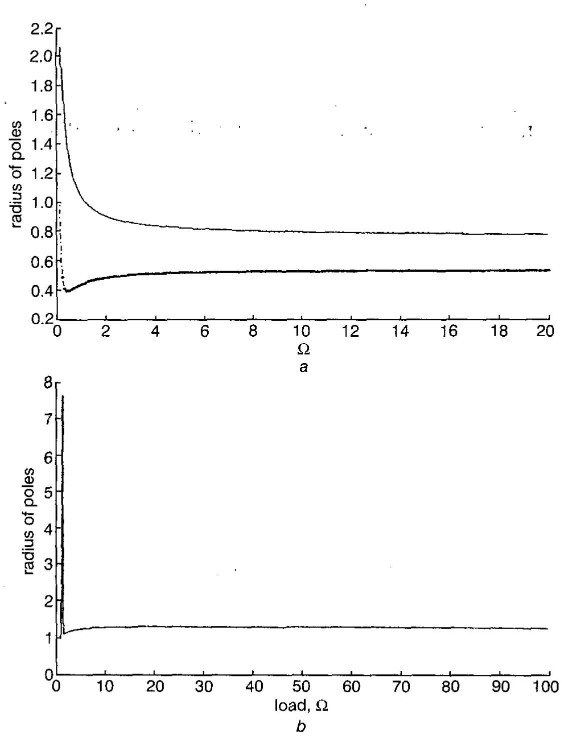
Fig. 3 *Controlled DC/AC converter system*

applied to the plant (eqn. 23), the closed-loop transfer function $H_t(z)$ without an RC controller becomes

$$H_t(z) = \frac{(b_1 + b_2z^{-1})}{(z + a_1 + a_2z^{-1})(m_1 + m_2z^{-1}) - (p_1 + p_2z^{-1})(b_1 + b_2z^{-1})} \quad (24)$$

When $L = L_n$, $C = C_n$, $R = R_n$, a deadbeat response $H_t(z) = H(z) = z^{-1}$ is achieved.

In practice, to enhance the robustness, $Q(z, z^{-1})$ can set to be $d_1z + d_0 + d_1z^{-1}$. This doesn't influence the above analysis results. According to the stability analysis in Section 2, the overall system is stablised if (i) all poles of $H_t(z)$ in eqn. 24 are inside the unit circle; (ii) $\|1 - k_rzH_t(z)\| < 1 \leq 1/|Q(z, z^{-1})|$. k_r can be larger because of the introduction of $Q(z, z^{-1})$.



ig. 4 *Stability analysis*
ı Radius of poles of $H_t(z)$
ʼ Maximum $\|e^{jw}H_t(e^{jw})\|$

3.5 Simulation and experiment

Our simulation and experimental studies are carried out using the schematic diagram of a controlled DC/AC converter system shown in Fig. 3.

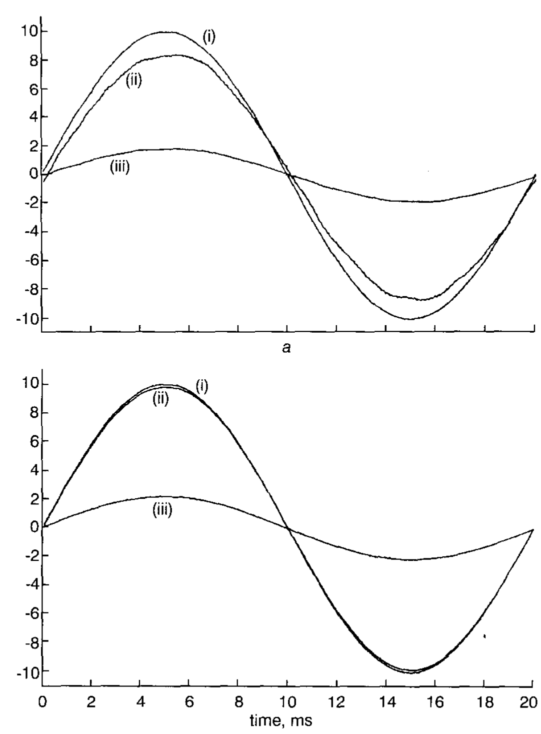


Fig. 5 *Steady-state simulation results with resistance load $R = 4.7\Omega$*
a OSAP controlled reference voltage $y_d(t)$, output voltage $v_c(t)$, output current $i_o(t)$
b Reference voltage $y_d(t)$, output voltage $v_c(t)$, output current $i_o(t)$ with OSAP plus RC
(i) $y_d(t)$, (ii) $v_c(t)$, (iii) $i_o(t)$

In particular, the DC/AC converter has the parameter values: $C_n = 700\,\mu$F; $L_n = 600\,\mu$H; $R_n = 2\,\Omega$; $C = 800\,\mu$F, $L = 700\,\mu$H; $y_d(t)$ is 50 Hz, 10 V (peak) sinusoidal signal; $v_{dc} = 20$ V; $f = 50$ Hz; $f_c = 1/T = 6.25$ kHz; $d_0 = 0.9$, $d_1 = 0.05$.

As shown in Fig. 4*a*, with these above parameter values and when $R > 1\,\Omega$, all the poles of the closed-loop transfer function $H_t(z)$ in eqn. 24 without the RC controller are located inside the unity circle, the system is stable.

As shown in Fig. 4*b* , the maximum gain of $zH_t(z)$ in frequency domain is no more than 8. According to the stability condition $\|1 - k_r zH_t(z)\| < 1$ for RC control design, the system with RC controller is stable if $k_r \in (0, 0.25)$. We set $k_r = 0.03$.

3.5.1 Simulation results

Figs. 5 and 6 show the simulation results of the only OSAP controlled and RC plus OSAP controlled CVCF PWM DC/AC converter with resistance load and uncontrolled rectifier load, respectively. With OSAP controller, the peaks of tracking error $e(t)$ between output voltage and reference voltage are about 1.8 V in in Fig. 5*a* and about 2.3 V Fig. 6*a*. Figs. 5*b* and 6*b* show the RC controller force the output voltage to approach reference voltage under different loads and significantly reduce the tracking error, respectively.

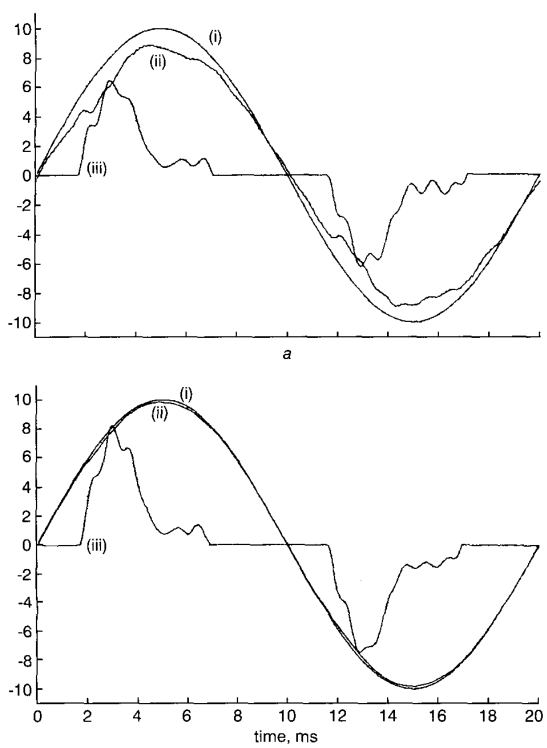


Fig. 6 *Steady-state simulation results with rectifier load $R_r = 4.7\Omega$, $C_r = 1470\,\mu F$*

a OSAP controlled reference voltage $y_d(t)$, output voltage $v_c(t)$, output current $i_o(t)$
b Reference voltage $y_d(t)$, output voltage $v_c(t)$, output current $i_o(t)$ with OSAP plus RC
(i) $y_d(t)$, (ii) $v_c(t)$, (iii) $i_o(t)$

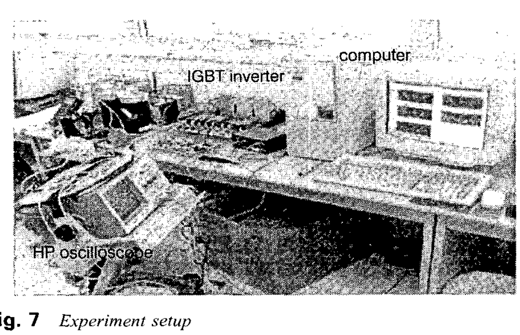


Fig. 7 *Experiment setup*

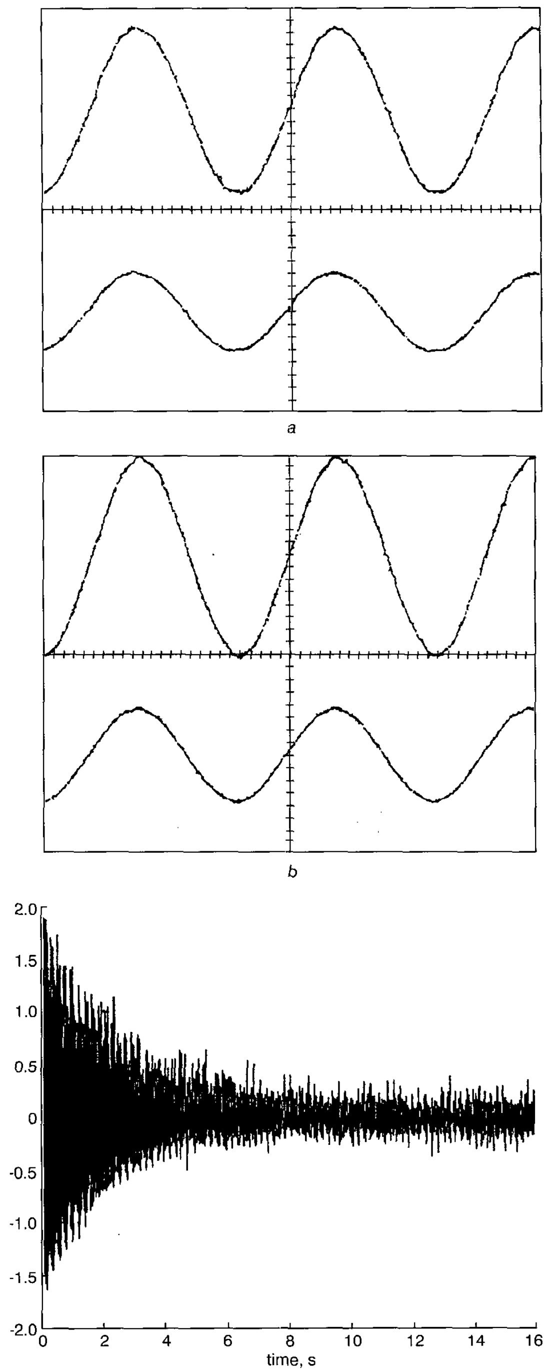


Fig. 8 *Experimental results with resistance load*

a Output voltage $v_c(t)$ (upper, 5V/division), output current $i_o(t)$ with OSAP controller
b Output voltage $v_c(t)$ (upper, 5V/division), output current $i_o(t)$ with OSAP plus RC
c Transient response of tracking error $e(t) = y_d(t) - v_c(t)$ with OSAP plus RC

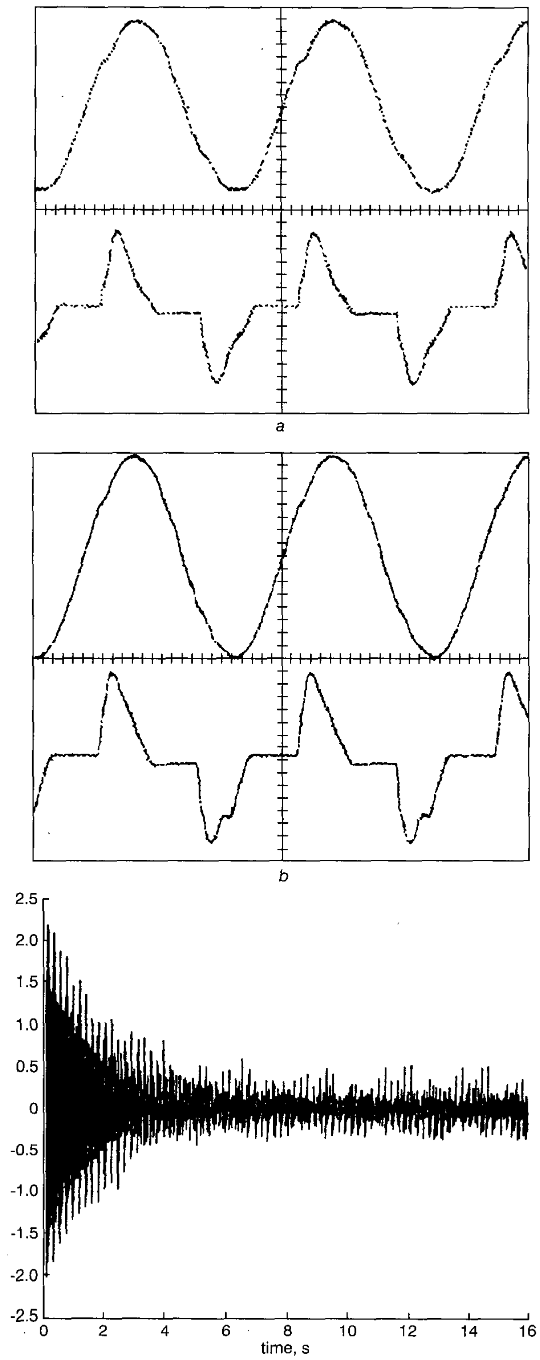
3.5.2 Experimental results

Experiment setup (Fig. 7) has been built for the converter system shown in Fig. 3. Facilities include DSPACE (DS1102) DSP development toolkits, H-bridge IGBT switches converter and a HP 546002 oscilloscope.

Fig. 8 shows the experimental results of the only OSAP and OSAP plus RC controlled CVCF PWM DC/AC converter under resistance load ($R = 4.7\,\Omega$). Output voltage is about 8.2 V with OSAP controller in Fig. 8*a*; output voltage approaches 10V with OSAP plus RC controller in Fig. 8*b*. Furthermore, Fig. 8*c* shows the tracking error $e(t)$ is reduced from about 1.8 V to be less than 0.3 V after about 4 s when RC controller is plugged into the OSAP controlled converter.

Fig. 9 shows the experimental results of the only OSAP and OSAP plus RC controlled CVCF PWM DC/AC converter under uncontrolled rectifier load ($C_r = 1470\,\mu$F, $R_r = 4.7\,\Omega$). Output voltage is about 7.7 V with OSAP controller and is distorted by current surge in Fig. 9*a*; output voltage approaches 10 V and has less distortion with OSAP plus RC controller in Fig. 9*b*. Furthermore, Fig. 9*c* shows the peak of tracking error $e(t)$ is reduced from about 2.3 V to be less than 0.4 V after about 4 s when RC controller is plugged into the OSAP controlled converter.

The experimental transient tracking errors are collected through A/D converters of DS1102 card and TRACE software. The residual tracking errors can be further reduced by improving the sampling frequency f_c.



ig. 9 *Experimental results with uncontrolled rectifier load*
Output voltage $v_c(t)$ (upper, 5V/division), output current $i_o(t)$ with OSAP ontroller
Output voltage $v_c(t)$ (upper, 5V/division), output current $i_o(t)$ with OSAP plus .C
Transient response of tracking error $e(t) = y_d(t) - v_c(t)$ with OSAP plus RC

4 Conclusion

A plug-in discrete time repetitive learning control scheme has been proposed for the CVCF PWM DC/AC converter systems. The periodic tracking errors caused by nonlinear load disturbances (such as rectifier load) and parameter uncertainties (ΔL and ΔC) are eliminated by the plug-in repetitive learning controller. It is shown that the proposed control scheme offers zero error tracking capability for the CVCF PWM DC/AC converter systems under different loads and parameter uncertainties. Minimised output voltage THD and fast response are achieved. Simulation and experimental results are provided to demonstrate the validity of the proposed control scheme.

5 References

1 KAWAMURA, A., HANEYOSHI, T., and HOFT, R.G.: 'Deadbeat controlled PWM inverter with parameters estimation using only voltage sensor', *IEEE Trans. Power Electron.*, 1988, **PE-3**, (2), pp. 118–125

2 GOKHALE, K.P., KAWAMURA, A., and HOFT, R.G.: 'Dead deat microprocessor control of PWM inverter for sinusoidal output waveform synthesis'. *IEEE Power electronics specialist conference*, 1985, pp. 28–36

3 KAWAMURA, A., and ISHIHARA, K.: 'Real time digital feedback control of three phase PWM inverter with quick transient response suitable for uninterruptible power supply'. Industry Applications Society Annual Meeting, 1988, pp. 728–734

4 CARPITA, M., and MARCHESONI, M.: 'Experimental study of a power conditioning system using sliding mode control', *IEEE Trans. Power Electron.*, 1996, **PE-11**, (5), pp. 731–733

5 JUNG, S.L., and TZOU, Y.Y.: 'Discrete feedforward sliding model control of PWM inverter for sinusoidal output waveform synthesis'. *IEEE Power electronics specialist conference*, 1994, pp. 552–559

6 KAWAMURA, A., and HOFT, R.G.: 'Instantaneous feedback controlled PWM inverter with adaptive hysteresis', *IEEE Trans. Ind. Appl.*, 1984, **IA-20**, (4), pp. 769–775

7 HSIN, Y., LONGMAN, R., SOLCZ, E., and DE JONG, J.: 'Experiments bridging learning and repetitive control', Advances in the Astronautical Sciences part 2 1997, **95**, pp. 671–690

8 TOSHIHARU, S., and TOSHIRO, O.: 'An iterative learning control law for dynamical systems', *Automatica*, 1991, **27**, (4), pp. 729–732

9 ARIMOTO, S., KAWAMURA, S., and MIYAZAKI, F.: 'Bettering operation of dynamic system learning: A control theory for servo mechanism or mechatroncs systems'. Proc of 23th CDC, 1984, pp. 1064–1069

10 ARIMOTO, S., KAWAMURA, S., and MIYAZAKI, F.: 'Learning control theory for dynamical systems'. Proc of 24th CDC, 1984, pp. 1375–1380

11 WANG, D.: 'On D-type and P-type ILC designs and anticipatory approach' (Special issue on iterative learning control), *Int. J. Control*, **73**, (10), pp. 890–901

12 HARA, S., YAMAMOTO, Y., OMATA, T., and NAKANO, M.: 'Repetitive control system: a new type servo system for periodical exogenous signals', *IEEE Trans. Autom. Control*, 1988, **AC-33**, pp. 659–667

13 FRANCIS, B.A., and WONHAM, W.M.: 'The internal model principle of control theory', *Automatica*, 1976, **12**, pp. 457–465

14 INOUE, T.: 'High accuracy control of a proton synchrotron magnet power supply'. Proc. 8th World Congress of IFAC, 1981 Kyoto, Japan, pp. 3137–3142

15 INOUE, T.: 'High accuracy control of servomechanism for repeated contouring'. Proc. 10th Annual symp. *Incremental motion control system and devices*, 1981, pp. 258–292

16 TOMIZUKA, M., TSAO, T., and CHEW, K.: 'Analysis and synthesis of discrete-time repetitive controllers', *Trans. ASME: J. Dynam. Syst. Meas. Control*, 1988, **110**, pp. 271–280

17 COSNER, C., ANWAR, G., and TOMIZUKA, M.: 'Plug in repetitive control for industrial robotic manipulators'. Proceedings of the IEEE International Conference on *Robotics and automation*, 1990, pp. 1970–1975

18 CHEW, K.K., and TOMIZUKA, M.: 'Digital control of repetitive errors in disk drive systems'. Proceedings of American Control Conference, 1989, pp. 540–548

19 MANAYATHARA, T.J., TSAO, T.C., BENTSMAN, J., and ROSS, D.: 'Rejection of unknown periodic load disturbances in continuous steel casting process using learning repetitive control approach', *IEEE Trans. Control Syst. Technol.*, 1996, **4**, (3), pp. 259–265

20 BROBERG, H.L., and MOLYET, R.G.: 'Correction of period errors in a weather satellite servo using repetitive control'. 1st IEEE Conference on *Control application*, Dayton, OH, Sep. 1992, pp. 682–683

21 HANEYOSHI, T., KAWAMURA, A., and HOFT, R.G.: 'Waveform compensation of PWM inverter with cyclic fluctuating loads', *IEEE Power electronics specialist conference*, 1987, pp. 745–751

22 TZOU, Y.Y., OU, R.S., JUNG, S.L., and CHANG, M.Y.: 'High-performance programmable AC power source with low harmonic distortion using DSP-based repetitive control control technique', *IEEE Trans. Power Electron.*, 1997, **PE-12**, pp. 715–725

23 TOMIZUKA, M.: 'Zero phase error tracking algorithm for digital control', *Trans. ASME: J. Dynam. Syst. Meas. Control*, 1987, **109**, (2), pp. 65–68

24 BROBERG, H.L., and MOLYET, R.G.: 'Reduction of repetitive errors in tracking of periodic signals: theory and application of repetitive control'. 1st IEEE Conference on *Control application*, Sep. 1992, Dayton, OH, pp. 1116–1121