Adaptive Repetitive Control with Feedforward Scheme for Grid-Connected Inverters

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Abstract— This paper is concerned with control of grid connected inverters using odd harmonic repetitive scheme. Due to the inherent long convergence period of repetitive control, a feedforward loop of the fundamental component of the grid voltage is used to make sure that the inverter's output current is bounded during the convergence period. The well know deficiency of repetitive control in rejecting harmonics when used with varying grid frequency has been solved by using an adaptive scheme in which the switching and sampling frequencies vary with respect to the grid frequency. The main qualities of the proposed controller are its effectiveness, simplicity and straightforward of implementation using a Digital Signal Processor. Simulation and practical results are presented to demonstrate the controller capabilities.

Keywords—grid-connected inverters; repetitive control; adaptive frequency control.

NOMENCLATURE

n	Number of samples on one fundamental cycle
N_{cpu}	PWM counter period
N_{cpu_o}	Nominal PWM counter period
T_{cpu}, f_{cpu}	DSP clock period and frequency
T_g, f_g	Grid voltage fundamental period and frequency
T_{inv}, f_{inv}	inverter modulating signal period and frequency
T_d	DSP computational time delay
T_s, f_s	Sampling period and frequency
T_{so}, f_{so}	Nominal sampling period and frequency
T_{sw}, f_{sw}	Switching period and frequency

I. INTRODUCTION

In grid-connected inverters repetitive controllers (RC) [RC references] have the advantage of much higher gain at the fundamental frequency and its harmonic than that provided by classical PID controllers and their derivatives [PID references]. Hence, they are capable of producing significantly higher quality output current, as demanded by national and international standards [harmonic

standards references], over a wider power output range than classical PID. They are also much simpler to implement in practice than proportional resonance controllers [PR references].

However, using RC alone in grid-connected inverters may not provide satisfactory transient response. The conventional RC structure [27]-[29], which has the ability to reject both even and odd harmonics has an inherent control delay of one fundamental period and thus suffers from slow dynamic response. Odd-harmonic RC [18]-[20], [30],[31], which makes use of the fact that most AC systems have only odd harmonics in their spectrum, has an inherent time delay of half the period of a fundamental cycle, which makes its dynamic response faster than that of conventional RC. But that may not be sufficient to deal with disturbance caused by grid voltage swell or sag could result in the inverter generating output current that is much higher than the reference current. Relying on RC alone to deal with such disturbance means that the current error could be too high during the RC convergence time, which could cause an over current fault.

Another shortcoming of RC is that its gain falls very rapidly either side of its resonant frequencies. Hence, if the grid frequency deviates slightly from its nominal value, the high resonant gains will not be lined up with the grid fundamental frequency and its harmonics causing serious deterioration in the controller's performance. Hornik and Zhong [24],[25] proposed voltage control of grid connected inverters with a frequency adaptive mechanism that is based on H_∞ repetitive control. The internal model of the system consists of a delay unit $e^{-T_D s}$ cascaded with a low-pass filter, $W(s) = \omega_c / (s + \omega_c)$. The adaptive mechanism is based on varying the cut-off frequency of the filter ω_c according to the varying grid frequency. According to the authors, this mechanism is effective for a grid frequency variation of only $\pm 0.2\text{Hz}$. If the grid variation is higher than this limit, the controller delay T_D needs to be changed. However, for low sampling frequency, an adaptive delay is impossible to be implemented without further deterioration of the controller performance [24].

To overcome the above deficiencies, this paper proposes an odd-harmonic repetitive controller, with a feed forward loop to speed its transient response and a fast adaptive frequency mechanism. The frequency adaptive feature is based on varying both switching and sampling frequencies according to the grid frequency and thus keeping the number of delay samples of the RC constant. The design results in fast convergence to steady state, within one fundamental cycle. The novelty of the paper is twofold: improving the transient response of the RC by using a feedforward loop of the grid voltage and improving the steady state performance of the RC by using an adaptive scheme that tracks the variation in grid frequency. The main qualities of the proposed controller are its effectiveness, simplicity and straightforward of implementation using a Digital Signal Processor.

The rest of the paper is organized as follows: section II describes the controller structure. Section III describes the proposed frequency adaptive mechanism. Finally, simulation and practical results are presented in sections IV and V, respectively.

II. CONTROLLER STRUCTURE AND DESIGN

Fig. 1a shows the circuit diagram of a conventional 2-level grid-connected inverter with LCL filter [3],[7]. The inverter parameters used in this paper are listed in Table I. The selection of the filter component values is a trade-off between inductor size, IGBT switching losses, resonant frequency, and the degree of filter attenuation at the switching frequency. The switching frequency has been chosen to be 8 kHz which is limited by the maximum IGBTs power dissipation. The L_1 inductance of 350 μ H has been chosen to limit the ripple current to a value of ~25% of the rated current (140A) as dictated by the ratings and losses in the IGBT modules. The values of L_2 and C have been chosen so the impedance of L_2 at the switching frequency is several times bigger than the impedance of C in order to prevent the ripple current from being injected into the grid. The computational time delay is determined by the speed of the DSP in processing the real time calculations.

The block diagram that represents the control scheme implemented in each phase is shown in Fig. 1b. The controller consists of an outer loop of the output current and an inner loop of the capacitor current to provide active damping. This structure increases the degree of freedom in designing the controller compared with the widely used one-feedback loop of L_1 current because two controller gains (K_C and K) can be optimized instead of only one controller gain.

A plug-in repetitive controller transfer function (G_{RC}) is implemented to reject current harmonics caused by grid voltage harmonics. The controller diagram is simplified as shown in Fig. 1c where $D_g(z)$ represents the grid disturbance and $G_p(z)$ is the open loop transfer function in the discrete time domain that takes into account the zero order hold and time delay effects and is given as

$$G_p(z) = \frac{Z \left(\frac{1 - e^{-sT_s}}{s} e^{-sT_d} \frac{1}{L_1 L_2 C s^3 + (L_1 + L_2)s} \right)}{1 + Z \left(\frac{1 - e^{-sT_s}}{s} e^{-sT_d} \frac{K_C L_2 C s^2}{L_1 L_2 C s^3 + (L_1 + L_2)s} \right)}, \quad (1)$$

where Z is the z-transform operator, T_s is the sampling time, T_d is the computational time delay, and K_C is the capacitor current loop gain. The transfer function of the odd harmonic repetitive controller is given in [30] as

$$G_{RC}(z) = -\frac{K_R z^m Q(z) z^{-n/2}}{1 + Q(z) z^{-n/2}}, \quad (2)$$

where $Q(z)$ is a low pass filter, K_R is the repetitive controller gain, n is the number of samples in one fundamental cycle and z^m is a non-causal phase lead unit.

The repetitive controller is capable of rejecting the disturbance of the fundamental harmonic and its multiples. However, the convergence time of the repetitive controller is relatively long and therefore a feedforward loop of the fundamental nominal component of the grid voltage is implemented to reduce the current error during the convergence time of the RC.

Fig. 1. Grid connected inverter (a) and its controller (b) and simplified controller (c)

TABLE I
INVERTER PARAMETERS

Description	Symbol	Value
Inverter side filter inductor	L_1	350 μ H
Filter capacitor	C	80 μ F
Grid side filter inductor	L_2	50 μ H
Nominal grid voltage	V_{go}	230 V (rms)
Inverter dc voltage	V_{dc}	700 Vdc
Nominal grid frequency	f_{go}	50Hz
Nominal switching frequency	f_{swd}	8kHz
Nominal sampling frequency	f_{so}	16kHz
Computational time delay	T_d	10 μ s

The controller design involves the determination of K_C , K , $Q(z)$, K_R , and m . The error signal E (see Fig. 1c) can be expressed as

$$E(z) = \frac{I_o^*(z) + D_g(z)G_p(z)}{1 + KG_p(z) + G_p(z)G_{RC}(z)} \quad (3)$$

Substituting (2) in (3) and rearranging gives

$$\frac{E(z)}{I_o^*(z) + D_g(z)G_p(z)} = \frac{(1 + z^{-n/2}Q(z))}{(1 + KG_p(z))(1 - z^{-n/2}R(z))}, \quad (4)$$

where

$$R(z) = Q(z) \left(\frac{K_R z^m G_p(z)}{1 + KG_p(z)} - 1 \right). \quad (5)$$

Equation (4) can be represented by the block diagram in Fig. 2 which consists of three cascaded transfer functions: $T_1(z)$, $T_2(z)$, and $T_3(z)$. $T_1(z)$ is the closed loop transfer function without the repetitive controller and its stability can be guaranteed by choosing K_C and K . The stability of the second transfer function $T_2(z)$ can be guaranteed by choosing a stable low pass filter $Q(z)$. The stability of the third transfer function $T_3(z)$ which contains a positive feedback loop can be guaranteed using the small gain theorem, i.e., the error signal will be bounded if the magnitude of the open loop transfer function is less than 1 for all values of frequencies. Hence

$$\|R(z)\|_\infty < 1. \quad (6)$$

Fig. 2. Block diagram of the system error

A. Selection of K_C and K

The selection of the capacitor current loop gain K_C and the output current loop gain K is determined to give good stability margins. By analysing $G_p(z)$ in Matlab SISO Design Tool, the gains K_C and K have been set to 5 and 3, respectively. This selection gives gain margin of 5.6dB, phase margin of 51 deg, damping ratio $\zeta = 0.33$, and settling time $t_s = 0.5$ ms.

B. Selection of $Q(z)$

$Q(z)$ is a zero-phase low pass filter that is used to attenuate the high frequency resonant peaks of the RC, to ensure stability. It has the following structure [32].

$$Q(z) = \alpha_1 z + \alpha_0 + \alpha_1 z^{-1}, \quad (7)$$

where, α_0 and α_1 are set to 0.5, and 0.25, respectively [32].

The bode diagram of the open loop transfer function $(K + G_{RC}(z))G_p(z)$ is shown in Fig. 3; it confirms the system's stability as the high gains near the crossover frequency have been attenuated and the system has positive stability margins.

Fig. 3. Bode diagram of the open loop transfer function

$$(K + G_{RC}(z))G_p(z), \text{ with } Q(z) = 0.25z + 0.5 + 0.25z^{-1}, m=1, K_R = 0.5 \text{ (GM} = 5.0\text{dB, PM} = 44.1 \text{ deg)}$$

C. Selection of K_R and m

The value of the RC gain K_R needs to be carefully selected as it is a key parameter for error convergence and system stability. A high repetitive controller gain K_R results in fast error convergence but the feedback system becomes less stable. The non-causal phase lead unit of m is normally used to compensate for any delay or phase lag introduced by the physical plant and controller transfer function. Implementing z^m is not possible practically unless it is cascaded with the delay units of the RC. The design criterion used in this paper is maximizing K_R so as to reduce the RC convergence time but at the same time minimizing $\|R(z)\|_\infty$ so as to increase the stability margins. By plotting $\|R(z)\|_\infty$ versus K_R and m it can be shown that when $m = 0$ the system is only stable when $K_R < 0.6$ ($\|R(z)\|_\infty = 0.99$ when $K_R = 0.6$). Increasing m does improve stability and the best value is $m = 3$ as this corresponds to the lowest possible $\|R(z)\|_\infty$. The minimum value for $\|R(z)\|_\infty$ occurs when $K_R = 2.8$ and $m = 3$ and hence they have been chosen for this design.

III. FREQUENCY ADAPTIVE REPETITIVE CONTROL

In order to keep the RC effective, it has to adapt to the varying grid frequency. The performance of the repetitive controller is not guaranteed unless the high resonant gains line up with the grid harmonics. Therefore, the time delay of the repetitive controller needs to adapt to the changes in grid voltage fundamental period. The number of delay samples n can be changed to cope with the changing frequency. However, unless the switching frequency (and hence the sampling frequency) are very high with respect to the fundamental frequency, this mechanism will not result in precise control of the time delay used by the repetitive controller. For example, if the sampling frequency is 4kHz, for a fundamental frequency of 50Hz, the number of samples per cycle is $4000/50 = 80$ samples. Each sample is equivalent to 0.253ms which means that the minimum change in grid

frequency that this scheme can deal with is approximately 0.63Hz. If the grid frequency deviates by only 0.15Hz from its nominal value, the RC gain at the 5th harmonic can reduce by about 20% of its nominal value. For 0.60 Hz deviation, the RC becomes completely ineffective. Therefore, varying the number of samples will not provide good tracking of grid frequency without deteriorating the RC performance.

The mechanism proposed in this paper is to change the switching frequency and the sampling frequency such that the ratio of the sampling frequency to the fundamental frequency remains constant and hence n does not need to change. The mechanism benefits from the high precision of the DSP clock used to implement the controller. Fig. 4 shows how the PWM carrier is implemented in the DSP. In this paper, the sampling frequency f_s is set to be twice the switching frequency such as $f_s = 2 f_{sw}$. A counter that is based on the Central Processing Unit (CPU) clock is set to count up and down periodically. The PWM counter period N_{cpu} is set to determine the required sampling period. Hence

$$N_{cpu} = \frac{T_s}{T_{cpu}}, \quad (8)$$

where T_s and T_{cpu} are the sampling and CPU clock periods, respectively. The number of samples n per one fundamental cycle is given by,

$$n = \frac{T_g}{T_s}, \quad (9)$$

where T_g is the grid voltage fundamental period. The inverter modulating signal period is given by

$$T_{inv} = nT_s. \quad (10)$$

Substituting (8) into (10) gives

$$T_{inv} = nN_{cpu}T_{cpu}. \quad (11)$$

According to (11), in order to vary the inverter's modulating signal period T_{inv} and hence the frequency f_{inv} while maintaining n constant, the number of counts N_{cpu} needs to be changed.

Fig. 5 shows the proposed controller of the PWM counter period N_{cpu} . The grid voltage fundamental period T_g is sensed (by a PLL or zero crossing detector) and divided by nT_{cpu} to calculate the demand PWM counter period N_{cpu}^* . The period error E_N is fed into a Proportional-Integrator (PI) controller to calculate ΔN_{cpu} which is added to the nominal PWM counter period N_{cpu_o} to produce N_{cpu} . The nominal PWM counter period N_{cpu_o} is calculated as

$$N_{cpu_o} = T_{so} / T_{cpu}, \quad (12)$$

where T_{so} is the nominal switching period.

To highlight the advantage of this mechanism over changing n , consider the case where the CPU frequency $f_{cpu} = 150\text{MHz}$, $n = 320$, the nominal switching frequency $f_{so} = 16\text{kHz}$, according to (12) the nominal PWM counter period $N_{cpu_o} = 150\text{MHz}/16\text{kHz} = 9375$. If the inverter modulating signal f_{inv} is controlled by varying n then by reducing n by 1, f_{inv} will be given by

$$f_{inv} = \frac{150\text{MHz}}{(320-1)*9375} = 50.1567 \text{ Hz} . \quad (13)$$

However, if f_{inv} is controlled by varying N_{cpu} as proposed here, by reducing N_{cpu} by 1 count, then f_{inv} will be given by

$$f_{inv} = \frac{150\text{MHz}}{320*(9375-1)} = 50.0053 \text{ Hz} . \quad (14)$$

It can be noticed that varying N_{cpu} gives 29 times more precision in controlling the frequency.

Fig. 4. PWM implementation in the DSP

Fig. 5. PWM period control

A. Frequency controller design

According to most of the grid codes of practice, the grid frequency variation has a maximum slope of 1 Hz/sec [12]. This is nearly equivalent to increasing or decreasing the grid period by about 0.4ms per second. The design objective is to track this

variation and maintain the error signal between the PWM counter period N_{cpu} and its demand N_{cpu}^* to the minimum of one count.

The deviation in grid frequency and hence the PWM counter period demand N_{cpu}^* will be modeled as a ramp function as:

$$N_{cpu}^* = \frac{D}{s^2}, \quad (15)$$

where D is the rate of change of N_{cpu}^* in count/sec. From Fig. 5 the error E_N (error signal between the PWM counter period N_{cpu} and its demand N_{cpu}^*) is given by

$$E_N(s) = \frac{N_{cpu}^* - N_{cpu-o}}{1 + (k_p + \frac{k_i}{s})}. \quad (16)$$

The steady state error E_{Nss} for a ramp input of N_{cpu}^* can be calculated by substituting (15) in (16) and using the final-value theorem,

$$E_{Nss} = \lim_{s \rightarrow 0} (sE_N) = \frac{D}{k_i}. \quad (17)$$

Maximum D is 184 count/sec (equivalent to 0.4ms per second). The integral gain k_i is chosen to give the minimum possible steady state frequency error which is one count. Therefore k_i is set to 184. The proportional gain k_p is normally set to deal with transient response to a step input, In this case, a step change in grid frequency is unlikely and the proportional gain is set to 10.

IV. SIMULATION RESULTS

Detailed simulation model of the 3-phase inverter presented in Fig. 1 has been built using the MATLAB SimPowerSystems. The system parameters were listed in Table I and the controller parameters are listed in Table II. Grid voltage harmonics were measured in the laboratory and similar values were included in the simulation model. Table III lists the amplitudes of the main grid harmonics. The total grid voltage THD was measured to be 1.9%.

Fig. 6a shows the output current without RC for a 14A (rms) demand. The THD is 14.2%. Fig. 6b shows the steady state output current with RC. The current THD is reduced to only 0.8%. The effectiveness of RC in improving the current THD is clearly noticed. Fig. 6c shows the output current harmonics with and without the RC.

Fig. 7a shows the output current during starting when the feedforward is deactivated. It can be noticed that during the first half cycle the output current is nearly five times the reference current. This is due to the disturbance caused by the fundamental component of the grid voltage. Fig. 7b shows the output current when the feedforward of the nominal fundamental component of the grid voltage is activated. The error in the first half cycle is small and it is due to the disturbance caused by the grid higher

harmonics. In Fig 7c, an error of 35V is suddenly introduced in the feedforward loop. This causes immediate change in the output current. The RC corrects for this error but it takes half fundamental cycle to converge. This shows the importance of the feedforward loop in complementing the RC.

To test the effectiveness of the adaptive frequency control, the grid frequency is set to start to change from 50Hz to 50.2Hz at simulation time $t=0.2\text{sec}$. The slope of change is 1Hz/sec. Fig. 8 shows the output current THD with and without activating the frequency adaptive control. Without the adaptive mechanism, the current THD increases as the frequency deviation increases and it reached 10%. Once the grid frequency reaches 50.2Hz and stops deviating, the THD dropped to 4%. The adaptive frequency control on the other hand was able to keep the output current THD at 0.3% all time.

TABLE II
CONTROLLER PARAMETERS

Description	Symbol	Value
CPU frequency	f_{cpu}	150MHz
Nominal CPU timer period	$N_{cpu,o}$	9375
Number of samples in one fundamental cycle	n	320
Output current feedback gain	K	3
Capacitor current feedback gain	K_C	5
RC gain	K_R	2.8
Noncausal phase lead	m	3
Frequency control proportional gain	k_p	10
Frequency control integral gain	k_i	184
Filter $Q(z)$ coefficient 1	α_0	0.5
Filter $Q(z)$ coefficient 2	α_1	0.25

TABLE III
MAIN GRID HARMONICS

Harmonic	% of fundamental
3	0.492
5	0.923
7	0.277
9	0.369
11	0.492
13	0.338
15	0.308
17	0.185
19	0.154

(a) Simulated steady state output current without RC (THD = 14.2%)

(b) Simulated steady state output current with RC (THD = 0.8%)

(c) Output current harmonics

Fig. 6. Simulated steady state output current and its harmonic spectrum with and without RC

(a) Simulated output current with RC but without the feedforward

(b) Simulated output current with RC and feedforward

(c) Output current with sudden change in the feedforward

Fig. 7. Simulated output current.

Fig. 8. Output current THD (Grid frequency started to change from 50Hz to 50.2Hz at $t = 0.2$ s with a slope of 1Hz/sec)

V. PRACTICAL IMPLEMENTATION AND EXPERIMENTAL RESULTS

The proposed controller was tested experimentally with the grid-connected inverter described in Fig. 1 and Table I. The control parameters were listed in Table II. The controller was implemented using a Texas Instrument TMS320F2812 32-bit fixed point DSP. The control algorithm was written in C programming language and compiled using Texas-Instrument Code Composer V5.3 tool.

The three-phase reference sine waves were generated internally by the DSP using lookup tables of $n = 320$ samples. The sine waves amplitude is set externally (by a setting in the user interface) and sent via Controller Area Network (CAN)-bus. The input DC is regulated by an external boost circuit to 700V dc so current can be injected into the 230Vrms grid.

The RC controller has been realized by programming as follow: from (2), the discrete transfer function that relates the RC output $Y(z)$ to the RC input $E(z)$ is given by

$$\frac{Y(z)}{E(z)} = -\frac{K_R z^m Q(z) z^{-n/2}}{1 + Q(z) z^{-n/2}}. \quad (18)$$

Substituting (7) in (18) and rearranging gives

$$Y(z) = X(z) \left(\alpha_1 z^{1+m-n/2} + \alpha_0 z^{m-n/2} + \alpha_1 z^{m-n/2-1} \right), \quad (19)$$

where $X(z)$ is given by

$$X(z) = -z^{-m} Y(z) - K_R E(z). \quad (20)$$

Equations (19) and (20) represent the indirect (standard) realization of digital controllers [33] and they are represented by the block diagram in Fig. 9. The RC controller is implemented in software by creating 3 arrays $x(i)$ (one per phase) each is 160 (320/2) entries long. At the discrete time i , the RC output $y(i)$ is calculated using the difference equation (21) and the array entry $x(i)$ is filled using the difference equation (22)

$$y(i) = \alpha_1 x(i+1+m-n/2) + \alpha_0 x(i+m-n/2) + \alpha_1 x(i+m-n/2-1), \quad (21)$$

$$x(i) = -y(i-m) - K_R e(i), \quad (22)$$

where $e(i)$ is the current error at discrete time i .

In order to implement the proposed frequency adaptive control, a high precision measurement of the grid voltage frequency is required. The grid voltage signal is sensed and the Positive-Going Zero Crossing (PGZC) is detected. The grid voltage fundamental period T_g is measured by calculating the number of samples between consecutive PGZC. In order to increase the measurement accuracy, the PGZC is detected every 15 fundamental cycles. In this way, the measurement error is reduced to $T_s/15$ (compared to T_s of the PGZC is detected every cycle). For the nominal sampling frequency of 16kHz, the measurement error is only $\pm 62.5\mu s/15 = \pm 4.16\mu s$, which is equivalent to $\pm 0.01\text{Hz}$.

Fig. 10a shows the output current when the RC is de-activated. The demand current is set to 15A (rms). The current THD is measured to be 13.0%. Fig. 10b shows the output current but when the RC is activated. The current THD is measured to be only 1.1%. In the system reported in [3] (which is similar to this one but without RC and $K(z)$ is a phase lag), such low THD was only achievable when the output current equals the rated current, i.e., 140A (rms). The RMS of the 50Hz fundamental current in Fig. 10a is 13A which is slightly less than the demanded RMS of 15A due to the deficiency of the pure proportional controller used. However, The RMS of the 50Hz fundamental current in Fig. 10b is 15A which means it follows very well the demanded current.

Fig. 10c shows the measured output current harmonics with and without the RC. The feedforward loop was activated all the time to avoid damaging high currents (see the simulation results in Fig. 7a). Fig. 10d shows the moment when the inverter connects to the grid. Before the main contactor closed, the inverter was producing output voltage that was matching exactly the grid voltage thanks to the feedforward loop of the nominal voltage of the fundamental component. This guaranteed smooth connection transient and low current when the contactor closed. To test the effect of the feedforward loop on the output current, an error of 35V is suddenly introduced in the feedforward loop as shown in Fig 10e. This causes immediate change in the output current. The RC corrects for this error but it takes half fundamental cycle to converge. The experimental result agrees with the simulation result in Fig 7c.

The grid frequency was monitored in the laboratory and the maximum deviation recorded was $\pm 0.1\text{Hz}$. The current THD was always maintained below 1.2%. In order to test the adaptive frequency controller to a higher grid frequency deviation, an AC voltage source would need to be used to emulate the grid.

Fig. 9. RC Implementation

(a) Output current without RC (10A/div)

(b) Output current with RC (10A/div)

(c) Experimental output current harmonics

(d) Inverter output voltage before and after connecting to the grid (5A/div)

(e) Output current with a sudden change in the feedforward (10A/div)

Fig. 10. Practical results

CONCLUSION

The design and practical implementation of a frequency adaptive odd-harmonic repetitive controller for a grid-connected inverter was discussed. The adaptive mechanism was found to be very effective in tracking the changes in grid frequency and hence maintaining the effectiveness of the repetitive controller. The proposed mechanism presents a straightforward implementation using a DSP system. The inherent long delay feature of RC has been overcome by implementing a feedforward loop at the fundamental frequency to ensure that the current is bounded during the convergence period of the repetitive controller. The Simulation and experimental results demonstrate the controller's capabilities.

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