Dual-Loop Power Control for Single-Phase Grid-Connected Converters with LCL Filter

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Abstract

Grid-connected converters have widely adopted LCL filters to acquire high harmonic suppression. However, the LCL filter increases the system order so that the design of the system stability would be complicated. Recently, sole-loop control strategies have been used for grid-connected converters with L or LC filters. But if the sole-loop control is directly transplanted to grid-connected converters with LCL filters, the systems may be unstable. This paper presents a novel dual-loop power control strategy composed of a power outer loop and a current inner loop. The outer loop regulates the grid-connected power. The inner loop improves the system stability margin and suppresses the resonant peak caused by the LCL filter. To obtain the control variables, a single-phase current detection is proposed based on PQ theory. The system transfer function is derived in detail and the influence of control gains on the system stability is analyzed with the root locus. Simulation and experimental results demonstrate the feasibility of the proposed control.

Key Words: Grid-connected converter, LCL filter, Power control, Power electronics, PQ theory, Root locus, Transfer function

NOMENCLATURE

L_1	Filter inductor
R_1	Inductor internal resistance
L_2	Filter inductor
R_2	Inductor internal resistance
C_f	Filter capacitor
V_{dc}	DC bus voltage
V_s	Utility grid voltage RMS
ω_0	Utility grid angular frequency
i_l	Converter output current
i_s	Grid-side current
v_s	Utility grid voltage
i_c	Filter capacitor current
v_c	Filter capacitor voltage
PLL	Phase-locked loop
PF	Power factor
VSC	Voltage source converter
THD	Total harmonic distortion
RMS	Root mean square

I. INTRODUCTION

Power electronic converters have been rapidly developed and widely applied in many fields. Distributed generation is one of these important application fields which uses power electronic converters to transmit energy from renewable energy resources to utility grids, such as solar, wind, micro-turbine,

and fuel cell, has won more and more attention [1]–[4]. Grid-connected converters which are the key component of distributed generators are mainly voltage source converters (VSCs).

The sinusoidal pulse width modulation technology has been broadly used in VSCs control systems for converting DC electricity to AC sinusoidal electricity. This modulation would generate harmonic at AC side. Generally, L, LC, LCL filters are connected at AC side in order to acquire sinusoidal AC electricity. A grid-connected VSC can attenuate high frequency harmonic effectively through an LCL filter which has potential benefits for a VSC to get higher harmonic suppression performance with lower switching frequency and less inductance in contrast with L and LC filters [5]–[7]. However, the LCL filter is three-order and the system order would be increased. The achievement of the system stability would be more complicated. A dual-loop control is adopted for grid-connected converters due to the system stability and the resonant damping caused by LCL filters [5], [6], [8], since a sole-loop control strategy is not enough to ensure the system stable.

In three-phase AC systems, the grid-connected current control directly uses a PI controller and can not achieve zero steady-state error. Generally, the three-phase current is transformed by the DQ synchronous reference frame from a stationary frame to a rotating frame in order to obtain its active and reactive power current components, which are DC values [9], [10]. The DC with a PI controller can implement zero steady-state error. But the DQ synchronous reference frame is not easy to be applied in single-phase AC systems.

Separate regulation of grid-connected active power and re-

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Fig. 1. System configuration.

active power for single-phase grid-connected converters which is helpful for supplying the power demand of local loads and controlling the power factor (PF) has been seldom reported in literatures at present. In [11] and [12], the DQ transformation methods applied in single-phase grid-tie converters control were proposed and were used to obtain the DC active and reactive power current components from the converters' output current, so that two PI controllers can be used to regulate them without zero steady-state error, respectively. However, both [11] and [12] do not mention these control system transfer functions and stability problems. In this paper, a dual-loop power control strategy for a grid-connected converter with an LCL filter is proposed to implement zero error power control based on PQ theory and its control system transfer function is derived. Besides, the design of the system stability is also studied in detail.

This paper is organized as follows. The configuration of the single-phase grid-connected converter system is described in Section II. The single-phase current detection based on PQ theory is discussed in Section III. Section IV analyzes the dual-loop power control strategy in detail, where the derivations of the control equivalence and the system transfer function are included. What is more, the system stability analysis and the regulation gains design are presented in Section IV. Simulation and experimental results are illustrated in Section V. Some conclusions are given in Section VI.

II. SYSTEM DESCRIPTION

Fig. 1 shows the configuration of the single-phase connected-grid converter system which consists of two parts: main circuit and grid-connected control. The main circuit is composed of a DC bus, a single-phase full-bridge voltage source converter (VSC), and an LCL filter. The DC bus could be served by renewable energy resources, such as photovoltaic, wind generator, storage battery, super capacitor, etc. The VSC is used to convert electricity energy from DC to AC and could be bi-directional. The sinusoidal pulse width modulation is introduced to acquire sinusoidal AC electricity. The LCL filter restrains the high-frequency switching harmonics. The connected-grid control will be discussed in detail in Section IV.

III. SINGLE-PHASE CURRENT DETECTION

This section serves Section IV. Firstly, v_s and i_s are delayed by one quarter of the fundamental period without consider-

TABLE I
COMPONENT PARAMETERS SETUP

Filter inductor L_1	4.5 mH
Inductor internal resistance R_1	0.16 Ω
Filter inductor L_2	2.5 mH
Inductor internal resistance R_2	0.11 Ω
Filter Capacitor C_f	15 μ F
DC bus voltage V_{dc}	400 V
Utility grid voltage RMS V_s	220 V
Utility grid angular frequency ω_0	314 rad/s
Carrier frequency	12.8 kHz

ing the background voltage harmonic, respectively. And the structured orthogonal two-phase voltage and current could be written as

$$\begin{cases} v_{s\alpha} = v_s = \sqrt{2}V_s \sin \omega_0 t \\ v_{s\beta} = -\sqrt{2}V_s \cos \omega_0 t \end{cases}, \quad (1)$$

$$\begin{cases} i_{s\alpha} = i_s = \sqrt{2} \sum_{n=1}^{\infty} I_{sn} \sin(n\omega_0 t - \theta_n) \\ i_{s\beta} = -\sqrt{2} \sum_{n=1}^{\infty} I_{sn} \cos(n\omega_0 t - \theta_n) \end{cases}, \quad (2)$$

where, V_s is the RMS of v_s ; ω_0 is the fundamental angular frequency; I_{sn} is the RMS of the current component of i_s at the angular frequency $n\omega_0$ ($n = 1, 3, 5, \dots$) and I_{s1} is the RMS of the fundamental current component; θ_n is the phase of the current component of i_s at the angular frequency $n\omega_0$.

According to the instantaneous reactive power theory [13], the instantaneous active power p and reactive power q can be calculated with (1) and (2), respectively,

$$\begin{bmatrix} p \\ q \end{bmatrix} = \begin{bmatrix} v_{s\alpha} & v_{s\beta} \\ v_{s\beta} & -v_{s\alpha} \end{bmatrix} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \end{bmatrix} = \begin{bmatrix} 2V_s \sum_{n=1}^{\infty} I_{sn} \cos[(n-1)\omega_0 t - \theta_n] \\ -2V_s \sum_{n=1}^{\infty} I_{sn} \sin[(n-1)\omega_0 t - \theta_n] \end{bmatrix}. \quad (3)$$

From (3), p and q can be decomposed into DC components $\bar{p}$, $\bar{q}$ and AC components $\tilde{p}$, $\tilde{q}$, respectively. $\bar{p}$ and $\bar{q}$ can be acquired from (3) through low pass filters (LPFs). Considering that the harmonic component of i_s is slight, $\tilde{p}$ and $\tilde{q}$ are small and the LPFs could be neglected.

$$\begin{bmatrix} \bar{p} \\ \bar{q} \end{bmatrix} = \begin{bmatrix} 2V_s I_{s1} \cos \theta_1 \\ 2V_s I_{s1} \sin \theta_1 \end{bmatrix} \quad (4)$$

where, $\bar{p}$ and $\bar{q}$ are the mean values of the instantaneous active and reactive powers which correspond to the power exchange between the DC bus and the utility grid. From (4), $\bar{p}$ and $\bar{q}$ are twice of the active and reactive powers in the actual single-phase circuit. Here we use two capital letters P_s and Q_s to represent the grid-connected active and reactive powers, respectively.

$$\begin{bmatrix} P_s \\ Q_s \end{bmatrix} = \frac{1}{2} \begin{bmatrix} \bar{p} \\ \bar{q} \end{bmatrix} = \begin{bmatrix} V_s I_{s1} \cos \theta_1 \\ V_s I_{s1} \sin \theta_1 \end{bmatrix}. \quad (5)$$

From (1) and (5), the instantaneous active power current i_{s1p} , the instantaneous reactive power current i_{s1q} , and har-

$$\begin{cases} I_{sp}^* = \sqrt{2}P_s^*/V_s \\ I_{sq}^* = \sqrt{2}Q_s^*/V_s \end{cases} \quad (11)$$

From (9) and Fig. 3(a), the following equation can be derived.

$$\begin{aligned} \begin{bmatrix} e_{sp} \\ e_{sq} \end{bmatrix} &= \begin{bmatrix} I_{sp}^* \\ I_{sq}^* \end{bmatrix} - \begin{bmatrix} I_{sp} \\ I_{sq} \end{bmatrix} = C_{\alpha\beta-pq} \begin{bmatrix} i_{s\alpha}^* \\ i_{s\beta}^* \end{bmatrix} - C_{\alpha\beta-pq} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \end{bmatrix} \\ &= C_{\alpha\beta-pq} \begin{bmatrix} i_{s\alpha}^* - i_{s\alpha} \\ i_{s\beta}^* - i_{s\beta} \end{bmatrix} = C_{\alpha\beta-pq} \begin{bmatrix} e_{s\alpha} \\ e_{s\beta} \end{bmatrix} \end{aligned} \quad (12)$$

From (12), Fig. 3 (b) is equivalent to (a). The time-domain expression of (12) can be written as

$$\begin{cases} e_{sp}(t) = e_{s\alpha}(t) \sin \omega_0 t - e_{s\beta}(t) \cos \omega_0 t \\ e_{sq}(t) = -e_{s\alpha}(t) \cos \omega_0 t - e_{s\beta}(t) \sin \omega_0 t \end{cases} \quad (13)$$

The time-domain expression of the PI_(s) output v_p , v_q in Fig. 3 (b) can be expressed as

$$\begin{cases} v_p(t) = e_{sp}(t) * g_s(t) \\ v_q(t) = e_{sq}(t) * g_s(t) \end{cases}, \quad (14)$$

where, the symbol $*$ denotes the convolution multiplication and $g_s(t)$ is the time-domain expression of $G_s(s)$.

The references $i_{c\alpha}^*$, $i_{c\beta}^*$ of the current inner loop can be obtained by v_p , v_q through $C_{pq-\alpha\beta}$.

$$\begin{cases} i_{c\alpha}^*(t) = v_p(t) \sin(\omega_0 t) - v_q(t) \cos(\omega_0 t) \\ i_{c\beta}^*(t) = -v_p(t) \cos(\omega_0 t) - v_q(t) \sin(\omega_0 t) \end{cases} \quad (15)$$

The s-domain expression of $i_{c\alpha}^*$, $i_{c\beta}^*$ can be obtained after carrying out the Laplace transform for (15).

$$\begin{cases} I_{c\alpha}^*(s) = \mathcal{L}[v_p(t) \sin \omega_0 t] - \mathcal{L}[v_q(t) \cos \omega_0 t] \\ I_{c\beta}^*(s) = -\mathcal{L}[v_p(t) \cos \omega_0 t] - \mathcal{L}[v_q(t) \sin \omega_0 t] \end{cases}, \quad (16)$$

where, the symbol $\mathcal{L}$ denotes the Laplace operator.

It is well-known that the spectrum of convolution of two signals in time-domain is equal to the product of spectrum of the two signals in s-domain. Substituting (13) into (14) and then taking Laplace transform, the s-domain expression of v_p , v_q can be expressed as

$$\begin{cases} V_p(s) = \mathcal{L}[e_{s\alpha}(t) \sin \omega_0 t - e_{s\beta}(t) \cos \omega_0 t] G_s(s) \\ V_q(s) = \mathcal{L}[-e_{s\alpha}(t) \cos \omega_0 t - e_{s\beta}(t) \sin \omega_0 t] G_s(s) \end{cases} \quad (17)$$

According to the frequency-shifting property of convolution in s-domain, like (18), (19) can be derived by substituting (17) into (16).

$$\begin{cases} \mathcal{L}[f(t) \sin \omega_0 t] = \frac{j}{2} [F(s + j\omega_0) - F(s - j\omega_0)] \\ \mathcal{L}[f(t) \cos \omega_0 t] = \frac{1}{2} [F(s + j\omega_0) + F(s - j\omega_0)] \end{cases} \quad (18)$$

$$\begin{cases} I_{c\alpha}^*(s) = \frac{1}{2} E_{s\alpha}(s) [G_s(s + j\omega_0) + G_s(s - j\omega_0)] \\ \quad - \frac{j}{2} E_{s\beta}(s) [G_s(s + j\omega_0) - G_s(s - j\omega_0)] \\ I_{c\beta}^*(s) = \frac{j}{2} E_{s\alpha}(s) [G_s(s + j\omega_0) - G_s(s - j\omega_0)] \\ \quad + \frac{1}{2} E_{s\beta}(s) [G_s(s + j\omega_0) + G_s(s - j\omega_0)] \end{cases} \quad (19)$$

Substituting (10) into (19), the following equation can be obtained.

$$\begin{bmatrix} I_{c\alpha}^*(s) \\ I_{c\beta}^*(s) \end{bmatrix} = \begin{bmatrix} K_{sp} + \frac{K_{si}s}{s^2 + \omega_0^2} & -\frac{K_{si}\omega_0}{s^2 + \omega_0^2} \\ \frac{K_{si}\omega_0}{s^2 + \omega_0^2} & K_{sp} + \frac{K_{si}s}{s^2 + \omega_0^2} \end{bmatrix} \begin{bmatrix} E_{s\alpha}(s) \\ E_{s\beta}(s) \end{bmatrix} \quad (20)$$

B. Current Inner Loop

There are several variables which can be used as the controlled variable for the inner loop, like i_l , u_l (denotes the voltage across L_1 and R_1), i_c , and v_c . However, only i_c as the controlled variable is useful for damping the resonance caused by the LCL filter and improving the system stability margin [6], [14]. Here the inner loop uses a proportional controller $P_{(c)}$, since the steady-state error of the inner loop does not affect the regulation precision of the outer loop.

C. Controller Gains Design and Stability Analysis

Assume that V_{dc} is constant and the switching frequency is high enough, the VSC could be equivalent to be a proportional element K . And the output of the inner loop could be considered as the voltage v_l at the VSC AC-side. According to (20), the dual-loop control structure in Fig. 2 can be equivalent to this structure of the dual-loop current control shown in Fig. 4. Here $E_{s\beta}(s)$ is thought of zero. The transfer function of a PI controller $G'_s(s)$ takes place of $G_s(s)$ and its expression is written as (21). The s-domain function in Fig. 4 can be derived and expressed as (22). The open-loop and closed-open transfer functions of the grid-side current i_s can be derived as (23) and (24), respectively, since the grid voltage v_s could be regarded as a disturbance and $V_s(s)$ is set as zero.

$$G'_s(s) = K_{sp} + \frac{K_{si}s}{s^2 + \omega_0^2} \quad (21)$$

The component parameters setup in Fig. 1 is listed in TABLE I. Fig. 5 shows the root locus plots of (24) accomplished in MATLAB. The cross type denotes the closed-loop pole and the ring denotes the closed-loop zero. In Fig. 5 (a), the increase of K_{sp} from 0.1 to 0.5 can improve the system stability, since the poles near the imaginary axis move away. When K_{sp} goes beyond 2, however, there are a couple of poles at the right side of the imaginary axis and the system could be unstable. Fig. 5 (b) and (c) show that both the increases of K_{si} and K_{cp} are unhelpful for the system stability. But the larger the outer-loop integral gain K_{si} , the less the system steady-state error, which could be even removed. What is more, the increase of the inner-loop proportional gain K_{cp} can enlarge the system open-loop gain and damp the resonant peak caused by the LCL filter shown in Fig. 6. Amplitude and phase margins are

Fig. 4. Structure diagram of equivalent dual-loop current control.

$$E_s(s)G'_s(s)K_{cp}K = [L_1C_f s^2 + (R_1 + K_{cp})C_f s + 1] V_s(s) + (a_3 s^3 + a_2 s^2 + a_1 s + a_0 s^0) I_s(s), \quad (22)$$

where, $a_3 = L_1 L_2 C_f$; $a_2 = (R_1 + K_{cp}K)L_2 C_f + R_2 L_1 C_f$; $a_1 = L_1 + L_2 + R_1 R_2 C_f + R_2 K_{cp} K C_f$; $a_0 = R_1 + R_2$.

$$G_{so}(s) = \frac{I_s(s)}{E(s)} = \frac{K_{cp}K (K_{sp}s^2 + K_{si}s + K_{sp}\omega_0^2)}{a_3 s^5 + a_2 s^4 + (a_1 + a_3 \omega_0^2) s^3 + (a_0 + a_2 \omega_0^2) s^2 + a_1 \omega_0^2 s + a_0 \omega_0^2}. \quad (23)$$

$$G_{sc}(s) = \frac{G_{so}(s)}{1 + G_{so}(s)} = \frac{K_{cp}K (K_{sp}s^2 + K_{si}s + K_{sp}\omega_0^2)}{a_3 s^5 + a_2 s^4 + (a_1 + a_3 \omega_0^2) s^3 + (a_0 + a_2 \omega_0^2 + K_{cp}K) s^2 + (a_1 \omega_0^2 + K_{cp}K K_{si}) s + (a_0 + K_{cp}K K_{sp}) \omega_0^2}. \quad (24)$$

Fig. 5. Root locus plot of the closed-loop transfer function with different control gains. (a) $0.1 < K_{sp} < \infty$, $K_{si} = 100$, $K_{cp} = 1$, (b) $K_{cp} = 0.5$, $10 < K_{si} < \infty$, $K_{cp} = 1$, (c) $K_{cp} = 0.5$, $K_{si} = 100$, $0.2 < K_{cp} < \infty$.

10 dB and 73° when K_{cp} is set as 0.2, respectively. When K_{cp} is 1, both margins are 10 dB and 39° . The decrease of K_{cp} would increase the phase margin. With the compromise of the system stability and resonant damping, K_{sp} , K_{si} , and K_{cp} are set as 0.5, 100, 1, respectively.

V. SIMULATION AND EXPERIMENT

A. Simulation

The simulation model in Fig. 1 is built in MATLAB/SIMULINK. The component parameters setup is listed in TABLE I. The reference active and reactive powers P_s^* , Q_s^* are decided by real-time power of distributed generators, energy dispatch centers, or local loads. Firstly, P_s^* is set as 1 kW and Q_s^* is zero. The spectrum of the grid-side current is acquired in Fig. 7 (a) and (b) when the inner loop proportional gain K_{cp} is set as 0.02 and 1, respectively. The increase of K_{cp}

Fig. 6. Bode diagram of the open-loop transfer function with three different inner-loop proportional gains at $K_{cp} = 0.5$, $K_{si} = 100$.

can reduce the THD of the grid-side current from 2.72 % to 1.50 % and can damp the resonant peak caused by the LCL filter which is resonant at 1 kHz.

Fig. 8 displays the simulation results in that case the initial

Fig. 7. Spectrum of the grid-side current with different inner loop proportional gains ($K_{cp} = 0.5$, $K_{si} = 100$). (a) $K_{sc} = 0.02$, (b) $K_{cp} = 1$.

values of P_s^* , Q_s^* are set as 1 kW and 500 Var, respectively. And then Q_s^* varies by 1 kVar at 0.2 s and P_s^* is unchanged. Both actual grid-connected active and reactive powers completely track the references in steady-state condition and the coupling between the active power and the reactive power is slight in transient process in Fig. 8(a). And the regulation of the grid-side active and reactive power current is corresponding to the grid-connected active and reactive power displayed in Fig. 8(a) and (b). The dynamic response is good by observing the waveform of the grid-side current at 0.2 s in Fig. 8(c).

B. Experiment

To verify the analysis above, the prototype is built in lab and its photos are shown in Fig. 9. The DC bus is served by a photovoltaic array whose peak power is 2 kWp in Fig. 9(a). The power module PM50B5LA060 is used as the VSC. The proposed control strategy is implemented in TMS320F2812 DSP in Fig. 9(b). The reference reactive power Q_s^* is set as 150 Var, considering that this power converter is regarded as an inductive load from the view of the utility grid and the grid-connected power factor (PF) is affected very lightly. The reference active power P_s^* is mainly decided by the illumination intensity and the ambient temperature which are uncertain in experiment.

The experimental results are shown in Fig. 10 and Fig. 11. The grid-connected reactive power is separately controllable in Fig. 10(a) and Fig. 11(b). And both PFs are also ensured beyond 0.97. The THDs of the grid-side current in two conditions are below 5 % and are in accordance with IEEE Standard.

Fig. 8. Simulation results. (a) Grid-connected power tracking effect, (b) grid-connected active and reactive power current tracking effect, (c) waveforms of the utility grid and the grid-side current.

Fig. 9. Prototype photos. (a) Photovoltaic array, (b) control system.

(a)

(b)

(c)

Fig. 10. Experimental Results when the actual power is beyond half of the normal rated power. (a) Power information, (b) waveforms of the utility grid voltage and grid-side current, (c) the grid-side current spectrum.

VI. CONCLUSIONS

This paper proposes the dual-loop power control strategy for single-phase grid-connected converters. The design of the proposed control system is studied in detail. The influence of the controller gains on the system stability is analyzed with the root locus of the system closed-loop transfer function. The outer loop proportional gain plays a more important role in the system stability than the other two regulation gains. The increase of the inner proportional gain can damp the resonant peak caused by the LCL filter and is useful for reducing the THD of the grid-side current. The validity of the analysis is confirmed by simulation and experimental results. What is more, the grid-connected active and reactive power can be regulated separately. This capability is helpful for realizing multifunctional converters in renewable energy resources grid-connected generation, such as reactive power compensation and power factor correction. And the capability is also valuable to the single-phase microgrid operation.

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(a)

(b)

(c)

Fig. 11. Experimental Results when the actual power is near a quarter of the normal rated power. (a) Power information, (b) waveforms of the utility grid voltage and grid-side current, (c) the grid-side current spectrum.

REFERENCES

- [1] J. M. Carrasco, L. G. Franquelo, J. T. Bialasiewicz, E. Galvan, R.C. P. Guisado, M. A. M. Prats, J. I. Leon, and N. Moreno-Alfonso, "Power-electronic for the grid integration of renewable energy sources: a survey," *IEEE Transactions on Industrial Electronics*, Vol. 53, No. 4, pp. 1002-1006, Aug. 2006.
- [2] A. A. Ferreira, J. A. Pomilio, G. Spiazzi, and L. de Araujo Silva, "Energy management fuzzy logic supervisory for electric vehicle power supplies system," *IEEE Transactions on Power Electronics*, Vol. 23, No. 1, pp. 107-115, Jan. 2008.
- [3] F. -S. Pai and S. -J. Huang, "Design and operation of power converter for microturbine powered distributed generator with capacity expansion capability," *IEEE Trans. Energy Convers.*, Vol. 23, No. 1, pp. 110-118, Mar. 2008.
- [4] L. Zhang, X. Yang, W. Chen, and X. Yao, "An isolated soft-switching bidirectional buck-boost inverter for fuel cell applications," *Journal of Power Electronics*, Vol. 10, No. 3, pp. 235-244, May 2010.
- [5] E. Twining and D. G. Holmes, "Grid current regulation of a three-phase voltage source inverter with an LCL input filter," *IEEE Trans. Power Electron.*, Vol. 18, No. 3, pp. 888-895, May 2003.
- [6] P. C. Loh and D. G. Holmes, "Analysis of multiloop control strategies for LC/LCL-filtered voltage-source and current-source inverters," *IEEE Trans. Ind. Electron.*, Vol. 41, No. 2, pp. 644-654, Mar. /Apr. 2005.
- [7] J. Dannehl, F. W. Fuchs, and P. B. Thøgersen, "PI state space current control of grid-connected PWM converters with LCL filters," *IEEE Trans. Power Electron.*, Vol. 25, No. 9, pp. 2320-2330, Sep. 2010.
- [8] F. Liu, Y. Zhou, S. Duan, J. Yin, B. Liu, and F. Liu, "Parameter design of a two-current-loop controller used in a grid-connected inverter system with LCL filter," *IEEE Trans. Ind. Electron.*, Vol. 56, No. 11, pp. 4483-4491, Nov. 2009.
- [9] Yun Wei Li, D. M. Viathgamuwa, and P. C. Loh, "A grid-interfacing power quality compensator for three-phase three-wire microgrid applications," *IEEE Trans. Power Electron.*, Vol. 21, No. 4, pp. 1021-1031, Jul. 2006.

- [10] P. Rodriguez, J. Pou, J. Bergas, J. I. Candela, R. P. Burgos, and D. Boroyevich, "Decoupled double synchronous reference frame PLL for power converter control," *IEEE Trans. on Power Electron.*, Vol. 22, No. 2, pp. 584-592, Mar. 2007.
- [11] U. A. Miranda, M. Aredes, and L. G. B. Rolim, "A DQ synchronous reference frame control for single-phase converters," *IEEE 36th Power Electronics Specialists Conference*, pp. 1377-1381, 2005.
- [12] B. Crowhurst, E. F. El-Saadany, L. El Chaar, and L. A. Lamont, "Single-phase grid-tie inverter control using DQ transform for active and reactive load power compensation," *IEEE International Conference on Power and Energy*, pp. 489-494, 2010.
- [13] H. Akagi, Y. Kanazawa, and A. Nabae, "Instantaneous reactive power compensators comprising switching devices without energy storage components," *IEEE Trans. Ind. Appl.*, Vol. IA-20, No. 3, pp. 625-630, May/Jun. 1984.
- [14] J. Dannehl, F. W. Fuchs, S. Hansen, and P. B. Thøgersen, "Investigation of active damping approaches for PI-based current control of grid-connected pulse width modulation converters with LCL filters," *IEEE Trans. Ind. Appl.*, Vol. 46, No. 4, pp. 1509-1517, Jul./Aug. 2010.

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