A Current Decoupling Control Scheme for LCL-Type Single-Phase Grid-Connected Converter

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This work was supported by the Guangdong Natural Science Foundation under Grant 2015a030313478.

ABSTRACT For better performance, different current control schemes are used in LCL-type grid-connected converters. However, these schemes are mostly used in three-phase grid-connected converters and multiple current sensors are needed. To address these issues, a current decoupling control scheme for single-phase grid-connected converters in rotating reference frame is proposed in the paper. Because an extended state observer is adopted in the control scheme, only one current sensor is needed and the complicated coupling relationships between the $d$ axis components and $q$ axis components can be eliminated completely. Moreover, to improve the dynamic and steady-state performance of converters, a state feedback closed-loop system structure and nonlinear controllers are adopted. The current decoupling control scheme is validated in an LCL-type single-phase grid-connected converter. The experimental results show that a fast-dynamic response under load change and voltage sag can be achieved besides preferable static performances with the proposed control scheme.

INDEX TERMS Current decoupling control, grid-connected inverter, LCL, state feedback.

I. INTRODUCTION
Compared with the L-type filters, LCL-type filters have many advantages, such as better filtering performance, smaller size and lower cost and LCL filters have been widely used in grid-connected converters [1]–[3]. Because LC-type filters belong to the higher-order filter, it is necessary to find an effective control scheme to reduce its impact on grid-connected current and ensure the stability of the control system [4]–[10]. Hence, different control schemes are presented. These control schemes can be categorized into four major classes, namely, proportional-resonant (PR) control schemes [6], [7], direct power control (DPC) schemes [11], [12], current control schemes [13], [14] and active damping control [15], [16]. Although a sinusoidal reference signal can effectively be tracked with PR control schemes and the coupling problem caused by coordinate transformation is avoided, the independent control of active current and reactive current cannot be realized and it is difficult to satisfy the reactive power requirement for voltage support when the fault occurs in the power grid. Since the performance of single-phase converters with DPC schemes is deeply related to the accuracy of the model of converters, the application of DPC schemes is limited in grid-connected converters. Active damping control schemes are proposed to suppress the resonant peak caused by LCL-type filter. Because active damping control schemes belong to virtual impedance control schemes, the value of virtual impedance is greatly influenced by the coupling terms between the $d$ axis components and $q$ axis components. In addition, most of active damping control schemes require at least two current sensors. Therefore, current control schemes are chosen in most applications. Current control schemes are well-known efficient control strategies for power converters. To obtain a better control effect, most of the current control schemes are implemented in a rotating reference frame. In the rotating reference frame, there is a strong coupling between $d$ axis components and $q$ axis components, which seriously affects the performances of LCL-type grid-connected converters [17], [18]. If the coupling terms are ignored as external disturbances, the independent control of the $d$ axis components and $q$ axis components can be realized in traditional current control schemes. However, this causes modeling distortion and reduces the performances of grid-current. In [19], by analyzing the transfer function of grid-connected converters, coupling compensation terms are introduced in a current control scheme. Compared with
ignoring the coupling term, this control scheme can partly improve the accuracy of the model and ensure the quality of grid current, but it cannot completely eliminate the coupling impacts between the $d$ axis components and $q$ axis components and at least two current sensors are required. Additionally, to decrease the number of current sensors, a control strategy is proposed in [14] and the signal errors between the grid current and its feedback value are controlled by the PI controller. Because the coupling terms between the $d$ axis components and $q$ axis components are not eliminated, there is a relatively slow dynamic response in converters and the overshoot of grid current is not restrained efficiently. Unfortunately, almost all of these studies are focus on three-phase grid-connected converters.

In order to extend current decoupling control scheme to single-phase grid-connected converters, eliminate the coupling terms between the $d$ axis components and $q$ axis components and decrease the number of current sensors, a decoupling control scheme with one current sensor for single-phase grid-connected converters with LCL filter is proposed in the paper. Through analysis of the relations between the $d$ axis components and $q$ axis components, a decoupling method for converters is presented. With extended state observers, tracking differentiators and feedback controllers, the independent control of the following, the first-order differential and the second-order differential signal of grid current is achieved and only one current sensor is needed in the control scheme. The simulation and experimental results confirm the excellent performance of the proposed scheme. Theoretical evaluations are verified through a performance comparison between the proposed control scheme and the other three control schemes.

II. MODEL IN ROTATING REFERENCE FRAME

A single-phase LCL-type single-phase grid-connected converter is depicted in Fig. 1. The converter consists of an LCL filter and an H-bridge. The H-bridge is made of four active switches $S_1$-$S_4$ and is designed to generate an alternating voltage $u_d$. $L_1$, $L_2$, $R$, $C$, and $C_{dc}$ are the inverter-side inductor, the grid-side inductor, the grid-side resistance, the filter capacitor at AC-side and the DC-link capacitance, respectively. $i_d$, $i_q$, and $i_f$ are the inverter-side current, the grid current, and the current through the capacitor $C$. $u_d$, $u_q$, and $U_{dc}$ are the grid voltage, the voltage across the capacitor $C$ and the DC-link voltage.

Usually, grid-connected converters are located near the grid, for convenience, the grid inductance and the grid resistance are integrated with $L_2$ and $R$ in the paper.

From FIGURE 1, four equations are required to completely describe the state of the converter:

$$\begin{align*}
\frac{di_d}{dt} &= u_d/L_1 - u_c/L_1 \\
\frac{di_q}{dt} &= u_q/L_1 - \omega i_d - u_{cq}/L_1 \\
\frac{di_f}{dt} &= i_c/L_f \\
i_1 &= i_d + i_q
\end{align*} \tag{1}$$

FIGURE 1. Configuration of the single-phase LCL-type grid-connected converter.

FIGURE 2. Block diagram of the converter in the rotating reference frame.

By generating a fictitious orthogonal current of grid current and a fictitious orthogonal voltage of grid voltage, the following equations are obtained to describe the state of the converter in the rotating reference frame:

$$\begin{align*}
\frac{di_{1d}}{dt} &= u_{1d}/L_1 + \omega i_{1q} - u_{cd}/L_1 \\
\frac{di_{1q}}{dt} &= u_{1q}/L_1 - \omega i_{1d} - u_{cq}/L_1 \\
\frac{di_{2d}}{dt} &= u_{cd}/L_2 + \omega i_{2q} - R i_{2d}/L_2 - u_{sd}/L_2 \\
\frac{di_{2q}}{dt} &= u_{cq}/L_2 - \omega i_{2d} - R i_{2q}/L_2 - u_{sq}/L_2 \\
\frac{du_{cd}}{dt} &= i_{cd}/C + \omega u_{eq} \\
\frac{du_{cq}}{dt} &= i_{cq}/C - \omega u_{ed} \\
i_{1d} &= i_{1d} + i_{2d} \\
i_{1q} &= i_{1q} + i_{2q}
\end{align*} \tag{2}$$

where $i_{1d}$, $i_{2d}$, $i_{1q}$, $i_{2q}$, $u_{cd}$, $u_{sd}$ and $u_{rd}$ are the $d$ axis component of $i_1$, $i_2$, $i_c$, $u_d$ and $u_r$ respectively. $i_{1q}$, $i_{2q}$, $i_{eq}$, $u_{eq}$, $u_{cq}$ and $u_{eq}$ are the $q$ axis component of $i_1$, $i_2$, $i_c$, $u_d$ and $u_r$. $\omega$ is the angle frequency of grid voltage.

According to equations (2), the block diagram of the LCL-type single-phase grid-connected converter in the rotating reference frame is given in FIGURE 2.

As shown in FIGURE 2, there are complicated coupling relationships between the $d$ axis components and $q$ axis components in the rotating reference frame.

III. DECOUPLING METHOD

The principle of current decoupling is demonstrated using the inductor $L_1$ as an example.

From equation (2), two differentiating equations are used to describe the state of the inductor $L_1$, as follow:

$$\begin{align*}
u_{rd} - u_{cd} &= L_1 \frac{di_{1d}}{dt} - \omega L_1 i_{1q} \\
u_{rq} - u_{eq} &= L_1 \frac{di_{1q}}{dt} + \omega L_1 i_{1d}
\end{align*} \tag{3}$$
After some manipulations, the state equations of equation (3) are obtained as

\[
\begin{align*}
\mathbf{x} &= \mathbf{Ax} + \mathbf{Bu} \\
\mathbf{y} &= \mathbf{Cx}
\end{align*}
\]

(4)

where \(\mathbf{A} = \begin{bmatrix} 0 & \omega \\ \omega & 0 \end{bmatrix}\), \(\mathbf{B} = \begin{bmatrix} 1/L_1 & 0 \\ 0 & 1/L_1 \end{bmatrix}\), \(\mathbf{C} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}\), \(\mathbf{x} = [i_{1d} \ i_{1q}]^T\), \(\mathbf{y} = [i_{1d} \ i_{1q}]^T\), \(\mathbf{u} = [u_{rd} - u_{cd} \ u_{rq} - u_{cq}]^T\).

The transfer function matrix of equation (4) is given below:

\[
\mathbf{G}(s) = \frac{\mathbf{y}(s)}{\mathbf{u}(s)} = \begin{bmatrix} \frac{s}{L_1(s^2 + \omega^2)} & \frac{s}{L_1(s^2 + \omega^2)} \\ \frac{s}{L_1(s^2 + \omega^2)} & \frac{s}{L_1(s^2 + \omega^2)} \end{bmatrix}
\]

(5)

The transfer function matrix shown in equation (5) is not a diagonal matrix, which indicates that the converter is a coupling system. Fortunately, the coupling terms for inductor \(L_1\) between the \(d\) axis components and \(q\) axis components can be decoupled by an introduction of the feedback matrix. When the feedback matrix \(\mathbf{H}\) is introduced, the transfer function matrix of equation (3) is modified as:

\[
\mathbf{G}'(s) = \mathbf{C}(\mathbf{sI} - \mathbf{A} + \mathbf{BH})^{-1}\mathbf{B}
\]

(6)

where \(\mathbf{I}\) is an identity matrix.

To make sure \(\mathbf{G}'(s)\) is a diagonal matrix, the feedback matrix \(\mathbf{H}\) is given by:

\[
\mathbf{H} = \begin{bmatrix} 0 & \omega L_1 \\
-\omega L_1 & 0 \end{bmatrix}
\]

(7)

After some substitutions and manipulations, the transfer function matrix for inductor \(L_1\) after decoupling is obtained as

\[
\mathbf{G}(s) = \frac{\mathbf{y}(s)}{\mathbf{u}(s)} = \begin{bmatrix} \frac{1/(sL_1)}{1/(sL_1)} & 0 \\ 0 & \frac{1/(sL_1)}{1/(sL_1)} \end{bmatrix}
\]

(8)

where \(\mathbf{u}' = [u_{rd} - u_{cd} - \omega L_1 i_{1q} \ u_{rq} - u_{cq} + \omega L_1 i_{1d}]^T\).

If the same decoupling strategy is adopted for \(L_2\) and \(C\), the coupling terms between the \(d\) axis components and \(q\) axis components can be eliminated completely. The procedure of the decoupling strategy is demonstrated in FIGURE 3.

As shown in FIGURE 3, the perfect elimination of these undesired coupling terms can be achieved by injecting decoupling terms with the same amplitude and opposite phase angles into \(i_{1d}, i_{2d}, u_{cd}\) and \(i_{1q}, i_{2q}, u_{cq}\). The model of the inverter after decoupling is shown in FIGURE 4 and the transfer function matrix of the inverter after decoupling are given by

\[
\mathbf{G}_{\text{INV}}(s) = \begin{bmatrix} \frac{1}{b_3 s^3 + b_2 s^2 + b_1 s + b_0} & 0 \\ 0 & \frac{1}{b_3 s^3 + b_2 s^2 + b_1 s + b_0} \end{bmatrix}
\]

(9)

where \(b_3 = CL_1 L_2, b_2 = CRL_1, b_1 = L_1 + L_2, b_0 = R\).

The matrix shown in equation (9) is a diagonal matrix, which indicates that the control system of the converter has been decoupled.

It is noted that there is an inevitable deviation between the actual values of LCL filter and the calculated values used in FIGURE 3. It is necessary to analyze the influence of the deviation on decoupling performance. Take the inductor \(L_1\) for example, if the calculated values of \(L_1\) is set to \(\lambda L_1\), the feedback matrix \(\mathbf{H}\) is modified as:

\[
\mathbf{H}' = \begin{bmatrix} 0 & \lambda \omega L_1 \\
-\lambda \omega L_1 & 0 \end{bmatrix}
\]

(10)

The modulus of the transfer function between the \(d\) axis components and \(q\) axis components is obtained from equations (4) and (10):

\[
|\mathbf{G}_{\text{couple}}(s)| = \left|\frac{1 - \lambda}{L_1[\omega^2(1 - \lambda^2) - \omega^2]}\right|
\]

(11)

The curve between the modulus of \(\omega L_2 \ G_{\text{couple}}(s)\) and \(\lambda\) is shown in FIGURE 5.

As shown in FIGURE 5, the performance of the proposed decoupling strategy is associated with the value of \(\lambda\). The closer the value of \(\lambda\) is to 1, the better the performance of decoupling. Because the actual values of the LCL filter are closed to that used in the decoupling strategy, most of the coupling terms between the \(d\) axis components and \(q\) axis components are eliminated. For example, when \(\lambda\) is in (0.9,1.1), the modulus of \(\omega L_2 \ G_{\text{couple}}(s)\) is in (0,0.1).

By simplifying equation (2), the inverter-side current \(i_1\) and the voltage across capacitor \(C\) are calculated with the grid current \(i_2\), as follow:

\[
\begin{align*}
i_1 &= L_2 C d^2 i_2/d^2 t + CRdi_2/dt + i_2 \\
u_c &= L_2 di_2/dt + Ri_2
\end{align*}
\]

(12)
This shows that the injecting decoupling terms can be obtained only with the following, the first-order differential and second-order differential signal of grid current. If these compensations are move to the input terminal of inductor $L_1$, the equivalent block diagram of the decoupling strategy shown in FIGURE 3 can be obtained. The equivalent block diagram of current decoupling strategy for $d$ axis components is given in FIGURE 6.

$$u_{rsd} = -(1 - L_1 C \omega)u_{sd}$$
$$u_{rsq} = -(1 - L_1 C \omega)u_{sq}$$

(13)

Because the value of $L_1 C \omega$ is almost equal to zero, the equation (7) is simplified as

$$u_{rsd} = u_{sd}$$
$$u_{rsq} = u_{sq}$$

(14)

The block diagrams of the grid voltage feed-forward compensation are demonstrated in FIGURE 7.

**FIGURE 5.** The curve between modulus of $\omega L_2 G_{couple}(s)$ and $\lambda$.

**FIGURE 6.** Equivalent block diagram of decoupling strategy for the $d$ axis components.

**FIGURE 7.** Block diagram of the grid voltage feed-forward compensation, (a) $d$ axis, (b) $q$ axis.

**FIGURE 8.** Structure of the proposed control scheme.

**IV. CURRENT DECOUPLING CONTROL SCHEME**

Based on the decoupling method, a current decoupling control scheme is developed in the paper. The detailed structure of the control scheme is demonstrated in FIGURE 8.

As shown in FIGURE 8, the control scheme consists of six parts: $d$-$q$ transformation module, phase-locked loop (PLL) module, extended state observer (ESO) module, controller, tracking differentiator (TD) module and an SVPWM module.

The transform matrices in $dq$ transformation module are given in equations. (15) and (16).

$$\begin{bmatrix} d \\ q \end{bmatrix} = \begin{bmatrix} \cos \theta & \sin \theta \\ -\sin \theta & \cos \theta \end{bmatrix} \begin{bmatrix} \alpha \\ \beta \end{bmatrix}$$

(15)

$$\begin{bmatrix} \alpha \\ \beta \end{bmatrix} = \begin{bmatrix} \cos \theta & -\sin \theta \\ \sin \theta & \cos \theta \end{bmatrix} \begin{bmatrix} d \\ q \end{bmatrix}$$

(16)

The PLL is used to obtain the phase angle $\theta$ and the angular frequency $\omega$ of the single-phase system and the input signal
of the PLL is an output signal of a voltage sensor that is used to acquire the voltage of the grid. There are many kinds of PLL, such as EPLL, Hilbert, all-pass filter, and second-order generalized integrator (SOGI) [12], [20], [21]. The SOGI is an advanced and popular phase-locked loop technique and the characteristic transfer function of SOGI in s-domain can be described as

\[
\begin{aligned}
    x_\alpha(s) &= \frac{\xi \omega s}{s^2 + \xi \omega + \omega^2} x(s) \\
    x_\delta(s) &= \frac{\omega^2}{s^2 + \xi \omega + \omega^2} x(s)
\end{aligned}
\]  

where \( \xi \) is the damping factor, the outstanding performance of SOGI depends on the selection of damping factor \( \xi \). The SOGI can provide a satisfactory harmonic filtering capability. Hence, the SOGI can effectively extract the phase angle and the angular frequency of the single-phase system from the original signals with harmonics, so it is adopted in the proposed control scheme.

The ESO module is used to obtain the estimation of the following, the first-order differential and second-order differential signal of the grid current quickly and accurately. The structure of the ESO module is proposed by Han [22] and the ESO are effectual to restrain high-frequency noise and accurately. During the whole observation process, The ESO shows a fast and stable performance without static error. Hence, the SOGI can effectively extract the phase angle and the angular frequency of the single-phase system from the original signals with harmonics, so it is adopted in the proposed control scheme.

The ESO module is used to obtain the estimation of the following, the first-order differential and second-order differential signal of the grid current quickly and accurately. The structure of the ESO module is proposed by Han [22] and the ESO are effectual to restrain high-frequency noise signal. Hence, the ESO has been successfully applied in many fields [23]–[25]. According to the structure of transfer function of converters, an ESO for \( d \) axis current component is constructed and the expression of the ESO is given by:

\[
\begin{aligned}
    \xi_d &= x_{1d} - i_{2d} \\
    \dot{x}_{1d} &= x_{2d} - \alpha_1 \epsilon_{1d}^{1/3} \\
    \dot{x}_{2d} &= x_{3d} - \alpha_2 \epsilon_{2d}^{1/3} \\
    \dot{x}_{3d} &= -\alpha_3 \epsilon_{3d}^{1/3}
\end{aligned}
\]

where \( x_{1d}, x_{2d} \) and \( x_{3d} \) are the estimations of the following, the first-order differential and second-order differential signal of grid current, respectively. \( \alpha_1 - \alpha_4 \) are the correction coefficients.

The TD module is used to obtain the following, the first-order differential and second-order differential signal of current reference. The expression of TD for \( d \) axis component is given by

\[
\begin{aligned}
    \dot{v}_{1d} &= v_{2d} \\
    \dot{v}_{2d} &= -r \times \text{sgn}(v_{1d} - i_{2d, \text{ref}} + v_{2d} |v_{2d}| / 2r)
\end{aligned}
\]

where \( r \) is a characteristic parameter reflecting the variation feature of TD, \( i_{2d, \text{ref}} \) is the input of TD, \( v_{1d} \) and \( v_{2d} \) are the output following signal and the output first-order differential signal of \( i_{2d, \text{ref}} \), \( \text{sgn} \) stands for sign function.

Because the classical PID controller is composed of the linear sum of the proportion of error, the integral of error and the differential of error, this inevitably brings the contradiction in speediness and overshoot. To address this issue, a nonlinear controller based on the state errors of following, the first order differential and second-order differential signal of grid current is used. Since the simple linear weighted sum form in the classical PID controller is replaced, not only the efficiency of information processing but also the control performances are improved. Hence, if the three signals could be controlled independently, the grid current, the voltage across the capacitor \( C \) and the inverter-side current are controlled effectively. To get better performance, the controller is designed to be high control gain when the error is small and to be low control gain when the error is large. The expression of the controller is given by

\[
y(t) = m_1 \text{fal}(\epsilon_1, a, \delta) + m_2 \text{fal}(\epsilon_2, a, \delta) + m_3 \text{fal}(\epsilon_3, a, \delta)
\]

where \( \text{fal}(\epsilon, a, b) = \begin{cases} |\epsilon|^a \text{sgn}(\epsilon) & |\epsilon| > \delta \\ \epsilon / \delta^{1-a} & |\epsilon| \leq \delta \end{cases} \). \( \epsilon_i \) (i = 1, 2, 3) are the three errors between the reference value and the output of the state observer. \( a \) is a constant in the interval [0, 1], \( \delta \) is a constant that affects the width of the linear interval. \( m_1, m_2, m_3 \) are function control gains. The output characteristic of the function \( \text{fal}(\epsilon, a, b) \) is shown in FIGURE 9.

Furthermore, the driving signals for S1-S4 are generated by a single-phase SVPWM module. Because the voltage of the DC-link capacitance must be used in the SVPWM module, another voltage sensor is used to acquire the voltage of the DC-link capacitance.

V. SIMULATION AND EXPERIMENT RESULTS

A. SIMULATION RESULTS

1) PERFORMANCES OF ESO.

The dynamic response of the estimation of the following, the first-order differential and second-order differential signal the for \( d \) axis component of grid current in FIGURE 10 (a), (b) and (c), respectively.

\[
\text{FIGURE 9. Output characteristic of the function } \text{fal}(\epsilon, a, b).
\]

It is shown in FIGURE 10 (a), the ESO has good estimation performance and the following signal of grid current reference obtained by TD can almost be reproduced. The simulation results in FIGURE 8 (b) and (c) show that the first-order differential signal and the second-order differential signal of grid current reference obtained by TD can be tracked quickly and accurately. During the whole observation process, The ESO shows a fast and stable performance without static error. Although there is a slight overshoot in the estimations of the first-order differential signal and the second-order differential signal, the two estimations can still track the reference signal in one working period. Hence, the ESO can make the control
FIGURE 10. Dynamic response of the estimation of the following, the first-order differential and second-order differential signal of grid current, (a) the following signal, (b) the first-order differential signal, (c) the second-order differential signal.

system of the inverter have good adaptability and robustness, and the performances of the inverter can be improved.

2) PERFORMANCES OF TD.

To verify the performance of the TD, a square wave signal is used and defined it as follows:

\[
v_d = \begin{cases} 
1 & 0.02N < t < 0.01 + 0.02N \\
0 & 0.01 + 0.02N \leq t < 0.02(N + 1) 
\end{cases}
\]  

(21)

where \( N \) is a positive integer.

The dynamic response of TD is given in FIGURE 11 and the bode diagram of the TD is shown in FIGURE 12. In FIGURE 11, \( v_{1d} \), \( v_{2d} \), and \( v_{3d} \) are the following, the first-order differential and second-order differential signal of the reference signal. From FIGURE 11, it can be seen that the TD not only can track the reference signal quickly and without overshoot but also the first-order and second-order differential signals of the reference signal are obtained simultaneously.

From the bode diagram of the TD, it can be concluded that the frequency characteristic of TD is similar to that of the band-pass filter and the high-frequency noise signals effectively are suppressed. This means that the differential signal distortion of the input signal is settled.

3) PERFORMANCES OF THE CONTROLLER.

In order to demonstrate the performance of the proposed ICC scheme, a simulation model of a single-phase converter is built in Matlab/Simulink and the nominal parameters of the simulation model are shown in Table 1. The simulation results for the time-domain response of the proposed scheme are demonstrated in FIGURE 13.

The results of FIGURE 13 show the time-domain response of the proposed control scheme under 20% voltage sag at 0.085 s, 25% voltage swell at 0.195 s, 10% third harmonic.
FIGURE 13. Simulation results for the response of the proposed control scheme under voltage sag, voltage swell, third harmonic voltage injection and grid current reference step-change events.

FIGURE 14. THD of the grid current, (a) without third harmonic voltage injection, (b) with 10% third harmonic voltage injection.

voltage injection at 0.28s and a grid current reference step change at 0.5s. From FIGURE 13, it can be seen that the steady-state waveform of the grid current is both smooth and have no shock with the proposed controller. The simulation results show a fast dynamic response under voltage sag, voltage swell and grid current reference step change can be achieved in single-phase grid-connected converters with the proposed control scheme.

The THDs of the grid current with the proposed control scheme are given in FIGURE 14. From FIGURE 14, it can be seen that the THD of grid current is lower with and without third harmonic voltage injection. Furthermore, it is found that the third harmonic voltage injection had almost no effect on grid current from the THD of grid current. It is implied that the proposed control scheme has better immunity to the harmonic voltage of grid voltage.

B. EXPERIMENT RESULTS

An experimental prototype is designed to demonstrate the performance of the proposed control scheme in the laboratory. In the experimental prototype, an LCL-type single-phase converter is designed and the converter consists of an H-bridge power circuit, sampling circuit, signal processing circuit, gate signal driving circuit and protection circuit. The power switches in the H-bridge power circuit are International Rectifier IRG4PH50UD. A Honeywell CSNE151 Hall current sensors and two voltage differential operational amplifier circuits are used in the sampling circuit. The photo of the experimental prototype is shown in FIGURE 15 and the nominal parameters of the experimental prototype are shown in Table 2. In the experimental prototype, a Texas Instruments TMS320F28035 microcontroller is used as a control chip and a Tektronix MDO3014 scope and a Tektronix A622 current probe are used to measure and analyze the experiment waveforms.

In order to verify the performance of the proposed control scheme, a traditional PI control scheme [26], a PI control scheme with capacitor-current-feedback active damping [27] and a PR control scheme [28] with capacitor-current-feedback active damping are used as comparison schemes. The two active damping control schemes with capacitor-current-feedback are chosen because some active damping control schemes are too complicated to implement on the experimental prototype and the performances of LCL type grid-connect inverters with these active damping control schemes are similar to those with control schemes proposed in [26] and [27]. The PI controller parameters are the same in the two PI control schemes, and the proportional parameter of PR controller and PI controller are set to the same value.

FIGURE 16 exhibits the experimental waveforms of the grid voltage and the phase current in steady-state scenarios when the four control schemes are used.

Because the resonance of LCL filters is not well damped for PI control scheme without active damping, the grid voltage for the control scheme is set to 30 V in the experiment. As seen in FIGURE 16 that the steady-state waveforms with the later three control schemes are very similar except the...
TABLE 3. Performance comparisons of the four control schemes in the nominal power.

| Control scheme                              | THD   | Power factor |
|---------------------------------------------|-------|--------------|
| PI control scheme without active damping    | 20.9% | 0.970        |
| PI control scheme with capacitor-current-feedback active damping | 3.32% | 0.996        |
| PR control scheme with capacitor-current-feedback active damping | 4.49% | 0.996        |
| Proposed control scheme                     | 3.59% | 0.998        |

FIGURE 16. Experimental waveforms of the grid voltage and the line current in steady-state scenarios when different control schemes are used, (a) Traditional PI control scheme, (b) PI control scheme with capacitor-current-feedback active damping, (c) PR control scheme with capacitor-current-feedback active damping, (d) proposed control scheme.

As shown in Table 3, a unity power factor can be achieved with the proposed control scheme. The THD of the grid current with the proposed control scheme is slightly higher than that under PI control scheme with capacitor-current-feedback active damping. The main reason for this is that the compensating coupling terms are formed by the value generated by ESO rather than the actual measured values. Because the same value of the proportional parameter is used in PR controller and PI controller and the value is not the optimal parameter value for PR controller, the THD of grid current with the PR control scheme is higher than that under PI control scheme with capacitor-current-feedback active damping and the proposed control scheme.

TABLE 4. The transition times of the four control schemes under 100%.

| Control scheme                              | TRANSITION TIME |
|---------------------------------------------|-----------------|
| PI control scheme without active damping    | about 10 ms     |
| PI control scheme with capacitor-current-feedback active damping | about 10 ms    |
| PR control scheme with capacitor-current-feedback active damping | about 10 ms    |
| Proposed control scheme                     | about 5 ms      |

FIGURE 17. Experimental results for the dynamical response of the four schemes under 100% load change, (a) Traditional PI control scheme, (b) PI control scheme with capacitor-current-feedback active damping, (c) PR control scheme with capacitor-current-feedback active damping, (d) proposed control scheme.

FIGURE 18. Experimental results for the dynamical response of the four schemes under 20% voltage sag, (a) PI control scheme without active damping, (b) PI control scheme with capacitor-current-feedback active damping, (c) PR control scheme with capacitor-current-feedback active damping, (d) proposed control scheme.

The dynamical responses of the four control schemes have also been experimentally verified under 100% load change (from 50% nominal power to 100% nominal power) and 20% voltage sag (from 100% nominal voltage to 80% nominal voltage). The experimental waves are shown in FIGURE 17 and FIGURE 18, respectively. The transition times of the converter under load change are listed in Table 4.

As shown in Table 4, the transition time of the converter with the proposed control scheme is about 5 ms (1/4 of the period of grid voltage) and the transition time of the converter with the other three control schemes is about 10 ms (1/2 of the period of grid voltage) under load change and voltage sag. This is because the current decoupling control is achieved in the proposed schemes.

As shown in FIGURE 18, the dynamical responses of the latter three schemes are very similar. This is a result of using similar grid voltage feed-forward compensation.

VI. CONCLUSION

A current decoupling control scheme based on extended state observer module, nonlinear controller and tracking differentiator module has been proposed, analyzed, and experimentally realized. Unlike current decoupling control
schemes for three-phase converters, the control scheme proposed in the paper is designed for a single-phase grid-connected converter. In particular, the coupling terms between the $d$ axis components and $q$ axis components in rotation reference frame can be eliminated completely and only one current sensor is used in the control scheme.

The control scheme has been realized and successfully tested in an LCL-type single-phase grid-connected converter. The THD of the grid current and power factor of the converter with the proposed control scheme and the other three control schemes are measured, pointing out a unity power factor can be achieved and The THD is kept low with one current sensor. The dynamical response under 100% load change pointing out the transition time of the converter with the proposed control scheme is about 5 ms, which is about 1/2 the time with other control schemes.

By the analysis of the deviation between the actual values of the LCL filter and the calculated values and the fluctuation of grid voltage, an equivalent block diagram of decoupling control scheme is about 5 ms, which is about 1/2 the time with other control schemes.

The THD of the grid current and power factor of the converter with the proposed control scheme have been numerically implemented by the Simulink tools of MATLAB, obtaining simulation results in good agreement with the corresponding experiments.

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