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LCL-Filters for Grid-Connected Inverters

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ABSTRACT: This paper examines the damping characteristics for LCL-filters for three-phase voltage source inverters (VSIs). In particular, it is found that there is an inherent damping term embedded in the feedback loop when converter current is used for implementing closed-loop control. This extra damping term can surely kill the resonance presented by LCL-filters. Therefore, in this paper characteristic of LCL filters will be discussed. Since maintaining a good power quality is important for the reliable operation of the system and loads, these filters will be applied to a three-phase grid system. Then, the inverter output will be filtered in order to obtain low voltage and current distortion. Theoretical analysis is then presented and lead to a general design guideline, which suggests a way of choosing the values of grid- and converter-side inductors, so that optimum damping can be naturally achieved by solely using converter current control. Simulation results are finally provided to validate the theoretical findings developed in this paper.

KEYWORDS: Current control, grid-connected inverters, LCL-filter, resonance damping.

I. INTRODUCTION

Growing with the increased adoption of renewable energy for electric power generation, three-phase voltage source inverter (VSI) has, to date, become an important interfacing topology for channeling renewable and other clean energy to the utility grids. For smoothing the currents injected into the grid, a single inductor L is usually used and connected in series with the output ports of the converter to perform filtering function. However, it is known that in high-power applications where the switching frequency of the converter is always limited by the associated switching losses, such a simple configuration may lead to bulky and costly passive filters, and may also slow down the dynamic response of the system if the designed converter needs to comply with the stringent grid codes specified. In order to overcome the aforementioned issues, some researchers have focused on design and control of three-phase grid-connected VSIs with higher order LCL-filters. Being a third-order system, LCL-filters can provide much better ripple and harmonic attenuation over the higher frequency range using smaller passive elements. Therefore, they are more suited for high-power conversion systems and have already been widely employed in wind farms of over hundreds of kilowatts [1]–[3]. Despite this prominent advantage, which is not possessed by the single L-filter, the inclusion of LCL-filters with three-phase VSIs complicates the current control design. The underlying reason for this is related to the theoretical fact that at certain frequencies, the resulting network may appear to have zero or very low impedance, inferring resonances, and hence closed loop instability. A straightforward method for dampening this resonance is to add real passive damping resistors in series with the filter capacitors. An analytical design approach for passively damping LCL-filters was discussed in [1]. Moreover, the passively damped filter may downgrade to a second order system, which will definitely compromise its attenuation factor at higher frequencies, and hence neutralize those benefits originally introduced by the undamped filters.

To provide an answer, this paper comprehensively explores the inherent damping characteristic of an LCL-filter, which so far has not yet been discussed in the literature. It specifically shows that the current control loop has an inherent damping term embedded when the converter current is sensed for feedback. This inherent damping term can be used for optimal damping of the LCL resonance without demanding additional passive or active damping, so long as the filter component values are designed appropriately. Such inherent damping unfortunately does not exist when the grid current is measured for feedback control, hence leading to the general conclusion that converter current feedback is more stable than grid current feedback. Upon drawing that conclusion, the next practical issue to address is the likely deviation of passive element values caused by aging, temperature variation, and system uncertainty. Simulation and experimental results for validating those findings are provided a at the end of the paper.

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II. SYSTEM MODELING AND DESCRIPTION

Fig. 1 shows the typical circuit diagram of a three-phase VSI connected to the grid through an LCL-filter. The equivalent series resistances (ESRs) of its converter-side inductor L_{inv}, grid-side inductor L_g , and filter capacitor C_f are all neglected here.

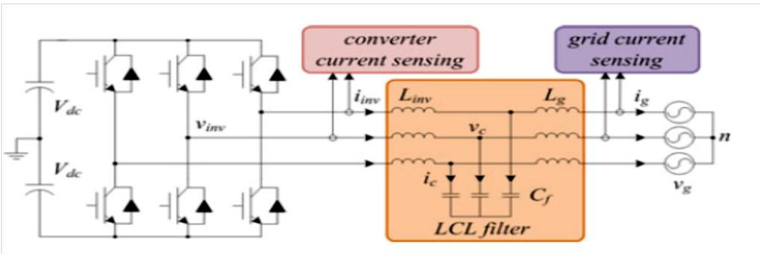

Fig.1. Three-phase grid-connected inverter with an LCL-filter.

Not including ESRs would, in principle, nullify all internal damping, causing the idealized system to be more unstable. Another assumption made here is that the ac supply voltages contain only positive-sequence fundamental component, which then means that they can be treated as short circuits with zero impedance when performing system stability and harmonic analyses. With both assumptions incorporated, Fig. 1 simplifies to the per-phase equivalent circuit shown in Fig. 2, whose s-domain transfer functions, representing the plant model $G_p(s)$ at non fundamental frequencies, can be readily derived as where v_{inv}, i_g, and i_c represent the converter pole voltage, grid current, and filter capacitor current in the time domain, respectively, while their capitalized notations are for representing them in the s-domain.

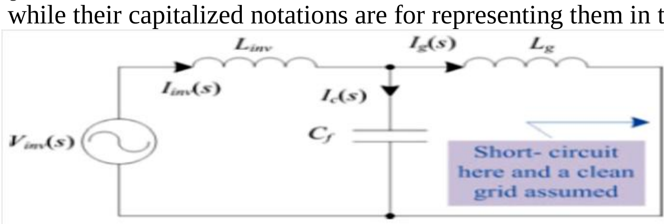

Fig.2. Simplified per phase equivalent circuit for stability analysis.

$$G_p(s) = \frac{I_g(s)}{V_{\rm inv}(s)} = \frac{1}{L_{\rm inv} L_g C_f s^3 + (L_{\rm inv} + L_g)s} \quad (1)$$

$$\frac{I_c(s)}{I_g(s)} = L_g C_f s^2 \quad (2)$$

Instead, it is noted that the main impact introduced by delays is to significantly reduce the system phase over the full frequency range, which certainly can be accounted for by dropping the phase of the open-loop transfer function derived with the PWM converter treated as a pure amplifier gain. This gain does not need be drawn explicitly, but can be combined with the current regulator, whose transfer expression in the synchronous dq frame is written as

$$G_c(s) = K_p V_{\rm dc} \left(1 + \frac{1}{\tau_c s}\right). \quad (3)$$

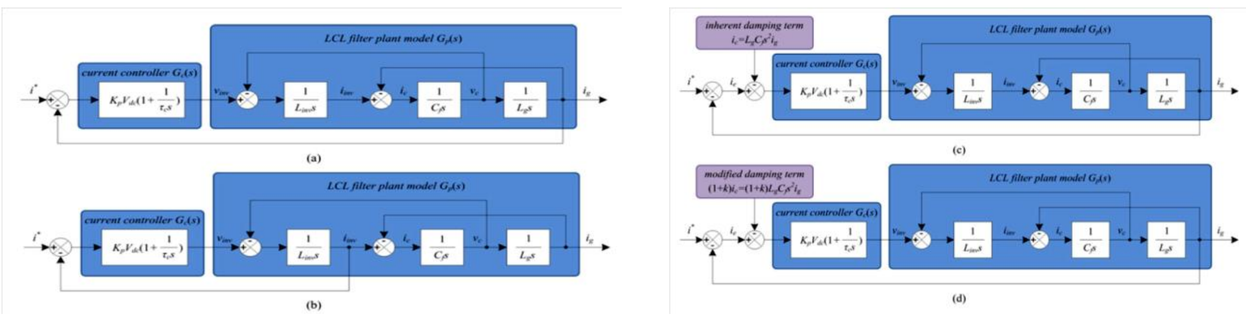

Fig.3. Block diagram of (a) grid current feedback control, (b) converter current feedback control, (c) equivalent converter current feedback control, and (d) modified equivalent converter current feedback control.

Equation (3) is no doubt a proportional-integral (PI) controller with K_p, τ_c, and V_{dc} representing its proportional gain, integral time constant, and half the dc-link voltage, respectively.

A. Grid Current Feedback

Fig. 3(a) shows the control block diagram of LCL-filter-based three-phase VSIs with grid current feedback. Combining (1) and (3), it is easy to derive the system open-loop gain as

$$\frac{I_g(s)}{I_e(s)} = G_c(s)G_p(s) = \frac{K_p V_{dc}(s + 1/\tau_c)}{L_{inv} L_g C_f s^4 + (L_{inv} + L_g)s^2} \quad (4)$$

Where i_e represents the current reference in the time domain and its capitalized notation denotes the s-domain. The third-order term in the characteristic equation is absent, indicating that the overall system is difficult to be closed-loop stable. Even if the ESRs of passive elements are included, they are generally insufficient to guarantee a well-damped plant. The plant response is thus dynamically slow with a small control bandwidth. Improvement to it can only be introduced by adding either external passive or active damping to limit the otherwise infinite gain at the LCL resonance frequency.

B. Converter Current Feedback

The converter current can alternatively be sensed for feedback control as shown in Fig. 3(b). Since the grid current is not directly controlled, the q-axis current reference should be set to $\omega_o C_f v_g$ instead of zero, where ω_o is the fundamental angular frequency. This is to again ensure unity power factor operation immediately before the grid terminal. Although Fig. 3(b) accurately represents the system, its adoption for open-loop stability analysis is not convenient because the feedback variable is tapped from the middle of the open-loop path rather than at the end. This fortunately can be resolved by noting that the converter current can be expressed as the summation of grid current and filter capacitor current, which when incorporated, gives rise to the modified control block diagram shown in Fig. 3(c). Comparing Fig. 3(c) with Fig. 3(a) reveals that converter current feedback is equivalent to grid current feedback, except for an additional—i_c term added to the forward path of the former. Using this modified notation, it is easy to obtain the following relationship.

$$\frac{I_g(s)}{I_e(s) - I_c(s)} = G_c(s)G_p(s) = \frac{K_p V_{dc}(s + 1/\tau_c)}{L_{inv} L_g C_f s^4 + (L_{inv} + L_g)s^2}. \quad (5)$$

Substituting (2) into (5) and performing slight manipulation then yields

$$\frac{I_g(s)}{I_e(s)} = \frac{K_p V_{dc}(s+1/\tau_c)}{L_{inv} L_g C_f s^4 + K_p V_{dc} L_g C_f s^3 + (K_p V_{dc} L_g C_f/\tau_c + L_{inv} + L_g)s^2}. \quad (6)$$

Comparing the denominators of (4) and (6), it is clear that the latter has an additional s^3 term, which would provide damping by shifting two of its poles further into the left half plane, while the remaining two poles remain at the origin. The extent of damping introduced can be tuned by adjusting the controller gains and passive parameters to size the coefficient of s^3 found in the denominator of (6) appropriately. As an example, to arrive at the optimum damping factor of $\zeta = 0.707$ recommended in most control texts, tuning should be done until the following condition is met

$$\frac{K_p V_{dc}}{L_{inv}} = 2\zeta\omega_{res} = 2\zeta\sqrt{\frac{K_p V_{dc} L_g C_f/\tau_c + L_{inv} + L_g}{L_{inv} L_g C_f}} \quad (7)$$

where ω_{res} , being the undamped resonance frequency, has been replaced by the relatively complex square-root term written on the right of (7). Since the integral term of the PI controller usually does not impact the resonance frequency significantly, this expression can be further simplified as

$$\omega_{res} \approx \sqrt{\frac{L_{inv} + L_g}{L_{inv} L_g C_f}} = \frac{1}{\sqrt{L_p C_f}} \quad (8)$$

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III. OPTIMUM DAMPING CONTROL OF LCL-FILTER

Previous research works on LCL-filtered systems usually assume that the filter can be approximated by an inductor, whose value is written as $L = L_{inv} + L_g$ in the low frequency range toward the left of the resonance frequency. From those plots, it is clear that the LCL-filter has roughly the same magnitude response as an L-filter over the low frequency range and under various damping factors. Their phase response is, however, quite different with the maximum phase lag of an L-filter being π/2 radius, while that of an LCL-filter always exceeding this value. The amount in excess of the LCL-filter would grow larger as the damping factor and operating frequency increase.

Because of the differences in phase, the first design recommendation suggested here is to optimally tune the damping factor of LCL resonance. Optimum damping control is deemed important since an under or lightly damped system will lead to serious transient oscillations when excited by a step change, while an over damped system will severely decrease the system phase margin, and hence dynamic response. The second recommendation made here is closely related to the dependence of the system crossover frequency ω_c with reference to its damping factor and resonance frequency.

$$\varphi_{LCL} \approx \frac{\pi}{2} + \arctan\left(\frac{2\zeta\alpha}{1-\alpha^2}\right). \qquad (9)$$

When optimum damping of LCL resonance is considered ($\zeta = 0.707$), the recommended value of crossover frequency is $\alpha = 0.3$. In this case, the system may have an acceptable bandwidth and the phase lag introduced by the LCL-filter is about 0.64π (115◦), which is not very severe degradation compared with that of L-filter (90◦). Sometimes, the damping factor can be slightly decreased to 0.5 so that higher crossover frequency can be obtained and hence faster current control loop. With ωc defined, it is now possible to design the optimum damping control for LCL-filter. Referring back to Fig. 4, it is obvious that the magnitude response of LCL-filter under various damping factors is basically the same as that of an L-filtered system in the low frequency range. Therefore, it is reasonable to approximate this LCL-filter as a single L (where $L = L_{inv} + L_g$) to find the relationship between K_p and ω_c . According to [17], this relationship can be depicted by the following equation

$$K_p \approx \frac{\omega_c L}{V_{dc}} = \frac{\omega_c(L_{inv} + L_g)}{V_{dc}}. \qquad (10)$$

Combining (7) and (10) gives the following relationship between L_{inv} and L_g

$$2\xi L_{inv} = \alpha(L_{inv} + L_g)\frac{L_{inv}}{L_g} = \frac{\alpha}{2\zeta - \alpha}. \qquad (11)$$

Equation (11) suggests a theoretical design criterion based on which optimum damping of LCL-filter can be achieved with its inherent damping characteristic. It should be noted that the performance of optimum damping will not be affected by the proportional gain of PI controller as long as (11) is satisfied. Therefore, the system crossover frequency can be placed at a desired frequency range so that the system dynamic response is fast enough. Moreover, (11) also reveals that the filter capacitance will not impact the performance of resonance damping, so this value should be first determined in LCL-filter design as will be presented in a later section. Nevertheless, (11) is only used for proving the concept that it is possible to achieve optimum damping control of LCL-filter through its own inherent damping characteristic. This design approach is, however, not recommended for practical implementation. On one hand, (11) usually leads to much smaller value on converter-side inductor ($L_{inv} : L_g = 3 : 11$ for $\alpha = 0.3$ and $\zeta = 0.7$), so it will not comply with the optimum design method of LCL-filter proposed in [19]–[21], where it has been stated that the grid- and converter-side inductances should be the same so as to maximize the filtering capability of inductors.

A straightforward method to have a tunable damping factor is to measure the filter capacitor current and feedback to the current reference with a proportional gain k [22]. With this simple modification, the equivalent block diagram with converter current feedback control is shown in Fig. 3(d) which gives

$$\frac{I_g(s)}{I_e(s) - (1+k)I_c(s)} = G_c(s)G_p(s) = \frac{K_p V_{dc}(s + 1/\tau_c)}{L_{inv}L_g C_f s^4 + (L_{inv} + L_g)s^2}. \qquad (12)$$

Again, with (2) substituted into (12), the decoupled system open-loop gain will be, as shown in (13) at the bottom of this page.

$$\frac{I_g(s)}{I_e(s)} = \frac{K_p V_{dc}(s + 1/\tau_c)}{L_{inv}L_g C_f s^4 + (1+k)K_p V_{dc} L_g C_f s^3 + [(1+k)K_p V_{dc} L_g C_f/\tau_c + L_{inv} + L_g]s^2}. \qquad (13)$$

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The second term of the denominator in (13) dominates the damping factor of *LCL*-filter and should satisfy the following relationship for optimum damping control

$$2\zeta\omega_{\rm res} = \frac{(1+k)K_p V_{\rm dc} L_g C_f}{L_{\rm inv} L_g C_f} = \frac{(1+k)K_p V_{\rm dc}}{L_{\rm inv}}. \quad (14)$$

It should be noted that, in practical implementation, k is usually less than the value computed from (14), as the system internal resistance can also provide a certain degree of damping to the LCL resonance.
Substituting (10) into (14) gives (15) for computing the feedback gain of i_c

$$k = \frac{2\zeta L_{\rm inv}}{\alpha(L_{\rm inv} + L_g)} - 1. \quad (15)$$

where ω_n is the nominal angular frequency and Q is the quality factor that mainly determines the dynamic response and selectivity of this notch filter. In order to have a fast transient process without high oscillations, Q factor here is chosen to be 0.7 and its implementation is demonstrated in Fig. 4.

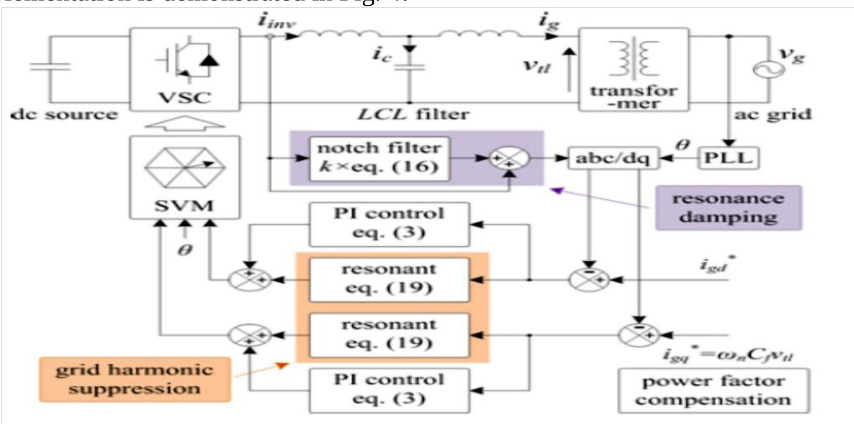

Fig.4. Implementation of overall control block diagram for experimental tests.

TABLE I
SPECIFICATIONS OF THE CONVERTER AND PARAMETERS USED FOR SIMULATION

Elements	Parameters	Values
Converter	Nominal power	1.5 kVA
	Switching frequency f_s	5 kHz
	Deadtime	2 μs
	Dc-link voltage V_{dc}	260 V
Grid	Grid voltage V_g	120 V
	Line frequency f_o	50 Hz
	Grid impedance + transformer leakage inductance L_g	1.3 mH
LCL-filter	Converter-side inductor L_{inv}	1.3 mH
	Filter capacitor C_f	15 μF
Controller	Proportional gain K_p	0.06
	Integral time constant τ_c	0.005
	Damping gain k	1.2

IV. SIMULATION RESULTS

Simulation was then conducted in MATLAB/PLECS environment to verify the effectiveness of the proposed design approach, and all the parameters are chosen to be the same as those in the aforementioned design The simulated result is then presented in Fig. 7(a), where it is quite clear that the system can be kept stable under both steady and transient states. When the current reference is forced to jump from 5 to 10 A, the inverter can respond very fast and there is no oscillation or current overshoot observed during current step change, which successfully confirms the concept of

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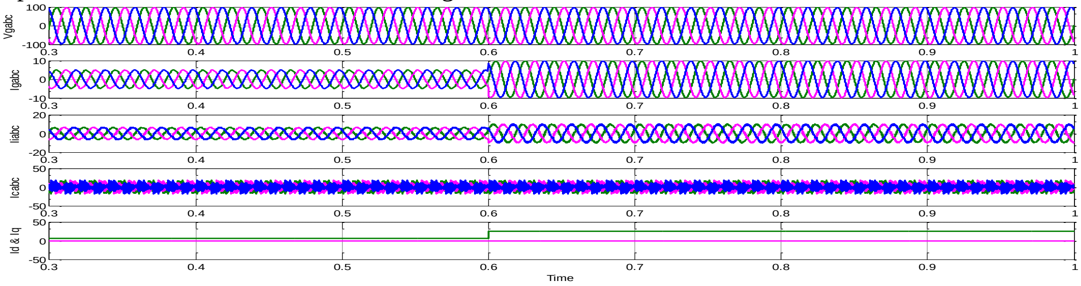
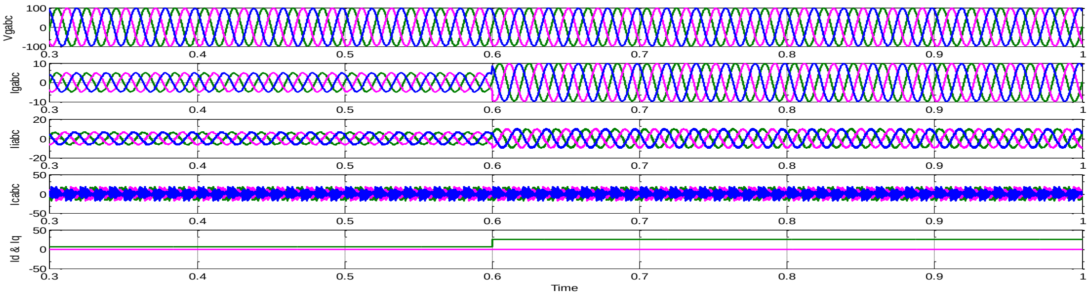
inherent resonance damping of LCL-filter. Since this type of LCL-filter no longer satisfy the design criterion suggested by (11), the resonance of LCL-filter will be improperly damped, causing serious transient oscillations as can be visualized from Fig. 7(b).On the other hand, small converter side inductance leads to very large current ripples (approaching 8A under nominal operation), as revealed by the simulation results shown in Fig. 7(a). This may result in significant power losses and even cause overheating of converter-side inductor.

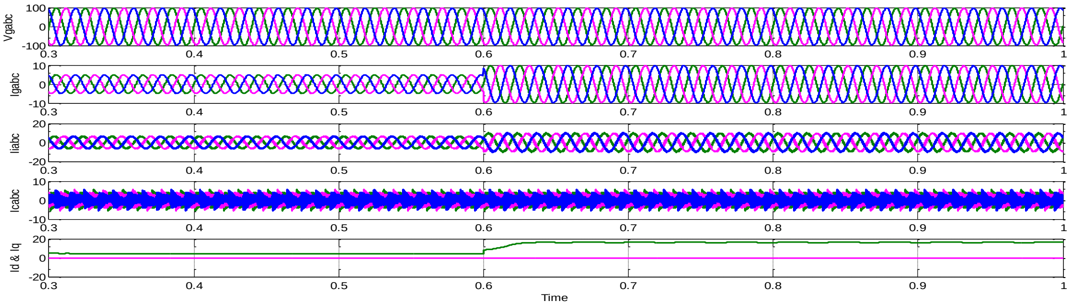


(a)

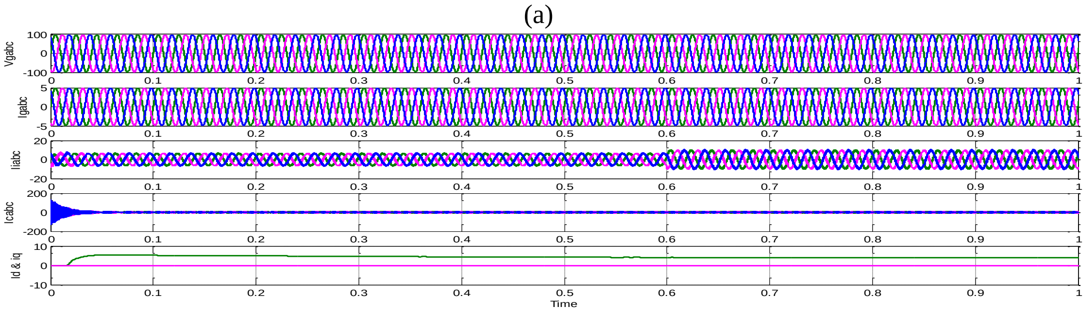


(b)

Fig.7. Simulated results of a three-phase inverter connected to the grid through an *LCL*-filter with (a) the proposed design method and (b) mismatched inductance.



(a)



(b)

Fig. 8.Simulated results of *LCL*-filter-based inverter using (a) converter current control and (b) grid current control with insufficient passive damping.

The design of the second set of LCL-filter is relatively simple, and the inductance is equalized to achieve maximum filtering of switching harmonics. The final values of passive elements of LCL-filter are chosen as C_f=15μF (4.5%) and L_{inv} = L_g = 1.36mH (4.4%). Again, only the converter current is measured for implementing the feedback control.

Owing to the inherent damping characteristic of LCL-filter, the harmonic components in the converter current can help stabilize the system without relying on any other damping techniques. To achieve an optimally damped system, an additional damping term i_c must be estimated from i_{inv} and subsequently added to the control loop through a gain k to tune the damping factor. The calculation of k can be found by (15). Consequently, the simulated waveforms are presented in Fig. 8(a), the total harmonic distortion (THD) of i_g is less than 1%, and there is also no overshoot current observed in its dynamic process, confirming the good damping performance of the proposed control scheme. One observation to be highlighted here is that it takes roughly half line cycle for the grid current reaching the steady-state after current step-up change as shown in the fifth trace of Fig. 8(a). This relatively slow dynamic response is solely caused by the notch filter, which is used for extracting the harmonic components of converter current, rather than the current controller, and it will not raise any stability problem. However, this value of resistance is still deemed as insufficient damping and transient oscillations in the grid currents are quite obvious as shown in Fig. 8(b).

V. CONCLUSION

Filters are main parts of a renewable energy system. First- order passive filters are L type which are generally use for controlling grid-connected inverter. The disadvantage of this type of filters is their big size. Another type of the passive filters is LC filters (second-order). Because of the big size of the inductor, the size of this filter is large. Moreover, time delay and resonance frequency are another drawbacks of LC filters. Compared with a first-order and second-order filters, a third-order LCL filter has lower coast and smaller size in applications above several kilowatts. However, resonance frequency is still as a problem of these filters. This paper has made an in-depth investigation of two different current control schemes for LCL-filter-based three-phase grid-connected inverters. It has been shown that using the converter current for feedback control will result in a more stable closed-loop control system due to the unique inherent damping characteristic of LCL-filters. Then, a general design criterion of LCL-filter has been proposed that allows the optimum damping achieved with a desired system control bandwidth.

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