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# Virtual Resistor Active Damping with Selective Harmonics Control of LCL-Filtered VSCs

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**Abstract**—LCL filter is widely adopted for strict standard compliance of grid-tied voltage source converters (VSCs). The third order low-pass filtering provides great attenuation for the high frequency harmonics generated by the power electronics guaranteeing low output currents noise injection into the grid. A major concern of the implementation of the LCL-filter is to safeguard the system stability by providing effective damping of the filter resonances. Active damping methods are preferred because it does not result in substantial power losses as it would result by the utilization of passive damping circuits. Capacitor-current active damping (CCAD) technique can be effective while realizing only a proportional feedback. However, the suitable feedback gain for maintaining stability remains to be identified. Another control issue related to the grid-tied VSC is the harmonics compensation of the currents due to the grid voltage distortion. Therefore, this paper proposes an improved resonator with phase compensation to suppress the harmonics distortion while maintaining stability with properly design capacitor current feedback. The capacitor feedback gain for stability is analytically derived in this paper. The proposed control scheme is verified by both simulation and experimental results.

**Index Terms**—Capacitor-current feedback Active damping (CCAD), Harmonics resonator, LCL filter, VSC

## I. INTRODUCTION

Pulse-Width Modulated (PWM) voltage source converters (VSCs) are the main building blocks for interfacing distributed generation (DG) units, power flow control and emerging smart loads with hybrid ac-dc distribution grids [1]–[4]. Continuous PWM strategy such as Sinusoidal PWM (S-PWM) is widely used for the modulation of VSCs due to the simplicity and better harmonics performance, instead of discontinuous PWM strategies [5], [6]. LCL filter is preferably adopted in VSCs for attenuating high frequency harmonics in the output waveforms. However, special attention must be paid to the LCL filter resonance which usually needs to be damped to safeguard the system stability. In renewable generation applications the grid-connected VSCs are generally controlled in Current-Control Mode (CCM). The choice of the converter or grid-side current control is influenced by the resonance of LCL filter based on

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the ratio between the resonance frequency  $f_r$  and the PWM sampling frequency  $f_s$  [7]. Recently, studies on the single-loop control revealed that the grid-side current control can be more stable if the the resonance frequency  $f_r$  lies between  $\frac{1}{6} f_s$  (critical frequency) and  $\frac{1}{2} f_s$  (Nyquist frequency) if the symmetric sampling and single-update PWM pattern is adopted [8]. For the double-update case, the critical frequency shifts to  $\frac{1}{4} f_s$  [9]. Correspondingly, stability with converter-side current control can be ensured without any additional means if the  $f_r$  is found below the critical frequency.

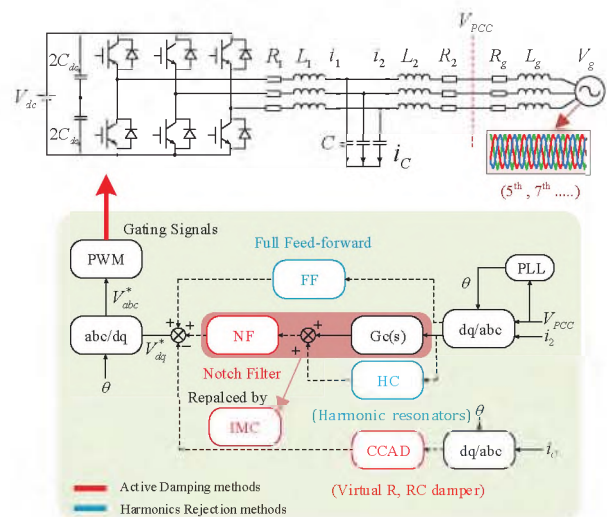


Fig. 1. Three-phase grid-tied voltage source converter with a LCL filter.

Achieving stable control is possible while operating in the previously described unstable control region defined by the chosen LCL filter and sampling frequency. However, a filter resonance damping measures becomes necessary for guaranteeing enough stability margin for the system. Passive damping solutions, i.e., parallel or series connection of resistor with the capacitor and/or inductor offers satisfactory performance but has consequence on the overall size and system efficiency [10]. On the other hand, the active damping schemes can improve the stability while not influencing the efficiency [7]–[9]. These methods can be classified into two main categories [8], [11], [12]: 1) multi-loop control methods; and 2) filter-cascaded methods.

The filter-based active damping methods, shown in Fig.1, are mainly realized by cascading the filter in the current controller loop. The notch filter (NF) method typically employs a second order digital filter which creates an anti-resonance at the pre-set LCL filter resonance frequency to improve the system attenuation around this frequency [11], [13]. Other filter-based methods, i.e, IMC (Internal Model Control) is derived based on the model of the control plant and are studied in [14]. However, both the NF and IMC model are sensitive to system parametric variations, e.g, the change of the LCL filter component values with the current bias and/or the intrinsic unknown grid impedance (represented by  $L_g$  and  $R_g$  in Fig.1). The adaptive or self-commissioned filter methods are proposed in [15] to improve the robustness against the circuit parametric uncertainties and dynamic variations, however, at the cost of control complexity. On the other hand, the multi-loop active damping method uses the available state signals in the circuit through feedback control to form an equivalent virtual resistor in series or parallel with the LCL filter elements, e.g., the filter capacitor and inductors. While both capacitor voltage and current feedback based control have been reported, cost of the voltage sensor can be lower compared to the high-bandwidth current ones. On the other hand, capacitor voltage active damping (CVAD) [16] requires the implementation of the differentiators in the feedback loop, which is difficult to implement in practice. By contrast, capacitor current active damping (CCAD), shown in Fig.1, only requires a proportional feedback [17]. If the influence of the digital control delay is considered, the proportional feedback of the capacitor current may result in a negative virtual resistor in some cases [12], leading to an undesired non-minimum phase system. However, this negative virtual resistor can be avoided if the resonance frequency is smaller than  $\frac{1}{6} f_s$  with maximum possible delay.

Another control issue in a grid-connected VSC system is the control of the current harmonics due to the distorted grid voltage. Full feed-forward (FF) scheme of the grid voltage is reported to be able to effectively compensate the harmonics by fully canceling out the influence of the grid voltage [18]. However, full FF scheme needs to implement the first and second-order differentiators which is hard to precisely discretize in the digitally controlled system. In contrast, harmonic resonant controller is widely adopted to realize the selective harmonics compensation, which enlarges the loop gain at the tuned frequency. From the control point of view, the harmonics current can be suppressed due to the infinitely large grid impedance at the harmonics frequencies [2], [19], [20]. However, a resonator controller requires phase compensation [20], [21] to ensure enough phase margin due to the system delay. In this paper, an improved harmonics resonator with phase compensation along with a design guideline for capacitor current active damping is presented. The main contributions are 1) The analytical derivation of suitable capacitor current feedback gain ensuring system stability 2) The derivation of an improved resonator controller with phase compensation.

Section II presents the system description and basic control scheme. The analysis of capacitor current active damping

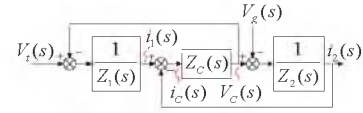


Fig. 2. Block diagram of the control plant of LCL-filtered VSC.

and harmonics control are introduced in Section III and IV respectively. Simulations and experimental results are carried out in Section V. Conclusion and future work are presented in Section VI.

## II. SYSTEM MODELING AND CONTROL

### A. Circuit Modeling

The VSC connected to the grid with a LCL filter is depicted in Fig. 1. The described grid is assumed to be a stiff grid which contains low-order harmonics such as 5<sup>th</sup>, 7<sup>th</sup>, 11<sup>th</sup> and 13<sup>th</sup>. Therefore, the grid interface components  $R_g$  and  $L_g$  are neglected in this paper and hence,  $V_{PCC} = V_g$ . The filter resistors  $R_1$  and  $R_2$  represent the internal resistance of the converter-side inductor  $L_1$  and grid-side inductor  $L_2$ , respectively. Additionally, the equivalent resistance due to the conduction losses of the VSC is incorporated into  $R_1$ . Fig.2 shows the average switched model of the control plant of a VSC with LCL filter, where  $V_t(s)$  is the VSC terminal voltage. The  $Z_1(s)$ ,  $Z_2(s)$  and  $Z_C(s)$  are the impedance of  $L_1$ ,  $L_2$  and  $C$  and given as:

$$Z_1(s) = sL_1 + R_1 \quad (1)$$

$$Z_2(s) = sL_2 + R_2 \quad (2)$$

$$Z_C(s) = \frac{1}{sC} \quad (3)$$

The transfer function relating the grid-side output current  $i_2$  and the converter terminal voltage  $V_t$  is:

$$G_{i_2}(s) = \frac{i_2(s)}{V_t(s)} = \frac{1}{\alpha s^3 + \beta s^2 + \gamma s + \delta} \quad (4)$$

where, constants are given by  $\alpha = L_1 L_2 C$ ,  $\beta = R_2 L_1 C + R_1 L_2 C$ ,  $\gamma = R_1 R_2 C + L_1 + L_2$  and  $\delta = R_1 + R_2$ . Considering that  $R_1$  and  $R_2$  are negligible, (4) can be simplified and given by (5).

$$G_{i_2}(s) = \frac{1}{L_1 L_2 C s} \left( \frac{1}{s^2 + \omega_r^2} \right) \quad (5)$$

where, the resonance frequency ( $\omega_r$ ) of the LCL filter is given by (6).

$$\omega_r = \sqrt{\frac{L_1 + L_2}{L_1 L_2 C}} \quad (6)$$

### B. Basic Control of the VSC

The block diagram of the grid-side current control of VSC without any active damping and harmonics compensation is shown in Fig. 3.

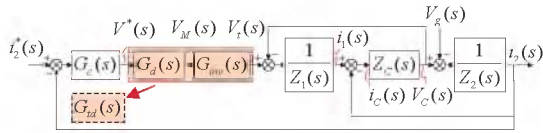


Fig. 3. Block diagram of the basic VSC grid current control scheme without active damping and harmonic suppression.

The compensator  $G_c(s)$  used for current control is typically a PR (proportional-resonant) controller in stationary ( $\alpha\beta$ ) frame or a PI controller in rotating ( $dq$ ) frame [2], [19].

$$G_{c(\alpha\beta)}(s) = k_p + \frac{k_r s}{s^2 + \omega_1^2} \quad (7)$$

$$G_{c(dq)}(s) = k_p + \frac{k_i}{s}$$

The control and computation delay in the digitally controlled system is noted as  $G_c(s)$ . The mechanism of the computation delay depends on the sampling and update of the implemented PWM [17]. If the symmetric sampled PWM with single update (sampling happens at the upper peak of the triangular wave) is adopted, the computation delay is given by  $G_d(s) = e^{-sT_s}$ .  $f_s$  and  $f_{sw}$  are the sampling and switching frequency respectively such that  $T_s = \frac{1}{f_s}$  and  $T_{sw} = \frac{1}{f_{sw}}$ .  $G_{inv}(s)$  is the transfer function representing the PWM process, which is given by (8).

$$G_{inv}(s) = \frac{K_{pwm} G_h(s)}{T_s} \quad (8)$$

where, the PWM gain  $K_{pwm}$  is the ratio of the dc voltage  $V_{in}$  and amplitude of the triangular carrier wave  $V_{tri}$  and taken as 1 to simplify the analysis.

The Zero-Order Hold (ZOH) characteristics of the PWM process is modeled as an equivalent transfer function  $G_h(s)$  as approximated by (9) using frequency domain analysis.

$$G_h(s) = \frac{1 - e^{-sT_s}}{s} \approx T_s e^{-\frac{1}{2}sT_s} \quad (9)$$

Combining (8) and (9), the total delay  $G_{td}$  can be expressed as (10).

$$G_{td}(s) \approx e^{-\frac{3}{2}sT_s} \quad (10)$$

### III. CAPACITOR CURRENT FEEDBACK ACTIVE DAMPING

#### A. Virtual Resistor

The allowable gain while ensuring stable operation of a VSC is limited when no active damping is used, as shown in Fig. 4 where the maximum controller gain is limited to  $K_{ad} = 0.5$ . Table. I lists the used parametric specifications.

TABLE I  
SYSTEM PARAMETERS

$L_1, L_2$	$R_1, R_2$	$C$	$V_{DC}$	$k_p$	$k_r$	$k_{th}$	$\omega_c$
1.5 mH	0.2 $\Omega$	20 $\mu$ F	700 V	5	2500	1000	2

As a consequence, the small bandwidth of the closed-loop system leads to undesired slow system response. Consider the equivalent circuit of the LCL filter with capacitor current  $i_c$

shown in Fig. 5. The feedback of  $i_c$  is equivalent to adding an impedance ( $Z_{ad} = R_{ad} + jX_{ad}$ ) in parallel with the filter capacitor [12].

In CCAD, the current control loop is modified using feedback of  $i_c$  via gain  $K_{ad}(s)$  as shown in Fig. 6. The relation between  $Z_{ad}$  and  $K_{ad}(s)$  is described by (11).

$$Z_{ad}(s) = \frac{L_1}{K_{ad}(s)G_{td}(s)C} \quad (11)$$

Without consideration of the delay  $G_{td}(s)$ , the equivalent impedance behaves like a virtual resistor if the proportional feedback gain  $K_{ad}$  is used [12], which is expressed by (12):

$$R_{ad} = \frac{L_1}{K_{ad}C} \quad (12)$$

However, the proportional feedback results in the virtual impedance if the system delay is included. By substituting (10) into (12), the impedance can be derived as (13):

$$Z_{ad}(s) = \frac{L_1}{K_{ad}C} e^{-sT_s} = R_{ad} e^{-sT_s} \quad (13)$$

Substituting  $s = j\omega$  yields (14).

$$Z_{ad}(j\omega) = R_{ad} [\cos(\frac{3}{2}\omega T_s) + j\sin(\frac{3}{2}\omega T_s)] \quad (14)$$

As noted in (14), the delay influences both the resistive and inductive (or capacitive) parts. The imaginary term may alter the LCL filter resonance frequency and influence the precision of the control [16]. Meanwhile, it can be proven that the resistive part becomes negative when the frequency  $f$  ( $\omega/2\pi$ ) is above  $\frac{1}{6}f_s$ , which leads to a non-minimum phase behavior of the grid-side current [12]. Hence, the proportional feedback gain of the capacitor current should be properly designed. Furthermore, the identification of the range of the gain  $K_{ad}$  which guarantees the stability of the system is essential for designing a well-tuned controller.

#### B. Range of Feedback Gain for Stability of the System

After the capacitor current feedback to the control loop, the equivalent control plant  $G'_{i_2}$  can be expressed by (15):

$$G'_{i_2}(s) = \frac{1}{L_1 L_2 C s^3 + G_{td}(s) K_{ad} L_2 C s^2 + L_2 s} \quad (15)$$

$$= \frac{1}{L_1 L_2 C s^2 + K_{ad} G_{td}(s) / L_1 s + \omega_c^2}$$

Based on the previously discussed control elements of the controlled system, the open-loop transfer function from the reference input to the output grid-side current is derived in (16):

$$G_{oi}(s) = G_c(s) G_d G'_{i_2}(s) \quad (16)$$

$$= \frac{G_c(s) G_{td}(s)}{L_1 L_2 C s (s^2 + K_{ad} G_{td}(s) / L_1 s + \omega_c^2)}$$

In order to derive the analytical formulations from the delay-influenced system, the delay  $G_d(s)$  can be approximated with the rational polynomial form in (17):

$$G_d(s) = e^{-sT_d} = \left( \frac{1 - \frac{T_d s}{2n}}{1 + \frac{T_d s}{2n}} \right)^n \quad (17)$$

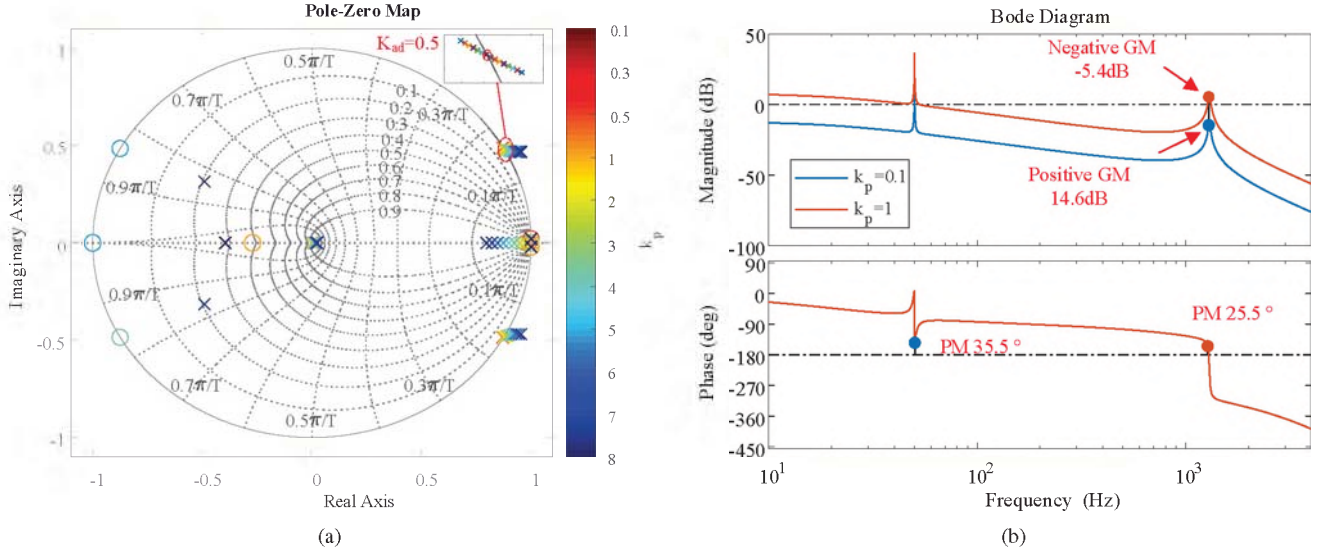


Fig. 4. Control loop without active damping (a) Poles-zeros map (b) Bode Plot.

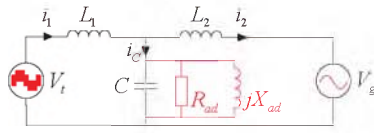


Fig. 5. Equivalent circuit of the LCL-filter based integration of the grid-connected VSC.

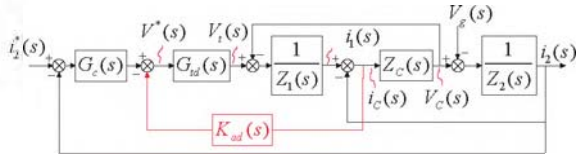


Fig. 6. Block diagram using CCAD control.

where  $n=2$  can give a precise approximation. Therefore, it can be noted that the pole of the open-loop can be located in the right-half part of the  $s$  plane or outside the unit circle of the corresponding  $z$  plane, with the increase of the feedback gain. Thus, the critical  $K_{ad,c}$  is defined as the gain which proceeds a imaginary pole  $s = j\omega_x$ , namely

$$\text{pole}(G_{ol}(j\omega_x)) = 0 \quad (18)$$

Substitute (16) and (17) into (18),  $\omega_x$  is solved as:

$$\omega_x = \sqrt{\frac{-b - \sqrt{b^2 - 4ac}}{2a}} \quad (19)$$

$$a = \left(\frac{T_d}{4}\right)^4, \quad b = -\frac{3}{8}T_d^2, \quad c = 1,$$

and  $K_{ad,c}$  hence is expressed in (20):

$$K_{ad,c} = \frac{(\omega_x^2 - \omega_r^2)(1 + \omega_x^2 T_d^2 / 16)^2 L_1}{\omega_x^2 T_d^2 (1 - \omega_x^2 T_d^2 / 16)} \quad (20)$$

From (19) and (20), it is observed that the imaginary pole  $\omega_x$  is the new resonance frequency shifted by the delay and it is solely determined by the value of the delay. Meanwhile, the maximum feedback gain  $K_{ad}$  allowing non-right-half-plane (RHP) poles is restrained by the LCL filter parameters and delay together. Above  $K_{ad,c}$ , the system can be still stable despite the open-loop poles located in RHP. Since the open-loop system has one pair of conjugate poles in the RHP, the boundary of this gain can be derived by forcing the open-loop system encircle the point  $(-1,0)$  once anti-clock-wisely in the half nyquist plot, where  $\omega$  increases from 0 to  $\infty$ . Since the  $-180^\circ$  crossing occurs at  $\omega_x$ , it can be mathematically expressed as:

$$|G_{ol}(j\omega_x)| \geq 1 \quad (21)$$

The upper boundary  $K_{ad,u}$  is then derived in (22).

$$K_{ad,u} = \frac{(\Delta\omega^2 + \omega_x^2 - \omega_r^2)(1 + \omega_x^2 T_d^2 / 16)^2 L_1}{\omega_x^2 T_d^2 (1 - \omega_x^2 T_d^2 / 16)} \quad (22)$$

$$\Delta\omega^2 = \frac{\sqrt{k_p^2 + (k_r/\omega_x)^2}}{\omega_x L_1 L_2 C} \approx \frac{k_p}{\omega_x L_1 L_2 C}$$

The expression (22) shows that the maximum feedback gain for achieving stability of the close-loop system is altogether determined by the controller parameters, the time delay and the LCL filter parameters. The difference between this gain and the previously derived  $K_{ad,c}$  points out that the controlled VSC with LCL filter can be still stable with the control of the current controller despite the open-loop poles in the RHP. However, such an unstable open-loop system is not preferred in practice and thus  $K_{ad,c}$  is taken as the maximum set value:

$$K_{ad-max} = K_{ad,c} \quad (23)$$

The minimum value of  $K_{ad}$  can be obtained by forcing the open-loop system to not encircle the point  $(-1,0)$  in the nyquist

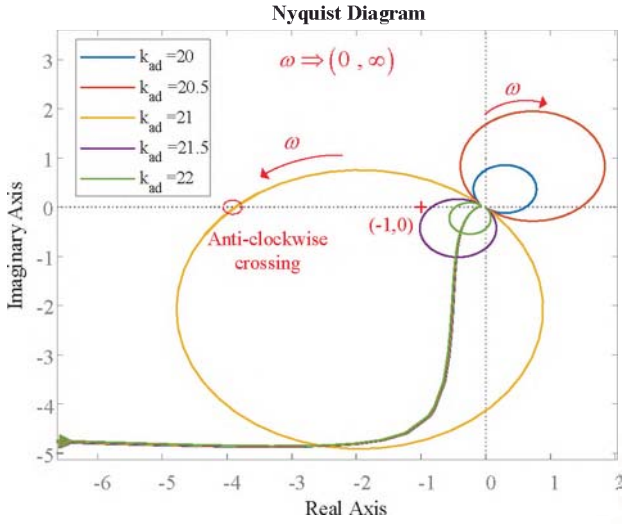


Fig. 7. Nyquist plot of the open-loop system (half-plot).

plot. The  $-180^\circ$  occurs at the resonance frequency  $\omega_r$ , hence it can be mathematically expressed in (24):

$$|G_{ol}(j\omega_r)| \leq 1 \quad (24)$$

which yields (25).

$$K_{ad-min} = \frac{\sqrt{k_p^2 + (k_r/\omega_r)^2}}{\omega_r^2 L_2 C} \approx \frac{k_p}{\omega_r^2 L_2 C} \quad (25)$$

Specifically,  $K_{ad-min}$  is equal to  $k_p/2$  when  $L_1 = L_2$  is satisfied. Therefore, the suitable range of feedback gain is:

$$K_{ad} \in (K_{ad-min}, K_{ad-max}) \quad (26)$$

Based on the parameters in Table.I, the pzmap of the close-loop system and the bode plot of the open-loop system are drawn in Fig.8, when the controller gain  $k_p$  is chosen as 5.  $K_{ad-c}$  and  $K_{ad-u}$ , as well as  $K_{ad-min}$ , are indicated in Fig.8a, which coincide with the calculated values based on the derived equations. Besides, Fig.8b shows the bode plot of the open-loop system with different  $K_{ad}$ , which presents the shifted resonance and the negative  $-180^\circ$  crossing. Fig.7 demonstrates the nyquist plot of the open-loop system with the  $K_{ad}$  around  $K_{ad-c}$  and  $K_{ad-u}$ . As the gain exceeds  $K_{ad-c}$ , a negative  $-180^\circ$  along with a resonance ( $>1$ dB) occurs in the bode plot which corresponds the one anti-clockwise encirclement of  $(-1,0)$  in the half nyquist plot. Meanwhile, the open-loop system has one conjugate pair of RHP poles. According to the nyquist stability criterion, the close-loop system is stable as the numbers of anti-clockwise encirclement equal the number of open-loop RHP poles. However, after  $k_{ad}$  becomes larger than  $K_{ad-u}$ , the resonance is smaller than 1 dB which means there is no encirclement of  $(-1,0)$ . Hence, the close-loop system has the same numbers of RHP poles as the open-loop system has.

### C. Stability Margins

In order to achieve stable control, an appropriate gain and phase margin should be guaranteed by properly setting up the feedback gain  $K_{ad}$  and controller gain  $k_p$ . The gain and phase margins of the controlled system are given by (27):

$$\begin{aligned} GM &= -20 \lg |G_{ol}(j\omega_r)| \\ PM &= \pi + \angle |G_{ol}(j\omega_{co})| \end{aligned} \quad (27)$$

where  $\omega_{co}$  is the gain crossover frequency where the open-loop gain  $|G_{ol}(j\omega)|$  reaches 1 and the phase crossover frequency is equal to  $\omega_r$ . The gain margin can be further expressed in (28):

$$\begin{aligned} GM &= -20 \lg \left| \frac{\sqrt{k_p^2 + (k_r/\omega_r)^2}}{K_{ad}\omega_r^2 L_2 C} \right| \\ &\approx -20 \lg \left| \frac{k_p}{K_{ad}\omega_r^2 L_2 C} \right| \\ &\approx 20 \left( \lg \left( \frac{K_{ad}}{k_p} \right) + \lg(\omega_r^2 L_2 C) \right) \end{aligned} \quad (28)$$

It can be noted that for a given LCL filter, the phase margin is proportional to  $K_{ad}$  and inversely proportional to  $k_p$  in logarithmic scale. To express the phase margin analytically, the gain crossover frequency is approximated in (29):

$$\begin{aligned} \omega_{co} &\approx \sqrt{\frac{-b + \sqrt{b^2 - 4ac}}{2a}} \\ a &= \left( \frac{K_{ad}}{L_1} \right)^2, \quad b = \omega_r^4, \quad c = -\left( \frac{k_p}{L_1 L_2 C} \right)^2 \end{aligned} \quad (29)$$

The phase margin can be further derived in (30):

$$\begin{aligned} PM &= \frac{\pi}{2} - \omega_{co} T_d - \tan^{-1} \left( \frac{k_r}{\omega_{co} k_p} \right) \\ &\quad - \tan^{-1} \left( \frac{\omega_{co} K_{ad} \cos(\omega_{co} T_d)}{L_1 (\omega_r^2 - \omega_{co}^2 - \omega_{co} \sin(\omega_{co} T_d))} \right) \end{aligned} \quad (30)$$

The derived relations are plot in Fig.9. It is noteworthy that both the gain and phase margin drop with the increase of  $K_{ad}$ . Usually, a gain margin with 10 dB guarantees a good stability. The corresponding  $K_{ad}$  is derived as 8, which indicates a phase margin of  $69^\circ$ .

### IV. HARMONICS COMPENSATION BASED ON RESONANT CONTROLLER

Harmonics resonators are widely used for suppressing the harmonics caused by the grid distortions by simply introducing single or multiples resonators tuned at the corresponding harmonics frequencies in parallel with the basic current controller. The generalized non-ideal resonators with two integrators [19] in  $\alpha\beta$  frame are expressed in (31):

$$G_h(s) = \sum_{h=3,5,7\dots} K_{ph} + \frac{2K_{rh}\omega_c s}{s^2 + 2\omega_c s + (h\omega_1)^2} \quad (31)$$

where  $K_{rh}$  and  $\omega_c$  represent the resonant controller gain and cut-off frequency (or damping factor) respectively. The bandwidth is adjusted by setting  $\omega_c$  appropriately. A small

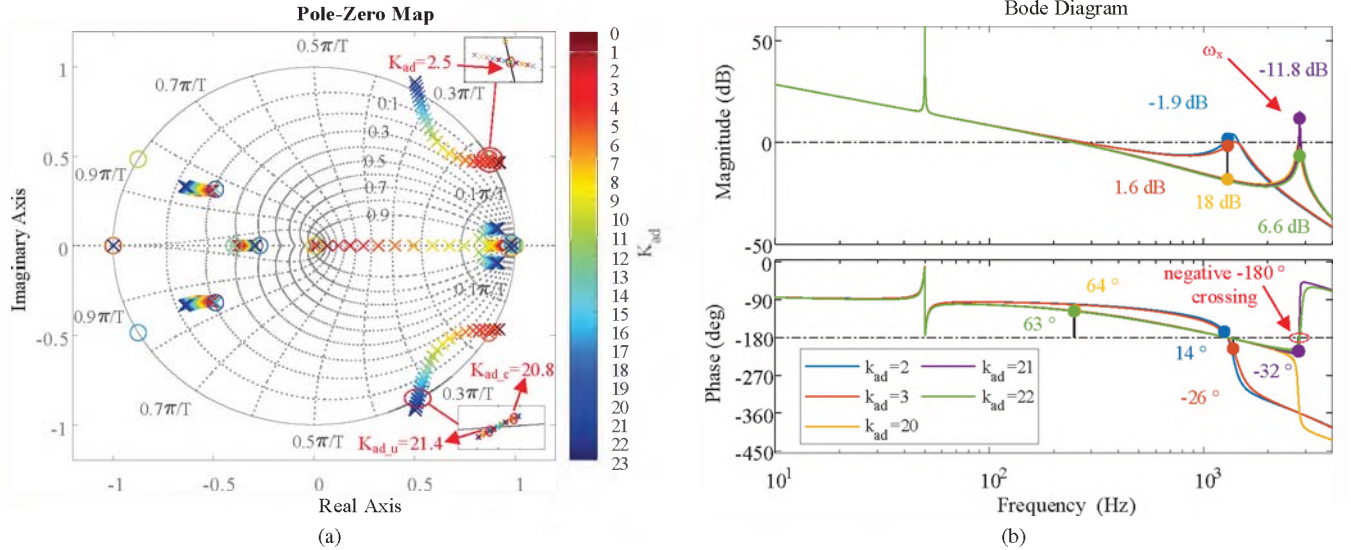


Fig. 8. Control loop with Capacitor-current feedback ( $k_p = 5$ ) (a) Poles-zeros map (b) Bode Plot.

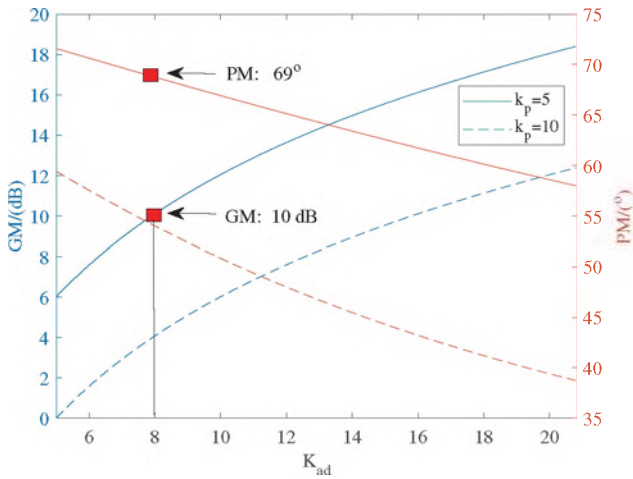


Fig. 9. Gain and Phase Margin versus  $K_{ad}$ .

cut-off frequency is usually chosen for high rejection performance [19]. One typical problem associated with the resonators is the phase deterioration caused by the delay. This can be improved by certain phase compensation methods, for example the introduction of a lead-filter [2] cascaded with the resonant controller. The generalized form of a lead filter ( $F_{lead}$ ) is given by (32)

$$F_{lead} = \frac{s/\omega_f + 1}{s\alpha/\omega_f + 1} \quad (32)$$

where  $\omega_f$  and  $\alpha$  are the tuning parameters determined by the required phase compensation  $\phi_{max}$  at the specified frequency  $\omega_{max}$ , which is described in (33).

$$\begin{aligned} \sqrt{\alpha} &= \tan\left(\frac{\pi}{4} - \frac{\phi_{max}}{4}\right) \\ \omega_f &= \omega_{max}\sqrt{\alpha} \end{aligned} \quad (33)$$

The bode plot in Fig.10 shows that the phase at the resonance frequency is improved by the lead filters and the phase margin is hence increased.

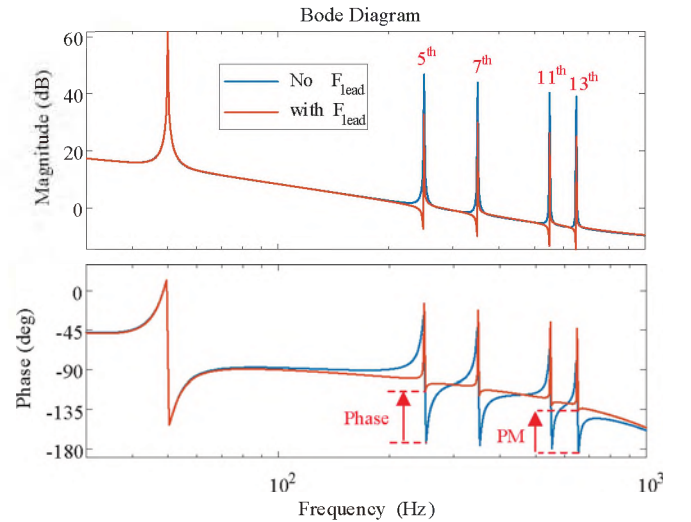


Fig. 10. Bode plot of the open-loop with harmonics resonators.

## V. SIMULATION AND EXPERIMENTAL VERIFICATION

### A. Simulation Results

To verify the accuracy of the identified range of the capacitor-current feedback gain, simulations are conducted in MATLAB/SIMULINK. The operating parameters are listed in Table I. The system operates in inverter mode at a power  $P_{ac} = 3$  kW with a feedback current control loop connected to a 400 V 50 Hz 3-phase grid.

Fig.11 shows the grid-side current waveform with different  $K_{ad}$  values under an ideal sinusoidal voltage. At  $K_{ad} = 2.6$ ,

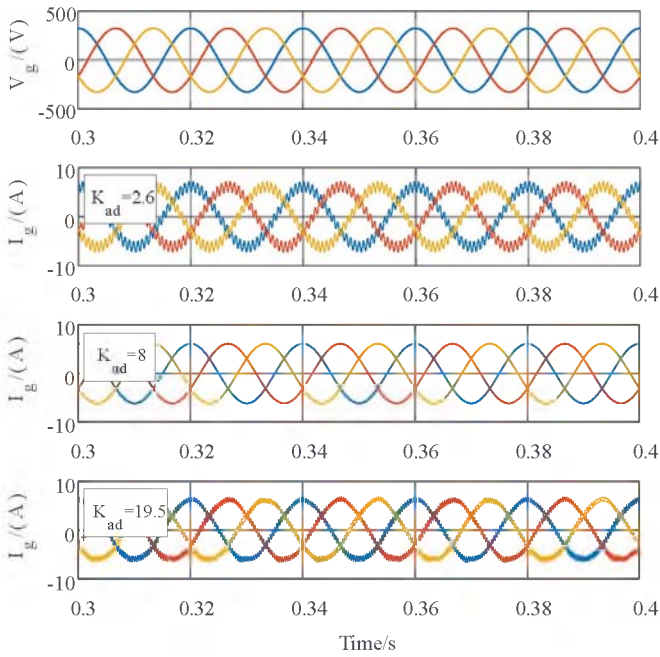


Fig. 11. Simulation results: grid voltages and grid-side currents under different  $K_{ad}$ .

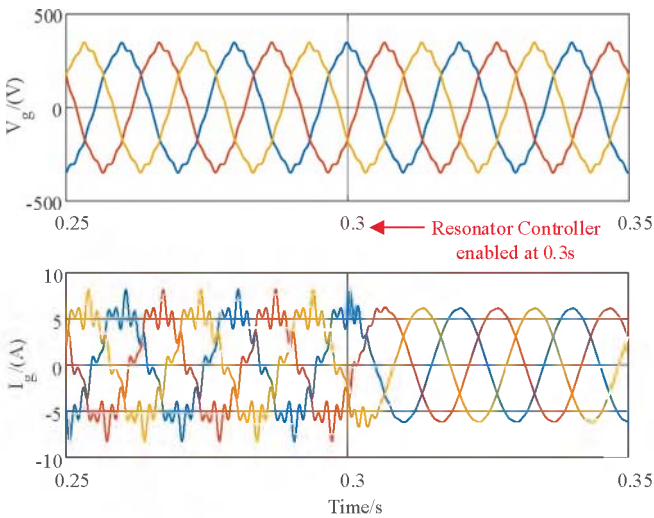


Fig. 12. Simulation results: grid voltages and grid-side currents with the proposed harmonics control.

which is close to the theoretical value 2.5, the grid -side current becomes marginally stable and the resonance occurs in the waveform with the frequency of  $\omega_r$  (1300Hz). With  $K_{ad} = 8$ , at which the system is supposed to have a GM of 10dB and PM of  $69^\circ$ , the grid-side current follows close to a sinusoidal 50Hz waveform. After the  $K_{ad}$  increases to 19.5, the new resonance component arises in the grid-side current, the frequency of which is around the shifted resonance frequency  $\omega_x$  (2800Hz). Therefore, the stability range identified from the simulation is close to the theoretical

one [2.5,20.8].

The simulation of the harmonics control is shown in Fig.12, where the distorted grid voltage containing low-order frequency harmonics with amplitudes of 5% of the fundamental rated voltage at the 5<sup>th</sup>, 7<sup>th</sup>, 11<sup>th</sup>, 13<sup>th</sup> is considered. The proposed resonator parameters are listed in Table.I and the lead filters are designed to provide  $60^\circ$  at each harmonics frequency. The harmonics control is enabled at 0.3s and the results show that the harmonics components in the grid-side current is quickly rejected by the resonators and the current achieves a satisfactory THD of about 0.5%.

### B. Experimental Verification

The experimental part of this paper is conducted in a commercial two-level VSC, i.e. Triphase/PM5F30C. The system is set to operate as a shunt active filter generating 1000 VAR reactive power to the distorted grid. The grid voltage is depicted in Fig.13, which contains in relation to the fundamental frequency component: 0.5% 5<sup>th</sup>, 2% 7<sup>th</sup>, 0.5% 11<sup>th</sup>, 0.3% 13<sup>th</sup>. As expected, the grid-side current is severely distorted due to the considerable filter capacitor, which leads to a quite large low-order harmonics in the current. Hence, the results in Fig.13 show the same tendency as the simulation results in Fig.12. Note that the system is still stable even without the capacitor-current feedback ( $K_{ad} = 0$ ). This is due to the quite large internal resistance of the LCL filter and inherent damping caused by the inefficient semiconductors in the Triphase system (high switching and conduction losses). The resonance (at  $\omega_r = 1300$ Hz) in the grid-side current at  $K_{ad} = 0$  shows that the system is close to the marginally stable condition. At  $K_{ad} = 7$ , only low-order harmonics exist in the grid-side current. After  $K_{ad}$  reaches 13, the system becomes marginally stable and the resonance (at  $\omega_x = 2800$ Hz) quickly arises. The maximum  $K_{ad}$  is smaller than the theoretical one and this can be explained by the fact that the delay and the damping characteristics in the system are different from the simulated system. In the final paper the parameters will be tuned accordingly so both simulation and experimental results are better matched. Fig.14 presents the grid-side current with and without the harmonics control under  $K_{ad} = 7$ . After the implementation of the proposed resonator, the THD of the grid-side current decreased from 32% to 7.5%, which is still quite large. This might be because that the insufficient phase compensation limits the further increase of  $k_{rh}$  and  $\omega_c$ .

## VI. CONCLUSION AND FUTURE WORK

This paper derives the boundaries of the capacitor-current feedback gain to realize effective active damping. Besides, a harmonics resonator cascaded with lead filter is proposed to compensate the selective harmonics components in the grid-side current due to the distorted grid voltage. Both simulation and experimental results show the performance of the proposed control scheme. However, the practical results deviates from the theoretical and simulation results due to the unidentified damping and delay in the practical system. The effectiveness of the resonator is so far limited by the current achieved phase



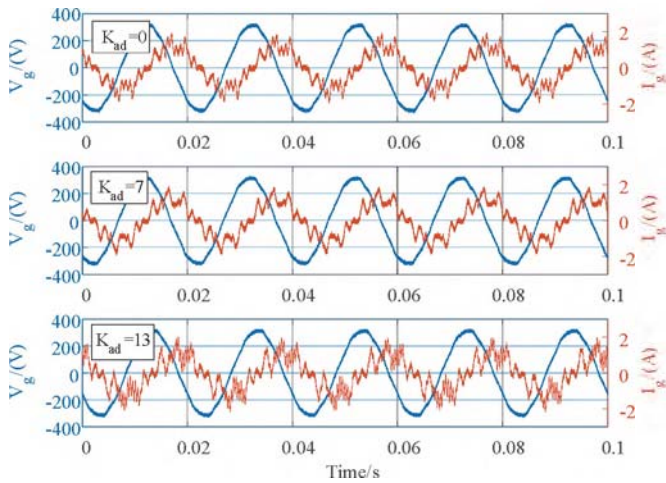


Fig. 13. Preliminary experimental results: Grid-side current under different  $K_{ad}$ .

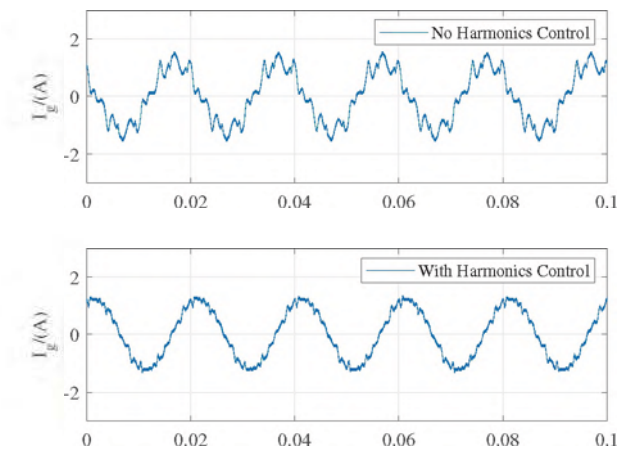


Fig. 14. Preliminary experimental results: Grid-side current with harmonics control under distorted grid voltage.

margin. However, in the final paper additional improved experimental results will be included addressing those differences. Therefore, future work will focus on more effective phase compensation methods and other harmonics control techniques i.e, multiple-axis control and full grid voltage feed-forward strategy.

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