

CURRENT SOURCE BASED ON H-BRIDGE INVERTER WITH OUTPUT LCL FILTER

Vojtech Blahnik^{*,**} — Jakub Talla^{*} — Zdenek Peroutka^{*,**}

The paper deals with a control of current source with an LCL output filter. The controlled current source is realized as a single-phase inverter and output LCL filter provides low ripple of output current. However, systems incorporating LCL filters require more complex control strategies and there are several interesting approaches to the control of this type of converter. This paper presents the inverter control algorithm, which combines model based control with a direct current control based on resonant controllers and single-phase vector control. The primary goal is to reduce the current ripple and distortion under required limits and provides fast and precise control of output current. The proposed control technique is verified by measurements on the laboratory model.

Key words: LCL output filter

1 INTRODUCTION

Many grid connected power electronic systems, such as Statcoms [1], UPFCs [2], and distributed generation system converters (*eg* renewable energy sources [3]), use a voltage source inverter connected to the supply network through a filter. This filter, typically a series inductance, reduces the switching harmonics entering to the distribution network. An alternative filter is a LCL, which can achieve reduced levels of harmonic distortion at lower switching frequencies and with less inductance value. This approach was massively deployed for three-phase power converter (for example [4]). However, single-phase power converter uses LCL filters rarely, because it requires more complex control strategies. Until now, several methods have been proposed in literature for the control of single-phase inverters. Hysteresis control and delta modulation controls are robust and simple, but its variable switching frequency expects special adaptation likes [5]. Widely used control technique is vector control based on simple PI controllers [6] and [7], but these methods show a very slow dynamic behaviour for single-phase system. Very promising control techniques are based on predictive controllers [8] and [9], this kind of controllers is high sensitive to the system model accuracy. Simple and very effective control method is direct current control based on PR controller [10]. They are capable to track sinusoidal references of arbitrary frequencies of both, positive and negative sequences, with zero steady state error.

The objective of this research was design of perspective control for single-phase controlled current source with output LCL filter for current ripple reducing. Designed control must precisely control generated current with arbitrary power factor in full power range. Furthermore it is necessary to eliminate the current distortions caused

by dead-times and other power converters non-linearity, see *eg* [11]. This paper describes in detail the enhanced control strategy and presents the experiments made on developed small-scale converter prototype with maximal output power 2 kVA. Tests are performed for the most problematic power converter conditions, such as are requirements for generating capacitive and inductive reactive current. The laboratory small-scale prototype consist of single-phase ac source ($v_{ac} = 230 \text{ V}_{\text{RMS}}$), symmetrical LCL filter ($L_1 = L_2 = 800 \mu\text{H}$, $R_{L1} = R_{L2} = 100 \text{ m}\Omega$, $C = 60 \mu\text{F}$), single-phase converter based on IGBT technology ($f_{\text{switch}} = 10 \text{ kHz}$, $t_{\text{deadtimes}} = 3.3 \mu\text{s}$) and dc source ($V_{dc} = 420 \text{ V}$). The current source power circuit is shown in Fig. 1.

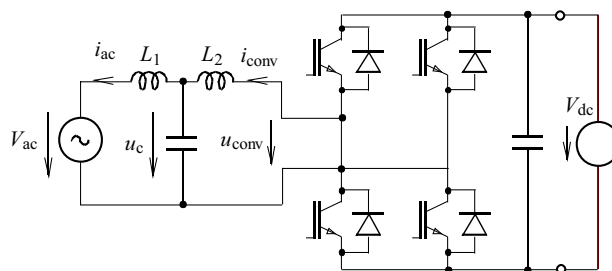


Fig. 1. Power circuit of controlled current source with LCL filter

2 CONVERTER CONTROL STRATEGY

The proposed control of current source must provide fast and well control of output current (i_{ac}) with a precisely defined phase shift against ac source (v_{ac}). That is the reason for using an accurate voltage synchronization method intended to single-phase systems. In this case,

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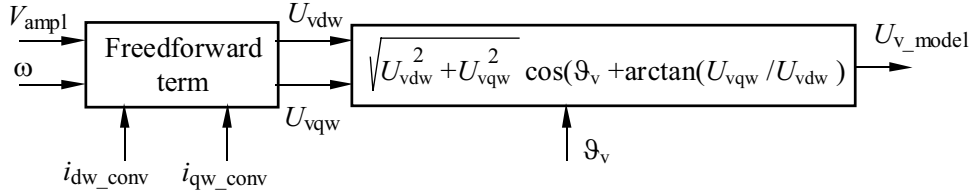


Fig. 2. Feedforward calculation — model part of the control structure

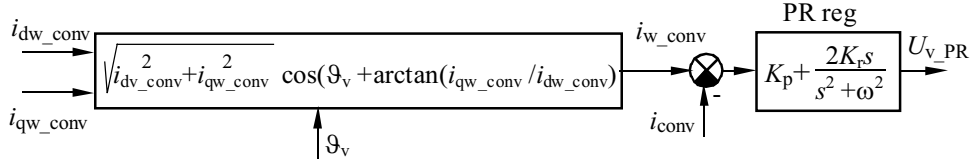


Fig. 3. Direct current control based on PR controller — feedback part of the control structure

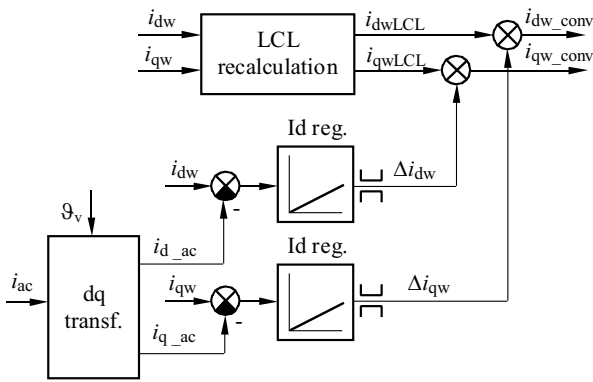


Fig. 4. Converter current components recalculation — compensation of LCL filter influence

the sliding DFT (discrete Fourier transformation) synchronization has been chosen. The output signals of DFT voltage synchronization are the position of voltage vector (V_{ac}) and the voltage amplitude (V_{ampl}). Signal V_{ampl} is used to calculate mathematical model of converter and signal ϑ_v is used across whole control structure.

The feedforward and direct current control were chosen to achieve very fast and precise current control. The feedforward compensation uses knowledge of ac source voltage components (ϑ_v and V_{ampl}), required converter current components (i_{dw_conv} , i_{qw_conv}) and simplified model of LCL filter (there is only calculation with inductance part L_1 and L_2 of the filter). The output signals (U_{vdw} and U_{vqw} , are in d, q virtual revolving references frame linked to space vector of ac voltage) are calculated by equations (1) and (2). The final feedforward term (U_{v_model}) is calculated by equation (3) as a shown in Fig. 2. The direct current control is implemented by proportional resonant (PR) controller with the pass-frequency $\omega = 2\pi 50$ Hz (pass-frequency is equal to the frequency of ac source), this type of controller is described in [12]. The controller provides control of converter current (i_{conv}) to required value (i_{w_conv}), which is calculated from requirements for converter current com-

ponents (i_{dw_conv} , i_{qw_conv}). This calculation is computed by equation (4) and the full direct current control structure with PR controller is illustrated in Fig. 3.

$$U_{vdw} = -\omega(L_1 + L_2)i_{qw_conv}, \quad (1)$$

$$U_{vqw} = V_{ampl} + \omega(L_1 + L_2)i_{dw_conv}, \quad (2)$$

$$U_{v_model} = \sqrt{U_{vdw}^2 + U_{vqw}^2} \cos(\vartheta_v + \arctan \frac{U_{vqw}}{U_{vdw}}), \quad (3)$$

$$i_{w_conv} = \sqrt{i_{dw_conv}^2 + i_{qw_conv}^2} \cos(\vartheta_v + \arctan \frac{i_{qw_conv}}{i_{dw_conv}}). \quad (4)$$

To achieve precise control of output current (i_{ac}) it is important to compensate the auxiliary effects of LCL filter. In this case, voltage losses and output current phase shift are compensated. The compensation of LCL filter is realized by using of forward LCL model calculation and by simplified vector control. The mathematical model inaccuracies are compensated by two integral controllers (Id reg. and Iq reg.) for active and reactive current part (Δi_{dw} and Δi_{qw}). The mathematical model of LCL filter is based on two basic equations (5) and (6). By subsequent modifications and transformation into d, q “virtual” revolving system are obtained the final equations (7) and (8). These equations are used for the conversion of output currents requirements (i_{dw} , i_{qw}) to feedforward currents requirements (i_{dwLCL} , i_{qwLCL}). The final converter currents requirements are signals i_{dw_conv} and i_{qw_conv} . These signals are calculated by sum of feedforward currents i_{dwLCL} and i_{qwLCL} with feedback signals Δi_{dw} and Δi_{qw} from integral controllers. The resulted signals (i_{dw_conv} and i_{qw_conv}) are requirements for converter current i_{conv} , as illustrated in Fig. 4. The dq transformation block uses principle for the fast decomposition (this method is not resistant against disturbances) like [8].

$$i_{conv} = i_{ac} + C \frac{du_c}{dt}, \quad (5)$$

$$u_c = V_{ac} + R_1 i_{ac} + L_1 \frac{di_{ac}}{dt}, \quad (6)$$

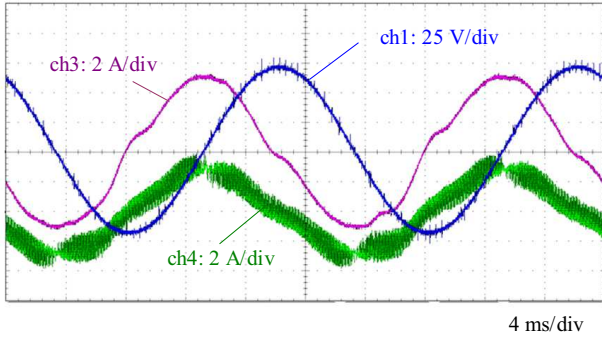


Fig. 5. Current source steady-state — without harmonic compensation (current components: $i_{dw} = 5$ A, $i_{qw} = 0$ A), ch1: ac source voltage v_{ac} [25 V/div], ch3: output current i_{ac} [2 A/div], ch4: current on converter side i_{conv} [2 A/div]

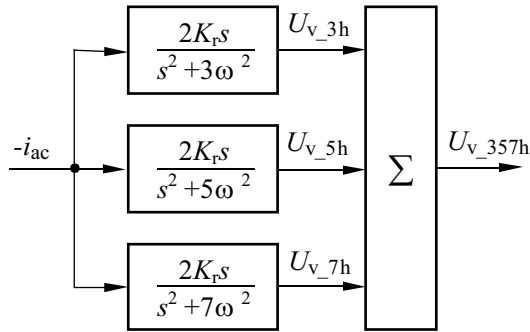


Fig. 6. Harmonic compensation — compensation by three resonant controllers (150 Hz, 250 Hz, 350 Hz)

$$i_{qwLCL} = \omega R_1 C i_{dw} + (1 - \omega^2 L_1 C) i_{qw} \quad (8)$$

The behavior of classical IGBT inverter is nonlinear including dead time effects influences and current dependent nonlinear voltage drops. It leads to output currents distortion as shown in Fig. 5. These distortions appear as a characteristics harmonics components at output current spectra. For that reason harmonics components are compensated by resonant controllers with selective pass frequency (150 Hz, 250 Hz and 350 Hz) as presented in Fig. 6. More information about low-frequency harmonic compensation by resonant controllers was published *eg* in [13].

The final designed control algorithm is composed of these five parts: voltage synchronization, feedforward calculation, compensation of LCL filter influence, harmonic compensation and direct current control. The final modulation signal (U_{PWM}) is input signal for PWM. This UPWM signal is a sum of three signals U_{v_model} , U_{v_357h} , U_{v_PR} . The complete block control diagram for controlled current source is depicted in Fig. 7. The zero vectors alternating method described in [11] for PWM modulation is used. This PWM method ensures double of current ripple frequency with the comparison to switching frequency of IGBTs.

3 EXPERIMENTAL RESULTS

The current source behaviour was tested on small-scale converter prototype with output power 2 kVA. The power circuit of this prototype is shown in Fig. 1. Converter is

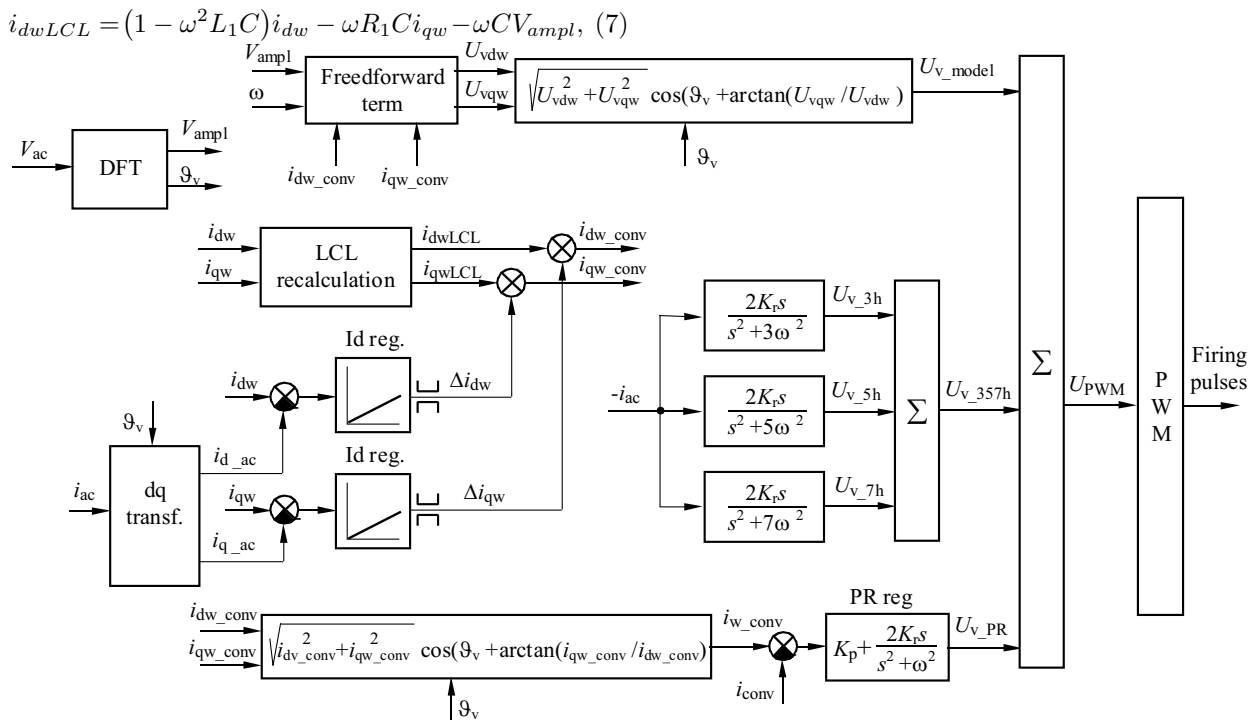


Fig. 7. Designed control for controlled current source with output LCL filter

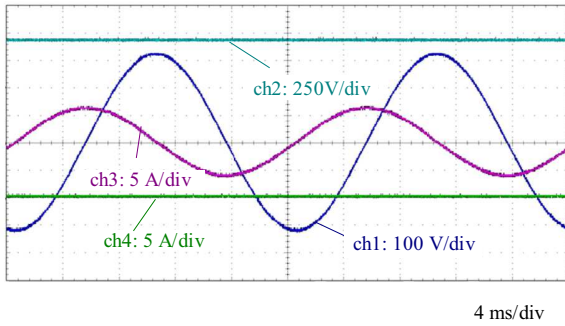


Fig. 8. Current source behaviour under blocking firing signals, ch1: ac source voltage v_{ac} [100 V/div], ch2: dc voltage v_{dc} [250 V/div], ch3: output current i_{ac} [5 A/div], ch4: current on converter side i_{conv} [5 A/div]

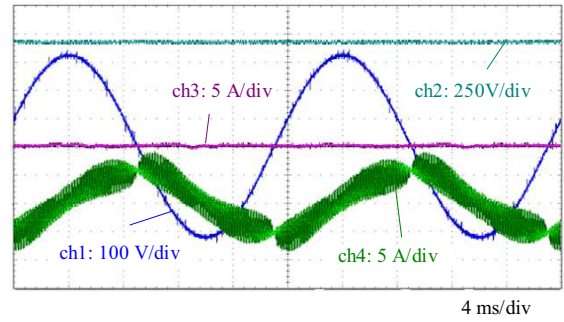


Fig. 9. Current source under steady-state (current components: $i_{dw} = 0$ A, $i_{qw} = 0$ A), ch1: ac source voltage v_{ac} [100 V/div], ch2: dc voltage v_{dc} [250 V/div], ch3: output current i_{ac} [5 A/div], ch4: current on converter side i_{conv} [5 A/div]

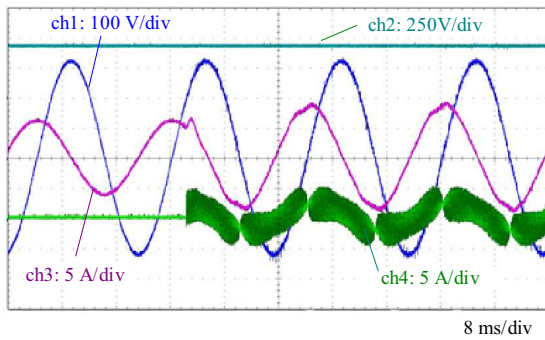


Fig. 10. Current source start-up (current components: $i_{dw} = 9$ A, $i_{qw} = 0$ A), ch1: ac source voltage v_{ac} [100 V/div], ch2: dc voltage v_{dc} [250 V/div], ch3: output current i_{ac} [5 A/div], ch4: current on converter side i_{conv} [5 A/div]

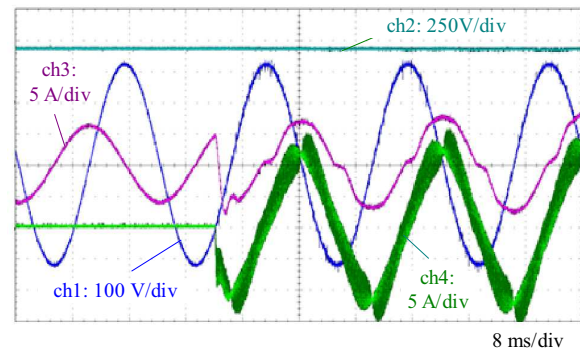


Fig. 11. Current source start-up (current components: $i_{dw} = -9$ A, $i_{qw} = 0$ A), ch1: ac source voltage v_{ac} [100 V/div], ch2: dc voltage v_{dc} [250 V/div], ch3: output current i_{ac} [5 A/div], ch4: current on converter side i_{conv} [5 A/div]

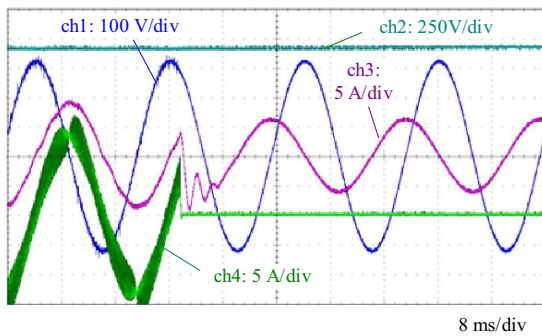


Fig. 12. Current source shutdown (current components: $i_{dw} = -9$ A, $i_{qw} = 0$ A), ch1: ac source voltage V_{ac} [100 V/div], ch2: dc voltage v_{dc} [250 V/div], ch3: output current i_{ac} [5 A/div], ch4: current on converter side i_{conv} [5 A/div]

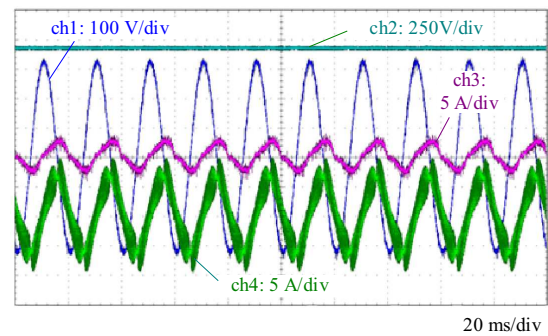


Fig. 13. Current source steady-state for low value of current (current components: $i_{dw} = -1$ A, $i_{qw} = 0$ A), ch1: ac source voltage V_{ac} [100 V/div], ch2: dc voltage v_{dc} [250 V/div], ch3: output current i_{ac} [2 A/div], ch4: current on converter side i_{conv} [5 A/div]

fed by dc source ($V_{dc} = 420$ V) and the output LCL filter is connected directly to terminals of ac source ($v_{ac} = 230$ V). The converter current is rippling with 20 kHz for 10 kHz switching frequency (caused by zero vector alternating modulation) and output current (i_{ac}) is almost without ripple (causing by correct function of LCL filter). The current source control has been implemented in the fixed-point digital signal processor Texas Instruments TMS320F2812 with sampling frequency 50 μ s (that is absolutely sufficient computing time for control algorithm implementation).

Figure 8 presents converter behaviour during connection to ac source (ch1-ac source voltage) without switching (converter firing signals are blocked). The converter current (ch4 i_{conv}) is zero, but the output current (ch3 i_{ac}) reach the value around 5 A. If the output current is required zero the converter must generates approximately the same value of converter current with opposite polarity as it is shown in Fig. 9. For required reactive current $i_{dw} = 9$ A, the start-up sequence of converter and steady state is shown in Fig. 10. The converter transient is very fast, but control algorithm spends more time for current distortion elimination. These current deformations

are removed by resonant controllers (harmonic compensation part). The problematic state is captured in Fig. 11 where the start-up sequence of converter for current requirement $i_{dw} = -9$ A is presented. In this case, the current distortion elimination takes several periods and stabilization (steady-state) is shown in Fig. 12. After converter shutdown, the current oscillation is very well visible (current $i_{ac, ch3}$ in Fig. 12). The oscillation frequency is equal to natural frequency of LCL filter. The most problematic converter states are small currents requirements (from 0 A to 2 A). The main problem came from current measurement accuracy, converter non-linearity and current control response. Converter behaviour under requirements for 1 A is depicted in Fig. 13.

4 CONCLUSION

The paper presents new control structure of current source with LCL output filter. The output LCL filter provides low output current (i_{ac}) ripple compared with converter current (i_{conv}). However, systems incorporating LCL filters require more complex control strategies. The paper describes control algorithm based on structure with a feedback direct current control based on resonant controllers and feedforward single-phase vector control as a LCL filter compensator. Designed control provides fast dynamic control of LCL filter output current and resonant controllers compensates 3th, 5th, and 7th harmonics of the output current caused by nonlinear behavior of IGBT converter. The proposed control technique is verified by experiments made on developed small-scale converter prototype with 2 kVA output power.

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