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Active Power Filter with a Reduced Number of Current Sensors

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Abstract. In this paper, a modification of a parallel active power filter (APF) control is presented. Its purpose is to decrease the cost of the hardware by minimizing the number of current sensors, normally present in conventional topologies. The approach employs the direct current control (DCC), a method for predictive control of currents in polyphase systems, developed by the authors. Since in traditional filter topologies, currents for compensation of distorted load currents are produced by an inverter-like circuit, DCC, as in similar approaches, is used for filter current generation. In this particular case, a variant of DCC with a modulated voltage during a sampling interval is used. However, following a relatively simple algorithm, only a grid current can be used in DCC, without any need for filter and load current measurement, thus eliminating the necessity of installing corresponding current sensors. In addition, robustness of the approach was proved by examining the effect of alteration of the filter model parameters. Simulation results show validity of the proposed method.

Keywords: active power filter, direct current control, current sensors, modulation

Aktivni močnostni filter z zmanjšanim številom tokovnih merilnikov

Povzetek. V članku je predstavljena modifikacija vodenja paralelnega aktivnega močnostnega filtra (APF), ki zniža stroške aparaturne opreme zaradi manjšega števila potrebnih tokovnih sond. Pristop temelji na uporabi metode za neposredno regulacijo toka (angl. direct current control – DCC), namenjene za prediktivno regulacijo tokov v večfaznih sistemih. Tokove za kompenzacijo popačenih bremenskih tokov generira aktivni filter, ki ima razsmerniško topologijo, zato je za regulacijo filtrskega toka možno uporabiti metodo DCC, ki se je že izkazala pri drugih aplikacijah s podobno topologijo. V predstavljenem primeru je uporabljena različica metode DCC, kjer se napetost modulira znotraj intervala vzorčenja. V predlaganem pristopu lahko za DCC zaradi njenega prediktivnega značaja in poznavanja parametrov filtrske veje uporabimo samo omrežne tokove. Zato merjenje filtrskih tokov in pripadajoče tokovne sonde niso več potrebni. V nadaljevanju prispevka je opravljena še analiza robustnosti metode glede na vpliv spremembe parametrov filtrskega modela. Simulacijski rezultati potrjujejo veljavnost predlaganega pristopa.

Ključne besede: aktivni močnostni filter, neposredna regulacija toka, tokovni merilniki, modulacija

1 Introduction

Strict norms imposed on quality of the electrical power supplied by the grid are challenged by an increasing usage of switching power devices in industrial applications and home appliances. As a consequence, different harmonics are present in the current drawn from the grid both due to the converter switching actions and/or the non-linear character of the load. An additional problem is caused by unbalanced loads in the three-phase systems. A very effective way of compensating for these distortions is by using shunt (parallel) active power filters (APF). They are basically power electronics, inverter-like devices that act as high-dynamic current sources. These filter currents, added to the load currents, cause the line supply currents to be sinusoidal and in phase with the supply voltages [1-6]. It is needless to say that control of APFs requires fast on-line tracking of actual currents and fast adaptation according to the implemented algorithm.

In this paper, a method for controlling the three-phase filter currents, based on the previous work by the authors [7-11], is presented. Its main advantage is in reduction of the number of current sensors present in traditional APF topologies, thus decreasing the total cost of the hardware. Therefore, instead of mounting six current sensors, only three are necessary. Still, the current generation algorithm remains fast, thus providing excellent filter current dynamics.

2 Active power filter

Figure 1 shows a typical topology of a parallel APF [3]. Independently generated currents in an APF are added to each of the three line phases. Their purpose is to compensate for load harmonics, fundamental reactive power and possible phase unbalances (e.g. loss of one phase, etc.). Unlike the classical scalar approaches, the vector approach deals with vector sums of currents in all phases – space vectors. Therefore, the current vector equation for the filter space vector is

$$\mathbf{i}_F = \mathbf{i}_S - \mathbf{i}_l \tag{1}$$

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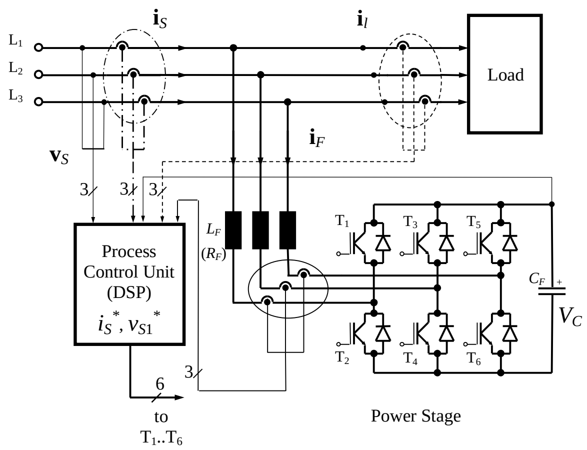

Figure 1. Topology of a three-phase, three-wire parallel active power filter.

Filter currents must instantly adapt themselves to the anomalies on the load side so as to produce the grid phase currents $\mathbf{i}_S$ to be sinusoidal and in phase with the grid voltage, containing only the fundamental harmonic. Consequently, from the grid side, the load behaves as being a pure resistive one.

In the classical approach, in order to fully perform the task of compensating harmonics and phase unbalances, line voltages as well as filter and load currents have to be measured in each phase with an additional measurement of the filter capacitor voltage. Filter currents measurement is also necessary for their effective control through the filter three-phase bridge voltage source inverter (VSI). The capacitor acts like a pseudo-constant voltage DC source that is chopped by the transistors, thus impressing the desired current.

Of course, the amount of the filter compensation current is dictated by the desired line current, determination of which forms a very important part of the APF algorithm. According to the above mentioned demands, the load-filter circuitry should behave like a resistive load with conductance

$$G_{(1,2,3)}(t) = \frac{i_{S(1,2,3)}(t)}{v_{S1(1,2,3)}(t)}\,. \tag{2}$$

To provide a sinusoidal waveform of supply line current $\mathbf{i}_S$, it is reasonable to take into account the supply line voltage fundamental $\mathbf{v}_{S1}$, since the line voltage is usually heavily distorted. It has been shown, that the line voltage fundamental can be calculated on-line without any special synchronization to the grid [3, 8].

Since there should be no active power in the filter branch (except for covering losses) and the supply line currents should be balanced, symmetrical substitutive conductance is calculated [7] as:

$$G(t) = G_1(t) = G_2(t) = G_3(t) = $$
$$= \frac{\int_{t-T^*}^{t} (v_{S(1)} i_{l(1)} + v_{S(2)} i_{l(2)} + v_{S(3)} i_{l(3)})\, d\tau}{\int_{t-T^*}^{t} (v_{S(1)}^2 + v_{S(2)}^2 + v_{S(3)}^2)\, d\tau}\,. \tag{3}$$

Sinusoidal line current reference $\mathbf{i}_S^*$ is now obtained easily:

$$i_{S(1,2,3)}^*(t) = v_{S1(1,2,3)}(t) \cdot G(t) \quad \text{or} \quad \mathbf{i}_S^* = \mathbf{v}_{S1} \cdot G\,. \tag{4}$$

In [7], filter current reference $\mathbf{i}_F^*$ was further calculated, because the actual filter current $\mathbf{i}_F$ was measured and controlled directly. Moreover, as it is evident from (3), the load current ($\mathbf{i}_l$) measurement was required. The advanced method, proposed in this paper, should rely only upon the measurement of supply line current $\mathbf{i}_S$. Assuming fast operating current control, one can expect the actual line currents to be sinusoidal. Therefore, the substitutive conductance becomes

$$G(t) = \frac{\int_{t-T^*}^{t} (v_{S(1)} i_{S(1)} + v_{S(2)} i_{S(2)} + v_{S(3)} i_{S(3)})\, d\tau}{\int_{t-T^*}^{t} (v_{S(1)}^2 + v_{S(2)}^2 + v_{S(3)}^2)\, d\tau}\,. \tag{5}$$

However, the supply line current amplitude must be additionally controlled to provide filter capacitor voltage V_C to be constant. Thus, the superior control of the active power flow from the supply grid is obtained. It can be carried out simply by the filter capacitor voltage error ($V_C^* - V_C$) slightly increasing the substitutive conductance G. Certainly, the negative value of the error (when the capacitor voltage V_C is too high) decreases the value of G and consequently reduces the active power flow from the supply grid. Some filter control methods rely solely on this approach [2], thus further reducing the number of sensors needed and providing a moderate dynamic response.

After comparing (5) and (3), we can now conclude that the load current measurement becomes obsolete (dashed lines in Figure 1). Instead, line currents have to be measured (dashed-dotted lines), without any gain in cost reduction. In order to achieve this goal, the task of the presented approach will be to eliminate the filter currents measurement needed for current control (solid lines in Figure 1).

As already explained, the main control variables of a parallel APF are the filter currents or their resulting space vector. In the presented approach, these currents are controlled using direct current control (DCC).

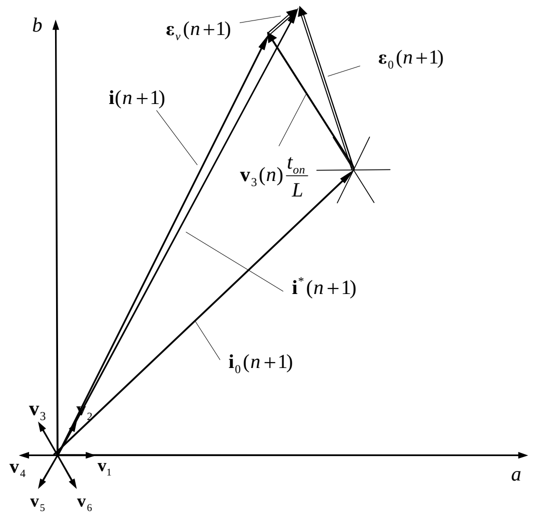

Figure 2. Principle of current generation using DCC II.

3 Direct current control

Direct current control is a predictive method developed for the fastest possible current control. The method was also altered for stator flux control that is more suitable for application in electrical drives, where back EMF occurs [10].

DCC is ideal for passive *R-L* three-phase loads where only the inverter voltage is impressed. However, it works equally well if additional voltages (e.g. back EMF, additional sources) are present, providing their easy detection/measurement. Besides the high dynamics, minimization of the current ripple is of a great concern. However, this goal is achieved on account of a higher switching frequency of inverter transistors leading to higher switching losses. For this purpose, two variants have been proposed; one suitable for loads with lower dynamics (higher time constants) yielding low switching losses, and the other with a reduced current ripple but with higher switching frequency (suitable for high dynamic loads). In the presented paper, only the second variant (DCC II) will be examined.

Due to the discrete nature of the microprocessor used for control, the basic current equation for a three-phase *R-L* load, when fed by voltage **v**, has to be written in its discretized form. Thus, the current at the end of sampling interval Δt can be generally written as [8]

$$\mathbf{i}(n+1) = \underbrace{\mathbf{i}(n)\left(1-\frac{R}{L}\Delta t\right)}_{i_0(n+1)} + \mathbf{v}(n)\cdot\frac{\Delta t}{L} \tag{6}$$

Note that currents and voltages are represented by space vectors, which are spatial sums of temporal phase values. Voltage vector **v** depends on the type of the supply. In modern three-phase systems, a transistor full bridge is usually used, which gives a possibility of generating six distinct active voltage space vectors [8]. If the load has a passive character, this is the only voltage applied. However, in a variety of applications, additional voltage pseudo sources can be present, e.g. back EMF in electrical drives. The goal of DCC is to minimize the error between the reference current (i.e. supply line current reference in (4)) and the actual current (6) at the end of the sampling interval.

$$\boldsymbol{\varepsilon}_0(n+1) = \mathbf{i}^*(n+1) - \mathbf{i}\ (n+1) \tag{7}$$

Figure 2 shows relations between space vectors. Since the current vector at instant (n+1) hardly matches the desired (reference) current in direction and size when one of the six active voltage vectors is applied, DCC II offers a possibility of minimizing the initial current error ε_0. From (6), this task is performed by first selecting the voltage vector pointing into direction of the current error ($\mathbf{v}_3$ in Figure 2) and then calculating the optimal time t_{on}, in which the active vector should be applied, yielding the final current error ε_v. The approach was explained in detail in [8-11].

4 APF with a reduced number of current sensors

According to Figure 1, equation (6) applied to the active power filter can be re-formulated as

$$\mathbf{i}_F(n+1) = \underbrace{\mathbf{i}_F(n)\left(1-\frac{R_F}{L_F}\Delta t\right) + \mathbf{v}_S(n)\cdot\frac{\Delta t}{L_F}}_{\mathbf{i}_{F0}(n+1)} + \mathbf{v}_F(n)\cdot\frac{t_{on}}{L_F} \tag{8}$$

Here, the impressed voltage (**v** in (6)) consists of two parts: the first part is sinusoidal grid voltage $\mathbf{v}_S$ applied throughout entire sampling interval Δt, whereas the second part is filter (inverter) voltage $\mathbf{v}_F$, through which the filter current is to be controlled. The latter is applied throughout pre-calculated interval t_{on}. Although changing, the grid voltage is measurable and predictable. Therefore, it will be aggregated to the first part of (8) under the term denoted as $\mathbf{i}_{F0}(n+1)$.

Now, the first simplification can be made. The right term in brackets containing an *R-L* time constant represents a natural decrease of the filter current. Due to the short microprocessor sampling interval Δt, low filter resistance R_F and relatively high filter inductance L_F, it can be neglected (in experimental setup: Δt = 78.125 µs, R_F = 90 mΩ and L_F = 2.6 mH yield $0.0027\cdot\mathbf{i}_F(n)$).

An additional simplification, crucial for the elimination of the filter current sensors, is introduced by presuming that the change of load current $\mathbf{i}_l$ during the sampling interval is negligible due to the usually much higher load time constant vs. short sampling interval. Therefore, from (1):

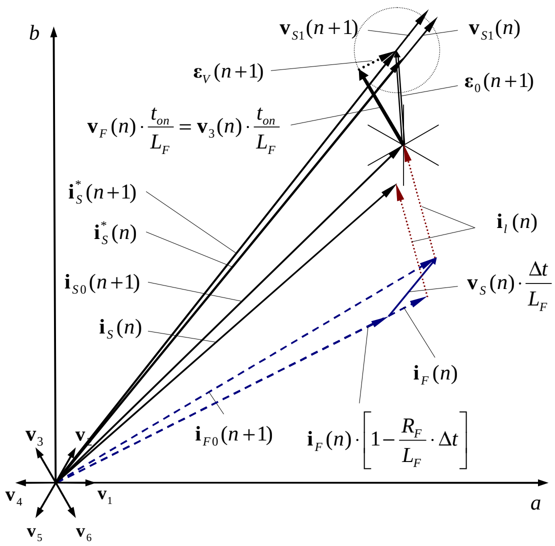

Figure 3. DCC approach with a modulated active voltage vector minimizes the predicted current error $\boldsymbol{\varepsilon}_V(n+1)$.

$$
\begin{aligned}
\mathbf{i}_F(n) &= \mathbf{i}_S(n)-\mathbf{i}_l(n)\\
\mathbf{i}_F(n+1) &\approx \mathbf{i}_S(n+1)-\mathbf{i}_l(n)
\end{aligned}
\tag{9}
$$

After inserting (9) into (8), an important conclusion can be drawn: the value of $\mathbf{i}_{S0}(n+1)$ can be predictably controlled without measuring the filter current:

$$
\mathbf{i}_S(n+1)=\underbrace{\mathbf{i}_S(n)+\mathbf{v}_S(n)\cdot\frac{\Delta t}{L_F}}_{\mathbf{i}_{S0}(n+1)}+\mathbf{v}_F(n)\cdot\frac{t_{on}}{L_F}
\tag{10}
$$

The first part of (10) – $\mathbf{i}_{S0}$, like in (8), is a term that can be established at the beginning of the sampling interval through the measurement of grid currents and voltages. If there is no filter voltage impressed ($t_{on}=0$), this will be the final supply line current at the end of the interval, producing an error

$$
\boldsymbol{\varepsilon}_0(n+1)=\mathbf{i}_S^*(n+1)-\mathbf{i}_{S0}(n+1)
\tag{11}
$$

with respect to the reference current calculated from (4), as shown in Figure 3.

It is now obvious that the supply line currents can be controlled directly by the filter voltage vectors, necessitating no filter current measurement. Therefore, if an active filter voltage vector is impressed during pre-calculated interval t_{on}, the final predicted supply line current error becomes

$$
\boldsymbol{\varepsilon}_V(n+1)=\underbrace{\mathbf{i}_S^*(n+1)-\mathbf{i}_{S0}(n+1)}_{\boldsymbol{\varepsilon}_0(n+1)}-\mathbf{v}_F(n)\cdot\frac{t_{on}}{L_F}
\tag{12}
$$

Among six possible filter (inverter) active voltage space vectors to form $\mathbf{v}_F$, the one pointing nearest to the direction of the initial pre-calculated error $\boldsymbol{\varepsilon}_0(n+1)$ should be chosen. As described in [7, 11], after adapting the nomenclature to its application in APF, voltage application interval t_{on} is calculated as

$$
t_{on}=\frac{9L_F}{4V_C}\left(\varepsilon_{0a}(n+1)\cdot\frac{v_{Fa}}{V_C}+\varepsilon_{0b}(n+1)\cdot\frac{v_{Fb}}{V_C}\right)
\tag{13}
$$

Vectors $\mathbf{v}_F$ and $\boldsymbol{\varepsilon}_0$ are decomposed into their components in an a–b stationary reference frame. L_F and V_C are the filter inductance and voltage across the capacitor, respectively (Figure 1).

5 Results

The proposed filter control with the advanced DCC method was simulated with the filter parameters $C_F = 1000\ \mu\text{F}$, $L_F = 2.6$ mH and $R_F = 90\ \text{m}\Omega$. The supply grid line-to-neutral voltage V_S was 230 V (RMS). Filter capacitor voltage V_C was controlled at 720 V. As a nonlinear load a single-phase thyristor rectifier was used connected between two phase terminals, thus simulating an extremely unbalanced load. Figure 4 shows the supply line voltages and load currents, where unbalance is obvious since there is no current in phase 3 of the nonlinear load. Harmonic distortion of load currents is significant (THD = 21.4%).

In Figure 5, the filter branch currents and compensated supply line currents are presented. Since there is no current in phase 3 of the load, the current in phase 3 of the filter branch takes a sinusoidal form and actually represents the supply line current in phase 3. As the compensated supply line currents are sinusoidal (harmonic distortion is reduced to THD = 2.6%) and synchronized with the corresponding supply line voltages, the fundamental reactive power is successfully compensated as well.

Figure 6 presents the trajectory of compensated supply line current $\mathbf{i}_S$ for steady-state conditions within one grid cycle (20 ms). It is clearly shown how individual active/zero filter voltage vectors $\mathbf{v}_F$ are activated. The circular shape of the trajectory meets the requirements.

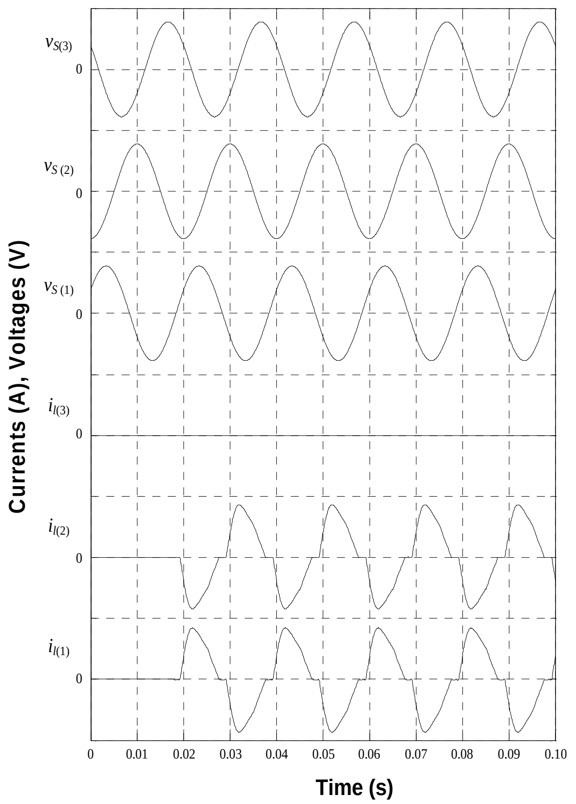

Figure 4. Supply line voltages ($v_{S(1)}$, $v_{S(2)}$, $v_{S(3)}$) and load currents ($i_{l(1)}$, $i_{l(2)}$, $i_{l(3)}$) during the load transient (k_v = 400 V/div, k_i = 50 A/div).

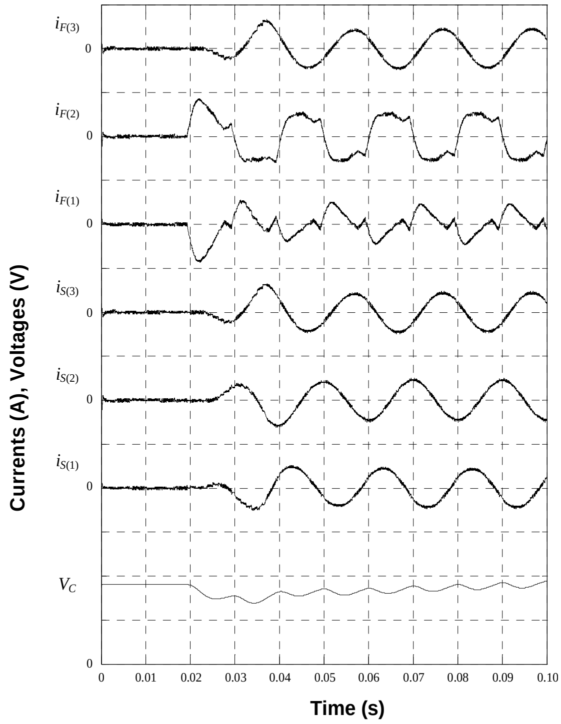

Figure 5. Filter currents ($i_{F(1)}$, $i_{F(2)}$, $i_{F(3)}$), compensated supply line currents ($i_{S(1)}$, $i_{S(2)}$, $i_{S(3)}$) and filter capacitor voltage (V_C) during the load transient (k_v = 400 V/div, k_i = 50 A/div).

6 Sensitivity to parameter variations

Although the initial values of the parameters are known, the main problem arises from their possible alteration during the filter operation. The filter inductance and resistance are the only parameters that can be altered while not being directly measured.

Inductance can change due to saturation and resistance due to temperature. Although these effects are not as drastic as those in electrical drives, it is still interesting to observe their possible impact on the presented method. Figure 7 shows the performance of the filter with its model parameters exceeding the actual ones by 20% (e.g. in saturation). As it can be seen when compared to Figure 6, the method still proves to be very robust, showing practically no difference.

7 Conclusion

In this paper, an approach to predictive current control in APF is presented. Unlike the standard topologies with load and filter currents measurement, the proposed one requires only measurement of line currents through corresponding sensors, thus reducing the hardware cost. A fairly precise calculation of conductance, needed for line reference current determination, can also be provided through the line current measurement. Moreover, instead of directly controlling the filter currents, line currents can be indirectly controlled, providing a fast and accurate current control algorithm. For this purpose, the already verified method, called the direct current control, was applied. The issue of parameter variation during filter operation was also addressed, showing no significant

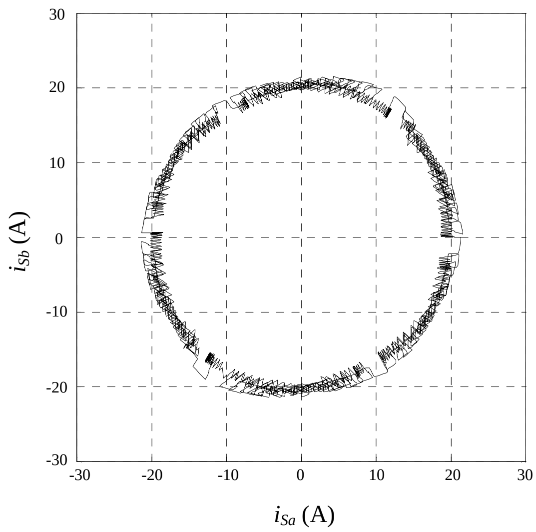

Figure 6. Orthogonal components of compensated supply line currents (i_{Sa}, i_{Sb}) in steady-state during one grid cycle (20 ms).

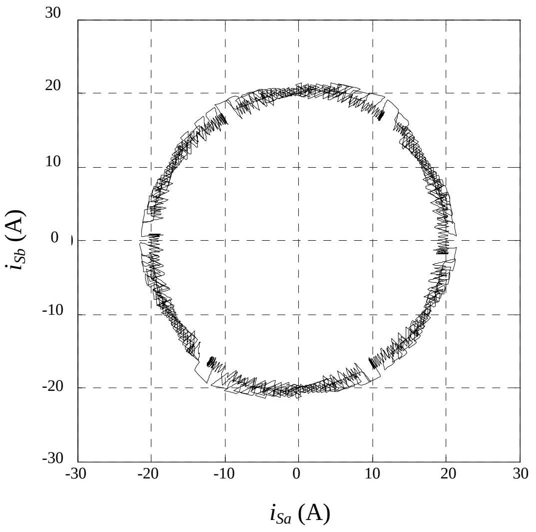

Figure 7. Orthogonal components of compensated supply line currents (i_{Sa}, i_{Sb}) in steady-state during one grid cycle (20 ms) for $L_F^{\text{model}} = 1.2\ L_F^{\text{actual}}$.

impact on the performance of the method. Our future work in this field will be focused on confirming the obtained results on a specially adapted experimental setup.

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