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Real-Time Grid Impedance Estimation Using a Converter

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Abstract—This paper presents two real-time methods for estimating the grid inductance and resistance using a three-phase converter. In the methods, a single-frequency rotating voltage signal with a small magnitude is injected to the converter voltage reference. The converter measures the currents and voltages at the point of common coupling (PCC). The grid impedance is estimated from the measured data using two alternative methods: an estimator based on the sliding discrete Fourier transform (SDFT) and an adaptive grid observer. Dynamic performance of these methods is compared by means of simulations and experiments in the case of a changing grid impedance.

Index Terms—Adaptive observer, grid converter, impedance estimation, sliding DFT.

I. INTRODUCTION

Grid converters are widely used as interfaces between renewable energy sources and the grid. In this interface, impedances play an important role since the ratio of the converter output impedance to the grid impedance determines the stability of the grid converter system [1]. Furthermore, the grid impedance affects the converter control performance [2]. Thus, its estimate can be used to optimize control tuning [3]. Due to the local nature of the grid impedance, its estimate can also be applied to detect the status of the grid, such as an islanding condition [4], [5].

Several methods for estimating the grid impedance have been proposed, see [3]–[20] and the papers cited therein. For the impedance estimation, grid converters have been utilized to inject an excitation signal to the grid [3]–[16]. On the contrary, the impedance has been estimated passively using existing power system disturbances [17], [18]. From these approaches, the active injection-based methods are more deterministic. In the estimation, the discrete Fourier transform (DFT) is often utilized [5]–[11], [20]. Although the DFT can be recursively calculated in real time with the sliding DFT (SDFT) [21], impedance estimators based on the SDFT have not been presented.

As an alternative to the DFT, model-based impedance estimation methods have been proposed in [3], [15]–[19]. In addition, a method combining the DFT and system models exist [14]. If the operating point of the converter changes during the estimation, DFT-based estimation may become biased due to spectral leakage [6]. On the contrary, model-based approaches can tolerate [18] or even require [3], [15] operating point changes. However, the model-based methods are more difficult to tune than the DFT-based methods [3], [18].

This paper presents an SDFT-based impedance estimator and an adaptive grid observer for real-time grid impedance estimation. The SDFT-based impedance estimation method is derived combining the method in [21] with the state-of-the-art impedance estimation methods. In the observer-based method, an inductive-resistive grid model is used. The nonlinear estimation-error dynamics of the observer are linearized and analyzed. Based on the analysis, simple tuning rules are developed. Finally, these two methods are compared through simulations and experiments.

II. SYSTEM MODEL

Fig. 1 shows a control system for a three-phase converter, augmented with real-time grid impedance estimation. The estimation feature is considered as a module that can be plugged in to an existing control system. The converter is connected to the grid at the point of common coupling (PCC), and the grid is modeled as a voltage source e_g^s , inductance L_g , and resistance R_g . The voltage at the PCC is $u_g^s = u_{g\alpha} + ju_{g\beta}$, where the superscript s refers to the stationary reference frame. A dynamic model of the inductive-resistive grid is

$$L_g \frac{di_g^s}{dt} = u_g^s - R_g i_g^s - e_g^s \quad (1)$$

where i_g^s is the grid current. The voltage e_g^s is unknown, but it is modeled as a rotating vector $e_g^s = e_g e^{j\omega_g t}$, where e_g is the magnitude and ω_g is the angular frequency. Harmonic components are considered as disturbances. A dynamic model for the grid voltage is

$$\frac{de_g^s}{dt} = j\omega_g e_g^s \quad (2)$$

III. SDFT-BASED IMPEDANCE ESTIMATOR

A straightforward way to estimate the grid impedance Z_g at a single frequency ω_e at a time is to inject an excitation signal v_e^s to the converter voltage reference $u_{c,ref}^s$, collect the PCC voltage u_g^s and current i_g^s samples to a buffer, and carry out the DFT on the data. Then, the impedance is obtained as the ratio of the voltage and current spectral components at the injection frequency [5], [6], [20]. The excitation signal can be pulsating or rotating [4]. In this paper, a rotating signal $v_e^s = v_e e^{j\omega_e t}$ with the magnitude of v_e is selected to excite all three phases evenly. The injection frequency is an interharmonic, and it is assumed that the injection-frequency component is not present in the grid voltage or current spectrum before the injection.

Fig. 1. Online grid impedance estimation as a part of the converter control system.

Fig. 2. (a) SDFT-based impedance estimator. A low-pass filter (LPF) is used to reduce measurement noise. (b) Internal structure of the SDFT module.

The estimation result is then more reliable than in the case of an unknown injection-frequency component pre-existing in the spectrum [5]. If needed, impedances at harmonic frequencies can be obtained by interpolating [8]. Instead of voltage injection, controlled single-frequency current injection could be used, if the existing control system is able to control the injection-frequency current.

The DFT can be calculated recursively in real time with SDFT methods [21]. An SDFT-based impedance estimator is illustrated in Fig. 2(a) and an SDFT module in Fig. 2(b). The SDFT requires a buffer of N samples and resonators for each frequency component of interest. In the case of single-frequency excitation, only one resonator per SDFT module is needed to extract the injection-frequency voltage component $u_{g,e}^s$ and current component $i_{g,e}^s$ from the PCC voltage u_g^s and current i_g^s . The structure can be augmented for simultaneous multi-frequency estimation by adding a resonator per an additional injection frequency in parallel with the first resonator.

To parametrize the SDFT module, the following steps are needed:

- 1) Frequency resolution f_{res} is selected so that the grid fundamental frequency f_g is integer multiple of f_{res} .
- 2) Buffer length is calculated as $N = 1/(f_{res}T_s)$, where T_s is the sampling period.
- 3) Harmonic number of the injection frequency with respect to the frequency resolution is calculated $h = \omega_e/(2\pi f_{res})$. If the number is non-integer either the injection angular frequency ω_e or the frequency resolution has to be re-selected.
- 4) Resonator coefficient is calculated as $c_e = \exp(-j2\pi h/N)$.

In this paper, the resolution of $f_{res} = 10$ Hz is selected. It enables injection frequencies multiple of 10 Hz. Therefore, spectral leakage of the fundamental frequency component and its harmonics is avoided. Other possible choices could be, e.g., 5 Hz and 25 Hz for the 50-Hz grid. The former choice increases the buffer size and the latter restricts more the possible injection frequencies. Since the buffer size is N samples, feasible information at the SDFT output is obtained after $t = NT_s = 1/f_{res}$, which can be considered as a settling time of the SDFT. A low-pass filter (LPF) can be added after the SDFT stage to reduce the effect of measurement noise on the estimates.

The estimated impedance $\hat{Z}_g$ is complex-valued. If an inductive-resistive grid is assumed, the resistance estimate $\hat{R}_g$ and the inductance estimate $\hat{L}_g$ are obtained from the real and imaginary parts of $\hat{Z}_g$, respectively. However, the SDFT-based impedance estimator is not restricted to this assumption.

IV. ADAPTIVE GRID OBSERVER

Fig. 3 shows an adaptive observer designed for impedance estimation. The observer has two main functions: a state observer for the grid current and parameter adaptation laws for L_g and R_g . The input signals for the observer are the measured grid current i_g^s and the voltage u_g^s at the PCC.

Fig. 3. Adaptive grid observer.

A. Structure

The observer is formulated based on the grid model (1) and (2) in a rotating coordinate system. These coordinates are marked without a superscript and tied to the excitation signal v_e^s , i.e., the signal in the rotating coordinates is a DC quantity

$$v_e = e^{-j\vartheta_e} v_e^s = v_e + j0 \quad (3)$$

where $\vartheta_e = \arg\{v_e^s\}$ is the rotation angle of the excitation signal. The coordinate system could also be grid-voltage oriented or stationary but excitation-signal coordinates are selected to obtain simple parameter adaptation laws for L_g and R_g as explained later. The observer is

$$\frac{d\hat{\mathbf{x}}}{dt} = \underbrace{\begin{bmatrix} -\frac{\hat{R}_g}{\hat{L}_g} - j\omega_e & -\frac{1}{\hat{L}_g} \\ 0 & j(\omega_g - \omega_e) \end{bmatrix}}_{\hat{\mathbf{A}}} \hat{\mathbf{x}} + \underbrace{\begin{bmatrix} \frac{1}{\hat{L}_g} \\ 0 \end{bmatrix}}_{\hat{\mathbf{B}}} u_g + \underbrace{\mathbf{K}_o}_{\hat{\mathbf{C}}} (i_g - \hat{i}_g) \quad (4)$$

$$\hat{i}_g = \begin{bmatrix} 1 & 0 \end{bmatrix} \hat{\mathbf{x}}$$

where $\hat{\mathbf{x}} = [\hat{i}_g, \hat{e}_g]^T$ is the estimated state vector and $\mathbf{K}_o = [k_1, k_2]^T$ is the observer gain. The selection of the gain is discussed in Section IV-B.

Parameter adaptation laws for the inductance and resistance estimation are formulated by integrating current estimation error $\tilde{i}_g = i_g - \hat{i}_g$ as

$$\hat{L}_g = k_L \int \text{Im} \{ \gamma \cdot \text{LPF} \{ \tilde{i}_g \} \} dt \quad (5)$$

and

$$\hat{R}_g = k_R \int \text{Re} \{ \gamma \cdot \text{LPF} \{ \tilde{i}_g \} \} dt \quad (6)$$

where γ is a unity-gain rotation operator, and k_L and k_R are the adaptation gains. The selection of the gains and rotation operator is discussed in Section IV-D. In the adaptation laws, the current estimation error is filtered with a low-pass filter (LPF). The filter is designed to pass through the injection-frequency current component (DC-valued signal in excitation-signal coordinates), but to attenuate the fundamental-frequency component (a rotating vector in excitation-signal coordinates). A first-order LPF is used

$$G_f(s) = \frac{\alpha_f}{s + \alpha_f} \quad (7)$$

where α_f is the bandwidth. In the design example presented in this paper, the LPF filter bandwidth is 10 Hz and the injection frequency is considered to be above the grid frequency of 50 Hz such as 110 Hz.

Fig. 4. Linearized inductance and resistance estimation loops. Blue blocks represent the inductance and resistance adaptation laws.

B. Linearization and the Observer Gain

The estimation error of the state vector is $\tilde{\mathbf{x}} = \mathbf{x} - \hat{\mathbf{x}}$, where $\mathbf{x} = [i_g, e_g]^T$ is the actual state vector. The dynamic model of the estimation error is

$$\frac{d\tilde{\mathbf{x}}}{dt} = (\hat{\mathbf{A}} - \mathbf{K}_o \mathbf{C}) \tilde{\mathbf{x}} + (\mathbf{A} - \hat{\mathbf{A}}) \mathbf{x} + (\mathbf{B} - \hat{\mathbf{B}}) u_g \quad (8)$$

where $\mathbf{A}$ and $\mathbf{B}$ are the actual system matrices. The dynamic model (8) is nonlinear with respect to $\hat{L}_g$. For the analysis and gain selection of the observer, it can be linearized as presented in the Appendix. As a result, a linearized model for the estimation error is obtained

$$\frac{d(\delta\tilde{\mathbf{x}})}{dt} = (\hat{\mathbf{A}}_0 - \mathbf{K}_o \mathbf{C}) \delta\tilde{\mathbf{x}} + \mathbf{B}_L \delta\tilde{L}_g + \mathbf{B}_R \delta\tilde{R}_g \quad (9)$$

where $\delta\tilde{\mathbf{x}}$ is the small-signal deviation of the estimation error and $\hat{\mathbf{A}}_0$, $\mathbf{B}_L$, and $\mathbf{B}_R$ are the system matrices at an operating point. Inputs $\delta\tilde{L}_g$ and $\delta\tilde{R}_g$ are the small-signal deviations of the estimation errors $\tilde{L}_g = L_g - \hat{L}_g$ and $\tilde{R}_g = R_g - \hat{R}_g$, respectively.

The characteristic polynomial of (9) is $\det(s\mathbf{I} - \hat{\mathbf{A}}_0 + \mathbf{K}_o \mathbf{C})$. With the observer gain $\mathbf{K}_o$, the polynomial can be parameterized as

$$\det(s\mathbf{I} - \hat{\mathbf{A}}_0 + \mathbf{K}_o \mathbf{C}) = s^2 + 2\zeta_o \omega_o s + \omega_o^2 \quad (10)$$

where ζ_o and ω_o are the design parameters. Then, simple closed-form expressions for the elements of $\mathbf{K}_o$ are obtained from (10)

$$\begin{aligned} k_1 &= 2\zeta_o \omega_o - \hat{R}_g / \hat{L}_g + j(\omega_g - 2\omega_e) \\ k_2 &= (\omega_g \omega_e - \omega_e^2 - \omega_o^2) \hat{L}_g + j(\omega_e - \omega_g)(\hat{R}_g + \hat{L}_g k_1) \end{aligned} \quad (11)$$

A suitable value for the damping ratio is around $\zeta_o = 0.7 \dots 1$ to avoid ringing in the estimation error dynamics. The natural frequency ω_o can be set at least a decade higher than the bandwidths of the parameter adaptation subsystem in order to avoid interaction between different parts of the observer.

For the design of parameter adaptation laws, the linearized model (9) is examined at the injection frequency ω_e . The injection frequency corresponds to the zero frequency in the selected rotating coordinates. Thus, the injection-frequency gains from the inductance and resistance estimation errors to the grid current estimation error $\tilde{i}_g = i_g - \hat{i}_g = \mathbf{C}\tilde{\mathbf{x}}$ can be calculated inserting $d(\delta\tilde{\mathbf{x}})/dt = 0$ in (9). The current estimation error at the injection frequency is

$$\delta\tilde{i}_g = \mathbf{G}_L \delta\tilde{L}_g + \mathbf{G}_R \delta\tilde{R}_g \quad (12)$$

Fig. 5. Vector diagram: operation point current $i_{g0,e}$; current estimation error δi_g and its orthogonal components; and rotated current estimation error $\gamma \delta i_g$ and its components (dashed lines). The rotation operator is $\gamma = \exp(-j\theta)$.

where the injection-frequency gains are

$$\begin{aligned} G_L &= \frac{\omega_e(\omega_e - \omega_g) i_{g0,e}}{L_{g0} \omega_o^2} \\ G_R &= -j \frac{(\omega_e - \omega_g) i_{g0,e}}{L_{g0} \omega_o^2} \end{aligned} \quad (13)$$

The gains are functions of the grid inductance L_{g0} and injection-frequency grid current component $i_{g0,e}$ at the operating point. Moreover, the gains translate the inductance and resistance estimation errors to the current estimation error that is the input for the parameter adaptation laws as shown in Fig. 4. Furthermore, the expressions (12) and (13) point out that the components caused by the inductance and resistance estimation errors on the current estimation error are orthogonal, i.e. $G_L = j\omega_e G_R$, which is illustrated in Fig. 5.

C. Injection-Frequency Operating-Point Current

The injection-frequency gains (13) are functions of the operating point current component $i_{g0,e}$ that has to be considered in the tuning of the adaptation laws. The effect of $i_{g0,e}$ on the linearized system is further analyzed. It is assumed that the grid voltage e_g does not contain any injection-frequency component. Then, $i_{g0,e}$ originates from the excitation signal v_e (3) and the control system actions trying to reject the signal as a disturbance. The effect of the control system on $i_{g0,e}$ can be reduced, e.g., by adding band-reject filters at the injection frequency in the control loops [4].

Control system effects are further analyzed considering a current-controlled converter. A block diagram of a two-degree-of-freedom current control loop is shown in Fig. 6. When current control can be described with this structure, the transfer function from the excitation signal v_e^s to the grid current becomes

$$\frac{i_g^s(s)}{v_e^s(s)} = \frac{Y_c(s)}{1 + F_c(s)Y_c(s)} \quad (14)$$

where Y_c is the open-loop transfer function from the voltage reference $u_{c,ref}^s$ to the grid current i_g^s and F_c is the feedback current controller. Moreover, the injection-frequency component of i_g at the operating point is

$$i_{g0,e} = \frac{Y_c(j\omega_e)}{1 + F_c(j\omega_e)Y_c(j\omega_e)} v_e \quad (15)$$

Fig. 6. Current control system. The current reference $i_{g,ref}^s$ goes through a reference pre-filter F_r , and the feedback current controller is F_c . The open-loop system is modeled via admittances Y_c and Y_g .

Typically, the controller F_c is known, but the open-loop admittance is unknown due to unknown grid impedance. However, considering inductive components only, an approximation for the open-loop admittance is

$$Y_c(j\omega_e) \approx \frac{1}{j\omega_e(L_g + L_t)} e^{-j\omega_e T_d} \quad (16)$$

where L_t is the total series inductance of the LCL filter chokes and $e^{-j\omega_e T_d}$ is the phase-shift due to delay $u_c^s(t) = u_{c,ref}^s(t - T_d)$ in the voltage reference chain. The delay of $T_d = 1.5 \cdot T_s$ is considered in this work.

Combining (13), (15), and (16), the injection-frequency gains become

$$\begin{aligned} G_L &= \frac{\omega_e(\omega_e - \omega_g)}{L_{g0} \omega_o^2} \frac{Y_c(j\omega_e)}{1 + F_c(j\omega_e)Y_c(j\omega_e)} v_e \\ &\approx \frac{\omega_e(\omega_e - \omega_g)}{L_{g0} \omega_o^2} \frac{v_e}{j\omega_e(L_{g0} + L_t)e^{j\omega_e T_d} + F_c(j\omega_e)} \end{aligned} \quad (17)$$

$$\begin{aligned} G_R &= \frac{-j(\omega_e - \omega_g)}{L_{g0} \omega_o^2} \frac{Y_c(j\omega_e)}{1 + F_c(j\omega_e)Y_c(j\omega_e)} v_e \\ &\approx \frac{j(\omega_g - \omega_e)}{L_{g0} \omega_o^2} \frac{v_e}{j\omega_e(L_{g0} + L_t)e^{j\omega_e T_d} + F_c(j\omega_e)} \end{aligned} \quad (18)$$

The gains in this form give a rough description of the steady-state effect of the inductance and resistance estimation errors and the excitation voltage v_e on the current estimation error, i.e., $\tilde{i}_g = G_L \tilde{L}_g + G_R \tilde{R}_g$. The selected approximations are more reasonable at the frequencies where the inductive behaviour of the system is dominant. Nevertheless, they afford an approach to select gains k_L , k_R , and the rotation operator γ of the parameter adaptation laws (5) and (6) explained in the following section.

D. Adaptation Gains

As Fig. 4 shows, the parameter adaptation laws together with the linearized model can be seen as a multiple-input multiple-output system. To reduce coupling between the inductance and resistance estimation, the estimation-error $\tilde{i}_g$ is rotated in the adaptation laws (5) and (6) with the rotation operator

$$\gamma = \exp(-j\theta) \quad (19)$$

Since the resistance is adapted from the real part of the current estimation error (6), the rotation angle $\theta = \arg\{G_R\}$ is selected to align the estimation-error component $\gamma G_R \tilde{R}_g$ originating from the resistance on the real axis as illustrated

in Fig. 5. The inductance error component $\gamma \mathbf{G}_L \tilde{L}_g$ is aligned on the imaginary axis. The rotation angle is calculated as

$$\theta = \arg \left\{ \frac{j(\omega_g - \omega_e)}{j\omega_e \omega_o^2 L_0 (L_0 + L_t) e^{j\omega_e T_d} + \omega_o^2 L_0 \mathbf{F}_c(j\omega_e)} \right\} \quad (20)$$

where L_0 is the initial guess of the unknown grid inductance.

After the rotation, the inductance and resistance adaptation loops are considered as two decoupled single-input single-output systems. Then, the closed-loop transfer functions for the inductance and resistance estimation are

$$\frac{\hat{L}_g(s)}{L_g(s)} = \frac{|\mathbf{G}_L| G_f(s) k_L}{s + |\mathbf{G}_L| G_f(s) k_L} \quad (21)$$

$$\frac{\hat{R}_g(s)}{R_g(s)} = \frac{|\mathbf{G}_R| G_f(s) k_R}{s + |\mathbf{G}_R| G_f(s) k_R} \quad (22)$$

A simple approach to select gains k_L and k_R is to approximate these transfer functions as first order systems. When $G_f(s) = 1$ is assumed, the transfer functions can be expressed as

$$\frac{\hat{L}_g(s)}{L_g(s)} = \frac{\alpha_L}{s + \alpha_L}, \quad \frac{\hat{R}_g(s)}{R_g(s)} = \frac{\alpha_R}{s + \alpha_R} \quad (23)$$

with the bandwidths of α_L and α_R , respectively. Comparing (21) and (22) with (23) results in the gains

$$k_L = \frac{\alpha_L}{|\mathbf{G}_L|} = \frac{\alpha_L \omega_o^2 L_0 |j\omega_e (L_0 + L_t) e^{j\omega_e T_d} + \mathbf{F}_c(j\omega_e)|}{\omega_e |v_e(\omega_e - \omega_g)|} \quad (24)$$

$$k_R = \frac{\alpha_R}{|\mathbf{G}_R|} = \frac{\alpha_R \omega_o^2 L_0 |j\omega_e (L_0 + L_t) e^{j\omega_e T_d} + \mathbf{F}_c(j\omega_e)|}{|v_e(\omega_g - \omega_e)|} \quad (25)$$

The assumption $G_f(s) = 1$ is reasonable when α_L and α_R are, e.g., one decade smaller than the bandwidth α_f of G_f . A practical upper limit for the LPF bandwidth is $\alpha_f < |\omega_e - \omega_g|$, since the fundamental-frequency component has to be attenuated in the adaptation loops.

The gains k_L and k_R and the angle θ are functions of excitation signal parameters v_e and ω_e and system parameters ω_g , L_t , T_d , $\mathbf{F}_c(j\omega_e)$, L_0 . From these parameters, the initial guess L_0 is the only uncertain parameter. If $L_0 < L_g$, the estimators are slower (lower effective bandwidth) than in the case of $L_0 = L_g$, and vice versa if $L_0 > L_g$. Nevertheless, an inaccurate initial guess $L_0 \neq L_g$ does not cause steady-state errors to the parameter estimates. Generally, it is better to underestimate L_0 than overestimate it in order to avoid interactions between the different parts of the grid observer¹. If the control system is not effectively rejecting the excitation signal v_e , then the gain calculation can be simplified by inserting $\mathbf{F}_c(j\omega_e) = 0$ in (20), (24), and (25). This is the case, if a band-reject filter at ω_e is used in the current feedback as in [4].

¹The grid observer can be analyzed as a whole, if the linearized model (9) is combined with the estimators (5), (6), and (7) to create a full closed-loop model.

E. Discretization for Implementation

The observer matrices are discretized for the implementation using zero-order-hold equivalent method. The discretized observer is

$$\hat{\mathbf{x}}(k+1) = \hat{\mathbf{A}}_d \hat{\mathbf{x}}(k) + \hat{\mathbf{B}}_d \mathbf{u}_g(k) + \mathbf{K}_{o,d} [\hat{\mathbf{i}}_g(k) - \hat{\mathbf{i}}_g(k)] \quad (26)$$

where the matrices are

$$\begin{aligned} \hat{\mathbf{A}}_d &= e^{\hat{\mathbf{A}} T_s} \\ &= e^{-j\omega_e T_s} \begin{bmatrix} e^{-\hat{R}_g T_s / \hat{L}_g} & \frac{1}{\hat{R}_g + j\omega_g \hat{L}_g} (e^{-\hat{R}_g T_s / \hat{L}_g} - e^{j\omega_g T_s}) \\ 0 & e^{j\omega_g T_s} \end{bmatrix} \\ \hat{\mathbf{B}}_d &= \left(\int_0^{T_s} e^{\hat{\mathbf{A}} \tau} d\tau \right) \hat{\mathbf{B}} \\ &= \begin{bmatrix} (1 - e^{-\hat{R}_g T_s / \hat{L}_g} e^{-j\omega_e T_s}) / (\hat{R}_g + j\hat{L}_g \omega_e) \\ 0 \end{bmatrix} \end{aligned} \quad (27)$$

Furthermore, the observer gain $\mathbf{K}_{o,d}$ in the discrete-time implementation can be calculated by selecting desired poles of the characteristic polynomial $\det(z\mathbf{I} - \hat{\mathbf{A}}_d + \mathbf{K}_{o,d}\mathbf{C})$ as in the continuous-time case (10). Alternatively, a simple mapping $\mathbf{K}_{o,d} = \mathbf{K}_o T_s$ can be used. For simplicity, the mapping is selected in this paper. The LPF (7) and integrators in (5) and (6) are discretized using Euler's method $s \rightarrow (z-1)/T_s$. The outputs of the integrators ($\hat{L}_g$ and $\hat{R}_g$) are limited to positive values in order to avoid division by zero in (27).

V. COMPARISON

The both impedance estimators, the SDFT-based method and the adaptive grid observer, can be run online in parallel with an existing control system and provide parameter estimates in real time. However, some of their properties are different or even contrary. The adaptive grid observer requires a selection for the gains and an initial guess of the inductance L_0 , but does not require a large buffer for the current and voltage samples as the SDFT. Moreover, the injection frequency is restricted to multiples of frequency resolution f_{res} with SDFT, but can be more freely selected with the observer. Simulations and experiments are further used to compare the estimators.

A. Simulations

Dynamic behaviour of the impedance estimation methods is first compared by means of simulations. A 400-V 12.5-kVA three-phase grid converter system in weak grid conditions is considered, and the designed current control bandwidth is 200 Hz [22]. The observer tuning parameters are $\omega_o = 2\pi \cdot 1000$ rad/s, $\zeta_o = 1$, $\alpha_f = 2\pi \cdot 10$ rad/s, $\alpha_L = \alpha_R = 2\pi \cdot 2$ rad/s, and $L_0 = 0.4$ p.u. The sampling frequency is 10 kHz, and the buffer size for the SDFT is $N = 1000$ complex-valued samples per measured space vector. This leads to frequency resolution of 10 Hz. The selected base values (1 p.u.) are: voltage $\sqrt{2/3} \cdot 400$ V, current $\sqrt{2} \cdot 18$ A, power 12.5 kVA, and angular frequency $2\pi \cdot 50$ rad/s.

Fig 7(a) shows the results when the methods are started at $t = 0.5$ s. During the test, the converter injects the power

Fig. 7. Simulation results: (a) estimated inductance and resistance (b) estimated inductance and resistance when the active power p_g at the PCC is fluctuating.

Fig. 9. Experimental results: (a) estimated inductance and resistance (b) grid phase currents $i_{g,a}$, $i_{g,b}$, and $i_{g,c}$ during the estimation; and (c) spectra of the PCC voltage and current during the estimation.

Fig. 8. Experimental test setup. The converter control software and the estimation methods are running on the dSPACE DS1006 processor board where the sampling of the measured signals is synchronized with the pulse width modulation (PWM) and two samples per carrier are taken.

of $p_g = 0.5$ p.u. to the 50-Hz grid and an excitation signal v^s at an inter-harmonic frequency of $\omega_e = 2\pi \cdot 110$ rad/s is added to the converter voltage reference with the magnitude of $v_e = 0.01$ p.u. At $t = 2$ s, the actual impedance is reduced to half of its original value. Even though the both methods

under comparison find the actual impedance, the SDFT-based estimates converge slightly faster. Fig. 7(b) compares the methods under changing operation point. The PCC power p_g has a sinusoidal component (0.25 p.u.) at 5 Hz from $t = 0.1$. As the figure shows, the SDFT-based method cannot estimate the impedance correctly when the operation point is not constant.

B. Experiments

The estimators were experimentally tested. A diagram of the setup used in the experiments is shown in Fig 8. The

Fig. 10. Experimentally estimated inductance and resistance.

inductive-resistive grid impedance is emulated with external chokes connected between the PCC and the 50-Hz electric power system. The control system and tuning parameters of the estimator equal the ones used in the simulations. A first-order LPF having a cut-off frequency of 10 Hz is used with the SDFT to reduce the measurement noise. The converter injects the active power of $p_g = 0.5$ p.u. to the grid during the estimation.

Fig. 9(a) shows the estimated inductance and resistance. The estimation starts at $t = 0.5$ s. The estimated inductance behaves smoothly and it converges to 22.2 mH (0.55 p.u.). The estimated resistance 1.4Ω (0.11 p.u.) is more noisy than the inductance. After 1.9 s, impedance was reduced by by-passing 10 mH choke in all of the three phases. Both methods under comparison detect the inductance change well. Generally, the SDFT-based estimates converge more rapidly. Fig. 9(a) also shows reference values (black dashed lines) for the inductance and the resistance that were calculated using the DFT as in [6]. The reference values were calculated as averages of 10 individual DFT results. The reference inductance is obtained from the impedance as $\tilde{L}_g = \text{Im}\{\tilde{Z}_g\}/\omega_e$. As the figure shows, the SDFT-based and observer-based estimates are very close or overlap the conventional DFT-based estimates.

Fig. 9(b) shows grid phase currents and Fig. 9(c) shows the spectra of the PCC voltage and current during the estimation process. It can be seen that the disturbance components at the PCC current and voltage are small (0.01 p.u.) at the injection frequency. The inter-harmonic injection frequency of 110 Hz was selected to provide an estimate close to the fundamental frequency but to avoid bias due to the typical grid-voltage harmonics. Different injection frequencies, such as 90, 110, 130, 170, 190, and 210 Hz, were also experimentally tested to verify that the presented estimation methods are not limited to the selected frequency.

The measurement was repeated with different initial guesses of the grid impedance. Fig. 10 shows the estimation results when the initial guess of the grid inductance is lower $L_0 = 0.3$

p.u. compared to the case in Fig. 9(a) where $L_0 = 0.4$ p.u. The figures show that the observer-based estimates converge slower when the initial guess is smaller. The slower response is explained with the lower estimator gains (24) and (25) that are functions of L_0 . If the initial guess is higher than 0.4 p.u., the behaviour is opposite since the gains become higher. Furthermore, with a too high initial guess $L_0 > 0.75$ p.u. (too high gains) the stability of the observer is lost after the impedance step when the actual inductance L_g is only 0.3 p.u.

VI. CONCLUSION

This paper presents two methods for real-time grid impedance estimation: an SDFT-based estimator and an adaptive grid observer. These methods are compared with simulations and experiments. According to results, steady-state accuracy of the both methods is the same. In the case of a impedance change, the convergence rate of the SDFT-based estimator is faster than that of the observer. Moreover, the SDFT does not need tuning as the observer. The gains of the observer are derived using a linerized model but depends on the initial guess of the grid inductance. Nevertheless, the observer can be easily implemented in a converter control system since it does not require long buffers as the SDFT. Since an excitation signal with a low magnitude can be used in the estimation, the disturbance caused to the power system is minor. Locally, the estimated grid impedance can be used, e.g., to detect islanding conditions. In remote monitoring of converters, it may provide added value when analyzing a converter system status or possible grid-converter interactions.

APPENDIX LINEARIZATION

For the linearization [23], the nonlinear system (8) is rewritten

$$\frac{d\tilde{\mathbf{x}}}{dt} = \mathbf{f}(\tilde{\mathbf{x}}, \mathbf{x}, \mathbf{u}_g, L_g, R_g, \tilde{L}_g, \tilde{R}_g) \quad (28)$$

where $\tilde{R}_g = R_g - \hat{R}_g$, $\tilde{L}_g = L_g - \hat{L}_g$ and constant ω_g and ω_e are assumed. The system has an operating point (equilibrium state) $\{\tilde{\mathbf{x}}_0, \mathbf{x}_0, \mathbf{u}_{g0}, L_{g0}, R_{g0}, \tilde{L}_{g0}, \tilde{R}_{g0}\}$ where

$$\mathbf{f}(\tilde{\mathbf{x}}_0, \mathbf{x}_0, \mathbf{u}_{g0}, L_{g0}, R_{g0}, \tilde{L}_{g0}, \tilde{R}_{g0}) = 0 \quad (29)$$

The condition (29) is satisfied when $\tilde{\mathbf{x}}_0 = \tilde{R}_{g0} = \tilde{L}_{g0} = 0$. Generally, the operating point can be a time-varying trajectory if multifrequency signals are considered. To simplify design of the adaptive grid observer, only injection-frequency signals are considered and the fundamental-frequency components are neglected in the linearized model. This assumption is reasonable due to the LPF in the parameter adaptation laws (5) and (6). In the vicinity of the operating point, the dynamic model for the estimation error is

$$\begin{aligned} \frac{d(\delta\tilde{\mathbf{x}})}{dt} = & \left[\frac{\partial \mathbf{f}}{\partial \tilde{\mathbf{x}}} \right]_0 \delta\tilde{\mathbf{x}} + \left[\frac{\partial \mathbf{f}}{\partial \mathbf{x}} \right]_0 \delta\mathbf{x} + \left[\frac{\partial \mathbf{f}}{\partial \mathbf{u}_g} \right]_0 \delta\mathbf{u}_g + \left[\frac{\partial \mathbf{f}}{\partial L_g} \right]_0 \delta L_g \\ & + \left[\frac{\partial \mathbf{f}}{\partial R_g} \right]_0 \delta R_g + \left[\frac{\partial \mathbf{f}}{\partial \tilde{L}_g} \right]_0 \delta \tilde{L}_g + \left[\frac{\partial \mathbf{f}}{\partial \tilde{R}_g} \right]_0 \delta \tilde{R}_g \end{aligned} \quad (30)$$

where the small-signal deviation is marked with δ , e.g., $\delta\tilde{\mathbf{x}} = \tilde{\mathbf{x}} - \tilde{\mathbf{x}}_0$. The partial derivatives, evaluated at the operating point, are

$$\begin{aligned} \left[\frac{\partial \mathbf{f}}{\partial \tilde{\mathbf{x}}} \right]_0 &= \hat{\mathbf{A}}_0 - \mathbf{K}_o \mathbf{C} \\ \left[\frac{\partial \mathbf{f}}{\partial \tilde{\mathbf{u}}} \right]_0 &= \left[\frac{\partial \mathbf{f}}{\partial \tilde{\mathbf{u}}_g} \right]_0 = 0 \\ \left[\frac{\partial \mathbf{f}}{\partial \tilde{L}_g} \right]_0 &= \left[\frac{\partial \mathbf{f}}{\partial \tilde{L}_g} \right]_0 = \begin{bmatrix} \frac{R_{g0}}{L_{g0}^2} & \frac{1}{L_{g0}^2} \\ 0 & 0 \end{bmatrix} \mathbf{x}_0 + \begin{bmatrix} -\frac{1}{L_{g0}^2} \\ 0 \end{bmatrix} \mathbf{u}_{g0} \\ \left[\frac{\partial \mathbf{f}}{\partial \tilde{R}_g} \right]_0 &= \left[\frac{\partial \mathbf{f}}{\partial \tilde{R}_g} \right]_0 = \begin{bmatrix} -\frac{1}{L_{g0}} & 0 \\ 0 & 0 \end{bmatrix} \mathbf{x}_0 \end{aligned} \quad (31)$$

Since $\mathbf{x}_0 = [\mathbf{i}_{g0}, \mathbf{e}_{g0}]^T$, the partial derivatives for the inductance and resistance terms are further reduced to

$$\begin{aligned} \left[\frac{\partial \mathbf{f}}{\partial \tilde{L}_g} \right]_0 &= \begin{bmatrix} (R_{g0}/L_{g0}^2)\mathbf{i}_{g0} + (\mathbf{e}_{g0} - \mathbf{u}_{g0})/L_{g0}^2 \\ 0 \end{bmatrix} = \mathbf{B}_L \\ \left[\frac{\partial \mathbf{f}}{\partial \tilde{R}_g} \right]_0 &= \begin{bmatrix} -\mathbf{i}_{g0}/L_{g0} \\ 0 \end{bmatrix} = \mathbf{B}_R \end{aligned} \quad (32)$$

Since the injection-frequency current component $\mathbf{i}_{g0,e}$ is only considered, $\mathbf{i}_{g0} = \mathbf{i}_{g0,e}$ and the PCC voltage can be expressed as

$$\mathbf{u}_{g0} = \mathbf{e}_{g0} + j\omega_e L_{g0} \mathbf{i}_{g0,e} + R_{g0} \mathbf{i}_{g0,e} \quad (33)$$

Inserting (33) into (32), the partial derivatives $\mathbf{B}_R$ and $\mathbf{B}_L$ become

$$\mathbf{B}_R = \begin{bmatrix} -\frac{\mathbf{i}_{g0,e}}{L_{g0}} \\ 0 \end{bmatrix}, \quad \mathbf{B}_L = \begin{bmatrix} -\frac{j\omega_e \mathbf{i}_{g0,e}}{L_{g0}} \\ 0 \end{bmatrix} \quad (34)$$

Finally, when it is assumed that L_g and R_g are constant or slowly varying, i.e., $\delta L_g = \delta R_g = 0$, the linearized model becomes

$$\frac{d(\delta\tilde{\mathbf{x}})}{dt} = (\hat{\mathbf{A}}_0 - \mathbf{K}_o \mathbf{C})\delta\tilde{\mathbf{x}} + \mathbf{B}_L \delta\tilde{L}_g + \mathbf{B}_R \delta\tilde{R}_g \quad (35)$$

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