Rapid Multivariable Identification of Grid Impedance in DQ Domain Considering Impedance Coupling

Matias Berg, Student Member, IEEE, Henrik Alenius, Student Member, IEEE, and Tomi Roinila, Member, IEEE

Abstract—Identifying grid-impedance at the point of common coupling is essential for the adaptive control and the online stability analysis of grid-connected converters. A balanced three-phase system is commonly modeled by d and q components in the synchronously reference frame. In identification of the synchronous reference-frame impedance components, errors may occur due to the coupling of the system impedances; for example, a measurement injection that is intended to perturb only the d-channel current may also perturb the q-channel current thus distorting the impedance measurements. Traditionally sequentially performed measurements, where different injections are performed one after another at same frequencies, have been required to tackle the impedance coupling. However, the sequential measurements are prone to changes in the operating conditions between the measurements. The present paper proposes a method to simultaneously obtain all the grid-impedance components within a single measurement cycle with no coupling effect. In the method, two orthogonal binary injections are simultaneously injected into the d and q current references of the inverter controller. Then, a frequency-domain interpolation technique is applied to adjust the measured current and voltage responses. As a result, the impedance coupling is avoided in the measured grid impedance. The proposed technique is validated by experimental measurements.

Index Terms—frequency response, identification, three-phase VSI, grid-impedance, MIMO

I. INTRODUCTION

Replacing conventional energy sources with renewable sources has been recognized as an important tool for retarding climate change [1]. The remarkable feature of renewable energy sources is that they commonly need to be connected through power-electronic interfaces to the grid. Therefore, grid-connected three-phase converters are increasingly becoming a vital factor in the operation of distribution grids. However, the grid-connected converters are prone to adverse impedance-based interactions between the grid impedance and other converters, which disturbs the power quality and may even introduce stability issues [2]–[6]. In order to prevent the adverse interactions, information on the grid impedance is often required for improved control design or adaptive control.

Accurate identification of the grid impedance or the converter impedance generally requires an external injection to the system. The injection can be made by a grid-connected converter [7]–[9] or by a separate measurement device [10], [11]. The grid-impedance identification may rely on the assumption of the grid-impedance shape. The grid-impedance is occasionally assumed to be a series resistor-inductor (RL) circuit [12], [13]. In [13], the identification method was based on a step response, and while in [12] the method was based on a current injection at a single frequency. However, the grid-impedance may vary significantly from a series RL circuit [14]–[16]. Therefore, identification at a wide frequency range may be more advantageous, and it is required for an accurate multi-variable stability analysis, such as with the generalized Nyquist criterion [11], [17].

Recent studies have presented a number of wideband techniques to measure the grid impedance [7]–[9], [18], [19]. In such techniques the impedance is measured by injecting a broadband perturbation such as a pseudo-random binary sequence (PRBS), for example, into the inverter controller reference. The resulting grid voltages and currents are measured, and Fourier methods are applied to obtain the grid impedance. The PRBS exhibits multiple favorable characteristics for use in online impedance identification, such as periodic and deterministic nature, and the lowest possible peak factor [20].

In the synchronous-reference frame, a three-phase converter system is a multiple-input-multiple-output system and consists of direct (d) and quadrature (q) channels [10], [11] that must be identified for a proper analysis. Both channels can be identified simultaneously in a brief time by using orthogonal binary injections [7]–[9]. In practice, however, the d and q channels are cross-coupled. Due to the cross-coupling and the interaction between the system impedances, an impedance coupling occurs. The impedance coupling may cause an intended d-channel current injection to appear also in the current q channel. Therefore, for example the response in the voltage d component would not only be caused by the intended d-channel current but also by q-channel current. A thorough analysis of the impedance coupling was performed in [21].

It was reported in [21] that the orthogonal injections applied into dq-domain systems in [7]–[9], [22] can give erroneous results if the impedance coupling effect is overlooked. However, the authors in [21] made a hasty generalization regarding the issue of the orthogonal injections because a DC-DC converter system that has not been reported to have d and q channels was analyzed in [22]. To avoid the impedance-coupling effect, a time-domain parametric identification method was proposed in [21] where uncorrelated binary sequences were used to perturb the system. However, the presented uncorrelated broadband injections, whose design algorithms were not given, differ

This work was supported in part by Business Finland Project SolarX. M. Berg*, H. Alenius, and T. Roinila are with the Faculty of Information Technology and Communication Sciences, Tampere University, Finland. *e-mail: matias.berg@tuni.fi.
greatly from the analytically derived uncorrelated sequences that have been used in the field of process identification in [23], [24]. The orthogonal-sequence-generation methods from [23], [24] have been applied in identification of power electronic systems in [7]–[9], [22] and are used in the present paper.

In tackling the impedance-coupling effect without parametric model, sequential independent injections have been proven to be an adequate method for obtaining the accurate multivariable impedance [10], [11], [25], [26]. By applying this method, the impedance-coupling effect can be taken into account in the calculation of impedance components. However, applying multiple sequential measurements relies on the assumption that the impedance does not change between the injections [10].

The present paper proposes a novel measurement procedure that utilizes orthogonal binary broadband injections in the elimination of impedance interactions. An orthogonal sequence is injected to the current reference d component, and a second orthogonal sequence is injected to the current reference q component. The result of the q-channel measurement is interpolated to the frequencies of the d-channel measurement. The result is used as an independent measurement. However, obtaining an accurate interpolation result is not straightforward due to the random phase behavior of the binary sequences; the results must be divided by the spectrum of the injected sequence before the interpolation. The original orthogonal binary sequence and the interpolation result are independent of each other, and can be used to accurately identify the grid impedance. As a result, the accurate multivariable impedance can be identified non-parametrically with simultaneous broadband excitations.

The remainder of this paper is structured as follows. Section II presents the system and the broadband injections. Section III describes the impedance coupling in the measurement and revises an existing method to obtain accurate results in the presence of the impedance coupling. The novel identification procedure is derived and simulated in Section IV. Experimental verification of the proposed technique is presented in Section V. Conclusions are drawn in the final section.

II. CONVERTER SYSTEM AND IDENTIFICATION SEQUENCES

This section begins by introducing the parameters and the controller structure of the grid-connected converter. Then, characteristics and the generation of orthogonal broadband binary injections are revised.

A. Grid-connected inverter

A block diagram of the system under study is shown in Fig. 1. The traditional synchronous-reference-frame phase-locked loop (PLL) is used to synchronize the rotating synchronous reference frame to the grid voltages at the point of common coupling [27]. The converter-side inductor currents are controlled with PI controllers. The perturbation signal, which is required for the grid impedance identification, is injected to the inductor current reference. The system and control parameters are given in Table I.

The diagram in Fig. 1 shows that the grid-impedance is identified from the output voltage (v) and the output current (i) measurements. Therefore, the filter capacitor ($C_i$) is not included in the impedance that is identified. A small error in the traditional measurement arises from the current angle shift at the filter capacitor.

B. Orthogonal broadband binary injections

Pseudorandom binary sequences are broadband signals that can be used for fast frequency response measurements because multiple frequencies are injected simultaneously. One of the most common of such signals is the maximum-length binary sequence (MLBS), which has two levels and can be generated with low computational effort by using shift registers [23]. The sequence is periodic, with a length of $N = 2^n - 1$, where $n$ is the length of the shift register. [23]. The lowest frequency with energy is $f_{gen}/N$, where $f_{gen}$ is the generation frequency. When considering up to the $f_{gen}$, the frequencies at which the MLBS has energy can be given as:

$$f_{k_{MLBS}} = kf_{gen}/N, \quad k = 1, 2, 3...N$$ (1)

where $k$ denotes the sequence number of the spectral line.

An inverse repeat sequence (IRS) can be generated from the MLBS by applying the Hadamard modulation discussed in [23]. The IRS generated from a $N$-bit-long MLBS has a length of $2N$, and the IRS has energy at frequencies:

$$f_{k_{IRS}} = (2k - 1)f_{gen}/2N, \quad k = 1, 2, 3...N$$ (2)

The excited frequencies of the IRS fall exactly between the frequencies of the MLBS; therefore, the MLBS and the IRS are orthogonal. They can be used for simultaneous identification of multiple transfer functions from a multiple-input and multiple-output system [7], [28]. The advantage is that both measurements are done simultaneously under the same conditions [29].

Fig. 2a shows a sample to the MLBS and IRS time-domain waveforms and 2b shows the amplitude and the phase of the MLBS and IRS whose magnitude is scaled by 1.2 for illustrative purposes. The present paper uses $n = 9$ and, therefore, the lengths of the MLBS and the IRS are 511 bits and 1022 bits, respectively. The generation frequency $f_{gen}$ is 4 kHz, and the amplitude of both sequences, $A_{MLBS}$ and $A_{IRS}$ are
Figure 2: (a) Part of the MLBS and the IRS in time domain and (b) frequency-domain amplitude and phase of the used MLBS and IRS. The IRS is set to have larger amplitude for illustration.

$A_{IRS}$ is 0.5. The energy of the MLBS drops to zero at $f_{gen}$, and the practical measurement bandwidth is up to 0.45$f_{gen}$ [9]. Fig. 3, where $\bar{z}$ denotes the unit delay, shows how the sequences are generated with shift registers.

### III. Impedance Coupling in Measurements

This section begins by inspecting the equivalent small-signal model of the grid-connected converter to analyze the effect of the impedance coupling on the measurement injections. Then, an existing measurement procedure that is not affected by the impedance coupling is revised.

A. Impedance-based interaction

Fig. 4 shows the equivalent small-signal circuit diagram of the grid-connected converter. The dynamic model of the converter has been derived in [30], but now the system is fed from a voltage source instead of photovoltaic panel. The inductor current reference, the output current, and the output voltage vectors are denoted by $\hat{i}^{ref}_L$, $\hat{i}$, and $\hat{v}$, respectively. The perturbations in the grid-voltage $\hat{v}_g$ are assumed to be zero here because we are analyzing a perturbation caused by the inverter and identifying the grid-impedance. The transfer matrices $G_{co}$, $Y$, and $Z$ that denote the unterminated control-to-output dynamics, the inverter output admittance, and the grid impedance describe the relations between the variables:

$$
\begin{bmatrix}
\hat{i} \\
\hat{i}_d \\
\hat{i}_q \\
\end{bmatrix} =
\begin{bmatrix}
G_{co-d} & G_{co-qd} & \hat{i}^{ref}_L \\
G_{co-dq} & G_{co-q} & \hat{i}_L \\
\end{bmatrix}
\begin{bmatrix}
\hat{v} \\
\hat{v}_s \\
\end{bmatrix}
$$

(3)

Figure 4: Equivalent small-signal circuit diagram of the grid-following converter control-to-output-dynamics-affected injection.

#### Table I: Grid-connected inverter parameters and operating point values.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>$V_{DC}$</td>
<td>413 V</td>
<td>Grid voltage rms</td>
<td>$V_g$</td>
<td>120 V</td>
</tr>
<tr>
<td>Output current d component</td>
<td>$I_d$</td>
<td>10.6 A</td>
<td>Output current q component</td>
<td>$I_q$</td>
<td>0.71 A</td>
</tr>
<tr>
<td>Synchronous frequency</td>
<td>$\omega_s$</td>
<td>2π/50 rad/s</td>
<td>Switching frequency</td>
<td>$f_s$</td>
<td>8 kHz</td>
</tr>
<tr>
<td>Filter capacitor capacitance</td>
<td>$C_f$</td>
<td>10 µF</td>
<td>Filter inductance</td>
<td>$L$</td>
<td>2.5 mH</td>
</tr>
<tr>
<td>$C_f$ ESR and damping resistor</td>
<td>$r_{CF}$</td>
<td>1.81 Ω</td>
<td>$L$ ESR</td>
<td>$r_L$</td>
<td>0.065 Ω</td>
</tr>
<tr>
<td>Filter inductance 2</td>
<td>$L_2$</td>
<td>0.1 mH</td>
<td>$L_2$ ESR</td>
<td>$r_{L2}$</td>
<td>0.022 Ω</td>
</tr>
<tr>
<td>Line inductance</td>
<td>$L_{line}$</td>
<td>8.83 mH</td>
<td>Line inductance ESR</td>
<td>$r_{Lg}$</td>
<td>0.262 Ω</td>
</tr>
<tr>
<td>Transformer inductance</td>
<td>$L_T$</td>
<td>0.507 mH</td>
<td>Transformer inductance ESR</td>
<td>$r_T$</td>
<td>0.417 Ω</td>
</tr>
<tr>
<td>Current controller P gain</td>
<td>$K_{p,c}$</td>
<td>0.0149</td>
<td>Current controller I gain</td>
<td>$K_{i,c}$</td>
<td>23.4</td>
</tr>
<tr>
<td>PLL P gain</td>
<td>$K_{p,pll}$</td>
<td>0.0120</td>
<td>PLL I gain</td>
<td>$K_{i,pll}$</td>
<td>0.0144</td>
</tr>
</tbody>
</table>
Figure 5: Frequency response of the simulation identifications grid-impedance a) d component and b) qd component. $Z_{vd}^{\text{id}}$ and $Z_{vd}^{\text{eq/id}}$ are based on the assumption that only the intended channel is perturbed.

\[
\hat{i} = \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} = \begin{bmatrix} Y_d & Y_{qd} \\ Y_{dq} & Y_q \end{bmatrix} \begin{bmatrix} \hat{v}_d \\ \hat{v}_q \end{bmatrix}
\]

\[
\hat{v} = \begin{bmatrix} \hat{v}_d \\ \hat{v}_q \end{bmatrix} = \begin{bmatrix} Z_d & Z_{qd} \\ Z_{dq} & Z_q \end{bmatrix} \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix}
\]

where subscripts d and q denote the direct and quadrature components, respectively. The transfer matrix from the current reference-to-output voltage is denoted as follows:

\[
\hat{v} = \begin{bmatrix} \hat{v}_d \\ \hat{v}_q \end{bmatrix} = \begin{bmatrix} G_{cv} & -G_{cv} \cdots \hat{i}_{ref} \\ G_{cv} \cdots & \hat{i}_{ref} \end{bmatrix}
\]

The dq-frame impedance matrix of the identified impedance is given as follows

\[
Z = \begin{bmatrix} r_{eq} + sL_{eq} & -L_{eq}\omega_s \\ L_{eq}\omega_s & r_{eq} + sL_{eq} \end{bmatrix}
\]

where $s$ is the Laplace variable, $L_{eq} = L_2 + L_{\text{line}} + L_T$, and $r_{eq} = r_2 + r_{\text{line}} + r_T$. The parameter descriptions and values are given in Table I. In order to demonstrate a strong impedance-based interaction, a relatively high line inductance (8.83 mH) is used.

The grid impedance elements $Z_d$ and $Z_{dq}$ are identified by injecting the MLBS on top of the d current reference. To avoid the effect of harmonic voltages, the number of averaged periods is chosen to be 80 throughout this paper so that an integer amount of fundamental cycles are included in the experimental measurement data [18]. Then, the responses in the voltages and currents are measured, and by computing the ratios of $\hat{v}_d$ to $\hat{i}_d$ and $\hat{v}_q$ to $\hat{i}_d$ the impedance elements are obtained. Fig. 5a shows the experimental measurement result of $Z_{vd}^{\text{id}}$. The result clearly has an error compared to the reference grid impedance below 150 Hz. Fig. 5b shows the identified $Z_{vd}^{\text{eq/id}}$ that is erroneous in the whole frequency range from 150 Hz to 1250 Hz. The errors are caused by the impedance coupling.

In order to study the effect of the impedance coupling, an ideal current injection into an equivalent dq-frame impedance is first considered. The circuit diagram of such a system is shown in Fig. 6a. The elements that cross-couple the d and q channels of the equivalent grid impedance can be seen as current-dependent voltage sources. Both d and q channels are injected using ideal current sources. At the desired frequencies, $\hat{i}_q$ can be kept zero while $\hat{i}_d$ is be perturbed. To illustrate this, the circuit elements that are dependent on $\hat{i}_q$ are grayed out. The ratio of $\hat{v}_d$ to $\hat{i}_d$ results in $Z_d$:

\[
\frac{\hat{v}_d}{\hat{i}_d} = Z_d
\]

In practical systems there are no ideal current sources. The perturbations are injected into the converter current references, and the responses in the currents and voltages are affected by the converter dynamics and the grid. Fig. 6b represents a
case where a perturbation is added to $i_{Ld}^{ref}$ and $i_{Lq}^{ref}$ is not perturbed. The circuit elements that are affected only by $i_{Lq}$ are grayed out. The cross couplings in the converter output admittance are modeled by the voltage-dependent current sources. In contrast to the previous case, the response in $i_q$ is not zero because the interaction among the circuit elements forms a path from $i_{Ld}^{ref}$ to $i_q$. It is highlighted in red that $Z_{qd} i_q$ may have a nonzero value unlike in the case of the ideal current injection. Therefore, the ratio of $v_d$ to $i_d$ may not result in $Z_d$:  

$$
\frac{v_d}{i_d} = \frac{Z_d i_d + Z_{qd} i_q}{i_d} \neq Z_d
$$  (9)

Similarly, the other impedance elements $Z_{qd}$, $Z_{dq}$, and $Z_q$ cannot be identified by computing the ratios of the corresponding frequency domain voltage and current.

The transfer function from the current references to the output currents can be modeled by grid-impedance-load-affected control-to-output dynamics as [31]:

$$G_{co}^L = (I + YZ)^{-1} G_{co}$$  (10)

where superscript "L" denotes that the transfer matrix is load-affected. The dynamics from $i_{Ld}^{ref}$ to $i_q$ are modeled by transfer function $G_{co-dq}^L$ that is shown in more detail in Appendix A. The gain of the transfer function may be low depending on the parameters and the impedance coupling may have no significant effect. For example, if the grid impedance was only resistive

$$Z = \begin{bmatrix} Z_d & Z_{qd} \\ Z_{dq} & Z_q \end{bmatrix} = \begin{bmatrix} r_{eq} & 0 \\ 0 & r_{eq} \end{bmatrix},$$  (11)

the current-dependent voltage sources in Fig. 6b would disappear and no impedance coupling occurs. However, no such assumption can be made if an unknown system is identified.

Fig. 7 shows the Bode plots of the modeled unterminated and the load-affected control-to-output dynamics $d$ component and $dq$ component transfer functions. In the load-affected case, where the line inductance ($L_q$) and the transformer inductance ($L_T$) are connected in series with the grid-side filter inductor $L_2$, the responses from $i_{Ld}^{ref}$ to $i_d$ and $i_q$ have a gain that is approximately the same magnitude. From 105 Hz to 200 Hz and 1250 Hz to 1410 Hz, the difference in gains of $G_{co-dq}^L$ and $G_{co-dq}^L$ is small. Consequently, an injection to $d$-component reference at these frequencies results not only in a response in $i_d$, but also in $i_q$ that cannot be neglected. Therefore, the result from $v_d$/$i_d$ would not give $Z_d$ because $v_d$ is also affected by $Z_{qd} i_q$. These observations give background for the error in the measurement results in Fig. 5. In [21], the impedance coupling was illustrated by a block diagram and the exact erroneous measurement result was derived step-by-step. Similarly to the present paper, the system consisted of an inverter and a grid-impedance. However, an external device was used to inject the current perturbations.

### B. Existing solution

The impedance coupling can be avoided by using multiple independent injections in the measurement [10], [11], [17], [25], [26]. Here this is accomplished by first injecting the MLBS to $i_{Ld}^{ref}$, while the injection to $i_{Lq}^{ref}$ is zero. The second MLBS injection is made to $i_{Lq}^{ref}$ while the injection to $i_{Ld}^{ref}$ is zero. This method is denoted here by "2MLBS" because it is implemented by two sequential MLBSs. All currents and voltages are recorded from both injections. It has to be assumed that the grid-impedance does not change between the measurements and two sets of equations can be written [10]:

$$
\begin{align*}
V_{d1} &= Z_d I_{d1} + Z_{q1} I_{q1} \\
V_{q1} &= Z_q I_{d1} + Z_{dq} I_{d1}
\end{align*}
$$  (12)

$$
\begin{align*}
V_{d2} &= Z_d I_{d2} + Z_{q2} I_{q2} \\
V_{q2} &= Z_q I_{d2} + Z_{dq} I_{d2}
\end{align*}
$$  (13)

where subscript 1 and 2 denote the first and the second injection, respectively. The capital letters are used to denote the discrete Fourier transform (DFT) of the variables. The equations can be presented in matrix form [10]:

$$
\begin{bmatrix}
V_{d1} \\
V_{q1}
\end{bmatrix} = \begin{bmatrix} Z_d & Z_{q1} \\ Z_{dq} & Z_q \end{bmatrix} \begin{bmatrix} I_{d1} \\
I_{q1}
\end{bmatrix}
$$  (14)  

Eq. 14 can be easily solved for the impedance elements [10]:

$$
\begin{bmatrix}
Z_d & Z_{q1} \\ Z_{dq} & Z_q \end{bmatrix}^{-1} = \begin{bmatrix} V_{d1} & V_{d2} \\ V_{q1} & V_{q2} \end{bmatrix} \begin{bmatrix} I_{d1} \\
I_{q1}
\end{bmatrix}
$$  (15)

Figs. 8 shows the resulting impedance elements based on the two independent measurements ($Z_d^{2MLBS}$, $Z_q^{2MLBS}$, $Z_{dq}^{2MLBS}$, and $Z_{qd}^{2MLBS}$) and results obtained by the simple ratios of voltages to currents (the injection for measurement of $Z_{dq}^{2MLBS}$ and $Z_{qd}^{2MLBS}$ was implemented by adding the IRS on top of the q current reference). The undesired impedance-based interaction is clearly no longer present in the measurement results. However, the main shortfall of the "2MLBS" method is the requirement of two sequential independent injections at the same frequencies. As a consequence, the measurement is more prone to changes in the operating point between the
measurements. Below, an identification method is proposed that is based on two orthogonal injections, which allows the independent measurements to be performed simultaneously. As a result, similar results to the conventional method can be achieved without having to perform sequential injections.

IV. NOVEL METHOD

This section begins by inspecting a simple interpolation method to obtain an independent measurement at the MLBS frequencies from the IRS measurement. It is noted that the pseudorandom behavior in the IRS phase causes problems. To tackle this problem, a procedure where transfer functions from the injected sequence to the voltages and currents are used in the interpolation is proposed. The disturbance rejection capability of the proposed technique is demonstrated at the end of the section.

A. Interpolation of IRS result in frequency domain

The objective is to obtain two sets of independent measurements at the same frequencies. However, the orthogonal injections, MLBS and IRS, have different frequency vectors, by definition. The solution is to interpolate the results from either of the simultaneously injected MLBS or IRS to the frequencies of another in the frequency domain. The interpolation of the IRS according to (1) and (2), the interpolation becomes a straightforward arithmetic mean.

The frequency domain I\(_d\), V\(_d\), I\(_q\), and V\(_q\) are interpolated to the frequencies of the MLBS, f\(_{MLBS}\), from the frequencies of the IRS, f\(_{IRS}\). The interpolation of V\(_q\) is given as an example:

\[
V_{q2}^{\text{intrpl}} = \frac{V_{q2} - f_{IRS}}{k} + \frac{V_{q2}^{f-IRS-k+1}}{2}
\]

which is the average of the complex-valued frequency bins, and superscript "intrpl" is used to denote that the variable is a result of the interpolation. The interpolated values are substituted into (15), and the impedance elements based on the interpolation (Z\(_{d}^{\text{intrpl}}\), Z\(_{q}^{\text{intrpl}}\), Z\(_{dq}^{\text{intrpl}}\), and Z\(_{q}^{\text{intrpl}}\)) are solved.

Fig. 9a compares the simulated identification results of the interpolation method (Z\(_{d}^{\text{intrpl}}\)) to the traditional measurement method (Z\(_{d}^{2MLBS}\)). In the case of the d component, the result is acceptable. Z\(_{q}^{\text{intrpl}}\), Z\(_{dq}^{\text{intrpl}}\), and Z\(_{q}^{\text{intrpl}}\) are shown in Figs. 9b, 9c, and 9d, respectively. While the impedance coupling is no longer present, it can be seen that the interpolated results clearly have deviations from the traditional method.

The response in V\(_q\) originates from the injected IRS. As shown in Fig. 2b, the phase behavior of the IRS is pseudorandom. Fig. 10a shows a Bode plot of the \(f_{Lq2}^{\text{ref-IRS}}\) and its interpolation (\(f_{Lq2}^{\text{ref-intrpl}}\)). The figure clearly shows that the interpolation result of the complex-valued bins gives mostly no reasonable result. This is due to the large changes in the...
functions from the q current reference (Iq2) and its interpolation, (b) Vq2 and its interpolation, (c) Vq2/Iq2 and its interpolation, and (d) comparison of Vq2/Iq2 and G_{TV-intrpl}. The results are from a Simulink simulation.

Figure 10: (a) $I_{q2}^{ref}$ and its interpolation, (b) $V_{q2}$ and its interpolation, (c) $V_{q2}/I_{q2}^{ref}$ and its interpolation, and (d) comparison of $V_{q2}/I_{q2}^{ref}$ and $G_{TV-intrpl}$. The results are from a Simulink simulation.

Because the transfer functions are based on the system dynamics, it can be assumed that there are no abrupt changes within the resolution of the measurement. The grid-impedance d component ($Z_d$) is used as an example. The solution to the grid-impedance d component from (15) is calculated open:

$$Z_d = \frac{V_{d1}I_{q2} - I_{q1}V_{d2}}{I_{d1}I_{q2} - I_{q1}I_{d2}}$$  \hspace{1cm} (18)

The current and the voltages related to the second independent measurement can be replaced by the corresponding transfer functions from the q current reference ($I_{q2}^{ref}$) to the voltages (6) and the currents (3):

$$Z_d = \frac{V_{d1}G_{co-q}I_{q2}^{ref} - I_{q1}G_{cv-q}V_{q2}^{ref}}{I_{d1}G_{co-q}I_{q2}^{ref} - I_{q1}G_{cv-q}I_{q2}^{ref}}$$  \hspace{1cm} (19)

Because the injected IRS to $I_{q2}^{ref}$ and the responses at the IRS frequencies are known, the unknown transfer functions can be solved at the frequencies of the IRS as

$$\left[ \begin{array}{c} G_{IRS}^{co-qd} \\ G_{IRS}^{cv-qd} \\ G_{IRS}^{co-q} \\ G_{IRS}^{cv-q} \end{array} \right] = \left[ \begin{array}{c} I_{IRS}^{d2} \\ V_{IRS}^{q2} \\ I_{d1} \\ V_{q2}^{q2} \end{array} \right] \frac{1}{I_{q2}^{ref-IIRS}} \left[ \begin{array}{c} G_{IRS}^{co-qd} \\ G_{IRS}^{cv-qd} \\ G_{IRS}^{co-q} \\ G_{IRS}^{cv-q} \end{array} \right]^{-1}$$  \hspace{1cm} (20)

The equation also includes $G_{IRS}^{cv-q}$ because it is needed in order to identify $Z_q$ and $Z_{dq}$. Because the transfer functions are at the frequencies of the IRS, Eq. (19) cannot be directly applied. However, the solved transfer function can be interpolated to the frequencies of the MLBS by using (17). The interpolated transfer functions can be used to replace the second independent injection results in (15), which gives solutions to all impedance elements:

$$\left[ \begin{array}{c} Z_{TF-intrpl}^{d} \\ Z_{TF-intrpl}^{dq} \\ Z_{TF-intrpl}^{d1} \\ Z_{TF-intrpl}^{q1} \end{array} \right] = \left[ \begin{array}{c} V_{d1}G_{cv-qd}^{TF-intrpl} I_{q1}G_{cv-q}^{TF-intrpl} \end{array} \right]^{-1}$$  \hspace{1cm} (21)

The solution to $Z_d$ is given as:

$$Z_d^{TF-intrpl} = \frac{V_{d1}G_{cv-qd}^{TF-intrpl} I_{d1}G_{cv-q}^{TF-intrpl} - I_{q1}G_{cv-q}^{TF-intrpl}}{I_{d1}G_{cv-q}^{TF-intrpl} - I_{q1}G_{cv-q}^{TF-intrpl}}$$  \hspace{1cm} (22)

Fig. 11 presents Simulink simulation results obtained by the proposed measurement procedure that is presented as a flow chart in Fig. 12. The identified $Z_d$, $Z_{qd}$, $Z_{dq}$, and $Z_q$ are shown in Figs. 11a, 11b, 11c, and 11d, respectively. The results obtained by the proposed method (TF-intrpl) are almost as accurate as those obtained by the traditional method (MLBS).
deviations caused by the phase problem that is not present in the proposed transfer-function interpolation method.

In order to provide more validation of the result, a fit ratio is computed as follows:

\[
FR = \left(1 - \frac{\sum_{k=1}^{K} |Z_{\text{reference}}(k) - Z_{\text{identified}}(k)|^2}{\sum_{k=1}^{K} |Z_{\text{reference}}(k)|^2}\right) \times 100\% \tag{23}
\]

where \(Z_{\text{reference}}\) is the reference to which the identification result \(Z_{\text{identified}}\) is compared, and \(K\) equals 256 when \(n = 9\) (because 256th element of the frequency vector is close to 2 kHz that is the reasonable measurement bandwidth). Here the fit ratio of the identification results obtained by the proposed technique to the reference model of the grid-impedance is computed. For \(Z_{\text{d,TF-intrpl}}\), \(Z_{\text{q,TF-intrpl}}\), \(Z_{\text{e,TF-intrpl}}\), and \(Z_{\text{q,TF-intrpl}}\), the fit ratios are 99.96 %, 99.47 %, 99.63 %, and 99.95 %, respectively. These numbers further validate the observations from the Bode plots in Fig. 11.

C. Disturbance rejection

The benefit of the proposed method is that both channels are injected and the impedance elements measured under the same operating conditions which provides rejection against effects of disturbances in the system during the measurement. In order to put the disturbance rejection to test, 5th, 7th, 11th, and 13th harmonics are added in the grid voltages during the measurement, and the identification results between the proposed method and the traditional method are compared.

Fig. 13 shows the grid-voltages and illustrates how the harmonics in the voltages affect the traditional and the proposed technique. The phase voltages are denoted by \(v_{g-a}\), \(v_{g-b}\), and \(v_{g-c}\), respectively. In the traditional method, the harmonics affect the measurement of the first MLBS when the d channel is injected. This implies that the DFT of the d channel will be distorted. The measurement of the q channel injection is not affected by the harmonics.

In the case of the proposed method in which the orthogonal sequences are applied, the effect of the harmonics in the grid voltages is different. The length of the IRS equals two times the length of the MLBS. Therefore, the proposed method and the traditional method have the same measurement time.
Figure 14: Comparison of the simulated identification results with the existing method and the proposed method under temporary harmonics in the grid voltages during the identification.

Figure 15: Experimentally identified grid impedance including the filter capacitor. (a) $Z_d$, (b) $Z_{qd}$, (c) $Z_{dq}$, and (d) $Z_q$.

Table II: Comparison of the fit ratio of the simulated results obtained by proposed method (TF-intrpl) and traditional (2MLBS) method to reference under grid voltage harmonics.

<table>
<thead>
<tr>
<th>Impedance (Fig. 14)</th>
<th>Traditional method (2MLBS)</th>
<th>Proposed method (TF-intrpl)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_d$</td>
<td>99.95 %</td>
<td>99.96 %</td>
</tr>
<tr>
<td>$Z_{qd}$</td>
<td>99.46 %</td>
<td>99.05 %</td>
</tr>
<tr>
<td>$Z_{dq}$</td>
<td>96.98 %</td>
<td>98.96 %</td>
</tr>
<tr>
<td>$Z_q$</td>
<td>99.98 %</td>
<td>99.95 %</td>
</tr>
</tbody>
</table>

However, in the proposed method, both channels are simultaneously injected and measured over the whole measurement time. Therefore, the harmonics affect the measurement of both channels but the time average is over a longer period of time in the case of the IRS. In the case of the MLBS, two sequences are averaged for every IRS period.

Fig. 14 shows simulated identification results in the case of the additional harmonics. In order to facilitate the comparison, fit ratios (23) are calculated for both methods (shown in Table II). The direct components ($Z_d$ and $Z_q$) are equal with both methods. In the case of $Z_{qd}$, the traditional method is slightly better than the proposed method. In the case of $Z_{dq}$, the results obtained by the traditional method have the largest deviations among all the results with both methods. In the proposed method, due to the good averaging capability, large deviations are more efficiently avoided than in the traditional method.

V. VERIFICATION OF THE PROPOSED TECHNIQUE

This section provides experimental verification for the proposed technique. First, a grid impedance that has a resonance is identified by the grid-connected inverter. Then, the output admittance of a grid-connected PV inverter is identified.

A. Identification of Grid Impedance Including Filter Capacitor

The used measurement setup is shown in Fig. 1. However, the identification procedure is changed so that the filter inductor current measurement is used for the identification. Therefore, the filter capacitor is included in the grid impedance, and the grid impedance becomes a parallel LC circuit with a resonance.
Fig. 15 shows that the method that is based on ratios of the voltages to the currents clearly gives the results that are not the impedance elements of an LC circuit in the dq frame. In the dq frame, a parallel LC resonance has two peaks that are separated by two times the fundamental frequency. Using the measurement method based on ratios of the voltages and currents, the resonance peaks appear combined in the measured impedance. Furthermore, in the direct components \( (Z_d \text{ and } Z_q) \) there are additional resonances at around 120 Hz. The existing method and the proposed method give practically identical results, and the interpolation does not cause problems.

The fit ratio is computed according to (23), and the results from the traditional method (2MLBS) are used as the reference to which the proposed method (TF-intrpl) is compared. Table III shows the fit ratios that clearly verify the applicability of the proposed method.

### B. Identification of PV Inverter Output Admittance

To further verify the proposed measurement procedure, the output admittance of a photovoltaic (PV) inverter is measured. Fig. 16 shows a block diagram of the system. The inverter is similar to the inverter in Fig. 1; however, the inverter is fed by a PV panel emulator instead of a DC voltage source. Therefore, a DC-voltage controller that produces the reference to the current \( d \) component is required. For the admittance identification, the currents and voltages are measured at the terminals of the isolation transformer; the transformer impedance is included to the converter output admittance that is identified. A measurement (meas.) PLL is used to synchronize the voltages at the transformer. The MLBS and IRS are added to the \( d \) and \( q \) voltage references of the grid emulator that is connected to the transformer over a line inductance. The interaction of the line impedance and converter impedance may cause an intended \( d \)-channel injection to appear in the \( q \)-channel and the impedance coupling occurs. The values of the passive components are given in Table I, and the controller parameters of the PV inverter and the measurement PLL that is used in the identification are given in Appendix B.

The measurement procedure is slightly modified compared to the previously used procedure because now the inverter output admittance is identified. Eq. (21) becomes

\[
\begin{bmatrix}
Y_d & Y_{dq} & Y_q
\end{bmatrix} = \begin{bmatrix}
I_{d1} & G_1 & I_{d2}
I_{q1} & G_3 & I_{q2}
\end{bmatrix} \begin{bmatrix}
V_{d1} & G_2
V_{q1} & G_4
\end{bmatrix}^{-1}
\]

where \( G_1, G_2, G_3, \) and \( G_4 \) are TF-interpolated transfer functions from \( V_{q2}^\text{ref} \) to \( I_{d2}, V_{d2}, I_{q2}, \) and \( V_{q2} \), respectively. The transfer functions are calculated with respect to the reference \( q \)-voltage \( V_{q2}^\text{ref} \) because the IRS is injected to the \( q \)-voltage reference of the grid-emulator. The amplitudes of the injected MLBS and IRS are 10 V, and the length of the shift register \((n)\) is increased from the previously used 9 to 10 in order to improve the frequency resolution of the MLBS from 7.8 Hz to 3.9 Hz.

Fig. 17 shows the identified PV inverter output admittance obtained by different methods. Clearly, the method based on ratios of two variables deviates from the traditional method (2MLBS) and the proposed (TF-intrpl). As Fig. 17a. shows, the resolution of the proposed method is sufficient to reveal the resonance at 23 Hz caused by the DC-voltage control bandwidth. The negative incremental resistance region caused by the PLL whose bandwidth is 20 Hz is visible in \( Y_{d,q} \) in Fig. 17d.

Table III shows the fit ratio of the proposed methods to the traditional method. In (23), \( K \) equals 512 because \( n = 10 \). In the case of PV inverter, there is more deviation in the fit ratio than in the case of the grid impedance. The reason for this deviation is that the relatively high line impedance damps the voltage perturbations that are implemented by the grid emulator. Furthermore, the converter admittance has a low magnitude resulting in a low current response that is prone to noise. Both 2MLBS and TF-intrpl methods suffer from the noise in the frequency range from 100 Hz to 1000 Hz. Nevertheless, the proposed method tackles the impedance coupling problem. The results could be enhanced by averaging.

### Table III: Fit ratio of the proposed method (TF-intrpl) to the (2MLBS) method in experiments.

<table>
<thead>
<tr>
<th>Impedance (Fig. 15)</th>
<th>Fit ratio (FR)</th>
<th>Admittance (Fig. 17)</th>
<th>Fit ratio (FR)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( Z_{d} )</td>
<td>99.7 %</td>
<td>( Y_{d} )</td>
<td>98.4 %</td>
</tr>
<tr>
<td>( Z_{dq} )</td>
<td>99.2 %</td>
<td>( Y_{dq} )</td>
<td>96.0 %</td>
</tr>
<tr>
<td>( Z_{dq} )</td>
<td>99.3 %</td>
<td>( Y_{dq} )</td>
<td>91.6 %</td>
</tr>
<tr>
<td>( Z_{q} )</td>
<td>99.7 %</td>
<td>( Y_{q} )</td>
<td>94.7 %</td>
</tr>
</tbody>
</table>
over a longer period of time or by increasing the injection amplitude.

The experiments verify the applicability of uncorrelated broadband injections in the presence of impedance coupling. The benefit of the proposed procedure lies in the short measurement time due to the use of the orthogonal binary perturbations. Furthermore, the method ensures that each channel is injected with the system in the same operating conditions, which may not be the case if sequential perturbations are applied. On the other hand, the method requires more signal processing than the existing method based on sequential injections. The reference signal must be recorded in the same reference frame as the voltages and currents which can set a limit for the PLL bandwidth. Another drawback of the interpolation is the introduction of small error. However, the interpolation error can be kept small by using an adequate frequency resolution.

VI. CONCLUSIONS

The identification of the grid impedance at the point of common coupling of grid-connected converters is advantageous for the controller design and the adaptive control. An injection can be added to the converter current reference, and the identification can be made from the responses in the terminal currents and voltages. The interaction of the converter output impedance and the grid impedance causes an error in the impedance identification if an element of the multiple-input-multiple-output impedance is identified under an assumption that the injected perturbation affects only the intended channel.

In this work, orthogonal binary injections are used to obtain two independent sets of measurements from a grid-connected converter system. Through the use of interpolation, one set of measurements can be moved to the frequencies of the other as an independent measurement. However, due to the pseudorandom phase behavior of the binary sequences, inferior results are obtained if the measured variables are interpolated directly in the frequency domain. In the proposed method, the variables are divided by the injected sequence and the obtained transfer functions are interpolated. The two acquired sets of measurements can be used to accurately identify the grid impedance by using existing methods. The presented method allows simultaneous measurement of the two independent measurement sets. The benefit lies in the fast measurement process and the mitigation of the possibility of excessive disturbances or changes in the grid impedance during the measurement.

REFERENCES


Matias Berg (S’17) received the B.Sc. (Tech.) and M.Sc. (Tech.) degrees in Electrical Engineering from Tampere University of Technology in 2015 and 2017, respectively. He is currently a doctoral student at Tampere University, and his research interests include dynamic modeling of grid-following and grid-forming converters.

Henrik Alenius (S’18) received his M.Sc. degree in electrical engineering from Tampere University of Technology, Tampere, Finland, in 2018. Since then, he has worked as a doctoral student with the Faculty of Information Technology and Communication Sciences at Tampere University. His research interests include impedance-based interactions in grid-connected systems, broadband methods in impedance measurements, and stability analysis of multi-parallel inverters.

Tomi Roinila (M’10) received the M.Sc.(Tech.) and Dr.Tech. degrees in automation and control engineering from Tampere University of Technology, Tampere, Finland, in 2006 and 2010, respectively. He is currently an Assistant Professor in Tampere University, Finland.

His main research interests include modeling and control of grid-connected power-electronics systems, analysis of energy-storage systems, and modeling of multi-converter systems.