Wide Bandwidth System Identification of AC System Impedances by Applying Perturbations to an Existing Converter

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Abstract—A new approach is developed using an existing single-phase power electronics inverter to make a wideband impedance measurement at its interface using a pseudo-random binary sequence (PRBS) voltage perturbation and digital network analyzer techniques. Since the PRBS is an approximation to white noise, all frequencies of interest can be excited simultaneously. The perturbation at the interface is measured and cross-correlation techniques are applied to construct the wideband impedance of the system under test. Online monitoring of interface impedances is a key enabler for a number of smart-grid related capabilities, such as grid health monitoring, active filter re-tuning, identifying system interaction, and adaptive control of grid-connected switching converters. Knowledge of the converter's surroundings enables smarter control actions, which lead to improved stability, performance and reliability of the smart grid.

I. INTRODUCTION

In recent times there has been significant interest in the wideband measurement of power system impedances [2-7]. With the increase of nonlinear harmonic-producing loads and high-switching-frequency power electronics converters, the power distribution system is becoming a wideband network; fundamental frequency impedances are no longer enough to describe the entire system. Knowledge of harmonic and high-frequency impedances is necessary in order to predict harmonic propagation and voltage distortion at various locations on the power network, to identify harmonic polluters and to predict resonances. Moreover, it can improve the effectiveness and stability of active filters, in particular if the impedance can be measured in real time at the active filter location. Additionally, it allows the detection of high-impedance ground faults and of slight system unbalance over wide frequencies. Note that Phasor Measurement Units (PMUs) cannot provide these wideband measurements, because they are restricted to measurements at 60Hz and its harmonics, since they rely on spectral content that already exists in the system.

Previous works have used dedicated hardware platforms for system identification [1, 2, 4, 6]; however, it would be desirable to be able to perform the wideband impedance measurement using existing equipment present in the system to minimize cost and complexity. We propose to use an existing power electronics inverter to inject a small-amplitude white-noise perturbation and use cross-correlation based digital network analyzer techniques to measure the system impedances. Although not a passive method, the injected perturbation magnitude is small in comparison to system voltage levels and is active only while a measurement is being performed. Existing small-signal injection techniques for AC impedance measurement [2-7] use a composite of reference steps [3-5] or a swept sinusoidal injection [6] to elicit a transient response of the system. The proposed method takes advantage of the fact that white noise injection excites all frequencies with equal magnitude, which provides the most information, reduces measurement time, improves the signal to noise ratio (SNR) of the measurement and simplifies post processing [12].

This work is an extension of a previous study [9,10] of online network identification using existing power converters in DC systems using correlation-based analysis. Good matching was achieved in previous experiments, but modifications to the basic technique are required to extend the method to single phase AC systems. This work also leverages some of the existing improvements to the methodology such as, windowing, blue noise excitation, and oversampling [10]. Also new improvements are made to reduce injection times and simplify estimation calculations.

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II. IMPEDANCE IDENTIFICATION

A switching converter operating at steady state can be considered a linear time invariant system to small-signal disturbances (1). The sampled system can be described by:

\[ y[n] = \sum_{k=1}^{\infty} h[k] u[n-k] + v[n] \]  \hspace{1cm} (1)

where \( y[n] \) is the sampled output signal, \( u[n] \) is the sampled input signal, \( h[n] \) is the discrete-time system impulse response, and \( v[n] \) represents unwanted disturbances such as switching and quantization noise. The cross-correlation of the input control signal \( u[n] \) and the output signal \( y[n] \) is defined in (2) as:

\[ R_{uy}[m] = \sum_{n=1}^{\infty} u[n] y[n+m] \]

\[ = \sum_{n=1}^{\infty} h[n] R_{uu}[m-n] + R_{uv}[m] \]  \hspace{1cm} (2)

where \( R_{uu}[m] \) is the auto-correlation of the input signal, \( R_{uy}[m] \) is the input-to-output cross-correlation, and \( R_{uv}[m] \) is the input-to-disturbance cross-correlation. White noise input exhibits the following properties:

\[ R_{uu}[m] = \delta[m] \]
\[ R_{uv}[m] = 0 \]  \hspace{1cm} (3)

where \( \delta[m] \) is the discrete impulse signal. These properties allow simplification of equation (2) such that the input-to-output cross-correlation becomes the discrete-time system impulse response (4). Taking the DFT yields the input-to-output transfer function (5)

\[ R_{uy}[m] = h[m] \]
\[ G_{vy}[e^{j\omega}] = DFT\{h[m]\} \]  \hspace{1cm} (4)  \hspace{1cm} (5)

The measurement of impedance requires a high-bandwidth voltage or current perturbation at the interface and measurements of both voltage and current. Here, the perturbation is accomplished through injection of a white noise approximation (PRBS) into the appropriate actuator command signal. This actuator is the inverter itself (Fig. 1), so that impedances looking outward from the inverter can be measured. Since impedance is defined as the ratio of voltage variations to current variations, this ratio must be constructed from two measurable transfer functions: control-to-voltage transfer function, \( G_{vd}[e^{j\omega}] \), and control-to-current transfer function, \( G_{id}[e^{j\omega}] \). This construction is shown in equation (6).

\[ \frac{G_{vd}[e^{j\omega}]}{G_{id}[e^{j\omega}]} = \left( \frac{\hat{v}[e^{j\omega}]}{\hat{i}[e^{j\omega}]} \right) = \hat{Z}[e^{j\omega}] \]

Extending (7) we can conclude

\[ \frac{DFT\{R_{uy}[m]\}}{DFT\{R_{uv}[m]\}} = \left( \frac{DFT\{u[n]\}}{DFT\{i[n]\}} \right)^{\ast} DFT\{y[n]\} \]

\[ = \frac{DFT\{v[n]\}}{DFT\{i[n]\}} = Z[e^{j\omega}] \]  \hspace{1cm} (8)

Although not deeply discussed in this paper, when attempting to implement identification on an embedded platform, it is beneficial to eliminate any unnecessary calculations. When the desired result is the frequency response of impedances, there are shortcuts that can be made to reduce calculation time. First, correlation is a time domain technique that when explicitly calculated requires \( O(N^2) \) operations. The DFT is then applied to find the frequency domain response. These operations are carried out to find two quantities, \( G_{vd}[e^{j\omega}] \) and \( G_{id}[e^{j\omega}] \), and the ratio of these quantities is taken to find \( Z[e^{j\omega}] \). As with the convolution theorem, it is found that DFT of the cross correlation is equivalent to the complex conjugate of the DFT of the input multiplied by the DFT of the output response:

\[ DFT\{R_{uy}[m]\} = (DFT\{u[n]\})^{\ast} DFT\{y[n]\} \]  \hspace{1cm} (7)

Similarly, for identification of impedances, the DFT of the converter voltage divided by the DFT of the converter current form the equivalent impedance, therefore computationally efficient FFT’s can be utilized to estimate system impedances.

In summary, the converter merely acts as a power amplifier of the test sequence, and the sensing and signal processing techniques used provide an equivalent result as that of a Network Analyzer.

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III. OVERCOMING CHALLENGES IN IMPEDANCE IDENTIFICATION

A. Reducing the Length of Injection

As noted in [10], when applying the cross-correlation technique described above to finite-length sampled data, property (3) is only approximately satisfied, i.e., the autocorrelation $R_{uu}[m]$ is not an ideal discrete impulse due to non-cancellation of terms. Previously, this was corrected by applying a Gaussian window in order to de-emphasize the introduced error. However, rather than padding the sampled data with zeros, if a circular cross-correlation is used, it can be found that (3) holds. Fig. 2 shows the standard auto correlation on the left and the circular auto correlation on the right. For the cross correlation to be circular, we must impose that the PRBS injection be periodically extended over the sampling interval. This simply means that the sampling interval must include the entire PRBS chosen with the addition of the first bit repeating at the end of the sequence. This eliminates the need for the Gaussian window averaging that was previously needed and also reduces the need for multi-period injection.

![Figure 2. Cross Correlation (left) vs. Improved Circular Correlation Technique (right)](image)

B. Reducing Injection Amplitude and Improving SNR

Another major challenge is choosing an appropriate perturbation magnitude over the frequency range of interest taking into account the converter’s input/output filter. The attenuation introduced by this filter dictates a minimum injection amplitude, so that the response can be measured through the filter’s attenuation.

1) Appropriate ADC Selection

Since the measurements are made with ADC’s, their effective resolution is also a determining factor as to the injection magnitude. For example, 8-bit ADC’s have 256 discrete values that can be measured over its range. Assuming the converter is grid tied with a measurement range of ±200V, the ADC resolution is 1.56V. This means that the response on the output must be at least 1.56V to even be measured, let alone measured accurately. This can be deemed an unacceptable disturbance depending on the situation. However, as most modern microcontrollers have higher resolution ADCs, the quantization errors can be minimized. With a 12-bit ADC, the resolution increases to 97.6mV and with a 16-bit ADC, the resolution is 6mV enabling both accurate measurements and smaller injected disturbances.

<table>
<thead>
<tr>
<th>ADC Bits</th>
<th>Min. Measurable Voltage (±200V full-scale)</th>
<th>Effective Resolution (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>8</td>
<td>1.56 V</td>
<td>48.165 dB</td>
</tr>
<tr>
<td>12</td>
<td>97.6 mV</td>
<td>72.247 dB</td>
</tr>
<tr>
<td>16</td>
<td>6.10 mV</td>
<td>96.230 dB</td>
</tr>
</tbody>
</table>

2) Alternative Perturbation Signals

Since the converter’s output filter tends to attenuate higher frequency signals, measuring response at high frequencies can be challenging. If white noise injection is used, in order to increase the measurable content at high frequencies, the injection magnitude has to be increased at all frequencies which may cause significant output distortion. However, alternative perturbation sources can be used that have non-flat signal spectra. An example of frequency dependent signal spectra is blue noise, which contains larger magnitude high frequency content. Using blue noise, the low frequency injection amplitude can be reduced while still providing enough excitation at high frequencies, thereby reducing the distortion seen at the output compared to white noise injection.

![Figure 3. Discrete Blue noise implementation](image)

![Figure 4. Using Blue noise excitation to bring high frequency measurements above the noise floor](image)
A blue noise approximation can be formed as a linear combination of the PRBS signal and its first-order discrete derivative as shown in Fig. 3. The discrete-time output equation is presented in (9) and its z-transform is shown in (10).

\[ y[n] = k_1(x[n] - x[n-1]) + k_2x[n] \quad (9) \]

\[ \frac{Y(z)}{X(z)} = (k_1 + k_2) - k_1z^{-1} \quad (10) \]

The DC gain of this filter is \( k_2 \) and the high-frequency gain is \( 2k_1 + k_2 \). By choosing constants \( k_1 \) and \( k_2 \), the user can independently adjust the low-frequency and high-frequency test signal magnitude. It is then possible to ensure that the high-frequency signal injection magnitude is sufficiently large to be above the noise floor without making the low frequency magnitude excessively large, as shown in Fig. 4.

**IV. EXTENSION TO AC SYSTEMS**

In the present work only single phase linear grid impedances are considered, so that a linearization step is not required. For DC-DC converters, the steady-state duty cycle is constant and the PRBS perturbation is added to this constant value. For AC systems, the steady-state duty cycle varies sinusoidally in synchronism with the AC voltage (SPWM). The perturbation signal is injected in addition to the nominal SPWM duty command for the inverter. The major difficulty is ensuring synchronization with the AC system; notice however, that grid connected inverters already have the necessary controls in the form of PLL’s or other techniques to achieve phase synchronization.

Fig. 5 illustrates a simulation using Simulink SimPower Systems block set. An H-Bridge IGBT inverter switching at 12 kHz is connected with a filtering inductor of 10mH to a simplified representation of an AC system consisting of an ideal voltage source and an R-L line impedance of 10Ω and 10mH. Initially the PRBS signal is disabled and the inverter operates under steady-state conditions. The identification procedure is initiated at simulation time \( t=1.0s \), when the PRBS is enabled, and is added to the SPWM reference, providing wideband excitation to the system under test (Fig. 6). Looking at the inverter’s output current (point of common coupling (PCC) current in Fig. 6), clearly the additional PRBS does not add a significant amount of noise to the system. Applying the cross correlation between the input and output voltage and current and taking the DFT, the output impedance of the system is obtained (Fig. 7). Excellent matching is achieved even in the presence of ZOH sampling and non-linear switching noise. A caveat is that the maximum identifiable frequency is limited to half of the converter’s sampling rate (Nyquist frequency), 6 kHz in this case. The minimum identifiable frequency is the inverse of the time duration of the PRBS injection. Clearly this can be tailored to different applications where low frequencies may or may not be of interest.

In DC-DC converter based identification, the subsystem under test operating in steady-state was assumed to have no existing spectral content except at DC; therefore, it was assumed that the only AC excitation (besides low-amplitude noise) was due to the PRBS perturbation introduced in the switching converter control signal. In grid connected applications, there exists an established system operating point that varies with location and time. We find that, since the PRBS does not fully approximate white noise, the correlation of the PRBS to any disturbances does not equal zero, as assumed in (3). When calculating impedances under grid-tied conditions, special care must be given to deal with these existing signals. If the AC source voltage were precisely known, it would be possible to subtract this signal from the measured PCC voltage. However, since the grid-
tied converter only has access to the PCC, exact knowledge of the AC source voltage cannot be assumed. Since we cannot use this information to null the existing signal spectrum, an approximation must be made.

There are two reasonable options to achieve accurate impedance estimation under pre-existing operating points.

1) Before the inverter goes online, using the PCC sensed quantities, measure the existing voltage spectrum at the PCC to establish a baseline of the subsystem operating conditions. After the converter has injected the test sequence, subtract off the existing spectrum from the spectrum obtained under injection. This will leave only the converter’s response. Accurate knowledge of existing signal spectrum magnitude and phase is needed.

2) Since the converter injects a wideband signal spectrum, accurate measurements are obtained at off-harmonic frequencies, where there is generally little to no existing spectrum. Simply disregard all fundamental and harmonic frequencies and extrapolate using the off-fundamental response. Since the excitation spectrum is uniform across the injection frequency range, very good approximations can be made assuming the impedance does not wildly vary around the fundamental and harmonic frequencies.

The output filter consists of a 1mH series inductor and a 200µF parallel capacitor. As previously discussed, the filter choice is important in order to provide a balance between switching ripple attenuation and excitation attenuation. Since the perturbation applied by the converter must be able to pass through this LC filter, a compromise was made in order to not excessively attenuate the perturbation signal. After the LC filter, an auto transformer is used to step the inverter output up to grid levels (60Hz 120V RMS). For safety reasons, a 1:1 isolation transformer is used at the grid tie. The inverter is fed from a 100V 10A DC source.

B. Control

A TI F28335 DSP Control Card was used due to its modular nature, allowing for future expansion into higher power setups. A custom docking board with current and voltage sensing as well as digital I/O was designed and constructed to provide flexible measurement capability. The control scheme used for the inverter is a deadbeat current controller fed by a sinusoidal current reference. The current reference is created from the PLL which is implemented using a Second Order Generalized Integrator (SOGI) [13] that introduces an orthogonal coordinate system. A DQ transformation is applied, and the D-axis feeds a voltage controlled oscillator. The inverter is modulated using a unipolar scheme that doubles the effective frequency of the current ripple.

VI. EXPERIMENTAL RESULTS

For this paper, only low power levels (50-100W) were used. This was deemed to be a good starting level because enough power is used to provide real world data while still operating at low voltages. Future work will increase power into the kW level for single phase applications and will eliminate the step-down auto transformer. Also, in order to provide accurate impedance estimations, the DSP’s ADC’s were not used. Instead the estimation is constructed by measuring the output voltage and current response with an Agilent 4395A Vector Spectrum Analyzer. Future work will focus on full implementation with a DSP.

Since an AC power source was not available allowing a very well defined source and source impedance, a different approach was used (Fig. 8). Using the auto transformer with a step up ratio of 10, any grid impedance is effectively reduced by a factor of 100. This allows very well controlled passive linear impedances to be inserted between the inverter and transformer (“insertion impedance” element in Fig. 8). The inverter will see the series combination of the inserted impedance, the auto transformer impedance, and 1/100th of the sum of the isolation transformer and grid impedance. This allows the load to be approximated as the insertion impedance, if it is significantly larger than that of the auto transformer. The insertion impedance can therefore be measured with an Impedance Analyzer when non-grid tied and compared to the impedance estimated by the converter. Three different insertion loads were tested:
A. Case of R-L Insertion Impedance

![Figure 10. R-L Insertion Impedance](image)

The first insertion impedance to be tested was an R-L circuit, consisting of a 1mH air core inductor in series with a 2.5Ω wire wound power resistor (Fig. 10). This impedance can be considered as an approximation to a line impedance model at low frequencies. Fig. 11 shows the inductor current (top), output current (middle) and output voltage (bottom) before and after injection, where the vertical line indicates the start of the injection. It is important to note that since the LC filter is present, much of the injected test signal does not appear on the output current and voltage waveforms. This is desirable in that very little noise is injected into the grid, but it can cause SNR problems if low bit count ADCs are used, as discussed in Section III.

Fig. 12 illustrates the ADC resolution issue. The output response is measured with an 8-Bit oscilloscope and compared to the Vector Spectrum analyzer. We can see that the oscilloscope captures the voltage and current spectrum in a very loose band with significant superimposed quantization noise due to its inability to precisely measure the response, while the Vector Spectrum analyzer is able to measure the response in a very tight range with low quantization noise due to its superior resolution. The impedance (bottom) is calculated for both cases and is compared to a third measurement, where the impedance is measured directly using the Agilent 4395A in Impedance Analyzer mode. The 8-Bit resolution causes corruption in the estimated impedance, especially at higher frequencies where the SNR is small.

![Figure 11. Inductor current (top 2A/Div), output current (middle 2A/Div), output voltage (bottom 5V/Div) before and after test sequence injection](image)

![Figure 12. Measured voltage spectra (top), current spectra (middle), and impedance (bottom) using an Oscilloscope, Spectrum Analyzer, and Impedance Analyzer.](image)

![Figure 13. Output voltage spectra (top), current spectra (middle), and impedance (bottom) for three different cases: 1) Inverter operating non-grid-tied with 0A ref, 2) non-grid-tied with 5A ref, 3) grid-tied with 5A ref.](image)
It can also be observed in the current spectra of Fig. 13 that content around the 21st harmonic is observed in the third case, which was not present in the second case. This was actually determined to be background content that was pre-existing on the grid.

The impedances are constructed for the three different cases and compared with the Impedance Analyzer’s result (black). We can see that all three instances have very good matching at low frequencies, but the impedance identification with sinusoidal steady-state current (third case) seems to diverge at high frequencies. This error stems from the current spectrum hitting the noise floor before the voltage spectrum. In this instance, we can see that with a purely RL insertion impedance, very little current can pass at high frequencies. The first case, where no sinusoidal current reference is established, still achieves accurate results because no large fundamental is present, so little to no sensor attenuation is required.

This discussion also suggests an interesting mode of operation for improved measurement accuracy, where the inverter is grid tied, but uses a 0A current reference to temporarily reduce the fundamental signal so any injected signal will not be dominated by the fundamental current reference. After an injection/measurement cycle normal operation can then be restored.

B. Case of R-L-C Insertion Impedance

Next, an insertion impedance with resonant peak was chosen (peak chosen for ~3kHz) for a more “interesting” impedance (Fig. 14).

![R-L-C Insertion Impedance](image)

**Figure 14.** R-L-C insertion impedance with 3kHz resonant peak.

Fig. 15 illustrates two cases of interest. First, the spectra are measured when the converter was not active (red). In both the voltage (top) and current (bottom) spectra there is noticeable content. With the converter active (blue) we can see that the existing spectra actually dominate the converters injected signal. Without proper care, this existing content can corrupt results. As previously discussed either disregarding these frequencies, or taking into account this background spectra will provide accurate impedance estimation as observed (bottom).

![Spectra Comparison](image)

**Figure 15.** Injected spectra vs Existing Spectra for voltage (top), current (middle), and estimated impedance vs Impedance Analyzer measurement.

C. Case of No Insertion Impedance

Finally, no insertion impedance was used in an attempt to measure beyond the output step up and isolation transformers. Fig. 16 shows the results under three different operating conditions. The green trace is the inverter operating with no grid tie and the output of the isolation transformer shorted. The blue trace is the inverter grid tied with a 5% injection strength, and the red trace is with a 10% injection strength. Looking at the impedance estimations for all three cases, we can observe that they all match that of the Impedance Analyzer’s measurement of the impedance of the two transformers. From this we can conclude that the measured impedance in the grid tied case is still dominated by the impedance of the two transformers.

Clearly, since the step up transformer effectively reduces any impedance on the other side by a factor of 100 ($n^2$) it not feasible to measure grid impedance at the lower power levels currently in use.

![Impedance Comparison](image)

**Figure 16.** Voltage (top), Current (middle), and estimated impedance (bottom) with no insertion impedance for various operating conditions.
VII. CONCLUSION

Some interesting conclusions can be drawn from this work. First, addressed are some of the practical improvements that can be used to achieve better results and simultaneously decrease injected disturbances. Also discussed is the need for a baseline measurement of the existing spectral content so this information can be taken into account when performing a grid impedance measurement using the proposed method. Although beyond the scope of this work, knowledge of this existing content can also be used for monitoring or active filtering purposes.

Also very clear is the need for appropriate sensors. Accurate impedance estimation is not possible when low resolution ADCs are used unless very large disturbances can be tolerated. In future work all sampling and signal processing will be implemented into the DSP.

It is apparent that using a low power inverter with interfacing transformers cannot provide insight into actual grid impedance quantities. Such an arrangement simply cannot provide enough power to see beyond the transformer.

The low-power experimental setup allowed investigating and addressing several SNR-related issues and the methodology was proven with various controlled inserted impedances.

Future work will involve higher power levels and investigate direct grid ties as well as transformer isolated grid ties. Also, an extension to three phase impedance identification in both ABC and DQ0 coordinate systems and their uses will be explored.

REFERENCES


