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Comparative Analysis of the Effects of Integral Action and Disturbance Feedforward on Current Control of Voltage-Source Converters

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Keywords
Converter control, Discrete-time, State-space model, Voltage source converter (VSC).

Abstract
This paper presents a comparison between integral action and disturbance feedforward in current control of grid converters. The current controllers are designed directly in the discrete-time domain with the objective of similar reference-tracking and disturbance-rejection performance under nominal conditions. Partial compensation of time delays is included in the controller designs. The analytically and experimentally compared properties, in addition to reference tracking and disturbance rejection, are noise sensitivity and robustness to grid impedance variations. The controllers are found to have comparable dynamic performance, although the realized disturbance-rejection performance of the integrator-based controller is slightly better. The disturbance-feedforward-based controller is found to be less susceptible to current measurement noise at the cost of having an additional entry point for noise through the voltage measurement. The integrator-based controller is found more robust to grid impedance variations.

1 Introduction
The energy sector is in the middle of a revolution, where an increasing share of the electricity produced worldwide is being generated from renewable energy sources, such as solar panels and wind turbines. These renewable energy sources are typically interfaced to the electric grid through voltage-source converters (VSCs) [1]. A substantive effort has been put into developing current controllers for grid-interfaced converters, and a wide variety of different control designs have been proposed. Despite the breadth of different control structures, the majority of the control designs seem to revolve around the use of proportional (P) error term in combination with a subset of integral action (I), resonant action (R), and disturbance feedforward. Such designs find use in both current control applications [2, 3, 4] as well as in voltage controllers based on the cascade controller structure [5].

Typically, integral action is associated with the elimination of steady-state reference-tracking error due to modeling mismatches and plant uncertainty. On the other hand, disturbance feedforward is used to enhance disturbance rejection capability. Despite these two concepts having seemingly different purposes, in [2], they were shown to have a similar effect on the dynamic performance of current-controlled converters under nominal conditions. Furthermore, the disturbance-feedforward-based controller was shown to have better tolerance for time delays, thus being preferred over the integrator-based controller in view of passivity-oriented design [2].

On the other hand, several different approaches for partially compensating system time delays have been proposed, e.g., with a lead-lag compensator [3] or by means of state-feedback control [6]. Even though the design and analysis of controllers has been conventionally carried out in the continuous-time domain, controllers employing time-delay compensation are often designed directly in the discrete-time
domain. This is due to the ease of augmenting the delays originating from computation and pulse-width modulation (PWM) in the discrete-time model [7].

As the inclusion of time-delay compensation fundamentally alters the controller structure and might not restore the loss of passivity due to the time delays, it is natural to wonder whether the results in [2] generalize to delay-compensated systems. To extend the analysis of [2] to systems of the aforementioned type, this paper analyzes the relation of integral action and low-pass-filtered disturbance feedforward for systems wherein time delays have been taken into account in the direct discrete-time control design. The comparison is carried out using the state-space modeling framework, from which both control designs can be extracted from. The main contribution of this paper is a thorough comparative analysis based on reference-tracking and disturbance-rejection performance, as well as noise sensitivity. In addition, robustness to grid impedance variations is analyzed. The results are validated experimentally.

2 System Model

A three-phase current-controlled converter equipped with an $L$ filter is considered, cf. Fig. 1(a). The system is analyzed using space vectors in synchronous coordinates, e.g., the converter current is $i_c = i_{cd} + j i_{cq}$. Signals in stationary coordinates are denoted with a superscript $s$, e.g., the converter voltage is $u_c^s = u_{c,a}^s + j u_{c,b}^s$. The converter is connected to an electric grid modeled as a Thévenin equivalent circuit consisting of an impedance $Z_g^s(s)$ and a voltage source $e_g^s$. The dynamics of grid synchronization are omitted from the analysis and low synchronization bandwidth is used. The dc-link voltage $u_{dc}$ is assumed constant. An ideal grid is assumed in the controller design, i.e., $Z_g^s(s) = 0$. Consequently, the continuous-time plant model in stationary coordinates is given by

$$\frac{d}{dt} i_c^s = -\frac{R_f}{L_f} i_c^s + \frac{1}{L_f} (u_c^s - u_g^s)$$

where $L_f$ is the filter inductance, $R_f$ models the resistive losses in the inductor and the converter, and $u_g^s$ is the voltage at the point of common coupling (PCC). In the following, the resistive losses are omitted, i.e., $R_f = 0$. Assuming that both $u_c^s$ and $u_g^s$ remain constant over the sampling period $T_s$ and that the sampling is synchronized with the PWM, the hold-equivalent plant model in synchronous coordinates rotating at the grid angular frequency $\omega_g$ is obtained as

$$i_c(k + 1) = \delta i_c(k) + \gamma u_c(k) - \gamma u_g(k)$$

where $\delta = \exp(-j\omega_g T_s)$ and $\gamma = \delta T_s / L_f$. The hold-equivalent model includes a half-sampling-period delay originating from the PWM, i.e., a delay of $0.5T_s$ [7]. In addition, the finite computational resources of the digital processor executing the control algorithm give rise to a one-sample delay, i.e., $u_c(k+1) = u_{c,ref}(k)$, where $u_{c,ref}$ is the converter voltage reference. This delay, together with the PWM delay, brings the total delay to one and half sampling periods, i.e., $1.5T_s$. Additional, significantly shorter delays originating from, e.g., measurements, are omitted for simplicity. A block diagram representation of the current controller, from which the controllers under comparison can be obtained from, is shown in Fig. 1(b). As in [2], the disturbance voltage is filtered with a simple first-order low-pass filter (LPF). In accordance with the figure, the generalized control law for the system can be written as

$$u_{c,ref}(k) = k_5 i_{c,ref}(k) - k_1 i_c(k) - k_2 u_c(k) + k_3 x_1(k) + k_4 u_f(k)$$

where $i_{c,ref}$ is the current reference, $k_5$ is the reference feedforward gain, $k_1$ is the converter current feedback gain, $k_2$ is the converter voltage feedback gain for delay compensation, $k_3$ is the integral gain, and $k_4$ is the disturbance feedforward gain. In addition, $x_1$ and $u_f$ are the integral and filtered disturbance feedforward states

$$x_1(k + 1) = x_1(k) + \dot{i}_{c,ref}(k) - i_c(k)$$

$$u_f(k + 1) = e^{-\omega_n T_s} u_f(k) + (1 - e^{-\omega_n T_s}) u_g(k)$$
respectively. In (5), $\omega_{lf}$ denotes the desired LPF bandwidth. Combining the hold-equivalent plant model (2), the computational delay, and both the integral and filtered disturbance feedforward states into a state-space model, one obtains

$$
\begin{bmatrix}
  i_{c}(k+1) \\
  u_{c}(k+1) \\
  x_{1}(k+1) \\
  u_{1}(k+1)
\end{bmatrix}
= \Phi \begin{bmatrix}
  x(k) \\
  u_{c,ref}(k) \\
  u_{g}(k) \\
  i_{c,ref}(k)
\end{bmatrix}
+ \begin{bmatrix}
  0 \\
  1 \\
  0 \\
  0
\end{bmatrix}
\begin{bmatrix}
  0 \\
  1 \\
  0 \\
  0
\end{bmatrix}
\begin{bmatrix}
  1 \\
  0 \\
  1 - e^{-\omega_{lf}T_{s}} \\
  e^{-\omega_{lf}T_{s}}
\end{bmatrix}
\begin{bmatrix}
  0 \\
  0 \\
  0 \\
  1
\end{bmatrix}
$$

(6)

Placing the control law (3) into the system model (6) yields the closed-loop model of the system

$$
x(k + 1) = (\Phi - \Gamma_{c} K_{i}) x(k) + \Gamma_{c} u_{g}(k) + (\Gamma_{c} K_{i} + \Gamma_{r}) i_{c,ref}(k)
$$

(7)

where $K_{i} = [k_{1}, k_{2}, -k_{1}, -k_{i}]$. The state-feedback gains $k_{1}$, $k_{2}$, and $k_{i}$ can be solved, e.g., analytically from the state-space model (6) or numerically using direct pole placement for which tools are readily available, e.g., in MATLAB [7]. The choice of feedforward gains $k_{1}$ and $k_{i}$ depends on the employed controller. From the closed-loop model (7), the general reference-tracking transfer function and the output admittance, containing both the integrator and the disturbance feedforward, can be obtained as

$$
G_{cl}(z) = \frac{i_{c}(z)}{i_{c,ref}(z)} = \frac{\gamma_{c} k_{1}(z - 1 + k_{1}/k_{1})}{(z - \delta)(z + k_{2})(z - 1) + \gamma_{1}(z - 1) + \gamma_{k_{1}}}
$$

(8)

$$
Y_{cl}(z) = \frac{i_{c}(z)}{u_{g}(z)} = \frac{\gamma(z - 1)[(z - \omega_{lf}T_{s})(z + k_{2}) - (1 - \omega_{lf}T_{s}) k_{1}]}{(z - \omega_{lf}T_{s})(z - \delta)(z + k_{2})(z - 1) + \gamma_{k_{1}}(z - 1) + \gamma_{k_{1}}}
$$

(9)

respectively. This general closed-loop system has a fourth-order characteristic polynomial. One of the poles is defined by the LPF bandwidth $\omega_{lf}$ and the three other poles can be set freely with an appropriate choice of the feedback gains $k_{1}$, $k_{2}$, and $k_{i}$. The pole originating from the LPF is canceled in (8), as the disturbance feedforward path does not affect reference-tracking properties.

Moreover, the sensitivity to current and voltage measurement noise is of interest. Noise sensitivity can be analyzed by including additive perturbation signals to the current and voltage measurements, as shown in gray in Fig. 1(a). The addition of current measurement noise to the system model corresponds to adding a noise component $i_{c,n}$ to the actual current $i_{c}$ in the control law (3) and the integrator (4). Similarly, voltage measurement noise can be included in the model by adding a noise component $u_{g,n}$ to the actual voltage $u_{g}$ in the LPF (5). The effect of current and voltage measurement noises are then characterized through their effect on the controller output $u_{c,ref}$. The transfer functions from $i_{c,n}$ and $u_{g,n}$ to $u_{c,ref}$.
defined as the current and voltage noise sensitivity functions, respectively, are obtained as

\[
Z_{in}(z) = \frac{u_{s,ref}(z)}{i_{cl}(z)} = \frac{-k_1(z-\delta)[z-1+k_1/k_1]}{(z-\delta)(z+k_2)(z-1)+\gamma k_1(z-1)+\gamma k_1} \tag{10}
\]

\[
G_{un}(z) = \frac{u_{s,ref}(z)}{ug(z)} = \frac{(1-e^{-\omega_T z})k_1(z-\delta)(z-1)+\gamma k_1(z-1)+\gamma k_1}{(z-\delta)(z+k_2)(z-1)+\gamma k_1(z-1)+\gamma k_1} \tag{11}
\]

respectively.

3 Comparative Analysis

The generalized control law (3) contains both the integrator and the disturbance feedforward together with the time-delay compensation. In this section, the integrator-based \((k_f = 0)\) and disturbance-feedforward-based \((k_f = 0)\) controllers are analyzed with regard to their dynamic performance, noise sensitivity, and robustness to grid impedance variations. The basis of comparison is similar design of reference-tracking and disturbance-rejection performance.

3.1 Integrator-Based Controller

The integrator-based controller is obtained from (8)-(11) by omitting the disturbance feedforward path, i.e., by setting the feedforward gain \(k_f = 0\). Furthermore, the feedback gains \(k_1, k_2,\) and \(k_f\) are selected to yield roots \(p_1, p_2,\) and \(p_3\) for the characteristic polynomial. The reference feedforward through the gain \(k_f\) creates a zero in the reference-tracking transfer function, and it can be placed at \(p_3\) by choosing the gain as \(k_f = k_f/(1-p_3)\). As a result, the closed-loop transfer functions are obtained as

\[
C_{cl}^{int}(z) = \frac{\gamma k_1(z-1+k_1/k_1)}{(z-\delta)(z+k_2)(z-1)+\gamma k_1(z-1)+\gamma k_1} = \frac{(1-p_1)(1-p_2)}{(z-p_1)(z-p_2)} \tag{12}
\]

\[
Y_{cl}^{int}(z) = \frac{\gamma(z-\delta)(z+k_2)}{(z-\delta)(z+k_2)(z-1)+\gamma k_1(z-1)+\gamma k_1} = \frac{z(z-\delta)(az+b)}{\gamma(z-p_1)(z-p_2)(z-p_3)} \tag{13}
\]

\[
Z_{in}^{int}(z) = \frac{-k_1z(z-\delta)[z-1+k_1/k_1]}{(z-\delta)(z+k_2)(z-1)+\gamma k_1(z-1)+\gamma k_1} = \frac{z(z-\delta)}{\gamma(z-p_1)(z-p_2)(z-p_3)} \tag{14}
\]

\[
C_{un}^{int}(z) = 0 \tag{15}
\]

where the coefficients \(a\) and \(b\) in (14) are functions of the desired characteristic polynomial roots and the system parameters\(^1\).

3.2 Disturbance-Feedforward-Based Controller

The disturbance-feedforward-based controller is obtained from (8)-(11) by omitting the integrator, i.e., by setting the integral gain \(k_i = 0\). Furthermore, the remaining feedback gains \(k_1\) and \(k_2\) are selected to yield roots \(p_1\) and \(p_2\) for the characteristic polynomial. The LPF pole \(\exp(-\omega_T T_s)\) is set at \(p_3\), i.e., the third root of the characteristic polynomial for the integrator-based controller. Lastly, the reference feedforward gain \(k_f\) is selected to yield unity gain for the reference-tracking transfer function at the fundamental frequency, which corresponds to the dc component in synchronous coordinates. As a result, one obtains the reference-tracking transfer function and the current noise sensitivity function as

\[
C_{cl}^{ff}(z) = \frac{\gamma k_1}{(z-\delta)(z+k_2)+\gamma k_1} = \frac{(1-p_1)(1-p_2)}{(z-p_1)(z-p_2)} \tag{16}
\]

\[
Z_{in}^{ff}(z) = \frac{-k_1z(z-\delta)}{(z-\delta)(z+k_2)+\gamma k_1} = \frac{\delta(p_1+p_2-p_1p_2-\delta^2)z(z-\delta)}{\gamma(z-p_1)(z-p_2)}. \tag{17}
\]

\(^1\)The coefficients can be written as \(a = (\delta+1)(p_1+p_2+p_3-\delta-1)-p_1p_2-p_1p_3-p_2p_3+\delta\) and \(b = p_1p_2p_3-\delta(p_1+p_2+p_3-\delta-1)\).
Based on the abovementioned parameterization, the output admittance is obtained as

\[
Y_{cl}^{ff}(z) = \frac{\gamma[(z - e^{-\omega_f T_s})(z + k_2) - (1 - e^{-\omega_f T_s})k_f]}{(z - e^{-\omega_f T_s})[(z - \delta)(z + k_2) + \gamma k_f]} = \frac{\gamma[(z - p_3)(z - p_1 - p_2 + \delta) - (1 - p_3)k_f]}{(z - p_1)(z - p_2)(z - p_3)}. \tag{18}
\]

The disturbance feedforward gain \(k_f\) enables the designer to freely select the location of one of the two zeros in (18). Typically, a zero \(z = 1\) is desired in the output admittance, as it results in perfect rejection of the dc component in synchronous coordinates, which corresponds to the fundamental frequency component in stationary coordinates. Following this design criterion, the disturbance feedforward gain is obtained as \(k_f = 1 - p_1 - p_2 + \delta\). As a consequence of selecting this zero location and setting the LPF pole at \(p_3\), the second zero in the numerator of (18) maps itself to \(z = p_1 + p_2 + p_3 - \delta - 1\), i.e., the location of the second zero in (13), and the output admittance becomes

\[
Y_{cl}^{ff}(z) = \frac{\gamma(z - 1)(z - p_1 - p_2 - p_3 + \delta + 1)}{(z - p_1)(z - p_2)(z - p_3)}. \tag{19}
\]

Furthermore, this choice of \(k_f\) results in the voltage noise sensitivity function

\[
C_{un}^{ff}(z) = \frac{(1 - e^{-\omega_f T_s})k_f z(z - \delta)}{(z - e^{-\omega_f T_s})[(z - \delta)(z + k_2) + \gamma k_1]} = \frac{(1 - p_3)(1 - p_1 - p_2 + \delta)z(z - \delta)}{(z - p_1)(z - p_2)(z - p_3)}. \tag{20}
\]

### 3.3 Comparison of the Resulting Systems

In the following, including the experiments, a 12.5-kVA three-phase converter system with nominal voltage of \(\sqrt{2}/3 \cdot 400\) V (1 p.u.) and current of \(\sqrt{2} \cdot 18\) A (1 p.u.) is considered. Fundamental angular frequency of the system is \(\omega_b = 2\pi \cdot 50\) rad/s (1 p.u.). The filter inductance is \(L_f = 5\) mH (0.125 p.u.) unless otherwise stated and the sampling period is \(T_s = 125\) μs. As an example design, the characteristic polynomials of the two systems are designed to have roots\(^2\) \(p_1 = 0\) and \(p_{2,3} = \exp(-\alpha_c T_s)\), where \(\alpha_c = 2\pi \cdot 400\) rad/s (8 p.u.).

#### 3.3.1 Reference Tracking

Under nominal conditions, the reference-tracking transfer functions (12) and (16) can be observed to be identical, which is also graphically depicted in the left column of Fig. 2. This is a natural consequence of the reference feedforward zero canceling the closed-loop pole originating from the integrator in the integrator-based controller, and due to the disturbance feedforward not affecting the reference-tracking dynamics in the disturbance-feedforward-based controller. On the other hand, as already well-established in the literature [2], the steady-state reference-tracking error of the disturbance-feedforward-based controller is nonzero under modeling errors. This reference-tracking error can be reduced by employing nonzero filter resistance \(R_f\) in the controller design stage. However, the tuning of this resistance value has to be often carried out using a trial-and-error method, since the modeling errors vary from system to system. On the contrary, the integrator-based controller achieves error-free reference tracking in steady state irrespective of the modeling errors. An additional challenge with the disturbance-feedforward-based controller is related to the digital LPF implementation. Since the LPF is digital, PWM ripple is not filtered out from the voltage measurement, which causes aliasing in the voltage measurement, e.g., with inductive-resistive grid impedances, resulting in a bias in the current. However, this challenge can be overcome, e.g., by using oversampling.

#### 3.3.2 Disturbance Rejection

While the output admittances (13) and (18) have identical poles and the same number of zeros, they are not identical by default, i.e., if a unity feedforward gain \(k_f = 1\) is used. With the choice of disturbance

\(^2\)The corresponding gains are obtained as \(K_f = [24.44 - 1.46, 0.54 - 0.039j, -2.91 - 0.11], 0\) and \(k_\alpha = 10.78 + 0.42\) for the integrator-based controller and as \(K_f = [10.75 - 1.57, 0.27 - 0.039j, 0, -1.27 + 0.039]\) and \(k_\alpha = 10.78 + 0.42\) for the disturbance-feedforward-based controller.
Fig. 2: Closed-loop reference-tracking (left) and discrete-time output admittance (right) frequency responses for the two controllers.

Fig. 3: Intersample output admittances of the integrator-based (left) and disturbance-feedforward-based (right) controllers. In addition, the intersample model is validated by means of simulations where a frequency-sweep utilizing single-frequency sinusoids is employed. The discrete-time output admittance is shown in grey for comparison. The range of frequencies extends up to the Nyquist frequency $f_s/2 = 4$ kHz.

Fig. 4: Zoomed-in view of the nonpassive region, i.e., the region of output admittance with negative real part, in the intersample output admittances of the integrator-based (left) and disturbance-feedforward-based (right) controllers for different locations of the third pole, which is defined as $p_3 = \exp(-\beta s T_s)$.

feedforward gain $k_f = 1 - p_1 - p_2 + \delta$ according to the above section, the resulting output admittance (19) equals (13), which is also graphically depicted in the right column of Fig. 2. However, these frequency responses reflect the system behavior in the discrete-time domain and under certain modeling assumptions, e.g., lossless system ($R_f = 0$) and zero-order hold (ZOH) of $u^s_g$. Consequently, the true output admittance, as observed from the output terminals of the converter system, differs from those predicted by (13) and (19). For obtaining a more detailed model of the output admittances of the two controllers, the intersample model is employed [8]. The intersample model provides a framework for modeling systems with both discrete- and continuous-time signals, and allows taking into account the characteristics of sampling in the current and voltage measurements. The intersample output admittances of the two controllers are shown in Fig. 3. The figures also include the discrete-time output admittance for comparison and the simulated output admittance obtained using a frequency sweep with single-frequency sinusoids. The simulation model is based on the system in Fig. 1(a). As can be observed from Fig. 3, the disturbance-feedforward-based controller is slightly more susceptible to grid-voltage disturbances in the low-frequency range, up to 1 kHz, while having comparable output admittance with the integrator-based controller for higher frequencies.
Fig. 5: Frequency responses of the current (left) and voltage (right) noise sensitivity functions for the two controllers.

By closely examining the phase behavior of the output admittances in Fig. 3, it can be observed that only very minor excursions outside of the range \([-\pi/2, \pi/2]\), i.e., the range of passive behavior where the real part of the output admittance is nonnegative [2], occur in the phase angle. Since the third pole \(p_3\) does not affect reference-tracking properties under nominal conditions, cf. (12) and (16), it can be used to shape the output admittance, and consequently, the passivity properties of the system. Fig. 4 explores the effect of varying the location of the third pole \(p_3 = \exp(-\beta_c T_s)\) on the passivity of the two systems. For the example design, the nonpassive region of the integrator-based controller is wider but shallower than the nonpassive region of the disturbance-feedforward-based controller. However, the controllers react similarly to the variations in the third pole, i.e., the width of the nonpassive region increases and the region becomes shallower as the value of \(\beta_c\) decreases.

Remark 1: If \(k_2 \neq 0\) and if the approach in [2] for the disturbance feedforward is applied, i.e., a unity gain \(k_f = 1\) is employed, neither of the zeros in (18) map to \(z = 1\), resulting in susceptibility to fundamental frequency disturbances. Hence, proper parameterization of the disturbance feedforward gain \(k_f\) is of paramount importance.

Remark 2: The discrete-time counterparts of the cases analyzed in [2], including the delays, can be obtained from (8) and (9) by setting \(k_1 = k_p + R_a - j\omega g L_f\), \(k_2 = 0\), \(k_l = k_p\), and using either \(k_i = k_i\) and \(k_f = 0\) or \(k_i = 0\) and \(k_f = 1\) to obtain the integrator-based or disturbance-feedforward-based controllers, respectively. The gains \(k_p\) and \(k_i\) refer to the proportional and integral gains of the PI controller and \(R_a\) to the active resistance used in [2].

3.3.3 Noise Sensitivity

The current and voltage noise sensitivity functions of the two controllers, on the other hand, are vastly different. Since the integrator-based controller does not utilize the PCC voltage measurement, the voltage noise sensitivity function (15) is naturally zero, whereas the PCC voltage measurement in the disturbance-feedforward-based controller creates an additional source of noise to the system, cf. (20). The current noise sensitivity function (14) of the integrator-based controller has an additional zero and pole in comparison to the corresponding sensitivity function (17) of the disturbance-feedforward-based controller. The noise sensitivity functions of the two controllers are shown in Fig. 5. The disturbance-feedforward-based controller can be observed to be less sensitive to current measurement noise at frequencies above 100 Hz. The greater sensitivity of the integrator-based controller is largely due to the considerably larger magnitude of its current feedback gain \(k_1\) as compared to the disturbance-feedforward-based controller. On the other hand, the additional noise source from the voltage measurements is only present in the disturbance-feedforward-based controller.

3.3.4 Robustness to Grid Impedance Variations

Initially, an inductive-resistive (LR) grid is assumed. In fact, as \(Y_{cl}^{int} = Y_{cl}^{ff}\) [cf. (13) and (19)], the robustness of the controllers to variations in LR-type grid impedance is identical in the discrete-time domain. This can be realized by modeling the converter as a Norton equivalent circuit, cf. Fig. 6(a), and the grid as a Thevenin circuit with series LR impedance \(Z_g\), cf. Fig. 6(b), in which case the stability of the
converter–grid interconnection can be analyzed through the product of the converter output admittance and the grid impedance, i.e., $Y_{cl}(z)Z_g(z)$ [9]. Fig. 7 shows the robustness map for LR grid impedance in terms of the lowest damping ratio $\zeta$ in the characteristic polynomial $1 + Y_{cl}(z)Z_g(z)$. The damping ratio $\zeta$ characterizes the decay of oscillations excited by perturbations in the system, i.e., the lower the damping ratio, the slower the decay of oscillations is. As can be observed from the figure, the controllers operate stably in the whole range of examined impedances. However, this shared behavior of the two controllers does not generalize to all grid impedances. Assuming that the grid impedance consists of a parallel capacitance and a series inductance, cf. Fig. 6(c), so that the $L$ filter and the grid impedance form an $LCL$ filter, the robustness maps of the two controllers are shown in Fig. 8 where $L_f = 2.8$ mH. The integrator-based controller can be observed to have wider range of stable operation with higher damping ratios overall.

### 4 Experimental Results

In the experimental setup, the 12.5-kVA converter under test is controlled using a dSPACE DS1006 processor board. The switching frequency of the converter is 4 kHz. The converter control system is synchronized to the PCC voltage using a phase-locked loop with 2 Hz bandwidth. The dc-link voltage ($u_{dc} = 650$ V) of the converter is maintained constant by another converter, and the converter under test is connected to a 50-kVA grid emulator (Regatron TopCon TC.ACS).
Fig. 9 shows the reference-tracking and disturbance-rejection performance of the two controllers under nominal conditions, i.e., $Z_s^g = 0$. In the reference-tracking case, a stepwise change from 0 p.u. to 0.2 p.u. is applied to the $d$-axis reference. Disturbance-rejection performance is demonstrated by applying a grid-voltage dip from 1 p.u. to 0.5 p.u. A nonzero resistance estimate of 0.10 p.u. is applied in the disturbance-feedforward-based controller tuning to improve reference-tracking accuracy. As can be observed, comparable dynamic performance is obtained with the two controllers. However, in agreement with the results from the intersample admittance models presented in Fig. 3, the response of the disturbance-feedforward-based controller to grid-voltage disturbance is slightly slower. Next, the reference-tracking and disturbance-rejection performance is examined when $L_f = 2.8$ mH with a parallel capacitor $C = 15$ µF (0.06 p.u.) and a series inductor $L = 6$ mH (0.15 p.u.) in the grid [cf. Fig. 6(c)]. The results are shown in Fig. 10. Similarly to the experiments under nominal conditions, comparable performance can be observed, although the disturbance-feedforward-based controller has slightly slower response to the grid-voltage dip. Lastly, the measurement noise sensitivity of the controllers is examined under nominal conditions. Fig. 11 shows the results of this experiment, where 100, 400, and 2000 Hz sinusoids with amplitude of 0.05 p.u. are injected to the current and voltage measurements at $t = 10, t = 30$, and $t = 40$ ms, respectively. The disturbance-feedforward-based controller is less susceptible to current measurement noise, but it has an additional entry point for the noise through the voltage measurement (cf. Fig. 5).
Fig. 11: Experimental results for current (left) and voltage (right) measurement noise sensitivity under nominal conditions. 100, 400, and 2000 Hz sinusoids with amplitudes of 0.05 p.u. are sequentially injected to the measurements, and the figures show the converter voltage reference $u_{c,ref} = u_{cd,ref} + j u_{cq,ref}$ responses.

5 Conclusion

This paper presented a comparative analysis of integral action and low-pass-filtered disturbance feedforward in current control of grid converters, wherein direct discrete-time design with partial time-delay compensation is employed. The basis of comparison is similar design of reference-tracking and disturbance-rejection performance. The two controllers are found to have comparable dynamic performance, although the realized disturbance-rejection performance of the integrator-based controller is slightly better. Furthermore, the disturbance-feedforward-based controller requires additional trial-and-error tuning of the resistance estimate to reduce steady-state reference-tracking error caused by modeling mismatches. The use of intersample output admittance model reveals that the passivity characteristics of the two controllers are different. The two controllers have comparable robustness to inductive-resistive grid impedance variations. However, the integrator-based controller is more robust to grid impedances consisting of a parallel capacitance and series inductance, i.e., different LCL filters. The disturbance-feedforward-based controller is found less sensitive to current measurement noise at the cost of having an additional entry point for noise through the voltage measurement.

References