Nonlinear Effect of Deadtime in Small-Signal Modeling of Power-Electronics System Under Low Load Conditions

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

Abstract—Deadtime is required to ensure that switches of a synchronous switching inverter leg never conduct at the same time. During deadtime, the current commutates to an anti-parallel diode that can cause a voltage error depending on the instantaneous current direction. To measure a frequency response from a system, external injections are commonly required to perturb the system. The perturbation can change the current direction at the frequency of the injection causing a voltage error at injection frequency due to the deadtime. The error depends on the perturbation amplitude, inductor current ripple and the fundamental current amplitude. This paper proposes a describing-function method to model the deadtime effect under low load conditions. It is shown that a nonlinear damping effect from the deadtime can occur under low load conditions and cannot be modeled with a resistor-like element. Real-time hardware-in-the-loop-simulation results are presented and used to demonstrate the effectiveness of the proposed method. Experimental measurements are used to verify the nonlinear deadtime effect.

Index Terms—frequency response, deadtime, describing function, nonlinearity

I. INTRODUCTION

Frequency-response analysis is among the most widely used techniques in dynamic analysis and controller tuning of power electronic systems. The basis of deriving the dynamic models lies in the linearity of the inspected system or linearizing the system around an operation point that yields linear frequency response. Impedance-based stability is an application that is based on measured or modeled frequency responses and has received a great deal of attention in the last few years [1]–[3].

A possible source of nonlinearity is the deadtime that is required to prevent shoot-through faults in synchronous switching power converters. Deadtime causes a voltage error that is dependent on the inductor current sign [4]. The inductor current sign can change due to an external disturbance that can be caused, for example, by a frequency-response measurement.

A typical frequency-response-measurement method of a power-electronics system is based on an external voltage/current injection that is placed, for example, on top of the nominal input voltage or current or control signal. Fourier methods are then applied to extract the spectral information between the desired input and output variables [5]. In a power-electronics system with the deadtime, a measurement injection can change the current sign at the injected frequency and cause a voltage error that acts as a damping.

The deadtime effect regarding the small-signal dynamics has gained some attention in the literature. Several studies have modeled the deadtime effect with a resistor-like element [6]–[10]. Three-phase converters were analyzed in [6] and [7]; and single-phase converters in [8] and [9]. In [10], the deadtime effect on dynamics of a synchronous switching buck converter was analyzed, focusing on the charging of the drain-source capacitance during the deadtime. The deadtime effect on dynamics of a quasi-square-wave flyback converter was analyzed in [11].

The importance of the ripple effect is shown in [8], and a describing-function-based method is used to solve a resistor-like element to model the deadtime effect. The model predicts damping to occur when the fundamental current amplitude is higher than half of the peak-to-peak current ripple. Furthermore, it is assumed that the inductor current amplitude is same as the injection current. This assumption in [8] is well-grounded because the output filter of the analyzed full-bridge inverter consists only of an inductor. However, if there was a resonance that increases the inductor current with respect to injection, this assumption would not be valid.

The present paper studies the deadtime effect under a low load condition where the inductor current fundamental component amplitude is lower than half the peak-to-peak ripple. In [8], the opposite assumption and operating point was used. Furthermore, in the present paper, the response in the inductor current that causes the deadtime effect is not considered small. Instead, we propose a describing-function-based method that models the output impedance as a function of injection amplitude and frequency. We show that resonance in the system can increase the inductor current so that a voltage error occurs or even saturates. Therefore, a moderate injection amplitude does not guarantee a linear operating region.

The remainder of this paper is structured as follows. Section II examines the deadtime effect and the amplitude dependency in frequency-response measurements. Section III derives a describing function model for the error. The proposed amplitude-dependent output impedance is introduced in Section IV. In Section V, hardware-in-the-loop (HIL) simulations and experimental measurements are used to verify the analyzed deadtime effect. Conclusions are drawn in the final section.

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II. DEADTIME EFFECT

This section begins by inspecting the voltage error that is caused by the deadtime. We then analyze and model how the deadtime effect is visible in frequency-response measurements and how the operating point and system parameters affect the deadtime effect. Throughout this paper, it is assumed that the frequency-response measurement is performed with a sine-sweep, where one frequency at a time is injected and measured.

A. Voltage error caused by deadtime

Fig. 1 shows a half-bridge inverter that is used to analyze the deadtime effect. The switches and diodes are analyzed with ideal components. However, in reality the required deadtime depends on the turn-off characteristics of the used semiconductor switches [12]. When the deadtime length is chosen, the changes in the switch characteristics with load current and temperature must be taken into account [13]. The deadtime is not changed according to operating conditions. In general, faster switching devices require shorter deadtime than lower switching devices. The operating point and component values are given in Table I. A deadtime value of 4 µs is applied in most parts of this work. This value (in a combination of the applied switching frequency of 10 kHz in the laboratory setup) was observed to efficiently demonstrate the nonlinear effect.

Table I: Operating point and component values.

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<th>Parameter</th>
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<td>$V_{DC}$</td>
<td>700 V</td>
<td>$I_o$</td>
<td>0-1.8 A</td>
</tr>
<tr>
<td>$L$</td>
<td>4 mH</td>
<td>$r_L$</td>
<td>10 mΩ</td>
</tr>
<tr>
<td>$C$</td>
<td>10 µF</td>
<td>$r_C$</td>
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</tr>
<tr>
<td>$\omega_s$</td>
<td>2π/60 rad/s</td>
<td>$V_{o,\text{rms}}$</td>
<td>120 V</td>
</tr>
<tr>
<td>$f_s$</td>
<td>10 kHz</td>
<td>$T_{\text{dead}}$</td>
<td>4 µs</td>
</tr>
</tbody>
</table>

The deadtime is used to delay the turn-on of switches $S_1$ and $S_2$, ensuring a period during which neither of the switches conducts. During deadtime, the current commutates to either of diodes $D_1$ or $D_2$ depending on the instantaneous current sign [4]. Therefore, the deadtime causes an instantaneous voltage error, $v_{\text{err}}$, in inverter voltage during the deadtime that can be given by

$$v_{\text{err}} = \text{sign}(i_L)V_{DC}$$  \hspace{1cm} (1)

where $V_{DC}$ and $i_L$ are the DC voltage and the inductor current, respectively. In order to facilitate the analysis, the error is averaged over a switching cycle:

$$v_{\text{err}}^{\text{avg}-Ts} = \frac{1}{T_s} \int_0^{T_s} v_{\text{err}}(t) \, dt$$  \hspace{1cm} (2)

where $T_s$ and $T_{\text{dead}}$ are the switching cycle and the deadtime lengths, respectively. The resulting maximum average voltage error is

$$V_{\text{max}}^{\text{err}} = \left| v_{\text{err}}^{\text{avg}-Ts} \right| = \left| \text{sign}(i_L) \frac{T_{\text{dead}}}{T_s} V_{DC} \right|$$  \hspace{1cm} (3)

which yields 28 V with the parameters of Table I. The fundamental component of a square wave with the amplitude of 28 V is $4/\pi \times 28 \text{ V} = 35.56 \text{ V}$. The value of the average voltage error has been analyzed in the literature earlier [4], [12].

Fig. 2 (a) shows the averaged voltage error over a 60 Hz fundamental cycle. The error has the same sign as the current. However, during the period of current zero crossings, the average voltage error is zero. In Fig. 2 (b), a perturbation is added to the inductor current, and a voltage error appears at the perturbation frequency due to the changes in the current zero crossings. The zero crossings period was modeled in [14], and a similar approach was used for modeling the deadtime effect in [9]. However, the present paper focuses on modeling the deadtime effect under a low load condition, where the 60 Hz fundamental current component is lower than half the peak-to-peak inductor current ripple.

Fig. 3 (a) and (b) show the inductor current and the averaged voltage error under low load conditions. In Fig. 3 (a), there is no fundamental 60 Hz component in the inductor current and in Fig. 3 (b) the fundamental component is lower than half the peak-to-peak current ripple. The average voltage error during the whole fundamental cycle is zero because the inductor current crosses zero during every switching cycle, and the instantaneous error has both signs during a switching cycle. Fig. 3 (c) illustrates a case where a perturbation is added to the current fundamental component, and the sum has a higher amplitude than half the peak-to-peak current ripple. Therefore,
Figure 3: Inductor current waveforms (red) and averaged voltage errors (blue) with different fundamental current amplitudes: (a) 0 A, (b) 1 A, and (c) 1 A with a perturbation of 1.5 A at 600 Hz.

Figure 4: Output impedance from a simulation without the deadtime with different injection amplitudes.

A voltage error appears when there are no zero crossing within a switching cycle or a zero current clamping occurs.

B. Perturbation amplitude dependency

The half-bridge inverter in Fig. 1 is simulated with Matlab/Simulink. A current injection is made with the output current, \( i_o \), to measure the output impedance, \( Z_o = \bar{v}_o / \bar{i}_o \).

The fundamental load current, a 60 Hz component, \( A_{\text{fund}} \), is provided by the current sink if required. In the following, the current sink draws no 60 Hz current component, and the current sink is used only for the injection. Fig. 4 shows the simulated measurement of the output impedance when no deadtime is applied in the switching as benchmark. The simulation is done two times with the injection amplitudes of 0.5 A and 2 A.

Figure 5: Output impedance, \( Z_o \), measured from a simulation with the deadtime and different injection amplitudes.

Figure 6: Output current-to-inductor current transfer function, \( G_{oL} \), measured from a simulation with different injection amplitudes.

A dynamic model of the system is created based on the parallel connection of the filter inductor impedance, \( Z_L \), and filter capacitor impedance, \( Z_C \):

\[
Z_C = r_C + \frac{1}{sC} \tag{4}
\]

\[
Z_L = r_L + sL \tag{5}
\]

\[
Z_o(s) = \frac{Z_C Z_L}{Z_C + Z_L} = \frac{r_C s^2 + \left( \frac{r_L r_C}{L} + \frac{1}{C} \right) s + \frac{r_L r_C}{C L}}{s^2 + \left( \frac{r_L + r_C}{L} \right) s + \frac{1}{C L}} \tag{6}
\]

where \( L, C, r_L \) and \( r_C \) denote the inductance, capacitance, inductor resistance and capacitor resistance, respectively. As
Fig. 4 shows, the model corresponds to the simulations according to the linear circuit theory.

Next, the output impedance simulation is repeated when a deadtime of 4 µs is applied in the switch control signals. Fig. 5 compares the results to the linear model, \( Z_o(s) \). It is apparent from Fig. 5 that the deadtime makes the system highly nonlinear, and the linear model cannot be used to model the system. With the injection amplitude of 0.5 A and 1 A, the resonance is damped. On the other hand, with the injection amplitudes of 2 A and 3 A, a damping appears at lower frequencies and the resonance is less damped. Therefore, a resistor-like element that is constant at all frequencies cannot be used to model the deadtime effect.

It is known that the error caused by the deadtime is dependent on the inductor current. Therefore, it is useful to look at the output current-to-inductor current dynamics. The transfer function can be given by

\[
G_{oL}(s) = \frac{Z_C}{Z_C + Z_L} = \frac{r_C s + \frac{1}{C L}}{s^2 + \frac{r_L + r_C}{L} s + \frac{1}{C L}} \tag{7}
\]

Fig. 6 compares the modeled and simulated \( G_{oL}(s) \). Similar to the case of the output impedance, the deadtime effect is visible and the linear model is not valid. Fig. 7 shows the absolute values of the inductor current and the voltage error with different injection amplitudes. It can be seen that the resonance amplifies the inductor current, which increases the voltage error. However, the voltage error has its maximum value, \( V_{err_{\text{max}}} \), according to (3). When the error saturates to its maximum value, its effect on the frequency response begins to diminish. It can be also noted that there is no voltage error at all if the inductor current amplitude is not high enough.

Fig. 8 illustrates how the error behaves as a function of the inductor current perturbation amplitude. For the simulations in Fig. 8, the filter capacitor was changed from 10 µF to 0.5 µF in order to reduce the reactive current fundamental component. All injections were made at 650 Hz. With zero fundamental current amplitude, there is a deadzone below which there is no voltage error. When the fundamental current amplitude is increased, the deadzone shortens until the fundamental amplitude is approximately half the peak-to-peak current ripple, \( \Delta i/2 \):

\[
\Delta i = \frac{V_{DC} T_s}{4L}. \tag{8}
\]

Half the peak-to-peak ripple, \( \Delta i/2 \), is 2.19 A with the parameters of Table I.

Because the gain of \( G_{oL}(s) \) is close to 0 dB at low frequencies, the injection amplitude must be higher than half the peak-to-peak ripple for the error to occur. This can be seen in Fig. 7. When the perturbation amplitude is high enough, the error saturates to the fundamental component of the square wave voltage error. The saturation begins when the perturbation amplitude equals the sum of half the peak-to-peak current ripple and the fundamental current. After this point, the increase in the perturbation amplitude affects the whole fundamental cycle; this can be seen from Figs. 7 and 8. Between the deadzone and the saturation is a slope region that can be seen well from Fig. 8 in the case where the fundamental current amplitude is 1 A. This behavior of the voltage error under low load conditions was illustrated in [9] but not analyzed.

Fig. 8 also shows the voltage error when the deadtime is 1 µs. The error is scaled by multiplying it by 4 in order to have the maximum error, \( V_{err_{\text{max}}} \), unchanged compared to the case with the deadtime of 4 µs according to (3). By inspecting the error as a function of the current with \( A_{\text{fund}} = 0 \), Fig. 8 shows that there is a voltage error on lower inductor current perturbations than \( \Delta i/2 \). This is partly due to the approximation of \( \Delta i/2 \) in (8) that gives the ripple when the duty cycle is 0.5 (that is, around the fundamental component zero crossing).

Furthermore, Fig. 8 reveals that the deadtime length not only affects the maximum voltage error, \( V_{err_{\text{max}}} \). With \( T_{\text{dead}} = 1 \mu s \), the error remains zero with higher inductor current response amplitudes. This can be explained with a zero current clamping effect, which causes a voltage error. For a zero current...
clamping to occur, it is sufficient for the current to drop to zero during the deadtime, and remain zero for the rest of the deadtime length. The maximum current change during the deadtime is approximated by

$$i_{\text{clamp}}^{\text{max}} = \frac{V_{\text{DC}}T_{\text{dead}}}{2L}.$$  (9)

If the zero is crossed for $i_{\text{clamp}}^{\text{max}}$ or fewer Amperes before the current slope direction changes, a zero current clamping occurs. The zero current clamping effect is the first source of voltage error rather than the main deadtime effect. The zero current clamping effect and the voltage error have been analyzed in [15].

III. DESCRIBING-FUNCTION MODEL

A describing-function model for the voltage error can be developed based on the observations from the voltage error as a function of the inductor current perturbation amplitude, current ripple, and the current fundamental component in Fig. 8. Therefore a describing-function model consisting of a deadzone, a slope, and a saturation illustrated in Fig. 9(a) must be created.

With lower current perturbation amplitudes, $A$, than $R_1$ there is no voltage error and with higher than $R_2$ the error saturates. The zero current clamping effect reduces $R_1$ by $i_{\text{clamp}}^{\text{max}}$ according to (9). $R_1$ and $R_2$ are defined by:

$$R_1 = \frac{\Delta i}{2} - A_{\text{fund}} - i_{\text{clamp}}^{\text{max}}$$  (10)

$$R_2 = \frac{\Delta i}{2} + A_{\text{fund}} \cos(\phi)$$  (11)

where $A_{\text{fund}}$ is the fundamental current amplitude. The reactive current produced by the capacitor with amplitude $A_{\text{react}}$ cannot be neglected because it affects $R_2$ if the inductor current fundamental component is not in phase with the ripple. Therefore, the reactive current component has to be taken into account:

$$A_{\text{react}} = V_c2\pi60C$$  (12)

Therefore, the fundamental current amplitude is given by:

$$A_{\text{fund}} = \sqrt{A_{\text{real}} + A_{\text{react}}}$$  (13)

The angle $\phi$ in (11) is given by

$$\phi = \tan^{-1}\left(\frac{A_{\text{react}}}{A_{\text{real}}}\right)$$  (14)

where $A_{\text{real}}$ is the 60 Hz component current drawn by the current sink. Between $R_1$ and $R_2$, the error increases from zero to $V_{\text{err}}^{\text{max}}$ by the slope $k$:

$$k = \frac{V_{\text{err}}^{\text{max}}}{R_2 - R_1}$$  (15)

Therefore, the error model, $N(A)$, can be created by summing two describing function models, $N(A)_1$ and $N(A)_2$, both modeling slope and saturation [16]:

$$N(A)_1 = -\frac{2k}{\pi} \left[ \sin^{-1}\left(\frac{R_1}{A}\right) + \frac{R_1}{A} \left(1 - \left(\frac{R_1}{A}\right)^2\right)^{\frac{1}{2}} \right]$$  (16)

$$N(A)_2 = \frac{2k}{\pi} \left[ \sin^{-1}\left(\frac{R_2}{A}\right) + \frac{R_2}{A} \left(1 - \left(\frac{R_2}{A}\right)^2\right)^{\frac{1}{2}} \right]$$  (17)

$$N(A) = \Re[N(A)_1 + N(A)_2].$$  (18)

Due to the deadzone, the sum of the slopes must be zero before $R_1$. Therefore, the slope of $N(A)_1$ is $-k$. Fig. 9(b) shows how $N(A)_1$ and $N(A)_2$ sum up to $N(A)$.

Figs. 10, 11, and 12 compare the proposed model, $N(A)_1A$ as function of $A$, to the simulated voltage error with fundamental current amplitudes of 0 A, 1 A, and 1.83 A, respectively. The fundamental amplitude of 1.83 A was chosen in order to have $R_1$ in (10) zero. The proposed model estimates the nonlinear error sufficiently well. However, regardless of modeling, the zero current clamping effect the model for $R_1$ in Fig. 10, and the approximation with a single slope in Fig. 12 are not completely accurate.

IV. AMPLITUDE AND FREQUENCY RESPONSE

This section deals with the output impedance that is affected by the nonlinear deadtime effect. Based on the describing-function model $N(A)$, an output impedance model that depends on both the injection amplitude and injection frequency is derived.

Due to the nonlinear voltage error, a transfer function must be defined as a function of the amplitude $A$ and frequency $\omega$. Therefore, the notation of the inductor and capacitor impedance subsystems are changed from $Z_L(s)$ and $Z_C(s)$ to $Z_L(j\omega)$ and $Z_C(j\omega)$, respectively. The effect of $N(A)$ is included in the output impedance, and therefore, becomes a function of the input amplitude in addition to the frequency. Therefore, the output impedance is expressed as $Z_o(A, j\omega)$.

Fig. 13 shows a frequency-domain circuit presentation of the nonlinear system. The capacitor and inductor impedances
are traditional linear elements. However, the nonlinear voltage error is in series with the inductor impedance as a reverse biased nonlinear current dependent voltage source. The equation based on the circuit can be given by

\[ |N(\text{IL}) + Z_L(j\omega) + Z_C(j\omega)|i_{L}(\text{IL}, \omega) - |Z_C(j\omega)|i_o(\text{IO}, \omega) = 0. \]  

\[ (19) \]

IO, VO and IL are phasors of the output current, output voltage and inductor current, respectively. IL must be solved for each input combination on IO and \( \omega \). The phasors are analyzed separately at different frequencies, \( \omega \). \( N(\text{IL}) \) is the analytical expression from (18). The equation can be solved, for example, by using Matlab \textit{fzero} function. The solved \( i_{L}(\text{IL}, \omega) \) is in a zero phase shift because the equation could be solved only for the absolute value. Thus, the output current \( i_o \) must be resolved in order to find out the relational phase shift:

\[ i_{o}(\text{IO}, \omega) = \frac{[N(\text{IL}) + Z_L(j\omega) + Z_C(j\omega)]i_{L}(\text{IL}, \omega)}{Z_C(j\omega)} \]

\[ (20) \]

As the next step, the output impedance \( Z_o(A, j\omega) \) can be solved:

\[ Z_o(A, j\omega) = \frac{-v_o(\text{VO}, \omega)}{i_o(\text{IO}, \omega)} \]

\[ = \frac{Z_C(j\omega)i_o(\text{IO}, \omega) - Z_C(j\omega)i_{L}(\text{IL}, \omega)}{i_{o}(\text{IO}, \omega)} \]

\[ (21) \]

Fig. 14 compares the modeled and the simulated output impedance, \( Z_o(A, j\omega) \). The model predicts well that, with low injection current amplitudes when the inductor current perturbation amplitude is small enough, the resonance is damped. On the other hand, the error saturates around the resonance with higher injection amplitudes and the undamped resonance becomes visible again. Higher injection amplitudes than \( R_1 \) cause errors at lower frequencies, which can be seen as damping.

Fig. 15 shows the modeled amplitudes of the voltage error and inductor current corresponding to the case of Fig. 14. As it can be seen the error saturates at the resonance in the case of injection amplitudes of 2 A and 3 A. Due to this, the damping effect is diminished as it can be seen from Fig. 14. Figs. 16 and 17 compare the model to the simulation when the fundamental load current is 1 A and the inductor current is 1.19 A. However, the effect is modeled only at the fundamental frequency of the square-wave-like voltage error.
caused DC capacitors of the used practical circuit. Secondly, a voltage injection is used to measure the output impedance in an HIL simulation. Thirdly, a practical laboratory setup is used to verify the nonlinear dependency on the injection amplitude.

A. Practical circuit

Fig. 18 shows the circuit diagram of the experimental setup. In the practical half-bridge inverter there are DC capacitors, $C_{DC}$, instead of the two ideal voltage sources that were used previously. Because the DC voltage source is a dynamic short circuit, a parallel connection of the DC capacitors, $Z_{par}^{C_{DC}}$, is visible in series with the inductor, and is therefore, included in the inductor impedance:

$$Z_{C_{DC}}(j\omega) = r_{C_{DC}} + \frac{1}{j\omega C_{DC}} \quad (22)$$

Figure 14: Modeled (lines) and simulated (circles) output impedance with different perturbation amplitudes and a fundamental current of 0.64 A.

Figure 15: Modeled and simulated inductor current amplitude and voltage error with different perturbation amplitudes and a fundamental current of 0.64 A.

Figure 16: Modeled (lines) and simulated (circles) output impedance with different perturbation amplitudes and a fundamental current of 1.19 A.

Figure 17: Modeled and simulated inductor current amplitude and voltage error with different perturbation amplitudes and a fundamental current of 1.19 A.

$$Z_{par}^{C_{DC}} = \frac{Z_{C_{DC}} C_{DC}}{Z_{C_{DC}} + Z_{C_{DC}}} \quad (23)$$

$$Z_L(j\omega) = Z_{par}^{C_{DC}} + r_L + j\omega L \quad (24)$$

where $r_{C_{DC}}$ is equivalent series resistance of the DC capacitor.

The analysis focuses on the anti-resonance in (24) because with the used laboratory equipment it was not possible to inject voltages that cause high enough current at the LC parallel resonant frequency to demonstrate the saturation effect.

For the sake of simplicity, the filter capacitor is not included in the output impedance. Therefore, the nonlinear circuit equation that must be solved for $I_L$ reduces to

$$|N(I_L) + Z_L(j\omega)|i_L(I_L, \omega) + v_o(V_O, \omega) = 0. \quad (25)$$
The output impedance can be solved from

\[ Z_o (A, j\omega) = -\frac{v_o (V_O, \omega)}{i_L (I_L, \omega)}. \]  

(26)

The parameters of the HIL simulation in this subsection are given in Table II.

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**B. HIL simulations**

The circuit of Fig. 18 is simulated with Typhoon HIL model 402. The phase locked loop (PLL) is omitted for simplicity. The inductor current amplitude, \( I_L \), and the capacitor current amplitude, \( I_C \), are 1.01 A and 0.64 A, respectively. The 60 Hz fundamental voltage component is provided by the ideal voltage source that has a very small series connected inductor and resistor to improve the simulation stability. The voltage source is also used to inject a sine sweep for the measurement. In order to demonstrate the nonlinear damping effect, high injection amplitudes are required.

Fig. 19 shows the real-time HIL simulated frequency response and the model around the anti-resonance at 150 Hz. It can be seen that the model clearly predicts the damping as the function of the perturbation amplitude. The most striking result from Fig. 19 is that the damping increases when the injection amplitude is increased from 0.5 V to 23 V. With injection amplitudes of 40 V and 50 V, there is less damping, which is predicted correctly by the model. A notable change in the damping can be seen even with the injection amplitudes 0.5 V, 2.3 V and 11 V, which are less than 10 percent of the fundamental 170 V component amplitude.

**C. Laboratory measurements and comparison to simulations**

The nonlinear damping by the deadtime is verified by laboratory measurements. Fig. 20 shows the laboratory setup for the circuit in Fig. 18. A low bandwidth PLL is used to synchronize the half-bridge converter to the 60 Hz voltage provided by the grid emulator. Frequency response of the output impedance is measured by injecting a sine sweep with grid-emulator. The measurement are performed separately with MOSFETs and IGBTs in order to see the effect of non-ideal switches. The used switch modules were PEB Sic 8024 module and PEB 8032 module by Imperix.

Due to the used isolation transformer, there is a voltage drop the value of which depends on the injected frequency, and the injected voltage over the capacitor is not the voltage over the grid emulator. Fig. 21 shows two injections as an example. The injected output voltage amplitude is not constant, and the injected voltage over the filter capacitor varies slightly between the IGBT and MOSFET measurements. The nominal values for the injections, 4V and 24 V, are chosen from the values at around 270 Hz. As it has been shown, the nonlinear deadtime effect is very amplitude sensitive. Therefore, the practical measurement is replicated in an HIL simulation by injecting a voltage that in reality was over the filter capacitor.

Figs. 22 and 23 show the measured laboratory measurements with MOSFETs and IGBTs compared to HIL simulations, respectively. The laboratory setup is shown in Fig. 20 and passive component parameters of the setup are given in Table III. As stated earlier, ideal switches with ideal diodes were used in the HIL simulations.

The HIL simulation gives a result that corresponds to the laboratory measurements in the case of the MOSFETs as shown in Fig. 22. Fig. 23 shows the converter output impedance with the IGBT. As the figure shows, there is more visible difference in damping between the simulation and laboratory measurement with low injection amplitudes compared to the case with the MOSFET. This implies that non-idealities with the IGBT cause more problems regarding modeling of the deadtime effect. On the other hand, the used deadtime of 4 \( \mu \)s is not relevant for the MOSFETs because it is unnecessary long. However, the purpose of the present paper is to study the damping that stems from the deadtime. With the
two highest injection amplitudes, the HIL simulation matches accurately to the laboratory measurements. Furthermore, with the highest injection amplitude, the damping decreases below the resonance frequency that is visible in both the laboratory measurements and HIL simulation. The non-idealities with MOSFETs could become more visible with higher switching frequencies.

Table III: System parameter values of the laboratory measurements.

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<td>( L )</td>
<td>1.2 mH</td>
<td>( r_L )</td>
<td>0.13 ( \Omega )</td>
</tr>
<tr>
<td>( C )</td>
<td>10 ( \mu )F</td>
<td>( r_C )</td>
<td>0.1 ( \Omega )</td>
</tr>
<tr>
<td>( C_{\text{DC}} )</td>
<td>500 ( \mu )F</td>
<td>( r_{\text{C-DC}} )</td>
<td>0.246 ( \Omega )</td>
</tr>
<tr>
<td>( \omega_n )</td>
<td>2(\pi)60 rad/s</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

VI. CONCLUSIONS

This paper has investigated the nonlinear damping caused by the deadtime effect under load conditions where the fundamental current component is less than half the peak-to-peak current ripple. The findings indicate that error caused by the deadtime can be modeled with a deadzone, slope, and saturation as a function of the inductor current amplitude. Due to the deadzone, the effect cannot be modeled with a resistor-like element, as was used earlier in the literature with the deadtime under different load conditions.

By using a describing function, we found a model for the error caused by the deadtime under a low load condition. The describing-function model in combination with linear circuit impedances was used to derive the output impedance of an inverter as a function of the injection amplitude. We observed that, even with a moderate measurement injection amplitude, a system resonance can increase the inductor current amplitude making the nonlinearities become visible. This can occur especially in simulations where ideal current source that can provide infinite voltage is used for the measurement injection. The proposed model works accurately with real-time HIL simulations that are becoming increasingly popular. Experimental measurements were provided for verifying the HIL simulations and the nonlinear deadtime effect. The present study represents the first occasion that the...
output impedance of a power electronic converter has been given as a function of the injection amplitude in addition to the injection frequency. Our work has some limitations regarding the nonidealities in practical semiconductor switches. In addition, low-order harmonics except the first one were assumed non-existent. For example, considerable harmonics produced by a nonlinear load could change the effect. Nevertheless, we believe our work can be the theoretical basis on the nonlinear analysis of the deadtime effect under low load conditions. On a wider level, research is also needed to determine how the nonlinear deadtime effect and the measurement result behave under broadband injections that are becoming increasingly popular [5], [17].

REFERENCES


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