Design of Damping Strategies for LC Filter Applied in Medium Voltage Variable Speed Drive †

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Abstract: In recent years, medium-voltage variable-speed drives have become popular in the industry. However, in some applications, the use of long cables can lead to overvoltages at the motor terminals, affecting the motor lifespan. Under such conditions, the use of passive filters is recommended. However, the use of inductive capacitive (LC) filters results in resonances, leading to control stability issues. This issue can be mitigated by introducing a resonance damping strategy. In this work, four damping strategies are implemented and designed: Passive damping, capacitor current feedback, capacitor voltage feedback, and notch filter-based damping. The paper performs a comparative study on the current control phase margin, current and voltage harmonic spectra, and overall changes in the control structure. Then, the effect of the damping strategy on the VSD control is evaluated, creating guidelines to support the selection of the appropriate damping strategy. The results indicate that capacitor voltage feedback stands out, since this strategy presents an interesting dynamic behavior while allowing the elimination of the passive damping losses at a relatively low cost.

Keywords: LC filter; resonance damping; induction machine drives

1. Introduction

In recent years, the implementation of variable speed drives (VSD) has increased in the industry and off-shore applications, such as oil production facilities [1]. VSD can improve the performance of processes and provide energy savings. Voltage source inverters with pulse width modulation (PWM) stand out among the several topologies present in the market. However, the PWM high dv/dt may cause various problems, such as bearing currents, reduced machine lifetime and hot spots in the winding isolation [2,3].

Furthermore, in some applications such as underground mining, oil drilling and subsea plants, a long cable is required to connect the inverter to the machine [4,5]. The combination of high dv/dt and long cables can lead to overvoltage at the motor terminals, due to the wave reflection phenomenon [6]. A high-power filter, shown in Figure 1, is usually installed at the Neutral Point Clamped (NPC) inverter output to mitigate the mentioned problems [7] and provide a quasi-sinusoidal voltage.

With the introduction of the filter, the VSD control can be implemented using the inverter side currents \( i_{fa}, i_{fb}, i_{fc} \) [8,9], or the motor side currents \( i_{sa}, i_{sb}, i_{sc} \) [10,11]. The latter approach has the advantage of directly using a variable set required for motor field-oriented control, thus improving the orientation.

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Figure 1. Medium voltage drive topology investigated in this paper. The resistor $R_d$ is employed only in the passive damping strategy.

Furthermore, the LC filter has an inherent resonance frequency, which may cause instability in the control loops [3]. It is possible to lower the control bandwidth and improve the control stability, by adopting a frequency lower than the filter resonance. In addition, different converter topologies may improve the harmonic response. In a modular multilevel converter, the inherent arm inductor can be tuned with the filter capacitor to create a low-pass filter in the converter output, which results in a highly sinusoidal output voltage [12] and mitigates the wave reflecting in long-cables and overvoltage problems in the motor terminal.

However, there is still the need for damping strategies to mitigate the resonance frequency. Passive damping (PD) is a popular approach that adds a resistor into the circuit. The resistor can be added in parallel [13], or series with the capacitor [3]. The latter is more popular. The main drawback of this approach is the increase in the system power losses [11]. An alternative to circumvent these losses is the application of active damping (AD). AD schemes are based on modifications in the control algorithm, which lead to additional measurements.

In the literature, several AD approaches have been proposed. Reference [10] proposes a genetic algorithm to auto-adjust the pole placement of the AD. Reference [11] proposes a virtual resistance-based damping. In [14], a scheme with different virtual resistances is employed in the quadrature and direct axis. The tuning is based on small-signal analyses. In [15], the authors proposed an AD based on a capacitor voltage feedback scheme using a high-pass filter.

Damping schemes have been extensively investigated for grid-connected inverters for photovoltaic and wind energy conversion systems. For example, reference [16] presents a survey on the main AD schemes for grid-connected inverters. However, for medium voltage (MV) drive application, there is a lack in the literature regarding the comparison of different damping techniques used in the converter. In this paper, the initial study presented in [17] is extended, as four different damping strategies are considered. The PD adds a resistor in series with the filter capacitor. The other three are AD strategies, including one based on capacitor current feedback, with a gain [18]. In addition, in [19], the capacitor voltage is implemented as feedback, passing through to a derivative filter. Finally, a notch filter is implemented in the control, tuned for the resonance frequency of the filter [20]. A study, presented in [21], proposes an improvement for a synchronous capacitor-current-feedback AD strategy, which is implemented based on arbitrary pole assignment. A different technique, presented in [22], proposes a variable time delay in the control of a VSD. The additional delay is capable of reducing the resonance created by the LC filter. Another technique to mitigate the LC filter resonance phenomenon can be achieved by a DC-link voltage feedback control method and an additional high-pass filter to obtain an effective damping current [23]. Besides that, the authors in [24] point out that many AD strategies introduce a multi-loop feedback, and the work proposed a
q-axis current reference tracking for motor that suppress the filter resonance simultaneously, reducing the resonance in the LC filter.

This paper compares four different damping strategies and their impact on the system control, and develops guidelines for decision-making involving multiple damping strategies analysis. Therefore, it allows the selection of the better suitable damping for a specific topology and presents the advantages and disadvantages of each implemented strategy. This paper aims to present a methodology for the implementations of different damping strategies and compare their benefits and effectiveness.

This paper is outlined as follows. Section 2 describes the VSD system description and, the LC filter and control design. The damping strategies description and design are shown in Section 3. The case study is described in Section 4. Section 5 presents the obtained results and discussion. The conclusions are stated in Section 6.

2. Materials and Methods
2.1. LC Filter Design

The LC filter design considers that the motor behaves as an inductor and associates it with the filter inductor for low frequencies. The long cable inductance may be disregarded if its inductance value is considerably small. The equivalent circuit is presented in Figure 2.

![Figure 2](image_url)

Figure 2. The equivalent circuit for the filter represents the motor behavior as an inductor.

In Figure 2, $v_f$ represents the converter voltage output, $L_f$ is the filter inductance, $L_{ls}$ is the stator leakage inductance, $L_{lr}$ is the rotor leakage inductance and $v_f$ is a voltage source representing the output voltage of the converter. Using the presented circuit, an equivalent inductance is given by [11]:

$$L_{eq} = \frac{L_f(L_{ls} + L_{lr})}{L_f + L_{ls} + L_{lr}},$$

(1)

The relation between the switching frequency and the filter resonance frequency is given by $r_f = f_{sw} / f_{res}$. Besides, it indicates a trade-off between harmonic attenuation and acceptable closed-loop response. According to [11], a value of $r_f = 3$ is recommended. If this ratio is increased, the filter will have a low resonance frequency, which may be inside the control bandwidth and impact its response. If it is decreased, the filter response can interact with the switching harmonics, which compromises attenuation and instability [25]. Finally, to adjust the resonance frequency to the desired value, the required filter capacitance is given by:

$$C = \frac{1}{L_{eq}(2\pi f_{res})^2}.$$

(2)

2.2. Control Design

The indirect field oriented control (IFOC) [26] is employed as a control scheme for the induction motor drive. A control block diagram is shown in Figure 3.

The inner control loop regulates the direct ($i_{sd}$) and quadrature ($i_{sq}$) stator currents components [26]. The direct current control loop employs a proportional–integral (PI) controller with gains $k_{pd}$ and $k_{id}$, which regulates the current $i_{sd}$, which creates the rotor magnetic flux ($\lambda_{dr}$), thus indirectly regulating it for a settled reference. On the other hand, for the quadrature current adjustment, a proportional control with gain $k_{pq}$ is applied.
The reference \( i_{dq}^{*} \) is computed by the motor speed control loop. The approach is based on explicit feedback commands [27]. The measured angular speed \( (\omega_m) \) and its reference \( \omega_m^* \) are employed in a proportional control with gain \( k_p \) and a friction factor compensation, with a gain \( D \), equal to the friction factor. For the angular position control, a PI controller is implemented, with \( k_{pa} \) and \( k_{ia} \) gains. The angular position \( (\theta_m) \) and its reference \( \theta_m^* \), are obtained from \( \omega_m \) and \( \omega_m^* \), respectively.

In Figure 3, \( L_m \) is the Magnetizing Inductance, \( R_r \) and \( R_s \) are the rotor and stator resistance. \( P \) is the number of Poles for the IM, and \( S \) is the estimated value for the slip. \( \omega_e \) is the electrical angular frequency, and \( \omega_s^* \) is its estimated value. The current control output is the inverter voltage reference for the direct \( (v_d^*) \) and quadrature \( (v_q^*) \) axis. The electrical angle \( (\theta e^*) \) is applied in the axis transformation, to obtain the voltage reference for the converter \( (v_f) \). A delay compensation \( (f_f^{-1}) \) is applied to avoid possible instability in the field orientation scheme [28]. The final network output \( (v_f^*) \) is delivered to a space vector modulator.

The current control tuning is performed following the block diagram in Figure 4. The block diagram output is the stator current \( (i_s) \), and the input is \( i_s^* \). \( i_e \) is the capacitor current, and \( v_e \) is the stator voltage. \( G_e(s) \) is the controller transfer function, and \( G_f(s) \) is the digital implementation delay. \( Z_f(s) \) and \( Z_e(s) \) denote the inductive and capacitive filter impedance given, respectively, by:

\[
Z_f = L_f s + R_f, \tag{3}
\]

\[
Z_e = \frac{1}{Cs}. \tag{4}
\]

The IM impedance is represented by:

\[
Z_m = L_s s + R_s', \tag{5}
\]

where \( R_s \) is the stator resistance, and \( L_s' \) is the stator transient inductance, given by [26]:

\[
L_s' = L_s - \frac{L_{m}^2}{L_r}. \tag{6}
\]
Accordingly, the plant transfer function is given by:

\[ G_p(s) = \frac{I_c(s)}{V_f(s)} = \frac{Z_c}{Z_f Z_m + Z_c Z_m + Z_f Z_c}. \]  

(7)

The transfer function \( G_p(s) \) is a third-order system [16]. The Padé approximation is used to simplify the system transfer function [19]. When this operator is employed, the following approximation is obtained for \( s \to 0 \), which results in a tuning formula:

\[ G_p(s) \approx \frac{1}{s(L_f + L_f' + C R_f R_f') + R_f + R_f'}. \]  

(8)

When the high order terms are neglected, the following transfer function is obtained:

\[ G_p(s) \approx \frac{1}{sL_T + R_T}, \]  

(9)

where, \( L_T = L_f + L_f' \) and \( R_T = R_f + R_f' \). Therefore, this simplification is applied in the current control designs.

\( G_d(s) \) represents one sampling period delay, derived from computational delay, and a half period delay from modulation [29]. The delay can be represented as:

\[ G_d(s) = e^{-1.5 T_s} s \approx \frac{1}{s + 1.5 T_s}. \]  

(10)

With the presented model, the system open loop transfer function is given by:

\[ G_{olD}(s) = k_{pd} k_{id} s + k_{id} s + \frac{1}{sL_T + R_T + 1 + 1.5 T_s}. \]  

(11)

where the subscript “D” refers to the direct axis while “Q” refers to the quadrature axis. With the application of the pole placement technique, the following tuning formulas are derived [30]:

\[ k_{pd} = L_T \omega_i, \]  

(12)

\[ k_{id} = R_T \omega_i, \]  

(13)

which leads to an initial damping of 0.707 and a band-pass frequency of \( f_s/20 \), where \( f_s = 1/T_s \) and \( \omega_i \approx 2 \pi f_s/20 \) [16].

Only the proportional gain is employed in the quadrature control. Thus, the close loop transfer function is given by:

\[ G_{clQ}(s) = \frac{k_{pq}}{s^2 + s \left( \frac{1}{1.5 T_s} + \frac{R_T}{L_T} \right) + \frac{k_{pq} + R_T}{1.5 T_s L_T}}. \]  

(14)

As \( R_L/L_T \ll 1/(1.5 T_s) \), the following tuning formula can be obtained:

\[ k_{pq} = L_T \omega_i - R_T. \]  

(15)

For speed control, the considered proportional gain is obtained, using the dynamic stiffness method [27],

\[ b_a = J \omega_3 - D. \]  

(16)

While for the position loop, the proportional and integral gains are, respectively,

\[ k_{pd} = (b_a + D) \omega_2. \]  

(17)
where $\omega_1 = \omega_i / 20$, $\omega_2 = \omega_i / 10$ and $\omega_3 = \omega_i / 10$ are the cut-off frequencies for the control loop. These values are selected to allow the speed control to be slower than the current control.

2.3. Delay Compensation

The digital implementation adds a delay derived by the system sampling, while the PWM adds a half sample delay. Thus, the mentioned effects can cause instability when the ratio between the IM nominal speed and the switching frequency is low. A method can be implemented in the control output to reduce instability, as presented in [28]. This is specially important in MV drives, where the switching frequency is relatively low, due to the limitations of high voltage/high power semiconductor devices. The delay model is given by:

$$f_e = \frac{e^{i(1.5T_s\omega^*_e)}}{K(\omega^*_e, T_s)},$$

where

$$K(\omega^*_e, T_s) = \text{sinc}\left(\frac{\omega^*_e T_s}{2}\right) = \frac{2}{\omega^*_e T_s} \sin\left(\frac{\omega^*_e T_s}{2}\right).$$

To implement a compensating effect and reduce the delay effect, the inverse function of the delay model ($1/f_e$) is implemented in the $v'_f$ of the control. The equation is given by:

$$v'_f = e^{-\omega_d\omega^*_e} v'_f K(\omega^*_e, T_s),$$

for the purpose of achieving a proper implementation in the simulation, the equation is manipulated, to obtain the following response for $v'_{fd}$ and $v'_{fq}$:

$$\begin{cases} v'_{fd} &= K(\omega^*_e, T_s)(v'_f \cos \omega_d + v'_{fq} \sin \omega_d), \\ S v'_{fq} &= K(\omega^*_e, T_s)(v'_f \cos \omega_d) - v'_{fd} \sin \omega_d, \end{cases}$$

where $\omega_d = 1.5T_s\omega^*_e$. The implementation of the delay compensation network in the simulation is shown in Figure 5.

Figure 5. The diagram represents the delay compensation, $f_e^{-1}$, applied in the controller output, $v'_f$.

3. Damping Schemes

The modeling and design of each damping strategy are presented in this section. Figure 6 exhibits the individual block diagram for each damping strategy, the damping implementation is presented in red. The implemented strategies are the PD, capacitor current feedback (CCF), capacitor voltage feedback (CVF) and the notch filter (NF), respectively.
3.1. Passive Damping (PD)

In the passive damping strategy, a resistor is added in series with the filter capacitor [3]. Even though this strategy reaches its goal, it also increases the system losses. Figure 6a shows the current control block diagram. The plant transfer function is presented as follows:

\[
G_p = \frac{Z_c + R_d}{Z_f Z_m + (Z_c + R_d) Z_m + Z_f (Z_c + R_d)}.
\]  

(23)

The resistance value is selected by adjusting the pole damping in the root locus, which is obtained in the z-plane of the plant closed-loop. The transfer function is given by:

\[
H(s) = \frac{G_c G_p G_d}{1 + G_c G_p G_d}.
\]

(24)

The value of the resistance is selected based on the root locus, ranging from 0 Ω to 3 Ω, with the goal of reaching the desired damping of 0.2.

The root locus is presented in Figure 7a, where the resistance increase moves the pole in places, and increases its damping. The selected value of resistance is \( R_d = 2.4 \) Ω.

It is important to point out that this technique affects the higher frequency poles related to the plant. On the other hand, the control poles, with lower frequencies, are maintained as observed in Figure 7a. Therefore, this technique causes a small impact on the control, according to the Padé approximation.

Similarly to Equation (7), the Padé approximation is applied to Equation (23), and the following transfer function is obtained:

\[
G_p(s) \approx \frac{1}{s(L_f + L_s + C R_s R_f) + R_f + R_s'}.
\]

(25)
this equation is similar to that obtained in Equation (9). Therefore, the control is kept unaffected by the implementation of the PD.

![Figure 7. Root locus for (a) the PD, (b) CCF, (c) CVF and (d) NF. The arrows indicated the respective increase of $R_d$, $K_c$, $K_v$ and $Q$, while the red circle indicates the pole placement for the selected damp.](image)

3.2. Capacitor Current Feedback (CCF)

In this method, the capacitor current is used as feedback with a gain to reduce the resonance [19]. Figure 6b shows the system block diagram, where $K_c$ is the CCF gain.

This strategy applies a control feedback. Therefore, no additional losses are added to the system. However, three extra current sensors are required. The plant transfer function is obtained similarly to that employed for passive damping,

$$G_p = \frac{Z_c}{Z_f Z_m + Z_c Z_m + Z_f Z_c - Z_m K_c z^{-1}}, \quad (26)$$

where the closed loop transfer function is given by:

$$H(s) = \frac{G_c G_p G_d M_f}{1 + G_c G_p G_d M_f}, \quad (27)$$

where $M_f$ is the closed loop created by the feedback:

$$M_f = \frac{Z_c}{Z_c - Z_m G_p G_d}. \quad (28)$$

The same tuning strategy presented for the PD is applied for the CCF. The root locus is shown in Figure 7b, with $K_c$ variation from 0 $\Omega$ to 3 $\Omega$. Therefore, $K_c = 2.3 \Omega$ is selected to provide a damping ratio similar to that in the passive damping scheme.

It is important to observe that the control pole places change, which indicates that high values of $K_c$ affect the closed-loop response.
Applying the Padé approximation similarly to the PD:

\[
G_p(s) \approx \frac{1}{s(L_f + L_s' + CR_s'(R_f + K_c)) + R_f + R_s'}
\]  

(29)

where the term \(CR_s'(R_f + K_c)\) presents an absolute value of 1.22 \(\mu\) and can be disregarded, thus resulting in the same approximation. Therefore, the CCF does not significantly affect the control.

### 3.3. Capacitor Voltage Feedback (CVF)

The capacitor voltage feedback works similarly to the CCF. However, it uses the capacitor voltage instead of the current as feedback. The CVF applies a filter to the measured voltage with derivative behavior [31]. As for the CCF, any additional losses are added, requiring three additional voltage sensors. The filter feedback transfer function is presented as follows:

\[
L(s) = K_v C \omega_{max} s + K_f \omega_{max} \frac{s}{s K_f + \omega_{max}},
\]  

(30)

where \(K_v\) is an adjusting constant, \(\omega_{max}\) is the frequency where the filter reaches the maximum phase, and \(K_f\) is given by:

\[
K_f = \sqrt{\frac{1 - \sin(\phi_{max})}{1 + \sin(\phi_{max})}},
\]  

(31)

where \(\phi_{max}\) is the filter maximum phase. Furthermore, according to [19], the angle value must be around \(70^\circ < \phi_{max} < 80^\circ\), for the design, \(\phi_{max} = 75^\circ\) is chosen. The value of \(\omega_{max}\) is chosen as the filter resonance frequency. Figure 6c presents the system block diagram. In this case, the following transfer function is obtained:

\[
G_p = \frac{Z_c}{Z_f Z_m + Z_c Z_m + Z_f Z_c + Z_m Z_c L_z},
\]  

(32)

and the closed loop transfer function is given by:

\[
H(s) = \frac{G_c G_p G_d M_{f2}}{1 + G_c G_p G_d M_{f2}},
\]  

(33)

where \(M_{f2}\) is the closed loop created by the feedback:

\[
M_{f2} = \frac{Z_c^2}{Z_c - Z_m G_p G_d}.
\]  

(34)

For the damping tuning, \(K_v\) is varied from 0 to 3, resulting in the root locus presented in Figure 7c. A damping value of 0.2 is chosen, obtaining a value of \(K_v = 2.6\).

As observed, the CVF present some changes to the control pole allocation. Considering that Equation (30) behaves as a gain of \(K_f = K_v C \omega_{max} K_f\) for lower frequencies, the Padé approximation is applied and the following transfer function is obtained:

\[
G_p(s) \approx \frac{1}{s(L_f + L_s'(1 + K_f) + CR_s' R_f) + R_f + R_s'},
\]  

(35)

where \(K_f = 0.01\) can be disregarded, indicating this damping strategy does not considerably affect the control.
3.4. Notch Filter (NF)

The Notch filter is implemented in the controller output, with the purpose of removing the resonant frequency from within the control itself. It is important to remark that the NF uses only a transfer function, therefore, any measurement is required [32]. The filter transfer function is given by:

$$N(s) = \frac{s^2 + \omega_m}{s^2 + 2Q\omega_m + \omega_m^2},$$  \hspace{1cm} (36)

where \(\omega_m\) is the cutoff frequency, and \(Q\) is the term used for tuning the damping. From the block diagram shown in Figure 6d, the plant transfer function is:

$$G_p = \frac{Z_c}{Z_f Z_m + Z_c Z_m + Z_f Z_c'},$$  \hspace{1cm} (37)

and in the case of NF, the closed loop transfer function is:

$$H(s) = \frac{G_c G_p G_d N}{1 + G_c G_p G_d N'},$$  \hspace{1cm} (38)

In this strategy, \(Q\) ranges from 0 to 0.7, and obtains the root locus presented in Figure 7d. The selected damping was around 0.2, which resulted in \(Q = 0.33\).

As for the Padé approximation, the obtained equation is:

$$G_p(s) \approx \frac{1}{s(L_f + L'_s + CK'_f R_f + KQ) + R_f + R'_s'},$$  \hspace{1cm} (39)

where the term \(K_q = \frac{2Q(R_f + R'_s)}{\omega_m}\) presents an absolute value of 36.8 \(\mu\) and can be disregarded, thus obtaining the same equation implemented in the control design in Equation (9).

4. Case Study

In this section, the VSD parameters used for the implementation are presented in Table 1. Induction motor parameters are employed in the simulation [33].

With the application of the methodology presented in section 2.1 for the filter design, and considering a value of 0.05 pu for the inductance, as suggested in [11], \(L_f = 1.88\) mH is obtained. If the ratio \(r_f = 3\) is selected, the obtained resonance frequency is \(f_{res} = 1333\) Hz. The resonance frequency and the equivalent inductance lead to a capacitance of \(C = 9.80\) \(\mu\)F.

The gains corresponding to the control loops are presented in Table 2, using the parameters in Table 1.

Table 1. The IM, LC filter and converter parameters employed in the simulations.

| Description          | Parameters | Value    |
|----------------------|------------|----------|
| Rated Power          | \(P_n\)    | 500 hp   |
| Rated Voltage        | \(V_f\)    | 2.3 kV   |
| Rated Speed          | \(\omega_{mp}\) | 1773 rpm |
| Rated Frequency      | \(\omega_e\) | 60 Hz    |
| Poles                | \(P\)      | 4        |
| Stator Resistance    | \(R_s\)    | 0.262 \(\Omega\) |
| Rotor Resistance     | \(R_f\)    | 0.187 \(\Omega\) |
| Stator Self-inductance | \(L_s\) | 0.1465 H |
Table 1. Cont.

| Description                  | Parameters | Value     |
|------------------------------|------------|-----------|
| Rotor Self-inductance        | $L_r$      | 0.1465 H  |
| Magnetizing Inductance       | $L_m$      | 0.1433 H  |
| Moment of Inertia            | $J$        | 11.062 kg m$^2$ |
| Friction Coefficient         | $D$        | 0.402 N ms|
| Switching Frequency          | $f_{sw}$   | 4 kHz     |
| Sampling Time                | $T_s$      | 0.25 ms   |
| Reactance/Resistance ratio   | $X_L/R$    | 40        |
| Filter Inductance            | $L_f$      | 1.88 mH   |
| Filter Capacitance           | $C$        | 9.80 µF   |

Table 2. Gains for each control loop considering the filter inductance.

| Control Loop | Proportional | Integral |
|--------------|--------------|----------|
| Direct Current| 10.3 Ω       | 329 Ω/s  |
| Quadrature Current | 9.85 Ω     | -        |
| Speed        | 347 N ms     | -        |
| Position     | 2181 Nm      | 137 Nm/s |

For the cable parameter and model, it is considered an operation in a mining facility, where a suitable cable length is 100 m [34]. The BITNER catalog [35] offers a variety of options for the cable selection, and with a limiting maximum voltage drop of 1%, the model GP5001 fulfills the requirements, with a cross-section of 16 mm$^2$ and voltage rating of 3.6 kV. The cable parameters are presented in Table 3.

A pi model is applied to represent the cable dynamics, the resistance, inductance and capacitance values are obtained from the cable length and the parameters in Table 3. The resulting values are 0.115 Ω for the resistance, 37.2 µH for the inductance and 5.5 nF for each capacitance of the branch.

Table 3. Parameters for the cable model.

| Unit Resistance | Unit Inductance | Unit Capacitance |
|-----------------|-----------------|------------------|
| Ω/km            | mH/km           | µF/km            |
| 1.15            | 0.372           | 0.11             |

5. Results

As mentioned in the damping design, some strategies may affect the closed-loop poles. The bode diagrams shown in Figure 8 are used to analyze the phase margin of each damping strategy for stability assessment.

A reference is needed to compare the damping strategies. The Padé approximation is chosen, as it is used for the control tuning, and any changes in the stability are undesired, with a phase margin of 66.3°. In Figure 8, it can be observed that the highest impact occurs in the NF strategy, with a phase margin of 62.5°, and variation of 3.8°, in comparison with the Padé approximation, which indicates that the control stability is maintained.

For further comparison, simulations were developed in Matlab/Simulink R2020b for operation with the parameters shown in Table 1. In Figure 9 is exhibited the speed response obtained, and Figure 10 shows the torque response. These responses are obtained for a speed ramp applied at 4 s, and a load step in 6 s.
Figure 8. Bode diagram presenting the open loop margin phase, for the transfer function of the (a) Padé approximation, (b) PD and (c) CCF, (d) CVF and (e) NF.

Figure 9. Rotor speed response for an acceleration at 4 s up to the rated speed, and rated torque applied at 6 s for the passive and active damping for: (a) PD, (b) CCF, (c) CVF and (d) NF. A zoom is presented when the load is added to the circuit in (e) PD, (f) CCF, (g) CVF and (h) NF.
For all implemented strategies, the VSD maintains its stability and presents a similar overshooting of 50 rpm when the load is applied. The control is capable of reestablishing the reference for the employed damping strategies, presenting no apparent difference.

In the steady state, the electrical torque developed by the IM presents undesirable oscillation. The worst case is presented by the NF, with an oscillation of 160 Nm. On the other hand, the others strategies present an oscillation of 95 Nm, 80 Nm and 100 Nm, respectively, for the PD, CCF and CVF damping strategies. The CVF presents the smallest oscillation, indicating better performance in that case.

For the comparison of each damping strategy performance, a simulation without any damping strategy is used as a basis to evaluate their damping effectiveness. The results are exhibit in Figure 11, with the motor voltage, stator current, and their respective Fourier spectra.
Both the voltage and current response present harmonics centered around the switching frequency, which can be confirmed by the spectra in Figure 11c,d, whose high magnitude peak is situated close to the filter resonance frequency of 1.3 kHz. The current presents a THD of 45.7%, and the voltage a THD of 228%. The stator currents for each damping strategy are shown in Figure 12, and its Fourier spectra, in Figure 13.

![Stator current for IM: (a) PD, (b) CCF, (c) CVF and (d) NF damping strategy.](image)

**Figure 12.** Stator current for IM: (a) PD, (b) CCF, (c) CVF and (d) NF damping strategy.

![Harmonic Spectrum for the stator current: (a) PD, (b) CCF, (c) CVF and (d) NF damping strategy.](image)

**Figure 13.** Harmonic Spectrum for the stator current: (a) PD, (b) CCF, (c) CVF and (d) NF damping strategy. The resonance frequency is indicated by a red line.

The damping strategies reduce the oscillation presented in Figure 11b at the filter resonance frequency. However, the NF damping strategy still maintains the presence of harmonics in such frequency, which increases the maximum THD presented for the stator current to 2.7%, the highest value in Figure 13. In comparison with the other strategies, with value of 1.4% for the PD, 1.3% for the CVF and the smallest THD for the CCF with 0.9%. Even so, the NF strategy still presents a considerable reduction.

The line-to-line voltage at the motor terminals is shown in Figure 14, and its Fourier spectra, in Figure 15.
Figure 14. Line-to-line voltage at the motor terminals for: (a) PD, (b) CCF, (c) CVF and (d) NF damping strategy.

Figure 15. Harmonic Spectrum for the line-to-line motor voltage: (a) PD, (b) CCF, (c) CVF and (d) NF damping strategy. The resonance frequency is indicated by a red line.

The filter introduction is capable of mitigating switching effects derived from the converter, as none of the strategies presented a considerable component around the switching frequency of 4 kHz. Besides that, the damping strategies present THD values of 2.5%, 2.7%, 2.7% and 11.0%, for the PD, CCF, CVF and NF, respectively. Furthermore, the resonance peak, observed in Figure 11d, is mitigated for each damping strategy. The NF presents the highest resonance peak due to the resonance component still present in the Fourier spectra. However, it is important to remark that the strategy still reduces considerably the resonance peak.

A brief comparison of each strategy is presented in Table 4. The PD strategy presents losses of 200 W under nominal condition, which accounts for 0.05% of the VSD power. The CCF requires three additional current sensors. In the NF strategy, additional sensors are not required, the resonance peak presents a minor reduction in comparison to the other strategies, as shown in Figure 14, and is sensitive to any resonance frequency changes. The
CVF presents an effective damping with no additional loses, and is cheaper than the CCF, being an appropriated solution for the studied VSD.

Table 4. Positive and negative characteristics for each damping strategy considered.

| Strategy | Positives                                                                 | Negatives                                      |
|----------|----------------------------------------------------------------------------|------------------------------------------------|
| PD       | -No additional measurements                                               | -Additional losses                             |
|          | -Closed-loop poles unaffected                                              |                                                 |
| CCF      | -No additional losses                                                      | -Require 3 extra current sensor                |
|          | -Barely affects the closed-loop poles                                      | -More expensive                                |
| CVF      | -No additional losses                                                      | -Require 3 extra voltage sensor                |
|          | -Voltage measurements is less expensive than current measurements          | -More expensive                                |
| NF       | -No additional losses                                                      | -Less effective damping                        |
|          | -No additional sensors                                                    | -Sensitive to resonance frequency changes      |

6. Conclusions

This paper presented damping strategies applied to a medium voltage induction motor with an LC filter connected to its output. The performance of the strategies is evaluated according to the effects on the VSD control structure and the harmonic response.

The impact on the control stability is evaluated through the phase margin, using as a basis the Padé approximation closed-loop transfer function. The phase margin is maintained for all the applied damping. Furthermore, the IM speed and torque response are used for analysis regarding the implemented strategy impact on the transient response. Harmonic analyses are presented for all four damping strategies, for comparing voltage and current spectra.

The results demonstrated that, despite the resonance effect caused by the filter, the damping strategies can reduce its effects considerably. This highly desired behavior indicated that, through an adequate design of the active damping strategy, resonance can be attenuated without impairing the VSD performance.

While the NF presented a less effective performance, it does not require any changes in the power circuit, since no resistors nor additional measurements are required. The four strategies can mitigate the resonance and maintain the system stability. The CCF and CVF are similar, and the main difference is the measured variable, which gives an advantage to the CVF, as its sensors are less expensive.

Further research may be applied to different damping strategies, filter and converter topologies, and IM with different nominal voltage and power. An experimental comparison may also be provided.

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Abbreviations

The following abbreviations are used in this manuscript:

AD Active damping
CCF Capacitor current feedback
CVF capacitor voltage feedback
IFOC Indirect field oriented control
IM Induction motor
MV Medium voltage
NF Notch filter
NPC Neutral Point Clamped
PD Passive damping
PWM Pulse width modulation
VSD Variable speed drives

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