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Passivity-Based Stability Assessment of Grid-Connected VSCs—An Overview

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Abstract—The interconnection stability of a grid-connected voltage-source converter (VSC) can be assessed by the passivity properties of the VSC input admittance. If critical grid resonances fall within regions where the input admittance acts passively, i.e., has nonnegative real part, then their destabilization is generally prevented. This paper presents an overview of passivity-based stability assessment, including techniques for space-vector modeling of VSCs whereby expressions for the input admittance can be derived. Design recommendations for minimizing the negative-real-part region are given as well.

Index Terms—Converter control, passivity, resonances, stabilization.

I. INTRODUCTION

THE PENETRATION of grid-connected converters, particularly voltage-source converters (VSCs), is currently rising rapidly [1]. As a consequence, the risk increases for destabilization of critical grid resonances [2]–[23]. These are resonances that are poorly damped and electrically located close to one VSC, or multiple VSCs, of high enough power rating in relation to the short-circuit ratio (SCR) of the grid.

Stability of a system comprising multiple grid-connected VSCs is generally difficult to analyze (e.g., using eigenvalues [13] or the Nyquist criterion [16]), particularly if the VSCs have different ratings and dynamic properties. On the other hand, the frequency-domain passivity theory offers an effective method for stability assessment [2]–[7].

Suppose that all VSCs, as well as all other power-system components, have a passive behavior, i.e., the real part of the input admittance (also called the conductance) is nonnegative for all frequencies. Then, the system is guaranteed to be stable regardless of the number of converters, because a network that consists solely of passive components—no matter how complex—is always stable.

However, the pure passivity of all components is impossible to obtain. VSCs have negative-conductance behavior in certain frequency ranges [2]. So do electrical machines (due to the induction generator effect [24]). Yet, if it can be ascertained that every grid-connected VSC has a nonnegative conductance in frequency regions where critical grid resonances appear—i.e., partial passivity—then it is unlikely that the VSCs will induce resonance destabilization. Grid codes based on this concept are enforced by several European administrations of electrified railways. It is typically required that nonnegative conductance above the fifth-harmonic frequency must be demonstrated for any new or retrofitted active-front-end rail vehicle to be approved [25], [26].

The main objective of this paper is to make a comprehensive overview of the passivity properties of the VSC input admittance. Important previously found results are highlighted. Modeling methods of three-phase VSCs for grid-interaction studies are reviewed, and design recommendations for minimizing the negative-real-part regions are given.

The stage is set in Section II, where a review of critical grid resonances is made, and the details of the assumed VSC control system are discussed. The passivity properties of the so-called inner input admittance, which results from just the current controller (CC) and the pulsewidth modulator (PWM), are considered in Section III. Requirements on the controller parameters and on the total time delay for obtaining passivity are presented. In Section IV, the consideration is extended to the total input admittance, which includes the impact of outer loops that feed into the CC, i.e., the phase-locked loop (PLL) and the direct-voltage controller (DVC). Here, the main novel result of this paper appears. It is shown how complex space vectors and complex transfer functions can be applied to assess the passivity properties of the total input admittance (even though the latter is $dq$ imbalanced [27]). The analysis method has similarities to that suggested in [10], [15], and [17], but the impact of the DVC is included in addition to the impact of the PLL. Moreover, the usage of space vectors rather than phase quantities obviates the need for deriving separate positive- and negative-sequence expressions.

II. PRELIMINARIES

A. Critical Grid Resonances

Grid resonances that are vulnerable to destabilization by VSCs can broadly be classified into two categories.

BibTeX:
1) Harmonic Resonances: Resonances in this category appear in the range from hundreds of hertz up to a few kilohertz. The first incident of harmonic resonance destabilization, as known to the authors, occurred in the Swiss single-phase rail grid in 1995 [28]-[31]. Harmonic resonances are caused by the inherent impedance characteristics of power lines and cables, often in conjunction with converter input filters, such as inductance–capacitance–inductance (LCL) filters.

2) Near-Synchronous Resonances: In this category, we find resonances in the range from the synchronous (fundamental) frequency \( f_1 \) up to roughly \( 2f_1 \), which typically appear in very weak grids, i.e., with an SCR approaching 1 p.u. [32].

Subsynchronous resonances [33], i.e., below \( f_1 \), can be classified in this category as well.

B. Causes of Resonance Destabilization

The causes of negative-conductance behavior of a VSC (in turn, possibly leading to resonance destabilization) are as follows:

1) the total, i.e., computation-plus-PWM, time delay \( T_d \);
2) the CC dynamics;
3) the dynamics of the outer controllers, i.e., the PLL and, provided that such are used, the DVC and the controller for the reactive power or the point-of-common-coupling (PCC)-voltage magnitude.

Causes 1) and 2) affect harmonic resonances, whereas causes 2) and 3) affect near-synchronous resonances.

C. System Model

In the following, the positive- and negative-sequence synchronous components are, for convenience, referred to as +1 and −1, respectively. Harmonics are referred to with their signed order in a similar fashion. Boldface letters are used to denote complex space vector and transfer functions that operate on complex space vectors. The derivative operator is denoted by \( s = d/dt \) (which shall be considered as the complex Laplace variable, where appropriate). Vectors and transfer functions referred to the stationary \( a\beta \) frame are denoted with the superscript \( s \), whereas vectors and transfer functions referred to the synchronous \( dq \) frame aligned with +1 are denoted without a superscript. We introduce \( T_s = 1/f_s \) as the sampling period, and \( \omega_1 = 2\pi f_1 \) and \( \omega_0 = 2\pi f_0 \) as the angular synchronous and angular sampling frequencies, respectively.

In the control system, the CC closes the innermost, and fastest, loop for the converter current \( i^c \) (see Fig. 1). Over-modulation is assumed not to occur, and switching harmonics are disregarded, allowing the PWM process to be modeled as lumped with the computational time delay into the total time delay \( T_d \). The converter is assumed to be equipped with an inductive input filter, with inductance \( L \) and a negligible resistance. Hence, the converter-current dynamics are in the \( a\beta \) frame governed by

\[
i^c = \frac{E^c - v^c}{sL}, \quad v^c = e^{-sT_d}v^c_{\text{ref}} \tag{1}\]

where \( E^c \) is the PCC voltage and \( v^c_{\text{ref}} \) is the reference vector to the PWM, by which the converter voltage \( v^c \) is generated. The \( dq \)-frame correspondence is obtained simply by substituting \( s \to s + j\omega_1 \) [27]

\[
i^d = \frac{E^d - v^d}{(s + j\omega_1)L}, \quad v^d = e^{-sT_d}v^d_{\text{ref}} \tag{2}\]

The grid impedance \( Z^s(s) \), which is assumed to be balanced (also known as symmetric [27]), adds the following relations in the \( a\beta \) and \( dq \) frames, respectively:

\[
v^g - Z^s(s)i^c = E^c - v^d \tag{3}\]

where \( Z(s) = Z^s(s + j\omega_1) \) and \( v^g \) is the stiff grid voltage. The PLL and the DVC feed into the CC with signals \( \omega(s) \) and \( \theta(s) \), respectively (see Section II-E for details). From these signals, together with \( i^d_{\text{ref}} \), the \( a\beta \)-frame converter-current reference is formed as \( i^c_{\text{ref}} = e^{j\phi} (i^d_{\text{ref}} + ji^q_{\text{ref}}) \). The DVC has as input the dc-link voltage \( v_d \), which is measured across the dc-link capacitor (with capacitance \( C_d \)).

D. CC

The two CC options that most frequently are suggested in the literature are here reviewed.

1) \( dq \)-Frame CC: The CC in this case given by

\[
v^c_{\text{ref}} = -e^{j\omega_1 T_d} [F_c(s)(i_{\text{ref}} - i) + j\omega_1 Li] \tag{4}\]

where \( F_c(s) \) is the controller transfer function and the angle-adjustment factor \( e^{j\omega_1 T_d} \) compensates the angle-displacement factor \( e^{-j\omega_1 T_d} \) in (2). For brevity, PCC-voltage feedforward [3] is not included, but a \( dq \) decoupling term \( j\omega_1 Li \) is added. This prioritizes control of +1 over −1.

A proportional (P)-plus-resonant (R) controller with reduced-order generalized integrators (ROGIs) [34] as R parts is a suitable choice

\[
F_c(s) = \alpha_c L \left( 1 + \sum_{h} \frac{a_h e^{j\phi_h}}{s - jh\omega_1} \right) \tag{5}\]

where \( \alpha_c \) is the ideal closed-loop-system bandwidth, \( a_h \) (with dimension angular frequency) is the individual gain factor of the R part for the harmonic order \( h \), and \( \phi_h \) is the compensation angle of that R part [35], [36]. Characteristically, R parts are included for +1 and −1 as well as for balanced harmonics, i.e., orders −5, +5, −7, +7, −11, +11, . . . [34]. These translate to \( h = 0, -2, \pm 6, \pm 12, \ldots \) in the \( dq \) frame. The R part at \( h = 0 \) reduces to a pure integrator. Selecting \( a_h \ll \alpha_c \) is recommended [37].
Combining (2) and (4) yields
\[ i = G_{ci}(s)i_{ref} + Y_1(s)E \] (6)
where the inner closed-loop system and the inner input admittance, respectively, are given by
\[ G_{ci}(s) = \frac{e^{-sT_d}F_c(s)}{[s + j\omega_1(1 - e^{-sT_d})]L + e^{-sT_d}F_c(s)} \] (7)
\[ Y_1(s) = \frac{1}{[s + j\omega_1(1 - e^{-sT_d})]L + e^{-sT_d}F_q(s)} \] (8)

By inner, it is meant that the outer control loops, i.e., the PLL and the DVC, are not yet considered. Owing to the \( dq \) decoupling and the angle-adjustment factor, \( G_{ci}(0) = 1 \) irrespective of \( F_c(0) \).

**Remark 1:** In (5), each ROGI pair at \( h = \pm 6, \pm 12, \ldots \) can, provided that \( a_h \) and \( \phi_h = -\phi_h \), be merged into a second-order generalized integrator (SOGI) [35], [36] as
\[ \frac{a_h e^{j\phi_h}}{s - jh\omega_1} = \frac{a_h e^{-j\phi_h}}{s + jh\omega_1} = 2a_h \cos \frac{\phi_h}{2} \] (9)

This reduces the total computational burden, since complex coefficients are avoided. If desired, the ROGI for \( -1 \) (i.e., \( h = -2 \)) can be replaced by an SOGI with \( h = 2 \) according to the right-hand side of (9).

2) \( \alpha \beta \)-Frame CC: An equivalent \( \alpha \beta \)-frame implementation of control law (4) can be obtained simply by substituting \( s \rightarrow \tilde{s} \), where
\[ \tilde{s} = s - j\omega_1. \] (10)

We get
\[ v_{ref}^s = -e^{j\omega_1 T_d}[F_c(\tilde{s})(i_{ref}^s - i^s) + j\omega_1 L i^s]. \] (11)
The correspondences to (7) and (8) too are obtained simply by substituting \( s \rightarrow \tilde{s} \)
\[ G_{ci}^s(\tilde{s}) = \frac{e^{-sT_d}F_c(\tilde{s})}{(s - j\omega_1 e^{-sT_d})L + e^{-sT_d}F_c(\tilde{s})} \] (12)
\[ Y_1^s(\tilde{s}) = \frac{1}{(s - j\omega_1 e^{-sT_d})L + e^{-sT_d}F_q(\tilde{s})}. \] (13)

**Remark 2:** It should be observed that the angle-adjustment factor and the \( dq \) decoupling term remain in (11), which is not common practice in \( \alpha \beta \)-frame control. However, both are useful in this case as well, as they for \( +1 \) compensate the static voltage drop \( j\omega_1 L i^s \) across the filter inductor, thus giving prioritized control of \( +1 \). As a result, \( G_{ci}^s(\omega_1) = 1 \) irrespective of \( F_c(0) \).

**Remark 3:** Notice that, even for \( \alpha \beta \)-frame implementation, controller (5) is designed as referred to the \( dq \) frame, i.e., with \( h = 0, -2, \pm 6, \ldots \). It is then transformed to the \( \alpha \beta \) frame by the substitution \( s \rightarrow \tilde{s} = s - j\omega_1 \).

**Remark 4:** For \( \alpha \beta \)-frame control, the ROGIs for \( \pm 1 \) can be merged into an SOGI according to (9). However, replacing the ROGIs for harmonics with SOGIs only adds to the computational burden, as the system order doubles for each controlled harmonic [34].

3) **Bandwidth Selection:** The following two assumptions allow a simplified stability analysis of the current control loop to be carried out: 1) The \( R \) parts have negligible impact (which is reasonable, given the aforementioned recommendation \( a_q \ll a_c \)). 2) The imperfect \( dq \) decoupling that results from the time delay can be neglected, i.e., \( 1 - e^{-sT_d} \approx 0 \). Then, \( G_{ci}(s) = G_k(s)/[1 + G_k(s)] \), where
\[ G_k(s) = \frac{a_c e^{-sT_d}}{s} \] (14)
is the open-loop transfer function. Since \( |G_k(j\omega_c)| = 1 \), \( \omega_c \) is the crossover frequency. Thus, the phase margin is given by
\[ \phi_m = \pi + \arg G_k(j\omega_c) = \frac{\pi}{2} - \omega_c T_d. \] (15)

With the total time delay expressed in the sampling period as \( T_d = nT_s = 2\pi n/\omega_s \), a bandwidth selection recommendation can be obtained as
\[ a_c \leq \left( \frac{\pi}{2} - \phi_m \right) \frac{\omega_s}{2\pi n}. \] (16)
The selection recommendation \( a_c \leq \omega_s/10 \) of [38] is obtained as a special case of (16), e.g., for \( \phi_m = \pi/5 = 36^\circ \) and \( n = 1.5 \).

**E. Outer Control Loops**

To save space, the loop that, via \( i_{ref}^s \), controls the reactive power or the PCC-voltage magnitude is disregarded, and a constant \( i_{ref}^s \) is considered (see Fig. 1).

1) **PLL:** The purpose of the PLL is to track the rotation of the PCC voltage vector, thereby aligning the \( dq \) frame (with angle \( \theta \) relative to the \( \alpha \beta \) frame) with the +1 component of \( E^s \), whose magnitude is \( E_0 \). The PLL uses the imaginary part of \( \mathbf{E} = e^{-j\theta}\mathbf{E}^s \) as input signal, which is fed to the PLL controller \( F_p(s) \). This is typically a P-integral controller, which can be expressed as
\[ F_p(s) = \frac{a_p}{E_0} \left( 1 + \frac{a_p}{s} \right) \] (17)
where the normalization of the input signal is made by the division by \( E_0 \), and where the gains (with dimension angular frequency) typically are selected as \( a_p < a_p \ll a_c \). To the PLL-controller output, \( \omega_1 \) is added, and the sum signal is then integrated to form the transformation angle as
\[ \theta = \frac{1}{s} [F_p(s) \text{Im}\{\mathbf{E}\} + \omega_1]. \] (18)
The PLL thus forces \( \text{Im}\{\mathbf{E}\} \) to zero in the steady state (disturbances disregarded), leaving \( \mathbf{E} = E_0 \). Assuming power-invariant space-vector scaling or per unit values, the complex converter input power is given by \( \mathbf{S} = \mathbf{E}i^s \) [39]. With \( \mathbf{E} = E_0 \), we thus have
\[ \mathbf{S} = P + jQ = E_0(i_d - ji_q) \] (19)
which shows that \( i_d \) and \( -i_q \) are the active-power-producing and reactive-power-producing current components, respectively.
2) DVC: The purpose of the DVC is to make $v_d$ track its reference $v_d^{ref}$. With $W_d = C_d v_d^2/2$, the energy balance of the dc link can be expressed as

$$\frac{dW_d}{dt} = \frac{C_d d v_d^2}{2} = P - P_l$$

where $P = \text{Re} \{ S \}$ and $P_l$ is the load power, including the converter losses ($P_l < 0$ for inverter operation). Since $i_d$ is the active-power-producing current component, the following control law can be used:

$$i_d^{ref} = F_d(s)(W_d^{ref} - W_d), \quad W_d^{ref} = \frac{C_d (v_d^{ref})^2}{2}.$$  \hspace{1cm} (21)

This allows the DVC to be structurally similar to the PLL controller (17)

$$F_d(s) = \frac{a_d}{E_0} \left( 1 + \frac{a_d}{s} \right)$$

and a similar parameter selection recommendation applies, i.e., $a_d < a_d \ll a_c$ (a deviation of the value of $C_d$ used in the control system from the actual value does not give a static control error, but effectively alters $a_d$).

### III. PASSIVITY PROPERTIES OF THE INNER INPUT ADMITTANCE

In this section, the impact of the outer control loops is disregarded, reducing the control system to the CC and the PWM. Yet, the important findings can be made by analyzing this reduced system. Because this system is linear and balanced, such an analysis is relatively straightforward [3]. The total time delay plays an important role at harmonic frequencies. Therefore, the elaborate discussions concerning this parameter are first made.

#### A. Total Time Delay

In digital VSC control systems, particularly for two-level VSCs, it is useful to sample the converter current synchronously in between switchings, i.e., coinciding with the peaks of the triangular carrier signal for suboscillation PWM, as shown in Fig. 2. Thereby, switching harmonics are suppressed from the samples, often allowing antialiasing filtering to be avoided [35]. There are two variants of this principle.

1) Double-Update PWM: In this variant, samples are taken both at the positive and negative peaks, as shown in Fig. 2. The sampling frequency is twice the switching frequency. The computation time of the CC is generally a fraction of $T_s$, but it is not negligible. For this reason, it is common practice to delay the update of the phase-voltage reference $v^{ref}_d$ (given here without a specific phase notation) by one sampling interval, as shown in Fig. 2 (dotted arrows). The current sample taken at time $t = kT_s$ updates $v^{ref}_d$ at $t = (k + 1)T_s$, and so on. Thereby, even short pulses, such as that about $t = (k + 3)T_s$, can be generated without any error due to computation. Each update of $v^{ref}_d$ affects just one switching event, which can occur minimum immediately and maximum $T_s$ after the update. The average (and unavoidable) PWM time delay is $0.5T_s$, which results in the total time delay (on average) $T_d = 1.5T_s$.

2) Single-Update PWM: In this variant, the sampling frequency is set equal to the switching frequency. Consequently, the value of $T_s$ is in this case twice that in double-update PWM (given that the same switching frequency is used in both cases). Current sampling is made either at the positive or negative peaks of the carrier signal, as shown in Fig. 3. Provided that the CC computation time is shorter than $0.5T_s$, reference update can be delayed just until the next peak of the opposite sign (rather than the entire period $T_s$) [44]. Each update of $v^{ref}_d$ affects two switching events, which occur symmetrically about the (in this example, negative) peak. The average PWM time delay taken over two such consecutive switching events is obviously $0.5T_s$, giving $T_d = T_s$.

#### B. Passivity Properties for Different Total Time Delays

In [2] and [5], it is shown that the negative-real-part region caused by the total time delay begins at the critical frequency $f_{crit}$ at which $\cos \omega T_d$ for $\omega = 2\pi f_{crit}$ changes sign from positive to negative, i.e., for $2\pi f_{crit} T_d = \pi/2$, giving

$$f_{crit} = \frac{1}{4T_d} = \frac{f_s}{4} \frac{T_s}{T_d}.$$  \hspace{1cm} (23)

For double-update PWM and single-update PWM, i.e., respectively, with $T_d = 1.5T_s$ and $T_d = T_s$, we obtain $f_{crit} = f_s/6$ and $f_{crit} = f_s/4$, respectively. Let us verify this numerically. A normalized switching frequency of 100 p.u. is considered (with the synchronous frequency as base frequency). This is a reasonable value, accounting for, e.g., a 5-kHz switching frequency at $f_1 = 50$ Hz. Consequently, $\omega_s = 200$ p.u. for double-update PWM and $\omega_s = 100$ p.u.
for single-update PWM. CC (11) is used, with \( \alpha_c = 8 \) p.u. Equation (15) shows that this selection yields \( \phi_m = 61^\circ \) for single-update PWM with \( T_d = ts \) and \( \phi_m = 68^\circ \) for double-update PWM with \( T_d = 1.5ts \), which both are generous values. R parts (ROGs), with \( \alpha_h = 0.2 \) p.u. for all \( h \), are included for \(+1, -1, -5, +7, -11, +13\) in the \( \alpha\beta \) frame, i.e., for \( h = 0, -2, \pm 6, \pm 12 \) in (5). The R-part compensation angles are selected as a compensation of the time delay at the R-part frequency [4]

\[
\phi_h = h\omega_1 T_d. \tag{24}
\]

In Fig. 4, the real parts of the inner input admittance are shown in the frequency region up to the Nyquist frequency, i.e., \(-\omega_s/2 \leq \omega \leq \omega_s/2\) (since the continuous-time models used here do not account for the effects of aliasing and PWM, evaluation for frequencies above the Nyquist frequency is meaningless). As the solid curves show, for neither one of the PWM variants is a passive system obtained; negative-real-part regions appear above the respective critical frequencies predicted by (23). A proposal for eliminating the negative-real-part region up to the Nyquist frequency for \( T_d = 1.5Ts \) by a scheme based on PCC-voltage feedforward can be found in [5]. The scheme can easily be adapted to the case \( T_d = Ts \).

If the total time delay can be brought down to \( T_d = 0.5Ts \), i.e., just the PWM time delay, then (23) gives \( f_{\text{crit}} = f_s/2 \), i.e., the negative-real-part region is eliminated and a passive inner input admittance up to the Nyquist frequency is obtained, as verified in Fig. 4 (dashed curves). (This fact is hinted in [40, eq. (6)], but it is not shown explicitly. It is implicitly shown via the Nyquist criterion in [41] and [42].) In addition, \( \alpha_c \) can be increased, for a certain \( \phi_m \), by \( 1.5/0.5 = 3 \) for double-update PWM and by \( 1/0.5 = 2 \) for single-update PWM, as shown by (16).

It is generally easier to reduce \( T_d \) to (or close to) \( 0.5Ts \) for single-update PWM than for double-update PWM [43]–[45]. This is because for single-update PWM, the sampling instant can be shifted between the positive and negative peaks, as shown in Fig. 5 (where the computational time delay is \( \sim 0.2Ts \)). When \( v_{\text{ref}} \) suddenly decreases, the current-sampling instant is shifted from the negative to the positive carrier peak. This allows the short positive pulse about \( t = (k + 1)Ts \) to be generated without timing error.

For double-update PWM, the sampling instants can be shifted away from the peaks, but at the expense of a greatly increased harmonic content of the current samples [40].

**Example:** Single-update PWM with \( f_s = 10 \) kHz and \( T_d = Ts \) is implemented in the control system for an LCL-filter-equipped VSC operating with \( f_1 = 50 \) Hz. The resonant frequency is \( 2.1 \) kHz, i.e., below \( f_{\text{crit}} = f_s/4 = 2.5 \) kHz.

As can be observed in Fig. 6, the system is initially stable. At the center of the displayed time interval, the interrupt for current sampling is shifted, so that \( T_d = 1.5Ts \) is obtained, giving \( f_{\text{crit}} = f_s/6 = 1.7 \) kHz. The resonance now falls within the negative-real-part region, and as a result, growing oscillations commence.

**C. Passivity Properties About the R-Part Frequencies**

Fig. 7 shows the detail about \(-11\). For the compensation-angle selection (24), \( \text{Re}(Y^j(−j11ω)) \) is a local minimum, so a local negative-real-part region is avoided, as the solid curve shows. This holds for all R-part frequencies for which \( \cos(h\omega_1 T_d) > 0 \) and \( |h\omega_1| < \omega_s/2 \) [4].
in many publications on proportional-resonant (PR) controllers (including [34]), the feature of a compensation angle is not even included, implying \(\phi_h = 0\). In that case, negative-real-part regions appear about the R-part frequencies, as shown in Fig. 7 (dashed curve). Although the regions are narrow, the large negative values obtained for larger \(|h|\) may yet be enough to destabilize ill-located grid resonances [4]. Thus, the usage of a properly selected compensation angle is highly recommended.

D. Passivity Properties With PCC-Voltage Feedforward

In [3], it is shown that, if feedforward of the +1 component of \(E^i\) is combined with an R part for \(+1\) \((h = 0\) in (5)), then a negative-real-part region about \(+\omega_0\) results. Caution is thus advised. In addition, (24) needs to be modified, as shown in [4], to prevent negative-real-part regions about the R-part frequencies.

IV. PASSIVITY PROPERTIES OF THE TOTAL INPUT ADMITTANCE

The impact of the PLL and the DVC is now included. The complexity of analysis increases markedly, since the dynamics are nonlinear and imbalanced. Rather than resorting to the usage of a multivariable model, involving real space vectors and transfer matrices [2], [27], modeling is still made using complex space vectors and transfer functions. This method is akin to that in [10], [15], and [17]. The differences are that we elect to use the \(dq\) frame rather than a per-phase analysis (thus, obviating the need for deriving separate positive- and negative-sequence expressions), and that the impact of the DVC is considered in addition to the PLL.

A. Impact of the PLL for an \(a\beta\)-Frame CC

Since a nonlinear system is obtained, linearization must be made to allow transfer functions to be derived. For this sake, a perturbation \(\Delta E\) about the operating point \(E_0\) of the PCC voltage is considered. For a constant \(\omega_1\), this yields the \(a\beta\)-frame vector

\[
E^i = e^{j\omega_1 t} (E_0 + \Delta E).
\]

Introducing a perturbation also in the \(dq\)-frame angle, as \(\theta = \omega_1 t + \Delta \theta\), gives \(E = e^{-j\theta} E^i = e^{-j\Delta \theta} (E_0 + \Delta E)\). This relation can be linearized by approximating \(e^{-j\Delta \theta} \approx 1 - j\Delta \theta\)

and by neglecting cross terms between the perturbation quantities, yielding

\[
E = E_0 + \Delta E - j E_0 \Delta \theta. \tag{26}
\]

Substitution of (26) in (18) results in

\[
\Delta \theta = \frac{F_p(s) \text{Im}\{\Delta E\} - E_0 F_p(s) \Delta \theta}{s + E_0 F_p(s) \text{Im}\{\Delta E\}}. \tag{27}
\]

If the integral term of (17) is neglected, then \(G_p(s) = [a_p/(s + a_p)]/E_0\), i.e., \(a_p\) is the closed-loop PLL bandwidth.

The \(dq\)- and \(a\beta\)-frame CCs (4) and (11) are equivalent under the design premises stated, concerning the current control loop only. However, they differ concerning their PLL impact [15]. For an \(a\beta\)-frame CC, the \(dq\)-frame reference \(i_{ref}\) is transformed into the \(a\beta\) frame as

\[
i_{ref}^i = e^{j\Delta \theta} i_{ref} = e^{j(\omega_1 t + \Delta \theta)} (i_0 + \Delta i_{ref}) \tag{28}
\]

where \(\Delta i_{ref}\) is the perturbation about the mean value \(i_0 = i_{d0} + j i_{q0}\). Since a constant \(i_{q0}'\) is assumed, \(\Delta i_{ref} = \Delta i_{ref}'\), where \(\Delta i_{ref}'\) is the perturbation impact from the DVC. Equation (28) can be linearized as

\[
i_{ref}^i \approx e^{j\omega_1 t} [(1 + j\Delta \theta) i_0 + \Delta i_{ref}] = e^{j\omega_1 t} (i_0 + \Delta i_{ref}') \tag{29}
\]

where

\[
\Delta i_{ref} = \Delta i_{ref}' + j G_p(s) i_0 \text{Im}\{\Delta E\}. \tag{30}
\]

The PLL thus acts as an added reference perturbation (in the \(q\)-direction only in case \(i_0\) is real). The closed-loop impact of (30) can be determined by considering (6) for the perturbation quantities, i.e.

\[
\Delta i = G_{ci}(s) \Delta i_{ref} + Y_i(s) \Delta E. \tag{31}
\]

Before proceeding to include the DVC, let us discuss the impact just of the PLL by assuming \(\Delta i_{ref}' = 0\). Since (30) includes \(\text{Im}\{\Delta E\}\), it is obvious that the model no longer is balanced [27]. Yet, a complex space-vector model can be employed using the identity \(\text{Im}\{\Delta E\} = (\Delta E - \Delta E^*)/(2j)\) in (30), which gives

\[
\Delta i = Y_+(s) \Delta E + Y_-(s) \Delta E^*. \tag{32}
\]

where

\[
Y_+(s) = Y_i(s) - Y_-(s), \quad Y_-(s) = -\frac{G_{ci}(s) G_p(s) i_0}{2}. \tag{33}
\]

Equation (32) shows that, if \(\Delta E\) contains just one frequency component, e.g., \(\Delta E = \Delta E^* e^{j\omega t}\), then—because of the imbalance—two components appear in \(\Delta i\), as

\[
\Delta i = Y_+(j\omega) \Delta E^* e^{j\omega t} + Y_-(j\omega) \Delta E^* e^{-j\omega t}. \tag{34}
\]

Component \(\Delta i_+\), which may be called an image [27], is negative sequence in the \(dq\) frame, but as long as \(\omega < \omega_1\), it is positive sequence in the \(a\beta\) frame; the components there appear as sideband components of +1, at \(\omega_1 \pm \omega\).
Both components, in turn, affect $\Delta E$ via the negative feedback described by (3). If $Z(s)$ has a resonance at, or close to, -$\omega_0$, then $\Delta i_+$ will be amplified, giving a large amplitude $|\Delta E|_+$ (initially exponentially growing if the resonance gets destabilized). Because the grid is assumed to be balanced, \[ |Z'(-j\omega)| = |Z'(j\omega)|, \] whereas $|Z(-j\omega)| \neq |Z(j\omega)|$ for $\omega \neq 0$. Consequently, $\Delta i_+$ is amplified much less by the resonance than $\Delta i_-$. The dominant component of $\Delta E$ is still $\Delta E_+ e^{j\omega t}$. This component may produce active power of nonzero mean with $\Delta i_+$, according to $\text{Re}\{\Delta E_+ e^{j\omega t} \Delta i_+\} = \text{Re}\{\text{Re}[Y_+(j\omega)]\Delta E_+\}$, whereas its interaction with $\Delta i_-$ just produces active-power pulsations of the angular frequency $2\omega_0$. This motivates neglecting the impact of $\Delta i_-$ for a stability analysis of the converter-grid interconnection for a balanced grid. Caution is advised, though, since in certain degenerated cases the assumptions may not hold. Moreover, for interaction with an imbalanced grid, which is the case, e.g., for subsynchronous torsional interaction [33] and analysis of multiple-converter systems, a multivariable model, as in [2], must be used.

Similar conclusions are drawn in [10], [15], and [17], though using somewhat different motivations and a different linearization method. In addition, the per-phase analysis is used, which per-phase analysis is used, which

\[ \frac{\text{d}}{\text{d}t} = \frac{\text{d}y}{\text{d}t} + \text{d}z, \]

...simplifies the passivity properties of $G_{\text{cl}}(s)$ given by (40) with the occurrences of resonance destabilization found by simulation in MATLAB. The grid impedance is selected as a parallel–series inductive–capacitive impedance

\[ Z'(s) = sL_g \left( sL_s + \frac{1}{sC_g} \right) = \frac{sL_g(s^2L_sC_s + 1)}{s^2(L_g + L_s)C_s + 1} \]

i.e., the angular resonant frequency in the $\alpha\beta$ frame is $\omega_{\text{res}} = 1/(L_g + L_s)C_g^{1/2}$. $L_g = L = 0.1$ p.u. and $L_s = 1$ p.u. are selected, whereas $C_g$ is varied in order to obtain the desired $\omega_{\text{res}}$, as explained in the following.

In the control system—which uses an $\alpha\beta$-frame CC—an R part is included only for $+1$, i.e., for $h = 0$ in (5), with $a_h = 0.5$ p.u. The PLL and the DVC have the parameters $a_p = a_d = 0.2$ p.u. and $a_{\text{ref}} = a_d = 0.05$ p.u., except where noted otherwise. The grid voltage is adjusted so as to give $E_0 = 1$ p.u., the reactive-power exchange is zero, i.e., $i_{\text{q}} = 0$, and the system is considered lossless. The last two assumptions yield $i_0 = P_i/E_0$.

1) Example 1: Inverter Operation: A power injection $P_i = -0.9$ p.u. into the dc link is considered. This scenario could account for, e.g., a photovoltaic inverter. To allow the PLL to quickly track variations in the PCC-voltage angle, $a_p = 2$ p.u. is used in this example (see [10] for a similar selection). The angular resonant frequency $\omega_{\text{res}}$ is adjusted...
Fig. 8. Inverter operation with $\omega_{\text{res}} = 1.38$ p.u. (a) Converter-current components. (b) Input-admittance real parts, where the dashed curve accounts for rectifier operation with $P_I = 0.9$ p.u. (c) DFT of the converter current. (d) DFT of the PCC voltage.

Fig. 9. Rectifier operation with $\omega_{\text{res}} = 1.34$ p.u. (a) Converter-current components. (b) Input-admittance real parts, where the dashed curve accounts for inverter operation with $P_I = -0.9$ p.u. (c) DFT of the converter current. (d) DFT of the PCC voltage.

Fig. 10. Input-admittance real parts for (a) $\alpha\beta$-frame CC and (b) $dq$-frame CC for (solid lines) $\alpha_p = \alpha_d = 0$ p.u. and (dashed lines) $\alpha_p = \alpha_d = 0.2$ p.u.

until marginally stable operation is reached, i.e., a constant-amplitude oscillation occurs, as shown in Fig. 8(a). This correlates well with Fig. 8(b), where the real part of $Y_+^s(s) = Y_+(s - j\omega_1)$ is shown: $\omega_{\text{res}} = 1.38$ p.u. is located near the upper boundary of the negative-real-part region. Fig. 8(c) shows the discrete Fourier transform (DFT) modulus of $i_s$ in a logarithmic scale. The components $|i_s^+|$ and $|i_s^-|$, located symmetrically about $\omega_1 = 1$ p.u., at $\omega_{\text{res}}$ and $2\omega_1 - \omega_{\text{res}}$, respectively, can be observed. The corresponding components in $E^s$ are shown in Fig. 8(d). It can be noted that the lower sideband component has roughly 30 dB smaller amplitude than the upper sideband component, and therefore can be neglected, as discussed in Section IV-A.

The PLL gives a negative impact for inverter operation [2], which can be deduced by the multiplication by $i_0$ of $G_p(s)$ in (40). The dashed curve in Fig. 8(b) shows that a much smaller negative-real-part region is obtained for rectifier operation with the same parameter values. For inverter operation, $\alpha_p$ should not be made larger than necessary to obtain acceptable dynamic performance, in order to minimize the negative-real-part region.

2) Example 2 (Rectifier Operation): A power draw of $P_I = 0.9$ p.u. from the dc link is now considered. This scenario could account for, e.g., a back-to-back ac motor drive. To give sufficiently small variations in $i_{dq}$ for variations that may occur in $P_I$ (though not included in the simulation), a larger $\alpha_d = 0.5$ p.u. is this time used. The results shown in Fig. 9 are very similar to those in Fig. 8, although the negative-real-part region this time is slightly smaller. Consequently, a slightly lower $\omega_{\text{res}}$ is needed to give marginally stable operation.

The DVC gives negative impact mainly for rectifier operation [2], which can be deduced by the multiplication by $-i_0^d$ of $G_d(s)$ in (40). The dashed curve in Fig. 9(b) shows that a much smaller negative-real-part region is obtained for inverter operation with the same parameter values. For rectifier operation, $\alpha_d$ should not be made larger than necessary to obtain acceptable dynamic performance, in order to minimize the negative-real-part region.

Remark 6: It is interesting to note in (40) that, for $G_p(s) = G_d(s)$ and a real $i_0$, terms $G_p(s)i_0$ and $-G_d(s)i_0^*$ cancel in (40). Thus, the identical selections of the PLL controller (17) and the DVC (22) remove the dependence of $i_{dq}$ from the input-admittance characteristics, giving the same passivity properties for both inverter and rectifier operations. For moderate gain selections, then very small negative-real-part regions are obtained, as shown in Fig. 10(a).

Remark 7: If a $dq$-frame CC is used, a widening of the negative-real-part region tends to result, as shown in Fig. 10(b). It is, therefore, generally preferable to use an $\alpha\beta$-frame CC.
V. CONCLUSION

An overview of methods for stability assessment based on the passivity properties of the VSC input admittance was presented. The modeling and analysis method in [10], [15], and [17] was generalized. Design recommendations for minimizing the negative-real-part regions were derived. These recommendations, which, in terms of passivity properties, clarify the key results of papers cited in the text, can be summarized as follows:

1) make the total time delay $T_d$ as small as possible. If it can be made equal just to the PWM time delay 0.5Ts, [40]–[42], then a positive real part of the inner input admittance is obtained up to the Nyquist frequency;

2) use proper selection of the R-part compensation angles $\phi_R$ according to (24), particularly if R parts are included for higher harmonic orders [4];

3) do not select the bandwidths of the outer loops, i.e., the PLL and the DVC, unnecessarily large, this particularly applies for the PLL in inverter operation and the DVC in rectifier operation [2], [11], [23];

4) use an $\alpha\beta$-frame CC to reduce the PLL impact [46].

REFERENCES


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