

Article

High-Frequency Oscillation Suppression Strategy for Interconnected Grid-Forming Energy Storage and Grid-Following HVDC Systems Based on Impedance Analysis

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Abstract

Interconnecting grid-following (GFL) and grid-forming (GFM) converters help stabilize weak grids with a high penetration of power electronics, but dynamic interactions can induce high-frequency oscillations (HFOs). The HFO mechanism within interconnected grid-forming battery energy storage (GFM-BESS) and grid-following VSC-HVDC systems remains unresolved, with viable suppression strategies yet to be established. To address this, this study utilizes harmonic linearization to derive small-signal frequency-domain impedance models for both devices. The models are developed for high-frequency-oscillation analysis: the GFM-BESS model retains all control loops, while the GFL-HVDC model retains the current loop and the phase-locked loop, with the slow outer loop omitted as justified within the high-frequency band. Frequency-coupled 2×2 impedance matrices are established to capture the frequency coupling introduced by the power loops and the phase-locked loop. Stability is established by the argument principle, the exact Nyquist encirclement count and the roots of the closed-loop characteristic function. At a nominal control delay of 50 μ s, the voltage-loop proportional gain has two finite stability boundaries: insufficient proportional action fails to damp the integral-loop mode, whereas excessive proportional action destabilizes the delay-affected filter mode. The resulting two-sided admissible interval is $0.0771 < k_{vp} < 0.5261$ p.u.; a robust engineering range of 0.15–0.30 p.u., with $k_{vp} = 0.22$ p.u. as the nominal value, is recommended. Frequency-sweep results validate the impedance models, while the independently evaluated root locus and Nyquist count give identical stability classifications. Time-domain gain-step tests further verify both sides of the interval: changing k_{vp} from 0.60 to 0.22 suppresses the HFO, whereas a further decrease of 0.15 p.u. to 0.07 re-excites it.



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1. Introduction

Driven by carbon-neutrality targets, wind and photovoltaic generation are expanding rapidly and progressively replacing fossil fuels [1]. As China's renewable resources and load centers are distributed inversely [2], long-distance, large-capacity transmission is required to optimize energy allocation [3]; voltage-source-converter high-voltage direct-current (VSC-HVDC) transmission is, therefore, widely used for offshore-wind integration and bulk power delivery. Under grid-following (GFL) control, however, VSC-HVDC

interacts adversely with weak AC grids and raises stability concerns. By contrast, the grid-forming battery energy storage system (GFM-BESS) provides inertia, supports renewable integration, and strengthens grids with high power-electronic penetration [4,5]. Interconnecting GFM-BESS with GFL-HVDC is, thus, a promising way to reinforce such grids.

This interconnection introduces many power-electronic devices whose control loops, interacting with one another or with the grid, can inject negative damping and excite wide-band oscillations from the sub-synchronous to the high-frequency band [6]. Such events are well documented: a sub-synchronous oscillation near 20 Hz on a series-compensated wind farm in Texas in 2009 [7]; high-frequency resonances in VSC-HVDC projects, such as the 2017 Luxi 1270 Hz event and the Yu-E commissioning resonances near 1800 Hz and 700 Hz [8,9]; and repeated high-frequency oscillations (HFOs) during no-load energization of the Zhangbei Kangbao station in 2020 [10]. Several of these events stemmed from control parameters that passed the initial design checks but proved sub-optimal for the as-built grid impedance; the Luxi event, for example, was traced to a poorly tuned voltage feed-forward loop [9]. Because a converter's impedance changes as the network is reconfigured, parameters that are sound in the design stage may produce negative damping after commissioning, and the resulting HFOs can damage equipment and trigger wide-area outages [11].

Wide-band-oscillation analysis is generally divided into time-domain and frequency-domain methods [12]. State-space (time-domain) analysis locates the system eigenvalues accurately, but it requires the manufacturers' confidential internal parameters and becomes intractable for large systems [13]. Frequency-domain (impedance) analysis relies only on the externally measurable terminal characteristic, which can be obtained without internal parameters and reduces the modeling effort; with its clear physical interpretation, it is widely used for the stability analysis of complex systems [14–17].

Existing impedance-based HFO studies fall into two groups. The first group examines interactions between grid-following devices, or between a grid-following VSC-HVDC and a weak AC grid [8,9]; here, both terminals are current-source-type, so no voltage-source subsystem participates in the resonance. The second group examines a single grid-forming converter tied to the grid or to a wind farm [18–21], where the grid-forming device faces a passive or current-source network rather than another actively controlled current-source converter. Moreover, grid-forming studies that retain the inner loops usually stop at the low- or sub-synchronous band [20,21], whereas VSC-HVDC HFO studies retain the current loop and PLL but omit the grid-forming voltage/power loops [8,9]. On the suppression side, the reported remedies are largely hardware- or add-on-based—additional damping branches, series/parallel filters, or feed-forward reshaping [9,10,18]—rather than a retuning of the existing controllers. Consequently, the case in which an actively controlled voltage-source converter (GFM-BESS) and an actively controlled current-source converter (GFL-HVDC) are interconnected and interact in the high-frequency band has received little attention, and its HFO mechanism and a hardware-free suppression route remain unclear. Table 1 summarizes the position of the present study relative to representative prior work along the four dimensions of converter configuration, retained control loops, stability assessment and suppression design.

To address this gap, this paper studies the high-frequency interaction of an interconnected GFM-BESS and GFL-HVDC system, and its specific contributions are as follows. (i) Converter configuration: unlike the current-source interconnected or single-grid-forming settings of prior work, an actively controlled voltage-source converter and an actively controlled current-source converter are interconnected, and the composite network is decomposed into a voltage-source subsystem and a current-source subsystem to expose the cross-subsystem interaction. (ii) Retained control loops: the GFM-BESS model retains the VSM power loops together with the voltage/current inner loops, feed-forward and control

delay, and the GFL-HVDC model retains the current loop, voltage feed-forward, control delay and PLL; because the study is confined to the high-frequency band, the slow GFL power loop is not modeled, and the resulting error is bounded in Section 2.3. (iii) Stability assessment: frequency-coupled 2×2 impedance matrices are established for both converters and the mirror-frequency coupling is quantified and shown to be negligible in the high-frequency band; on this basis, stability is determined rigorously through the argument principle, right-half-plane pole counts and exact Nyquist encirclements. (iv) Suppression design and guideline: the voltage-loop characteristic function is written in a form affine in k_{vp} , so every imaginary-axis crossing is obtained directly from the exact exponential-delay model. This gives both a lower and an upper gain boundary and, therefore, a reproducible two-sided design interval, rather than a monotonic “larger gain is better” rule. The resulting retuning strategy requires no added filter, damping branch or auxiliary control. The remainder of the paper is organized as follows: Section 2 establishes the impedance models; Section 3 analyses the cause of the high-frequency oscillation; Section 4 proposes the gain-selection guideline and validates the impedance models; and Section 5 concludes the paper.

Table 1. Position of the present study relative to representative impedance-based HFO work.

Ref.	Configuration	Retained Loops	Stability Assessment	Suppression
[8,9]	GFL-HVDC–AC grid	Current loop, PLL	SISO Nyquist	Added filter/ feed-forward
[18,19]	Single GFM–grid/ wind farm	Inner loops	Nyquist/eigenvalue	Damping branch
[20,21]	Single GFM–grid	Inner + outer loops	Eigenvalue (low freq.)	Parameter design
This work	GFM-BESS–GFL-HVDC (interconnected)	VSM power + inner loops, delay, PLL	RHP pole count + Nyquist encirclement	Proportional-gain retuning (hardware-free)

2. Frequency-Domain Impedance Models of the GFM-BESS and the GFL-HVDC for High-Frequency Analysis

Figure 1 shows a schematic diagram of the interconnected system that contains the GFM-BESS and the GFL-HVDC. To faithfully reproduce the distribution of components in a real power system, the GFM-BESS and the renewable source are connected to the sending end of the flexible-DC link, while a voltage source emulates the remaining equivalent voltage sources and impedances of the grid, such as synchronous machines. The GFM-BESS adopts virtual-synchronous-machine (VSM) control, and the grid-side converters of both the flexible-DC sending terminal and the renewable source adopt grid-following control.

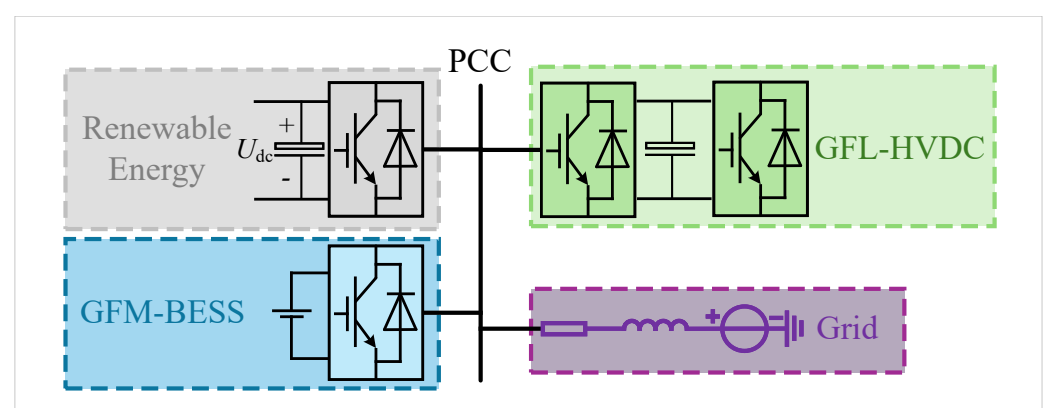


Figure 1. Schematic diagram of the interconnected GFM-BESS and GFL-HVDC system.

2.1. Frequency-Domain Impedance Model of the GFM-BESS

The complete control structure of the GFM-BESS is shown in Figure 2. In the figure, v_{abc} and i_{abc} are the three-phase voltage and current at the filter output of the GFM-BESS, e_{abc} and i_{abc} are the three-phase voltage and current at the GFM-BESS terminal, L_f and C_f are the filter inductance and capacitance, and L_g and R_g are the inductance and resistance of the connected grid. Following the harmonic-linearization approach, a small perturbation is applied to phase A, so the phase voltage and phase current are written as

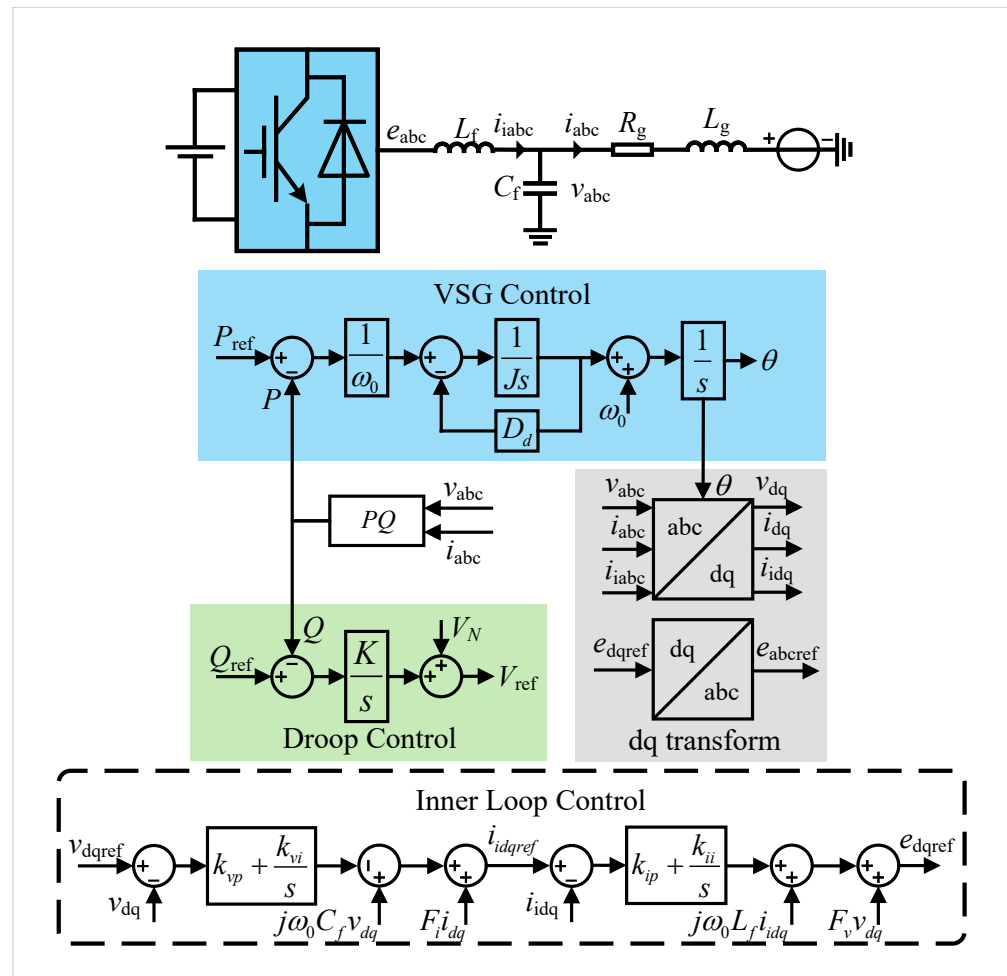


Figure 2. Control architecture of the GFM-BESS. All control-loop gains and the associated voltage/current variables in the inner-loop blocks are expressed in per-unit (p.u.), consistent with the values listed in Appendix A.

$$v_a(t) = V_1 \cos(\omega_1 t) + V_p \cos(\omega_p t + \varphi_{v,p}) \quad (1)$$

$$i_a(t) = I_1 \cos(\omega_1 t + \varphi_{i,1}) + I_p \cos(\omega_p t + \varphi_{i,p}) \quad (2)$$

where V_1 and V_p are the amplitudes of the fundamental and positive-sequence perturbation voltage, respectively; $\varphi_{v,p}$, $\varphi_{i,1}$ and $\varphi_{i,p}$ are the phase angles of the perturbation voltage, the fundamental current and the perturbation current, respectively; I_1 and I_p are the amplitudes of the fundamental and perturbation current; $\omega_1 = 2\pi f_1$ is the fundamental angular frequency; and $\omega_p = 2\pi f_p$ is the angular frequency of the perturbation. In the stationary $\alpha\beta$ frame, the three-phase voltage and current become

$$v_{\alpha\beta}(f) = \begin{cases} V_1, & f = f_1 \\ V_p e^{j\varphi_{v,p}}, & f = f_p \end{cases} \quad (3)$$

$$i_{\alpha\beta}(f) = \begin{cases} I_1 e^{j\varphi_{i,1}}, & f = f_1 \\ I_p e^{j\varphi_{i,p}}, & f = f_p \end{cases} \quad (4)$$

where the specific expressions for the $v_{\alpha\beta}$ and $i_{\alpha\beta}$ components are given by:

$$v_{\alpha}(f) = \begin{cases} 0.5V_1, & f = \pm f_1 \\ 0.5V_p e^{\pm j\varphi_{v,p}}, & f = \pm f_p \end{cases} \quad (5)$$

$$v_{\beta}(f) = \begin{cases} \mp 0.5jV_1, & f = \pm f_1 \\ \mp 0.5jV_p e^{\pm j\varphi_{v,p}}, & f = \pm f_p \end{cases} \quad (6)$$

$$i_{\alpha}(f) = \begin{cases} 0.5I_1 e^{\pm j\varphi_{i,1}}, & f = \pm f_1 \\ 0.5I_p e^{\pm j\varphi_{i,p}}, & f = \pm f_p \end{cases} \quad (7)$$

$$i_{\beta}(f) = \begin{cases} \mp 0.5jI_1 e^{\pm j\varphi_{i,1}}, & f = \pm f_1 \\ \mp 0.5jI_p e^{\pm j\varphi_{i,p}}, & f = \pm f_p \end{cases} \quad (8)$$

According to the instantaneous power theory, the output power of the GFM-BESS can be written as:

$$P_e = 1.5(v_{\alpha}i_{\alpha} + v_{\beta}i_{\beta}) \quad (9)$$

$$Q_e = 1.5(v_{\beta}i_{\alpha} - v_{\alpha}i_{\beta}) \quad (10)$$

Substituting (3) and (4) into (9) and (10) yields the frequency-domain expression of $P_e(f)$ and $Q_e(f)$ as:

$$P_e(f) = \begin{cases} 0.75(V_1 I_1 e^{j\varphi_{i,1}} + V_1 I_1 e^{-j\varphi_{i,1}}), & \text{dc} \\ 0.75(V_1 I_p e^{\pm j\varphi_{i,p}} + V_p I_1 e^{\pm j(\varphi_{v,p} - \varphi_{i,1})}), & f = \pm(f_p - f_1) \end{cases} \quad (11)$$

$$Q_e(f) = \begin{cases} 0.75j(V_1 I_1 e^{j\varphi_{i,1}} - V_1 I_1 e^{-j\varphi_{i,1}}), & \text{dc} \\ 0.75j(\pm V_1 I_p e^{\pm j\varphi_{i,p}} \mp V_p I_1 e^{\pm j(\varphi_{v,p} - \varphi_{i,1})}), & f = \pm(f_p - f_1) \end{cases} \quad (12)$$

The active-power loop of the GFM-BESS uses VSM control and the reactive-power loop uses droop control. Denoting the inertia and damping coefficients by J and D_d and the reactive-power-voltage droop coefficient by K , the perturbations produced in the phase angle and the amplitude of the voltage reference output by the power outer loop are

$$\Delta\theta(f) = -M(s) \Delta P_e(f) \quad (13)$$

$$\Delta V^*(f) = -N(s) \Delta Q_e(f) \quad (14)$$

where $M(s) = 1/(Js^2 + D_d s)$ and $N(s) = 1/(Ks)$ are the transfer functions of the active-power loop and the reactive-power loop, respectively. The output voltage reference calculated by the outer loop is fed into the inner loop. According to the inner-loop control block diagram of the GFM, the voltage output from the inner-loop controller to the converter terminal can be calculated as:

$$\left\{ \left[(V^* - v_{dq}) G_v(s) + F_i i_{dq} + j\omega_1 C_f v_{dq} - i_{dq} \right] G_i(s) + F_v v_{dq} + j\omega_1 L_f i_{dq} \right\} H_d(s) = e_{dq} \quad (15)$$

where

$$G_v(s) = k_{vp} + \frac{k_{vi}}{s} \quad (16)$$

$$G_i(s) = k_{ip} + \frac{k_{ii}}{s} \quad (17)$$

$$H_d(s) = e^{-sT_d} \quad (18)$$

where k_{vp} and k_{vi} are the proportional and integral gains of the voltage loop, respectively; k_{ip} and k_{ii} are the proportional and integral gains of the current loop, respectively; T_d is the control time delay; the subscript “dq” denotes the positive-sequence synchronous reference frame; and F_v and F_i represent the feed-forward coefficients of voltage and current, respectively. Since the dq and $\alpha\beta$ frames differ by the factor $e^{j\omega_1 t}$, multiplying (15) by $e^{j\omega_1 t}$ shifts every transfer-function argument from s to $s - j\omega_1$ (the constant decoupling and feed-forward coefficients are unchanged).

$$\left\{ \left[(v_{\alpha\beta}^* - v_{\alpha\beta}) G_v(s - j\omega_1) + F_i i_{\alpha\beta} + j\omega_1 C_f v_{\alpha\beta} - i_{\alpha\beta} \right] G_i(s - j\omega_1) + F_v v_{\alpha\beta} + j\omega_1 L_f i_{\alpha\beta} \right\} H_d(s - j\omega_1) = e_{\alpha\beta} \quad (19)$$

Formulating the KCL and KVL for the filter capacitor and inductor yields:

$$\begin{cases} i_{\alpha\beta} = i_{\alpha\beta} + sC_f v_{\alpha\beta} \\ v_{\alpha\beta} = e_{\alpha\beta} - sL_f i_{\alpha\beta} \end{cases} \quad (20)$$

Combining (3) and (20), the relationship among $v_{\alpha\beta}^*$, $v_{\alpha\beta}$, and $i_{\alpha\beta}$ is given by:

$$G_1(s) v_{\alpha\beta}^* + G_2(s) v_{\alpha\beta} + G_3(s) i_{\alpha\beta} = 0 \quad (21)$$

where

$$G_1(s) = G_v(s - j\omega_1) G_i(s - j\omega_1) H_d(s - j\omega_1) \quad (22)$$

$$G_2(s) = \left[-G_v(s - j\omega_1) G_i(s - j\omega_1) + F_v + j\omega_1 C_f G_i(s - j\omega_1) - sC_f G_i(s - j\omega_1) + j\omega_1 L_f C_f s \right] H_d(s - j\omega_1) - L_f C_f s^2 - 1 \quad (23)$$

$$G_3(s) = \left[(F_i - 1) G_i(s - j\omega_1) + j\omega_1 L_f \right] H_d(s - j\omega_1) - sL_f \quad (24)$$

Here, the terminal current $i_{\alpha\beta}$ is defined as positive when flowing *out of* the converter, consistent with (20), so the GFM-BESS is represented as a Thévenin source and its output impedance is $Z_{\text{GFM}} = -v_{\alpha\beta}/i_{\alpha\beta}$; the corresponding sign is retained explicitly in (30). In the stationary reference frame, the voltage reference vector $v_{\alpha\beta}^*$ can be expressed as:

$$v_{\alpha\beta}^* = V^* e^{j\theta} = e^{j\omega_1 t} (V_1 + V_1 j\Delta\theta + \Delta V^*) \quad (25)$$

Combining (13), (14) and (25), in the frequency domain, $v_{\alpha\beta}^*$ can be expressed as:

$$v_{\alpha\beta}^*(f) \Big|_{f=f_p} = G_4(s) V_p e^{j\varphi_{v,p}} + G_5(s) I_p e^{j\varphi_{i,p}} \Big|_{s=j\omega_p} \quad (26)$$

where

$$G_4(s) = 0.75jN(s - j\omega_1) I_1 e^{-j\varphi_{i,1}} - 0.75jV_1 M(s - j\omega_1) I_1 e^{-j\varphi_{i,1}} \quad (27)$$

$$G_5(s) = -0.75jN(s - j\omega_1) V_1 - 0.75jV_1 M(s - j\omega_1) V_1 \quad (28)$$

By substituting (26) into (21), the relationship between the voltage and current perturbation signals is obtained as:

$$[G_1(s)G_4(s) + G_2(s)]V_p e^{j\varphi_{v,p}} + [G_1(s)G_5(s) + G_3(s)]I_p e^{j\varphi_{i,p}} \Big|_{s=j\omega_p} = 0 \quad (29)$$

Therefore, considering both the inner and outer loops, the positive-sequence frequency-domain impedance of the GFM is given by:

$$Z_{\text{GFM}}(s) = \frac{G_1(s)G_5(s) + G_3(s)}{G_1(s)G_4(s) + G_2(s)} \quad (30)$$

In (30), G_1 – G_3 are set by the inner-loop PI gains, feed-forward coefficients, filter parameters and control delay, while G_4 , G_5 are set by the power-loop functions $M(s)$, $N(s)$ and the operating point. Retaining loops of different bandwidths, the model is valid for small-signal stability analysis over a wide frequency range.

Equation (30) is a single scalar transfer function relating the positive-sequence perturbation voltage to the positive-sequence perturbation current. Strictly speaking, however, the active- and reactive-power loops of the VSM process the instantaneous powers P_e and Q_e , which are *real* scalar quantities in the synchronous frame. A real signal has a conjugate-symmetric spectrum, so the response to a positive-sequence perturbation at f_p is necessarily accompanied by a mirror-frequency component at $f_n = f_p - 2f_1$ [22,23]. This frequency coupling is neglected in (30); whether that omission is admissible is examined in Section 2.3, after the GFL-HVDC model is derived.

2.2. Frequency-Domain Impedance Model of the GFL-HVDC

The sending-end converter of the VSC-HVDC system adopts a grid-following (GFL) control structure, as shown in Figure 3. On the DC side, the sending-end converter is regulated by a DC-link-voltage controller (DVC) whose closed-loop bandwidth is of the order of tens of hertz, comparable to that of the power loop, while the GFM-BESS is backed by a battery, which acts as a stiff DC source. The DC-link voltage is, therefore, treated as a constant, ideal source in the small-signal model. This simplification is admissible for the high-frequency band studied here: any DVC-induced impedance term is shaped by the DVC loop gain and rolls off as $f_{\text{bw}}^{\text{DVC}} / f_p$, so at the 2–5 kHz oscillation frequencies, it is more than two orders of magnitude below the current-loop impedance and does not reshape the high-frequency characteristic. The DVC/AC-current coupling introduced by the DC dynamics is confined to the sub- and medium-frequency range and lies outside the HFO band of interest. Neglecting the influence of the phase-locked loop (PLL) and assuming $\theta_{\text{PLL}} = 2\pi f_1 t$, the following expression can be obtained in the stationary reference frame based on the control block diagram:

$$\begin{cases} [-(i_m^* - i_m)G_{mi}(s - j\omega_1) - j\omega_1 L_m i_m]H_{md}(s - j\omega_1) + F_{mv}v_m H_{md}(s - j\omega_1) = e_m^* \\ e_m^* + sL_m i_m = v_m \end{cases} \quad (31)$$

where i_m^* denotes the current reference input to the inner loop; L_m represents the filter inductance of the GFL converter; F_{mv} is the voltage feed-forward coefficient; $G_{mi}(s)$ signifies the transfer function of the current loop proportional–integral (PI) controller; and $H_{md}(s)$ stands for the delay function, which reflects the total control loop latency introduced by

links such as sampling, computation, and communication in the digital control process. Specifically, $G_{mi}(s)$ and $H_{md}(s)$ can be expressed as:

$$G_{mi}(s) = k_{mip} + \frac{k_{mii}}{s} \quad (32)$$

$$H_{md}(s) = e^{-sT_{md}} \quad (33)$$

where k_{mip} and k_{mii} are the proportional and integral gains of the current loop, respectively, and T_{md} is the control time delay. According to (31), the frequency-domain impedance of the GFL converter without considering the PLL influence is obtained as:

$$Z_{\text{GFL}}(s) = \frac{v_m}{i_m} = \frac{[G_{mi}(s - j\omega_1) - j\omega_1 L_m] H_{md}(s - j\omega_1) + s L_m}{1 - F_{mv} H_{md}(s - j\omega_1)} \quad (34)$$

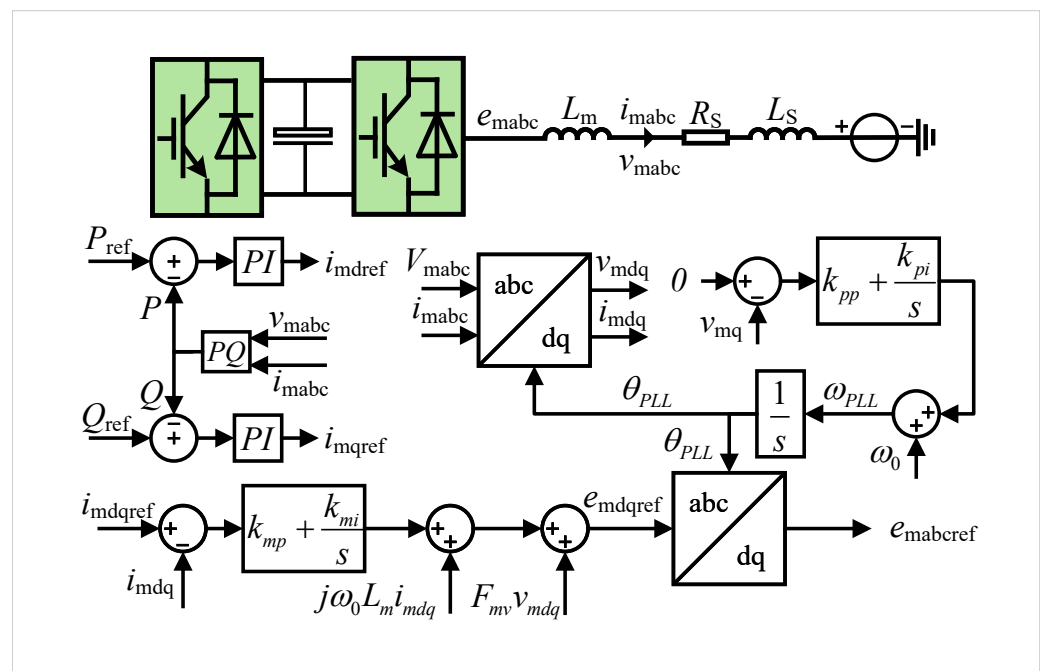


Figure 3. Control structure of the GFL-HVDC. As in Figure 2, all control-loop gains and inner-loop variables are expressed in per-unit (p.u.), consistent with Appendix A.

At this point, (34) does not include the transfer functions related to the phase-locked loop (PLL). However, when a small-signal perturbation occurs in the terminal voltage, the phase output of the PLL is also affected; thus, the influence of the PLL should be taken into account. Injecting a positive-sequence small-signal perturbation at frequency f_p at the terminal of the GFL converter yields:

$$v_{ma}(t) = V_{m1} \cos(\omega_1 t) + V_{mp} \cos(\omega_p t + \varphi_{mv,p}) \quad (35)$$

$$i_{ma}(t) = I_{m1} \cos(\omega_1 t + \varphi_{mi.1}) + I_{mp} \cos(\omega_p t + \varphi_{mi.p}) \quad (36)$$

where V_{m1} and V_{mp} are the amplitudes of the fundamental voltage and positive-sequence perturbation voltage, respectively; I_{m1} and I_{mp} are the amplitudes of the fundamental current and positive-sequence perturbation current, respectively; $\varphi_{mv,p}$ is the phase angle of the positive-sequence perturbation voltage; and $\varphi_{mi,1}$ and $\varphi_{mi,p}$ are the phase angles of the fundamental and positive-sequence perturbation currents, respectively. Transforming (35) and (36) into the stationary reference frame yields:

$$v_{m\alpha\beta}(f) = \begin{cases} V_{m1}, & f = f_1 \\ V_{mp} e^{j\varphi_{mv,p}}, & f = f_p \end{cases} \quad (37)$$

$$i_{m\alpha\beta}(f) = \begin{cases} I_{m1} e^{j\varphi_{mi,1}}, & f = f_1 \\ I_{mp} e^{j\varphi_{mi,p}}, & f = f_p \end{cases} \quad (38)$$

According to Figure 3, the phase angle output of the PLL will also generate a corresponding perturbation $\Delta\theta_{\text{PLL}}$, given by:

$$\theta_{\text{PLL}} = \theta_1 + \Delta\theta_{\text{PLL}} = 2\pi f_1 t + \Delta\theta_{\text{PLL}} \quad (39)$$

The Park transformation $\mathbf{T}(\theta_{\text{PLL}})$ can be expressed as:

$$\mathbf{T}(\theta_{\text{PLL}}) = \begin{bmatrix} \cos(\Delta\theta_{\text{PLL}}) & \sin(\Delta\theta_{\text{PLL}}) & 0 \\ -\sin(\Delta\theta_{\text{PLL}}) & \cos(\Delta\theta_{\text{PLL}}) & 0 \\ 0 & 0 & 1 \end{bmatrix} \mathbf{T}(\theta_1) \quad (40)$$

Applying the Park transformation $\mathbf{T}(\theta_1)$ to the three-phase voltages:

$$\begin{bmatrix} v_{md1} \\ v_{mq1} \end{bmatrix} = \mathbf{T}(\theta_1) \begin{bmatrix} v_{ma} \\ v_{mb} \\ v_{mc} \end{bmatrix} \quad (41)$$

Similarly, applying the Park transformation $\mathbf{T}(\theta_{\text{PLL}})$ to the three-phase voltages:

$$\begin{bmatrix} v_{md} \\ v_{mq} \end{bmatrix} = \mathbf{T}(\theta_{\text{PLL}}) \begin{bmatrix} v_{ma} \\ v_{mb} \\ v_{mc} \end{bmatrix} \quad (42)$$

According to the operating principle of the PLL, the relationship between V_{mq} and V_{mq1} satisfies the following expression:

$$v_{mq} = -\Delta\theta_{\text{PLL}} v_{md1} + v_{mq1} \quad (43)$$

The perturbation on the output angle of the PLL is:

$$\Delta\theta_{\text{PLL}}(f) = G_{mp}(s) V_{mp} e^{\pm j\varphi_{mv,p}} / 2, \quad f = \pm(f_p - f_1). \quad (44)$$

where $G_{mp}(s)$ is determined by subsequent equations. By combining the above equations:

$$v_{mq}(f) = [-G_{mp}(s) V_{m1} \mp j] V_{mp} e^{\pm j\varphi_{mv,p}} / 2, \quad f = \pm(f_p - f_1). \quad (45)$$

According to the control structure of the PLL, it holds that:

$$\Delta\theta_{\text{PLL}}(f) = H_{\text{PLL}}(s) v_{mq}(f) \quad (46)$$

where:

$$H_{\text{PLL}}(s) = \frac{1}{s} \left(k_{pp} + \frac{k_{pi}}{s} \right) \quad (47)$$

k_{pp} and k_{pi} denote the proportional and integral gains of the PLL, respectively. Combining (44), (45), and (46), the following can be obtained:

$$G_{mp}(s) = \frac{\mp j H_{PLL}(s)}{1 + V_{m1} H_{PLL}(s)} = \mp j G_{PLL}(s) \quad (48)$$

where:

$$G_{PLL}(s) = \frac{H_{PLL}(s)}{1 + V_{m1} H_{PLL}(s)} \quad (49)$$

When a small-signal perturbation occurs, its dynamics are governed by the current loop and the PLL, while the GFL power loop only sets the steady-state reference. Its closed-loop bandwidth $f_{bw} \approx k_{mpp}/2\pi$ is a few hertz (Appendix A), far below the HFO band, so the current reference $i_{mdq}^* = I_d + jI_q$ is treated as constant, with the resulting error bounded in Section 2.3. Transforming into the stationary frame yields:

$$i_{m\alpha\beta}^* = i_{mdq}^* e^{j\theta_{PLL}} = i_{mdq}^* e^{j(\theta_1 + \Delta\theta_{PLL})} = i_{mdq}^* e^{j\omega_1 t} (1 + j\Delta\theta_{PLL}) \quad (50)$$

Substituting (44) into (50) yields:

$$i_{m\alpha\beta}^*(f) = i_{mdq}^* G_{PLL}(s - j\omega_1) V_{mp} e^{j\varphi_{mv,p}/2} \Big|_{s=j\omega_p} \quad (51)$$

Transforming the variables from the dq reference frame to the $\alpha\beta$ reference frame, i.e., convolving both sides with $e^{j\theta_{PLL}}[f]$, yields:

$$\left[-\left(i_{m\alpha\beta}^* - i_{m\alpha\beta} \right) G_{mi}(s - j\omega_1) - j\omega_1 L_m i_{m\alpha\beta} \right] H_{md}(s - j\omega_1) + F_{mv} v_{m\alpha\beta} H_{md}(s - j\omega_1) = e_{m\alpha\beta}^* \quad (52)$$

By combining the above equations, the positive-sequence impedance of the GFL-controlled VSC-HVDC system considering the PLL influence is calculated as:

$$\begin{aligned} Z_{GFL}(s) &= \frac{v_{mp}(s)}{i_{mp}(s)} = \frac{G_{m1}(s) + G_{m2}(s)}{G_{m3}(s) + G_{m4}(s)} \\ &= \frac{[G_{mi}(s - j\omega_1) - j\omega_1 L_m] H_{md}(s - j\omega_1) + s L_m}{1 - F_{mv} H_{md}(s - j\omega_1) + \frac{1}{2} i_{mdq}^* G_{PLL}(s - j\omega_1) G_{mi}(s - j\omega_1) H_{md}(s - j\omega_1)} \end{aligned} \quad (53)$$

where:

$$G_{m1}(s) = [G_{mi}(s - j\omega_1) - j\omega_1 L_m] H_{md}(s - j\omega_1) \quad (54)$$

$$G_{m2}(s) = s L_m \quad (55)$$

$$G_{m3}(s) = 1 - F_{mv} H_{md}(s - j\omega_1) \quad (56)$$

$$G_{m4}(s) = \frac{1}{2} i_{mdq}^* G_{PLL}(s - j\omega_1) G_{mi}(s - j\omega_1) H_{md}(s - j\omega_1) \quad (57)$$

The GFL output impedance, thus, depends on the current-loop PI and feed-forward gains, the PLL PI gains, the filter inductance and the control delay.

Analogous to the GFM-BESS, the GFL current reference i_{mdq}^* is rotated into the stationary frame through the *real* PLL angle $\Delta\theta_{PLL}$; the resulting mirror-frequency term, dropped in the scalar impedance (53), is quantified in the next subsection.

2.3. Frequency Coupling and the Validity of the Scalar Model

The scalar impedances (30) and (53) are obtained by retaining only the positive-sequence line at f_p . To assess the error introduced by this simplification, the frequency-

coupled models of both converters are established below and the coupling strength is quantified across the frequency range of interest.

2.3.1. Origin of the Frequency Coupling

When only a positive-sequence voltage perturbation is injected, all inner-loop operators—the voltage/current PI regulators, the delay $H_d(s)$, the feed-forward terms and the $j\omega_1$ decoupling terms—act as linear time-invariant complex-vector operators and, therefore, map the positive-sequence line onto itself without transferring energy to any other line. Coupling appears only through control quantities that are *real* scalars, namely the instantaneous powers P_e, Q_e and the phase/amplitude corrections $\Delta\theta, \Delta V^*$ in the GFM-BESS, and the PLL angle $\Delta\theta_{\text{PLL}}$ in the GFL-HVDC [22,23]. A real signal in the synchronous frame has a conjugate-symmetric spectrum, so a component at $+(f_p - f_1)$ is inevitably paired with one at $-(f_p - f_1)$; after re-modulation by $e^{j\omega_1 t}$ the latter falls on the mirror line $f_n = f_p - 2f_1$. Following the modified sequence-domain definition of [22], the mirror (negative-sequence) variables are represented by their conjugates $\tilde{V}_n = V_n e^{j\varphi_{v,n}}$ and $\tilde{I}_n = I_n e^{j\varphi_{i,n}}$, so the terminal behavior of each converter is described by a linear 2×2 impedance matrix

$$\begin{bmatrix} \hat{V}_p \\ \tilde{V}_n \end{bmatrix} = \mathbf{Z}(s) \begin{bmatrix} \hat{I}_p \\ \tilde{I}_n \end{bmatrix}, \quad \mathbf{Z}(s) = \begin{bmatrix} Z_{pp}(s) & Z_{pn}(s) \\ Z_{np}(s) & Z_{nn}(s) \end{bmatrix}, \quad (58)$$

where $\hat{V}_p = V_p e^{j\varphi_{v,p}}$, $\hat{I}_p = I_p e^{j\varphi_{i,p}}$, and the off-diagonal terms Z_{pn}, Z_{np} quantify the frequency coupling.

2.3.2. Frequency-Coupled Model of the GFM-BESS

Retaining the mirror line, the full linearization of P_e and Q_e at $f = +(f_p - f_1)$ introduces the mixed quantity

$$A = V_1 \hat{I}_p + I_1 e^{-j\varphi_{i,1}} \hat{V}_p + V_1 \tilde{I}_n + I_1 e^{j\varphi_{i,1}} \tilde{V}_n, \quad (59)$$

in which the positive- and mirror-sequence unknowns appear together because the power convolution pairs the fundamental line with both perturbation lines. Consequently, the outer-loop voltage reference on the positive-sequence line becomes

$$v_{\alpha\beta}^* \Big|_{f_p} = G_4 \hat{V}_p + G_5 \hat{I}_p + G_6 \tilde{V}_n + G_7 \tilde{I}_n, \quad (60)$$

where G_4, G_5 are as in (27) and (28) and the two additional coupling functions are

$$G_6(s) = -0.75j[N(s - j\omega_1) + V_1 M(s - j\omega_1)]I_1 e^{j\varphi_{i,1}}, \quad (61)$$

$$G_7(s) = 0.75jV_1[N(s - j\omega_1) - V_1 M(s - j\omega_1)]. \quad (62)$$

Writing (21) on both the positive-sequence line ($s = j\omega_p$) and the mirror line ($s = j(2\omega_1 - \omega_p)$) and collecting the four unknowns yields

$$\mathbf{A}_{\text{GFM}} \begin{bmatrix} \hat{V}_p \\ \tilde{V}_n \end{bmatrix} + \mathbf{B}_{\text{GFM}} \begin{bmatrix} \hat{I}_p \\ \tilde{I}_n \end{bmatrix} = \mathbf{0}, \quad \mathbf{Z}_{\text{GFM}} = -\mathbf{A}_{\text{GFM}}^{-1} \mathbf{B}_{\text{GFM}}, \quad (63)$$

$$\mathbf{A}_{\text{GFM}} = \begin{bmatrix} G_1 G_4 + G_2 & G_1 G_6 \\ \bar{G}_1 \bar{G}_6 & \bar{G}_1 \bar{G}_4 + \bar{G}_2 \end{bmatrix}, \quad \mathbf{B}_{\text{GFM}} = \begin{bmatrix} G_1 G_5 + G_3 & G_1 G_7 \\ \bar{G}_1 \bar{G}_7 & \bar{G}_1 \bar{G}_5 + \bar{G}_3 \end{bmatrix}, \quad (64)$$

where $\bar{G}(s)$ denotes the mirror-line counterpart of $G(s)$, obtained by shifting every bare s to $s - 2j\omega_1$ (which leaves the $s - j\omega_1$ arguments of the control blocks unchanged) and conjugating the explicit imaginary constants ($\omega_1 \rightarrow -\omega_1$, $\varphi_{i,1} \rightarrow -\varphi_{i,1}$). When the coupling functions vanish ($G_6 = G_7 = 0$), $\mathbf{A}_{\text{GFM}}$ and $\mathbf{B}_{\text{GFM}}$ become diagonal and the leading diagonal element reduces exactly to the scalar impedance $Z_{pp} = -(G_1 G_5 + G_3)/(G_1 G_4 + G_2)$ of (30).

2.3.3. Frequency-Coupled Model of the GFL-HVDC

For the GFL-HVDC, the only coupling path is the real PLL angle acting on the constant current reference. The complete linearization gives $\Delta\theta_{\text{PLL}}|_{+(f_p-f_1)} = -\frac{j}{2}G_{\text{PLL}}(s - j\omega_1)(\hat{V}_p - \hat{V}_n)$, so the injected current reference on the two lines depends on both $\hat{V}_p$ and $\hat{V}_n$. Reusing the transfer functions G_{m1} – G_{m4} , the matrix model is

$$\mathbf{Z}_{\text{GFL}} = \mathbf{A}_{\text{GFL}}^{-1} \mathbf{B}_{\text{GFL}},$$

$$\mathbf{A}_{\text{GFL}} = \begin{bmatrix} G_{m3} + G_{m4} & -G_{m4} \\ -\bar{G}_{m4} & \bar{G}_{m3} + \bar{G}_{m4} \end{bmatrix}, \quad \mathbf{B}_{\text{GFL}} = \begin{bmatrix} G_{m1} + G_{m2} & 0 \\ 0 & \bar{G}_{m1} + \bar{G}_{m2} \end{bmatrix}. \quad (65)$$

Because $\mathbf{B}_{\text{GFL}}$ is diagonal, the normalized coupling admits the closed form

$$\frac{Z_{pn}}{Z_{pp}} = \frac{G_{m4}(s)}{\bar{G}_{m3}(s) + \bar{G}_{m4}(s)} \cdot \frac{\bar{G}_{m1}(s) + \bar{G}_{m2}(s)}{G_{m1}(s) + G_{m2}(s)}, \quad (66)$$

and since $G_{m4} \propto G_{\text{PLL}}(s - j\omega_1)$ rolls off at -20 dB/dec, the coupling decreases monotonically with frequency. Setting G_{m4} 's off-diagonal contribution to zero recovers the scalar impedance (53).

2.3.4. Quantification of the Coupling

To measure the coupling strength, the ratio

$$\kappa(\omega) = \max \left\{ \frac{|Z_{pn}(j\omega)|}{|Z_{pp}(j\omega)|}, \frac{|Z_{np}(j\omega)|}{|Z_{nn}(j\omega)|} \right\} \quad (67)$$

is evaluated with the parameters of Appendix A, as listed in Table 2. Near the fundamental frequency, the coupling is significant—for the GFL-HVDC, κ exceeds 0 dB below 200 Hz—which is consistent with the mirror-frequency behavior reported for VSMs [23]; in this low-frequency band, a 2×2 model and the generalized Nyquist criterion (GNC) are indeed required. In the high-frequency band, however, κ falls rapidly: above 500 Hz, it is below -23 dB for both converters, and at 2 kHz, which covers the upper gain-boundary mode identified in Section 3.3, it is no greater than -49.3 dB. The off-diagonal terms are, therefore, over two orders of magnitude smaller than the diagonal terms in the HFO band.

Table 2. Frequency-coupling ratio $\kappa(\omega)$ of the two converters.

f (Hz)	100	200	500	1000	2000	3349
GFM-BESS κ (dB)	−21.0	−47.1	−73.1	−88.0	−99.7	−110.4
GFL-HVDC κ (dB)	2.0	2.3	−23.4	−36.9	−49.3	−58.2

The same reasoning bounds the second approximation in the GFL-HVDC model—the constant-current-reference assumption of (53). The power loop rolls off above $f_{\text{bw}} \approx k_{\text{mpp}}/2\pi$, so the perturbation it imposes on the current reference scales as f_{bw}/f_p ; with $k_{\text{mpp}} = 0.2$ p.u., this is below 10^{-3} throughout the 0.7–2.0 kHz boundary-frequency range, comparable to or smaller than κ . The power-loop gains in Appendix A, therefore, fix the operating point but do not reshape the high-frequency impedance.

The consequence for stability follows from a first-order argument: when $\kappa \ll 1$, the eigenvalues of the loop-gain matrix $\mathbf{L} = \mathbf{Z}_{\text{sys1}} \mathbf{Z}_{\text{sys2}}^{-1}$ deviate from the diagonal (scalar) ratios only by a term of order κ . Since $\kappa \leq -49.3$ dB around the upper boundary mode, the generalized Nyquist eigenloci are accurately represented there by the scalar ratio $Z_{\text{sys1}} / Z_{\text{sys2}}$. The scalar impedances (30) and (53) are, therefore, used below, with the caveat that these simplifications hold only in the HFO band, not near the fundamental.

3. Mechanism Analysis of the High-Frequency Oscillation

The GFL-HVDC is equivalent to a current source in parallel with its impedance, and the GFM-BESS is equivalent to a voltage source in series with its wide-band impedance. Accordingly, from Figure 1, the impedance structure of the system is obtained as shown in Figure 4.

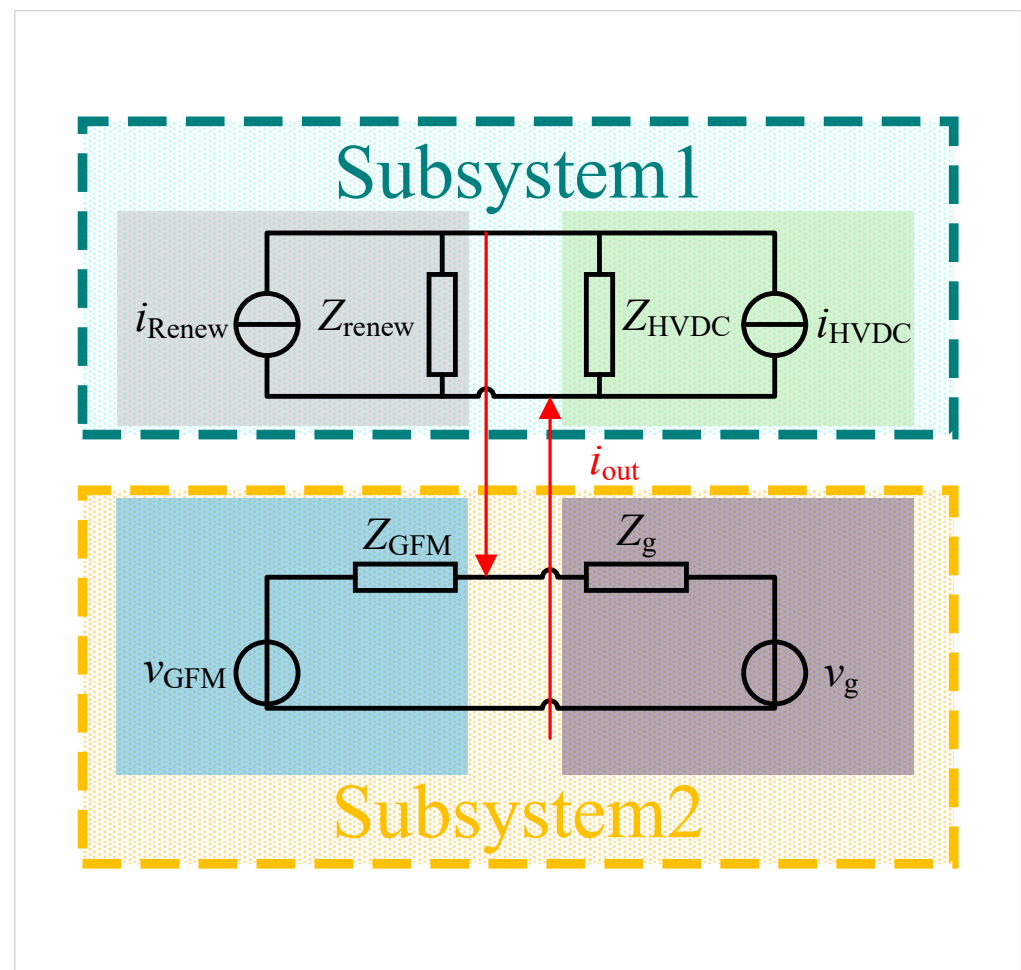


Figure 4. Schematic diagram of the impedance structure of the interconnected GFM-BESS and GFL-HVDC system.

Let the current source of the renewable plant be i_s with impedance Z_s , the current source of the VSC-HVDC be i_{HVDC} with impedance Z_{HVDC} , the voltage source of the GFM-BESS be v_{GFM} with impedance Z_{GFM} , and the voltage source of the grid be v_g with impedance Z_g . The GFM-BESS and the grid are combined into the voltage-source subsystem 1, and the renewable plant and the flexible-DC link are combined into the current-source subsystem 2. The equivalent impedance of subsystems can be expressed as

$$\begin{cases} Z_{\text{sys1}} = \frac{Z_{\text{GFM}}Z_g}{Z_{\text{GFM}} + Z_g} \\ Z_{\text{sys2}} = \frac{Z_s Z_{\text{HVDC}}}{Z_s + Z_{\text{HVDC}}} \end{cases} \quad (68)$$

From Figure 4, the output current i_{out} at the connection point of the subsystems is

$$i_{\text{out}} = \left(i_s + i_{\text{HVDC}} - \frac{v_g}{Z_{\text{sys2}}} + \frac{v_{\text{GFM}} - v_g}{(1 + Z_{\text{GFM}}/Z_g)Z_{\text{sys2}}} \right) \frac{1}{1 + Z_{\text{sys1}}/Z_{\text{sys2}}} \quad (69)$$

From (69), the system is stable if all poles of i_{out} lie in the left half-plane, i.e., the poles of i_{sys2} , v_{sys1} and $1/Z_{\text{sys2}}$ lie there and $Z_{\text{sys1}}/Z_{\text{sys2}}$ satisfies the Nyquist criterion. If the renewable plant, the flexible-DC link and the GFM-BESS are each stable stand-alone, the first three are automatic and v_{sys1} reduces to a Nyquist condition on Z_{GFM}/Z_g . Hence, the interconnected system is stable if and only if:

- (1) The renewable plant, the flexible-DC link and the GFM-BESS are each stable in stand-alone grid-connected operation;
- (2) The impedance ratio Z_{GFM}/Z_g satisfies the Nyquist stability criterion;
- (3) The impedance ratio $Z_{\text{sys1}}/Z_{\text{sys2}}$ satisfies the Nyquist stability criterion.

Conditions (2) and (3) must be applied in their full form rather than through a magnitude-crossing test alone. Let $L(s) = Z_{\text{sys1}}(s)/Z_{\text{sys2}}(s)$ be the minor-loop gain of Condition (3). By the Nyquist criterion, the number of closed-loop right-half-plane (RHP) poles is

$$Z = N + P, \quad (70)$$

where P is the number of RHP poles of $L(s)$ and N is the number of clockwise encirclements of the critical point $(-1, j0)$ by the contour of $L(j\omega)$ as ω traverses $(-\infty, \infty)$. Stability requires $Z = 0$, i.e., $N = -P$. Under Condition (1), each converter is stable in stand-alone operation, so Z_{sys1} and Z_{sys2} contain no RHP poles and $P = 0$; the closed-loop RHP-pole count then reduces to $Z = N$, and instability is certified whenever the Nyquist locus of L encircles $(-1, j0)$. A magnitude crossing with a phase difference exceeding 180° is only a symptom of such an encirclement and does not by itself certify instability. The pole count, the encirclement number and the resulting root locus are established quantitatively in Section 3.3 for both stable and unstable parameter sets.

Using the impedance models from Section 2 together with the parameters in Appendix A (Table A1), the high-frequency behavior of the interconnected system is examined below. Because Condition (3) governs the cross-subsystem interaction, the way in which each converter's control parameters reshape its high-frequency impedance is analyzed first, so as to identify the parameters that actually drive the HFO.

3.1. Sensitivity Analysis of the GFL-HVDC Control Parameters

Figure 5 shows how the GFL-HVDC impedance varies with its current-loop gains. As the proportional gain k_{mip} increases (Figure 5a), a resonant point and an abrupt phase jump appear in the medium-to-high-frequency band, reshaping the impedance where it interacts with the voltage-source subsystem. The integral gain k_{mii} (Figure 5b) has almost no effect, since k_{mii}/s vanishes at high frequency. Thus, k_{mip} is the key high-frequency parameter; whether a given reshaping stabilizes the interconnection is decided by the closed-loop assessment of Section 3.3.

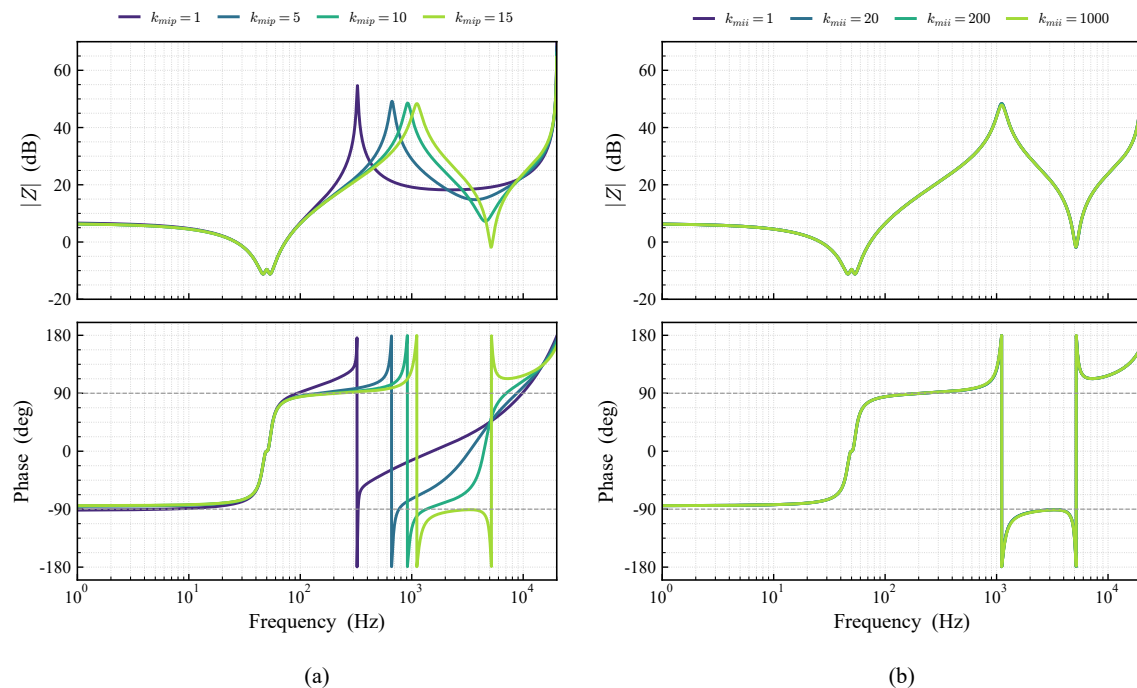


Figure 5. Impedance–frequency characteristics of the GFL-HVDC as the current-loop gains vary: (a) proportional gain k_{mip} ; (b) integral gain k_{mii} .

3.2. Sensitivity Analysis of the GFM-BESS Control Parameters

The GFM-BESS impedance is shaped by all four inner-loop gains, but the proportional voltage-loop gain k_{vp} is the parameter that directly moves the critical oscillatory branches when the remaining parameters are fixed. Increasing k_{vp} initially supplies damping to the mode created by the integral action; after a finite optimum, however, the higher voltage-loop crossover exposes the filter dynamics and the control delay, so the damping decreases and eventually becomes negative. Consequently, the dependence is non-monotonic and cannot be inferred from an impedance-magnitude reduction alone. The integral gain k_{vi} is retained at its Appendix A value, and the exact two-sided stability boundary in k_{vp} is derived below.

In summary, proportional-gain sensitivity is only a preliminary impedance observation. A valid gain recommendation must additionally establish internal stability of the GFM-BESS and satisfy the Nyquist condition of the interconnected system. The next subsection performs these checks through an exact imaginary-axis-crossing calculation, a root locus and the argument principle.

3.3. Rigorous Stability Assessment and Root-Locus Analysis

The impedance criterion must be applied only after the internal stability of each sub-system has been established. This prerequisite is particularly important when a controller gain is varied, because a Nyquist locus of the interconnection may appear benign even though the stand-alone converter has already acquired RHP poles. The assessment is, therefore, performed in two stages. First, the exact GFM-BESS voltage-loop characteristic function is used to determine every admissible value of k_{vp} . Second, the impedance-ratio Nyquist condition is applied over the internally stable interval.

For consistency between the physical filter parameters and the per-unit controller variables, define the base impedance and the dynamic per-unit filter coefficients as

$$Z_b = \frac{V_n^2}{P_n} = 2.304 \, \Omega, \quad \ell_f = \frac{L_f}{Z_b} = 8.6806 \times 10^{-4} \, \text{s}, \quad c_f = C_f Z_b = 2.3040 \times 10^{-5} \, \text{s}. \quad (71)$$

These coefficients follow directly from substituting $v = V_b v_{pu}$ and $i = I_b i_{pu}$ into the physical capacitor and inductor equations. They are, therefore, required whenever the controller gains and feedback variables are expressed in per unit. In particular, substituting L_f and C_f in SI units directly into a per-unit controller equation would mix two unit systems and artificially change the numerical scale of k_{vp} . Let

$$q = s - j\omega_1, \quad G_v(q) = k_{vp} + \frac{k_{vi}}{q}, \quad G_i(q) = k_{ip} + \frac{k_{ii}}{q}, \quad H_d(q) = e^{-qT_d}. \quad (72)$$

For the internal-stability test, the terminal-current perturbation and the outer-loop reference perturbation are set to zero. Substitution of (72) into (21), with the capacitor and inductor dynamics retained, gives the exact quasi-polynomial characteristic function

$$\Delta_v(s, k_{vp}) = \ell_f c_f s^2 + 1 - F_v H_d(q) + H_d(q) G_i(q) \left[G_v(q) + c_f q \right] - j\omega_1 \ell_f c_f s H_d(q). \quad (73)$$

Internal stability requires every zero of Δ_v to lie in the open left half-plane. Because k_{vp} enters (73) affinely, the characteristic function can be written as

$$\Delta_v(s, k_{vp}) = \Delta_0(s) + k_{vp} \Delta_1(s), \quad (74)$$

where

$$\begin{aligned} \Delta_0(s) &= \ell_f c_f s^2 + 1 - F_v H_d(q) + H_d(q) G_i(q) \left(\frac{k_{vi}}{q} + c_f q \right) \\ &\quad - j\omega_1 \ell_f c_f s H_d(q), \\ \Delta_1(s) &= H_d(q) G_i(q). \end{aligned} \quad (75)$$

Equations (74) and (75) permit the stability boundary to be found without approximating the exponential delay. If a root crosses the imaginary axis at $s = j\omega_c$, the corresponding gain must satisfy

$$k_{vp}(\omega_c) = -\frac{\Delta_0(j\omega_c)}{\Delta_1(j\omega_c)}, \quad \text{Im} \left\{ -\frac{\Delta_0(j\omega_c)}{\Delta_1(j\omega_c)} \right\} = 0, \quad k_{vp}(\omega_c) > 0. \quad (76)$$

Thus, the gain-boundary candidates are obtained by enumerating the positive real solutions of (76), rather than by following only one preselected root-locus branch. Candidate frequencies are bracketed on a logarithmic grid and refined by a scalar zero solver; the full characteristic spectrum is then continued between adjacent candidates so that a crossing belonging to any root branch cannot be omitted.

Because the stationary-frame characteristic function is complex, both the positive-frequency branch and its mirror-frequency branch must be checked. With the Appendix A parameters, $k_{vi} = 500$, $k_{ip} = 5$, $k_{ii} = 100$, $F_v = 1$ and the nominal delay $T_d = 50 \, \mu\text{s}$, the first stability-changing solutions of the exact exponential-delay Equation (76) are

$$\begin{aligned} k_{vp,-}^L &= 0.064473, & f_-^L &= -598.87 \, \text{Hz}, \\ k_{vp,+}^L &= 0.077118, & f_+^L &= 711.93 \, \text{Hz}, \\ k_{vp,+}^U &= 0.526146, & f_+^U &= 1974.87 \, \text{Hz}, \\ k_{vp,-}^U &= 0.557775, & f_-^U &= -1993.50 \, \text{Hz}. \end{aligned} \quad (77)$$

Here, the subscripts “+” and “−” denote the positive- and mirror-frequency branches, respectively, while “L” and “U” denote the lower and upper crossings. The two modal stability intervals and their intersection are

$$\begin{aligned}\mathcal{K}_+ &= (0.077118, 0.526146), \\ \mathcal{K}_- &= (0.064473, 0.557775), \\ \mathcal{K}_{\text{int}} &= \mathcal{K}_+ \cap \mathcal{K}_- = (0.077118, 0.526146).\end{aligned}\quad (78)$$

Thus, the positive-frequency branch supplies both active boundaries. The two mirror-frequency crossings are nevertheless retained in the calculation so that the final interval is an intersection over all HFO branches rather than the result of following one selected trajectory. Higher-frequency delay-induced candidates occur only after the system has already entered the unstable component and do not alter this connected admissible interval. The crossing direction follows from the implicit-function theorem:

$$\left. \frac{ds}{dk_{vp}} \right|_{s=j\omega_c} = - \frac{\Delta_1(j\omega_c)}{\partial \Delta_0(j\omega_c)/\partial s + k_{vp} \partial \Delta_1(j\omega_c)/\partial s}.\quad (79)$$

For the positive-frequency branch, the real parts of (79) are $-1.3401 \times 10^4 \text{ s}^{-1}/\text{p.u.}$ at the lower boundary and $+6.8174 \times 10^3 \text{ s}^{-1}/\text{p.u.}$ at the upper boundary. The corresponding mirror-branch values are $-1.3675 \times 10^4 \text{ s}^{-1}/\text{p.u.}$ and $+6.9804 \times 10^3 \text{ s}^{-1}/\text{p.u.}$ All four crossings are, therefore, transversal. At either lower crossing, increasing k_{vp} moves the corresponding mode into the LHP because the proportional action supplies the damping missing from the integral-loop mode. At either upper crossing, increasing k_{vp} moves a filter- and delay-affected mode into the RHP. This establishes analytically why the gain dependence is non-monotonic.

For an independent argument-principle check, define the parameter return ratio

$$L_v(s; k_{vp}) = k_{vp} \frac{\Delta_1(s)}{\Delta_0(s)}, \quad \Delta_v(s, k_{vp}) = \Delta_0(s) [1 + L_v(s; k_{vp})].\quad (80)$$

The RHP-zero count of the open characteristic function is evaluated directly on a Nyquist contour Γ_R :

$$P = \lim_{R \rightarrow \infty} \frac{1}{2\pi j} \oint_{\Gamma_R} \frac{\Delta'_0(s)}{\Delta_0(s)} ds = 2.\quad (81)$$

Consequently, L_v has $P = 2$ RHP poles. Using the convention that clockwise encirclements of $(-1, j0)$ are positive, the argument principle gives

$$Z = P + N,\quad (82)$$

where Z is the number of RHP zeros of Δ_v and N is the net clockwise encirclement number. Stability, therefore, requires $N = -2$, i.e. two net counter-clockwise encirclements. Table 3 reports the real parts of both HFO branches together with the Nyquist count. The corresponding root trajectories of the two branches are plotted in Figure 6, and the Nyquist loci of $L_v(j\omega; k_{vp})$ at the representative gains are shown in Figure 7. The exact exponential delay is used in (76), in the Nyquist evaluation and in the numerical continuation of both root trajectories; no Padé approximation is used for the reported boundaries or the plotted branches.

The root locus and the Nyquist calculation, therefore, give the same maximal internal-stability interval connected to the nominal design point:

$$0.077118 < k_{vp} < 0.526146.\quad (83)$$

Table 3. Real parts of the positive- and mirror-frequency HFO branches and Nyquist count as k_{vp} varies at $T_d = 50 \mu s$. Here, $P = 2$ for L_v , and stability requires $N = -2$ and $Z = P + N = 0$.

k_{vp}	$\sigma_+ (s^{-1})$	$\sigma_- (s^{-1})$	N	Z	Result
0.050	+338.9	+190.3	0	2	unstable
0.070	+93.6	−76.8	−1	1	unstable
0.100	−327.0	−542.1	−2	0	stable
0.150	−1193.4	−1559.1	−2	0	stable
0.220	−1955.9	−2418.6	−2	0	stable
0.300	−1619.8	−1975.1	−2	0	stable
0.400	−898.6	−1179.4	−2	0	stable
0.500	−180.1	−413.8	−2	0	stable
0.540	+93.9	−125.1	−1	1	unstable
0.600	+489.0	+289.3	0	2	unstable

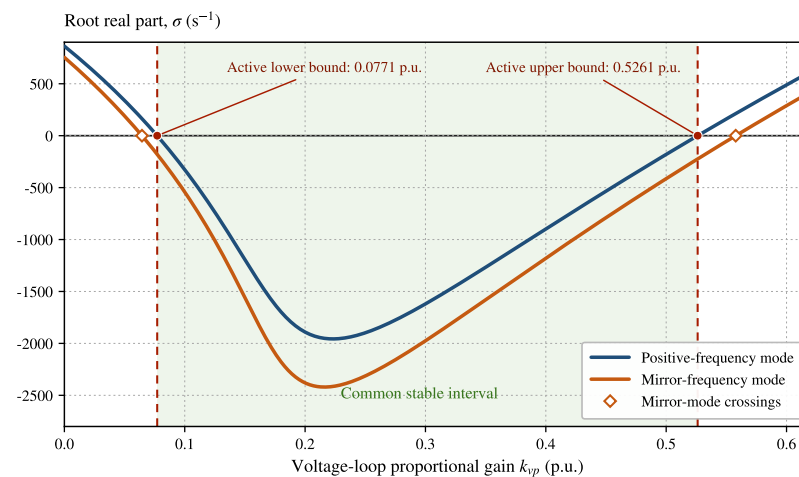


Figure 6. Exact root trajectories of the two GFM-BESS HFO branches as k_{vp} varies at $T_d = 50 \mu s$. Circular markers denote the two active crossings of the positive-frequency branch; diamond markers denote the two mirror-frequency crossings. The shaded interval is the common stable set (78). Slow non-HFO poles are omitted for clarity.

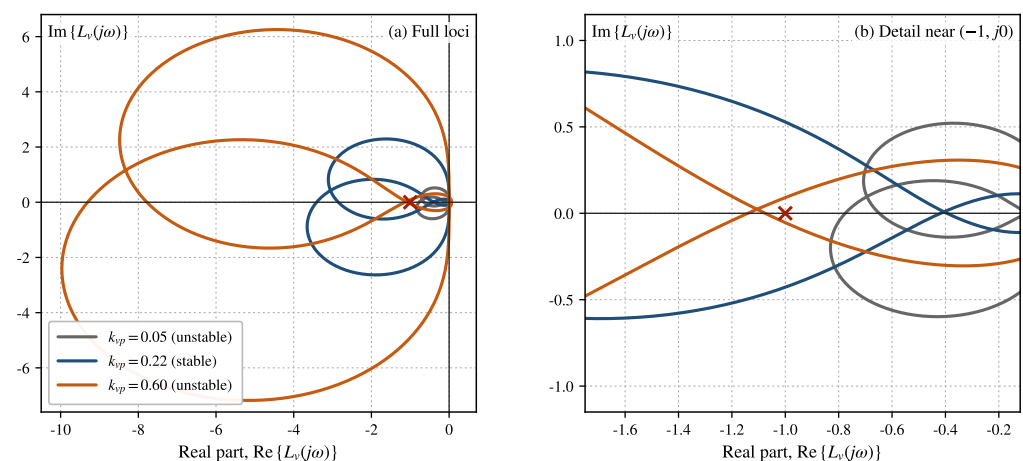


Figure 7. Nyquist loci of $L_v(j\omega; k_{vp})$: (a) full loci; (b) neighborhood of $(-1, j0)$. At $k_{vp} = 0.22$, the two net counter-clockwise encirclements give $N = -2$ and cancel the two RHP poles of L_v , whereas $k_{vp} = 0.05$ and 0.60 have zero net encirclement and retain two closed-loop RHP poles.

This internal-stability test is followed by the interconnection condition

$$Z_{\text{sys1}}(s) + Z_{\text{sys2}}(s) = 0 \quad (84)$$

or, equivalently, the generalized Nyquist condition on the loop-gain matrix $\mathbf{L}$. In the HFO band, the coupling bound of Section 2.3 permits the scalar ratio $L_Z = Z_{\text{sys1}}/Z_{\text{sys2}}$ to be used. For the fixed network parameters of Appendix A and $T_d = 50 \mu\text{s}$, the impedance-ratio Nyquist sweep introduces no additional RHP root over (83); hence, $k_{vp,+}^L$ and $k_{vp,+}^U$ in (77) are the two active gain boundaries. The final admissible interval is, therefore, the intersection of the internal-stability interval and the impedance-ratio Nyquist-stable set, which here equals (83).

4. Proposed Gain-Retuning Strategy and Model Validation

The analytical impedance models derived in Section 2 form the basis of the stability boundary and the gain-retuning rule. Their terminal characteristics are first verified against frequency-sweep results obtained from the detailed PSCAD model. The exact interval of Section 3.3 is then reduced by a damping margin to obtain an engineering range and a nominal gain. Finally, a two-step time-domain gain test verifies that moving from above the upper boundary to the nominal gain suppresses the HFO, whereas moving below the lower boundary re-excites it.

4.1. Frequency-Sweep Validation of the Impedance Models

To verify that the derived frequency-domain models reproduce the terminal characteristics of the detailed PSCAD simulation, small-signal frequency sweeps are performed independently for the GFM-BESS and the GFL-HVDC. At each perturbation frequency, the corresponding voltage and current responses are extracted to calculate the simulated impedance, which is then compared with the analytical impedance obtained from (30) and (53) using the same operating point and control parameters.

Figure 8 compares the analytical GFM-BESS impedance with the PSCAD frequency-sweep result. The two results agree closely in both magnitude and phase over the frequency range relevant to the HFO, including the impedance variation around the high-frequency resonance. This agreement confirms that the GFM-BESS model retains the control-loop and delay dynamics required to describe its high-frequency terminal behavior.

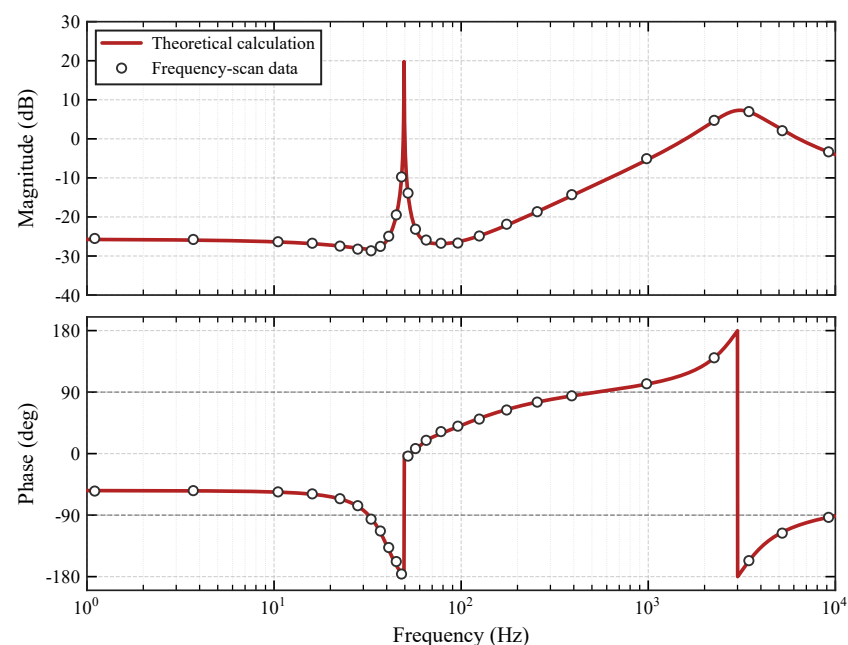


Figure 8. Comparison of the analytical impedance model of the GFM-BESS with the frequency-sweep result obtained from PSCAD.

Figure 9 presents the corresponding comparison for the GFL-HVDC. The analytical and PSCAD results exhibit consistent magnitude and phase characteristics throughout the investigated band, showing that the current loop, voltage feed-forward, control delay and PLL retained in the model capture the high-frequency response of the converter. The agreement in Figures 8 and 9 validates the two converter impedance models and supports their subsequent use in the stability assessment and suppression design.

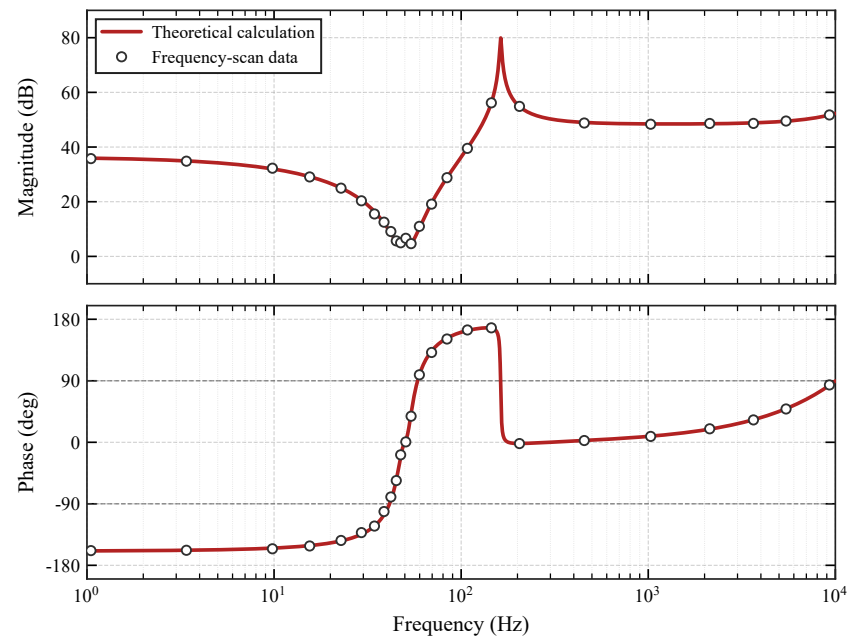


Figure 9. Comparison of the analytical impedance model of the GFL-HVDC with the frequency-sweep result obtained from PSCAD.

4.2. Gain Selection and Stability Margin

The exact mathematical interval (83) should not be used up to either boundary because modeling error, parameter drift and discrete implementation effects can move the imaginary-axis crossings. A damping-based engineering interval is, therefore, obtained from the high-frequency spectral abscissa

$$\alpha_h(k_{vp}) = \max_{\lambda \in \Lambda_h(k_{vp})} \text{Re}\{\lambda\}, \quad (85)$$

where Λ_h contains the oscillatory roots above 300 Hz. Over the exact stable interval, α_h reaches its minimum near $k_{vp} = 0.223$. Moreover, Table 3 shows that the root real part remains below approximately $-1.19 \times 10^3 \text{ s}^{-1}$ throughout

$$0.15 \leq k_{vp} \leq 0.30. \quad (86)$$

This interval is separated from both imaginary-axis crossings and is, therefore, adopted as the engineering range. The nominal value is selected near the maximum-damping point:

$$k_{vp}^{\text{nom}} = 0.22. \quad (87)$$

Although 0.22 appears numerically small when compared with the original trial value $k_{vp} = 5$, the two values are not close dynamically. From (71), the proportional voltage-loop frequency scale under the ideal inner-current-loop approximation is $\omega_{v,p} \approx k_{vp}/c_f$. Hence, $k_{vp} = 0.22$ corresponds to $\omega_{v,p} \approx 9.55 \times 10^3 \text{ rad/s}$, or approximately 1.52 kHz; therefore, it does not represent a slow voltage response. By contrast, $k_{vp} = 5$ is 22.7 times the nominal value and lies far above the exact upper boundary. Its apparent plausibility results from

comparing a per-unit controller gain with unnormalized physical filter coefficients. The gain scale in (83) is, thus, a consequence of the stated per-unit convention, not a numerical artefact of the root solver.

The selection procedure can be reproduced for another operating point as follows: (i) construct Δ_0 and Δ_1 from (75); (ii) enumerate every positive solution of (76); (iii) use a test point between successive crossings and the argument principle to identify the stable intervals; (iv) intersect these intervals with the impedance-ratio Nyquist-stable set; and (v) select the gain that minimizes (85), subject to a prescribed distance from both boundaries. This procedure explicitly checks all root branches and does not rely on an impedance-magnitude crossing or on the assumption that increasing k_{vp} is always beneficial.

The physical trade-off is also two-sided. Below $k_{vp,+}^L$, the proportional action is too weak to damp the mode associated with k_{vi}/q . Above $k_{vp,+}^U$, the voltage loop demands a response in the filter- and delay-affected frequency range; the resulting phase loss causes the approximately 2 kHz mode to cross into the RHP. Thus, a larger k_{vp} initially improves damping but eventually reduces it. This explains why the proportional gain must be selected inside a finite interval rather than at its largest implementable value. All results in this section use the nominal delay $T_d = 50 \mu\text{s}$; no enlarged-delay operating case is invoked.

4.3. Time-Domain Validation of the Gain-Retuning Strategy

The two-sided gain condition is further verified by applying two successive changes to k_{vp} in the detailed time-domain model. Figure 10 shows the complete response from 2.8 to 3.6 s. Before $t = 3.0$ s, $k_{vp} = 0.60$, which is greater than the active upper boundary $k_{vp,+}^U = 0.526146$; the system, therefore, exhibits an HFO. At $t = 3.0$ s, k_{vp} is reduced from 0.60 to the nominal value 0.22. Because 0.22 lies inside both modal stability intervals in (78) and near the maximum-damping point, the oscillatory component decays and the system returns to steady operation. At $t = 3.4$ s, k_{vp} is further decreased by 0.15 p.u., from 0.22 to 0.07. The resulting value is below the active lower boundary $k_{vp,+}^L = 0.077118$, and the HFO is consequently re-excited. The two transitions, therefore, verify both the upper and lower sides of the analytically derived interval.

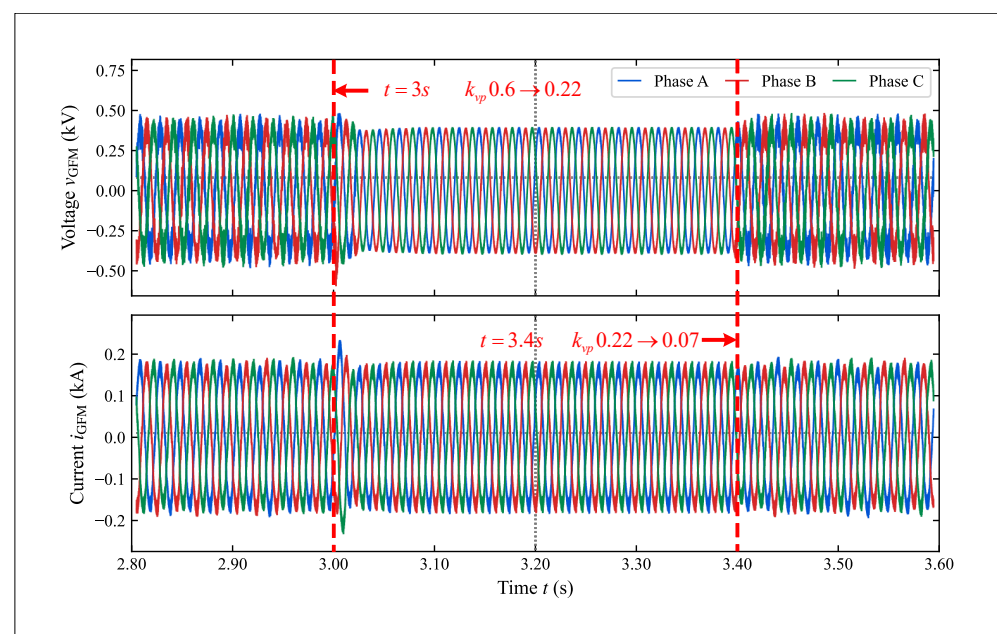


Figure 10. Time-domain validation of the two-sided voltage-loop-gain condition from 2.8 to 3.6 s. At $t = 3.0$ s, k_{vp} is reduced from 0.60 to 0.22 and the HFO is suppressed. At $t = 3.4$ s, it is further decreased by 0.15 p.u. to 0.07, and the HFO is re-excited.

Figure 11 enlarges the first transition over 2.9–3.1 s. Immediately after k_{vp} is changed to 0.22 at $t = 3.0$ s, the high-frequency component decreases rapidly. This local view confirms that the suppression observed in Figure 10 is a damped response associated with moving the critical roots into the left-half plane, rather than an apparent reduction caused by the scale of the complete time record.

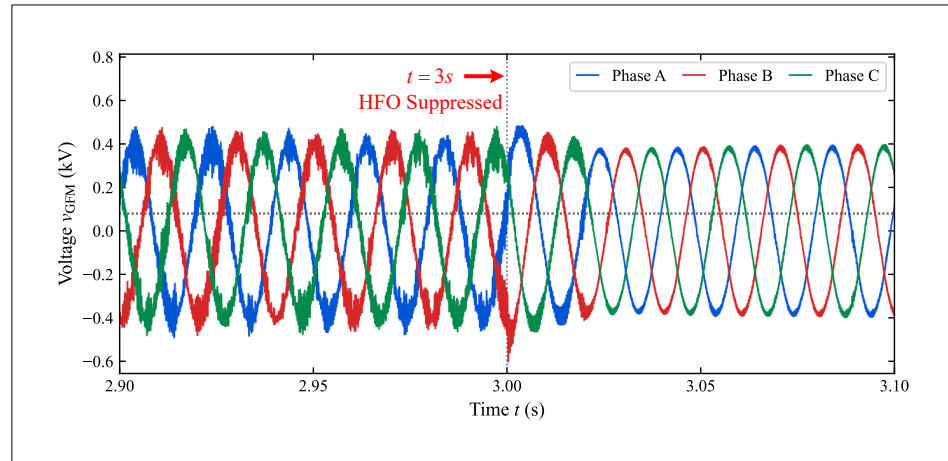


Figure 11. Enlarged view of the HFO-suppression interval from 2.9 to 3.1 s. Reducing k_{vp} from 0.60 to 0.22 at $t = 3.0$ s causes the high-frequency component to decay.

The complementary transition is shown in Figure 12 over 3.3–3.5 s. The system remains well damped while $k_{vp} = 0.22$. After the gain is decreased by 0.15 p.u. to 0.07 at $t = 3.4$ s, a high-frequency component develops again. This re-excitation is consistent with the positive-frequency root crossing the imaginary axis at $k_{vp,+}^L$ and completes the time-domain verification of the proposed retuning strategy.

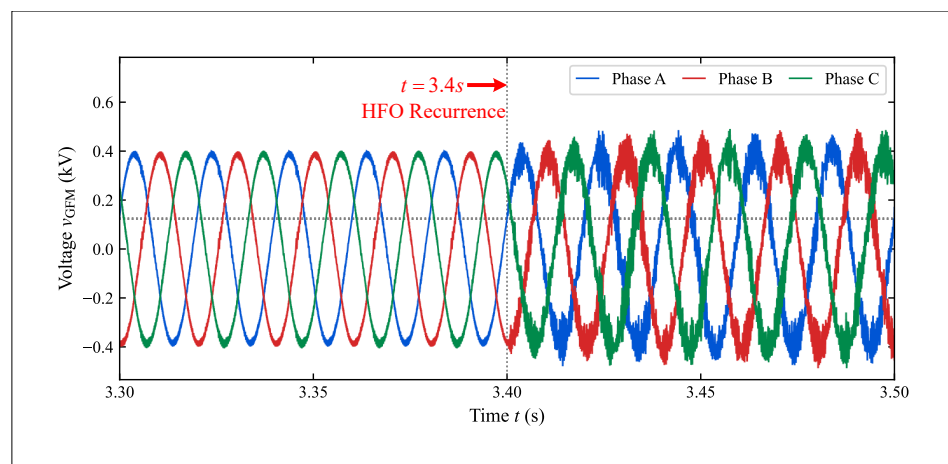


Figure 12. Enlarged view of the HFO re-excitation interval from 3.3 to 3.5 s. At $t = 3.4$ s, k_{vp} is decreased by 0.15 p.u. from 0.22 to 0.07, placing the gain below its lower stability boundary and causing the HFO to reappear.

5. Conclusions

This paper studied the high-frequency stability of an interconnected GFM-BESS and GFL-HVDC system by impedance modeling and rigorous controller-parameter analysis. The composite network was divided into voltage-source and current-source subsystems, and a frequency-coupled 2×2 impedance model was established; the mirror-frequency coupling was shown to be negligible in the HFO band, which permits a scalar-impedance assessment there. Internal converter stability was checked before applying the interconnection

Nyquist criterion. By writing the exact exponential-delay characteristic function in a form affine in k_{vp} , the crossings of both the positive- and mirror-frequency HFO branches were enumerated directly. At $T_d = 50 \mu\text{s}$, their modal intervals are $\mathcal{K}_+ = (0.077118, 0.526146)$ and $\mathcal{K}_- = (0.064473, 0.557775)$; hence, their intersection is $0.077118 < k_{vp} < 0.526146$. Exact root continuation and the argument-principle/Nyquist count independently give the same stable interval. The non-monotonic root motion shows that insufficient gain fails to damp the integral-loop mode, whereas excessive gain destabilizes the filter- and delay-affected mode. A robust range of $0.15 \leq k_{vp} \leq 0.30$ and a nominal value $k_{vp} = 0.22$ are, therefore, recommended. This two-sided rule replaces a monotonic high-gain recommendation and provides a reproducible hardware-free retuning procedure. Time-domain gain-step tests corroborate both sides of the interval: changing k_{vp} from 0.60 to 0.22 suppresses the HFO, whereas a further reduction of 0.15 p.u. to 0.07 drives the gain below the lower boundary and re-excites the oscillation. Future work will address coordinated wide-band damping in multi-converter systems and experimental validation under parameter uncertainty.

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Appendix A. System Parameters

Table A1. System parameters used in the analysis.

Symbol	Description	Value
<i>System</i>		
V_n	Rated voltage	480 V
ω_1	Rated angular frequency	$100\pi \text{ rad/s}$
<i>GFM-BESS</i>		
P_n/Q_n	Rated active / reactive power	100 kW/0 kVar
J	Virtual inertia constant	0.001 p.u.
D_d	Damping coefficient	0.25 p.u.
K	Reactive-power droop coefficient	20 p.u.
k_{vp}	Voltage-loop proportional gain	0.22 p.u.
k_{vi}	Voltage-loop integral gain	500 s^{-1}
k_{ip}	Current-loop proportional gain	5 p.u.
k_{ii}	Current-loop integral gain	100 s^{-1}
F_v/F_i	Voltage/current feed-forward coefficient	1/0.75
T_d	Control delay	$50 \mu\text{s}$
L_f/C_f	Filter inductance/capacitance	2 mH/10 μF

Table A1. Cont.

Symbol	Description	Value
GFL-HVDC		
P_{mn}/Q_{mn}	Rated active/reactive power	600 kW/0 kVar
k_{mpp}/k_{mpi}	Power-loop proportional/integral gain	0.2/0.05 p.u.
k_{mip}/k_{mii}	Current-loop proportional/integral gain	1/1 p.u.
k_{pp}/k_{pi}	PLL proportional/integral gain	50/900 p.u.
F_{mv}	Voltage feed-forward coefficient	1
T_{md}	Control delay	50 μ s
R_m/L_m	Filter resistance/inductance	0.00015 Ω /1 mH
Renewable source		
P_{sn}/Q_{sn}	Rated active/reactive power	550 kW/0 kVar
k_{spp}/k_{spi}	Power-loop proportional/integral gain	0.2/0.05 p.u.
k_{sip}/k_{sii}	Current-loop proportional/integral gain	0.2/1 p.u.
k_{pp}/k_{pi}	PLL proportional/integral gain	50/900 p.u.
F_{sv}	Voltage feed-forward coefficient	1
T_{sd}	Control delay	50 μ s
R_{sf}/L_{sf}	Filter resistance/inductance	0.00015 Ω /2 mH
AC grid		
R_g/L_g	Grid resistance/inductance	0.05 Ω /4 mH

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