

DC-dynamic-aware Parameter Design of Matching Control for Grid-forming Inverters

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Abstract—The increasing deployment of inverter-based resources (IBRs) has heightened the demand for grid-forming (GFM) inverters whose stability depends on both grid conditions and their direct current (DC)-side characteristics. Among various GFM schemes, matching-controlled (MC) GFM inverters, while offering inherent energy-consistent synchronization, exhibit strong sensitivity to DC-link voltage variations due to their power-synchronization mechanism. This leads to performance degradation, especially when parameters are tuned solely seen from an alternating current (AC)-side perspective, particularly under varying grid strengths or constrained DC sources. Although prior studies have analyzed the behavior of MC inverters through modeling and stability assessment, a systematic and quantitatively grounded parameter design that explicitly accounts for DC-side dynamics remains unexplored. To address this, a DC-dynamic-aware parameter design is proposed for MC-based GFM inverters. A large-signal model is established to capture AC- and DC-coupled dynamics and identify instability mechanisms during transient events. Parameter selection criteria for strong and weak grids are then derived by revealing how dominant constraints shift between the DC and AC sides. More importantly, a parameter-tuning method is proposed by combining steady-state power-limit boundaries with a constraint-aware region-of-attraction transient assessment. The effectiveness of the proposed design is verified by hardware-in-the-loop (HIL) experimental tests under strong- and weak-grid conditions, and the robustness of the selected parameter against variations in DC response time and DC-link capacitance is demonstrated.

Index Terms—Grid-forming (GFM) inverters, Matching control (MC), Direct current (DC)-link dynamics, Parameter tuning.

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NOMENCLATURE

MC	Matching control
SCR	Short circuit ratio
k_θ	Proportional gain of the MC
$k_{\theta,pu}$	Per-unit value of k_θ
$P_{e,i}^{max}$	Maximum power transmission capacity of the i -th inverter
P_i^{max}	Maximum output power capacity of the i -th inverter
S_i^{max}	Maximum DC power capacity of the i -th inverter
δ	Phase angle difference in the dq frame between the MC inverter and the grid

I. INTRODUCTION

THE increasing penetration of renewable energy sources (RESs) and power electronics converters is transforming modern power systems from synchronous generator (SG)-dominated structures into those with massive inverter-based resources (IBRs) [1]–[2]. As such, grid-forming (GFM) control has become essential for maintaining system stability, providing inertia, and ensuring robust frequency support [3]–[6]. While effective for grid support, most of the existing GFM control methods are designed without explicit consideration of direct current (DC)-source dynamics. That is, the alternating current (AC)-side frequency and power dynamics are not explicitly coupled with the DC-side energy availability. As a result, GFM inverters can exhibit power or frequency dynamics that are inconsistent with the instantaneous capability of their DC sources under fault conditions, thereby degrading stability, particularly under limited or fluctuating RES conditions [7]–[9].

By contrast, the matching control (MC) is developed based on a source-aware synchronization mechanism. That is, by directly coupling the inverter's AC frequency to its DC-side energy state, the MC establishes a physical analogy between the DC-source dynamics and the SG rotor motion [10]–[12]. This enables autonomous operation and exhibits a droop-like behavior. However, the MC scheme also introduces a fundamental challenge that DC-side dynamics, including DC-link voltage fluctuations and source current/power limits, are prone to propagating directly into the synchronization loop [13]–[17]. As a result, tuning the MC parameters solely based on the AC-side criteria can fail to guarantee global stability, especially under weak grids or when the DC source is close to its operational limits [18]–[19].

To address these issues, various studies have been performed to examine and improve the stability of MC GFM inverters. For example, in [20], a unified sequence-impedance

model was formulated to characterize the output behavior of such inverters. However, the analysis remains restricted to small-signal conditions, and the influence of DC-voltage dynamics is difficult to fully infer from an impedance description. The consideration of DC-source dynamics was included in [21], where the impact of MC parameters was discussed. Nonetheless, no systematic guideline for parameter selection is provided.

Furthermore, a generalized parameter-design procedure based on H_∞ optimization was outlined in [22]. This method employs a robust control strategy to minimize the worst-case impact of disturbances and model uncertainties, though it does not incorporate variations in grid strength. In [23], a detailed single-input-single-output (SISO) model was constructed for stability assessment and parameter tuning, but the derivation relies heavily on steady-state linearization, leaving transient

behaviors unaddressed. Further guidance on parameter selection across different operating conditions was provided in [24] to enhance damping, while the corresponding effects on virtual inertia and transients remain unexamined. Moreover, conditions ensuring the existence, uniqueness, stability, and boundedness of MC inverters were rigorously derived in [25]. Nevertheless, the validation is performed on systems that include both SGs and GFM inverters, and its applicability in an IBR-based system has not been fully verified.

Although the prior-art studies have explored the GFM characteristics of the MC inverter through modeling, stability analysis, and experimental verification, a quantitative, DC-dynamic-aware parameter design method for MC-based GFM inverters remains of interest. As summarized in Table I, three main limitations can be identified in state-of-the-art studies:

TABLE I
COMPARISON OF EXISTING STUDIES ON STABILITY ANALYSIS AND IMPROVEMENT OF MC-BASED GFM INVERTERS

Ref.	Method and main contribution	Focus and limitations
[20]	Sequence-impedance-based frequency-domain stability assessment	1) Frequency-domain stability analysis 2) DC dynamics and large-signal transients not explicit
[21]	MC strategy considering DC-voltage dynamics	1) DC-voltage dynamics considered 2) No systematic gain-selection guideline
[22]	Generalized GFM control design based on H_∞ optimization	1) Optimization-based control design 2) Grid-strength variation not explicit
[23]	SISO linearized model for stability assessment and parameter tuning	1) Local stability assessment 2) Limited transient evaluation
[24]	Damping enhancement for DC-link-voltage-synchronized control	1) Damping-oriented improvement 2) DC-side constraints not quantified
[25]	Theoretical conditions for existence, stability, and boundedness	1) Rigorous stability theory 2) Applicability to IBR-dominated systems not fully verified
This work	DC-dynamic-aware parameter design with large-signal modeling and quantitative gain bounds	1) DC- and AC-constraints included 2) Steady-state feasibility and transient robustness jointly assessed

1) The large-signal transient assessment is insufficiently addressed due to the reliance on small-signal or steady-state linearized models.

2) DC-link dynamics and DC-source power capability (i.e., energy availability) are not explicitly mapped into quantitative parameter boundaries.

3) The constraint shift between DC- and AC-side is not systematically incorporated into the parameter design.

To address these challenges, this paper proposes a parameter design for MC-based GFM inverters that explicitly incorporates DC-voltage dynamics. The main contributions are summarized as follows:

1) Large-signal modeling with explicit DC-voltage dynamics: A large-signal model is formulated to capture the DC-link dynamics, power transmission limitations, and grid interactions, enabling accurate characterization of transients beyond small-signal assumptions.

2) Parameter-selection principles under strong and weak grids: The analysis identifies the influence of DC- and AC-side constraints on transient stability, leading to distinct and physically interpretable design principles for selecting the control parameter.

3) A combined parameter-selection method based on both steady-state and dynamic performance: Steady-state power-limit boundaries are analytically derived to determine feasible gain intervals, and a constraint-aware region of attraction

(ROA)-inspired analysis is incorporated to evaluate transient recoverability under baseline DC-source parameters.

The improvement with the proposed design is validated through hardware-in-the-loop (HIL) experimental tests. The experimental results demonstrate improved transient stability and enhanced frequency regulation capability under various operating scenarios. The remainder of this paper is structured as follows. The control schemes and large-signal modeling are presented in Section II. The proposed parameter design method is detailed and analyzed in Section III, followed by the HIL experimental validations in Section IV. Finally, Section V concludes the paper.

II. LARGE-SIGNAL MODELING OF THE GRID-CONNECTED MATCHING-CONTROLLED INVERTER

A. Control Diagram

The configuration of a single grid-connected inverter is depicted in Fig. 1. The outer synchronization loop and the inner cascaded voltage-current control loops of the MC are shown in Figs. 2(a) and 2(c), respectively. The voltage reference v_{af}^* in MC loop is determined by:

$$v_{af}^* = \mu [-\sin \theta \quad \cos \theta]^T \quad (1)$$

where μ and θ are the magnitude and phase angle of the AC modulation voltage, respectively. Similar to the droop control where the frequency is regulated according to active power,

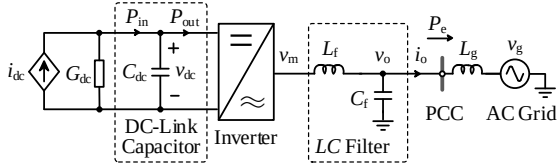


Fig. 1. Configuration of a single grid-connected inverter considering DC-link dynamics (PCC—point of common coupling; i_{dc} , G_{dc} —the current and equivalent conductance of the DC source; v_{dc} , C_{dc} —the DC-link voltage and capacitance; P_{in} , P_{out} —input and output power flow through the DC-link capacitor; v_m , v_o , i_o —the input voltage, output voltage and output current of the filter; L_f , C_f —the inductance and capacitance of the LC filter; P_e —input active power to the grid; L_g —the inductance of the transmission line; v_g —the AC grid voltage).

the frequency in the MC is adjusted based on the energy deviation through the DC-link capacitor, thereby establishing a source-aware synchronization structure. The relationship between the frequency ω and DC-link voltage v_{dc} in the synchronization loop of the MC shown in Fig. 2(a) is expressed as:

$$\omega - \omega^* = k_\theta (v_{dc} - V_{dc}^*) \quad (2)$$

where ω^* , k_θ , and V_{dc}^* are the references of angular frequency, proportional gain of the MC loop, and DC-link voltage, respectively. Subsequently, the AC voltage reference v_{dq}^* is derived and utilized in the inner control loops. Furthermore, since the dynamics of the grid are increasingly reshaped by GFM inverters that emulate SGs, the responses of the IBR-based grid in transients can be simplified as those of the virtual synchronous generator (VSG) control [15], [26]. The outer synchronization loop of the grid is shown in Fig. 2(b) as a general VSG control loop, and inner control loops are illustrated in Fig. 2(c). The voltage reference v_{abc}^* in the VSG control loop is given as:

$$v_{abc}^* = 2\omega M_f i_f \begin{bmatrix} \sin \theta & \sin\left(\theta - \frac{2\pi}{3}\right) & \sin\left(\theta - \frac{4\pi}{3}\right) \end{bmatrix}^T \quad (3)$$

where M_f and i_f represent the virtual mutual inductance magnitude and excitation current, respectively.

B. State-space Model with DC Dynamics

To investigate the impact of DC dynamics in the grid-connected system, the transients of the proportional-integral (PI) control modules are neglected since they are fast enough compared with the dynamics of the phase angle. Specifically, the following should hold:

$$v_{odq} = v_{odq}^* \quad (4)$$

$$i_{odq} = i_{odq}^* \quad (5)$$

$$\|v_{odq}\| = v_o^* \quad (6)$$

where v_{odq} and i_{odq} are the output voltage and current in the dq frame, respectively. The superscript “*” denotes the reference value, while $\|\cdot\|$ represent the amplitude. The phase angle difference δ in the dq frame between the MC inverter and the grid is expressed as:

$$\delta = \delta^c - \delta^s \quad (7)$$

$$\dot{\delta}^c = \omega^c - \omega^* \quad (8)$$

$$\dot{\delta}^s = \omega^s - \omega^* \quad (9)$$

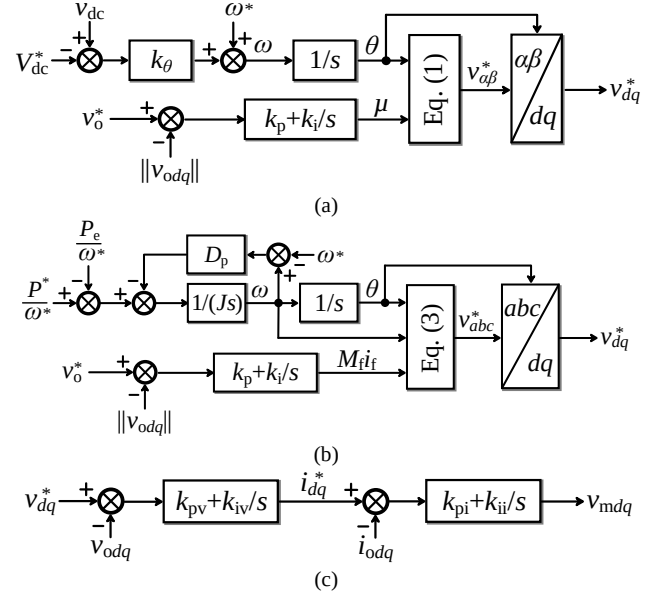


Fig. 2. Control schemes of the GFM inverter and the grid (V_{dc}^* , v_o^* , ω^* , and P^* —the references of DC-link voltage, AC voltage magnitude, angular frequency, and active power; ω and θ —angular frequency and phase angle, k_p and k_i —proportional and integral coefficients of the AC voltage magnitude controller, k_θ and μ —proportional gain and modulation AC voltage magnitude of the MC loop; J and D_p —virtual inertia and damping coefficients of the VSG control loop; M_f and i_f —virtual mutual inductance magnitude and excitation current of the VSG control loop; v_{abc}^* , $v_{\alpha\beta}^*$, and v_{dq}^* —voltage references in abc, $\alpha\beta$ and dq frames; v_{odq} and v_{mdq} —output voltage and modulation voltage in the dq frame; i_{odq} and i_{dq}^* —output current and its reference in the dq frame, k_{pv} and k_{iv} —proportional and integral coefficients of the voltage controller; k_{pi} and k_{ii} —proportional and integral coefficients of the current controller; $\|\cdot\|$ —amplitude of the signal). (a) Outer synchronization loop: MC loop. (b) Outer synchronization loop: VSG control loop. (c) Inner control loops: cascaded voltage and current control loops.

where δ^c and δ^s refer to the phase angles of the MC inverter and the grid in the dq frame, respectively. The relationship between θ in Fig. 2 and δ in (7) to (9) can be given as:

$$\dot{\delta} = \dot{\theta} - \omega^* \quad (10)$$

where θ is the phase angle of the MC inverter in the $\alpha\beta$ frame. Accordingly, the first- and second-order derivatives of δ can be obtained from (2), and (7)-(9) as:

$$\dot{\delta} = \omega^c - \omega^s = k_\theta (v_{dc} - V_{dc}^*) - \dot{\delta}^s \quad (11)$$

$$\ddot{\delta} = k_\theta \dot{v}_{dc} - \ddot{\delta}^s \quad (12)$$

in which $\ddot{\delta}^s$ and $\dot{v}_{dc}$ are the second-order derivative of δ^s and the derivative of v_{dc} . The dynamics of the grid is described by the swing equation as:

$$J_g \omega^* \dot{\delta}^s = -D_{p,g} \omega^* \dot{\delta}^s + P_g^* - P_{e,g} \quad (13)$$

where J_g and $D_{p,g}$ are the equivalent inertia and damping coefficients, respectively. Moreover, in (13), $P_{e,g}$ and P_g^* are the output active power and its reference, respectively. The derivative of δ^s can be derived from (11) as:

$$\dot{\delta}^s = k_\theta (v_{dc} - V_{dc}^*) - \dot{\delta} \quad (14)$$

The dynamics of the DC-link capacitor in Fig. 1 can be expressed as:

$$P_{in} - P_{out} = \frac{C_{dc}}{2} \dot{v}_{dc}^2 = C_{dc} v_{dc} \dot{v}_{dc} \quad (15)$$

where P_{in} and P_{out} are the input and output power flows through the DC-link capacitor C_{dc} in Fig. 1, respectively. Accordingly, the derivative of the DC-link voltage v_{dc} is

$$\dot{v}_{dc} = \frac{P_{in} - P_{out}}{C_{dc} v_{dc}} \quad (16)$$

with

$$P_{in} = v_{dc} i_{dc} - G_{dc} v_{dc}^2 \quad (17)$$

$$P_{out} = P_e = \frac{3E_g V_g}{2\omega^* L_g} \sin \delta = P_e^{\max} \sin \delta \quad (18)$$

where E_g and V_g are the voltage amplitude at the point of common coupling (PCC) and the grid voltage amplitude, respectively. G_{dc} is the equivalent conductance of the DC source, L_g is the grid inductance, and $P_e^{\max}$ refers to the maximum AC power transmission capacity.

The output current reference of the DC-source i_{dc}^* with the output limitation $i_{dc}^{\max}$ is controlled as:

$$i_{dc}^* = \begin{cases} k_{dc} (V_{dc}^* - v_{dc}) + \frac{P^*}{V_{dc}^*} + G_{dc} v_{dc}, & P_{out} < V_{dc}^* i_{dc}^{\max} \\ i_{dc}^{\max}, & P_{out} \geq V_{dc}^* i_{dc}^{\max} \end{cases} \quad (19)$$

where k_{dc} and P^* denote the proportional gain of the current control and the active power reference, respectively. With the consideration of the DC-voltage dynamics, the time delay between the DC-source current i_{dc} and its reference i_{dc}^* is modeled as a low-pass filter as:

$$i_{dc} = \frac{1}{\tau_{dc} s + 1} i_{dc}^* \quad (20)$$

where s refers to the complex frequency. τ_{dc} represents the time constant of the DC source and can be adjusted flexibly to investigate the impact of the DC dynamics. By adjusting C_{dc} , G_{dc} , τ_{dc} , and $i_{dc}^{\max}$, a general DC model that captures the dominant dynamics can be obtained for emulating various types of DC sources, such as PV systems, battery energy storage systems, and DC microgrids [12], [27]. Furthermore, the first-order assumption can fail to reveal the accurate characteristics under strongly nonlinear or piecewise operating modes. In this case, the model can be updated with a source-specific model, and the DC transients can be refined by incorporating nonlinear impedances, dynamics of the maximum power point tracking (MPPT), or current-limit switching if required. In this work, the aforementioned model, with the first-order assumption, is adopted to explore the DC dynamics.

Additionally, the output active power of the inverter and the grid is identical to the active power consumed by the load. Therefore, the following holds

$$P_{e,g} = P_L - P_e \quad (21)$$

with $P_{e,g}$, P_L , and P_e being the active power from the grid, to the load and from the inverter. The dynamics of the entire system can then be described by

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = k_\theta \frac{x_3 x_4 - \frac{3E_g V_g}{2\omega^* L_g} \sin x_1}{C_{dc} x_3} + \frac{D_{p,g}}{J_g} [k_\theta (x_3 - V_{dc}^*) - x_1] \\ \quad - \frac{1}{J_g \omega^*} \left(P_g^* - P_L + \frac{3E_g V_g}{2\omega^* L_g} \sin x_1 \right) \\ \dot{x}_3 = \frac{x_3 x_4 - G_{dc} x_3^2 - \frac{3E_g V_g}{2\omega^* L_g} \sin x_1}{C_{dc} x_3} \\ \dot{x}_4 = \frac{1}{\tau_{dc}} (i_{dc}^* - x_4) \end{cases} \quad (22)$$

with the state variables $\mathbf{x} = [x_1, x_2, x_3, x_4]^T$ being selected as

$$\begin{cases} x_1 = \delta \\ x_2 = \dot{\delta} \\ x_3 = v_{dc} \\ x_4 = i_{dc} \end{cases} \quad (23)$$

Furthermore, the input variables are given as $\mathbf{u} = [P_L, V_g, L_g, k_\theta]^T$. Subsequently, the system can be described by

$$\dot{\mathbf{x}} = \mathbf{f}(\mathbf{x}, \mathbf{u}) \quad (24)$$

which is the state-space model of the entire MC inverter considering the DC dynamics.

III. PROPOSED PARAMETER DESIGN

A. Consideration of Power Limits

The impact of the proportional gain k_θ in the MC loop is discussed in this section to investigate how it influences the virtual inertia and damping characteristics, and how these effects interact with the power limitations imposed by the DC source and the grid. The DC-side capacitance C_{dc} and conductance G_{dc} can be used by the MC loop to provide inertia and damping support, and the dynamics can be expressed as

$$J_v \dot{\omega} = T_{vm} - T_{ve} - D_v \omega \quad (25)$$

where $J_v = C_{dc}/k_\theta^2$ and $D_v = G_{dc}/k_\theta^2$ are the virtual inertia and virtual damping coefficients, respectively; T_{vm} and T_{ve} represent the virtual mechanical and electromagnetic torques [28], respectively. However, modifying the source-side parameters C_{dc} or G_{dc} is usually challenging in practical applications. By contrast, adjusting the control coefficient k_θ , in return, affects J_v and D_v , offering more flexibilities to enhance the frequency performance. That is, a flexible control coefficient k_θ can be introduced to re-establish the matching relationship between the angular frequency ω and the DC-link voltage v_{dc} , according to (2), thereby affecting the frequency regulation capability.

Specifically, a reduction of k_θ induces direct increases in J_v and D_v , effectively suppressing transient and reducing steady-

state frequency deviations. However, a lower k_θ also increases the DC power required for stable operation under transients. Consequently, arbitrary reduction of k_θ is infeasible, as an excessive decrease can drive the MC inverter to its power limit, thereby compromising the GFM capability. Therefore, determining feasible bounds for k_θ is essential to ensure that the inverter maintains sufficient virtual inertia while staying within both DC- and AC-side power limits.

$$P_i^{\max} = \min \{S_i^{\max}, P_{e,i}^{\max}\} \quad (26)$$

with

$$\begin{cases} S_i^{\max} = V_{dc,i}^* V_{dc,i}^{\max} \\ P_{e,i}^{\max} = \frac{3E_{g,i} V_{g,i}}{2\omega^* L_{g,i}} \end{cases} \quad (27)$$

where $S_i^{\max}$, $P_{e,i}^{\max}$, $P_i^{\max}$, $V_{dc,i}^*$, $V_{dc,i}^{\max}$, $E_{g,i}$, $V_{g,i}$ and $L_{g,i}$ are the maximum DC power capacity, maximum power transmission capacity, maximum power output capacity, DC-link voltage reference, DC-link current output limitation, PCC voltage, grid voltage, and transmission line inductance of the i -th inverter. This strategy ensures reasonable power sharing among all GFM inverters, provided the load step remains within the system's collective power capacity. It is worth noting that the relationship between $S_i^{\max}$ and $P_{e,i}^{\max}$ varies under different grid conditions. Precisely, $S_i^{\max}$ is less than $P_{e,i}^{\max}$ under a strong grid owing to a relatively low grid impedance $L_{g,i}$. On the contrary, $S_i^{\max}$ will be larger than $P_{e,i}^{\max}$ as the grid strength decreases, resulting in different considerations regarding the power limitation. The steady-state droop characteristics of the MC can be expressed as

$$P - P^* = \frac{k_{dc} V_{dc}}{k_\theta} (\omega^* - \omega) = D_p \omega^* (\omega^* - \omega) \quad (28)$$

where k_{dc} represents the DC-link current control loop parameter and D_p denotes the equivalent active power damping coefficient of the MC [15]. Accordingly, the value of k_θ can be determined based on the chosen D_p by (28). Furthermore, as indicated in (28), a larger k_θ leads to the decrease of D_p and less output power. Therefore, the available range of k_θ can be derived considering the power limits of both the MC inverter and the grid under a load step of ΔP_L .

B. Design Guideline

For practical implementation, the proposed design for k_θ is summarized as follows:

- 1) Specify the operating scenario and the corresponding power-capability constraints;
- 2) Derive the static feasible interval of k_θ from the steady-state power-sharing relationship;
- 3) Examine candidate values through the ROA-inspired transient assessment to determine the final parameter of k_θ considering both transient stability margin and frequency-support performance.

Following the above guideline, a load-step disturbance is considered to investigate the feasible operating range of k_θ . Without loss of generality, the MC inverter with the control parameter k_θ to be determined can be labeled as the first one

among the n inverters. Accordingly, other inverters are regarded as the grid side assets. Under a load step of ΔP_L , the power sharing between the MC inverter and the grid can be expressed as:

$$\Delta P_L = \Delta P_{L,1} + \Delta P_{L,g} = \Delta P_{L,1} + \sum_{i=2}^n \Delta P_{L,i} \quad (29)$$

in which $\Delta P_{L,1}$, $\Delta P_{L,g}$, and $\Delta P_{L,i}$ are the active power deviations of the MC inverter, the grid, and the i -th inverter. It is reflected by (29) how the MC inverter and the remaining GFM units collectively respond to a disturbance. Meanwhile, the relationship between k_θ and equivalent damping coefficient $D_{p,1}$ can be derived as:

$$k_\theta = \frac{k_{dc} V_{dc}}{\omega^* D_{p,1}} \quad (30)$$

1) Static boundary

Physically, a larger k_θ weakens the inverter's damping contribution, as shown in (30). As k_θ increases, the output power P_e of the MC inverter decreases accordingly, leading to greater power required from the grid side. Once the output active power of the grid increases to its maximum value $\Delta P_{L,g}^{\max}$, the value of k_θ also reaches the upper limit $k_\theta^{\max}$. Under this condition, the relationship of active power between the MC inverter $\Delta P_{L,1}$ and the grid $\Delta P_{L,g}$ can be expressed as:

$$\Delta P_{L,g} = \Delta P_{L,g}^{\max} = \sum_{i=2}^n (P_i^{\max} - P_i^*) \quad (31)$$

$$\Delta P_{L,1} = \Delta P_{L,1}^{\min} = \Delta P_L - \Delta P_{L,g}^{\max} \quad (32)$$

The damping coefficient can be designed as:

$$\frac{D_{p,1}^{\min}}{\Delta P_{L,1}^{\min}} = \frac{D_g}{\Delta P_{L,g}^{\max}} \quad (33)$$

where

$$D_g = \sum_{i=2}^n D_{p,i} \quad (34)$$

Accordingly, the upper limit of the MC proportional gain $k_\theta^{\max}$ can be derived as:

$$k_\theta^{\max} = \frac{k_{dc} V_{dc}}{\omega^* D_{p,1}^{\min}} \quad (35)$$

On the contrary, a smaller k_θ forces the MC inverter to provide more power for the same load change, potentially reaching its DC or AC power limits. Similarly, the lower limit of the MC proportional gain $k_\theta^{\min}$ can be calculated as

$$k_\theta^{\min} = \frac{k_{dc} V_{dc}}{\omega^* D_{p,1}^{\max}} \quad (36)$$

with

$$\Delta P_{L,1} = \Delta P_{L,1}^{\max} = P_1^{\max} - P_1^* \quad (37)$$

$$\Delta P_{L,g} = \Delta P_{L,g}^{\min} = \Delta P_L - \Delta P_{L,1}^{\max} \quad (38)$$

$$\frac{D_{p,1}^{\max}}{\Delta P_{L,1}^{\max}} = \frac{D_g}{\Delta P_{L,g}^{\min}} \quad (39)$$

Together, these static upper and lower bounds ensure that the MC inverter provides adequate inertia and damping, while operating within its power capabilities.

2) Transient assessment

The aforementioned static upper and lower limits of k_θ are determined from the power-sharing relationship in the steady state. However, the existence of an equilibrium point does not guarantee smooth and stable transients under the load-fault condition. Moreover, the adjustment of k_θ also affects J_v and D_v of the MC inverter, thereby affecting its frequency-support capability during dynamics. Although the steady-state bounds of k_θ provide useful guidance, they are not adequate or become ineffective to guarantee stable operation during disturbances. Therefore, transients under different grid strengths are further explored.

To fully investigate the impact of k_θ during transients, an IEEE 9-bus system [29] is used for the case study, with the topology shown in Fig. 3. The system specifications are listed in Table II to emulate a cluster of GFM inverters through aggregation [15], and the grid strength and capacity settings are listed in Table III. The grid-forming inverter (GFI) 1 in Fig. 3 employs the MC, and the VSG control strategy is adopted in GFIs 2 and 3 to form the grid dynamics. In this paper, the IEEE 9-bus system with homogeneous grid-side units (GFI 2 and GFI 3) is adopted as a representative scenario to facilitate analysis and validation. The short-circuit ratio (SCR) of the grid is calculated by

$$\text{SCR} = \frac{S_{sc}}{S_N} = \frac{3V_g^2}{2\omega^* L_g S_N} \quad (40)$$

where S_{sc} is the short-circuit capacity at the PCC. The SCR values are set to 3.50 and 1.60 to establish two representative cases with different dominant power constraints. These cases correspond to relatively higher and lower grid strengths and are denoted as the strong-grid and weak-grid conditions, respectively, for ease of discussion. For clarification, the maximum output powers of the MC inverter and the grid, considering both $S^{\max}$ and $P_e^{\max}$, are bolded in Table III. As listed in Table III, the dominant constraint comes from the DC-side power limits $S_1^{\max}$ and $S_g^{\max}$ under the strong-grid condition, and shifts to the AC-side transmission limits $P_{e,1}^{\max}$ and $P_{e,g}^{\max}$ when the grid strength decreases. The selected load-step values in Table III are designed to create sufficiently severe disturbances to explore the transient stability.

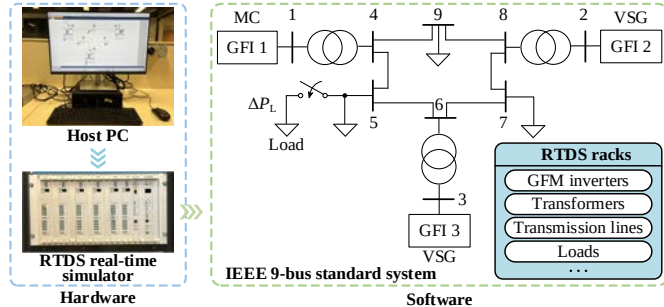


Fig. 3. HIL platform based on the real-time digital simulator (RTDS) with the IEEE 9-bus test system consisting of three GFIs.

TABLE II
SPECIFICATION OF THE SYSTEM FOR ANALYSIS AND
HIL EXPERIMENTAL TESTS

Symbol/unit	Value	Description
Base values		
S_N/MVA	100	Base power capacity
$\omega^*/(\text{rad/s})$	314	Base angular frequency
v_{dc}^*/V	2450	Base DC-link voltage
v_o^*/V_{rms}	1000	Base AC voltage (line-line)
Circuit		
τ_{dc}/s	0.05	Time constant of the DC source
G_{dc}/Ω^{-1}	0.83	Conductance of the DC source
C_{dc}/mF	8	DC-link capacitance
$L_f/\mu\text{H}$	200	Filter inductance
$C_f/\mu\text{F}$	300	Filter capacitance
Controller		
$k_{dc}/(\text{p.u.})$	100.0	Gain of the DC-source current control
$P^*/(\text{p.u.})$	1	Active power reference of the GFM control
$J/(\text{p.u.})$	0.5	Inertia coefficient of the VSG control
$D_p/(\text{p.u.})$	98.7	Damping coefficient of the VSG control
$k_p/(\text{rad}\cdot\text{s})$	0.001	Proportional coefficient of the voltage controller
k_i/rad	0.5	Integral coefficient of the voltage controller

TABLE III
SPECIFICATION OF THE GRID STRENGTH AND CAPACITY FOR
ANALYSIS AND HIL EXPERIMENTAL TESTS (UNIT: P.U.)

Symbol	Value
Case 1: strong grid	
SCR	3.50
ΔP_L	0.60
$S_1^{\max}$	1.40
$S_g^{\max}$	1.36
$P_{e,1}^{\max}$	1.70
$P_{e,g}^{\max}$	1.70
$k_{\theta,pu}^{\max}$	1.50
$k_{\theta,pu}^{\min}$	0.50
Case 2: weak grid	
SCR	1.60
ΔP_L	0.40
$S_1^{\max}$	1.40
$S_g^{\max}$	1.36
$P_{e,1}^{\max}$	1.22
$P_{e,g}^{\max}$	1.30
$k_{\theta,pu}^{\max}$	3.00
$k_{\theta,pu}^{\min}$	0.82

The ROA-inspired transient stability assessment [16], [30] is then conducted based on the proposed large-signal model in Section II-B, and the results are shown in Fig. 4. Since the DC dynamics are included in addition to the angle dynamics, the stability is assessed by projecting the post-disturbance trajectories onto the $\delta - \dot{\delta}$ plane for visualization. For each candidate $k_{\theta,pu}$, the stable equilibrium point (SEP), unstable equilibrium point (UEP), and representative post-disturbance trajectory are presented in Fig. 4. Under the strong grid, an additional DC-aware boundary is introduced to indicate the physical power-feasibility constraint imposed by the DC source.

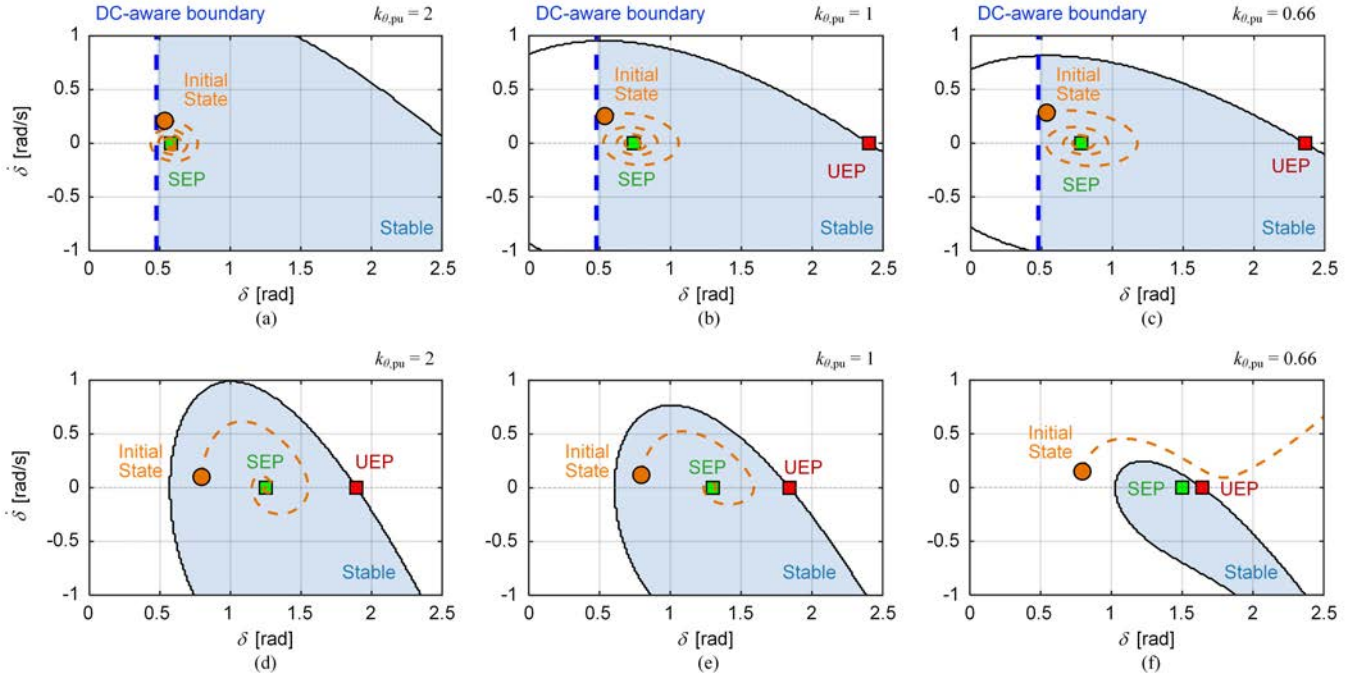


Fig. 4. Constraint-aware ROA-inspired transient stability analysis under different grid strengths and $k_{\theta,pu}$ values (SEP - stable equilibrium point, UEP - unstable equilibrium point). (a)-(c) Strong grid with SCR = 3.50 and $k_{\theta,pu} = 2, 1$, and 0.66, respectively. (d)-(f) Weak grid with SCR = 1.60 and $k_{\theta,pu} = 2, 1$, and 0.66, respectively.

To simplify the power sharing after the adjustment of k_{θ} , the per-unit value of k_{θ} is adopted as

$$k_{\theta,pu} = \frac{k_{\theta}}{k_{\theta,base}} \quad (41)$$

with $k_{\theta,base}$ resulting in a fair active power sharing between the MC inverter and the grid. That is, the equivalent damping coefficient of the MC in the system is twice that of the VSG-controlled GFIs (i.e., GFIs 2 and 3) when $k_{\theta,pu} = 1$.

Under the strong-grid condition, the dominant constraint arises from the DC power limits, as the AC-side transmission capability is sufficiently high. The maximum output power increments of the MC inverter $S_1^{\max}$ and the grid $S_g^{\max}$ are obtained as 0.40 and 0.36 p.u., respectively. Accordingly, the static upper and lower limits of $k_{\theta,pu}$ are derived to be 1.5 and 0.5 p.u. by (35) and (36). The ROA-inspired assessment under the strong grid is shown in Figs. 4(a) to 4(c). In this case, a DC-aware boundary is included to indicate the source-side power-feasibility limit. A trajectory crossing this boundary requires power from the grid $P_{e,g}$ beyond the available capability $P_{e,g}^{\max}$ and therefore becomes infeasible, even if it is still located inside the conventional projected ROA. This explains why an excessively large $k_{\theta,pu}$ can lead to instability under the strong-grid condition.

Under the weak-grid condition, the limiting factor shifts from the DC-side source capability to the AC-side power-transfer capability. The maximum output power increments of the MC inverter $P_{e,1}^{\max}$ and the grid $P_{e,g}^{\max}$ are 0.22 and 0.30 p.u., respectively, yielding the static lower and upper limits of $k_{\theta,pu}$ at 0.82 and 3.0 p.u. The projected stable regions under the weak-grid condition are shown in Figs. 4(d) to 4(f). Different from the strong-grid case, the dominant limitation is directly

related to the AC-side power-angle characteristic. Therefore, the projected ROA in the $\delta - \dot{\delta}$ plane can provide a more intuitive indication of the transient stability margin. It can be observed in Fig. 4 that $k_{\theta,pu} = 1$ and 2 remain inside the recoverable region, whereas $k_{\theta,pu} = 0.66$ loses synchronization. This result confirms that an excessively small $k_{\theta,pu}$, although beneficial for increasing the equivalent inertia and damping according to (25), can become unstable under weak-grid conditions when the required transient power approaches the AC transmission limit.

Based on the above static bounds and the ROA-inspired assessment, the final selection of $k_{\theta,pu}$ should ensure steady-state power feasibility while avoiding operation near the transient stability boundary. In this work, $k_{\theta,pu} = 1$ is selected as a representative design value. Under the strong-grid condition, it remains within the static feasible interval and avoids the DC-aware infeasible boundary, while providing more balanced transient response than the other tested candidates. Under the weak-grid condition, it also stays inside the recoverable region and avoids the AC-side power-transfer boundary. Compared with $k_{\theta,pu} = 2$, the selected value provides better steady-state frequency support according to the MC relationship in (2). Therefore, $k_{\theta,pu} = 1$ offers a practical compromise among steady-state feasibility, transient recoverability, and frequency-support performance under both grid conditions. It should be noted that with the proposed method, the conclusions are not restricted to this configuration and the method remains applicable to heterogeneous GFM systems, parallel MC inverters, and more complex grid topologies.

Furthermore, the interaction between the inner loops and the outer synchronization loop can become more apparent

under a weak grid. Hence, the stability region predicted by the reduced-order model can become overly optimistic, especially when the inner loops are not sufficiently fast [7], [9], [31]. For practical implementation, parameters close to the dynamic stability boundary should be avoided, and a more conservative k_θ should be selected if the inner-loop bandwidth is insufficient.

These observations confirm that the analytically derived bounds define the steady-state feasibility interval, while the ROA-inspired assessment further reveals transient recoverability and the dominant physical constraints across different grid strengths. Although the assessment is performed under the baseline DC-source parameters listed in Table II, variations in the source response time constant τ_{dc} and DC-link capacitance C_{dc} can further reshape the dynamics. To address this, their influences will be examined in Section IV as a robustness verification of the selected parameter.

IV. EXPERIMENTAL TESTS

HIL experimental tests are performed to validate the proposed parameter design. The experiments are conducted on an RTDS-based IEEE 9-bus test system, as shown in Fig. 3, in which the MC-based GFM operates alongside two VSG-controlled units. The configuration of the platform is illustrated in Fig. 3, and the results under a strong and weak grid with $\tau_{dc} = 0.05$ s and $C_{dc} = 8$ mF are shown in Figs. 5 and 6, respectively. To test the stability of the MC inverter with different values of $k_{\theta,pu}$, an extra load is connected at node 5 at t_1 and t_2 . For clarification, the parameters used in the HIL verification align with those in Tables II and III and are consistent with the analysis in Section III-B.

The dynamics under a strong grid with an SCR of 3.50 are benchmarked in Fig. 5, with $k_{\theta,pu}$ being set to 0.66, 1, and 2 p.u., respectively. It can be observed from Fig. 5 that the MC inverter with $k_{\theta,pu}$ of 2 p.u., which exceeds the upper limit (i.e., 1.50 p.u. as listed in Table III), is unable to maintain stable operation during the transient. The damping effect is insufficient due to the unreasonable tuning of $k_{\theta,pu}$, thereby leading to grid collapse. When $k_{\theta,pu}$ becomes 1 p.u., the MC inverter exhibits desirable dynamics with negligible oscillations. However, apparent oscillations and longer settling time are observed when $k_{\theta,pu}$ is further decreased to 0.66 p.u., due to a higher J_v as expressed in (25). Additionally, the DC-link voltage becomes more sensitive to frequency deviation, and the steady-state value decreases as $k_{\theta,pu}$ increases, as expressed in (2). Overall, the MC inverters employing $k_{\theta,pu}$ within the upper and lower limits maintain stable operation during the transient. Moreover, the dynamics are consistent with the analysis in Section III, therefore verifying the effectiveness of the proposed parameter tuning method.

Similarly, the transient responses under a weak grid with an SCR of 1.60 are shown in Fig. 6, with $k_{\theta,pu}$ being set to 0.66, 1, and 2 p.u., respectively. Unlike the strong grid condition in Fig. 5, the limiting constraint shifts from the DC power capability to the AC transmission limit, as the line

inductance increases. That is, the maximum power transmission capacity of the MC inverter $P_{e,1}^{\max}$ is less than the maximum DC power capacity $S_1^{\max}$, due to the increase of $L_{g,1}$ in (27). It can be derived from Fig. 6 that the MC inverter with $k_{\theta,pu}$ being 0.66 p.u. lost synchronization with the grid after the load step at t_2 , which is consistent with the analysis in Section III since the parameter limit is 0.82 to 3 p.u. in this case. This also indicates that the MC inverter with a smaller k_θ , although preferred for higher equivalent inertia and local damping in traditional tuning methods, can still become unstable under weak grid conditions when the transient power demand approaches the power transmission limit. When $k_{\theta,pu}$ is increased to 1 p.u., the system becomes stable after the load step, exhibiting smooth and fast dynamics without transient oscillations. Although a similar trend is observed in the dynamics under $k_{\theta,pu} = 2$ p.u., the steady-state frequency is lower, as shown in Fig. 6(a), indicating reduced frequency-support capability. The experimental results presented in Figs. 5 and 6 have validated the proposed parameter design under a fixed set of DC-source dynamics.

To further examine the robustness of the selected parameter against variations in DC-source dynamics, HIL tests are conducted by increasing the response time constant τ_{dc} from 0.05 s to 0.2 s, while the same candidate values of $k_{\theta,pu}$ are retained. The experimental results under strong- and weak-grid conditions are illustrated in Figs. 7 and 8, respectively. Under the strong-grid condition, as shown in Fig. 7, the MC inverter with $k_{\theta,pu} = 2$ p.u. loses stability after the load step, since this value exceeds the upper bound derived for the strong-grid case. In contrast, the MC inverter maintains stable operation under the slower DC-source response when $k_{\theta,pu} = 1$ or 0.66 p.u., although larger transient deviations and longer settling time are observed compared with the baseline case in Fig. 5. Specifically, the frequency and DC-link voltage exhibit larger transient deviations and longer settling time, implying that an increased τ_{dc} reduces the effective transient damping provided through the DC-side dynamics. Although synchronization is ultimately preserved under a strong grid, the reduced dynamic margin indicates that the system becomes more sensitive to parameter selection. The influence of a larger τ_{dc} becomes more pronounced under the weak-grid condition, as shown in Fig. 8. In this case, both $k_{\theta,pu} = 0.66$ and 2 p.u. fail to maintain synchronization after the load disturbance, whereas $k_{\theta,pu} = 1$ p.u. preserves stable operation. This result indicates that the parameter selected from the baseline ROA-inspired design yields a more robust compromise as the DC-source response slows.

The robustness against the reduced DC-link energy buffer is further evaluated by decreasing C_{dc} from 8 to 2 mF, as shown in Figs. 9 and 10. Under the strong-grid condition in Fig. 9, the reduced capacitance causes larger DC-link voltage fluctuations and more pronounced frequency oscillations. Nevertheless, the properly selected $k_{\theta,pu} = 1$ p.u. can still maintain stable operation, whereas the boundary-violating parameter exhibits a degraded or unstable behavior. This

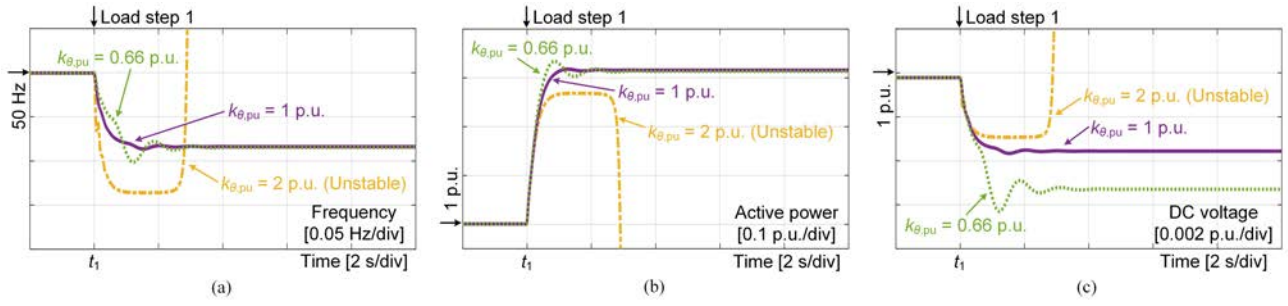


Fig. 5. HIL results of the MC inverter under a strong grid ($SCR = 3.50$) during load step 1 at t_1 (load step 1—a 0.6-p.u. load step as listed in case 1 of Table III) with $\tau_{dc} = 0.05$ s and $C_{dc} = 8$ mF. (a) Frequency. (b) Active power. (c) DC-link voltage.

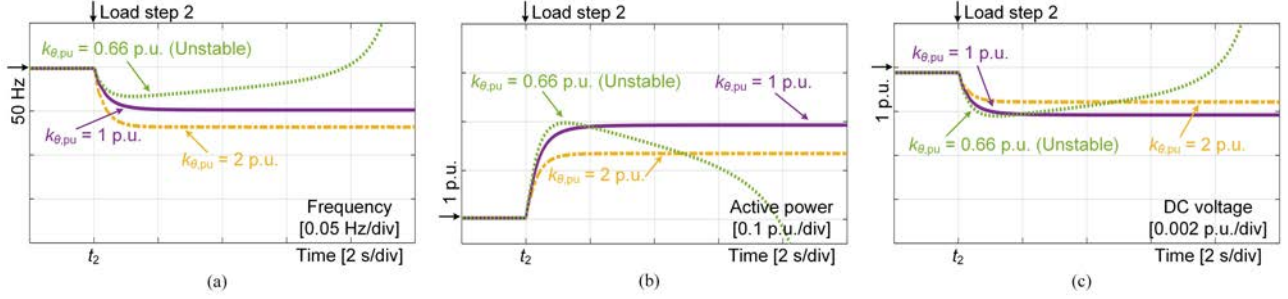


Fig. 6. HIL results of the MC inverter under a weak grid ($SCR = 1.60$) during load step 2 at t_2 (load step 2—a 0.4-p.u. load step as listed in case 2 of Table III) with $\tau_{dc} = 0.05$ s and $C_{dc} = 8$ mF. (a) Frequency. (b) Active power. (c) DC-link voltage.

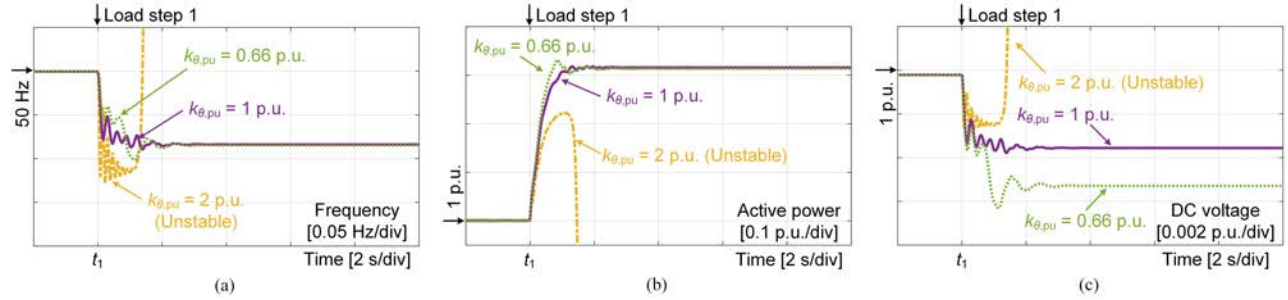


Fig. 7. HIL results of the MC inverter under a strong grid ($SCR = 3.50$) during load step 1 at t_1 (load step 1—a 0.6-p.u. load step as listed in case 1 of Table III) with $\tau_{dc} = 0.2$ s and $C_{dc} = 8$ mF. (a) Frequency. (b) Active power. (c) DC-link voltage.

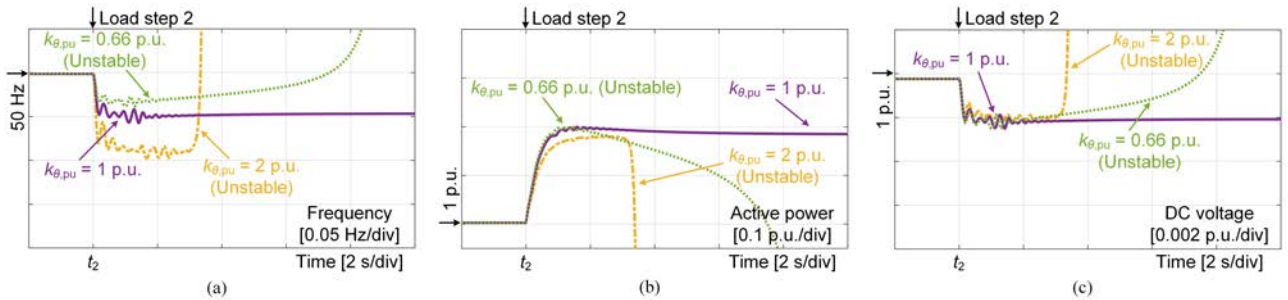


Fig. 8. HIL results of the MC inverter under a weak grid ($SCR = 1.60$) during load step 2 at t_2 (load step 2—a 0.4-p.u. load step as listed in case 2 of Table III) with $\tau_{dc} = 0.2$ s and $C_{dc} = 8$ mF. (a) Frequency. (b) Active power. (c) DC-link voltage.

behavior aligns with the physical interpretation that a smaller C_{dc} limits the instantaneous energy-support capability, thereby intensifying the coupling between DC-voltage dynamics and the MC synchronization mechanism. Under the weak-grid condition, as shown in Fig. 10, the reduced DC-link capacitance further aggravates the transient response. The cases with $k_{\theta,pu} = 0.66$ and 2 p.u. lose stability, while $k_{\theta,pu} = 1$ p.u. remains the most reliable choice among the tested parameters. This confirms that the proposed parameter-

selection method provides a practical and robust design guideline, even when the DC-side energy buffer is reduced. Furthermore, a more pronounced stability degradation is observed under weak-grid conditions when comparing Figs. 7 and 8, as well as between Figs. 9 and 10. This can be attributed to the larger phase-angle excursion and smaller stability margin under weak grids, as well as their greater sensitivity to transient power imbalances due to weaker synchronization capability [32].

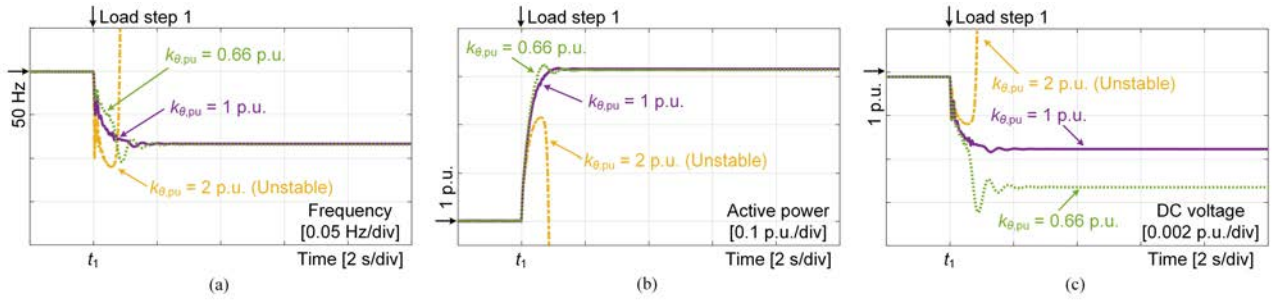


Fig. 9. HIL results of the MC inverter under a strong grid ($SCR = 3.50$) during load step 1 at t_1 (load step 1—a 0.6-p.u. load step as listed in Case 1 of Table III) with $\tau_{dc} = 0.05$ s and $C_{dc} = 2$ mF. (a) Frequency. (b) Active power. (c) DC-link voltage.

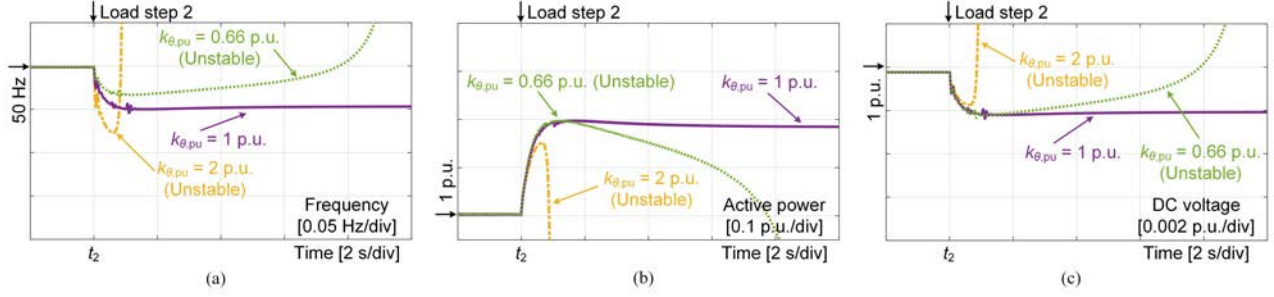


Fig. 10. HIL results of the MC inverter under a weak grid ($SCR = 1.60$) during load step 2 at t_2 (load step 2—a 0.4-p.u. load step as listed in Case 2 of Table III) with $\tau_{dc} = 0.05$ s and $C_{dc} = 2$ mF. (a) Frequency. (b) Active power. (c) DC-link voltage.

In all, the effectiveness of the proposed parameter-design method is confirmed across both strong- and weak-grid conditions. The static bounds and ROA-inspired assessment explain the stability differences observed under the baseline DC-source parameters, while the additional HIL tests with varied τ_{dc} and C_{dc} further demonstrate the robustness of the selected parameter. Selecting $k_{\theta,pu} = 1$ provides a practical compromise between frequency-support capability and transient stability in the case study.

V. CONCLUSION

This paper has presented a comprehensive methodology for designing the proportional gain k_{θ} in MC inverters by explicitly incorporating DC-link voltage dynamics into the analysis. A large-signal model was formulated to capture the interaction between DC-link voltage dynamics, source limitations, and grid behavior. With this model, analytical upper and lower bounds for the control gain were derived using DC-side power constraints and AC-side transmission limits. The proposed approach was validated through HIL experimental tests on an RTDS-based platform using the IEEE 9-bus test system. The results have demonstrated that the analytically derived bounds, combined with the ROA-inspired assessment, effectively explain the transient stability behavior under the baseline DC-source parameters and across different grid strengths. Additional HIL tests with increased DC-source response time and reduced DC-link capacitance further confirm the robustness of the selected control gain under variations in DC-side dynamics. In all, the findings provide practical guidance for tuning the MC parameter in IBR-based power systems, highlighting the importance of DC-dynamic-aware design to achieve robust frequency regulation and transient performance across various grid strengths.

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