

Modeling and control of power electronics for renewables

Edited by

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Modeling and control of power electronics for renewables

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Editorial: Modeling and control of power electronics for renewables

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KEYWORDS

power electronics, modeling, grid stability, inverter control, renewable power generation

Editorial on the Research Topic

Modeling and control of power electronics for renewables

Power electronics is used in a wide range of applications, including switched-mode power supplies, motor drives, active power filters, and renewable power generation (RPG). In recent years, power electronics has experienced a rapid increase in use as the grid interface of RPG systems. However, as the penetration level of renewable energy grows significantly, increasing challenges have emerged, e.g., weak grid stability issues, high-frequency circulations, active grid support function, and arc faults. As a result, further tasks have been brought to the table to enhance the modeling and control of power electronics, and in-depth discussions on these issues are urgently required to provide technical support for the global energy transition.

The main purpose of this Special Section is then to collect the recent advances in the modeling and control of power electronics converters, as well as to provide the researchers and engineers with how to analyze and model converter behaviors in order to improve their design and operation. A series of controls specifically designed for use with power converters to address emerging challenges like the weak grid stability issues, active grid support, arc detection, etc., are selectively archived in this Special Section. In general, this Research Topic provides an overview of the state-of-the-art in modeling and controlling power electronics for renewable energy power systems.

This Research Topic is divided into four thematic areas (“optimization and control of photovoltaic (PV) systems”, “technologies for enhancing grid stability”, “modeling for transient characteristics and electromagnetic interference (EMI) analysis”, and “cutting-edge intelligent control algorithms”), and this Special Section has collected 12 articles.

The first area focuses on optimizing and controlling photovoltaic systems. Zhang *et al.* presented “A ground current suppression method for systems with a large number of photovoltaic (PV) inverters,” in which a hardware-software approach is proposed. The hardware solution diverts ground current to the DC bus via filter capacitors, while

the software-based mitigation strategy (carrier phase shifting and frequency shifting) is synthesized to mutually cancel zero-sequence currents across multiple inverters. This method addresses escalating ground current issues in large-scale PV plants, and it has been experimentally validated to reduce leakage currents and stabilize grid-connected operation without the need for grid neutral-point access. Moreover, in the paper “An operating mode control method for photovoltaic (PV) battery hybrid systems,” Zhang et al. create a generalized control architecture for PV–battery hybrid systems that improves dynamic response performance by allowing smooth switching among six grid-connected and islanding modes without requiring control loop reconfiguration. In addition, a differential power processing (DPP) architecture with unit-power balancing control is designed by Ni et al. in their paper “Unit Power Rating Balancing for Differential Power Processing-Based Distributed Photovoltaic Systems.” By coordinating maximum power point tracking (MPPT) and unit balance point tracking (UBPT) units, they successfully reduce the maximum processing power of each power-processing unit in distributed PV systems.

The second part of this Special Section focuses on grid stability-enhancing technologies. The paper “Megawatt-level converter grid-forming control technology by adopting MPC for medium voltage distribution network,” Liu et al. proposes an enhanced grid-forming control strategy that integrates virtual admittance with model predictive control to improve several key grid-integration performance metrics such as the power quality, fault ride-through capability, dynamic response, and transient overcurrent limitations in weak grid scenarios. The technique can support carbon reduction in DC industrial parks by demonstrating stable and efficient current restriction under different grid strengths. Moreover, in order to greatly increase the stability margin of AC/DC hybrid grids, Zhou et al. propose a multi-resource collaborative damping control strategy that combines energy storage and high voltage direct current (HVDC) into a multi-input multi-output control (MIMO) system. The proposal is presented in their paper “Research on the multi-objective collaboration damping control strategy based on multi-resource” (Zhou et al.). Gao et al. presented “A Review of Dynamic Voltage Support for Power Grids with Large-Scale Penetration of Renewable Generation,” analyzing dynamic voltage stabilization mechanisms, aggregation modeling approaches for renewable plants, and innovations in coordinated control of multi-type reactive power sources using autonomous decentralized strategies. The review highlights challenges in transient voltage stability while emphasizing converter-based reactive support capabilities and future research directions for renewable-dominated grids. Additionally, a grid-forming control is used to modify reactive-power droop coefficient, significantly raising the static voltage stability limit of renewable energy power plants Wentao et al. “Research on static voltage stability enhancement for new energy station based on grid-forming control strategy”.

The third area focuses on modeling and analysis of transient behavior, EMI characteristics, and device reliability under extreme conditions. By creating a dynamic equivalent model of the current inner loop, the paper “The impact of current-loop control parameters

on the electromagnetic transient voltage performance of voltage-source converter,” by Ding et al. clarifies how control parameters affect transient-voltage behavior and proposes a bandwidth-optimization strategy to enhance high-frequency stability. In the paper of Liu et al. “Time domain analysis of flyback EMI based on distributed parameters theory,” a distributed-parameter model for flyback converters is established, explaining how leakage inductance and parasitic capacitance shape the EMI spectrum and guiding EMI-suppression design for high-density power supplies. Furthermore, Wen et al., in “Exploring the Performance of GaN Trench CAVETs from Cryogenic to Elevated Temperatures,” conduct extensive electrical characterization, demonstrating improved 2DEG mobility and reduced on-resistance at cryogenic temperatures, device survival up to 500°C, and severe degradation at 800°C, highlighting the potential of GaN CAVET for extreme-temperature power-electronics applications.

The last area focuses on cutting-edge intelligent control algorithms. Wang et al. “Load frequency optimal control of the hydropower-photovoltaic hybrid microgrid system based on the off-policy integral reinforcement learning algorithm,” apply off-policy integral reinforcement learning to water-solar complementary microgrid frequency control, overcoming the “curse of dimensionality” in traditional dynamic programming and enabling disturbance-adaptive regulation. Fen et al. “A Coordinated Power Quality Improvement Control Strategy for AC/DC Hybrid Distribution Networks based on Three-Phase Four-Leg Flexible Interconnection Converter,” propose a coordinated control strategy for a three-phase four-leg flexible interconnection converter that achieves full compensation of AC three-phase imbalance while also mitigating DC double-frequency ripples.

Oriented for addressing the emerging issues of RPG systems, i.e., from suppressing ground currents and managing grid-forming converter dynamics to analyzing high-frequency EMI, improving multi-resource damping, and ensuring static voltage stability, the contributions in this Research Topic present cutting-edge control, modeling, and optimization strategies. It is shown how advanced predictive, data-driven, and reinforcement-learning methods, combined with innovative hybrid converter designs, can improve PV-battery performance, coordinate AC/DC power quality, and streamline differential power processing. These works, which combine hardware-software co-design with flexible architectures, provide practical solutions for increasing efficiency, reliability, and stability across a variety of grid scenarios, and highlight the critical role of sophisticated power electronics control in advancing sustainable energy systems.

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An operating mode control method for photovoltaic (PV) battery hybrid systems

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Depending on the PV power, load power, and battery status, the system may operate in different modes. The control loop may have to switch between operating modes. In practice, it is difficult to implement control loop switching because the transition and dynamic process are difficult to control. As a result, this paper presents a generalized mode control method that avoids loop switching across modes. First, system structure and topology are introduced. The operating conditions for both grid-connected and off-grid modes are then divided into six sub-cases. Furthermore, the control architecture, control loop, and reference transition for various scenarios are described. Finally, an experimental platform is built, and the results are presented to verify the proposed method.

KEYWORDS

PV, battery, mode control, transition, grid-connected

1 Introduction

As part of the global green mission, an increasing number of renewable energy sources are being installed. As renewable energy sources gain traction, power electronic converters become increasingly popular (Zhang et al., 2023a). The combination of solar panels and energy storage will be a trend in future energy development, and many experts have conducted extensive research on photovoltaic and energy storage hybrid systems (Zhang et al., 2023b). The system operates in a variety of modes depending on the operational conditions of photovoltaic, energy storage, and the power grid. The seamless transition between modes has also become a research topic (Hmad et al., 2023).

Several papers investigate the transition between grid-connected and offline modes; one of the mode transition control strategies relies on a virtual switch located between two functional modes (Balaguer-Alvarez et al., 2014). During grid-connected mode, the inverter functions solely as a current source, utilizing the current controller loop. When an islanding condition is detected, the virtual switch is assigned to the voltage loop. As a result, the inverter operates as a voltage source (Qinfei et al., 2017). The anti-islanding method has a strong influence on the transition performance. The majority of islanding detection methods rely on continuous monitoring of system characteristics at the point of common coupling (PCC), including current, voltage, frequency, and harmonics (Ahmad et al., 2013). These transition approaches can be classified into three categories: passive, active, and hybrid (Koochi-Kamali and Rahim, 2016; Aillane et al., 2023) presents an improved mode transition approach. During the transition, a super-twisting algorithm approach was used to ensure load voltage and manage the inverter's

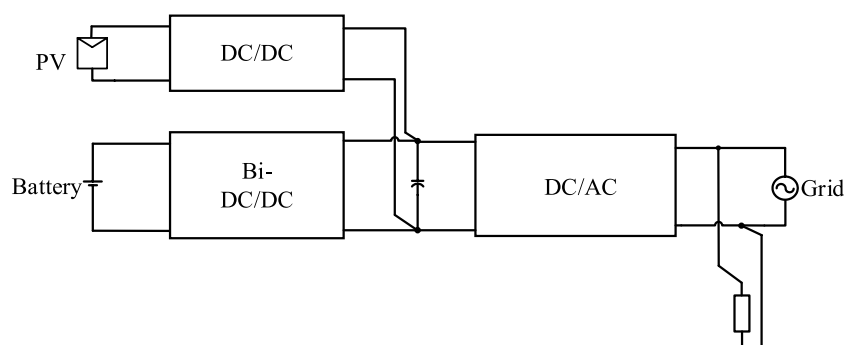


FIGURE 1
System structure.

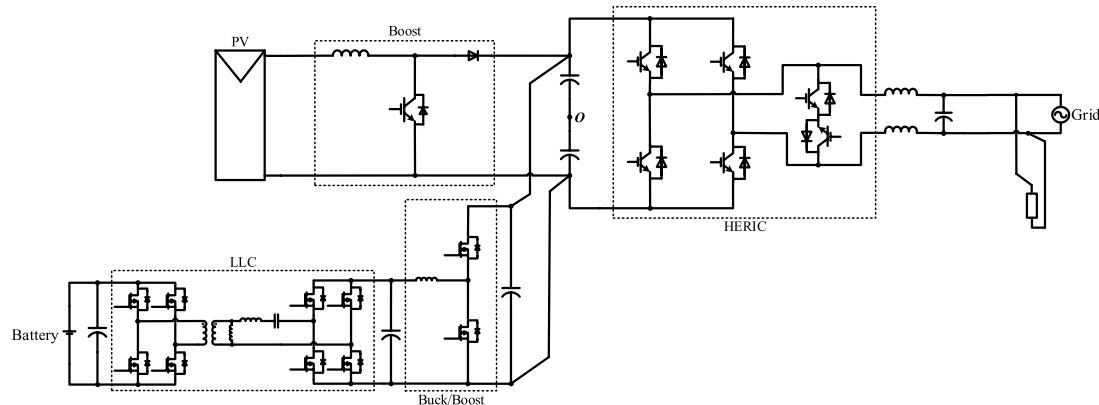


FIGURE 2
System topology.

nonlinear properties, resulting in a smooth transition with minimal disruption impacts. [Ashabani and Mohamed \(2014\)](#) investigates another switching-based method. Three different approaches are considered: power drooping current controlled, power drooping voltage controlled, and current drooping voltage controlled. Furthermore, to aid smooth mode transitions by dampening power, frequency, voltage, and current signals, an approach based on a supplement controller employing port-controlled hamiltonian (PCH) modeling and control was developed in ([Azimi and Lotfifard, 2021](#)). Another approach to achieving a seamless transition is to employ an extra distributed generation (DG) system, which is often a specialized storage unit ([Jihed et al., 2019](#)). The premise behind this technique entailed deploying a supplemental energy unit to relieve the transient, which, however, incurs additional expenditures.

The second set of solutions for dealing with the transition problem includes the use of an extra feedforward compensator ([Tran et al., 2013](#)). The basic idea behind this control structure group is to include a voltage loop in addition to the current loop when a grid fault occurs. In [Hwang and Park \(2013\)](#), an improved seamless transition based on a phase-locked loop (PLL) mechanism was developed for a three-phase grid-connected inverter. The control strategy entails modifying the PLL depending on the operating mode, synchronizing the output inverter voltage to the

grid voltage in grid-connected mode, and generating an angle at the required frequency in off-grid mode. In [Harirchi et al. \(2015\)](#), a similar PLL-based transfer was performed, with feedforward voltage used to mitigate the transition's negative effects on a three-phase grid-connected PV inverter.

Various research studies have developed transition control without reconfiguring the control structure, also known as unified control ([Yi et al., 2018](#)). The overall goal of this technique was to use the voltage control loop as a reference current generator when connected to the grid and as a voltage regulator when off-grid. Unlike previous transition structures, this technique does not necessitate any changes at the control level. Variants of the universal control system have been proposed as a solution to transition issues. In [Liu and Liu \(2014\)](#), the authors developed an indirect current control loop with an applied voltage loop for a three-phase inverter to ensure smooth transfer. In [Sowa et al. \(2021\)](#), a three-phase universal controller for flexible microgrids was presented, ensuring operation in all operating modes without the need for control structure reconfiguration. In [Yi et al. \(2018\)](#), the authors described a unified control and power management system for a hybrid PV-battery application that included both DC and AC charging buses. In [Li et al. \(2020\)](#), a non-linear-simplex method was proposed for determining the

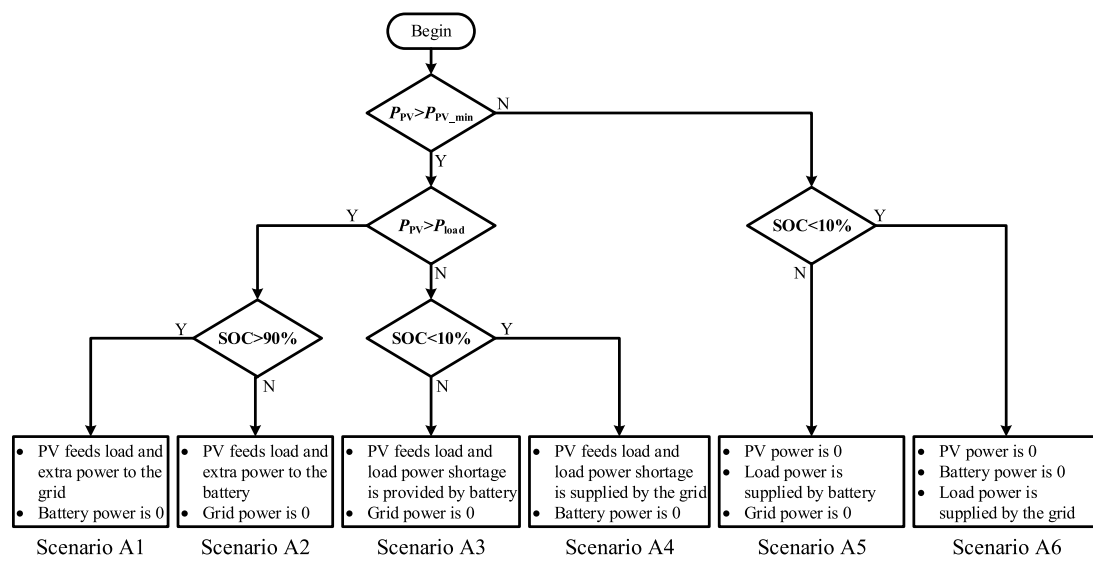


FIGURE 3
Six different scenarios under the grid-connected mode.

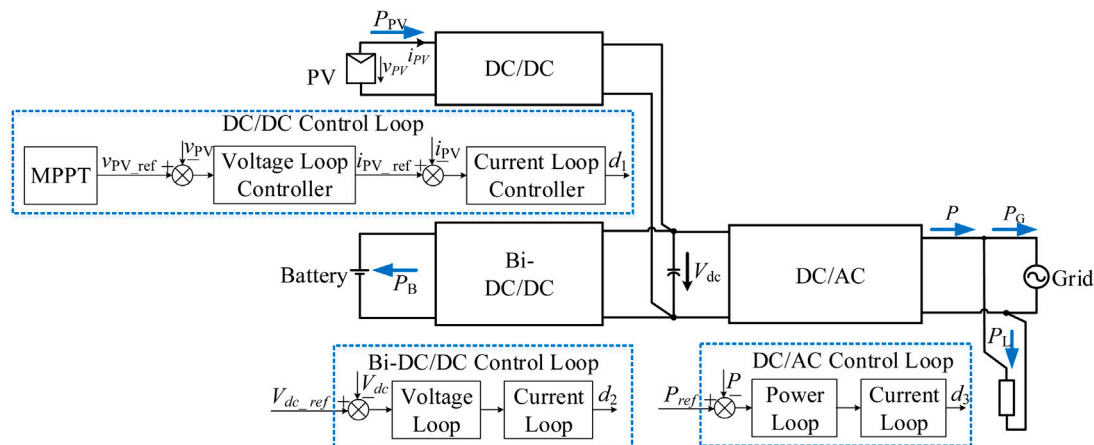


FIGURE 4
System control architecture under grid-connected mode.

optimal controller settings with the goal of reducing voltage variation and achieving a seamless state transition. In the literature (Singh et al., 2017; Jihed et al., 2019), a unified system with droop control approach is investigated. The concept of droop methods has been widely applied to the parallel operation of DG inverters with voltage and/or current control loops. However, one major disadvantage of this method is its poor dynamic performance.

PV-battery hybrid systems operate in a variety of scenarios based on PV power, load power, and battery status, and the control loop may need to switch. Because transition and dynamic processes are difficult to control, it is difficult to implement control loop switching in practice. As a result, this paper proposes a generalized mode control method that avoids loop switching in a variety of scenarios. The main contributions are as follows.

- 1) This paper describes a generalized operating mode control method. The conditions in grid-connected and off-grid modes are classified into six scenarios based on the values of PV power P_{PV} , load power P_{load} , battery state of charge (SOC), and so on. Furthermore, the control architecture, control loop, and reference transition for various scenarios are discussed.
- 2) In grid-connected mode, the battery-side DC/DC controls the DC-bus voltage, the PV-side DC/DC applies Maximum Power Point Tracking (MPPT) mode to maximize PV output power, and the grid-side DC/AC controls the output power. The control loops for DC/DC and Bi-DC/DC do not need to change in different operating scenarios; only the DC/AC output power reference does.
- 3) In off-grid mode, the battery-side DC/DC regulates the DC-bus voltage, the PV-side DC/DC uses power point tracking (PPT) mode to track the PV output power reference, and the

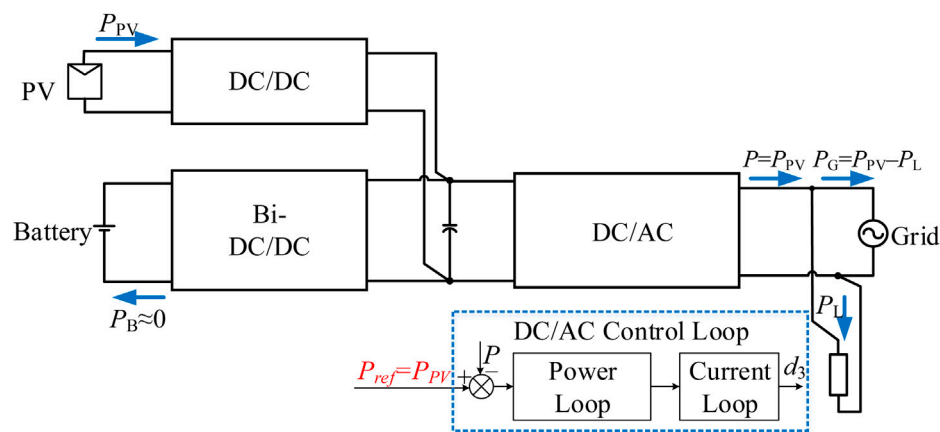


FIGURE 5
Scenario A1.

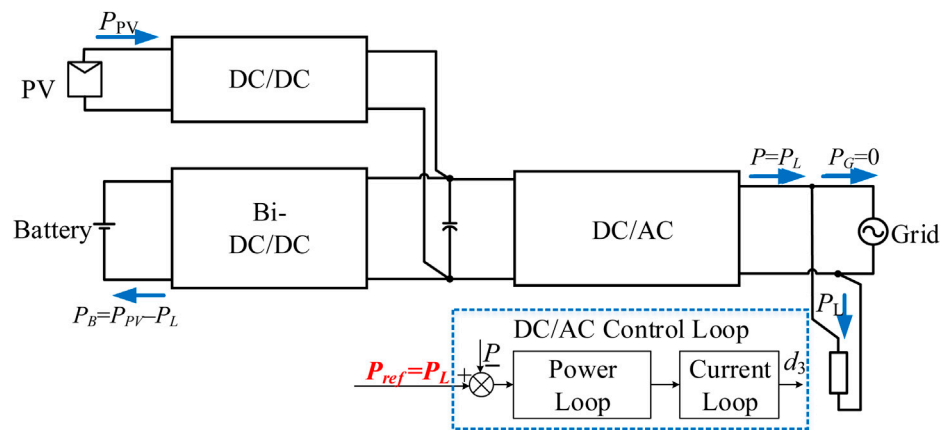


FIGURE 6
Scenario A2.

grid-side DC/AC regulates the output voltage. The DC/AC and Bi-DC/DC control loops do not need to change in different operating scenarios, while only the PV-side DC/DC power reference does.

- 4) An experimental platform is built, and the results are presented to support the proposed method.

2 System structure and topology

Figure 1 depicts the architecture of a typical PV-battery hybrid system, which is a common DC-bus structure. PVs, batteries, and the grid and load are connected to the DC-bus through DC/DC, bidirectional DC/DC, and DC/AC, respectively.

Figure 2 presents the system topology. The Boost topology is used for the PV-side DC/DC. A two-stage structure with LLC plus Buck/Boost is used for the battery-side Bi-DC/DC. The Highly Efficient Reliable Inverter Concept (HERIC) topology is chosen for the DC/AC side.

There are two operating modes for the system, which are grid-connected mode and off-grid mode. In the grid-connected mode, the grid is normal, and the system operates in current source mode, feeding power to the grid. In off-grid mode, the grid is not present, and the system operates in voltage source mode to ensure load voltage. Furthermore, depending on PV power, load power, and battery status, the system operates in different scenarios. Furthermore, the control loop may have to switch for different scenarios. Since the transition and dynamic process is difficult to control, it is relatively challenging to implement the control loop switching. Therefore, the following presents a generalized mode control method for avoiding loop switching for different scenarios.

3 System control

Before introducing specific methods, we set basic control prerequisites first. One of them is the priorities of power sources, including three points.

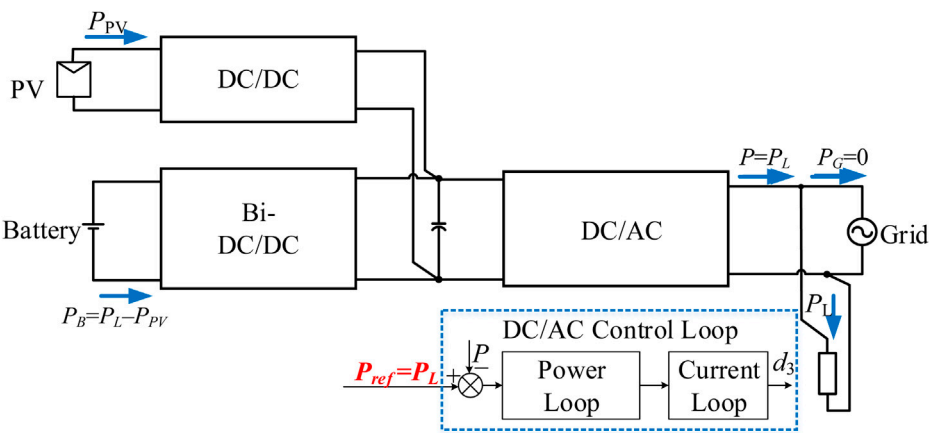


FIGURE 7
Scenario A3.

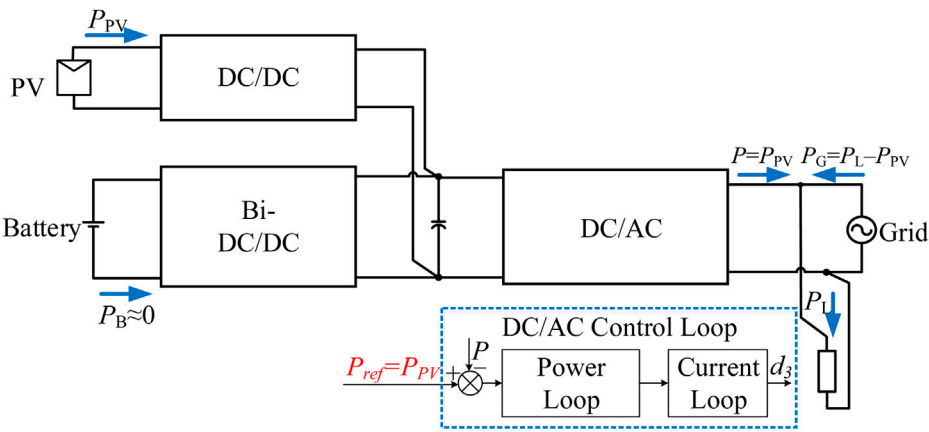


FIGURE 8
Scenario A4.

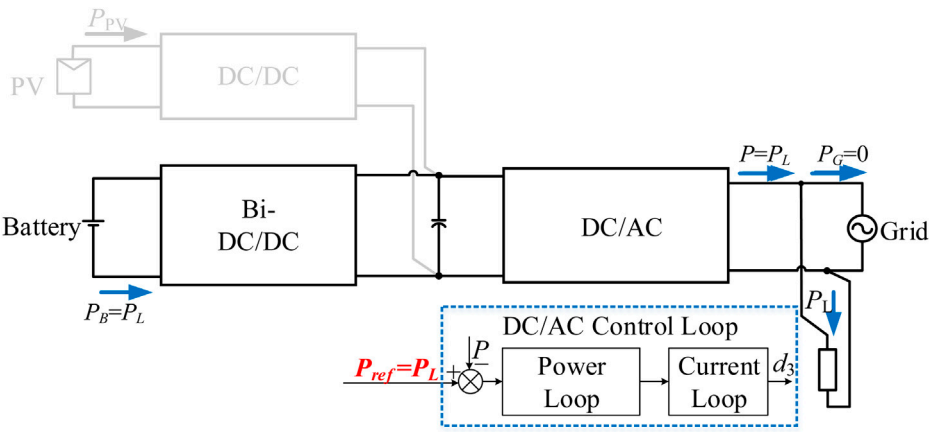


FIGURE 9
Scenario A5.

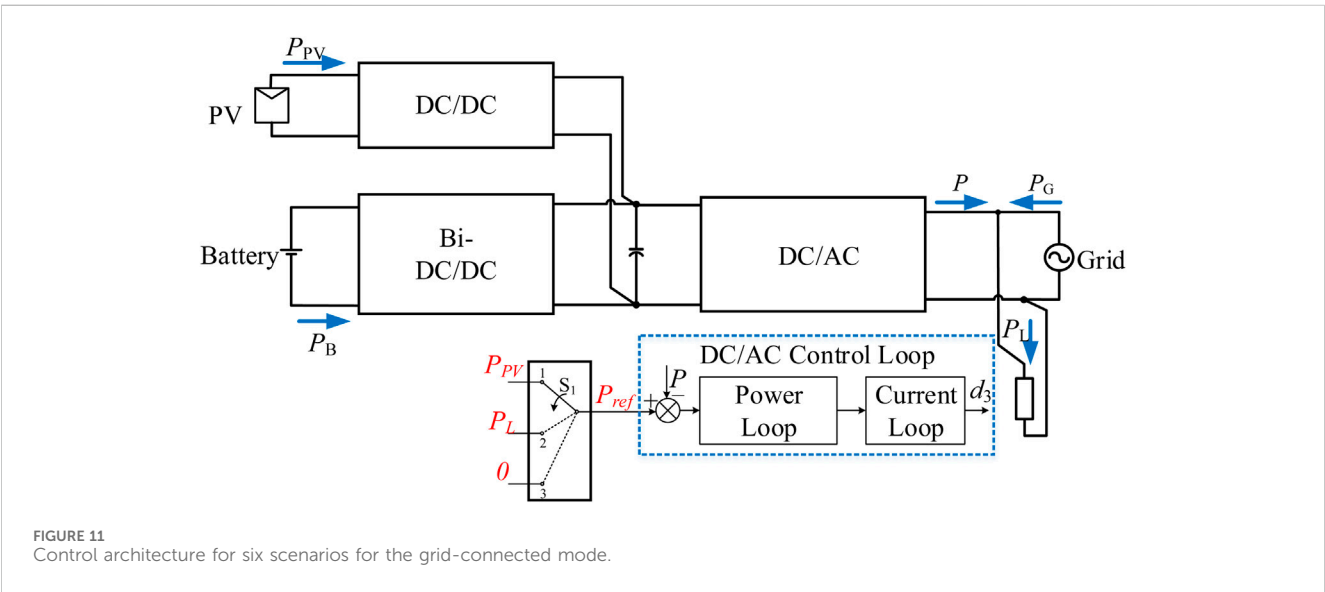
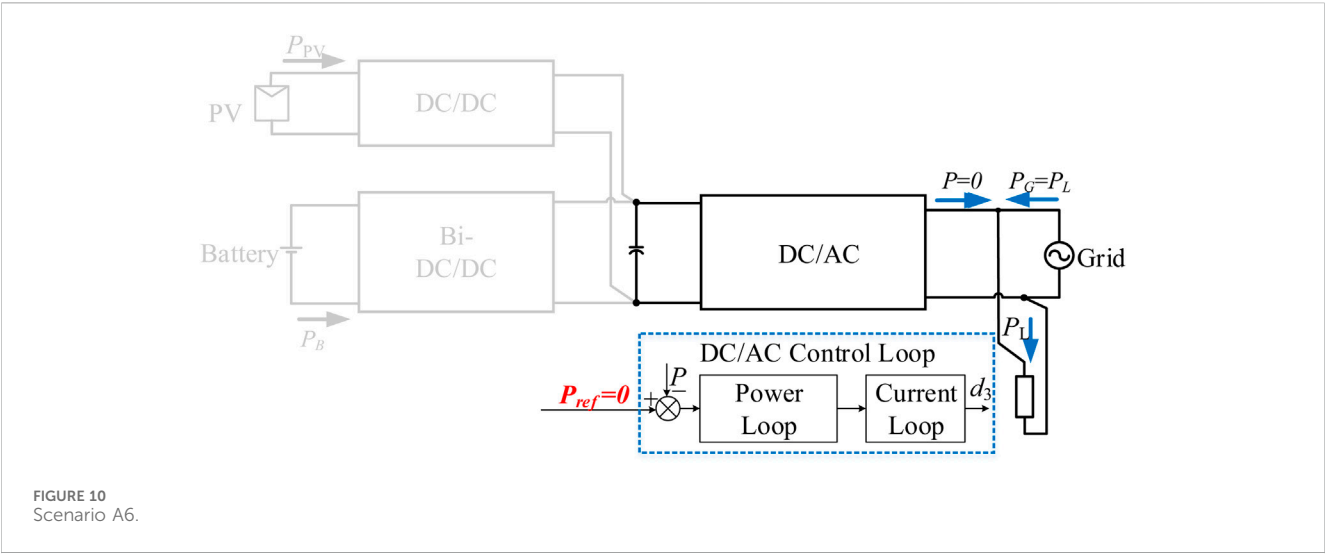


TABLE 1 Summary of six scenarios for the grid-connected mode.

Scenarios	S_1	DC/AC power reference P_{ref}
Scenario A1	"1"	P_{PV}
Scenario A2	"2"	P_L
Scenario A3	"2"	P_L
Scenario A4	"1"	P_{PV}
Scenario A5	"2"	P_L
Scenario A6	"3"	0

- 1) PVs are set as the first priority power source. The output power of PV modules should be prioritized to feed the load.
- 2) Batteries are set as the second priority power source. When there is a surplus or shortage of PV power to the load, the battery is then used to achieve power balance.

- 3) The grid is set as the third priority power source. When both the PV modules and the batteries reach their limits, the grid is then employed to power the load.

Furthermore, the output power of PV modules, the SOC of batteries, and the charging and discharging power of batteries are restricted as follows.

- 1) When the PV output power P_{PV} is less than the set minimum threshold P_{PV_min} , the PV is considered to have no power and the PV side is shut off; otherwise, the PV side should be used to generate power.
- 2) If the battery SOC is greater than 90%, the battery should not be charged further; if the battery SOC is less than 10%, the battery should not be drained further.

The following will analyze the grid-connected mode and off-grid mode, respectively.

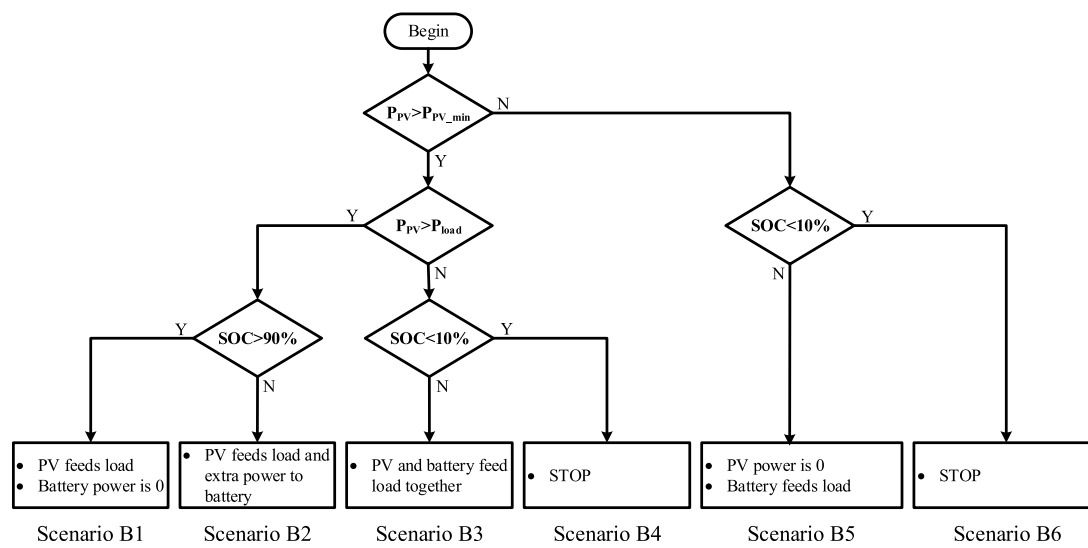


FIGURE 12
Six different scenarios under the off-grid mode.

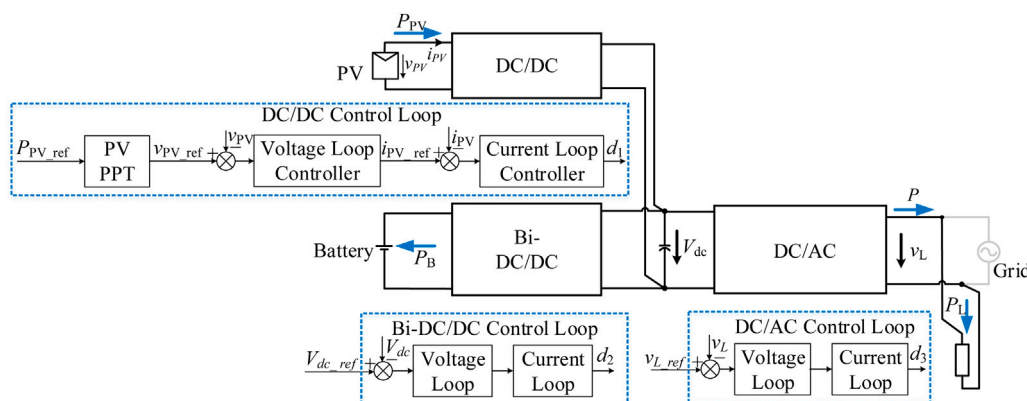


FIGURE 13
System control architecture under off-grid mode.

3.1 Grid-connected mode

Based on the values of PV power P_{PV} , load power P_{load} , battery SOC, etc., the operating conditions under the grid-connected mode are divided into six scenarios, as shown in Figure 3.

Figure 4 depicts the control architecture in grid-connected mode, where the battery-side DC/DC controls the DC-bus voltage V_{dc} , the PV-side DC/DC operates in MPPT mode to maximize the PV output power P_{PV} , and the grid side DC/AC controls the output power P . The advantage of the control architecture shown in Figure 4 is that the control loops for DC/DC and Bi-DC/DC do not need to change in different operating scenarios, with only the DC/AC output power reference P_{ref} needed to change. This reduces the control loops switching during different scenarios, making it simple to implement.

It is worth mentioning that, in this paper, v_{pv} is PV voltage, v_{pv_ref} is PV voltage reference, i_{pv} is PV current, i_{pv_ref} is PV current reference, V_{dc_ref} is DC-bus voltage reference, d_1 is output duty cycle of the DC/DC control loop, d_2 is output duty cycle of the Bi-DC/DC control loop, and d_3 is output duty cycle of the DC/AC control loop.

3.1.1 Scenario A1

In this scenario, $P_{PV} > P_{Load}$ and $SOC > 90\%$, where the PV power is larger than the load power and the battery cannot be charged. In the control loop, as shown in Figure 5, the DC/AC power reference P_{ref} is set to P_{PV} . This can result in that the DC/AC outputs the PV-side power P_{PV} and the battery power P_B is zero. Furthermore, the PV powers the load P_L , and the extra power $(P_{PV} - P_L)$ is fed to the grid, and the grid power is P_G .

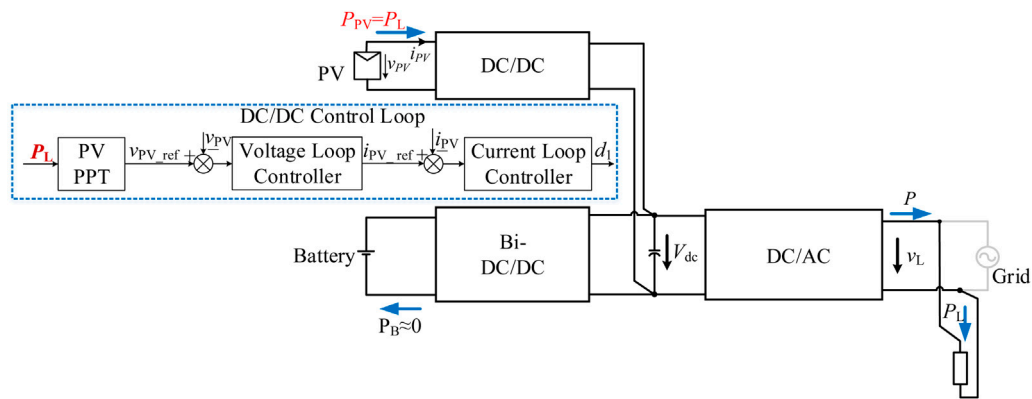


FIGURE 14
Scenario B1.

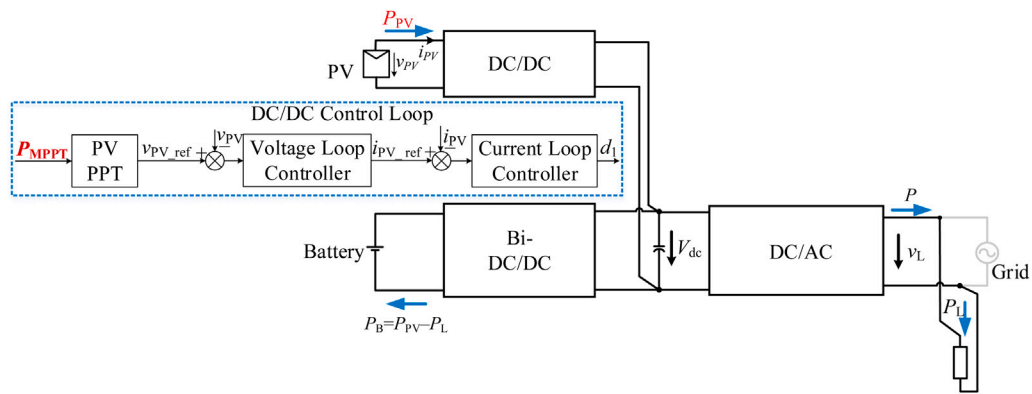


FIGURE 15
Scenario B2.

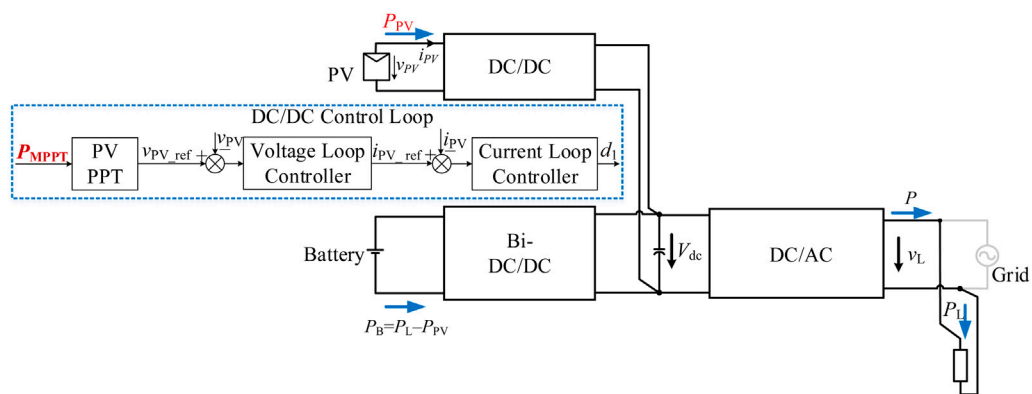


FIGURE 16
Scenario B3.

3.1.2 Scenario A2

In this scenario, $P_{PV} > P_{Load}$ and $SOC \leq 90\%$, where the PV power is higher than the load power and the battery can be charged. In the control loop, as shown in Figure 6, the DC/AC power

reference P_{ref} is set to P_L . This can result in that the DC/AC outputs the load power P_L and the power fed into the grid is 0. Furthermore, the battery automatically absorbs the remaining power generated by PV, which is $P_{PV} - P_L$.

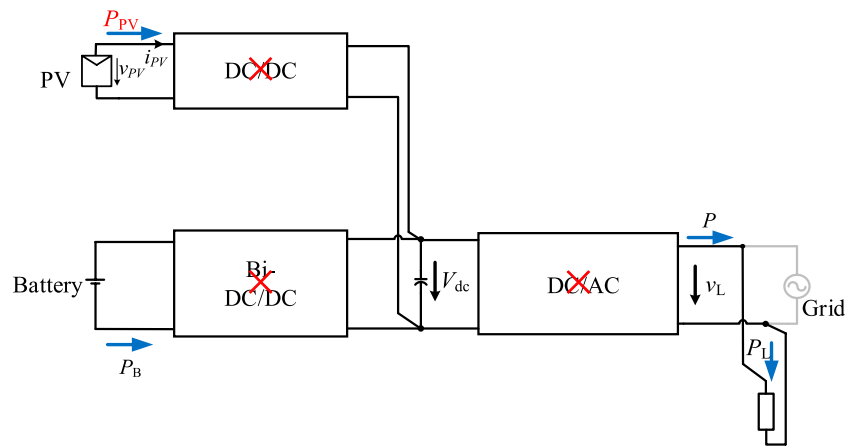


FIGURE 17
Scenario B4.

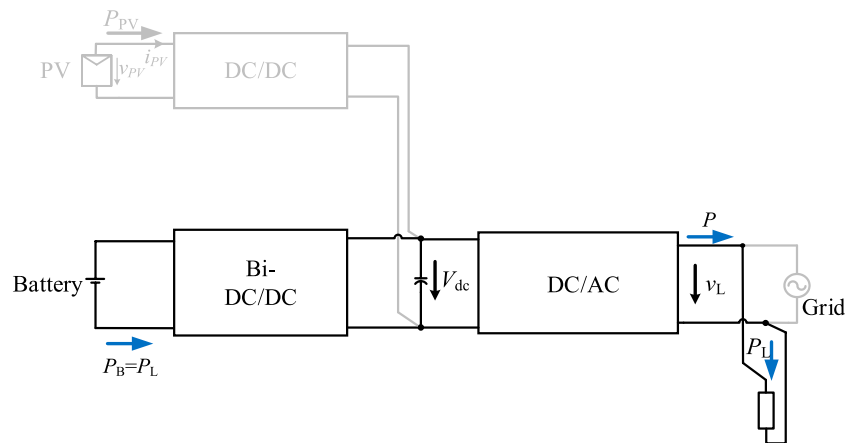


FIGURE 18
Scenario B5.

3.1.3 Scenario A3

In this scenario, $P_{PV} \leq P_{Load}$ and $SOC > 10\%$, where the PV power is less than the load power and the battery can be discharged. In the control loop, as shown in Figure 7, the DC/AC power reference P_{ref} is set to P_L . This can result in that the DC/AC outputs the load power P_L and the power fed into the grid is 0. Furthermore, the battery automatically provides the remaining power $P_L - P_{PV}$.

3.1.4 Scenario A4

In this scenario, $P_{PV} \leq P_{Load}$ and $SOC \leq 10\%$, where the PV power is less than the load power and the battery cannot be discharged. In the control loop, as shown in Figure 8, the DC/AC power reference P_{ref} is set to P_{PV} . This can result in that the DC/AC outputs the PV power P_{PV} and the power that the battery provides is 0. Furthermore, the grid provides the remaining load power $P_L - P_{PV}$.

3.1.5 Scenario A5

In this scenario, $P_{PV} \leq P_{PV_{min}}$ and $SOC > 10\%$, where the PV cannot generate power and the battery can be discharged to feed the load. In the control loop, as shown in Figure 9, the DC/AC power reference P_{ref} is set to P_L . This can result in that the DC/AC outputs the load power P_L and the power from the grid is 0. Furthermore, the battery automatically provides the load power P_L .

3.1.6 Scenario A6

In this scenario, $P_{PV} \leq P_{PV_{min}}$ and $SOC \leq 10\%$, where the PV cannot generate power and the battery cannot be discharged to feed the load. Thus, the load power can only be provided by the grid. In the control loop, as shown in Figure 10, the DC/AC power reference P_{ref} is set to 0. This can result in that the DC/AC outputs no power and the load power is from the grid.

The summary of six scenarios for the grid-connected mode is shown in Figure 11. From it, a switch S_1 is introduced to assign different DC/AC power references. In scenarios A1 and A4, the

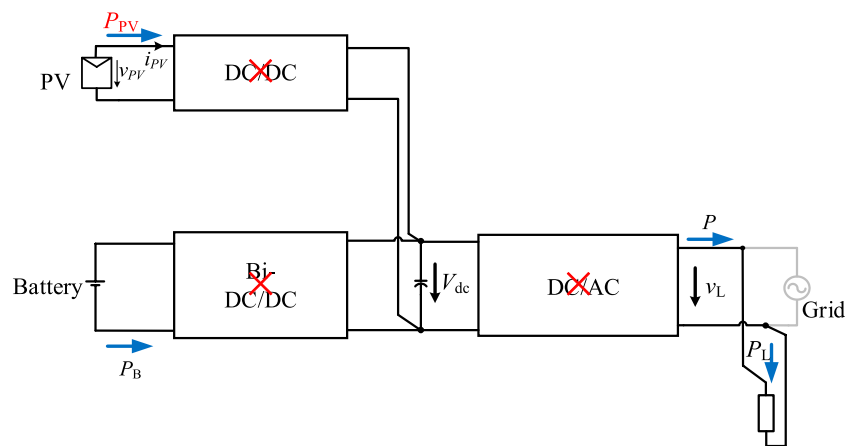


FIGURE 19
Scenario B6.

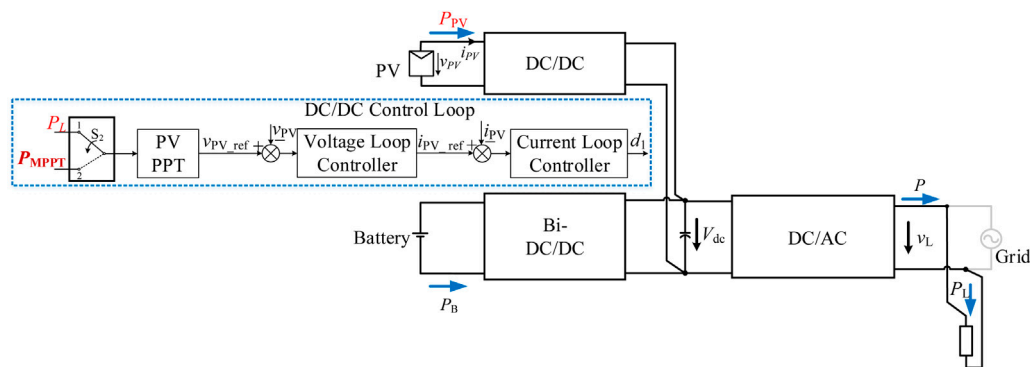


FIGURE 20
Control architecture for six scenarios for the off-grid mode.

TABLE 2 Summary of six scenarios for the off-grid mode.

Scenarios	S_2	DC/AC power reference
Scenario B1	"1"	P_L
Scenario B2	"2"	P_{MPPT}
Scenario B3	"2"	P_{MPPT}
Scenario B4	—	System stops
Scenario B5	—	PV stops
Scenario B6	—	System stops

switch S_1 is placed to "1". In scenarios A2, A3, and A5, the switch S_1 is placed to "2". In scenario A6, the switch S_1 is placed to "3". The details are presented in Table 1.

3.2 Off-grid mode

Based on the values of PV power P_{PV} , load power P_{load} , battery SOC, etc., the operating conditions under the off-grid

mode are divided into six different scenarios, as shown in Figure 12.

Figure 13 depicts the control mode in off-grid mode, where the battery-side Bi-DC/DC controls the DC-bus voltage V_{dc} , the PV-side DC/DC operates in power point tracking (PPT) mode to follow the PV output power P_{PV} , and the grid-side DC/AC controls the output voltage v_L . The advantage of the control architecture shown in Figure 13 is that the DC/AC and Bi-DC/DC control loops do not need to change in different operating scenarios, with only the PV-side DC/DC power reference P_{PV_ref} needed to change. This reduces the control switching during different scenarios, making it simple to implement.

3.2.1 Scenario B1

In this scenario, $P_{PV} > P_{Load}$ and $SOC > 90\%$, where the PV power is larger than the load power and the battery cannot be charged. In the control loop, as shown in Figure 14, the PV-side DC/DC power reference P_{PV_ref} is set to the load power P_L . This can result in that the PV side cannot implement MPPT function, and the PV-side power P_{PV} is limited to P_L . Therefore, the PV provides the load power P_L , and the power from the battery P_B is zero.

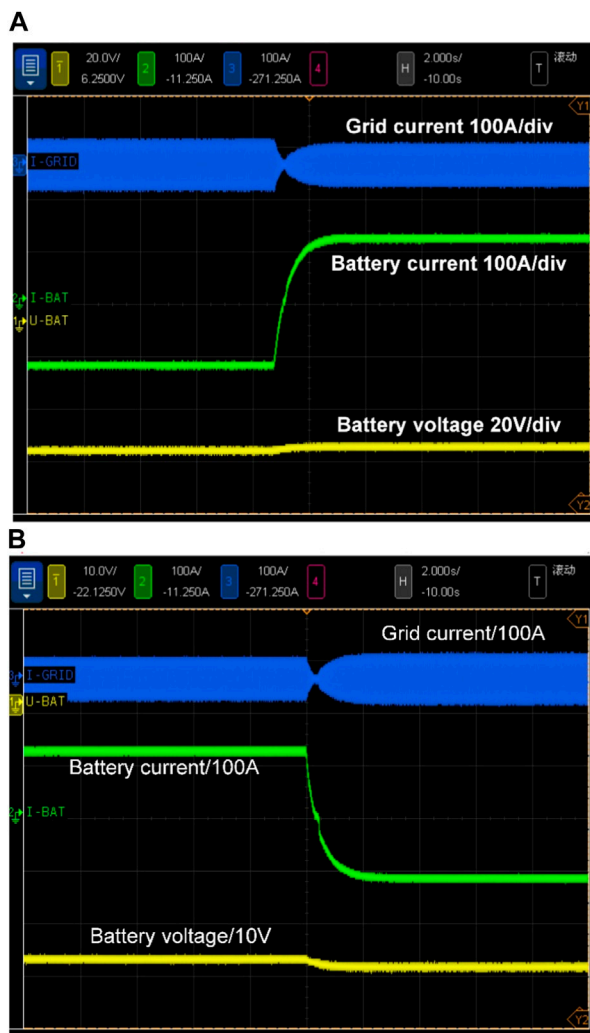


FIGURE 21
Experimental results for the grid-connected mode when there is no PV (A) battery charging to discharging; (B) battery discharging to charging.

3.2.2 Scenario B2

In this scenario, $P_{PV} > P_{Load}$ and $SOC \leq 90\%$, where the PV power is higher than the load power and the battery can be charged. In the control loop, as shown in Figure 15, the PV-side DC/DC performs MPPT function. This can result in that the PV-side DC/DC outputs the maximum power P_{MPPT} . Furthermore, the battery automatically absorbs the extra power generated by PV, which is $P_{PV} - P_L$.

3.2.3 Scenario B3

In this scenario, $P_{PV} \leq P_{Load}$ and $SOC > 10\%$, where the PV power is less than the load power and the battery can be discharged to the load. In the control loop, as shown in Figure 16, the PV-side DC/DC performs MPPT. This can result in that the PV-side DC/DC outputs the maximum power P_{MPPT} . Furthermore, the battery automatically provides the remaining power, which is $P_L - P_{PV}$.

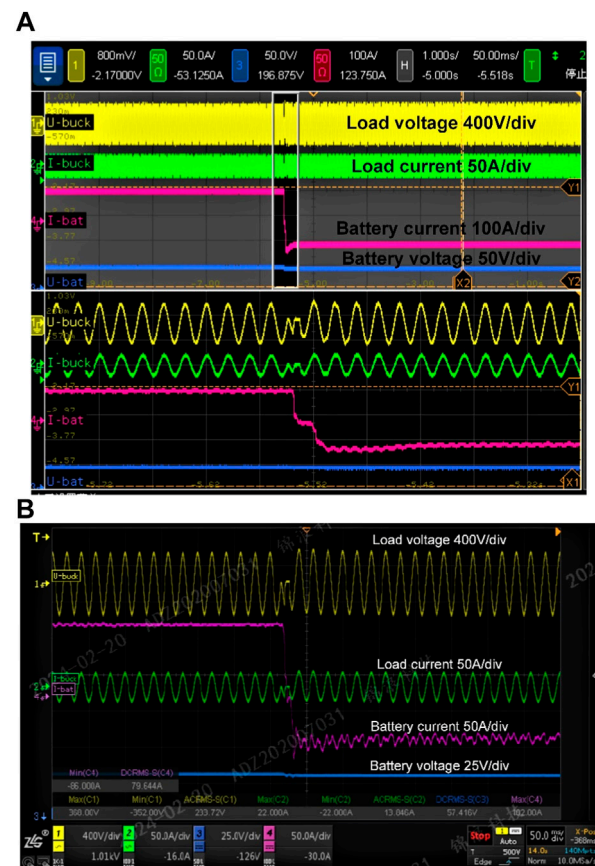


FIGURE 22
Experimental results between grid-connected mode and off-grid mode (A) grid-connected mode to off-grid mode; (B) off-grid mode to grid-connected mode.

3.2.4 Scenario B4

In this scenario, $P_{PV} \leq P_{Load}$, and $SOC \leq 10\%$, where the PV power is less than the load power and the battery cannot be discharged. Therefore, as shown in Figure 17, the system cannot operate.

3.2.5 Scenario B5

In this scenario, $P_{PV} \leq P_{PV_min}$ and $SOC > 10\%$, where the PV cannot generate power and the battery can be discharged to feed the load. The PV-side DC/DC stops. Furthermore, the battery automatically provides the load power P_L as shown in Figure 18.

3.2.6 Scenario B6

In this scenario, $P_{PV} \leq P_{PV_min}$ and $SOC \leq 10\%$, where the PV cannot generate power and the battery cannot be discharged to feed the load. Thus, as shown in Figure 19, the system cannot operate.

The summary of six scenarios for the off-grid mode is shown in Figure 20. From it, a switch S_2 is introduced to set different DC/DC power references. In scenarios B1, the switch S_2 is placed to “1”. In scenarios B2 and B3, the switch S_2 is placed to “2”. In other scenarios, PV-side DC/DC stops. The details are presented in Table 2.

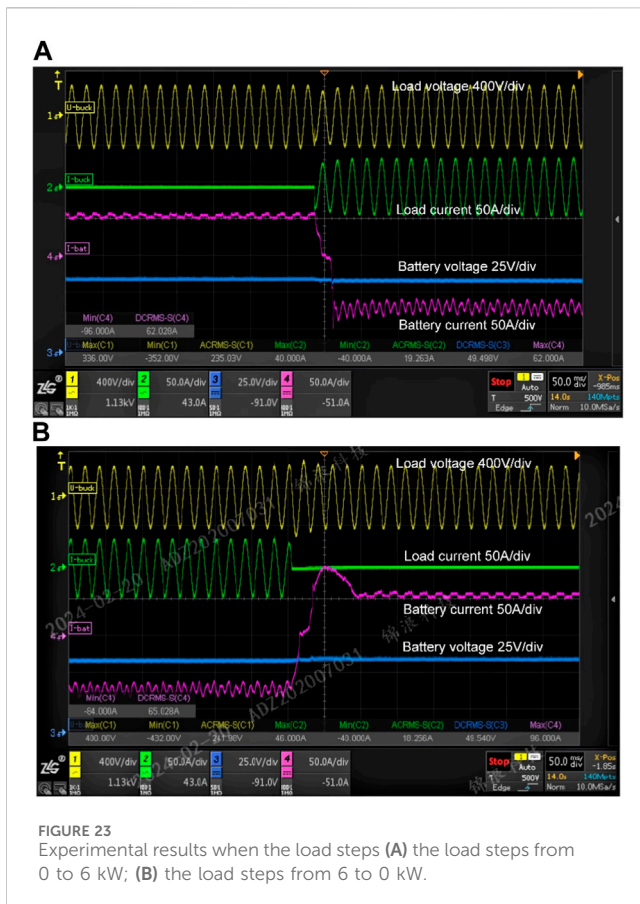


FIGURE 23
Experimental results when the load steps (A) the load steps from 0 to 6 kW; (B) the load steps from 6 to 0 kW.

4 Experimental verifications

A platform is built and Figures 21–23 are the experimental results. Figure 21 depicts the experimental results for the grid-connected mode with no PV and the battery switched between charging and discharging. From Figure 21A, the system operates smoothly when the battery is switched from charging to discharging and the current is adjusted from -120 to $+120$ A. From Figure 21B, the battery is transitioning from charging to discharging, and the entire process is seamless. The battery current is changed from $+120$ to -120 A, and the grid current is adjusted correspondingly. The proposed mode transition method ensures a smooth transition between these two operational modes.

Figure 22 depicts the experimental results of the transition between grid-connected mode and off-grid mode. From Figure 22A, the system operates smoothly and the load voltage remains stable when the system switched from grid-connected to off-grid mode. In grid-connected mode, the inverter functions as a current source. In the off-grid mode, however, the inverter functions as a voltage source. In addition, when the grid is normal, the battery is charging. When the grid is offline, the battery is switched to discharging mode to maintain the load voltage.

From Figure 22B, the system operates smoothly and the load voltage remains stable when the system is switched from off-grid to grid-connected mode. The inverter switches from a voltage source in off-grid mode to a current source in grid-connected mode. In addition, the battery is in discharging mode to maintain the load voltage when the grid is offline. When the grid resumes to normal

operation, the battery is switched to charging mode. The proposed mode transition method ensures a smooth transition between grid-connected mode and off-grid mode.

Figure 23 depicts the experimental results when the load steps. From Figure 23A, the load steps from 0 to 6 kW. When the load power is low, the PV power feeds into the battery and the battery is in the charging mode. When the load power is high, the PV and battery work together to power the load, and the battery is in discharging mode. From Figure 23A, the system operates smoothly when the load steps from 0 to 6 kW.

From Figure 23B, the system operates smoothly and the load voltage remains stable when the system steps from 6 to 0 kW. Furthermore, when the load is 6 kW, the PV and battery work together to supply the voltage. When the load drops to 0 kW, the PV feeds the battery and the battery is in the charging mode. The proposed mode transition method ensures a smooth transition between these two operational modes.

5 Conclusion

Depending on PV power, load power, and battery status, the system operates in various scenarios, and the control loop may need to change. Because the transition and dynamic processes are difficult to control, implementing control loop switching can be difficult. As a result, this paper presents a generalized mode control method that avoids loop switching in a variety of scenarios.

First, the system structure and topology are introduced, with the common DC-bus structure being used. The operating conditions in grid-connected and off-grid modes are then classified into six scenarios. Furthermore, the control architecture, control loop, and reference transition for various scenarios are discussed. In grid-connected mode, the battery-side DC/DC controls the DC-bus voltage, the PV-side DC/DC uses MPPT mode to maximize PV output power, and the grid-side DC/AC controls the output power. The control loops for DC/DC and Bi-DC/DC do not need to change in different operating scenarios; only the DC/AC output power reference P_{ref} does. Furthermore, P_{ref} equals P_{PV} in scenarios A1 ($P_{PV} > P_{Load}$ and $SOC > 90\%$) and A4 ($P_{PV} \leq P_{Load}$ and $SOC \leq 10\%$). P_{ref} equals P_L in scenarios A2 ($P_{PV} > P_{Load}$ and $SOC \leq 90\%$), A3 ($P_{PV} \leq P_{Load}$ and $SOC > 10\%$), and A5 ($P_{PV} \leq P_{PV_{min}}$ and $SOC > 10\%$). P_{ref} equals zero in scenario A6 ($P_{PV} \leq P_{PV_{min}}$ and $SOC \leq 10\%$). In off-grid mode, the battery-side DC/DC regulates the DC-bus voltage, the PV-side DC/DC uses PPT mode to track the PV output power reference, and the grid-side DC/AC regulates the output voltage. During operating mode switching, only the PV-side DC/DC power reference P_{PV} needs to change; the DC/AC and Bi-DC/DC control loops remain unchanged. P_{PV} equals P_L in scenario B1 ($P_{PV} > P_{Load}$ and $SOC > 90\%$). P_{PV} equals P_{MPPT} in scenario B2 ($P_{PV} > P_{Load}$ and $SOC \leq 90\%$) and B3 ($P_{PV} \leq P_{Load}$ and $SOC > 10\%$). PV-side DC/DC stops in scenario B5 ($P_{PV} \leq P_{PV_{min}}$ and $SOC > 10\%$). In B4 ($P_{PV} \leq P_{Load}$ and $SOC \leq 10\%$) and B6 ($P_{PV} \leq P_{PV_{min}}$ and $SOC \leq 10\%$) scenarios, the system stops. Finally, an experimental platform is built, and the proposed methodology is validated by the experimental data. According to the experimental findings, there are no surges during mode transitions with the proposed method and seamless transition among operating modes can be achieved.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

Author contributions

WZ: Writing—original draft, Methodology. YW: Conceptualization, Writing—original draft. PX: Software, Writing—original draft. DL: Supervision, Writing—review and editing. BL: Validation, Writing—original draft.

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Conflict of interest

Authors WZ, YW, PX, and BL were employed by Research Center, Ginlong Technologies Co., Ltd.

The remaining authors declare that the research was conducted in the absence of any commercial or financial relationships that could be construed as a potential conflict of interest.

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Load frequency optimal control of the hydropower-photovoltaic hybrid microgrid system based on the off-policy integral reinforcement learning algorithm

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With the promotion and development of clean energy, it is challenging to ensure the optimization of control performance in frequency control of the hydropower-photovoltaic hybrid microgrid system caused by the output power fluctuation of photovoltaic power generation. In this study, an optimal load frequency controller (LFC) for a hydropower-photovoltaic hybrid microgrid system was designed to improve the dynamic response when the load and photovoltaic output power are perturbed based on the off-policy integral reinforcement learning algorithm. First, a mechanism model of the hydropower-photovoltaic hybrid microgrid system was established. Next, the LFC problem was transformed into a zero-sum game control problem based on the characteristics of the power system. Subsequently, three neural networks were employed to approximate the Nash equilibrium solution of the zero-sum game with historical input and output data when the system dynamics are completely unknown. Finally, simulation experiments were conducted to verify the effectiveness and optimality of the proposed method. The introduction of this method provides a new perspective for frequency control for the hydropower-photovoltaic hybrid microgrid system.

KEYWORDS

hydropower-photovoltaic hybrid microgrid system, load frequency controller, off policy integral reinforce learning algorithm, data-based optimal control, neural networks

1 Introduction

With the development of the national economy and society, the contradiction between increasing energy demand and energy shortages has become increasingly obvious (Gilani et al., 2020; Patnaik et al., 2020; Zhang and Kong, 2022). Traditional thermal power generation causes problems such as the consumption of nonrenewable energy and excessive carbon emissions (Ahmad et al., 2018; Cowie et al., 2020; Olabi and Abdelkareem, 2022).

Abbreviations: LFC, load frequency controller; PV, photovoltaic; PID, proportional-integral-derivative; ADP, adaptive dynamic programming; IRL, integral reinforcement learning.

Hydropower and solar energy have attracted the attention of researchers owing to their renewable and environment-friendly nature (Gielen et al., 2019; Zepter et al., 2019).

However, photovoltaic (PV) power generation is intermittent, leading to unstable output power and microgrid frequency oscillations (Thirunavukkarasu and Sawle, 2021; Chen et al., 2022; Wu and Yang, 2023). To ensure the frequency stability of a microgrid, it is necessary to supplement controllable power sources, such as hydroelectric units or energy storage devices, to fill the power deficit, which can effectively maintain the microgrid frequency stability (Coban et al., 2022). The power quality of PV power systems can be improved by utilizing a control algorithm for controllable power sources, which is applied to obtain an optimal load frequency controller (LFC) system (Papaefthymiou et al., 2010; Ma et al., 2014; Dhundhara and Verma, 2020). Some researchers focus on the suppression of local load fluctuations and their interactions with the distribution system (Khalid et al., 2022). Additionally, the role of ancillary services and the integration of renewable energy should also be addressed upon introduction to minimize fluctuations and cover intermittency (Khalid et al., 2022; Rehman et al., 2024; Osman et al., 2022).

Owing to their simple structure and ease of implementation, proportional-integral-derivative (PID) control methods are widely used in microgrid LFC (Mohamed et al., 2020; Nisha and Jamuna, 2022). Ray et al. (2011) utilized a PI controller to regulate the frequency of a microgrid and achieve the required frequency ratings. Guha et al. (2021) designed a fractional-order PID method to solve the frequency stabilization problem of microgrid systems with uncertain parameters. Huang et al. (2021) used fuzzy reasoning in PID to improve the control performance of a hydraulic turbine regulation system.

Many practical power systems can only be partially modeled, and models of unknown parts are unavailable (Ganguly et al., 2018; Li et al., 2023; Wu and Yang, 2023). Dynamic characteristics of droop-controlled inverters are evaluated by a reduce-order small-signal transfer function model, which is designed on the basis of the Jordan continued-fraction expansion to provide a preprocessing method for real-time power system simulation (Wang et al., 2020). Therefore, owing to the insensitivity to the dynamics of the unmodeled parts of the controlled object, adaptive control methods have been proposed by continuously identifying system parameters to achieve the ideal control effect. Adaptive control methods can be used to resolve problems arising from parameter variations in the LFC of a power system. Zeng et al. (2015) designed a port-controlled Hamiltonian system that decomposed nonlinear control into stabilizing control with a given equilibrium point and proposed L-2 adaptive control for application to a hydroelectric generator unit. Fang et al. (2011) effectively improved the dynamic performance of the hydraulic turbine regulation process using an improved particle swarm optimization algorithm, which was applied to the optimal design of the parameters of a hydraulic turbine regulation system to achieve an optimal positive setting of the parameters. Tran et al. (2021) used a combination of second-order sliding film control and a state estimator for frequency regulation to reduce the number of overtones. Although these methods can achieve better control performance, they have not been widely popularized in practical power systems owing to their complexity and difficulty.

The adaptive dynamic programming (ADP) algorithm is an emerging intelligent control algorithm that solves the problem of dimensional disasters caused by the traditional dynamic programming (DP) method (Werbos, 1992; Vamvoudakis and Lewis, 2010; Lewis et al., 2012; Bellman and Dreyfus, 2015) and is suitable for systems with a high degree of nonlinearity. Shuai et al. (2020) used a hybrid ADP algorithm to achieve optimal operation of gas and electric systems. Xue et al. (2022) used ADP for the real-time scheduling of battery heat storage tank integrated heat and power systems, providing optimal economic operation strategies. The off-policy integral reinforcement learning (IRL) algorithm is proposed based on the theory of the ADP algorithm, which can explore system information with historical input and output data, thereby overcoming the difficulty of traditional ADP relays on neural network weights in the training process to find the continuous excitation function. Chai et al. (2017) used the game theory to solve multi-objective trajectory optimization problems for aerial vehicles. Song et al. (2019) proposed an off-policy IRL algorithm to solve an optimal control problem with partially known system dynamics. Based on the ADP algorithm, this paper proposes an integral reinforcement learning method that requires only the historical input-output data of the system, allowing for optimal solutions even when the system dynamics are completely unknown.

To the best of our knowledge, in the hydropower-photovoltaic hybrid microgrid system, the challenges of considering system disturbances and employing model-free methods for frequency control are quite evident. Traditional frequency control methods typically rely on a mathematical model of the system and assume that disturbances are known or predictable. However, in real microgrid systems, disturbances such as load variations and fluctuations in renewable energy output are often unpredictable, and obtaining an accurate model of the system can be difficult or complex. The existing reinforcement learning methods for frequency control in the hydropower-photovoltaic hybrid microgrid systems have not simultaneously addressed disturbances in the system and utilized the model-free approaches, which motivates our study. The focus of this paper is on how to abstract a hybrid power generation system with disturbances as a zero-sum game problem and solve it using the proposed model-free method. This approach provides a theoretical foundation and basis for the grid integration of a series of photovoltaic combined power generation systems. The main contributions of this article are as follows:

1. A hydropower-photovoltaic hybrid microgrid system model was constructed on the basis of the mechanistic modeling of the hydraulic turbine and photovoltaic power generation, meanwhile treating the photovoltaic power generation perturbed as the disturbance term.
2. Based on the power generation characteristics, the secondary frequency modulation control signal was used as the control vector, and the input system load frequency and solar energy power were used as the perturbation vectors of the hydropower-photovoltaic microgrid power system, which transforms the LFC problem into a zero-sum optimal control problem. By solving the Nash equilibrium of the zero-sum game, the optimal control rate and the maximum disturbance that the system can withstand can be obtained,

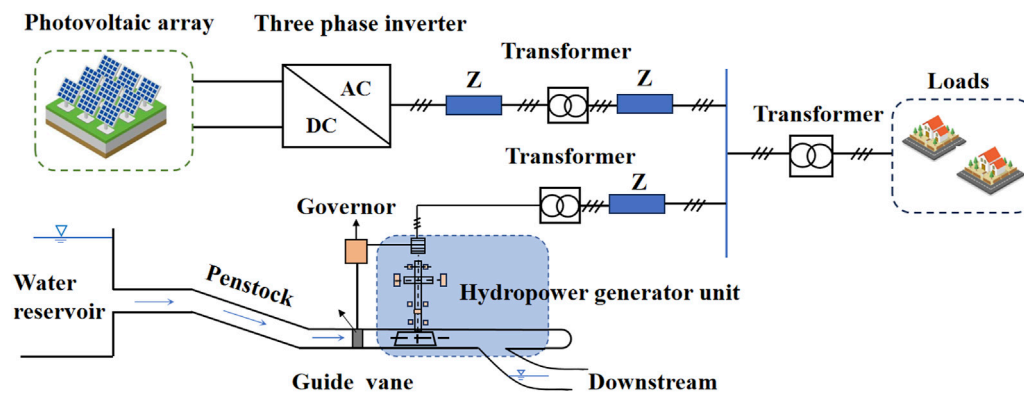


FIGURE 1
Main structure of the independent microgrid.

thereby controlling the load frequency of the hydropower-photovoltaic hybrid microgrid system.

3. An off-policy IRL algorithm was adopted to resolve the zero-sum optimal control problem in which three networks were employed to approximate the Nash equilibrium point of the zero-sum game to obtain the optimal LFC of the hybrid system. The proposed method overcomes the limitation of existing solution methods that require precise system model information.

2 Problem statement

The hydropower-photovoltaic microgrid power system effectively exploits the inherent frequency regulation advantages of hydropower units while integrating solar energy generation resources within the same regional grid. This hybrid system aims to enhance the overall frequency quality of the microgrid by balancing both renewable energy inputs and electrical load demand. However, such integration significantly increases the operational requirements for the Load Frequency Control (LFC) controller. In this study, Figure 1 outlines the core structure of the system: a power busbar connects hydropower units (HP), photovoltaic generation units (PV), and electrical loads. Specifically, the PV units are connected to the alternating current (AC) microgrid through direct current (DC) to alternating current (AC) conversion using DC/AC inverters.

The frequency stability of this isolated microgrid relies heavily on maintaining an active power balance within the network. Variations in electrical load and the intermittent, fluctuating output from photovoltaic sources can disturb this balance, leading to changes in system frequency. A central feature of the hydropower-photovoltaic microgrid system is the hydro-turbine generator, which is responsible for providing rotational reserves that help regulate frequency by adjusting the mechanical input to the turbine. This response compensates for any mismatch between generation and demand, ensuring system stability.

The hydropower units play a critical role in Load Frequency Control (LFC) tasks. The primary function of the LFC system is to regulate water flow into the turbines of the hydropower generators. It achieves this by dynamically adjusting the active power output of

the hydropower units in real-time, depending on the load and intermittent power output from the solar resources. This real-time control is vital for compensating fluctuations in both solar power production and load changes, stabilizing the generator speed, and ultimately controlling the frequency of the microgrid.

To improve the effectiveness of the system and minimize control costs, an advanced optimal load frequency controller was designed, utilizing an off-policy Inverse Reinforcement Learning (IRL) algorithm. This controller ensures the stability of the grid-connected voltage in the hydropower-photovoltaic microgrid by optimizing the dynamic allocation of power resources. In essence, it manages the trade-offs between ensuring grid frequency stability and maintaining operational cost efficiency, leading to a robust, reliable, and sustainable microgrid power system.

3 Materials and methods

3.1 Establishment of the hydro turbine group model

The hydro turbine group consisted of hydro turbines, governors, and generators. A turbine group model was established for each part.

The equations of moment and flow of the hydro turbine are expressed as Equation 1:

$$\begin{cases} \Delta m_t(t) = n_x \Delta x(t) + n_y \Delta y(t) + n_h \Delta h(t) \\ \Delta q(t) = n_{qx} \Delta x(t) + n_{qy} \Delta y(t) + n_{qh} \Delta h(t), \end{cases} \quad (1)$$

where $\Delta m_t(t)$ and $\Delta q(t)$ represent the increments in torque and flow, respectively; $\Delta y(t)$, $\Delta h(t)$, and $\Delta x(t)$ indicate the relative deviation variables of the guide vane opening, water head, and hydro turbine speed, respectively; n_x , n_y , and n_h represent the transfer coefficients of the hydro turbine torque to the speed, guide vane opening and water head, respectively; n_{qx} , n_{qy} and n_{qh} are transfer coefficients of the hydro turbine flow to the speed, guide vane opening and the water head, respectively. Under stable conditions, each transfer coefficient is regarded as a constant.

When $n = \frac{n_{xy}n_{yh}}{n_y} - n_{qh}$, the transfer function from the guide vane increment $\Delta y(t)$ to the torque increment $\Delta m_t(t)$ of the hydro turbine is as follows:

$$G_{ym}(s) = \frac{\Delta m(s)}{\Delta y(s)} = n_y \frac{1 - nT_w s}{1 + n_{qh}T_w s}, \quad (2)$$

where T_w indicates the inertia time constant of the water flow.

The hydro turbine governor can be simplified as a first-order inertial link by ignoring the nonlinear factors, as Equation 3:

$$G_y(s) = \frac{\Delta y(s)}{u(s)} = \frac{1}{T_y s + 1}, \quad (3)$$

where T_y represents the response time constant of the hydro-turbine governor. The corresponding differential equation is as follows:

$$\frac{d\Delta y(t)}{dt} = -\frac{1}{T_y} \Delta y(t) + \frac{1}{T_y} u(t) \quad (4)$$

Equation 2 is substituted into Equation 4 after the Laplace transformation. The differential equation is obtained by the Laplace transformation of Equation 2, and Equation 4 is substituted into the differential equation to obtain the hydro turbine differential equation as follows:

$$\frac{d\Delta m_t(t)}{dt} = \frac{1}{n_{qh}T_w} \left[-\Delta m_t(t) + \left(n_y + n_y n \frac{T_w}{T_y} \right) \Delta y(t) - n_y n \frac{T_w}{T_y} u(t) \right] \quad (5)$$

The second-order model of the generator includes the rotor rotation motion equation and the equation that characterizes the relationship between the power angle and speed, as follows:

$$\begin{cases} \frac{d\delta(t)}{dt} = \omega_0 \Delta x(t) \\ \frac{dx(t)}{dt} = \frac{1}{T_J} [\Delta m_t(t) - \Delta P_e(t) - D\Delta x(t)], \end{cases} \quad (6)$$

where δ indicates the generator power angle, which remains constant under stable operating conditions, and ΔP_e indicates the electromagnetic power increment, which is equivalent to the relative value of the load increment ΔP_L . $\omega_0 = 100\pi$ indicates the synchronous electrical angular velocity, D is the damping coefficient, and T_J indicates the inertia time constant of the

generator. ΔP_S is the sum of the incremental relative values of power supplied by all sources within the microgrid. From the Equation 6, we get $\Delta P_S - \Delta P_L = T_M \frac{d\Delta f(t)}{dt} + D_M \Delta f(t)$. T_M represents the equivalent inertia time coefficient of the system, which is equal to the weighted sum of the inertia coefficients of all the generators in the system. D_M represents the equivalent load-damping factor of the system.

By combining Equations 4–6, the following mathematical model of the hydro-turbine group can be obtained:

$$\begin{cases} \frac{d\delta(t)}{dt} = \omega_0 \Delta x(t) \\ \frac{d\Delta x(t)}{dt} = \frac{1}{T_J} [\Delta m_t(t) - \Delta P_e(t) - D\Delta x(t)] \\ \frac{d\Delta m_t(t)}{dt} = \frac{1}{n_{qh}T_w} \left[-\Delta m_t(t) + \left(n_y + n_y n \frac{T_w}{T_y} \right) \Delta y(t) - n_y n \frac{T_w}{T_y} u(t) \right] \\ \frac{d\Delta y(t)}{dt} = -\frac{1}{T_y} \Delta y(t) + \frac{1}{T_y} u(t) \end{cases} \quad (7)$$

3.2 Establishment of the photovoltaic model

PV panels convert solar energy into electrical energy based on PV effects. The main body of the frequency control in this study was the hydropower unit. Therefore, in this subsection, a first-order model with time constant T_{solar} was used to express the frequency characteristics of the PV power as follows:

$$\frac{d\Delta P_{PV}}{dt} = -\frac{1}{T_{solar}} \Delta P_{PV} + \frac{1}{T_{solar}} \Delta P_{solar}, \quad (8)$$

where T_{solar} represents the time constant of the PV power system, ΔP_{PV} represents the output power of the PV power system, and ΔP_{solar} represents the solar power. Solar power has obvious volatility, as PV panels are easily affected by light intensity and ambient temperature. Therefore, solar power is regarded as a disturbance term. A block diagram of the model is shown in the Figure 2.

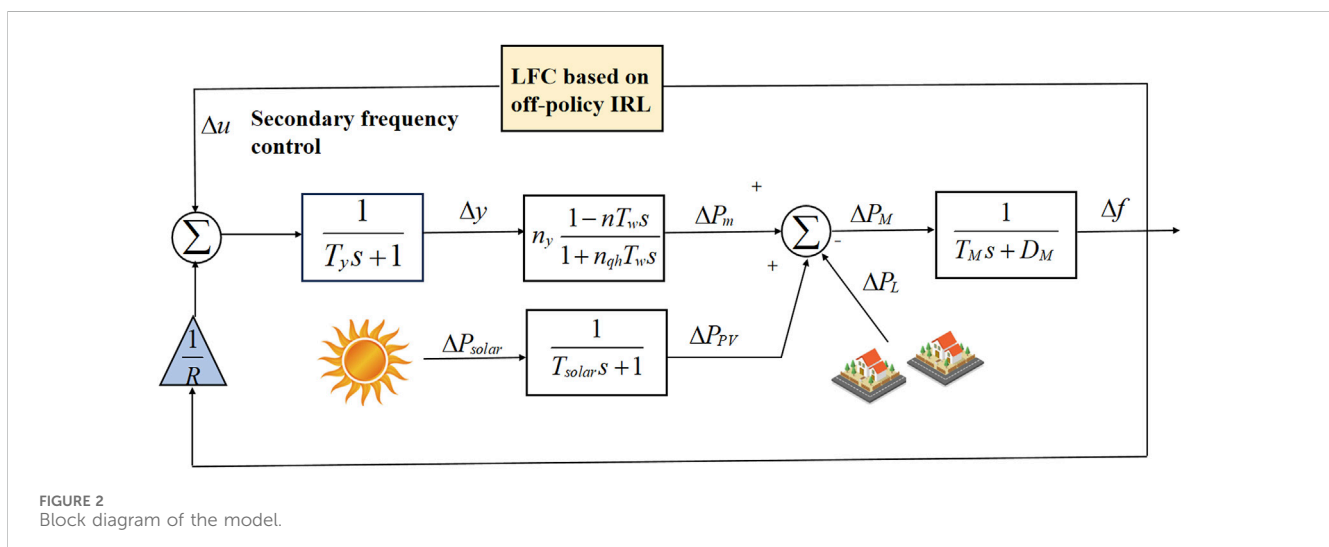


FIGURE 2
Block diagram of the model.

3.3 Establishment of the hydropower-photovoltaic microgrid power system model

The transient changes in the voltage and power angle of the system can be ignored in the frequency control analysis. Therefore, in the analysis process of LFC, $\Delta m_t(t) = \Delta P_m(t)$ is the incremental relative value of the output power of the hydro turbine group, the relative value of the electromagnetic power increment $\Delta P_e(t)$ is replaced by the relative value of the load increment $\Delta P_L(t)$, and the relative value of the speed deviation $\Delta x(t)$ is equal to the relative value of the frequency deviation $\Delta f(t)$.

By combining Equations 7, 8, the hydro-photovoltaic microgrid power system can be derived as Equation 9:

$$\begin{cases} \frac{d\delta(t)}{dt} = w_0 \Delta f(t) \\ \frac{d\Delta f(t)}{dt} = \frac{1}{T_f} [\Delta P_m(t) - \Delta P_L(t) - D \Delta f(t)] \\ \frac{d\Delta P_m(t)}{dt} = \frac{1}{n_{gh} T_w} \left[-\Delta P_m(t) + \left(n_y + n_y n \frac{T_w}{T_y} \right) \Delta y(t) - n_y n \frac{T_w}{T_y} u(t) \right] \\ \frac{d\Delta y(t)}{dt} = -\frac{1}{T_y} \Delta y(t) + \frac{1}{T_y} u(t) \\ \frac{d\Delta P_{PV}}{dt} = -\frac{1}{T_{solar}} \Delta P_{PV} + \frac{1}{T_{solar}} \Delta P_{solar} \end{cases} \quad (9)$$

Here, $x(t) = [x_1(t) \ x_2(t) \ x_3(t) \ x_4(t) \ x_5(t)]^T = [\delta(t) \ f(t) \ \Delta P_m(t) \ \Delta y(t) \ \Delta P_{PV}(t)]^T$ are state variable. The load frequency $\Delta P_L(t)$ and solar energy ΔP_{solar} of the input system are considered as the elements in the disturbance vector, and the PV output power fluctuation and load power change cause the power supply and demand of microgrid to lose balance. The system disturbance vector can be obtained as follows: $w(t) = [\Delta P_L(t) \ \Delta P_{solar}(t)]^T$; $u = u(t)$ is the control vector signal. Therefore, the load frequency model of the hydropower-photovoltaic system can be obtained as follows:

$$\dot{x} = Ax + Bu + Fw, \quad (10)$$

where the system state variable $x \in R^{5 \times 1}$ is the system control variable and $u \in R^{1 \times 1}$ is the disturbance variable $w \in R^{2 \times 1}$. The system matrix is expressed as follows:

$$A = \begin{bmatrix} 0 & w_0 & 0 & 0 & 0 \\ 0 & -\frac{D}{T_f} & \frac{1}{T_f} & 0 & 0 \\ 0 & 0 & -\frac{1}{n_{gh} T_w} & \frac{n_y}{n_{gh} T_w} \left(1 + n \frac{T_w}{T_y} \right) & 0 \\ 0 & 0 & 0 & -\frac{1}{T_y} & 0 \\ 0 & 0 & 0 & 0 & -\frac{1}{T_{solar}} \end{bmatrix},$$

$$B = \begin{bmatrix} 0 \\ 0 \\ \frac{n_y n}{n_{gh} T_y} \\ \frac{1}{T_y} \\ 0 \end{bmatrix}, F = \begin{bmatrix} 0 & 0 \\ -\frac{1}{T_f} & 0 \\ 0 & 0 \\ 0 & 0 \\ 0 & \frac{1}{T_{solar}} \end{bmatrix}$$

Thus far, the load frequency control problem of the hydropower-photovoltaic microgrid power system has been transformed into a zero-sum game optimal control problem, wherein the input of the governor was taken as the control variable and the load frequency and solar power were taken as the disturbance variables.

4 Results of the optimal controller based on off-policy IRL algorithm

The hydropower-photovoltaic microgrid power system model was established using Equation 10, where x, u and w are the state, control input, and disturbance input of the system, respectively. $x = 0$ is the equilibrium point of the hydropower-photovoltaic microgrid power system. The infinite-horizon performance index function can be designed as follows:

$$J(x(0), u, w) = \int_0^\infty \{r(x(t), u, w)\}, \quad (11)$$

where in the utility function can be described as Equation 12:

$$r(x(t), u, w) = x^T R x + u^T S u - w^T T w, \quad (12)$$

where the coefficient matrices R, S and T are real symmetric positive definite matrices. Please note that this performance indicator has clear economic significance, in which $x^T R x$ is the penalty cost when the quality of power supply at the node i deviates from the system's stable value and $u^T S u - w^T T w$ is the control costs and disturbance costs incurred by node i to reduce penalty costs.

The purpose of the zero-sum game is to solve for an optimal control that satisfies Equation 13.

$$V^*(x(0)) = \inf_u \sup_w (x(0), u, w) \quad (13)$$

The zero-sum game selects to minimize the player set u and maximize the player set w , the saddle point u^* and w^* must satisfy the following inequality Equation 14:

$$J(x, u^*, w) \leq J(x, u^*, w^*) \leq J(x, u, w^*) \quad (14)$$

When there is a unique set of solutions that satisfy the following Nash equilibrium condition Equation 15:

$$V^*(x) = \inf_u \sup_w J(x, u, w) = \sup_w \inf_u J(x(0), u, w), \quad (15)$$

the cost function of every player can be written as Equation 16:

$$V(x(t)) = \int_t^\infty \{x^T R x + u^T S u - w^T T w\} dt \quad (16)$$

Using the Leibniz formula and differentiating Equation 6, the Bellman equation of the zero-sum game can be obtained as follows:

$$H(x, \nabla V, u, w) = x^T R x + u^T S u - w^T T w + \nabla V^T (Ax + Bu - Fw), \quad (17)$$

where $\nabla V = \frac{\partial V}{\partial x}$. The control and disturbance inputs can be obtained as Equations 18, 19:

$$\frac{\partial H}{\partial u} = 2uS + \nabla V^T B = 0 \quad (18)$$

$$\frac{\partial H}{\partial w} = -2wT + \nabla V^T F = 0 \quad (19)$$

The optimal control policy u^* and the optimal disturbance w^* can be derived as follows:

$$u^* = -\frac{1}{2}S^{-1}B^T \nabla V \quad (20)$$

$$w^* = \frac{1}{2}T^{-1}F^T \nabla V \quad (21)$$

The Hamilton-Jacobi-Bellman equation can be obtained by substituting Equations 20, 21 into Equation 17 as follows:

$$0 = x^T R x + \nabla V^T A x - \frac{1}{4} \nabla V^T B S^{-1} B^T \nabla V + \frac{1}{4} \nabla V^T F T^{-1} F^T \nabla V \quad (22)$$

The following equations were used to update the control and disturbance policies as Equations 23, 24:

$$u^{[k+1]} = -\frac{1}{2}S^{-1}B^T \nabla V^{[k]} \quad (23)$$

$$w^{[k+1]} = \frac{1}{2}T^{-1}F^T \nabla V^{[k]}, \quad (24)$$

where the superscript $[k]$ represents the number of steps in the iteration process.

The Equation 11 can be transformed as Equation 25:

$$\dot{x} = Ax + Bu^{[k]} + Fw^{[k]} + B(u - u^{[k]}) + F(w - w^{[k]}) \quad (25)$$

The Equation 22 can be rewritten as follows:

$$\begin{aligned} & V^{[k]}(x(t + \Delta t)) - V^{[k]}(x(t)) \\ &= \int_t^{t+\Delta t} \Delta V^{[k]T} \dot{x} d\tau \\ &= \int_t^{t+\Delta t} \Delta V^{[k]T} (Ax + bu + Fw) d\tau \\ &\quad + \int_t^{t+\Delta t} \Delta V^{[k]T} (B(u - u^{[k]}) + F(w - w^{[k]})) d\tau \\ &= - \int_t^{t+\Delta t} (x^T R x + u^{[k]T} S u^{[k]} - w^{[k]T} T w) d\tau \\ &\quad + \int_t^{t+\Delta t} \Delta V^{[k]T} (B(u - u^{[k]}) + F(w - w^{[k]})) d\tau \end{aligned} \quad (26)$$

By deriving Equation 23, we get $\nabla V^{[k]T} B = -2u^{[k+1]}S$ and $\nabla V^{[k]T} F = 2w^{[k+1]}T$; upon substituting these parameters into Equation 26, the following equation was obtained:

$$\begin{aligned} & V^{[k]}(x(t + \Delta t)) - V^{[k]}(x(t)) \\ &= - \int_t^{t+\Delta t} (x^T R x + u^{[k]T} S u^{[k]} - w^{[k]T} T w) d\tau \\ &\quad + \int_t^{t+\Delta t} 2(u^{[k+1]T} S(u - u^{[k]}) - w^{[k+1]T} T(w - w^{[k]})) d\tau \end{aligned} \quad (27)$$

From Equations 26, 27, the system dynamic matrices A , B , and F are replaced. Equation 27 overcomes the difficulty of obtaining the dynamic information of the system in practical applications. $(V^{[k]}, u^{[k+1]}, w^{[k+1]})$ is a unique solution for the off-policy IRL algorithm. Three neural networks were employed to solve the solution $(V^{[k]}, u^{[k+1]}, w^{[k+1]})$; the expressions are as follows:

$$\hat{V}^{[k]} = P_V^T(x) \hat{\theta}_V^{[k]} \quad (28)$$

$$u^{[k]}(x) = (P_u^a(x))^T \hat{\theta}_{u_a}^{[k]} \quad (29)$$

$$w^{[k]}(x) = (P_w^b(x))^T \hat{\theta}_{w_b}^{[k]}, \quad (30)$$

$P_V^T(x)$, $P_u^T(x)$, and $P_w^T(x)$ satisfy $P_V^T(0) = 0$, $P_u^T(0) = 0$ and $P_w^T(0) = 0$, respectively, and are linearly independent. $\hat{V}^{[k]}$ is approached by a critical neural network (CNN), and $u^{[k]}(x)$ and $w^{[k]}(x)$ are approached by an action neural network (ANN) and a disturbance neural network (DNN), respectively. Here, $\hat{\theta}_V^{[k]}$, $\hat{\theta}_{u_a}^{[k]}$, and $\hat{\theta}_{w_b}^{[k]}$ indicate the weights of CNN, ANN, and DNN, respectively.

According to Equation 28, the residual can be written as follows:

$$\begin{aligned} \epsilon^{[k]}(x, u, w) &= \hat{V}^{[k]}(x(t)) - \hat{V}^{[k]}(x(t + \Delta t)) \\ &\quad - \int_t^{t+\Delta t} (x^T R x + u^{[k]T} S u^{[k]} - w^{[k]T} T w) d\tau \\ &\quad + \int_t^{t+\Delta t} 2(u^{[k+1]T} S(u - u^{[k]}) - w^{[k+1]T} T(w - w^{[k]})) d\tau \end{aligned} \quad (31)$$

Substituting Equations 28–30 into Equation 31 yields

$$\begin{aligned} g^{[k]}(x, u, w) &= [P_V(x(t)) - P_V(x(t + \Delta t))]^T \hat{\theta}_V^{[k+1]} - \int_t^{t+\Delta t} x^T R x d\tau \\ &\quad - \sum_{a=1}^n \int_t^{t+\Delta t} \hat{\theta}_{u_a}^{[k]T} P_u^a(x(t)) S (P_u^a(x(t)))^T \hat{\theta}_{u_a}^{[k]} d\tau \\ &\quad + \sum_{b=1}^m \int_t^{t+\Delta t} \hat{\theta}_{w_b}^{[k]T} P_w^b(x(t)) T (P_w^b(x(t)))^T \hat{\theta}_{w_b}^{[k]} d\tau \\ &\quad + 2 \sum_{a=1}^n \int_t^{t+\Delta t} \hat{\theta}_{u_a}^{[k]T} P_u^a(x(t)) S (P_u^a(x(t)))^T \hat{\theta}_{u_a}^{[k+1]} d\tau \\ &\quad - 2 \sum_{a=1}^n \int_t^{t+\Delta t} u^T S (P_u^a(x(t)))^T \hat{\theta}_{u_a}^{[k+1]} d\tau \\ &\quad - 2 \sum_{b=1}^m \int_t^{t+\Delta t} \hat{\theta}_{w_b}^{[k]T} P_w^b(x(t)) T (P_w^b(x(t)))^T \hat{\theta}_{w_b}^{[k+1]} d\tau \\ &\quad + 2 \sum_{b=1}^m \int_t^{t+\Delta t} w^T T (P_w^b(x(t)))^T \hat{\theta}_{w_b}^{[k+1]} d\tau. \end{aligned} \quad (32)$$

In order to simplify Equation 32, the following parameters are defined as Equations 33–39:

$$Q_{FA}(x(t)) = P_V^T(x(t)) - P_V^T(x(t + \Delta t)) \quad (33)$$

$$Q_{FB}(x(t)) = 2 \sum_{a=1}^n \int_t^{t+\Delta t} \hat{\theta}_{u_a}^{[k]T} P_u^a(x(t)) S (P_u^a(x(t)))^T d\tau \quad (34)$$

$$Q_{FC}(x(t), u) = 2 \sum_{a=1}^n \int_t^{t+\Delta t} u^T S (P_u^a(x(t)))^T d\tau \quad (35)$$

$$Q_{FD}(x(t)) = -2 \sum_{b=1}^m \int_t^{t+\Delta t} \hat{\theta}_{w_b}^{[k]T} P_w^b(x(t)) T (P_w^b(x(t)))^T d\tau \quad (36)$$

$$Q_{FE}(x(t), w) = -2 \sum_{b=1}^m \int_t^{t+\Delta t} w^T T (P_w^b(x(t)))^T d\tau \quad (37)$$

$$Q_M(x(t)) = \int_t^{t+\Delta t} x^T R x d\tau \quad (38)$$

$$\begin{aligned} Q_N(x(t)) &= \sum_{a=1}^n \int_t^{t+\Delta t} \hat{\theta}_{u_a}^{[k]T} P_u^a(x(t)) S (P_u^a(x(t)))^T \hat{\theta}_{u_a}^{[k]} d\tau \\ &\quad - \sum_{b=1}^m \int_t^{t+\Delta t} \hat{\theta}_{w_b}^{[k]T} P_w^b(x(t)) T (P_w^b(x(t)))^T \hat{\theta}_{w_b}^{[k]} d\tau \end{aligned} \quad (39)$$

Then, Equation 32 can be written as Equation 40:

$$\begin{aligned} g^{[k]}(x, u, w) &= Q_{FA}(x(t)) \hat{\theta}_V^{[k+1]} + Q_{FB}(x(t)) \hat{\theta}_{u_a}^{[k+1]} \\ &\quad - Q_{FC}(x(t), u) \hat{\theta}_{u_a}^{[k+1]} + Q_{FD}(x(t)) \hat{\theta}_{w_b}^{[k+1]} \\ &\quad - Q_{FE}(x(t), w) \hat{\theta}_{w_b}^{[k+1]} - Q_M(x(t)) - Q_N(x(t)) \end{aligned} \quad (40)$$

The Equations 41–43 are then generated to obtain the optimal solutions:

$$\hat{W}^{[k]} = \left[\hat{\theta}_V^{[k+1]}, \hat{\theta}_{u1}^{[k+1]} \dots \hat{\theta}_{ua}^{[k+1]}, \hat{\theta}_{w1}^{[k+1]} \dots \hat{\theta}_{wb}^{[k+1]} \right] \quad (41)$$

$$X_A^{[k]}(x, u, w) = \left[Q_{FA}(x(t)), Q_{FB}^1(x(t)) - Q_{FC}^1(x(t), u) \dots, Q_{FB}^n(x(t)) - Q_{FC}^n(x(t)), Q_{FD}^1(x(t)) - Q_{FE}^1(x(t), w), \dots, Q_{FD}^n(x(t)) - Q_{FE}^n(x(t), w) \right] \quad (42)$$

$$X_B^{[k]} = Q_M(x(t)) + Q_N(x(t)) \quad (43)$$

Finally, Equation 32 can be written as follows:

$$\vartheta^{[k]}(x, u, w) = X_A^{[k]}(x, u, w) \hat{W}^{[k]} - X_B^{[k]} \quad (44)$$

In order to solve weight $\hat{W}^{[k]}$, the residual error $\vartheta^{[k]}(x, u, w)$ is approximated to zero. The inner product is applied to solve $\hat{W}^{[k]}$ using Equation 45 as follows:

$$\langle d\vartheta^{[k]}(x, u, w) / d\hat{W}^{[k]}, d\vartheta^{[k]}(x, u, w) \rangle_D = 0, \quad (45)$$

upon substituting Equation 45 is substituted in Equations 44, 46 is obtained as follows:

$$\langle X_A^{[k]}, X_A^{[k]} \rangle_D \hat{W}^{[k]} - \langle X_A^{[k]}, X_B^{[k]} \rangle_D = 0 \quad (46)$$

$\hat{W}$ can be calculated as follows:

$$\hat{W}^{[k]} = \langle X_A^{[k]}, X_A^{[k]} \rangle_D^{-1} \langle X_A^{[k]}, X_B^{[k]} \rangle_D \quad (47)$$

Various numerical integrals in domain D were acquired to calculate $\langle X_A^{[k]}, X_A^{[k]} \rangle_D$ and $\langle X_A^{[k]}, X_B^{[k]} \rangle_D$. The Monte Carlo integration was used to resolve this calculation. When $\mu_D \triangleq \int_D d(x, u, w)$, and $Q_M \triangleq [(x_i, u_i, w_i) | (x_i, u_i, w_i) \in D, i = 1, 2, \dots, M]$, all of them are the sets sampled on domain D . M indicates the dimensions of the Q_M , which should be as wide as possible to ensure that the sample set provides full coverage of D . Therefore, $\langle X_A^{[k]}, X_A^{[k]} \rangle_D$ can be obtained as follows:

$$\begin{aligned} \langle X_A^{[k]}, X_A^{[k]} \rangle_D &= \int_D (X_A^{[k]}(x_i, u_i, w_i))^T (X_A^{[k]}(x_i, u_i, w_i)) d(x_i, u_i, w_i) \\ &= \frac{\mu_D}{M} \sum_{i=1}^M ((X_A^{[k]}(x_i, u_i, w_i))^T (X_A^{[k]}(x_i, u_i, w_i))) \\ &= \frac{\mu_D}{M} (y^{[k]})^T y^{[k]} \end{aligned} \quad (48)$$

in which, $y^{[k]} = [(X_A^{[k]T}(x_1, u_1, w_1), X_A^{[k]T}(x_2, u_2, w_2), \dots, X_A^{[k]T}(x_M, u_M, w_M))]$.

$$\begin{aligned} \langle X_A^{[k]}, X_B^{[k]} \rangle_D &= \frac{\mu_D}{M} \sum_{i=1}^M ((X_A^{[k]}(x_i, u_i, w_i))^T (X_B^{[k]}(x_i))) \\ &= \frac{\mu_D}{M} (y^{[k]})^T \beta^{[k]} \end{aligned} \quad (49)$$

where $\beta^{[k]} = [X_B^{[k]}(x_1), X_B^{[k]}(x_2), \dots, X_B^{[k]}(x_M)]^T$.

Upon substituting Equations 48, 49 in Equation 47, the following equation is obtained:

$$\hat{W}^{[k]} = \left[(y^{[k]})^T y^{[k]} \right]^{-1} (y^{[k]})^T \beta^{[k]} \quad (50)$$

The zero-sum problem can be solved using Algorithm 1 as follows:

Step 1: Start with the signals u and w as well as collecting the hydropower-photovoltaic cogeneration system data (x_p, u_p, w_p) to build the set Q_M ; then, calculate the $Q_{FA}(x), Q_{FB}(x), Q_{FC}(x, u), Q_{FD}(x), Q_{FE}(x, w), Q_M(x)$ and $Q_N(x)$.

Step 2: The values of cost function, control, and disturbance are set initial admissible weight vectors as $\hat{\theta}_V^{[0]}, \hat{\theta}_{u1}^{[0]} \dots \hat{\theta}_{ua}^{[0]}$, and $\hat{\theta}_{w1}^{[0]} \dots \hat{\theta}_{wb}^{[0]}$, separately.

Step 3: Calculate the $y^{[k]}$, and $\beta^{[k]}$ to renew the $\hat{W}^{[k]}$ by the Equation 50.

Step 4: Let $k = k + 1$, return to step 3, and go on.

Step 5: Until $\|\hat{W}^{[k+1]} - \hat{W}^{[k]}\| \leq \alpha$, where in α is a small positive constant. Then, the iteration is stopped, and $\hat{W}^{[k]}$ is used to acquire the control policy using Equation 41.

Algorithm 1 Off-policy IRL method to solve the optimal control problem.

It is worth mentioning that the input and output data of the hydropower-photovoltaic microgrid power system are necessary to solve the zero-sum problem when the system dynamics are completely unknown.

5 Discussion

The hydropower-photovoltaic microgrid power system model was established, the proposed Algorithm 1 was utilized to solve the LFC control, and the simulation was realized in the MATLAB platform. Simulation results verified that the microgrid can maintain frequency stability despite local load and PV disturbances. The control and disturbance curves eventually approach to 0, as shown in Figure 3. The Figure 3 illustrates the behavior of two variables over a period of 10 time steps, designated on the x -axis. The y -axis represents the Control Value ranging from -0.5 to 1.0 . The graph features two sets of trajectories for the control u and ω , each represented by both initial estimated values and adjusted values obtained using the Algorithm 1. The dashed and solid lines indicate the approximation curves under initial admissible control and Algorithm 1, respectively. It can be seen that the convergence speed of Algorithm 1 is better than that of the initial admissible control method. The frequency finally stabilized. For variable ω , it starts from a lower value and similarly converge towards zero. Overall, the obtained trajectories using Algorithm 1 exhibit a more rapid convergence towards 0 for both u and ω compared to their respective initial trajectories. Demonstrating the enhanced performance of Algorithm 1 over the initial admissible control method.

The weight convergence curves of the three networks are shown in Figures 4–6. These three figures illustrates the convergence of weights for every seven different networks θ_V, θ_u , and θ_ω over 15 iterations, represented on the x -axis. The y -axis denotes the Weight of Value Function, with values ranging from -10 to 15 . At the beginning of the iterations, the weights of these networks start from various initial values. Some weights exhibit significant fluctuations, particularly up to around the 5th iteration, where noticeable oscillations are apparent. As iterations proceed, the weights for all networks gradually stabilize and converge to constant values between 0 and 1, indicating that the values no

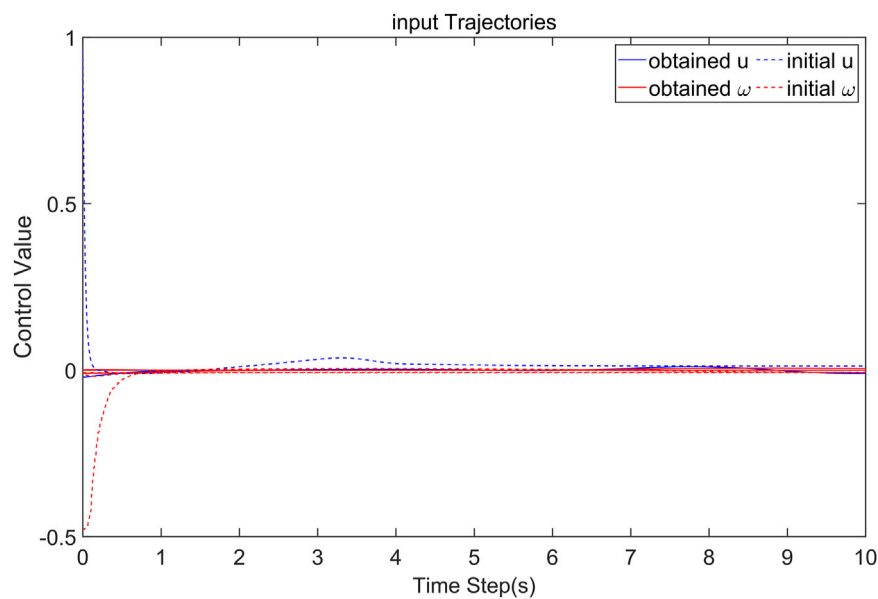


FIGURE 3
State traces of the system.

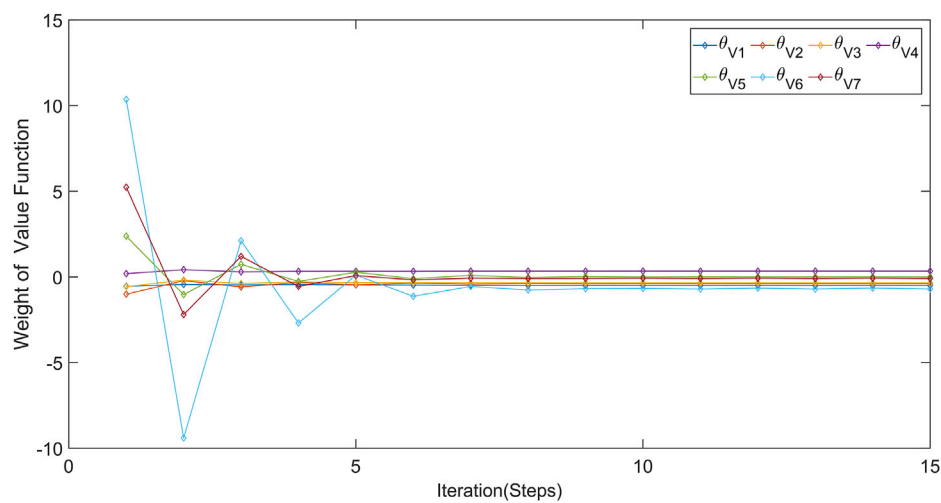


FIGURE 4
Convergence curves of the θ_V .

longer change significantly with additional iterations. By the 15th step, all weights have reached their steady-state values. The convergence of these weights within 15 iterations suggests that the optimization process is effective, allowing the system to arrive at an optimal control that approximates the Nash equilibrium point for the zero-sum game problem.

Compared to traditional Dynamic Programming methods, the proposed method effectively overcomes the “curse of dimensionality,” significantly reducing the computational burden when solving high-dimensional matrices. In contrast to previous reinforcement learning approaches for controlling the optimal

frequency of hydropower-photovoltaic microgrid power systems, this method incorporates the consideration of disturbance factors, providing a robust theoretical basis for the grid integration of hybrid power generation systems.

The desired voltage and current is 50 hz sinusoidal waves, such that the system is dynamic with high frequency. Yet the IRL method depends on a process to collect the control and states data from the system under a quasi-optimal control, which may lead to the power oscillation and need more time to turn the system from the transient state to steady state. Therefore, the limitation of this method is that it is currently only applicable to offline systems.

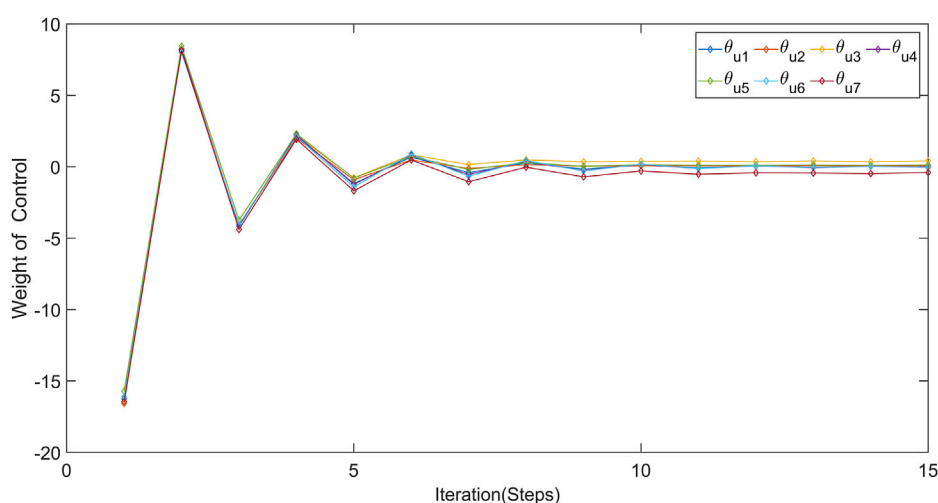


FIGURE 5
Convergence curves of the θ_u .

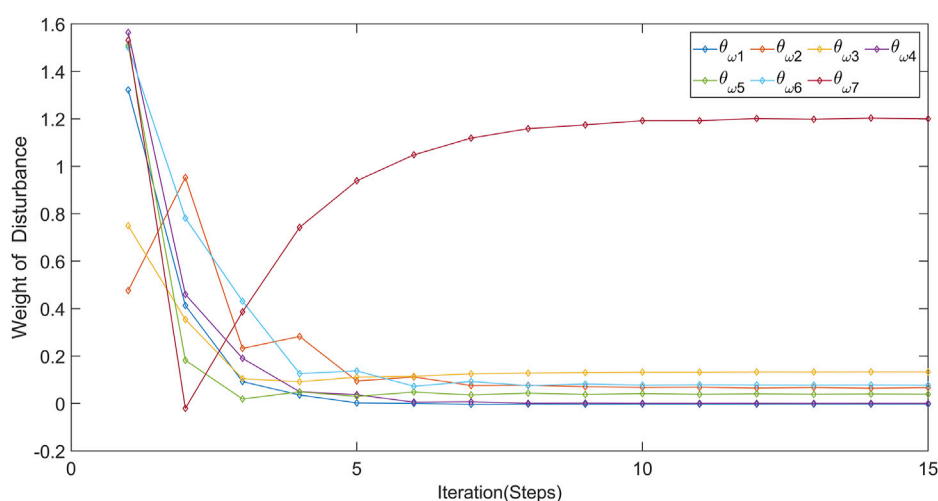


FIGURE 6
Convergence curves of the θ_w .

6 Conclusion

This paper focused on the hydropower-photovoltaic hybrid microgrid system and designed an optimal LFC using the IRL algorithm. First, the mechanism models of the hydro turbine generator and the photovoltaic generator were established, respectively. Second, a state-space model of the hydropower-photovoltaic hybrid microgrid system was developed, and based on the power generation characteristics, it was transformed in solving a zero-sum game problem. Third, the IRL algorithm was employed to approximate the Nash equilibrium point of the zero-sum game problem using three neural networks. Finally, the simulation experiments were conducted to verify the effectiveness of the proposed method.

Data availability statement

The raw data supporting the conclusions of this article will be made available by the authors, without undue reservation.

Author contributions

EW: Conceptualization, Writing–original draft, Writing–review and editing. LY: Methodology, Writing–review and editing. FZ: Writing–review and editing. XL: Writing–review and editing. JL: Writing–original draft. LS: Writing–original draft. MZ: Writing–original draft.

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Conflict of interest

Authors EW and LY were employed by Guoneng Qinghai Yellow River MaerDang Hydropower Development Co., Ltd.

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Research on static voltage stability enhancement for new energy station based on grid-forming control strategy

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The advent of novel power systems has given rise to a multitude of safety and stability concerns associated with the integration of emerging energy sources and power electronic equipment. The active support of the grid-forming control strategy represents an effective solution to the voltage and frequency stability issues associated with the weak damping and low inertia inherent to high-ratio new energy systems. Firstly, a static voltage stability index based on critical impedance is proposed for assessment of the static stability margin of a new energy grid-connected system, based on the static voltage stability theory of a traditional single-unit single-load system. Secondly, an analysis is conducted of the control principle of the grid-forming control converter and its impedance characteristics. In conclusion, a method for enhancing the static voltage stability margin of grid-connected new energy stations through parameter control of grid-forming converters is presented. The simulation verification of the single-feed and multi-feed systems demonstrates the efficacy and accuracy of the methodology presented in this study.

KEYWORDS

new power systems, high penetration of new energy, grid-forming control, converter impedance characteristics, static voltage stability

1 Introduction

The advent of a significant proportion of new energy has resulted in a shift in the grid structure, whereby traditional synchronous generator sets are no longer the dominant force. Instead, the grid is now co-dominated by both synchronous generator sets and non-synchronous generator sets, with the latter represented by new energy. The high proportion of new energy in the system gives rise to a series of frequency and voltage stability issues, largely attributable to the low inertia and damping characteristics inherent to the technology (Zhang et al., 2021; Chu and Teng, 2023; Linbin et al., 2020a; Zhou et al., 2024). The grid-forming control strategy, which has been the subject of much study, provides superior voltage and frequency support for high-ratio new energy systems (Mingxuan et al., 2022; Mahmoud et al., 2022; Gao et al., 2024; Chaoran et al., 2021). The transient stability of grid-forming converters has been the subject of extensive analysis in most studies (Jingyang et al., 2024; Teng and Xiongfei, 2021; Bo and Xiongfei, 2022). However, there is a paucity of research examining the impact of grid-forming converters on the static voltage stability of

the system. The term “static stability of power system voltage” is used to describe the long-term stability of voltage under a given operating condition for a given load and generation capacity (Yue et al., 2019; Ramirez and Murillo-Perez, 2006).

At present, some researchers are evaluating the static voltage stability of new energy power systems using the short circuit ratio (SCR) as a metric. However, most of these studies have identified a slight deficiency in their ability to accurately classify system stability based on the critical short circuit ratio (Li et al., 2021). Moreover, some studies employ the eigenvalue analysis and singular value decomposition methods for evaluation purposes. In Hao et al. (2021), the short-circuit ratio of a multi-feeder DC system is defined at the planning level, and the critical value of the metrics is determined through the sensitivity of the PV curves. However, no online analysis has been reported. In Jian et al. (2019), an online power/voltage stability metric is proposed for calculating the critical value of the metric by simplifying the multi-machine, multi-feed system into a single-machine system. Nevertheless, the method necessitates the acquisition of the Davening equivalent potentials of the nodes or the intra-generator potentials. In Linbin et al. (2020b), a generalised short-circuit ratio is proposed for the assessment of the static stability of the system based on a modal approach to decouple the multi-feed system; however, the calculation process is complex. A data-driven approach based on a moment-based spectral estimation method is proposed to facilitate an understanding of the voltage magnitude changes by synchronising the phasor measurement unit (PMU) data, with a view to taking appropriate measures to prevent system static voltage instability (Fan et al., 2021). A critical short-circuit ratio was proposed as a means of assessing the static voltage stability of power systems that integrate high-voltage direct current (HVDC) transmission (Lee and Andersson, 2016; Hao et al., 2020). In Adewuyi et al. (2019), a battery energy storage system (BESS) was employed for the optimal active and reactive power compensation of a continuously loaded power system, and a voltage stability assessment model was constructed that incorporates transmission line active and reactive current information. To address the increasingly prevalent problem of correlated uncertainty in power grids, a probabilistic modelling approach based on the inherent correlation of uncertain system parameters has been proposed as a means of achieving both static and dynamic voltage stability (Alzubaidi et al., 2024). The traditional continuous power flow (CPF) method is inadequate for accounting for the dynamic characteristics of renewable energy generation, resulting in significant errors in the estimation of renewable energy penetration into power systems. To address this limitation, an extended CPF method is proposed for This study considers the static voltage stability of a high-penetration new energy power system based on the extended Jacobi matrix (Tiezhu et al., 2024), considering the dynamic characteristics of the synchronous generator and the renewable energy generation in the AC constant-voltage control mode. To facilitate the analysis of the impact of the integration of inverter-based distributed generators in different control modes on static voltage stability, a system static voltage stability index is derived for the purpose of evaluating the load state of the distribution network from a complex current model (Yaran et al., 2021). To meet the requirements of voltage stability assessment and to address the issues of class imbalance and poor model generalisation in machine learning, an

online assessment method for static voltage stability is proposed in reference (ZhuJun et al., 2022). This method is based on the RUSBagging method, which addresses the issue of class imbalance to a certain extent and improves the assessment accuracy of a few class samples.

In conclusion, most studies on static voltage stability focus on the assessment of the static stability of new energy high-penetration systems. The calculation of related assessment methods is complex, and the accuracy of the results is difficult to guarantee. There is a paucity of studies examining the enhancement of static voltage stability in high-percentage new energy and high-percentage power electronic systems. Moreover, numerous scholars have investigated the transient stability characteristics and enhancement techniques of grid-forming converters, yet there is a paucity of in-depth research on static voltage stability. There is a research gap concerning the enhancement of static stability margins through the control of grid-forming parameters. The current methods for enhancing the static voltage stability margin, such as reinforcing the power grid infrastructure and implementing capacitance compensation, necessitate a significant investment in equipment addition or modification. Consequently, the deployment of these solutions in real-world scenarios is fraught with difficulties.

This paper examines the static stability limit power of single-unit single-load systems and single-unit single-feed systems from an impedance perspective, based on traditional static stability. It also determines the impedance magnitude corresponding to the static stability limit power of the new energy grid-connected system. Furthermore, it analyses the impedance characteristics of the grid-forming converter and demonstrates that altering relevant control parameters can modify the equivalent impedance, thereby magnifying limit output power and enhancing static voltage stability margin. Building upon this analysis, a novel method for enhancing static stability limits in new energy field stations based on grid-forming control is proposed. By adjusting control parameters, this approach effectively improves the static stability margin and enhances the grid-forming converter limit power at a low cost with practical applicability.

The contributions of this study are as follows:

- (1) Different from the traditional static voltage stability theory, the static voltage stability margin index based on critical impedance proposed in this paper has clear physical meaning and simple calculation.
- (2) Based on the control principle of the grid converter, the impedance characteristics of the converter are analysed using small signal perturbation method. It is found that the impedance can be modified by adjusting the sag coefficient of reactive voltage.
- (3) A method of increasing static voltage stability margin based on grid-forming control is proposed, which is simple, low cost and easy to be applied in practice.

The rest of this paper is organised as follows: Part II proposes a critical impedance-based static voltage stability margin assessment index for new energy grid-connected systems based on the traditional static voltage stability theory; Part III analyses the control principle of grid-forming converter and its impedance characteristics based on the small-signal model; Part IV elaborates the static voltage stability margin enhancement method for new

energy grid-connected systems based on the parameter control of converter; In part V, the method proposed in this paper is verified; Finally, the conclusion is made and the future work is expected.

2 Static stability limit theory analysis

2.1 Single machine with single load system

As shown in the figure is a single-machine, single-load system, E is the Davening-equivalent potential of the AC large system, Z_s is the system-equivalent impedance, and $P_L + jQ_L$ is the load power. After the power system is subjected to small perturbations, no spontaneous oscillations or non-periodic out-of-steps occur, and it can automatically recover to the initial operating state, and the power system is statically stable. When this system reaches the static critical stability, the maximum load power is shown in Equation 1.

$$P_{Emax} = \frac{EU_L}{Z_s} \quad (1)$$

When the load absorbed power exceeds this limit, then the system will be destabilised. Therefore, the static stable transmission power limit of the system should be enhanced as much as possible to improve the static stability margin and reduce the risk of static instability of the system as much as possible.

In this paper, the theory related to static stability of high percentage new energy systems is studied from the impedance perspective analysis, and the equivalent impedance of the load when the single-unit, single-load system reaches the static stability limit will be analysed in the following.

For a specific system, its parameters are certain, and when the load equivalent impedance is changed, its absorbed power will be changed. Let the load impedance be expressed as $Z_L \angle \alpha_L$, the system impedance be expressed as $Z_s \angle \alpha_s$, then the system output complex power can be expressed as

$$S_w = U_L I_L^* = \frac{Z_L (\cos \alpha_L + j \sin \alpha_L) E^2}{Z_L^2 + 2Z_L Z_s \cos(\alpha_L - \alpha_s) + Z_s^2} \quad (2)$$

where, U_L and I_L denote the load voltage and load current. Equation 2 reflects the complex power absorbed by the load from the system, it is not only related to the load impedance magnitude, but also affected by the impedance phase.

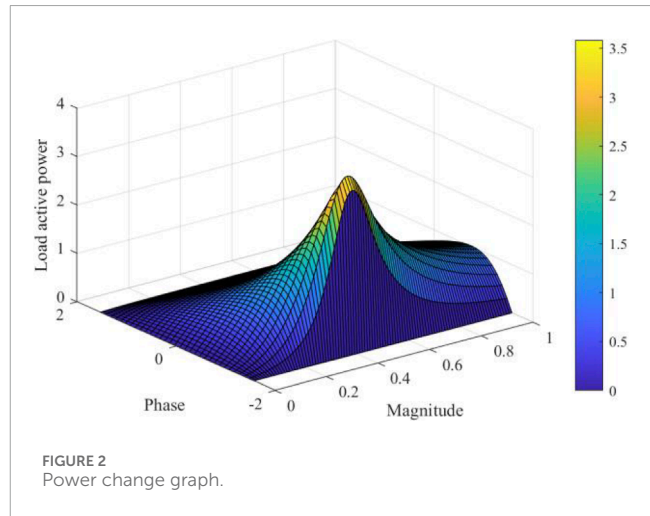
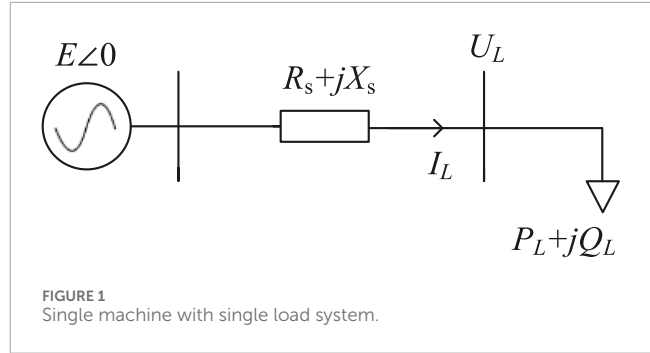
The active power absorbed by the load can be expressed as in Equation 3.

$$P_L = \frac{E^2 Z_L \cos \alpha_L}{Z_L^2 + 2Z_L Z_s \cos(\alpha_L - \alpha_s) + Z_s^2} \quad (3)$$

Taking the partial derivatives for the magnitude and phase angle respectively, we have

$$\begin{cases} \frac{\partial P_L}{\partial Z} = \frac{\cos \alpha_L}{\Delta^2} (Z_s^2 - Z_L^2) \\ \frac{\partial P_L}{\partial \alpha_L} = -\frac{(Z_s^2 + Z_L^2) \sin \alpha_L + 2Z_L Z_s \sin \alpha_s}{\Delta^2} \\ \Delta = Z_s^2 + Z_L^2 + 2Z_L Z_s \cos(\alpha_L - \alpha_s) \end{cases} \quad (4)$$

It can be shown in Equation 4 that the load power achieves its maximum value when and only when $Z_L = Z_s$ and $\alpha_L = -\alpha_s$, i.e.,



when the load impedance is equal in magnitude and opposite in phase to the system. The corresponding limit power is

$$P_{Lmax} = \frac{E^2}{4Z_s \cos \alpha_s} \quad (5)$$

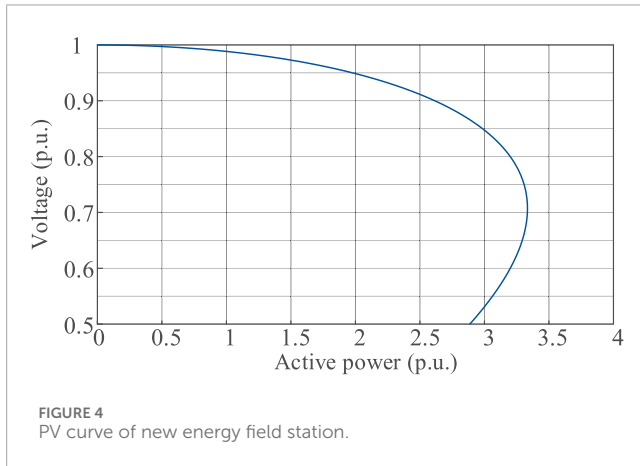
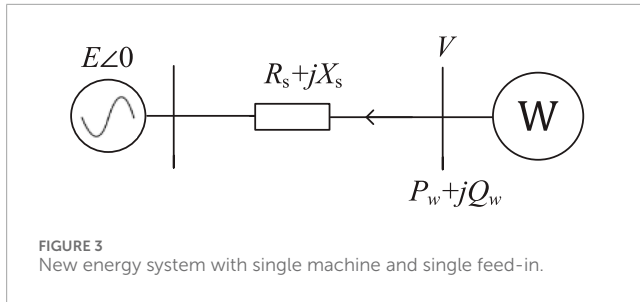
From Equation 5, reducing the system impedance and phase angle can improve the system limit output power.

SIMULINK is used to build a single machine and single load system as shown in Figure 1, the system potential $E = 1$, $R_s + jX_s = 0.4e^{j80^\circ}$, the load complex impedance is $Z_L e^{j\alpha}$, and the three-dimensional diagram of the absorbed active power with impedance magnitude and phase is shown below, which shows that when the load impedance is equal to the system impedance and the phase is opposite, the absorbed active power reaches the maximum value.

2.2 Single-machine and single-feed system

The new energy single-machine single-feed system is shown in Figure 2, where U_w , P_w , Q_w are the new energy station voltage, active output and reactive output, and the meaning of other symbols is the same as that in Figure 1.

Comparing Figures 3, 3, although the power transmission direction of the stand-alone single-load system is opposite to that of the new energy's stand-alone single-feeder system, they have similar structures, and therefore, they have similar properties.

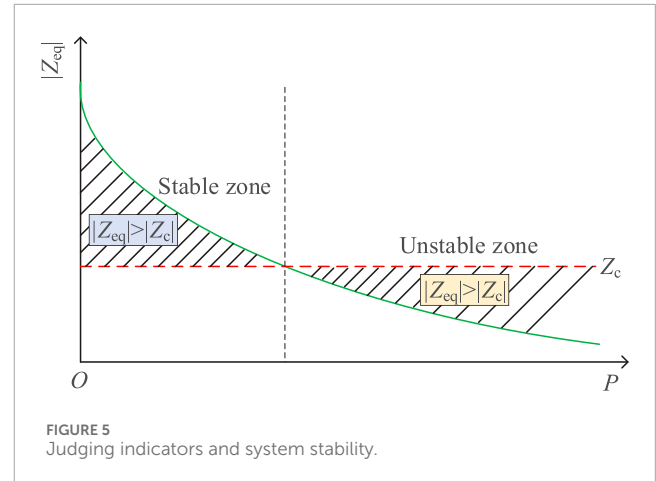


In the stand-alone single-load system, with the increase of the load power, a voltage dive will occur and lead to voltage instability, while in the stand-alone single-feeder system, a similar phenomenon will also occur. For example, the single machine single-feed system shown in Figure 3 is built in Simulink. Set $R_s + jX_s = 0 + 0.15j$ and $E = 1$ in Figure 3. Increase the active output of the new energy station W continuously, and the voltage and power of the new energy station is as follows:

As can be seen in Figure 4, with the increase of the new energy output power, the voltage of the new energy field station gradually decreases to reach the maximum output power of the new energy, and then continue to increase the active power, the voltage of the new energy field station will plummet, and there is a voltage destabilisation, which is similar to that of the stand-alone, single-feed-in system.

Further, considering the new energy single-feeder system, whether there is a similar single-feeder system, that is, the new energy unit equivalent impedance and the AC system amplitude is equal to the system is in a critical stable state, and can be based on the magnitude of the new energy unit equivalent impedance to determine the voltage stability of the system.

After the above analysis, the new energy grid-connected system has a similar conclusion with the load, when its equivalent impedance is equal to the system impedance, the new energy output power reaches the limit. When the equivalent impedance is greater than the system impedance, the power is less than the limit and the system is stable. When the equivalent impedance is smaller than the system impedance, the output power is greater than the stability limit, and the system is unstable. In this paper, the system impedance is called the critical impedance, based on which the static voltage



stability margin of the new energy station is defined.

$$Z_{\Delta} = Z_w - Z_{cr} \quad (6)$$

where Z_{Δ} is the static voltage stability margin expressed as an impedance vector, Z_w is the equivalent impedance of the new energy field station, and Z_{cr} is the system impedance or critical impedance.

Based on this index, static voltage stability determination criteria can be described as Equation 7.

$$\begin{cases} Z_{\Delta} > 0, & \text{stable} \\ Z_{\Delta} = 0, & \text{critical stable} \\ Z_{\Delta} < 0, & \text{unstable} \end{cases} \quad (7)$$

Figure 5 gives the curve of the impedance determination index defined in this paper versus the static voltage stability of the system.

3 Impedance characteristics of new energy via grid-forming converter for grid-connection

3.1 Grid-forming converter topology and control

Figure 6 shows a typical grid-forming converter topology and control method. Where: V_d is the DC side voltage of the converter, which can be identified as a constant value; e_a , e_b and e_c are the internal potentials of the converter; i_a , i_b and i_c are the output currents of the VSG; i_{ga} , i_{gb} and i_{gc} are the grid-connecting currents of the converter; v_a , v_b and v_c are the output terminals of the converter; and L_f , C_f and R_f are the filtering inductances, filtering capacitances and damping resistances of the converter, respectively; L_g and R_g are the equivalent line inductance and resistance of the grid, respectively; v_{ga} , v_{gb} and v_{gc} are the equivalent AC system voltages.

The active control strategy of the grid-forming converter draws on the inertia and primary frequency regulation characteristics of the synchronous generator; the reactive control strategy simulates the primary voltage regulation characteristics of the synchronous

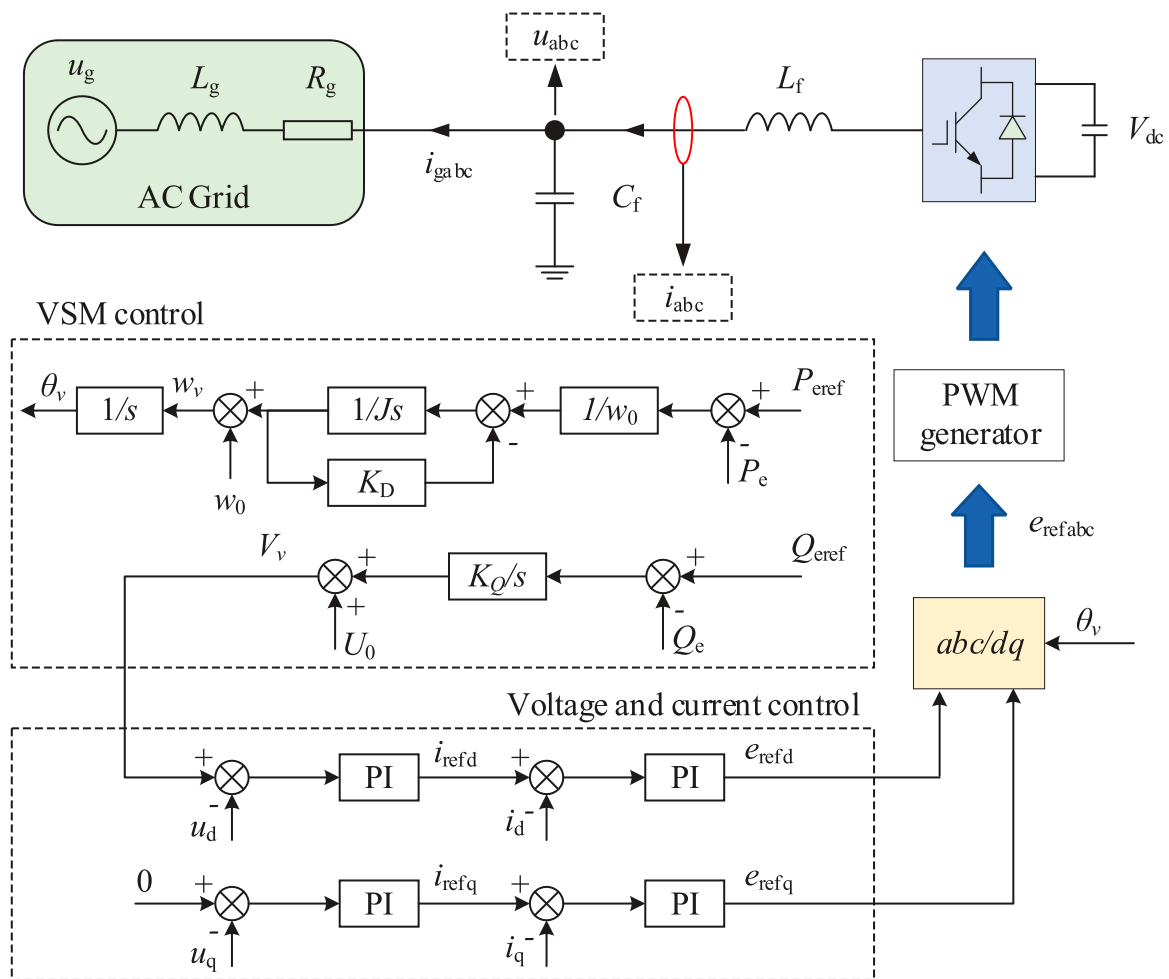


FIGURE 6
Main circuit topology and control method for VSGs.

generator. For the voltage-controlled converter, the following expressions are available:

$$\begin{cases} T_s + (\omega_N - \omega_v)D_p - T_e = Js\omega_v \\ T_e = \frac{P_e}{\omega_v} \approx \frac{P_e}{\omega_N} \\ T_e = \frac{P_s}{\omega_v} \approx \frac{P_s}{\omega_N} \\ -\omega_v = s\theta \\ Q_s + \sqrt{2}D_q(V_N - V) - Q_e = \sqrt{2}K_Q E_v \end{cases} \quad (8)$$

where: J is the virtual rotational inertia; ω_v and ω_N are the output angular frequency of the converter and the rated angular frequency of the grid, respectively; T_s and T_e are the given torque and the electromagnetic torque, respectively; D_p and D_q are the active damping coefficient and the reactive damping coefficient, respectively; P_s and Q_s are the active and reactive power given, respectively; P_e and Q_e are the instantaneous output active and reactive powers, respectively are the instantaneous output active power and reactive power; θ is the phase of the inverter's internal potential; K_Q is the reactive inertia coefficient; V_N and V are the

rated voltage RMS and output voltage RMS, respectively; and E_v is the rms value of the inverter's internal potential.

The output active power P_e and reactive power Q_e of the converter can be calculated according to the instantaneous power theory, which is given by Equation 9.

$$\begin{cases} P_e = 1.5(v_\alpha i_\alpha + v_\beta i_\beta) \\ Q_e = 1.5(v_\beta i_\alpha - v_\alpha i_\beta) \end{cases} \quad (9)$$

where: i_α , i_β are the output current of the converter in the $\alpha\beta$ coordinate system; v_α , v_β are the output voltage of the converter in the $\alpha\beta$ coordinate system.

The modulating waveform of VSG is jointly determined by the outputs of active and reactive controllers with the Equation 10.

$$\begin{cases} e_{ma} = \sqrt{2}E_v \cos \theta \\ e_{mb} = \sqrt{2}E_v \cos \left(\theta - \frac{2\pi}{3} \right) \\ e_{mc} = \sqrt{2}E_v \cos \left(\theta + \frac{2\pi}{3} \right) \end{cases} \quad (10)$$

3.2 Impedance characteristics of grid-forming converter control system

The converter is controlled in a stationary coordinate system and therefore cannot be modelled in a synchronous rotating coordinate system for small signal linearisation. In this study, the harmonic linearisation method is used to derive the positive sequence output impedance model of the converter. The f_p frequency positive sequence voltage perturbation is injected at the point of common coupling (PCC), and according to the 2-fold mirror frequency coupling effect, the main circuit can be decomposed into f_p frequency and $f_p - 2f_1$ frequency equivalent circuits, and its small-signal model is

$$\begin{cases} (sL_f + R_f)i_a[f] = e_a[f] - (s^2L_fC_f + sR_fC_f + l)v_a[f] \\ f = \pm f_p \end{cases} \quad (11)$$

$$\begin{cases} (sL_e + R_e)i_a[f] = v_a[f] - v_{ap}[f] \\ f = \pm(f_p - 2f_1) \end{cases} \quad (12)$$

where the expressions for the components of the filter capacitor voltage $v_a[f]$, output current $i_a[f]$, and inductor current $i_{La}[f]$ at different frequencies are shown in Equations 13–15.

$$v_a[f] = \begin{cases} V_1, & f = \pm f_1 \\ V_p, & f = \pm f_p \\ V_{p2}, & f = \pm(f_p - 2f_1) \end{cases} \quad (13)$$

$$i_a[f] = \begin{cases} I_1, & f = \pm f_1 \\ I_p, & f = \pm f_p \\ I_{p2}, & f = \pm(f_p - 2f_1) \end{cases} \quad (14)$$

$$i_{La}[f] = \begin{cases} I_{L1}, & f = \pm f_1 \\ I_{Lp}, & f = \pm f_p \\ I_{Lp2}, & f = \pm(f_p - 2f_1) \end{cases} \quad (15)$$

where: $V_1 = V_1/2$; $V_p = (V_p/2)e^{j\phi_{vp}}$; $V_{p2} = (V_{p2}/2)e^{j\phi_{vp2}}$; $I_1 = (I_1/2)e^{j\phi_{i1}}$; $I_p = (I_p/2)e^{j\phi_{ip}}$; $I_{p2} = (I_{p2}/2)e^{j\phi_{ip2}}$; $I_{L1} = (I_{L1}/2)e^{j\phi_{iL1}}$; $I_{Lp} = (I_{Lp}/2)e^{j\phi_{iLp}}$; $I_{Lp2} = (I_{Lp2}/2)e^{j\phi_{iLp2}}$. X_1, X_p , and X_{p2} are the amplitudes or initial phase angles of the fundamental frequency, the positive-order perturbation component, and the negative-order coupling component of the associated steady state value of X . The amplitudes or initial phase angles of the positive-order perturbation component and the negative-order coupling component of the associated steady-state value of X are the amplitudes or initial phase angles.

The positive sequence voltage perturbation passes through the power control loop causing a phase angle perturbation quantity $\Delta\varphi$ such that $\varphi = \varphi_1 + \Delta\varphi$, where $\varphi_1 = \omega_{1t}$ is the fundamental voltage phase angle.

$v_a[f]$, $i_a[f]$ and $i_{La}[f]$ are park transformed to obtain the voltage-current expression in the dq coordinate system as shown in Equations 16–21.

$$v_d[f] = \begin{cases} V_d, & f = 0 \\ \Delta v_{d0} + V_q\Delta\varphi, & f = \pm(f_p - f_1) \end{cases} \quad (16)$$

$$v_q[f] = \begin{cases} V_q, & f = 0 \\ \Delta v_{q0} - V_d\Delta\varphi, & f = \pm(f_p - f_1) \end{cases} \quad (17)$$

$$i_d[f] = \begin{cases} I_d, & f = 0 \\ \Delta i_{d0} + I_q\Delta\varphi, & f = \pm(f_p - f_1) \end{cases} \quad (18)$$

$$i_q[f] = \begin{cases} I_q, & f = 0 \\ \Delta i_{q0} - I_d\Delta\varphi, & f = \pm(f_p - f_1) \end{cases} \quad (19)$$

$$i_{Ld}[f] = \begin{cases} I_{Ld}, & f = 0 \\ \Delta i_{Ld0} + I_{Lq}\Delta\varphi, & f = \pm(f_p - f_1) \end{cases} \quad (20)$$

$$i_{Lq}[f] = \begin{cases} I_{Lq}, & f = 0 \\ \Delta \Delta i_{Lq0} - I_{Ld}\Delta\varphi, & f = \pm(f_p - f_1) \end{cases} \quad (21)$$

where: $V_d = V_1, V_d = 0, \Delta v_{d0} = V_p + V_{p2}, \Delta v_{q0} = \mp jV_p \pm jV_{p2}, I_d = I_1 \cos \phi_{i1}, I_q = I_1 \sin \phi_{i1}, \Delta i_{d0} = I_p + I_{p2}, \Delta i_{q0} = \mp jI_p \pm jI_{p2}, I_{Ld} = I_{L1} \cos \phi_{i1}, I_{Lq} = I_{L1} \sin \phi_{i1}, \Delta i_{Ld0} = I_{Lp} + I_{Lp2}, \Delta i_{Lq0} = \mp jI_{Lp} \pm jI_{Lp2}$.

After neglecting the high frequency components, the frequency domain expressions for active and reactive power of the converter are obtained by Equations 22, 23 using the convolution theorem in the frequency domain:

$$P_e[f] = \begin{cases} 1.5(V_d I_d + I_q V_q) f = 0 \\ 1.5(I_d \Delta v_{d0} + V_d \Delta i_{d0} + I_q \Delta v_{q0} + V_q \Delta i_{q0}) \\ f = \pm(f_p - f_1) \end{cases} \quad (22)$$

$$Q_e[f] = \begin{cases} 1.5(V_q I_d - I_q V_d) f = 0 \\ 1.5(I_d \Delta v_{q0} + V_q \Delta i_{d0} - I_q \Delta v_{d0} - V_d \Delta i_{q0}) \\ f = \pm(f_p - f_1) \end{cases} \quad (23)$$

According to the mathematical model of the power control loop in Equation 8, ignoring the small signal quantity of the quadratic term, we obtain the frequency domain expressions shown in Equations 24, 25 for φ and E_v

$$\varphi[f] = \begin{cases} \varphi_1, & f = 0 \\ G_{pp}(s)(I_d \Delta v_{d0} + V_d \Delta i_{d0} + I_q \Delta v_{q0} + V_q \Delta i_{q0}) \\ f = \pm(f_p - f_1) \end{cases} \quad (24)$$

$$E_v[f] = \begin{cases} E_{v0}, & f = 0 \\ G_{pq}(s)(I_d \Delta v_{q0} + V_q \Delta i_{d0} - I_q \Delta v_{d0} - V_d \Delta i_{q0}) + H_p(s) \Delta v_{d0} \\ f = \pm(f_p - f_1) \end{cases} \quad (25)$$

where: $G_{pq}(s) = -(3/2)/(K_s); H_p(s) = -D_q/(K_s); G_{pp}(s) = -(3/2)/[(J_s + D_p)\omega_n s]$.

The voltage-current control dual-loop voltage loop reference is generated by the power control loop, where $v_{dref}[f] = E_v[f]$ and $v_{qref}[f] = 0$, and is inverted by Park's coordinate inverse transformation to obtain the a-phase bridge-arm voltage response

$e_a[f]$ as shown in Equation 26

$$e_a[f] = \begin{cases} V_1 + (sL_f + R_f)I_1 \\ f = \pm f_1 \\ M_p(s \mp j\omega_1)V_p + N_p(s \mp j\omega_1)I_p \\ + (-G_{pi}(s \mp j\omega_1) + j\omega_n L_f)I_{LP} + L_p(s \mp j\omega_1)I_{p2} \\ f = \pm f_p \\ M_{p2}(s \pm j\omega_1)V_p + N_{p2}(s \pm j\omega_1)I_p \\ + (-G_{pi}(s \pm j\omega_1) + j\omega_n L_f)I_{LP2} + L_{p2}(s \pm j\omega_1)I_{p2} \\ f = \pm(f_p - 2f_1) \end{cases} \quad (26)$$

where $M_p(s \mp j\omega_1)$, $N_p(s \mp j\omega_1)$, $L_p(s \mp j\omega_1)$, $M_{p2}(s \pm j\omega_1)$, $N_{p2}(s \pm j\omega_1)$, $L_{p2}(s \pm j\omega_1)$ are the coefficients of voltage-current related components.

To eliminate the $f_p - 2f_1$ frequency coupling current, (Equation 26), is substituted into (Equation 12) to obtain I_{p2} with respect to the filter capacitor voltage and output current expressions:

$$I_{p2} = \frac{M_{p2}(s \mp j\omega_1)V_p + N_{p2}(s \mp j\omega_1)I_p}{D_{p2}(s \mp j\omega_1)} \quad f = \pm f_p \quad (27)$$

where $D_{p2}(s)$ can be described as Equation 28.

$$D_{p2}(s) = [(s \mp j\omega_1)L_e + R_e] - L_{p2}(s) + \{[(s \mp j\omega)L_f + R_f] - G_{pi}(s) - j\omega_n L_f\} \times \{[(s \mp j\omega_1)L_e + R_e](s \mp j\omega_1)C_f + 1\} \quad (28)$$

Bringing Equation 27 into Equation 26 and combining Equations 11, 12, the positive sequence impedance of the converter is obtained by solving as shown in Equation 29.

$$Z_p(s) = \frac{N_p(s - j\omega_1) + L_p(s - j\omega_1) \frac{N_{p2}(s - j\omega_1)}{D_{p2}(s - j\omega_1)} + [j\omega_n L_f - G_{pi}(s - j\omega_1) - (sL_f + R_f)]}{M_p(s - j\omega_1) + L_p(s - j\omega_1) \frac{M_{p2}(s - j\omega_1)}{D_{p2}(s - j\omega_1)} - 1 + [j\omega_n L_f - G_{pi}(s - j\omega_1) - (sL_f + R_f)]sC_f + sL_e + R_e} \quad (29)$$

By analyzing the impact of each control parameter in the impedance expression on the amplitude and phase of the converter impedance, it is evident that only the reactive voltage sag coefficient exerts significant influence. Therefore, this study selects the reactive voltage sag coefficient as the variable for regulating converter impedance. The influence of other relevant parameters on grid-forming converter impedance characteristics is minimal and can be disregarded. It should be noted that when compared with (Cespedes and Jian, 2014), under identical parameter conditions, the equivalent impedance of a grid-forming converter is lower than that of a grid-following converter, thereby reducing its static voltage stability margin to some extent after connection to the system. Consequently, urgent research into methods for improving static voltage stability in grid-forming converter grid-connection scenarios at this stage is warranted.

To verify the correctness of the conclusions of this paper based on the impedance characteristics of the converter, a simulation model of a single machine single-feed system as shown in Figure 3 is constructed. It is proved that the reactive voltage sag coefficient has a large influence on the magnitude and phase of

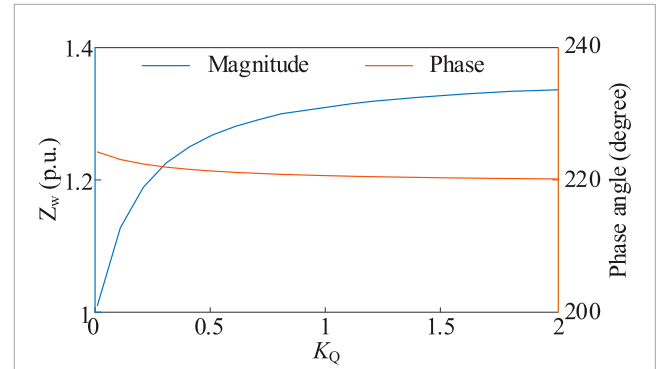


FIGURE 7 Effect of reactive voltage sag coefficient on converter impedance characteristics.

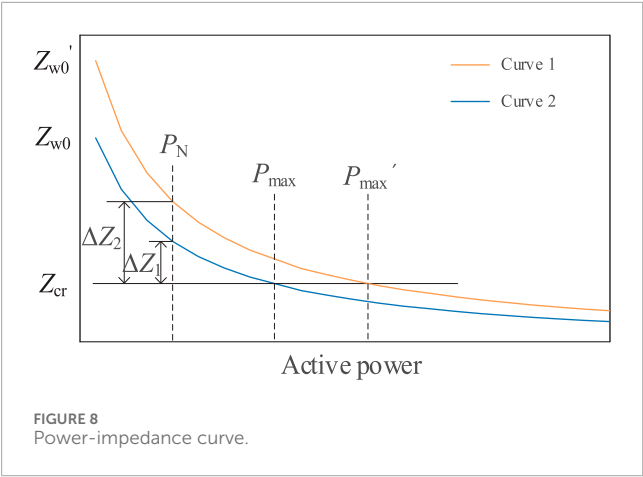
the impedance. The Figure 7 shows the influence of the reactive voltage sag coefficient on the impedance magnitude and phase of the field station.

It can be seen from Figure 7 that the reactive voltage sag coefficient has little effect on the phase of the power station impedance and a larger effect on the amplitude. Therefore, the influence of the control parameter on the impedance phase is no longer considered in this paper. The following section further proposes the static voltage stability limit enhancement strategy based on the matching of impedance parameters by changing the control parameter to affect the impedance replication of the converter.

4 Static stability margin improvement method based on converter parameter control

When the power transmitted by the new energy unit to the system increases, its equivalent impedance will gradually decrease, and the corresponding static stability voltage margin will be reduced. Based on the theoretical basis that the converter control parameters can affect its impedance characteristics, this paper proposes the method of static stability margin improvement as follows: Under the current working condition, the impedance at the current output power of the same power is improved by changing the control parameters. When the output power of the new energy is increased to the next level, the corresponding impedance is increased compared with that before the parameter change. Then, according to Equation 6, the static voltage stability margin of the new energy is increased ($\Delta Z_2 > \Delta Z_1$), and the limit output power is increased. Figure 8 shows the power impedance schematic curve when the static voltage stability margin is increased by the proposed method.

The vertical axis of the graph is the equivalent impedance magnitude of the converter, and the horizontal axis is the output power, the impedance magnitude decreases with the increase of power. Z_{cr} is the equivalent impedance when the power output reaches the critical limit. In the initial working condition, the impedance power curve and the critical impedance curve to



determine the limit power for $P_{\max}$, by changing control parameters to make its initial impedance from Z_{w0} increased to Z_{w0}' , the curve and the critical impedance curve to determine the limit power for $P_{\max}'$, obviously $P_{\max} > P_{\max}'$.

The objective of this paper is to enhance the static voltage stability margin of a new energy grid-connected system. This is achieved by controlling the parameters to increase the amplitude of the converter impedance at the output of the same power, thereby expanding its margin with the critical impedance and attaining a higher static stability margin.

5 Simulation verification

To validate the static stability assessment index and enhancement strategy proposed in this paper, this section analyses the static voltage stability margin and enhancement effect of the system before and after applying the strategy in this paper based on the new energy single-feed system and multi-feed system, in which the new energy station takes the wind farm as an example and adopts the grid-forming control strategy.

5.1 Single-feed new energy system

Firstly, the applicability of the strategy proposed in this paper is analysed in a simple single-fed system as shown in Figure 9, and the main parameters of the grid-forming converter are shown in Table 1. The AC system impedance is $2j$, and the system voltage is 1. All the simulations in this paper are carried out under the reference value of 100MVA, and the voltage level is set to 110V.

In the Table 1, V_d denotes the inverter DC voltage, L_f , R_f and C_f denote the filter inductance, resistance and capacitance on the AC side, respectively. L_e and R_e denote the line inductance and resistance. f_s is the switching frequency of the inverter. k_{ip} and k_{ii} denote current inner loop proportional integral coefficients. k_{vp} and k_{vi} denote voltage outer loop proportional integral coefficient. The meaning of the remaining variables has been described in the previous content.

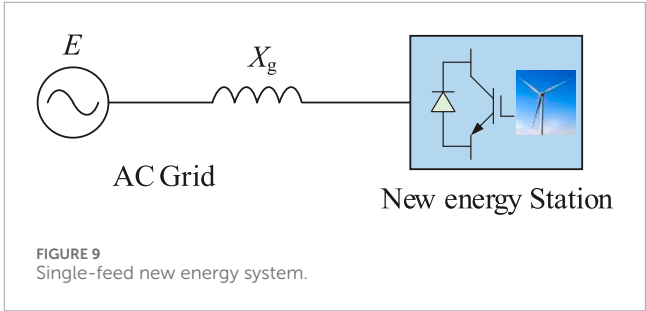


TABLE 1 Main parameters of grid-forming converter.

Parameter	Value	Parameter	Value
V_d/V	250	$J/kg \cdot m^2$	0.3
E_v/V	115	D_p	12
L_f/mH	3.2	K_Q	11
R_f/Ω	0.1	D_q	280
$C_f/\mu F$	18	k_{ip}	5
L_e/mH	0.065	k_{ii}	145
R_e/Ω	0.04	k_{vp}	4
V_N/V	110	k_{vi}	8
f_i/Hz	50	f_s/kHz	10

TABLE 2 Initial parameters of the station.

Station serial number	Output power (MW)	Equivalent impedance amplitude (p.u.)
Station 1	62.3	12.6
Station 2	17.8	44.2

In the system based on the initial parameters listed in Table 2, the outgoing power of the new energy field station is continuously increased by increasing the impedance amplitude and the corresponding impedance amplitude is recorded until the system becomes unstable and the static stabilization limit power is recorded. The equivalent impedance and limit power of the new energy field station at the original converter control parameters and after changing the parameters are given in Table 3. According to the results, the equivalent steady-state impedance of the new energy field station can be improved to a certain extent by changing the reactive voltage sag factor, i.e., the static voltage stability margin is improved. Although the improvement of the steady state margin at steady state is small, the improvement for the ultimate power is still considerable.

Figure 10A gives the curve of the PCC voltage with the output power of the new energy field station during the whole process

TABLE 3 Static stability indicators under different control parameters.

Parameter	Original parameter	Situation 1	Situation 2
Steady-state impedance (p.u.)	10.19	10.35	10.52
impedance margin (p.u.)	8.19	8.35	8.52
Critical power (MW)	142.9	164.5	193.7

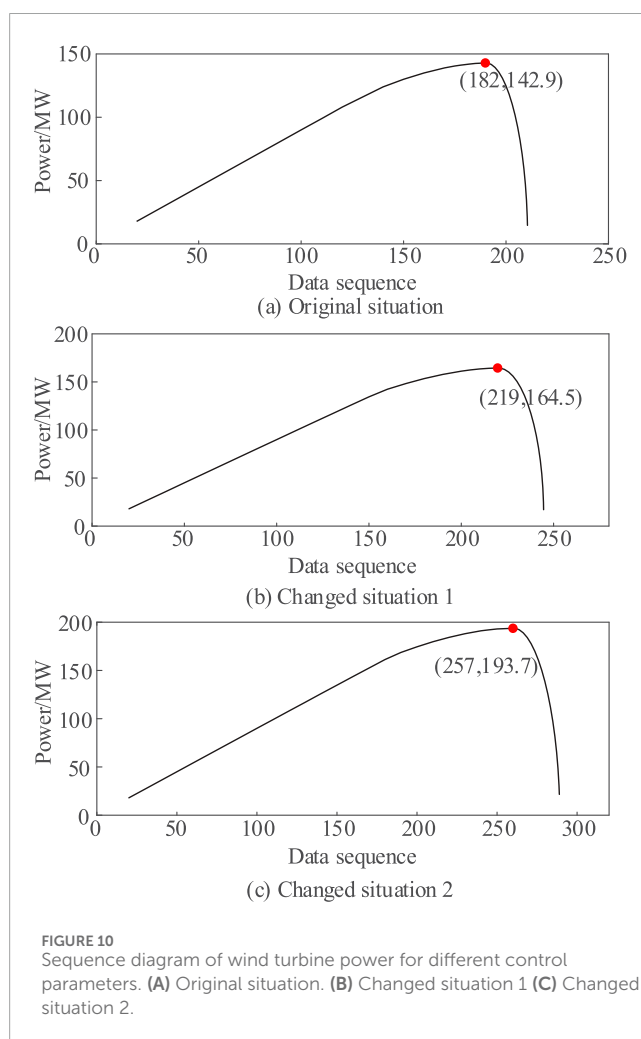


FIGURE 10 Sequence diagram of wind turbine power for different control parameters. (A) Original situation. (B) Changed situation 1 (C) Changed situation 2.

under the original control parameters, and the power decreases rapidly after reaching the maximum value, indicating that the phenomenon of static voltage destabilisation occurs. At this time, the corresponding voltage is the critical stabilisation voltage, and the output impedance of the new energy field station is 2.04 (equivalent to the high-voltage side of the system, the same below), which is equal to the size of the AC system impedance, which verifies the static voltage stabilisation method of critical stabilisation impedance proposed in this paper.

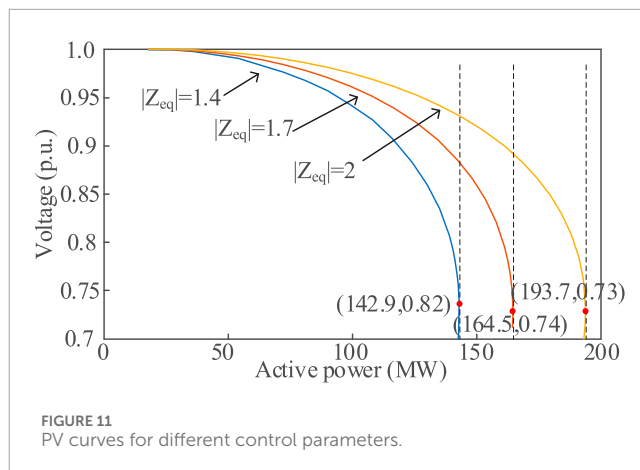


FIGURE 11 PV curves for different control parameters.

Now, the control parameters of the grid-forming converter are adjusted to verify whether the control strategy proposed in this paper can improve the new energy static voltage stability margin. The reactive voltage sag coefficients are changed to 15 and 22, at which time the voltage and impedance in the process of delivering power increase are recorded, which are given in Figures 10B, C, respectively. The power-voltage curves under three different parameter conditions are given in Figure 11. According to the results, the limiting output power of the new energy field station is increased after changing the control parameters, and the impedance amplitude of the field station is larger at the same power, and the static voltage stability margin characterised by impedance is higher. The converter impedances of the three curves at the critical point are 2.01 p.u. and 1.98 p.u., which are close to the critical impedance of 2 p.u., respectively.

5.2 Multi-feed new energy system

Before validating the effectiveness of the methodology of this paper in multi-feed-in systems, the changes in the limit power and the total power limit of the section are briefly analysed for each field station after applying the proposed strategy.

If the static stability margin of the current field station 2 is low, its static stability limit needs to be improved. When the control parameters are changed so that X_2 is increased, for the new energy field station 1, the equivalent system impedance is calculated by Equation 30.

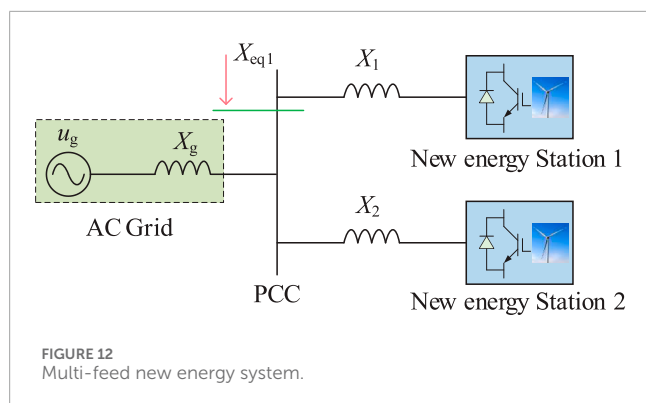
$$X_{eq1} = \frac{X_2 X_g}{X_2 + X_g} \quad (30)$$

The static stability margin of the new energy field station 1 is obtained by Equation 31.

$$Z_{\Delta} = X_1 - X_{eq} \quad (31)$$

The increase of the grid-connected impedance of station 2 will lead to the increase of the equivalent impedance X_{eq} , which will lead to the decrease of the static stability margin of station 1, thus increasing the risk of static voltage instability of the system.

Therefore, it is set in this paper that in a multi-site grid-connected system, the site with smaller static voltage stability margin



should be adjusted with priority; Secondly, after adjusting the parameters, it should be verified whether the margin of each site meets the requirements.

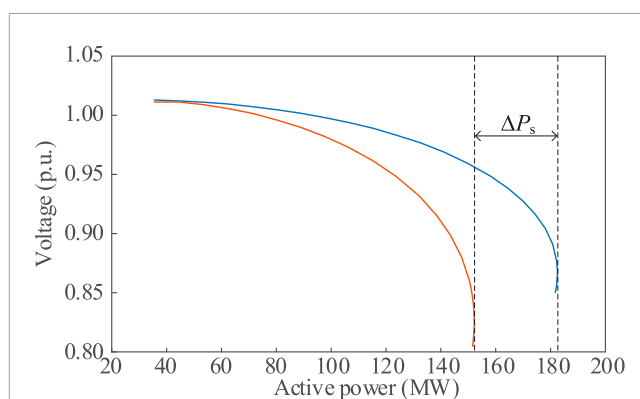
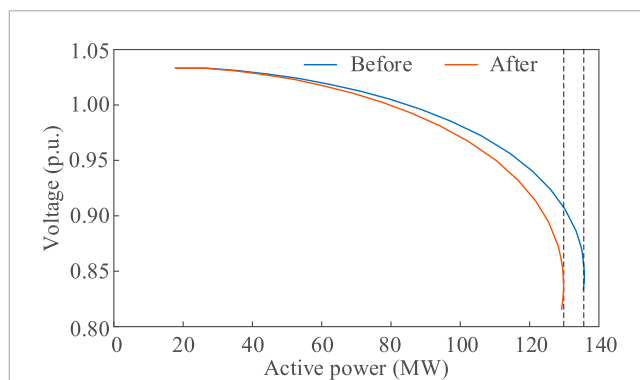
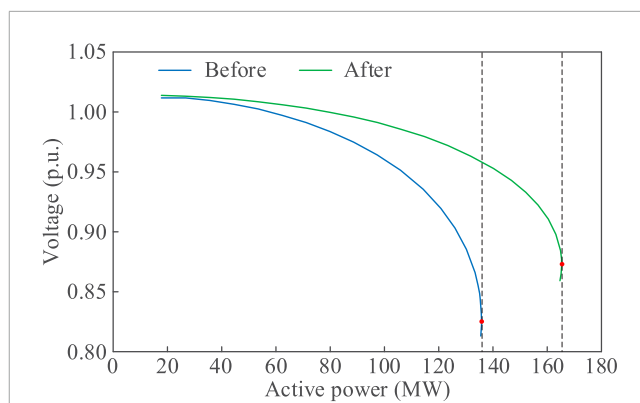
It is noted from Equation 32 that although increasing the impedance of a new energy station individually reduces the static voltage stability margins of other grid-connected stations to a certain extent, the overall static stability margins are improved from the point of view of the whole aggregation station, which can withstand a higher power delivery limit.

$$\begin{cases} Z_{\Delta 1} = X_1 - \frac{X_g X_2}{X_g + X_2} \\ Z_{\Delta 2} = X_2 - \frac{X_g X_1}{X_g + X_1} \\ Z_{\Delta g} = \frac{X_1 X_2}{X_1 + X_2} - X_g \end{cases} \quad (32)$$

The application of the method proposed in this paper in the multi-field station grid-connected system is now described. The new energy multi-feed-in system shown in Figure 12 is constructed, the converters are all network-controlled, and the initial control parameters of each station are the same (the same as the single-machine single-feed-in system). The power reference value of the new energy multi-feeder system is 100MVA, the voltage level is 350V, the impedance of the system is $1j$, and the relevant parameters of each station under the initial working condition are shown in the following table.

The power-voltage curves of the two stations and the pooling bus are shown in Figures 13–15, respectively, when the control parameters are changed so that the equivalent impedance of Station 1 is increased. After changing the parameters, the static stability margin of the new energy station 1 increases due to the increase of its own impedance, while the margin of the station 2 decreases slightly, but the limit power sent from the pooling bus increases compared with the previous one. The power curves of the pooling bus before and after the parameter change are shown in Figure 16.

As can be seen from Figures 13–15, changing the control parameters of a station can appropriately increase the static stability limit margin of that station, but it will cause the static voltage stability margin of other stations to slightly decrease. Overall, the static phone stability margin of the whole grid-connected system is improved greatly, which also proves the feasibility of the method proposed in this paper. The power sequence before



and after in Figure 16 shows that the static stability margin of the system is improved after the converter control parameters are changed.

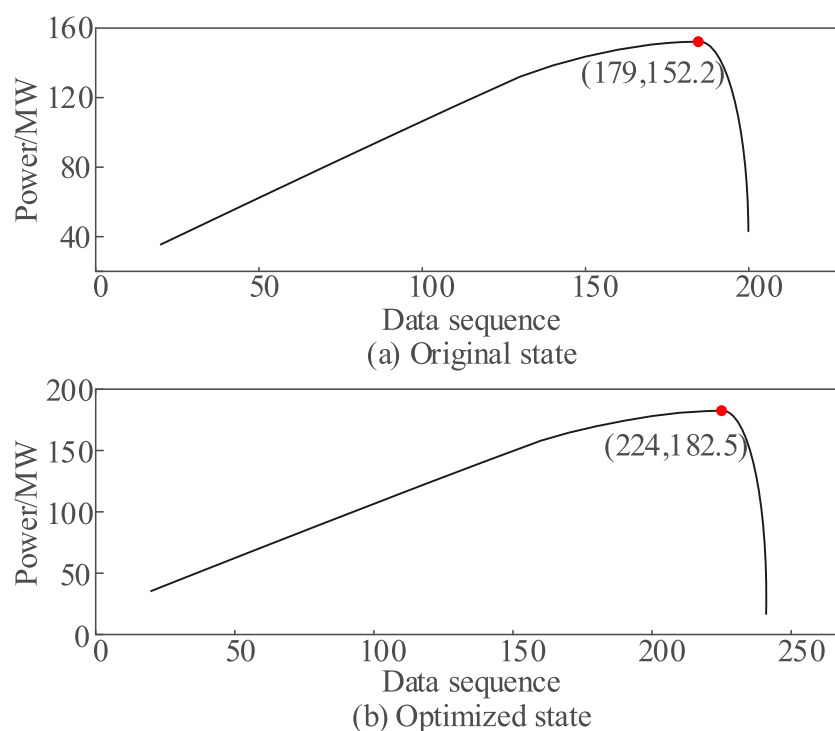


FIGURE 16
Power sequences at PCC. (A) Original state. (B) Optimized state.

6 Conclusion

This study employs an analytical approach to investigate the impedance characteristics of the grid-forming converter, with a view to identifying the control parameters that exert the most significant influence on its impedance amplitude. Based on these findings, a static voltage stabilization enhancement method is put forward, which is predicated upon the impedance characteristics of the grid-forming converter connected to the grid. The following conclusions are obtained:

- (1) The critical impedance-based static voltage stability limit evaluation index proposed in this paper can effectively realize the evaluation of the static voltage stability margin of new energy field stations connected to the grid, with clear physical meaning and simple calculation.
- (2) The adoption of the virtual synchronous machine control strategy by the grid-forming converter allows for the manipulation of the reactive voltage sag coefficient, which in turn affects the magnitude and phase of its impedance. Consequently, the static voltage stability margin of the system can be enhanced by modifying this parameter.
- (3) The static voltage stability margin enhancement strategy based on the impedance characteristics of grid-forming converters is applicable not only to simple single-fed new energy systems, but also to multi-site-fed systems. The strategy has a broad range of applications, straightforward implementation steps, and a minimal calculation volume. No additional equipment is necessary, and the financial outlay is minimal, facilitating practical implementation.

- (4) The present study examines the static stability of grid-forming converters that are controlled by a virtual synchronous generator. Subsequently, the grid-forming converters under other control strategies, including droop control and virtual oscillator, will be subjected to further analysis.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

Author contributions

SW: Writing–original draft, Writing–review and editing, Conceptualization. LG: Writing–original draft. GY: Writing–original draft. WQ: Writing–review and editing, Project administration. CH: Writing–review and editing, Supervision. HX: Writing–review and editing, Software. ZW: Writing–review and editing, Project administration.

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Conflict of interest

Authors SW, LG, GY, WQ, CH, HX, and ZW were employed by State Grid Jiangsu Electric Power Co., Ltd.

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The impact of current-loop control parameters on the electromagnetic transient voltage performance of voltage-source converter

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Since the large-scale integration of renewable energy sources into the AC grid has led to a relative decline in the voltage support capacity of the grid and the deterioration of the voltage dynamic at the grid connection point, especially under fast-scale conditions, the voltage disturbance has become more obvious. To improve the dynamic characteristics of the electromagnetic transient voltage at the grid connection point, this paper uses a practical dynamic damping method to analyze the impact of the converter current inner loop, feedforward voltage, and other links on the dynamic performance of the electromagnetic transient voltage. First, the current inner loop dynamic in the converter's synchronous coordinate is converted into an equivalent transfer function in the stationary coordinate, and the transfer function between the transient voltage disturbance at the grid connection point and the inner loop current output is established. On this basis, the Bode diagram and the vector diagram of the transfer function in the weak damping frequency band are used to analyze the dynamic damping of the current inner loop parameters and voltage feedforward filter parameters on the voltage disturbance at the grid connection point. The results indicate that moderately increasing the current inner loop bandwidth or reducing the feedforward filter bandwidth can help enhance the electromagnetic transient voltage stability of the grid connection point, but increasing the current inner loop bandwidth will worsen the low-frequency damping characteristics and reduce the feedforward filter bandwidth will still help increase low-frequency damping.

KEYWORDS

renewable energy, converter, transient voltage, dynamic damping, current loop

1 Introduction

The dynamic characteristics of frequency and voltage have always been critical characteristics of power quality in the power system. In traditional power systems, the focus is mainly on the electromechanical time-scale frequency dynamics dominated by the synchronous machine rotor, while in the electromechanical time-scale, voltage dynamics are often described using algebraic equations (Kundur, 1994; Wu et al., 2023).

Therefore, traditional power systems lack research and attention to faster scale voltage dynamics (Vu et al., 1995; Cheng et al., 2024), and the practical factor is that the synchronous machine excitation control system provides sufficient voltage support capability (Huang et al., 2018; Löf et al., 1993).

With the large-scale integration of power electronic devices such as photovoltaics, wind power, flexible DC, and flexible AC transmission into the power grid, the sensitivity of the power grid response to electromagnetic time-scale disturbances continues to increase. Taking the doubly fed induction generator (DFIG) as an example, the rapid fluctuation of grid voltage can trigger negative sequence magnetic potential fluctuations in the electromagnetic time scale of DFIG, making it difficult to achieve the low voltage ride through (LVRT) control goal smoothly (Xu et al., 2012; Li et al., 2022). In addition, multi-scale cascade control structures are commonly used in various new energy generation equipment types. For example, new energy converters commonly use control loops such as phase-locked loops (PLL), current inner loops, and power outer loops, which require the dynamic analysis of grid voltage affected by multi-scale cascade control of converters to be extended to more time scales (Zhou et al., 2014; Hu et al., 2023). In summary, as the scale of power electronic equipment integration continues to increase, the dynamic characteristics of electromagnetic time scale grid voltage will become one of the focuses of power quality issues (Milano, 2016; Hansen et al., 2019; Breyer et al., 2022).

The voltage stability analysis in traditional power systems often uses the quasi-steady state phasor method to study the relationship between reactive power and voltage amplitude in power flow equations (Xie et al., 2012; Simpson-Porco and Bullo, 2016; Song et al., 2019; Li et al., 2021; Chu and Teng, 2023; Chen et al., 2019; Chen et al., 2020). However, the multi-scale cascade control of power electronic equipment is aimed at controlling instantaneous values. Therefore, traditional voltage stability analysis models based on vector methods make it difficult to explain the impact of power electronic control loops on the dynamic characteristics of power grid voltage at the electromagnetic transient time scale (Zhao et al., 2016).

Generally, the voltage dynamics of the power grid are mainly affected by the reactive power or current output. For the widely used grid following converters, their electromagnetic transient scale exhibits external characteristics of the current source, and the control characteristics of instantaneous current will directly affect the dynamic characteristics of the power grid voltage. Zhao et al. (2016) conducted a preliminary analysis of the dynamic voltage characteristics at the electromagnetic scale, but the modeling process lacked detailed derivation, and the proposed scheme only improved the high-frequency characteristics while ignoring the potential adverse effects in the low-frequency range.

This article provides a detailed derivation of the transformation formula for the inner loop controller from the rotating coordinate system of the control system to the three-phase stationary coordinate system. Based on this, the influence of the bandwidth changes of the current inner loop and voltage feedforward filter on the high-frequency and low-frequency ranges of the grid voltage is analyzed.

Research has found that moderately increasing the bandwidth of the current inner loop or reducing the bandwidth of the

feedforward filter can help enhance the transient voltage stability at the grid connection point. However, increasing the current inner loop bandwidth will deteriorate the low-frequency damping characteristics, while reducing the bandwidth of the feedforward filter still helps to increase low-frequency damping.

2 Dynamic modeling of transient voltage scale

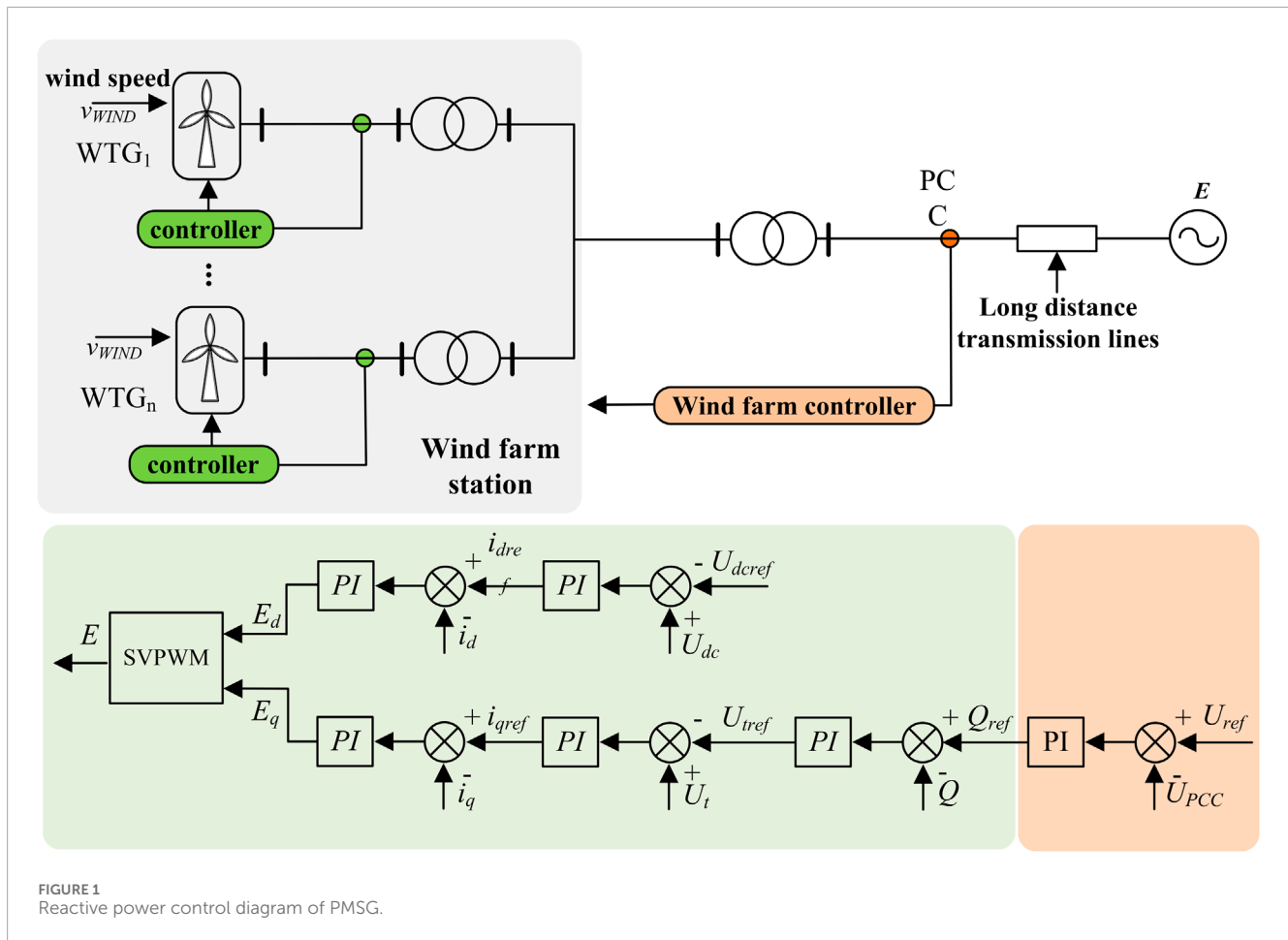
2.1 Multi-scale voltage control architecture

Due to the large number and diverse types of equipment in the power system, a layered multi-scale control structure is traditionally commonly used to meet the multi-time scale frequency and voltage regulation requirements during the operation of the large power grid. This article takes an example of a wind farm composed of direct-drive wind turbine generator. Below is a brief introduction to the voltage control architecture of the existing wind farm station on a time scale from slow to fast. The control structure diagram is shown in Figure 1.

The wind farm controller is responsible for detecting the amplitude of the high-voltage side voltage. For general onshore wind farms, the high-voltage side voltage level is generally around 220 kV. When the amplitude of the high-voltage side voltage deviates from the set value, the wind farm controller will calculate the reactive power demand and send it to the reactive power compensation equipment and wind turbines in the wind farm. Reactive power compensation equipment, such as static var generator (SVG), has a larger rated capacity, generally about three times that of wind turbines. The reactive power compensation equipment converts the reactive power demand output by the wind farm controller into the command value for voltage control on the medium voltage side. The voltage level on the medium voltage side is generally around 35 kV. The wind turbine also receives the reactive power command value issued by the wind farm controller and converts it into the command value of the low-voltage side voltage control, that is, the command value of the AC voltage outer loop. The low-voltage side voltage level is generally 0.69 kV.

The above voltage control at all levels covers multiple voltage levels such as station terminal voltage and wind turbine terminal voltage, covering voltage control issues from electromechanical to medium to long-term time scales on a time scale. These scale voltage controls also exist in traditional power systems. In addition, power electronic converters have a much faster response speed than traditional synchronous generators, thus also possessing faster scale voltage control that traditional power systems do not have, i.e., electromagnetic transient voltage/reactive power control at the current inner loop scale. The inverter controls the transient voltage in the electromagnetic transient time scale by controlling the inner loop current command, such as the q -axis current.

This article focuses on the transient voltage dynamic response problem at the electromagnetic transient scale. The following text will establish a transfer function model for the controller and transmission line at this time scale.



2.2 Dynamic equivalence model of inner loop scale

This article aims to analyze the instantaneous voltage dynamic characteristics of a voltage source converter (VSC) single-machine infinite bus system. Therefore, a single machine-connected system with an LC filter and AC line connected to the grid, as shown in Figure 2, is established to simplify the analysis process. Among them, U_1 and I_1 represent the internal potential and output current of the inverter, U_t represents the grid voltage, and I_C and I_2 represent the current flowing through the filter capacitor and the current flowing through the AC line, respectively.

The filtering capacitor at the top of Figure 2 serves as a low resistance path for high-frequency harmonics, reducing the harmonic component in the output current I_1 of the inverter. The filtering capacitor voltage U_t is used as the grid voltage of the converter, and its dynamic characteristics are the focus of this article.

To analyze the voltage dynamics at the grid connection point, it is necessary to establish a dynamic model of the VSC inverter control system. For the time scale of the current inner loop, the dynamic model needs to take into account the inner loop PI controller, voltage feed-forward (VFF) filter, and other links. Below Figure 2 is a schematic diagram of the control architecture in the time scale of the VSC current inner loop, where the subscript containing abc represents the three-phase AC signal and the subscript containing

dq represents the DC signal in the rotating coordinate system of the controller.

Below are the dynamic equivalent models for the current inner loop, voltage feedforward filter, and AC power grid, respectively. The main idea of modeling is to adopt a certain degree of approximation for the above links, and convert the transfer function of the rotating coordinate system (subscript dq) to the stationary coordinate system (subscript $\alpha\beta$), finally convert it to a three-phase stationary coordinate system (subscript abc).

The current inner loop of the inverter adopts proportional integral control (PI) based on the rotating coordinate system, as shown in Figure 2. Ignoring coupling terms ωL that have a relatively small impact on dynamic characteristics, the input-output and transfer function of the current inner loop can be expressed as follows:

$$\begin{bmatrix} u_{1d}^* \\ u_{1q}^* \end{bmatrix} = \underbrace{\begin{bmatrix} K_p + \frac{K_i}{s} & 0 \\ 0 & K_p + \frac{K_i}{s} \end{bmatrix}}_{G_{dq}(s)} \left\{ \begin{bmatrix} i_{1d}^* \\ i_{1q}^* \end{bmatrix} - \begin{bmatrix} i_{1d} \\ i_{1q} \end{bmatrix} \right\} \quad (1)$$

where the input and output variables have self-illumination in Figure 2, which will not be further introduced here. K_p and K_i represent the proportion and integration coefficients in PI control, s represents the Laplace operator, $G_{dq}(s)$ represents the transfer function matrix.

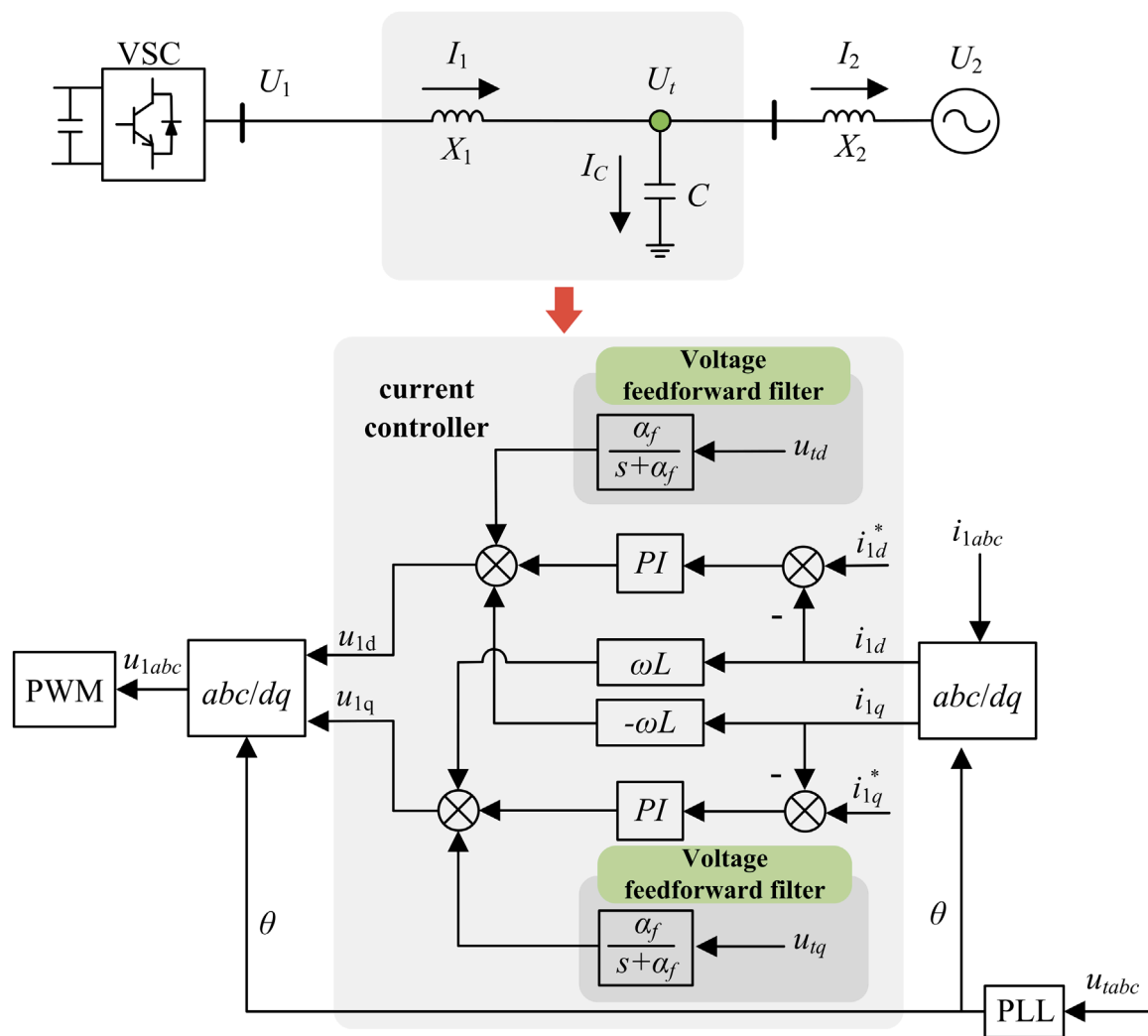


FIGURE 2
VSC grid-connected system.

There is a relationship between the product of time-domain functions and trigonometric functions and their Laplace transform (Zmood et al., 2001):

$$\begin{cases} L(f(t) \cos \omega t) = \frac{1}{2} [F(s + j\omega) + F(s - j\omega)] \\ L(f(t) \sin \omega t) = \frac{j}{2} [F(s + j\omega) - F(s - j\omega)] \end{cases} \quad (2)$$

where $f(t)$ represents the time-domain function, $Lf(t)$ represents the Laplace transform of $f(t)$, and F represents the result after the Laplace transform.

Taking the time-domain voltage signal as an example, there is a transformation relationship between the time-domain voltage signal in the dq coordinate system and the $\alpha\beta$ coordinate system as follows:

$$\begin{bmatrix} u_\alpha(t) \\ u_\beta(t) \end{bmatrix} = \begin{bmatrix} \cos \omega t & -\sin \omega t \\ \sin \omega t & \cos \omega t \end{bmatrix} \begin{bmatrix} u_d(t) \\ u_q(t) \end{bmatrix} \quad (3)$$

After substituting Equation 3 into Equation 2 and applying the same processing to the current signal, the transfer

function matrix $G_{dq}(s)$ in Equation 1 can be used to derive $G_{\alpha\beta}(s)$ in the $\alpha\beta$ coordinate system, which is shown in Equation 4.

$$G_{\alpha\beta}(s) = \begin{bmatrix} K_p + \frac{sK_i}{s^2 + \omega^2} & -\frac{\omega K_i}{s^2 + \omega^2} \\ \frac{\omega K_i}{s^2 + \omega^2} & K_p + \frac{sK_i}{s^2 + \omega^2} \end{bmatrix} \quad (4)$$

where ω represents the fundamental frequency.

The electromagnetic transient simulation algorithm of the power system establishes circuit equations in a three-phase stationary reference frame and uses numerical iteration methods to simulate the dynamic process of the system. Therefore, this paper simulates the electromagnetic transient simulation algorithm and establishes the system dynamic equations in a three-phase stationary reference frame. Then, by defining the transformation matrix T in Equation 5 between the $\alpha\beta$ coordinate frame and the abc coordinate frame, $G_{abc}(s)$ can be further obtained from $G_{\alpha\beta}(s)$, which is shown in

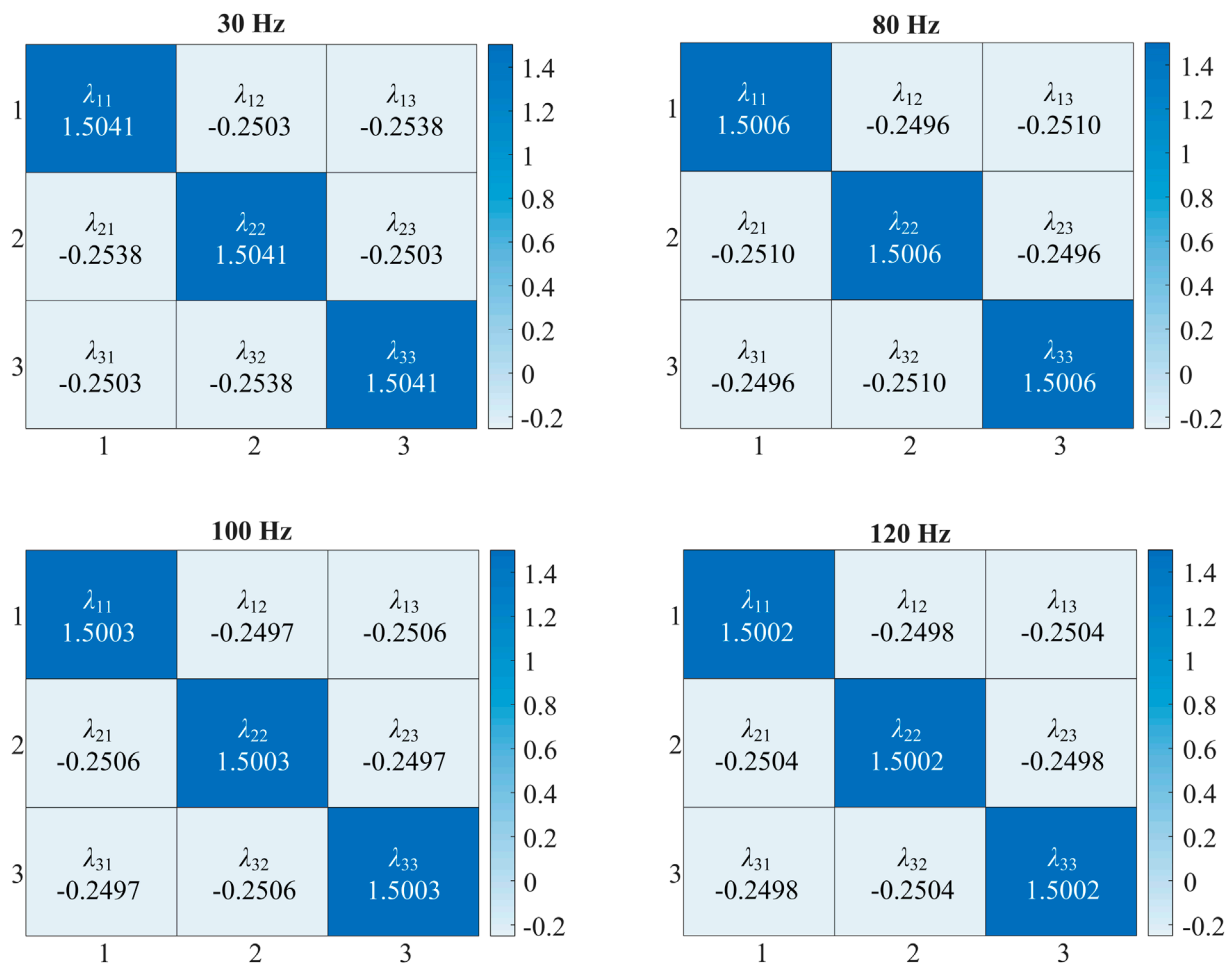


FIGURE 3
Distribution of elements in RGA.

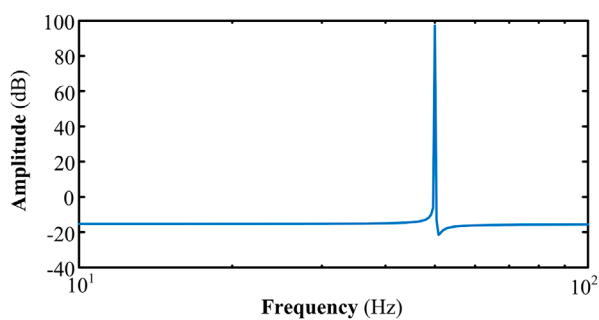


FIGURE 4
Bode diagram of $G_o(s)$.

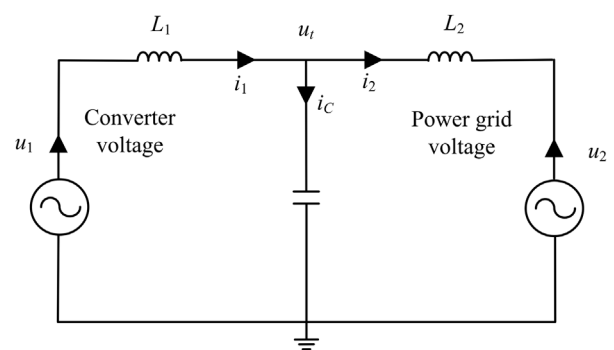


FIGURE 5
Small signal model structure diagram.

Equation 6.

$$T = \begin{bmatrix} \frac{2}{3} & -\frac{1}{3} & \frac{1}{3} \\ 0 & \frac{\sqrt{3}}{3} & -\frac{\sqrt{3}}{3} \end{bmatrix} \quad (5)$$

$$G_{abc}(s) = T^+ G_{\alpha\beta}(s) T \quad (6)$$

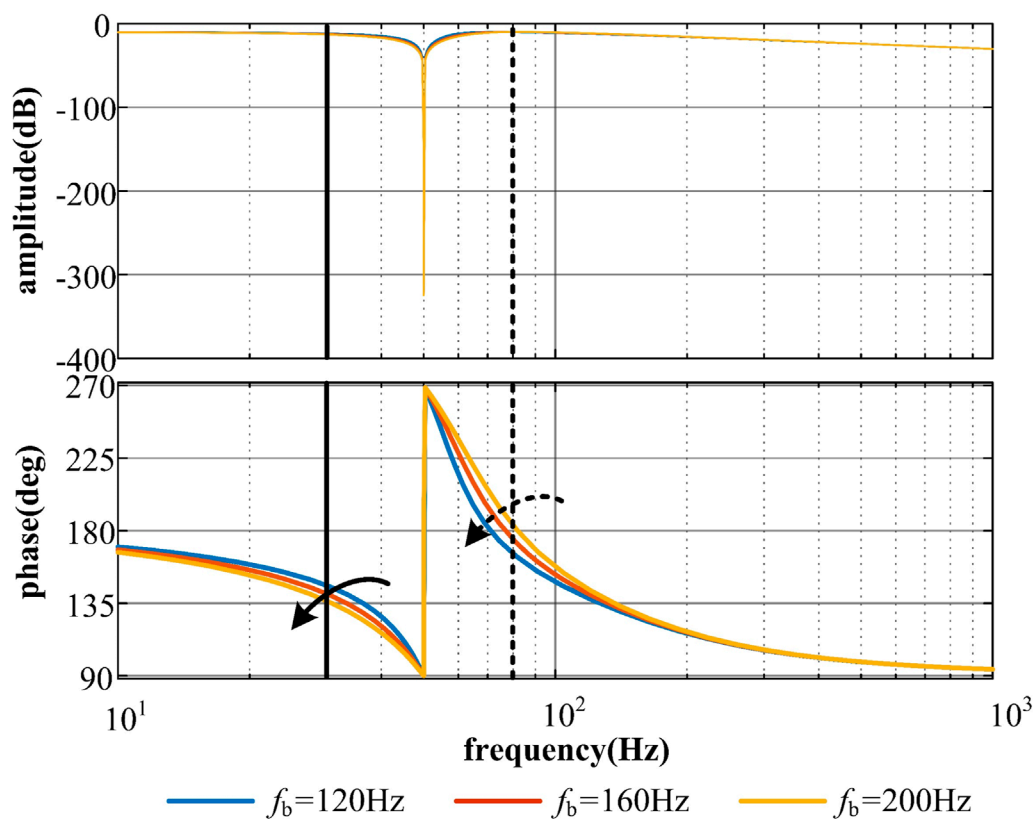


FIGURE 6
Bode plots of $G_{CC}(s)$.

where T^+ represents the pseudo inverse of matrix T . Equation 7 is a commonly used form of T^+ .

$$T^+ = \begin{bmatrix} 1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \quad (7)$$

According to Equation 6, $G_{abc}(s)$ is approximately a diagonally dominant matrix in the mid to high frequency range, with all diagonal elements being $G_c(s)$, as shown in Equation 8.

$$G_c(s) = \frac{1}{3} \left(K_p + \frac{sK_i}{s^2 + \omega^2} \right) \quad (8)$$

The nondiagonal elements are $G_o(s)$, which can be calculated by Equation 9.

$$G_o(s) = \frac{1}{6} \left(\pm \frac{\sqrt{3}\omega K_i}{s^2 + \omega^2} - \frac{sK_i}{s^2 + \omega^2} - K_p \right) \quad (9)$$

In order to simplify the analysis and make $G_{abc}(s)$ approximately decoupled between the three-phase instantaneous current control, this paper adopts the relative gain array (RGA) to analyze the priority of the matrix elements. Bristol proposed RGA (Bristol, 1966), and Grosdidier et al. (1985) further gave relevant proofs of the mathematical properties of RGA. The control theory states that a RGA can be formed for every multiple input multiple

output system. Its basic principle is: for a system with g inputs and k outputs, the relative gain λ_{ij} between the input u_j ($j = 1, 2, \dots, g$) and the output y_i ($i = 1, 2, \dots, k$) can be expressed by Equation 10:

$$\lambda_{ij} = \frac{\left(\frac{\partial y_i}{\partial u_j} \right)_1}{\left(\frac{\partial y_i}{\partial u_j} \right)_2} \quad (10)$$

where, the numerator represents the mode in which all controls are open-loop; the denominator represents the mode in which all controls are open-loop except for the closed-loop control of the u_j - y_i loop.

In this paper, the relative gain matrix can be expressed by Equation 11.

$$R_{RGA} = [\lambda_{ij}] = G_{abc}(s) \otimes (G_{abc}(s)^{-1})^T \quad (11)$$

where, $\otimes$ represents the Hadamard product of the matrix; $G_{abc}(s)$ represents the transfer function matrix at any frequency.

In order to evaluate the correlation of each element in the relative gain matrix, the heat maps of the transfer function $G_{abc}(s)$ at 30 Hz, 80 Hz, 100 Hz, and 120 Hz are plotted in Figure 3. It can be seen that the value of the diagonal element of RGA is much larger than the nondiagonal elements and the system can be approximately decoupled.

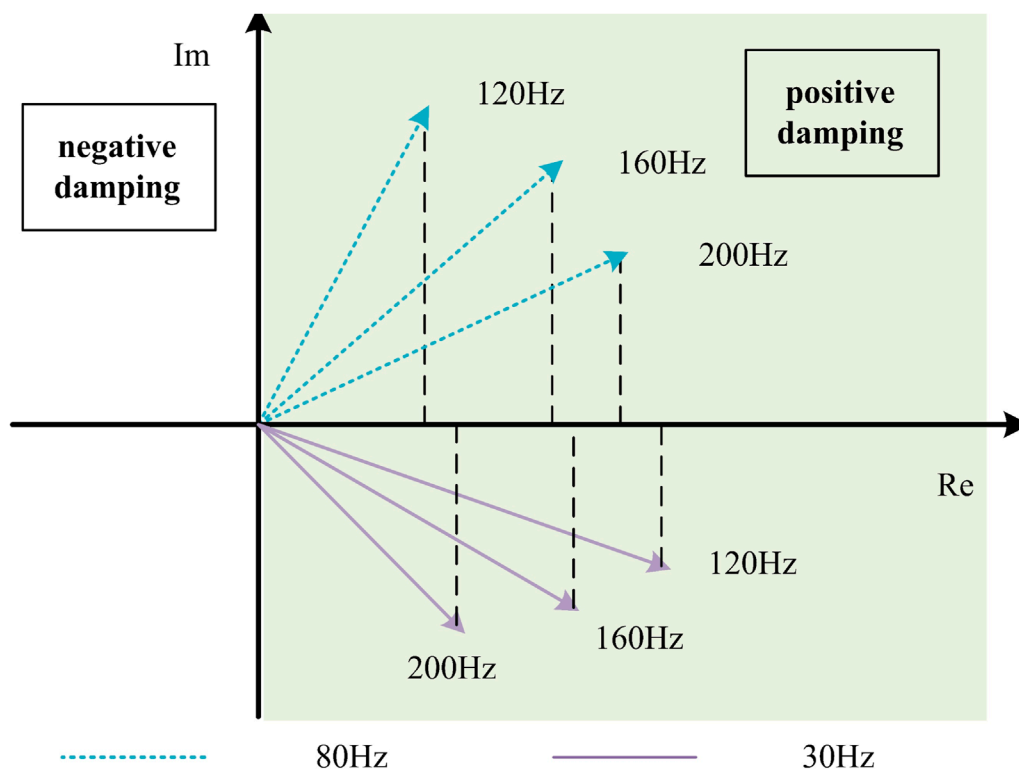


FIGURE 7
Phasor diagram of $G_{CC}(s)$.

In addition, according to the analysis, it can be seen that the influence of nondiagonal elements $G_o(s)$ near the fundamental frequency band is relatively small, as shown in Figure 4, so the $G_o(s)$ will be ignored in the following text (Erika and Holmes, 2003).

The above coordinate transformation aims to approximate decoupling between three-phase instantaneous current control, facilitating the use of a single-phase transfer function $G_c(s)$ to analyze the dynamic damping characteristics of the instantaneous voltage at the grid connection point of the inverter in the following text.

In addition, to simplify the analysis process, a first-order transfer function is used to approximate the sampling/calculation delay of voltage feedforward VFF and pulse width modulation (PWM), respectively (Jia et al., 2019). In the abc coordinate system, the above two first-order links are approximated as the following transfer functions.

$$G_I(s) \approx \frac{\alpha_f}{\alpha_f + s} \quad (12)$$

$$G_d(s) = \frac{1}{1 + T_{PWM}(s + j\omega)} \approx \frac{1}{1 + T_{PWM}s} \quad (13)$$

where $G_I(s)$ represents the VFF transfer function, which was originally constructed in the dq rotating coordinate system and is approximated as a first-order transfer function in the abc coordinate system; The approximation process of $G_d(s)$ and $G_I(s)$

is the same as, where T_{PWM} represents a sampling delay constant of 1.5 times.

According to Equations 8, 12, 13 above, the transfer function of the current inner loop time scale in the abc coordinate system can be obtained by Equation 14.

$$u_{1k} = \left[\frac{1}{3} \left(K_p + \frac{sK_i}{s^2 + \omega^2} \right) (i_k^* - i_k) + \frac{\alpha_f}{\alpha_f + s} u_{tk} \right] \cdot \frac{1}{1 + T_{PWM}s} \quad (k = a, b, c) \quad (14)$$

where the input and output variables also have self-illumination in Figure 2.

The dynamic model of the three phase symmetrical circuit is shown in Equation 15, mainly describing the differential equation between the instantaneous voltage and current on the LC filter and the line inductance.

$$\begin{cases} L_1 \frac{di_{1k}}{dt} = u_{1k} - u_{tk} - i_{1k}R_1 \\ C_1 \frac{du_{tk}}{dt} = i_{Ck} \\ L_2 \frac{di_{2k}}{dt} = u_{tk} - u_{2k} - i_{2k}R_2 \end{cases} \quad (k = a, b, c) \quad (15)$$

At this point, the dynamic equivalent model of the VSC grid-connected system with the current inner loop scale has been established, and the composition structure of the system is shown in the schematic diagram in Figure 5.

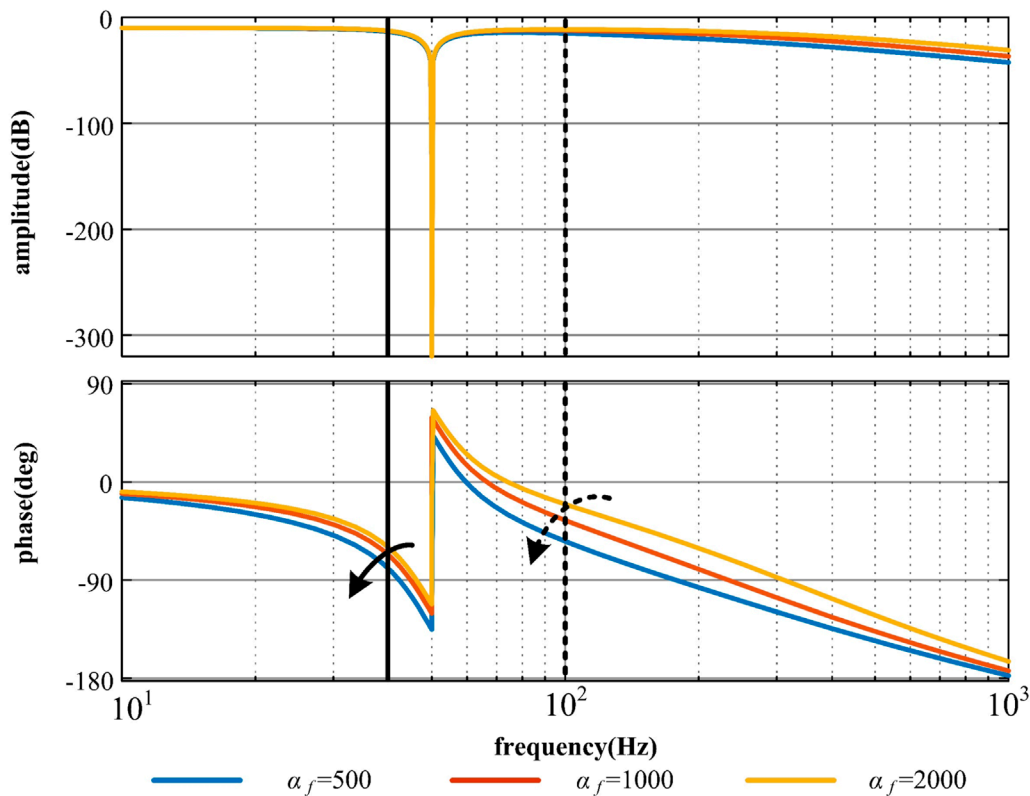


FIGURE 8
Bode plots of $G_{VFF}(s)$.

3 Dynamic damping analysis method for electromagnetic transient voltage

As shown in Figure 4, the transient voltage at the grid connection point of the inverter mainly refers to the dynamic voltage u_t on the filtering capacitor C . Obviously, the voltage on the filtering capacitor C satisfies the following differential equation.

$$C \frac{du_{tk}}{dt} = i_{Ck} = i_{1k} - i_{2k} (k = a, b, c) \quad (16)$$

Equation 16 reflects the dynamic generation mechanism of transient voltage at the grid connection point of the inverter; that is, the voltage on the filtering capacitor C cannot undergo sudden changes, but indirectly generates voltage changes through changes in injected currents on both sides. From another perspective, the output current i_1 of the inverter is generated by the deviation between the output voltage u_1 of the inverter and the voltage u_t of the grid connection point, while the injected current i_2 of the grid is generated by the deviation between the voltage u_t of the grid connection point and the voltage u_2 of the grid connection point.

According to the above analysis, it can be concluded that when the voltage u_t at the grid connection point is disturbed by external factors, the currents i_1 and i_2 on the inverter side and grid side will change. According to Equation 16, the changes in current i_1 and i_2 will also cause a transient response of the voltage u_t at the grid connection point.

The above analysis provides an analytical approach for the dynamic damping of transient voltage at the grid connection point, which involves establishing a small signal model between the voltage disturbance Δu_t at the grid connection point and the current disturbance Δi_1 at the output of the inverter. By observing the frequency domain characteristics of the open-loop system, the dynamic damping distribution of transient voltage at the grid connection point in different frequency bands can be determined.

By linearizing and approximating the dynamic model in Section 2.2, the transfer function between perturbations can be obtained.

$$\begin{cases} \Delta u_1 = -G_c(s)G_d(s)\Delta i_1 + G_t(s)G_d(s)\Delta u_t \\ \Delta i_1 = \frac{1}{sL_1 + R_1}(\Delta u_1 - \Delta u_t) \\ \Delta i_2 = \frac{1}{sL_2 + R_2}(\Delta u_t - \Delta u_2) \\ \Delta i_C = Cs\Delta u_t = \Delta i_1 - \Delta i_2 \end{cases} \quad (17)$$

Equation 17 consists of four equations, with a total of five variables, namely Δi_1 , Δi_2 , Δu_1 , Δu_2 , and Δu_t . When conducting stability analysis, the variation of voltage amplitude in the power grid is generally ignored, i.e., $\Delta u_2 = 0$. So, according to the first two equations in Equation 17, the transfer function between Δi_1 and Δu_t can be solved, which is shown in Equation 18.

$$i_1 = \frac{G_t(s)G_d(s)}{sL_1 + R_1 + G_c(s)G_d(s)}\Delta u_t - \frac{1}{sL_1 + R_1 + G_c(s)G_d(s)}\Delta u_t \quad (18)$$

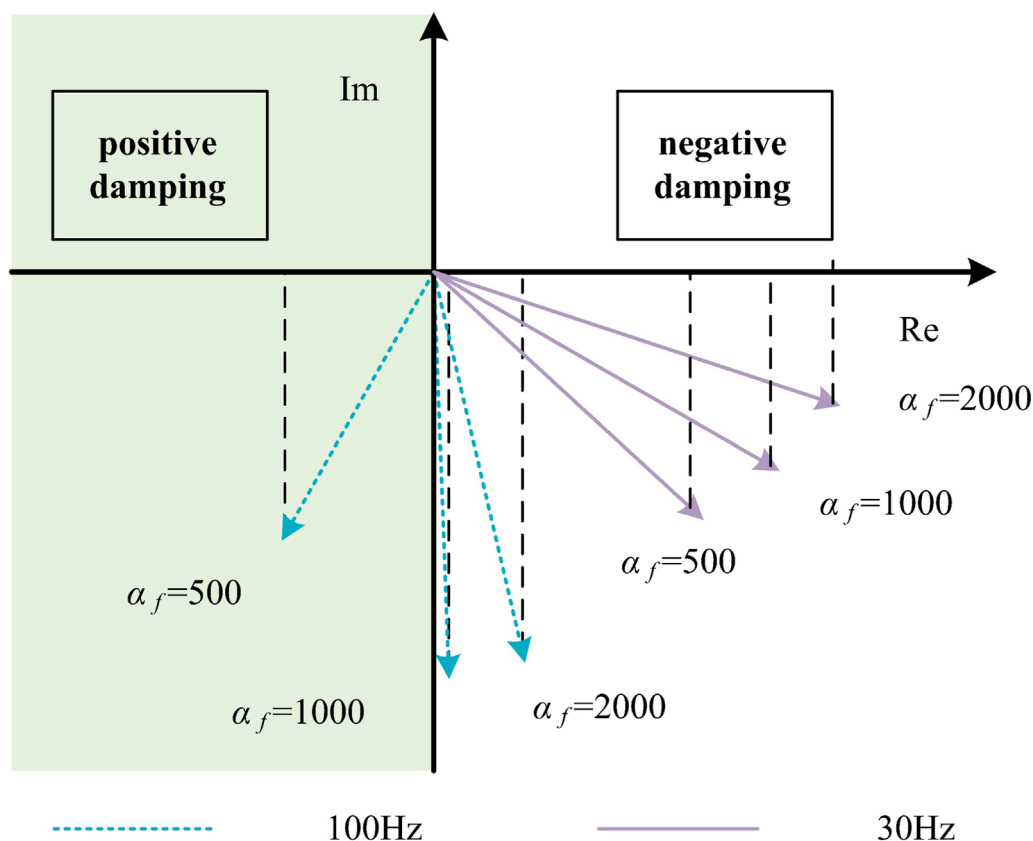


FIGURE 9
Phasor diagram of $G_{VFF}(s)$.

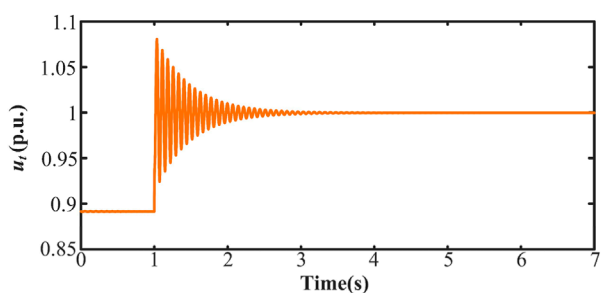


FIGURE 10
Terminal voltage of 120 Hz.

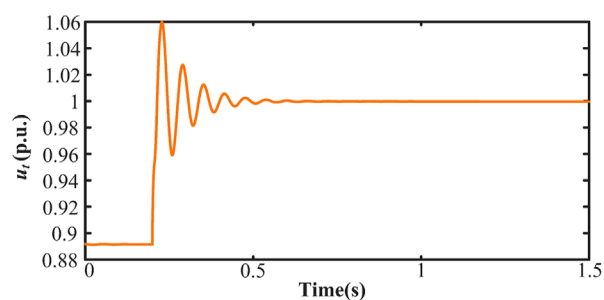


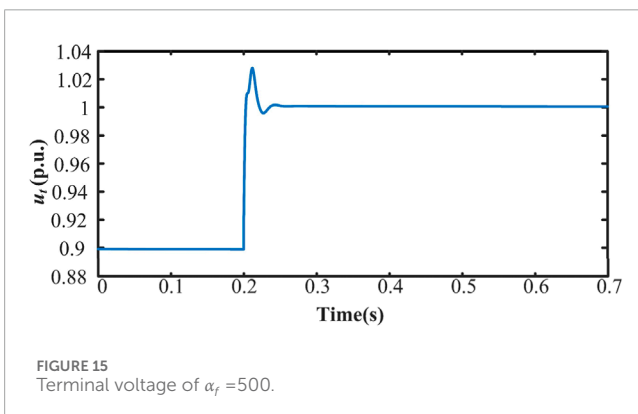
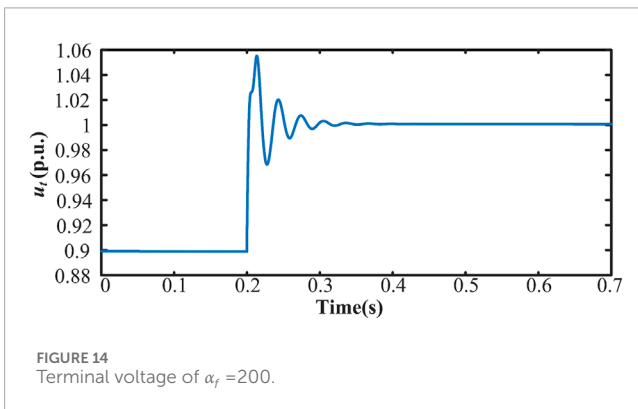
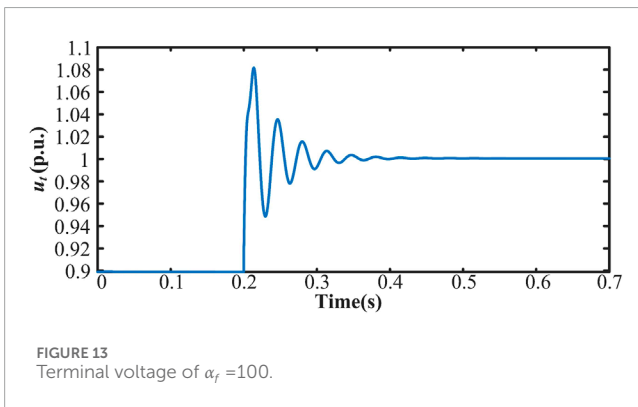
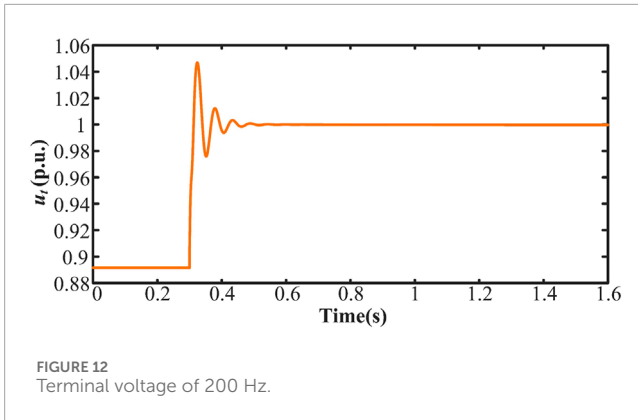
FIGURE 11
Terminal voltage of 160 Hz.

Equations 19, 20 define the transfer functions for analyzing the damping effects of the current inner loop and VFF on transient voltage separately.

$$G_{CC}(s) = \frac{1}{sL_1 + R_1 + G_c(s)G_d(s)} \quad (19)$$

$$G_{VFF}(s) = \frac{G_I(s)G_d(s)}{sL_1 + R_1 + G_c(s)G_d(s)} \quad (20)$$

From the above two definitions, it can be seen that when the influence of VFF is not considered, analyzing the dynamic damping characteristics represented by $G_{CC}(s)$ can analyze the influence of current inner loop parameters on transient voltage. When considering the influence of VFF, analysis $G_{VFF}(s)$ reveals the influence of VFF parameters on transient voltage.



4 Case study

4.1 Analysis of the influence of controller bandwidth

Here, we analyze the Bode diagram and vector diagram of the above transfer function $G_{CC}(s)$ and $G_{VFF}(s)$ at the current inner loop time scale, respectively, to reflect the influence of controller bandwidth on the dynamic damping of transient voltage at the grid connection point.

Firstly, analyze the dynamic damping impact of the current inner loop bandwidth on the transient voltage at the grid connection point. Figures 6, 7 show the Bode plot of the transfer function and its corresponding vector plots at 30 Hz and 80 Hz frequency points, respectively. Where f_b represents the bandwidth of the current inner loop, and in the case analysis, three types of inner loop control are used: 120 Hz, 160 Hz, and 200 Hz. Here, it can be considered that the time scale of the current inner loop corresponds to the region between 80 and 200 Hz. It can be seen from the dashed line in Figure 7 that as the inner loop bandwidth increases, $G_{CC}(s)$ provides greater positive damping at 80 Hz, which can effectively reduce the output current disturbance of the inverter caused by voltage disturbance at the grid connection point. However, it is also evident from the solid line in Figure 7 that as the inner loop bandwidth increases, $G_{CC}(s)$ provides smaller positive damping at 30 Hz, leading to a deterioration of low-frequency characteristics.

Next, analyze the dynamic damping of voltage feedforward filter parameters on transient voltage at grid points. Take α_f as 500, 1000, and 2000, respectively, and draw Bode plots of $G_{VFF}(s)$ and their corresponding vector plots at 30 Hz and 100 Hz frequency points, as shown in Figures 8, 9.

From the results in Figures 8, 9, it can be seen that increasing α_f can also increase negative damping in both the frequency range (100 Hz) corresponding to the current inner loop scale and the low-frequency range (30 Hz). Increasing α_f corresponds to reducing the bandwidth of the voltage feedforward filter.

4.2 Simulation verification

To verify the effectiveness of the above analysis, this paper built a VSC grid-connected system as shown in Figure 2 on the MATLAB/Simulink platform to observe the changes in PCC point voltage u_t under different current inner loop bandwidth and voltage feedforward filter parameter settings.

When the inner loop bandwidth is 120 Hz, a disturbance of 0.1 p.u. is added to the voltage at the grid connection point at 1 s. The voltage stabilizes after 3 s, as shown in Figure 10. When the inner loop bandwidth is 160 Hz, a disturbance of 0.1 p.u. is added to the voltage at the grid connection point at 0.2 s. The voltage fluctuation lasts for 0.55 s, as shown in Figure 11. As indicated in Figure 12, when the inner loop bandwidth is increased to 200 Hz, the voltage fluctuation lasts for 0.5 s under the same disturbance.

The above simulation proves that increasing the bandwidth of the current inner loop will increase the positive damping, which can effectively reduce the fluctuation duration caused by voltage disturbances.

Moreover, the influence of VFF parameters is analyzed below. When the bandwidth of the current inner loop is 200 Hz, the values of α_f are set to 100, 200, and 500, respectively. At $t = 0.2$ s, a 0.1 p.u. disturbance is introduced to the voltage of the PCC point, causing voltage fluctuation durations of 0.25 s, 0.17 s, and 0.07 s, as shown in Figures 13–15.

From the above results, it can be seen that as the parameters of the voltage feedforward filter increase, it can provide greater positive damping, and the time required for the grid voltage to recover to a steady state after being disturbed is shorter. Therefore, increasing the parameters of the voltage feedforward filter can effectively reduce the grid voltage disturbance.

5 Conclusion

This paper analyzes the impacts of the current-loop control parameters on the electromagnetic transient voltage performance of the voltage-source converter. The conclusion is that moderately increasing the current inner loop bandwidth or reducing the feedforward filter bandwidth helps to enhance the damping in the high-frequency range of the voltage dynamic. However, increasing the current inner loop bandwidth will deteriorate the low-frequency damping characteristics, while reducing the bandwidth of the feedforward filter still helps to increase low-frequency damping.

Data availability statement

The raw data supporting the conclusions of this article will be made available by the authors, without undue reservation.

Author contributions

XD: Conceptualization, Methodology, Validation, Writing–original draft. HC: Methodology, Investigation, Software,

Writing–review and editing. XZ: Conceptualization, Data curation, Formal Analysis, Supervision, Writing–original draft. QS: Supervision, Writing–review and editing. GW: Writing–review and editing, Conceptualization, Formal Analysis, Methodology, Project administration.

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Conflict of interest

Authors XD, HC, and XZ were employed by China Energy Engineering Group Zhejiang Electric Power Design Institute Co., Ltd.

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Supplementary material

The Supplementary Material for this article can be found online at: <https://www.frontiersin.org/articles/10.3389/fenrg.2024.1496400/full#supplementary-material>

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Research on the multi-objective collaboration damping control strategy based on multi-resource

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To improve the stability margin of the power grid under unconventional conditions and enhance the fault emergency support capability of AC/DC grids, this paper investigates multi-resource and multi-objective collaboration damping control strategy method based on a multi-input multi-output (MIMO) system model, which considers the multiple control impacts. Based on the analysis of system oscillation characteristics, appropriate feedback signals and controller installation locations are selected based on the modal ratio index. To realize the control effect, the HVDC and energy storage devices are used. Finally, the coupled robust controllers based on robust mechanism are designed out. Simulation results in PSCAD software show that, compared to traditional damping controllers, the controller designed is more suitable for system conditions with large perturbations and uncertainties.

KEYWORDS

damping control, multi-objective, HVDC, energy storage, control strategy

1 Introduction

Actively and steadily promoting carbon peaking and carbon neutrality (“dual carbon”) goals, and accelerating the construction of a safe, stable, economically efficient, coordinated supply and demand, flexible, and intelligent new power system to ensure energy supply and energy security is the mission of China’s energy and power industry development (Sheta et al., 2023). As the project with the largest investment and engineering volume in China’s western development strategy, the implementation of the “West-to-East Electricity Transmission” project has effectively alleviated environmental pollution and energy shortages in China’s southeast coastal areas. It is the right choice to achieve optimal allocation and efficient utilization of energy resources in China (Jiang et al., 2022).

Ultra-high voltage AC transmission technology has advantages such as strong networking capabilities, long transmission distances, and flexible landing points. It can connect adjacent provincial power grids, achieving time and seasonal complementarity, mutual support among wind, photovoltaic, hydro, and thermal power generation, as well as cross-regional and cross-basin compensation and adjustment (Li Q. et al., 2024). Ultra-high voltage DC transmission technology, with its advantages of low transmission loss, large transmission capacity, point-to-point transmission, strong controllability, and no stability issues (Tao et al., 2020), has increasingly become the mainstay of the “West-to-East Electricity Transmission” project.

Taking the East China Power Grid as an example, it has developed into a typical AC-DC hybrid power grid structure with multiple UHV DC feed-ins and UHV AC interconnections, as shown in Figure 11. As of April 2024, there are 13 UHV DC

transmission projects feeding into the East China Power Grid, with a total rated power of 78.8 million kilowatts, and the DC landing points are very dense. The UHV AC main grid structure in East China is based on the “Wandian Dongsong” project, forming a UHV AC loop network. The East China UHV AC-DC hybrid power grid has played a significant role in ensuring national energy security, promoting the achievement of the “dual carbon” goals, and facilitating the integration of renewable energy (Jiang et al., 2023a).

However, while the large-scale access of UHV AC and DC achieves energy mutual assistance and adequate power supply, it also brings many challenges to the operation and control of the large power grid. The flexible controllability of High Voltage Direct Current (HVDC) transmission has fundamentally changed the dynamic characteristics of the AC-DC hybrid power grid compared to traditional power systems (Wen et al., 2023).

The main manifestations are the continuous increase in power from outside the region and the continuous replacement of local conventional power sources by the development of renewable energy. This has led to a decreasing proportion of traditional power generation units being operational, resulting in characteristics such as low system inertia and a high degree of “hollowness” (Verma et al., 2023). Consequently, this brings about the issue of reduced oscillation damping ratios in the near-field of DC landing points, further increasing the risk of low-frequency oscillations in the power grid. As there are often multiple oscillation modes among units, the existing damping controller design methods are mostly in the form of single-input single-output (SISO) (Abd El-Kareem et al., 2021). Single-control-loop damping controllers do not reflect the interactive effects between multiple control loops, which can weaken or even worsen the effect of closed-loop control in complex power systems or when there are long time delays in wide-area signals.

In response to the aforementioned issues of low-frequency oscillation analysis and suppression, scholars at home and abroad have conducted extensive and in-depth research at various levels, including “source-grid-load-storage”. At the power source level, installing a power system stabilizer (PSS) (Zou et al., 2024), but PSS will be invalid if interzone modes occurs. At the load level, power output can be adjusted by regulating the load, thereby achieving system power balance. At the energy storage level, by coordinating with renewable energy devices, the power injected into the system can be changed to compensate for the randomness of renewable energy output. At the grid level, additional damping control based on renewable energy converters, HVDC, and FACTS has been a research hotspot. Additional damping control has a faster response speed and better suppression effect, and it is still applicable to sub/super-synchronous oscillations in power systems with large-scale renewable energy access. Its principle is to extract signals related to damping characteristics from the AC system as feedback signals, use them as inputs to the damping controller, and add the signals output by the transfer function to a certain control channel of the converter equipment, which can enhance the system damping (Wilches-Bernal et al., 2019). In 2007, China Southern Power Grid developed the world’s first multi-loop DC adaptive coordinated control system based on

wide-area information. This system utilizes the Wide-Area Measurement System (WAMS) of China Southern Power Grid, together with the GaoZhao and XingAn DC systems, to build a closed-loop control system, which can effectively suppress oscillations between Yunnan-Guangdong and Yunnan-Guizhou regions (Ming et al., 2011).

Inspired by traditional control methods, more advanced and mature analysis methods for additional damping control are used. Using modal analysis, literature (Ma et al., 2021) analyzes the main factors affecting the stability of a flexible DC system (VSC-HVDC) with DC circuit breakers. A robust damping controller based on Hoo mixed sensitivity theory is designed to address the stability issues caused by the circuit breaker system in the DC system. Focusing on wind power grid-connected systems, literature (Yantong et al., 2020; Jing et al., 2020) designs robust controllers based on static synchronous compensators and grid-side converters to suppress oscillations caused by wind turbine grid connection. Considering the time delay effects of WAMS, literature (Weng et al., 2013; Ma et al., 2017) designs a wide-area robust damping controller and verifies its robustness to transmission delays and external disturbances in a four-machine two-area test system. To address the interaction between the generator-side oscillation mode and the grid-side additional damping controller, literature (Hengfeng et al., 2016) selects controller parameters based on a fuzzy control strategy to achieve coordination between the generator and the grid, maximizing system damping under interval oscillations. This approach has been validated in the XiMeng UHV transmission system. Literature (Tang et al., 2015) applies heuristic dynamic programming algorithms to the coordinated control of static synchronous compensators (STATCOM) and doubly-fed induction generators, achieving improved damping effects. Literature (Bento, 2021) uses particle swarm optimization to optimize the design and parameter selection of wide-area damping controllers, ensuring their robustness through automatic optimization.

But the above analysis did not consider the unified suppress methods to decrease the multi-modes low frequency oscillations. As the multi-modes usually increase the control difficulty, it is better to design the control strategy in a unified form. In which form, not only the control effect can be enhanced, but also can be inner interact between different controllers can be simultaneously considered. Aiming to design the low frequency oscillation controller in a unified form, this paper utilizes the HVDC and energy storage to suppress multi-modes LFOs, namely the additional MIMO controllers are designed based on HVDC and energy storage control schemes.

The novelties of the paper are as follows:

- 1) To consider the inner interact between different controllers, the MIMO controllers are used for better control effect, which is realized based on HVDC and energy storage control devices.
- 2) During the controller design procedure, the robust control theory is used, which can not only enhance the control effect, the robustness of the controllers can also be guaranteed.
- 3) The controllers of HVDC and energy storage are simultaneously considered, which can control different control modes at the same time, and the multi-resource of the power system can be used thoroughly too.

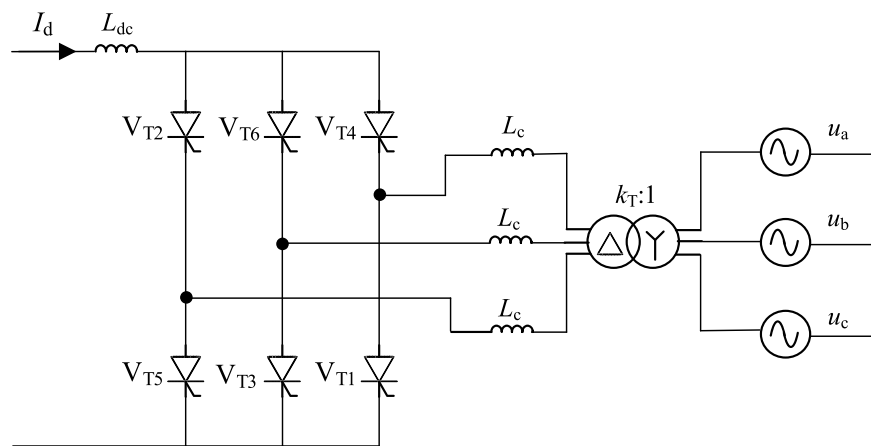


FIGURE 1
Schematic diagram of three-phase full-wave bridge circuit structure.

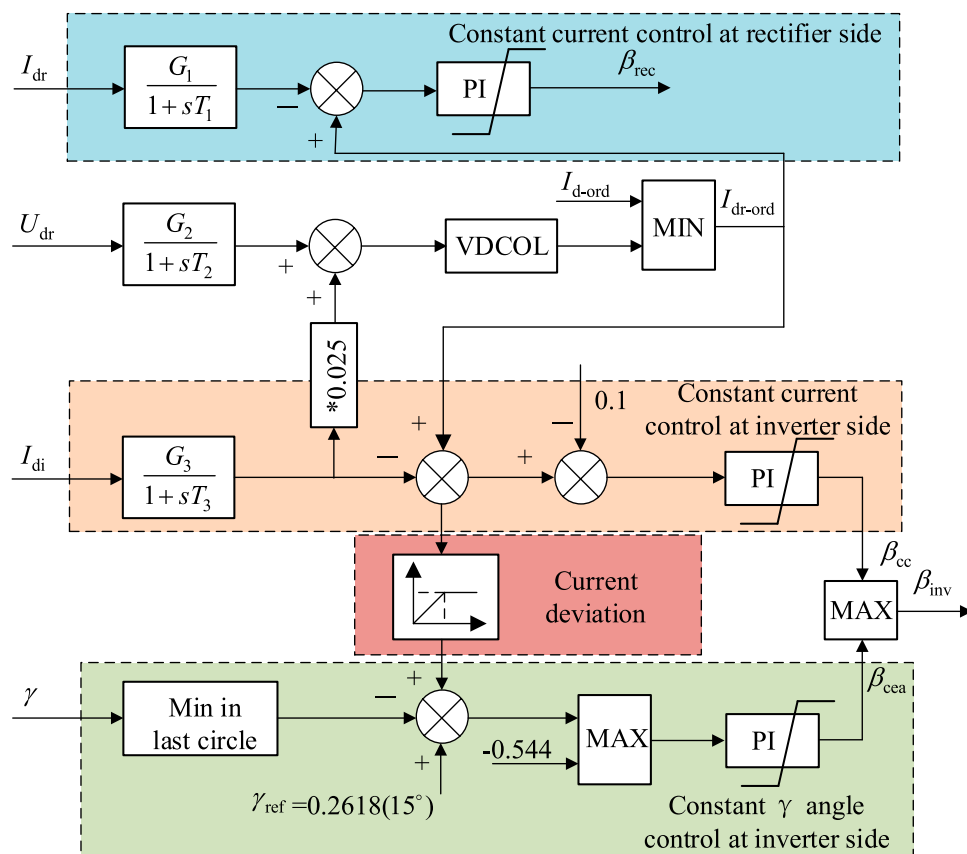
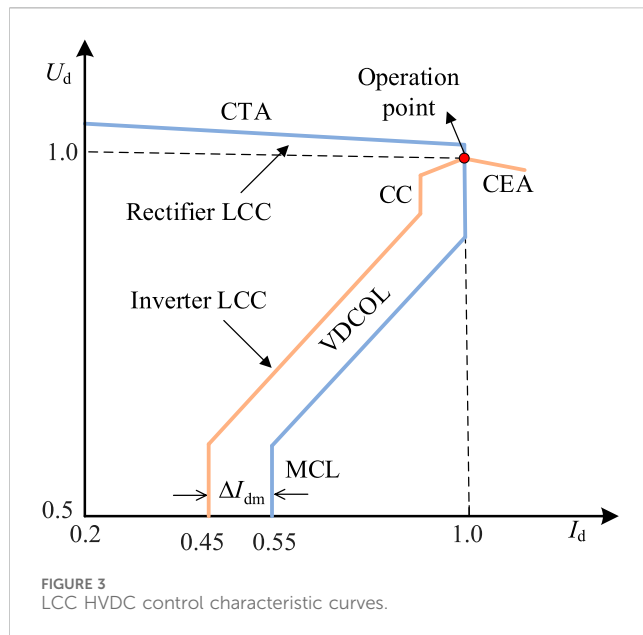


FIGURE 2
LCC HVDC control diagram.

The paper is structured as follows: Section 2 outlines the fundamental frameworks of the LCC-HVDC and the energy storage, the primary control of the two devices is also introduced in this section. Section 3 introduces the MIMO

controllers design method based on robust theory. Section 4 presents a case study backed by simulation results, validating the efficacy of our proposed strategy. Finally, Section 5 offers concluding remarks on this paper.



2 The basic structures of HVDC and energy-storage system

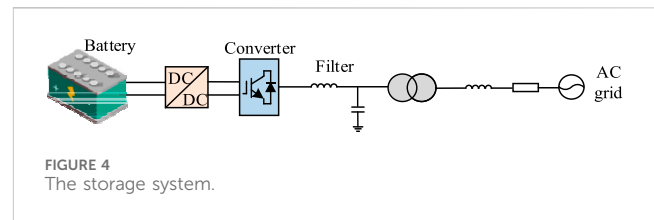
2.1 The basic structure of LCC-HVDC

Figure 1 is a schematic diagram of the three-phase full-wave bridge circuit structure (Jiang et al., 2023b), which is also the structure adopted by LCC in conventional DC systems. In this diagram, u_a , u_b , and u_c represent the three-phase AC phase voltages of the system, respectively. L_c represents the system's equivalent commutation inductance, L_{dc} represents the smoothing inductance, I_d represents the DC current, k_T represents the transformation ratio of the converter transformer, and VT_i ($i = 1-6$) represents the thyristor converter valve on the corresponding arm of the LCC.

During normal operation of the converter, one converter valve is conducting in both the upper and lower arms of the converter's commutation valves. The two conducting valves have adjacent numbers (considering VT_6 adjacent to VT_1). When the LCC undergoes a state transition, the already conducting valve in the upper or lower arm will be turned off, and the valve that was previously off will be turned on in sequence. The alternating conduction process of the valves in the upper/lower arms is known as the commutation process.

The control strategy of the LCC converter station consists of two parts: the control strategy for the rectifier side LCC and the control strategy for the inverter side LCC. The control structure is shown in Figure 2.

In Figure 2, I_{dr} and I_{di} represent the DC currents on the rectifier and inverter sides, respectively. U_{dr} and U_{di} represent the DC voltages on the rectifier and inverter sides, respectively. β_{rec} and β_{inv} represent the lead trigger angle command values for the rectifier and inverter sides, respectively. I_{d_ord} is the settable current command value, I_{dr_ord} is the current command value on the rectifier side, and γ_{ref} is the rated arc extinction angle. γ is



the system's arc extinction angle, which takes the minimum value from the previous cycle. β_{cc} and β_{cea} represent the lead trigger angle command values corresponding to constant current control and constant arc extinction angle control on the inverter side, respectively. The control mode corresponding to the larger of the two values determines the current control mode of the converter.

The CE control strategy employed in LCC converter stations falls between the CEA and CC control modes on the inverter side. Additionally, the VDCOL control mode is implemented on both the rectifier and inverter sides. Its purpose is to limit the DC current when the DC voltage is too low, thereby reducing the transmitted DC power. This aids in the rapid recovery of the DC system after a fault and prevents commutation failures. The control characteristic curves for the rectifier and inverter side LCCs are shown in Figure 3. Under normal system operation, the rectifier side LCC operates in constant current control mode, while the inverter side LCC operates in constant extinction angle control mode.

2.2 The basic structure of energy-storage system

The grid-connected structure of the energy storage system is shown in Figure 4. Due to the intermittency and volatility of solar energy, the output of the photovoltaic system fluctuates accordingly. To ensure a stable power supply, the energy storage system assists the photovoltaic system as the main control power source for the black start. By storing and releasing energy, it smooths out the photovoltaic output and ensures that the system provides continuous and stable power to the load.

Figure 5 represents the typical control structure of an ES system. In the diagram, i_B denotes the DC current output by the energy storage system; U_{dc} represents the DC voltage output by the DC/DC converter; f_{ref} is the system reference frequency, which goes through a phase-locked loop to obtain the phase angle θ ; u_{abc} stands for the measured value of the grid-side voltage; i_{l_abc} signifies the filter inductor current; u_{dq_ref} is the given outer-loop voltage reference signal; i_{dq_ref} represents the inner-loop current reference signal generated by the outer-loop control; and u_{dq} and i_{dq} are the voltage and current in the dq axis after coordinate transformation.

In the typical control structure of the energy storage system, the bidirectional DC/DC converter maintains a constant DC voltage by controlling the DC current output from the energy storage battery, enabling bidirectional energy transfer between the energy storage system and the grid. The PCS of the energy storage system employs

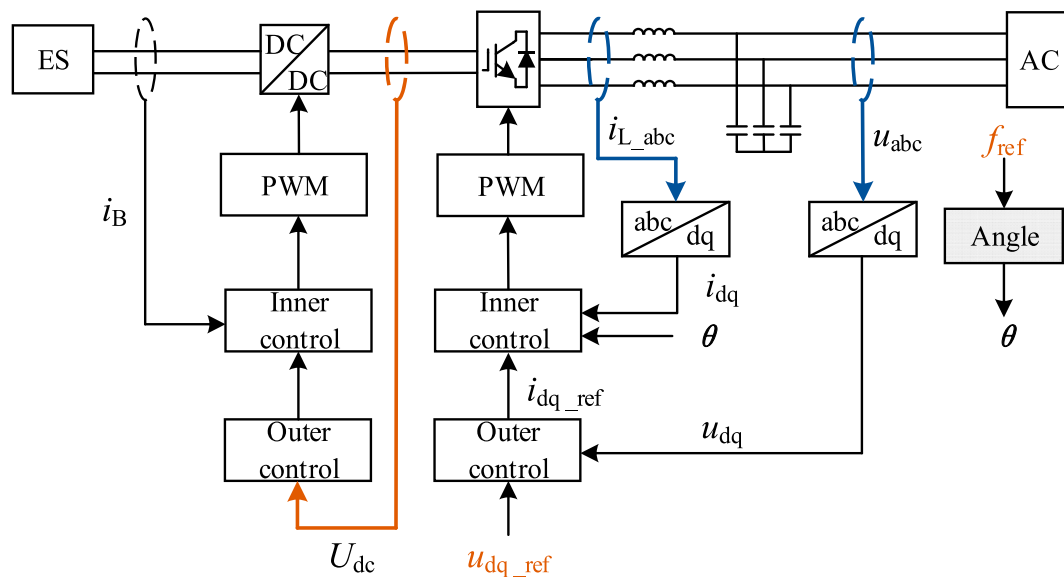


FIGURE 5
The control system of energy storage (ES) system.

V/f control to provide stable voltage and frequency support for grid restoration. Additionally, the energy storage system balances issues such as mismatches between system output and load power, offering stable and reliable electrical support to the grid.

3 The coordinated low frequency oscillation damping control strategy based on robust theory

In this section, by improving the existing robust damping control methods and considering the interaction between multiple control loops, a coupled robust damping control method based on a multi-input multi-output (MIMO) system model is proposed. Based on the analysis of system oscillation characteristics, appropriate feedback signals and controller installation locations are selected using the modal ratio index. The Total Least Squares-Estimation of Signal Parameters via Rotational Invariance Techniques (TLS-ESPRIT) algorithm is utilized to identify a reduced-order system model. Finally, a coupled robust controller is designed.

3.1 The improved robust controllers

For a MIMO system, when designing a robust damping controller using mixed H₂/H_∞ control theory, the system structure is shown in Figure 6 and Equation 1. In the figure: $G(s)$ represents the controlled system; $K(s)$ represents the control system; ε_i ($i = 1, 2, \dots, n$) represents the controller output; r represents the reference input; u represents the system input; y represents the system output; $W_1(s)$, $W_2(s)$, and $W_3(s)$ are weighting functions; $z_{\infty 1}$, $z_{\infty 2}$, and z_2 are reference outputs for measuring system performance; w_i ($i = 1, 2, \dots, n$) represents

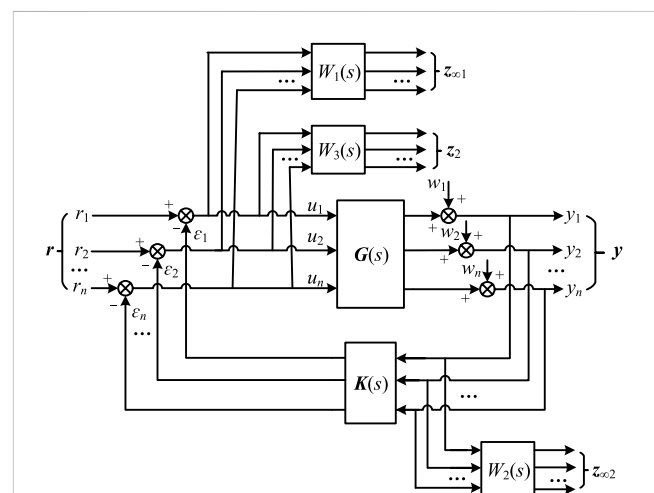


FIGURE 6
MIMO control system based on robust theory.

external disturbances to the system. The mode of Figure 6 is as follows (Apkarian et al., 2001):

$$\begin{cases} \dot{x} = Ax + B_1 w + B_2 u \\ z_{\infty} = \begin{bmatrix} z_{\infty 1} \\ z_{\infty 2} \end{bmatrix} = C_{\infty} x + D_{\infty 1} w + D_{\infty 2} u \\ z_2 = C_2 x + D_{21} w + D_{22} u \\ y = Cx + D_{y1} w + D_{y2} u \end{cases} \quad (1)$$

In the formula: x represents the system state variables; A represents the state matrix; B_1 represents the disturbance gain matrix; B_2 represents the control input matrix; C_{∞} , $D_{\infty 1}$, and $D_{\infty 2}$ are coefficient matrices for state variables, disturbance variables, and input variables related to the H_{∞} index, respectively; C_2 , D_{21} , and

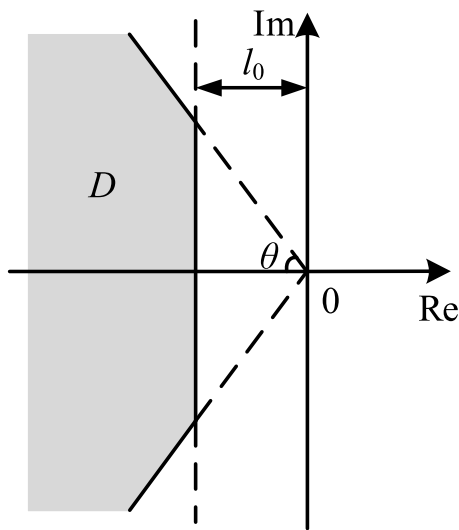


FIGURE 7
Pole configuration region.

D_{22} are coefficient matrices for state variables, disturbance variables, and input variables related to the H_2 index, respectively; D_{y1} and D_{y2} are direct transmission matrices corresponding to w and u , respectively; C represents the system output matrix.

The power grid itself has a complex structure, which inherently leads to uncertainties in modeling and disturbances. But if the poles of the system in Figure 6 are constrained in the shadow area plotted in Figure 7, the small stability can be guaranteed. In the figure, l_0 represents the minimum distance from the pole configuration region to the imaginary axis, and θ represents the maximum argument of the pole configuration region. This can be described using matrix inequalities as follows Equations 2, 3:

$$D = \{s \in \mathbb{C} \mid L + sM + s^*M^T < 0\} \quad (2)$$

$$\begin{cases} L = \begin{bmatrix} 2l_0 & 0 \\ 0 & 0 \end{bmatrix} \\ M = \begin{bmatrix} \sin \theta & -\cos \theta \\ \cos \theta & \sin \theta \end{bmatrix} \end{cases} \quad (3)$$

The above problem that make sure the poles in certain zones can be realized by the LMI toolbox, which can be found in the mathematics math software, namely the MATLAB.

The H_∞ performance index reflects the system's effectiveness in suppressing disturbances, which can use the following equation to calculate:

$$\|T_\infty(s)\|_\infty = p_k\{\sigma_{\max}(T_\infty(j\omega))\} \quad (4)$$

From Equation 4, it can be seen that it represents the peak value $p_k\{\cdot\}$ of the maximum singular value $\sigma_{\max}(\cdot)$ of the system's frequency response. From a time-domain perspective, it represents the steady-state gain in the worst-case scenario for sinusoidal inputs of any frequency. According to the bounded real lemma, the necessary and sufficient condition to satisfy the controller design objectives is the existence of a symmetric

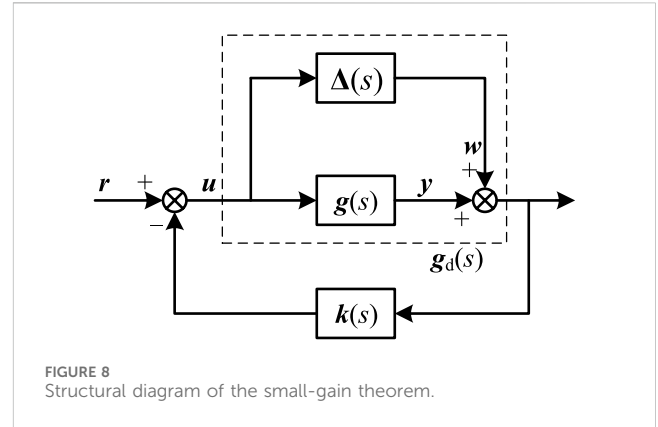


FIGURE 8
Structural diagram of the small-gain theorem.

positive definite matrix P_∞ , such that the following Equation 5 holds:

$$\begin{pmatrix} AP_\infty + P_\infty A^T & B_1 & P_\infty C^T \\ B_1^T & -\gamma_\infty I & D_{\infty 1}^T \\ CP_\infty^T & D_{\infty 1} & -\gamma_\infty E \end{pmatrix} < 0 \quad (5)$$

In the formula, E represents the identity matrix.

The H_2 performance can be calculated by Equation 6, which is as follow.

$$\|T_2(s)\|_2 = \sqrt{\frac{1}{2\pi} \int_{-\infty}^{\infty} \text{tr}(T_2^*(j\omega)T_2(j\omega))d\omega} \quad (6)$$

In the formula, $\text{tr}(\cdot)$ represents the trace of the matrix, and $[T_2^*(j\omega)]$ is the conjugate transpose of $T_2(j\omega)$. From Equation 6, it can be seen that the H_2 norm of $T_2(s)$ is equal to the root mean square value of the output energy of the system's impulse response. The necessary and sufficient condition to satisfy the controller design objectives is the existence of symmetric positive definite matrices P_2 and Q , such that Equations 7–9 hold.

$$\begin{pmatrix} A^T P_2 + P_2 A & P_2 B_1 \\ B_1^T P_2 & -E \end{pmatrix} < 0 \quad (7)$$

$$Q > CP_2 C^T \quad (8)$$

$$\|T_2(s)\|_2 < \sqrt{\lambda_{\max}(Q)} \quad (9)$$

The mixed H_2/H_∞ problem then can be described as the following Equation 10 optimization problem:

$$J = \min\{\alpha\|T_\infty(s)\|_\infty + \beta\|T_2(s)\|_2\} \quad (10)$$

where, $\alpha + \beta = 1$.

3.2 Small-gain theorem for robust stability of systems considering uncertainty

Considering the constantly changing operating conditions and working states of the actual power system, it will cause changes in the system's small perturbation offline identification model. Additionally, there is an unavoidable truncation error between the system's true transfer function model and the reduced-order transfer function model obtained through small perturbation identification methods. As shown in Figure 8, let $g(s)$ represent

the reduced-order transfer function matrix obtained using the identification method, $\mathbf{g}_d(s)$ represent the actual transfer function matrix of the system, and $\Delta(s)$ represent the identification error. Then we have Equation 11:

$$\Delta(s) = \mathbf{g}_d(s) - \mathbf{g}(s) \quad (11)$$

The small-gain theorem can be described as follows: when operating conditions change, the necessary and sufficient condition for the system shown in Figure 8 to remain stable is that Equation 12 holds true:

$$\|\mathbf{G}_{uw}(s)\|_{\infty} = \left\| \frac{\mathbf{u}}{\mathbf{w}} \right\|_{\infty} = \left\| -\mathbf{k}(s)(1 + \mathbf{k}(s)\mathbf{g}(s))^{-1} \right\|_{\infty} < \left\| \frac{1}{\Delta(s)} \right\|_{\infty} \quad (12)$$

In the formula, $\mathbf{G}_{uw}(s)$ represents the closed-loop transfer function from $\mathbf{w}$ to $\mathbf{u}$. The smaller the value of $\|\mathbf{G}_{uw}(s)\|_{\infty}$, the stronger the robustness of the controller to changes in operating conditions.

3.3 Coupled robust control for MIMO systems

In large-scale AC/DC interconnected power systems, multiple oscillation modes often coexist, requiring the use of multiple HVDC or FACTS auxiliary control functions to enhance damping for multiple oscillation modes. A common approach involves pairing multiple modulation signals with control loops, designing damping controllers using a specific damping channel to enhance damping for a particular oscillation mode. For the MIMO multivariable decentralized control system shown in Figure 6, if $u_i \rightarrow y_i$ ($i = 1, 2, \dots, n$) is treated as an independent SISO system, due to the inherent coupling in the electrical system, the decentralized robust damping controller designed accordingly does not reflect the interaction between multiple variables and may even worsen the damping effect of the controller. However, in complex control loop situations, it is sometimes difficult to eliminate the coupling that exists in the loop using controllers designed based on decoupling methods. Therefore, this chapter considers the interaction between multiple control loops in the MIMO system and proposes a coupled robust damping control method.

For a control loop consisting of any input variable u_j ($j = 1, 2, \dots, n$) and any output variable y_i ($i = 1, 2, \dots, n$), when all other loops are open, the transfer function from u_j to y_i is defined as g_{ij} . And let the transfer function matrix $\mathbf{G}$ be Equation 13:

$$\mathbf{G} = \begin{bmatrix} g_{11} & \cdots & g_{1n} \\ \vdots & \ddots & \vdots \\ g_{n1} & \cdots & g_{nn} \end{bmatrix} \quad (13)$$

The control system $\mathbf{K}$ is as Equation 14 shows.

$$\mathbf{K} = \text{diag}[k_{11} \quad k_{22} \quad \cdots \quad k_{nn}] \quad (14)$$

In the formula, k_{ii} ($i = 1, 2, \dots, n$) represents the transfer function of the controller from the output y_i to the reference input r_i .

$$\begin{cases} y_i = g_{ij}u_j + \bar{\mathbf{G}}_i^{ij}\bar{\mathbf{u}}^j \\ \bar{\mathbf{y}}^j = \bar{\mathbf{G}}_j^{ij}u_j + \bar{\mathbf{G}}^{jj}\bar{\mathbf{u}}^j \\ \bar{\mathbf{u}}^j = -\bar{\mathbf{K}}^{jj}\bar{\mathbf{y}}^j \end{cases} \quad (15)$$

If the reference input is set to zero, the system output variables can be expressed as:

In the formula, $\mathbf{u} = [u_1 \quad u_2 \quad \cdots \quad u_n]^T$, let $\bar{\mathbf{u}}^j$ represent the input variable column vector of $\mathbf{u}$ with the j -th row element removed; $\mathbf{y} = [y_1 \quad y_2 \quad \cdots \quad y_n]^T$, $\bar{\mathbf{y}}^j$ represent the output variable column vector of $\mathbf{y}$ with the i -th row element removed; $\bar{\mathbf{G}}_i^{ij}$ represent the row vector of the i -th row of $\mathbf{G}$ after removing g_{ij} ; $\bar{\mathbf{G}}_j^{ij}$ represent the column vector of the j -th column of $\mathbf{G}$ after removing g_{ij} ; $\bar{\mathbf{G}}^{jj}$ represent the matrix of $\mathbf{G}$ after removing the i -th row and j -th column; $\bar{\mathbf{K}}^{jj}$ represent the matrix of $\mathbf{K}$ after removing the i -th row and i -th column.

Based on Equation 15, after considering the coupling effect of other control loops on this control loop, the transfer function G_{ij} from u_j to y_i can be expressed as Equation 16 show:

$$G_{ij} = g_{ij} - \bar{\mathbf{G}}_i^{ij}\bar{\mathbf{K}}^{jj}\bar{\mathbf{G}}_j^{ij}(\mathbf{E} + \bar{\mathbf{K}}^{jj}\bar{\mathbf{G}}^{jj})^{-1} \quad (16)$$

Taking a two-input two-output system as an example, to reflect the interactive coupling that exists in multiple control loops, controllers k_{21} and k_{12} that reflect cross-coupling should be designed based on the system described by G_{12} and G_{21} , in addition to the controllers k_{11} and k_{22} that have already been designed.

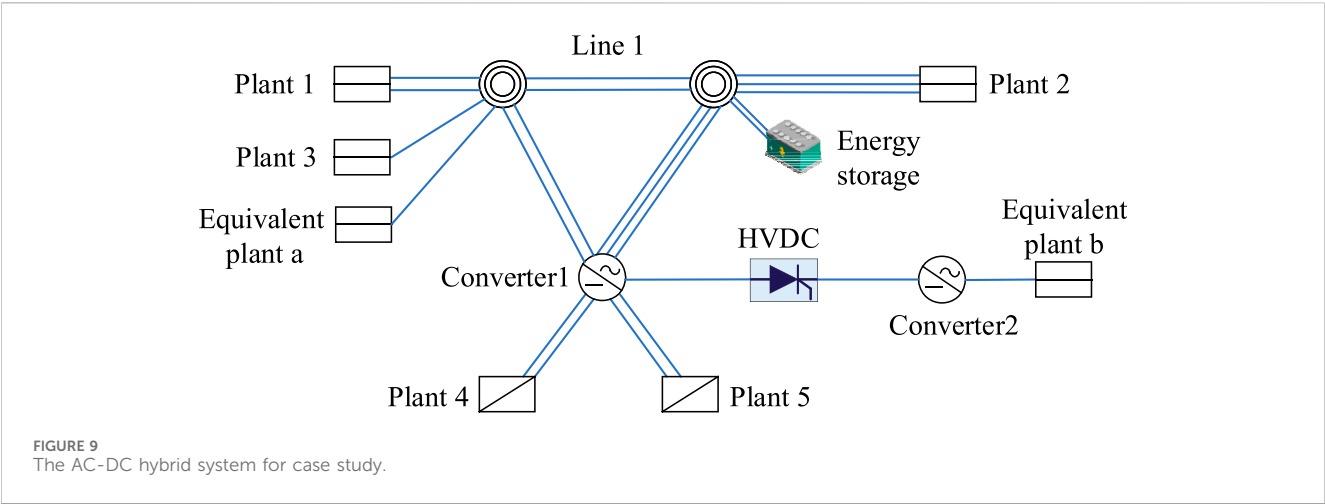
3.4 System identification and feedback signals selections

The TLS-ESPRIT algorithm can be used for both simulation data analysis and system identification. Compared to the Prony method, this algorithm demonstrates stronger resistance to noise and interference, effectively recovering harmonics and obtaining the attenuation characteristic parameters of signals. It is one of the most commonly used methods for offline identification of low-frequency oscillations in power systems. The general idea of this algorithm is to construct a Hankel matrix $\mathbf{H}$ using sampled data and then decompose it into signal subspace and noise subspace through singular value decomposition. This allows for the calculation of the signal's rotation operator, which can further be used to obtain the oscillation frequency, attenuation factor, and other characteristic parameters of each oscillation mode (Chen et al., 2019).

When designing controllers, to ensure good damping effects on the system's weakly damped modes and minimize coupling effects on other modes, the dominant mode ratio (DMR) is commonly used as an indicator to measure the damping effect per unit of control output. This metric assesses the relative strength of feedback signals and the damping performance of the controller on a specific oscillation mode. Compared to the residue method, this indicator eliminates the influence of dimensions, allowing for direct comparisons between different types of signals. The calculation formula for the dominant mode ratio (η) is as follows Equation 17 shows (He et al., 2007; Wu et al., 2024; Li Jiajun et al., 2024):

$$\eta = \frac{|c\phi_i||z_i(0)|}{\sum_{j=1}^n |c\phi_j||z_j(0)|} \quad (17)$$

In the formula: $|c\phi_i|$ represents the observability of the feedback signal for the dominant oscillation mode i ; $|z_i(0)|$ represents the



amplitude of the dominant oscillation mode i at the initial moment after perturbation; $|c\phi_j|$ represents the observability of the feedback signal for the j -th oscillation mode; $|z_j(0)|$ represents the oscillation amplitude of the j -th oscillation mode at the initial moment after perturbation; n represents the number of system oscillation modes.

Different perturbations in the system can cause variations in the initial amplitude $|z_k(0)|$ of different oscillation modes. Therefore, in this paper, when calculating the dominant mode ratio (DMR) indicator, the average value of DMR obtained after applying different perturbations to the system is used.

In summary, the steps for designing a controller are as follows: First, small perturbations are applied to the system at steady state, and the TLS-ESPRIT algorithm is used. Based on the dominant mode ratio indicator, appropriate feedback signals in the system are selected, and a MIMO system model is established corresponding to the HVDC and energy storage control channels. The system's transfer function is obtained and reduced in order. Then, using robust control methods and considering the coupling effects between different control loops, a coupled robust damping controller is designed. Finally, it is attached to the corresponding control loop, and its effectiveness is verified through actual system simulation.

And the detail MIMO system identification procedures can be described as follows: A step disturbance is applied at the setpoint of the DC power, and the changes in the speed differences of key generators are measured respectively. Subsequently, the TLS-ESPRIT algorithm is utilized to identify the transfer functions from the DC system to various generators. Concurrently, another step disturbance is imposed on the setpoint of the constant power control in the outer loop of the energy storage system, and the changes in the speed differences of the key generators are again measured individually. The TLS-ESPRIT algorithm is then employed to identify the transfer functions from the energy storage system to different generators. Ultimately, the MIMO (Multi-Input Multi-Output) transfer functions from both the DC transmission and the energy storage system to the speeds of key generators are obtained.

TABLE 1 The damping characteristics of the case study system.

Name	Parameters	Value
Plant1	Rated capacity	600 MW
Plant2	Rated capacity	400 MW
Plant3	Rated capacity	300 MW
Plant4	Rated capacity	600 MW
Plant5	Rated capacity	600 MW
Equivalent plant a	Rated capacity	1200 MW
Equivalent plant b	Rated capacity	1300 MW
HVDC	Rated capacity	1000 MW
	DC voltage	± 500 kV
Energy storage	Rated capacity	400 MW

TABLE 2 The damping characteristics of the case study system.

Oscillation Modes/Hz	Damping ratio/%
1.61	0.12
0.82	4.86

4 Case study

To validate the proposed strategy, simulations were conducted using PSCAD software. The AC-DC hybrid system, as illustrated in Figure 9, was established, and its parameters are presented in Table 1. In this hybrid HVDC system, energy storage and HVDC are used as low frequency oscillation damping controllers.

The oscillation modes of the system identified by the TLS-ESPRIT algorithm are shown in Table 2. Generally, modes with a damping ratio less than 5% are considered as weakly damped modes. As can be seen from the table, there are two weakly damped oscillation modes in the system, with oscillation frequencies of

TABLE 3 The feedback signals evaluation based on dominant mode ratio.

Feedback signals	Dominant mode ratio
$\Delta\omega_1$	0.32
$\Delta\omega_2$	0.28
$\Delta\omega_3$	0.12
$\Delta\omega_4$	0.19
$\Delta\omega_5$	0.10
ΔP_1	0.08
ΔP_2	0.07

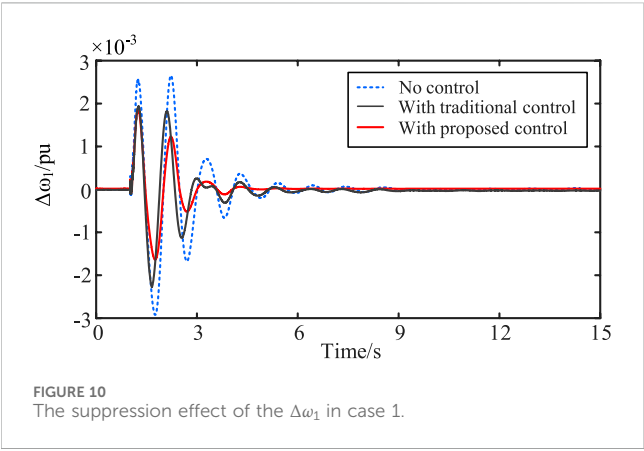


FIGURE 10 The suppression effect of the $\Delta\omega_1$ in case 1.

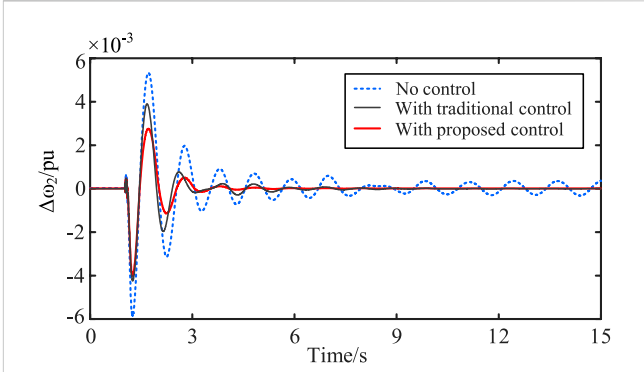


FIGURE 11 The suppression effect of the $\Delta\omega_2$ in case 1.

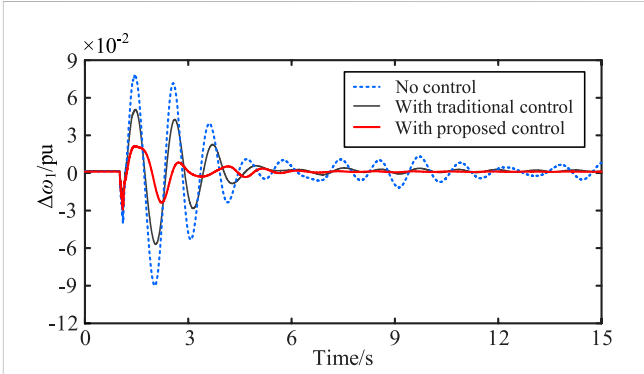


FIGURE 12 The suppression effect of the $\Delta\omega_1$ in case 2.

1.61 Hz and 0.82 Hz, respectively. Among them, the damping ratio of the oscillation frequency of 1.61 Hz is close to 0, which is the dominant oscillation mode of the system.

The feedback signals of the controllers are also evaluated based on dominant mode ratio index. The reason why we choose rotor speed and active power as the feedback signal is that the low frequency oscillations are related to the active power generated by plants. Such that the signals of rotor speed, active power can all be selected as the feedback signals. According to Table 3, it can be seen that the rotor speed deviation of plant1 and plant2, namely $\Delta\omega_1$ and $\Delta\omega_2$, are the most two suitable input signals for controller design, as they have bigger index values. The following parts will use the proposed method to suppress the low frequency oscillations, where the traditional PI controllers are also designed for comparison.

4.1 Case 1

At 1s, a single-phase grounding short-circuit fault occurs on Line 1, with a duration of 0.1s. We compare the feedback signals $\Delta\omega_1$ and $\Delta\omega_2$ under three conditions: without controller, with coupled robust damping controller, and with traditional controller. The suppression effect of the $\Delta\omega_1$ is shown in Figure 10, and the suppression effect of the $\Delta\omega_2$ is shown in Figure 11. In the figures, both $\Delta\omega_1$ and $\Delta\omega_2$ are per unit values.

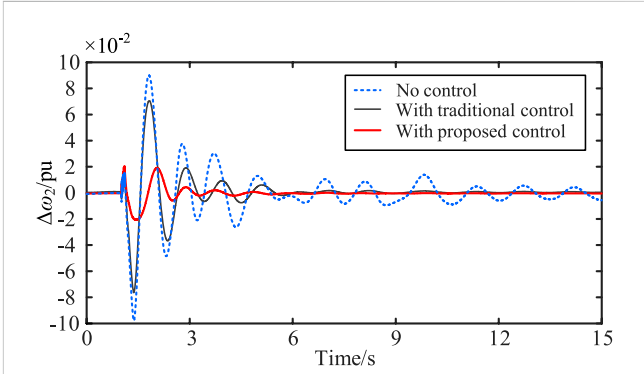
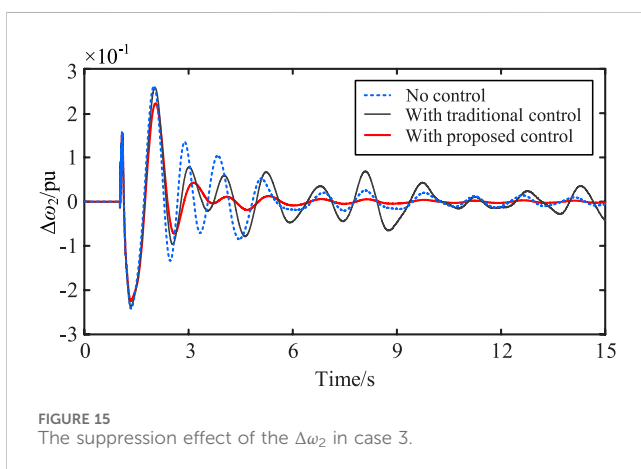
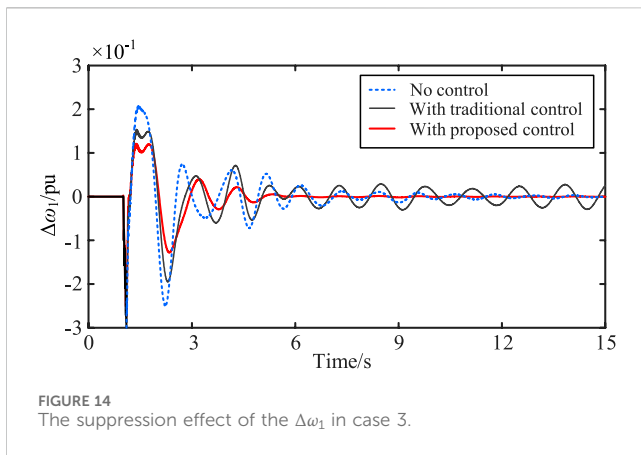


FIGURE 13 The suppression effect of the $\Delta\omega_2$ in case 2.

It is obvious that under case 1 condition, both the coupled robust controller proposed in this paper and the traditional classical controller can suppress low-frequency oscillation. However, the coupled robust controller demonstrates better recovery characteristics and has a more effective suppression effect compared to the traditional controller, which proves the advantages of the proposed method in this paper.



4.2 Case 2

At 1s, the generator in plant 3 is disconnected. The comparison of the suppression effects on the oscillation modes of $\Delta\omega_1$ and $\Delta\omega_2$ under this operating condition is shown in Figures 12, 13.

Similarly, under case 2 condition, both the coupled robust controller proposed in this paper and the traditional classical controller can also suppress low-frequency oscillation. And the coupled robust controller also can obtain better results. It proves that the robustness of the proposed method can be guaranteed when different faults happen.

4.3 Case 3

At 1 s, a three-phase short-circuit fault occurs in the system, which is cleared after 0.1 s. The comparison of the suppression effects of the $\Delta\omega_1$ and $\Delta\omega_2$ oscillation modes in this scenario is shown in the Figures 14, 15.

As shown in the Figures 14, 15, when the system experiences a more severe three-phase short-circuit fault compared to case 1 and case2, during the initial oscillation phase, the traditional damping controllers can suppress the low-frequency oscillation of the system. However, for a considerable period after the fault is cleared, the traditional control is unable to fully restore the system to a stable

operating state. This is because the two traditional controllers are independently designed based on the linearized models of two small-disturbance SISO (Single-Input Single-Output) systems, and they do not account for or eliminate the negative interactions between the multiple control loops in the MIMO (Multi-Input Multi-Output) system. As a result, the system may experience underdamped oscillations, which weaken the damping effect of the controllers. On the other hand, the controller designed based on MIMO control considerations not only considers the interactions between control loops during its design but also possesses robustness, ensuring stable operation of the system under various conditions.

5 Conclusion

This study explored the multi-resource and multi-objective collaboration damping control strategy based on HVDC and energy-storage system. Drawing from previous analyses and simulation results, several key conclusions can be summarized, as follows:

- (1) Due to the complex grid structure and numerous control devices in the actual system, it is difficult for decentralized damping controllers designed based on the decoupling of control loops to exert their advantages. In this paper, a coupled robust damping controller considering the interaction between multiple control loops is designed based on the MIMO system, avoiding decoupling in complex control loop situations. Compared to traditional damping control methods, this approach demonstrates good suppression effects under various operating conditions.
- (2) The controller is implemented through DC and energy storage systems, extending the control capabilities of existing system equipment. Simulation results indicate that compared to traditional damping controllers, the coupled robust controller is insensitive to system perturbations and can effectively suppress low-frequency oscillation of the system under various disturbances and failures, demonstrating strong robust performance.
- (3) The proposed control method can be extended to systems with more input and output signals, providing a reference for the stable operation and control of complex power grids, including scenarios such as energy storage systems equipped with renewable energy and VSC-HVDC transmission systems.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

Author contributions

BZ: Conceptualization, Formal Analysis, Investigation, Writing—original draft, Writing—review and editing. YX: Software,

Writing—original draft, Writing—review and editing. XS: Data curation, Writing—review and editing. XJ: Supervision, Writing—review and editing.

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Time domain analysis of flyback EMI based on distributed parameters theory

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The electromagnetic interference (EMI) is a key issue for most of electric or electronic devices, the researches of suitable analysis methods and solutions absorb much attentions. The research about electromagnetic interference (EMI) is normally based on two approaches (i.e., time domain analysis and frequency domain analysis). The frequency domain analysis is easier to apply theoretical model to analyze the circuit loop impedance of EMI noise which can provide guidance of EMI suppression, but it cannot reflect the real amplitude of EMI noise without the real waveforms of noise sources. The time domain analysis is usually based on circuit simulation to get the EMI noise directly, but it is time consuming. In order to combine the advantages of these two approaches, this paper proposes a high frequency equivalent circuit model of flyback converter and analyze the effect of each distributed parameters on the common mode (CM) and differential mode (DM) noise. The experimental result verified the accuracy and efficiency of the proposed method, which not only can correctly reflect the real waveforms of the circuit, but also can predict the EMI noise based on the constructed waveforms.

KEYWORDS

time domain, distributed parameters, flyback, EMI, equivalent circuit

1 Introduction

In switch-mode power supplies (SMPS), to realize high power density design of power converters, the distance between components should be reduced which could strengthen the electromagnetic field coupling among the components, making the EMI issues more prominent. Therefore, effectively analyzing the mechanisms of EMI noise and proposing corresponding effective solutions are crucial for achieving high power density designs in SMPS.

The essential reason of EMI noise arises from the voltage transitions and current ripples induced by switching actions of the power converters. Consequently, constructing the switching waveforms of the main components in the power converters can effectively reflect the characteristics of the EMI noises. However, since the impact of EMI noises span high frequency bands, various distributed parameters of the circuit significantly influence the switching waveforms, rendering the corresponding differential equations exceedingly complex. As a result, current theoretical analyses of EMI noise are typically based on frequency domain analysis. Time domain analysis is mainly based on circuit simulation, but building and debugging a circuit simulation system has certain complexities. Therefore, it is of great significance to construct time-domain waveforms theoretically to analyze and predict EMI noise (Trinchero et al., 2015; Yang et al., 2020; Fu, 2019; Zhu et al., 2020;

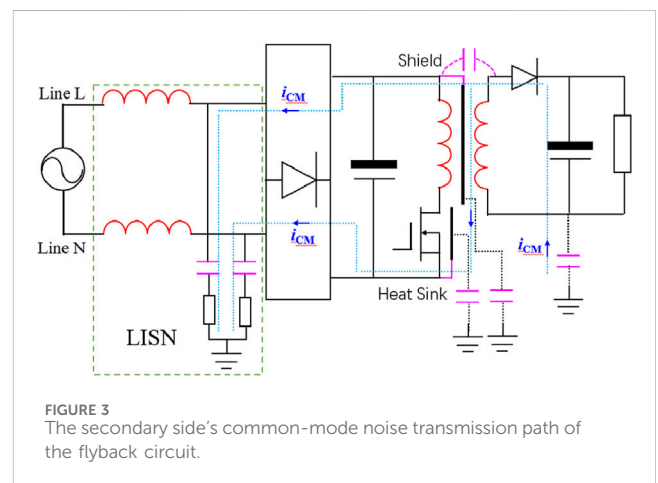
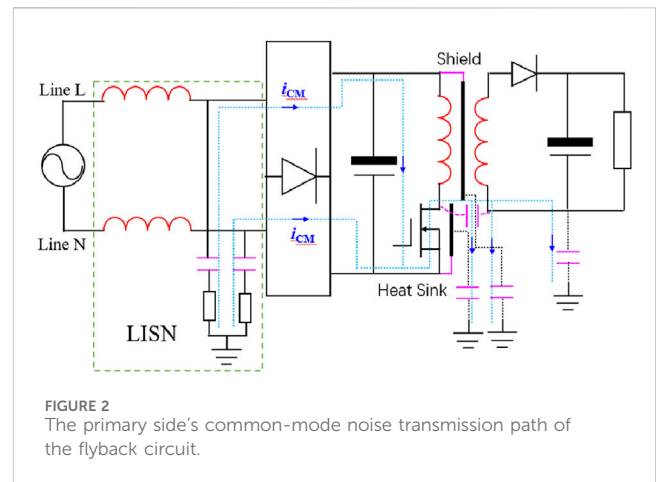
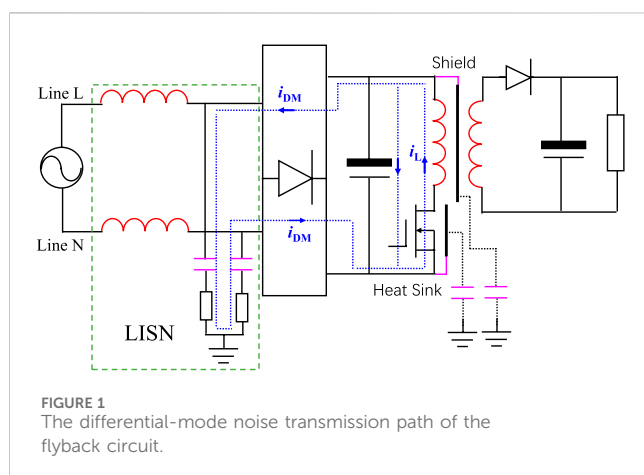
Takahashi et al., 2019; Ma et al., 2008; Liu et al., 2015; Emami et al., 2010; Kai et al., 2024; Fakhakh and Ammous, 2017; Wang et al., 2012; Liang et al., 2012).

This paper utilizes the flyback converter as a platform, and based on its high frequency equivalent circuit model, provides a detailed analysis and solution of the differential equations corresponding to each switching process. This establishes the theoretical solution model for the switching waveforms in the flyback circuit. Building upon this foundation, this paper thoroughly examines the impact of various distributed parameters of the flyback circuit on the waveforms of different components, thereby deducing their influence on EMI noises (differential mode and common mode noises). Finally, through experimental verification, it is confirmed that the theoretically constructed switching waveforms of the flyback converter are highly accurate and can be effectively used to analyze the impact of distributed parameters on EMI noises.

2 EMI noise mechanisms of flyback circuit

When analyzing EMI noise in switch-mode power supply circuits, it is common to categorize the noise into differential mode and common mode components due to their distinct causes and transmission paths. Differential mode noise primarily originates from current ripples in the power loop of the circuits, thus its main transmission path lies between the line L and line N within the power loop. The DM noise transmission path of the flyback circuit is illustrated in Figure 1. Common mode noise, on the other hand, mainly arises from potential jumps during the switching process, leading to ground displacement currents. Therefore, its transmission path involves the loops between the line L and ground, as well as the line N and ground. In a flyback circuit, both the MOSFET of primary side and the diode of secondary side experience potential jumps during the switching instant, which can generate common mode noise. Figures 2, 3 show the transmission path of common-mode noise generated from the primary side and secondary side separately.

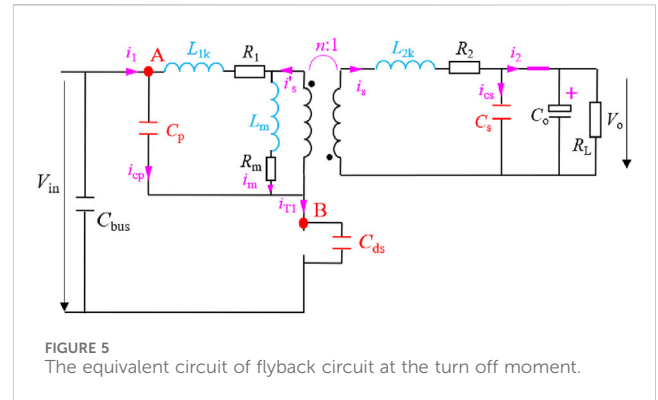
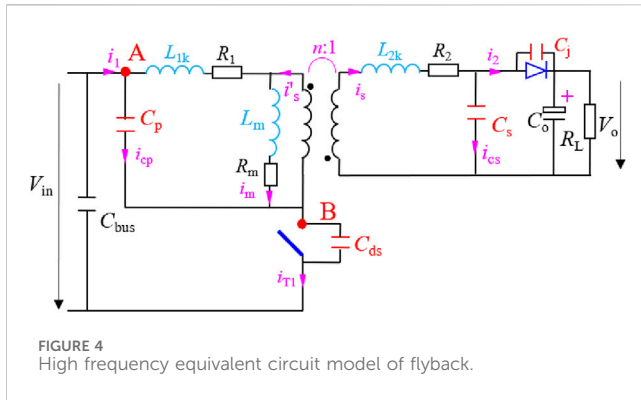
Based on the noise transmission paths shown in Figures 1–3, one can intuitively analyze the impact of circuit's distributed parameters



on EMI noise propagation in the frequency domain. However, this method only reflects the effect of distributed parameters on noise within a given frequency band. In the absence of the original noise's spectral information, this approach is suitable for qualitatively analyzing the impact of distributed parameters on EMI noise but does not facilitate predictive analysis of EMI during the design process of switch-mode power supplies. Generally, the waveforms at various points in the circuit are obtained via simulation and used for EMI prediction and design. However, constructing a waveform model that considers distributed parameters through circuit theory would be more beneficial for an in-depth analysis of these parameters' impact on EMI. Therefore, this paper will analyze the influence of distributed parameters on EMI noise sources by theoretically constructing the switching waveforms of the flyback circuit.

3 Theoretical model of flyback circuit waveforms

To construct relatively accurate switching waveforms for the flyback circuit, it is necessary to fully consider the effects of distributed parameters. For this purpose, the high-frequency equivalent circuit of the flyback transformer shown in Figure 4 is



adopted in this paper. This circuit not only takes into account the leakage inductance, parasitic capacitance, excitation inductance, winding resistance, and core equivalent resistance, but also considers the impact of the parasitic capacitance of the primary side switch and the secondary side diode. This approach can more comprehensively reflect the influence of distributed parameters on the switching waveforms. Due to the majority of current mainstream flyback power supplies operate in DCM (Discontinuous Conduction Mode), which is in order to reduce winding losses and improve efficiency, this paper will use DCM as an example. Based on Figure 4, we will conduct a detailed analysis and construct the switching waveforms corresponding to each switching process.

3.1 The waveforms of flyback circuit at turn off moment

At the turn-off moment of the flyback circuit, the primary switch is turned off and the secondary diode conducts (with a voltage drop V_D). The voltage of the magnetizing inductance L_m is clamped by the output voltage, causing resonance between the primary leakage inductance L_{1k} , the primary capacitance C_p , and the switch's drain-source capacitance C_{ds} , as shown in Figure 5 which represents the equivalent circuit during the turn-off period of the flyback circuit. Based on Figure 5, the Equation 1 for the turn-off moment can be derived. In solving for the magnetizing inductance voltage $V_{Lm_off}(t)$, the influence of the secondary side resistance R_2 and the magnetizing resistance R_m is neglected, assuming that the magnetizing current decreases linearly at this time.

By applying the Laplace transform to solve Equation 1, the time-domain equations for the switch capacitance voltage and magnetizing current at the turn-off moment can be obtained. Substituting these into Equation 2 allows further solution of the key circuit waveforms.

It is important to note that to completely solve Equation 1, the initial values of the switch junction capacitance voltage, the current through the switch, and the magnetizing current must be determined. At the turn-off moment, the current through the switch instantaneously transfers to the secondary side for “freewheeling,” meaning the magnetizing current remains constant (reaching its peak value I_{peak}) and the switch current becomes zero. The voltage across the switch junction capacitance C_{ds} cannot change instantaneously and remains at zero. Therefore, at the turn-off moment, the initial values of the switch junction capacitance voltage, the current through the switch, and the magnetizing current satisfy Equation 3.

It should be noted that the oscillation frequencies corresponding to each switching process are different. Since resistance is frequency-dependent, a fixed resistance value cannot represent the actual winding resistance (such as R_1 and R_2) or core equivalent resistance (such as R_m) for each process. These values need to be estimated based on the oscillation frequency. At the turn-off moment, the resonance frequency of the primary circuit can be estimated using Equation 4. These resistances mainly act as damping during the LC resonance process. As long as their values do not affect the nature of the resonance (such as underdamped, critically damped, and overdamped resonance), they will not impact the analysis of the influence of other distributed parameters (such as transformer leakage inductance and parasitic capacitance) on the switch waveform. Therefore, the estimated value range for these resistances can be quite broad.

$$\begin{cases} L_{C1} \cdot \frac{d^2 V_{cds_off1}(t)}{dt^2} + R_{C1} \cdot \frac{d V_{cds_off1}(t)}{dt} + V_{cds_off1}(t) = V_{in} + V_{F1} \\ L_m \frac{d i_{m_off1}(t)}{dt} + R_m \cdot i_{m_off1}(t) = -V_{F1} \\ L_{C1} = L_{1k} \cdot (C_p + C_{ds}) \\ R_{C1} = R_1 \cdot (C_p + C_{ds}) \\ V_{F1} = \frac{n \cdot (V_o + V_D)}{1 + n^2 \cdot L_{2k}/L_m} \end{cases} \quad (1)$$

$$\begin{cases} V_{cp_off1}(t) = V_{in} - V_{cds_off1}(t) \\ i_{cp_off1}(t) = C_p \frac{d V_{cp_off1}(t)}{dt} \\ i_{s_off1}(t) = n \cdot (i_{cp_off1}(t) + i_{m_off1}(t) - i_{1_off1}(t)) \end{cases} \quad (2)$$

$$\begin{cases} V_{cds_off1}(0) = 0 \\ \frac{d V_{cds_off1}(0)}{dt} = 0 \\ i_{m_off1}(0) = I_{peak} \end{cases} \quad (3)$$

$$f_{r1} = \frac{1}{2\pi \sqrt{L_{1k} \cdot (C_p + C_{ds})}} \quad (4)$$

3.2 The waveforms of flyback circuit at discontinuous interval

During the discontinuous conduction period of the flyback circuit, the secondary side diode turns off, and the magnetizing

inductance is no longer clamped by the output voltage. Instead, it participates in the resonance of the primary side circuit along with the primary leakage inductance L_{lk} , the primary winding capacitance C_p , and the switch drain-source capacitance C_{ds} . The resonant frequency can be calculated using Equation 5. As shown in Figure 6, the equivalent circuit during the discontinuous conduction period can be used to derive the corresponding circuit equations, as given in Equation 6. When the freewheeling time t_{off} of the secondary side diode is determined, the initial values of the voltages and currents across the switch and the secondary winding during this period satisfy Equation 7.

Using the Laplace transform to solve Equation 6, the time-domain equations for the voltages across the switch and the secondary winding during the discontinuous conduction period can be obtained. Substituting these time-domain equations into Equation 8 allows the key waveforms of the circuit during the discontinuous conduction period to be determined.

$$f_{r2} = \frac{1}{2\pi\sqrt{(L_{lk} + L_m) \cdot (C_p + C_{ds})}} \quad (5)$$

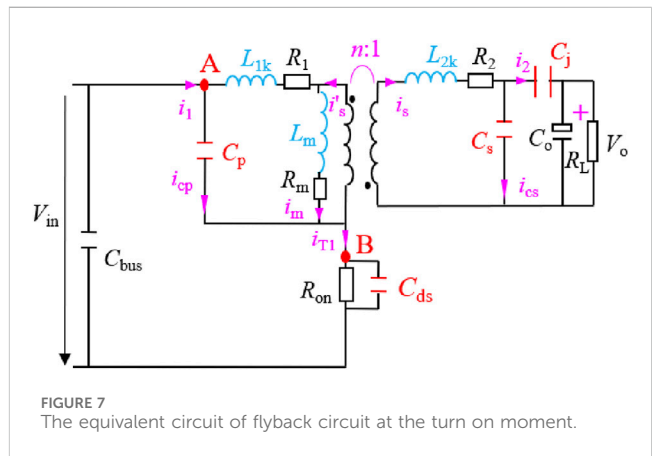
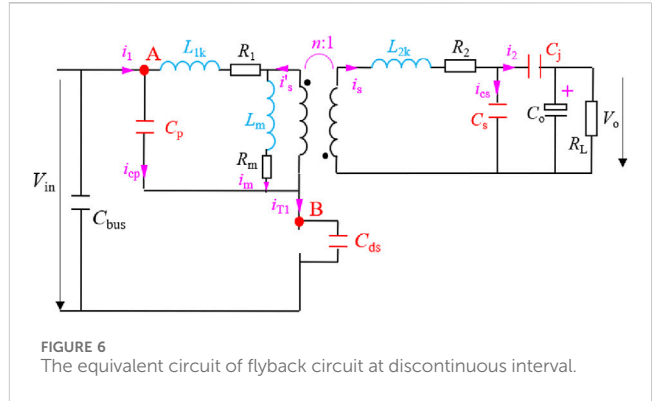
$$\begin{cases} L_{C11} \frac{d^2}{dt^2} V_{cds- off f2}(t) + R_{C11} \frac{d}{dt} V_{cds- off f2}(t) + V_{cds- off f2}(t) = V_{in} - L_{C12} \frac{d^2}{dt^2} V_{cs- off f2}(t) - R_{C12} \frac{d}{dt} V_{cs- off f2}(t) \\ L_{C21} \frac{d^2}{dt^2} V_{cds- off f2}(t) + R_{C21} \frac{d}{dt} V_{cds- off f2}(t) + L_{C22} \frac{d^2}{dt^2} V_{cs- off f2}(t) + R_{C22} \frac{d}{dt} V_{cs- off f2}(t) + V_{cs- off f2}(t) = 0 \\ L_{C11} = (L_{lk} + L_m) \cdot (C_p + C_{ds}) \\ L_{C12} = \frac{L_m(C_s + C_j)}{n} \\ L_{C21} = \frac{L_m(C_p + C_{ds})}{n} \\ L_{C22} = (L_{2k} + L_m/n^2) \cdot (C_s + C_j) \\ R_{C11} = (R_1 + R_m) \cdot (C_p + C_{ds}) \\ R_{C12} = \frac{R_m \cdot (C_s + C_j)}{n} \\ R_{C21} = \frac{R_m \cdot (C_p + C_{ds})}{n} \\ R_{C22} = (R_2 + R_m/n^2) \cdot (C_s + C_j) \end{cases} \quad (6)$$

$$\begin{cases} V_{cds- off f2}(0) = V_{cds- off f1}(t_{off}) \\ \frac{d}{dt} V_{cds- off f2}(0) = \frac{d}{dt} V_{cds- off f1}(t_{off}) \\ V_{cs- off f2}(0) = V_{cs- off f1}(t_{off}) \\ \frac{d}{dt} V_{cs- off f2}(0) = \frac{d}{dt} V_{cs- off f1}(t_{off}) \end{cases} \quad (7)$$

$$\begin{cases} V_{cp- off f2}(t) = V_{in} - V_{cds- off f2}(t) \\ V_{D- off f2}(t) = V_{cs- off f2}(t) - V_o \\ i_{1- off f2}(t) = C_{ds} \frac{d}{dt} V_{cds- off f2}(t) \\ i_{s- off f2}(t) = (C_s + C_j) \frac{d}{dt} V_{cs- off f2}(t) \end{cases} \quad (8)$$

3.3 The waveforms of flyback circuit at turn on moment

During turn on moment, the switch can be approximately represented as a conducting resistance R_{on} , and in view of its small value, it can be shorted. At this moment, to prevent a sudden change in the voltage across the primary parasitic capacitance of the transformer, a lead inductance L_{1w} , which



can be extracted by PCB simulation, is introduced for analysis. As shown in Figure 7, this is the equivalent circuit at the turn-on moment. The charging current also exhibits a steep spike due to the rate of change of the V_{cp} potential approaching a sudden change. Additionally, since the magnetizing inductance is large at the turn-on instant and the magnetizing current remains essentially constant, the current changes occurring in the circuit can be considered not to flow through the magnetizing loop. Therefore, the magnetizing inductance is effectively disconnected, which is why it is not reflected in the equivalent circuit shown in Figure 7.

Based on Figure 7, the Equation 9 can be derived. When the duration t_d of the discontinuous stage is determined, the initial values of the voltage and current across the primary and secondary winding parasitic capacitors at the moment of turn-on satisfy Equation 10. By applying the Laplace transform to solve Equation 9, the time-domain equations for the port voltages (i.e., parasitic capacitor voltages) of the primary and secondary windings at the moment of turn-on can be obtained. Substituting these into Equation 11 allows further solution of other key waveforms in the flyback circuit at this moment.

At the turn-on moment, a peak in the primary current occurs, and the secondary side also experiences voltage and current oscillations. However, at this time, the oscillation frequencies of the primary and secondary sides are different. Their resonance frequencies are determined by their circuit parameters respectively and can be estimated using Equation 12, where f_{rp}

represents the primary resonance frequency and f_{rs} represents the secondary resonance frequency.

$$\begin{cases} L_{1w} \frac{d}{dt} i_{1-on1}(t) + V_{cp-on1}(t) = V_{in} \\ L_{2k} \frac{d}{dt} i_{s-on1}(t) + R_2 \cdot i_{s-on1}(t) + V_{cs-on1}(t) + \frac{V_{Lm-on1}(t)}{n} = 0 \\ V_{Lm-on1}(t) = \frac{L_{1k}}{n} \frac{d}{dt} i_{s-on1}(t) + \frac{R_1}{n} i_{s-on1}(t) + V_{cp-on1}(t) \\ i_{cp-on1}(t) = i_{1-on1}(t) + \frac{i_{s-on1}(t)}{n} \\ i_{s-on1}(t) = (C_s + C_j) \frac{d}{dt} V_{cs-on1}(t) \end{cases} \quad (9)$$

$$\begin{cases} V_{cp-on1}(0) = V_{cp-off2}(t_d) \\ \frac{d}{dt} V_{cp-on1}(0) = \frac{d}{dt} V_{cp-off2}(t_d) \\ V_{cs-on1}(0) = V_{cs-off2}(t_d) \\ \frac{d}{dt} V_{cs-on1}(0) = \frac{d}{dt} V_{cs-off2}(t_d) \end{cases} \quad (10)$$

$$\begin{cases} i_{s-on1}(t) = (C_s + C_j) \cdot \frac{d}{dt} V_{cs-on1}(t) \\ i_{1-on1}(t) = C_p \frac{d}{dt} V_{cp-on1}(t) - \frac{i_{s-on1}(t)}{n} \\ V_{D-on1}(t) = V_{cs-on1}(t) - V_o \\ V_{cds-on1}(t) = 0 \end{cases} \quad (11)$$

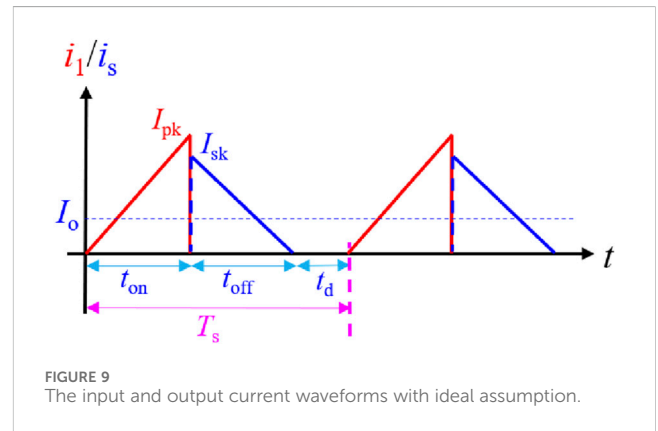
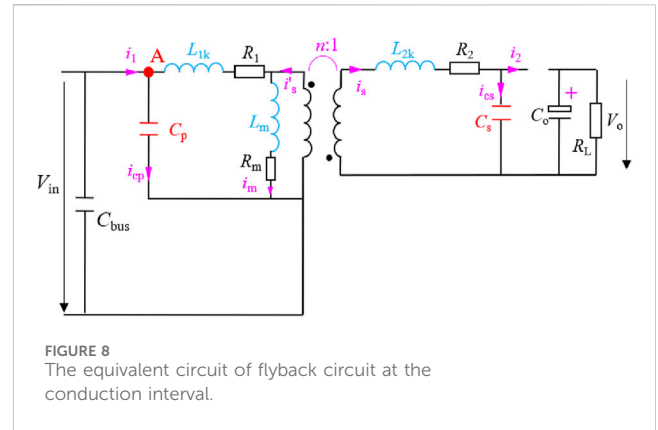
$$\begin{cases} f_{rp} = \frac{1}{2\pi\sqrt{L_{1w} \cdot C_p}} \\ f_{rs} = \frac{1}{2\pi\sqrt{\left(L_{2k} + \frac{L_{1k}}{n^2}\right) \cdot (C_s + C_j)}} \end{cases} \quad (12)$$

3.4 The waveforms of flyback circuit at conduction interval

This article suggests that during the conduction period, the voltage of primary-side capacitor V_{cp} is clamped by the input voltage, and the magnetizing current can be considered to steadily increase under constant voltage excitation, while the secondary-side current is zero. Therefore, the equivalent circuit during the conduction period needs to consider the influence of the magnetizing loop, as shown in Figure 8. Based on this, the corresponding circuit Equation 13 can be derived, which can be simplified to Equation 14. In DCM mode, it can be assumed that the initial value of the primary current $i_{1-on2}(0)$ is zero. However, when the oscillation time t_{rn} at the moment of turn-on is determined, Equation 15 can also be used to estimate $i_{1-on2}(0)$.

$$\begin{cases} L_{1k} \frac{d}{dt} i_{L1k-on2}(t) + R_1 \cdot i_{L1k-on2}(t) + L_m \frac{d}{dt} i_{m-on2}(t) + R_m \cdot i_{m-on2}(t) = V_{in} \\ i_{1-on2}(t) = i_{cp-on2}(t) + i_{L1k-on2}(t) \\ i_{m-on2}(t) = i_{L1k-on2}(t) + \frac{i_{s-on2}(t)}{n} \end{cases} \quad (13)$$

$$(R_1 + R_m) \cdot i_{1-on2}(t) + (L_{1k} + L_m) \cdot \frac{d}{dt} i_{1-on2}(t) = V_{in} \quad (14)$$



$$i_{1-on2}(0) = i_{1-on1}(t_{rn}) \quad (15)$$

3.5 The waveforms of flyback circuit during full period

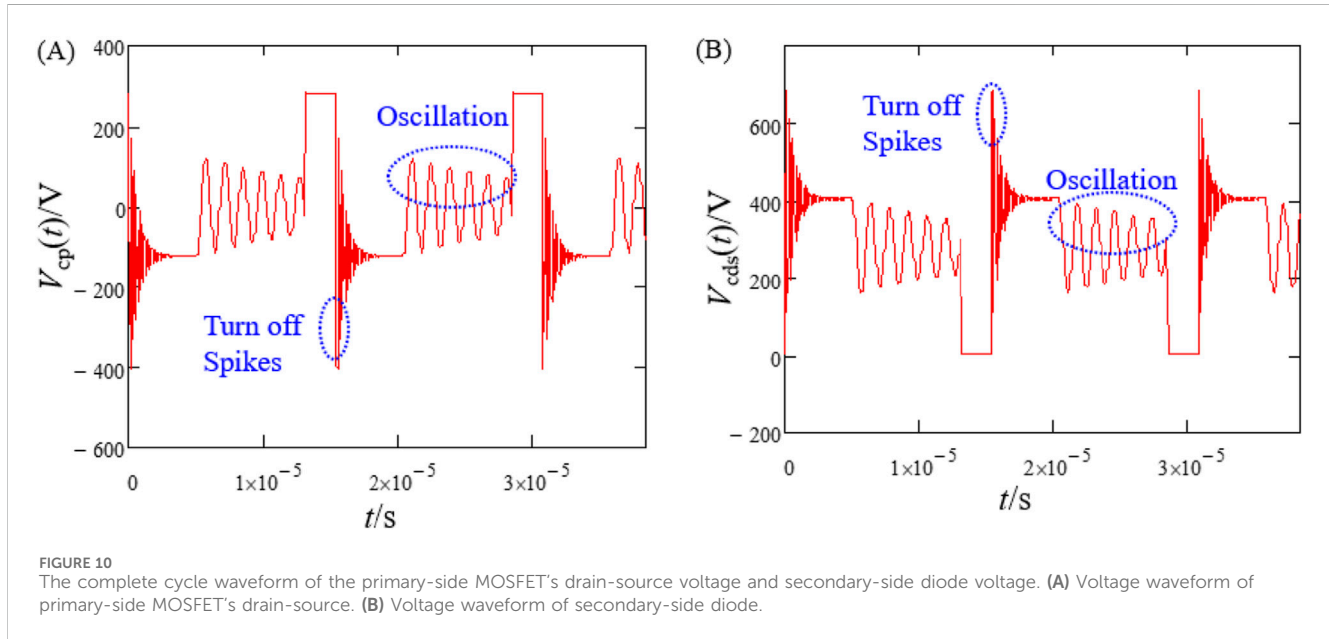
The previous sections have introduced the circuit equations and their solution methods for each switching process in the flyback circuit, which can be used to construct the main waveform characteristics of the circuit during each switching process. However, to fully draw the switch waveforms for the entire cycle, it is necessary to determine the duration of each switching process.

When analyzing the duration of each switching process, this article estimates using ideal primary and secondary current waveforms, as shown in Figure 9, ignoring the influence of winding resistances R_1 and R_2 on the switch waveforms. Although in the DCM mode of the flyback circuit, the circuit operation can be divided into four states: turn-off moment, discontinuous period, turn-on moment, and conduction period, considering that the waveform oscillation at the turn-on moment is relatively short, this article ignores its time proportion during analysis. Therefore, in Figure 9, only three time periods are shown: the conduction time t_{on} , the diode freewheeling time t_{off} , and the discontinuous time t_d .

Given the output power P_o and output voltage V_o , the average output current I_o can be calculated using Equation 16. According to Figure 9, I_o also satisfies Equation 17. During the period of t_{off} ,

TABLE 1 The distributed parameters of flyback circuit.

Parameters	L_{1k} (μH)	L_{2k} (μH)	L_m (μH)	C_p (pF)	C_s (pF)	C_{ds} (pF)	C_j (pF)	n
Value	9	0.225	430	5	647	100	100	10



when ignoring the effect of R_2 , Equation 18 can be derived. During the period of t_{on} , when ignoring the effect of R_1 , Equation 19 can be derived. The input peak current I_{pk} and the output peak current I_{sk} satisfy the rule of ampere-turns balance, as shown in Equation 20. By combining Equations 16, 20, the conduction time t_{on} and the output diode freewheeling time t_{off} can be determined, as shown in Equation 21. The discontinuous time t_d is then equal to the switching cycle minus t_{on} and t_{off} , as shown in Equation 22.

Based on the above analysis and combining the time-domain equations for each stage, the required circuit waveforms for the entire cycle can be written. Taking the voltage of the primary-side MOSFET and the secondary-side diode as examples, their equations for the entire cycle can be represented by Equations 23, 24, respectively.

$$I_o = \frac{P_o}{V_o} \quad (16)$$

$$I_o = \frac{I_{sk} \cdot t_{off}}{2 \cdot T_s} \quad (17)$$

$$\left(L_{2k} + \frac{L_m}{n^2}\right) \cdot \frac{I_{sk}}{t_{off}} = V_o + V_D \quad (18)$$

$$(L_{1k} + L_m) \cdot \frac{I_{pk}}{t_{on}} = V_{in} \quad (19)$$

$$n \cdot I_{pk} = I_{sk} \quad (20)$$

$$\begin{cases} t_{off} = \sqrt{\frac{2 \cdot P_o \cdot T_s \cdot (L_{2k} + L_m/n^2)}{V_o \cdot (V_o + V_D)}} \\ t_{on} = \frac{2 \cdot I_o \cdot T_s \cdot (L_{1k} + L_m)}{n \cdot V_{in} \cdot t_{off}} \end{cases} \quad (21)$$

$$t_d = T_s - t_{on} - t_{off} \quad (22)$$

$$V_{cds}(t) = \begin{cases} V_{cds_off1}(t) & 0 \leq t \leq t_{off} \\ V_{cds_off2}(t) & t_{off} \leq t \leq t_{off} + t_d \\ 0 & t_{off} + t_d \leq t \leq T_s \end{cases} \quad (23)$$

$$V_D(t) = \begin{cases} V_{D_off1}(t) & 0 \leq t \leq t_{off} \\ V_{D_off2}(t) & t_{off} \leq t \leq t_{off} + t_d \\ V_{D_on1}(t) & t_{off} + t_d \leq t \leq t_{off} + t_d + t_{rn} \\ V_{D_on2}(t) & t_{off} + t_d + t_{rn} \leq t \leq T_s \end{cases} \quad (24)$$

To specifically draw the switch waveforms, it is accepted to assume the following parameters: an output power P_o of 28 W, an input voltage V_{in} of 380 V, an output voltage V_o of 12 V, a switching frequency f_s of 65 kHz, primary and secondary winding turns ratio of 40:10, a load resistance R_o of 5.1 Ω , a secondary-side diode forward voltage V_D of 0.5 V, and relevant distributed parameters as shown in Table 1.

According to the analysis in Section 3-1, the values of R_1 , R_2 , and R_m are affected by frequency. Based on the given parameters, the resonant frequency f_{r1} of the primary circuit at the turn-off moment can be calculated as 5.18 MHz. Accordingly, it is assumed that at this time, $R_1 = 30 \Omega$, $R_m = 2 \Omega$, and $R_2 = 5 \Omega$. During the discontinuous period, the primary resonant frequency f_{r2} is 741 kHz, and it is assumed that at this time, $R_1 = 10 \Omega$, $R_m = 1 \Omega$, and $R_2 = 3 \Omega$. At the turn-on moment, it is assumed that $R_1 = 30 \Omega$, $R_m = 30 \Omega$, $R_2 = 5 \Omega$, $L_{1w} = 10 \text{ nH}$, and $R_{1w} = 5 \Omega$. During the conduction period, it is assumed that $R_1 = 30 \Omega$, $R_m = 30 \Omega$, and $R_2 = 1 \Omega$.

Based on the given parameters and combining Equations 23, 24, the complete cycle waveforms of the switch transistor's drain-source voltage and the secondary-side diode voltage can be plotted, as shown in Figure 10.

4 The distributed parameters influence on differential and common-mode noise

According to the analysis in Section 3, the distributed parameters of the flyback circuit have a significant impact on various voltage and current waveforms. The spikes and oscillations in these waveforms are also the fundamental causes of EMI noise generation. Therefore, performing Fourier decomposition on these voltage and current waveforms can reveal the distribution of the noise spectrum they generate.

According to the analysis in Section 3-1, the generation mechanism of differential-mode noise is mainly due to the current ripple in the primary power loop. Therefore, to analyze the impact of distributed parameters on differential-mode noise, Fast Fourier Transform (FFT) can be performed on the primary current i_1 . Common-mode noise is primarily caused by ground distribution currents arising from potential jumps in the circuit. Thus, to analyze the impact of distributed parameters on common-mode noise, FFT can be performed on the primary switch transistor voltage V_{cds} and the secondary diode voltage V_D .

4.1 The distributed parameters influence on differential mode noise

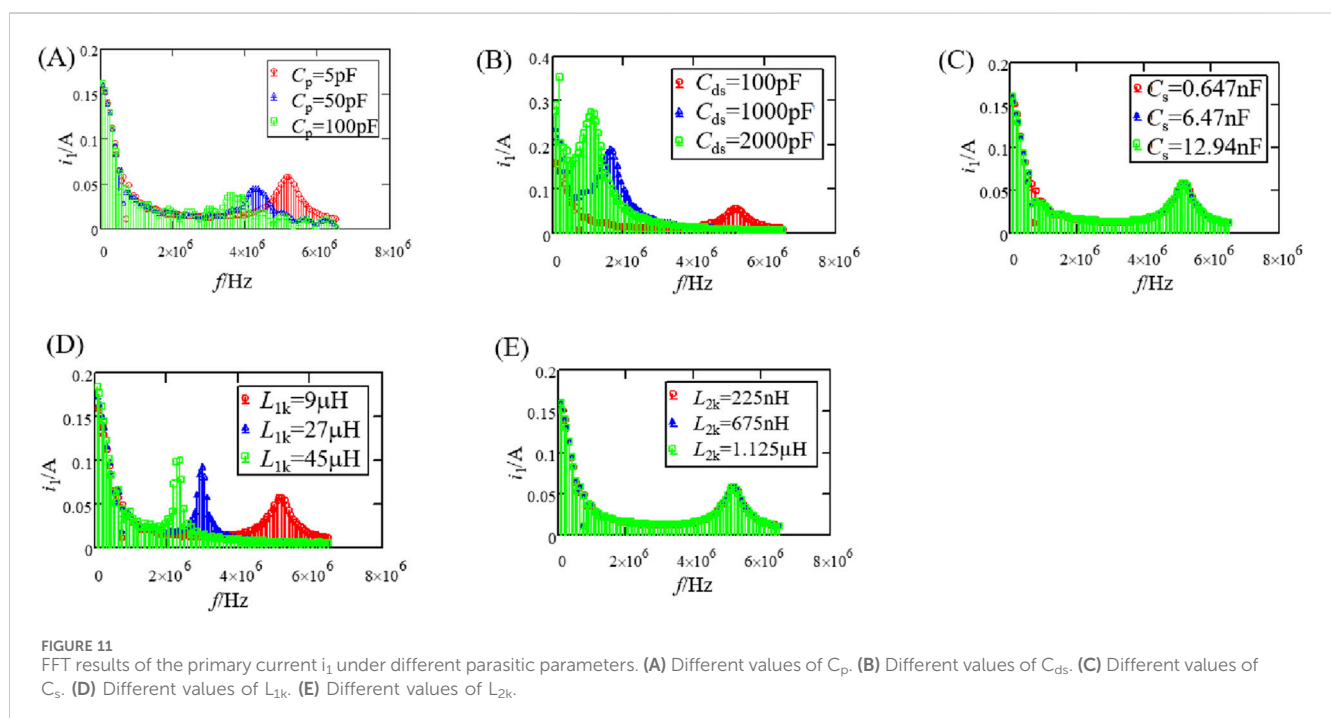
Based on the analysis in Section 3, different primary current waveforms corresponding to various distributed parameters can be constructed and subjected to FFT decomposition to obtain the corresponding noise spectrum, as shown in Figure 11. It can be concluded that an increase in capacitance C_p , C_{ds} , and

primary leakage inductance L_{1k} will worsen the differential-mode noise in the low-frequency range but improve it in the high-frequency range. It is observed that due to the smaller value of C_p , its impact is much less than that of C_{ds} . Moreover, as C_p , C_{ds} , and L_{1k} increase, the frequency corresponding to the noise peaks in the mid-to-high frequency range of differential-mode noise decrease. This undoubtedly requires the filter to enhance its low-frequency filtering performance, increasing the difficulty of filter design. The distributed parameters L_{2k} and C_s of the secondary circuit have essentially no effect on differential-mode noise.

4.2 The distributed parameters influence on common mode noise

Similarly, by constructing the primary switch transistor voltage waveforms and secondary diode voltage waveforms corresponding to different distributed parameters, and performing FFT decomposition to obtain the corresponding common-mode noise spectrum, as shown in Figures 12, 13, the following conclusions can be drawn:

- (1) Due to changes in the secondary winding leakage inductance L_{2k} and parasitic capacitance C_s , the FFT results of V_{cds} and V_D remain essentially unchanged. Therefore, the secondary circuit parameters have essentially no effect on common-mode noise.
- (2) When the primary winding parasitic capacitance C_p and leakage inductance L_{1k} increase, the frequency corresponding to the peaks of common-mode noise in the mid-to-high frequency range decrease, while the high-



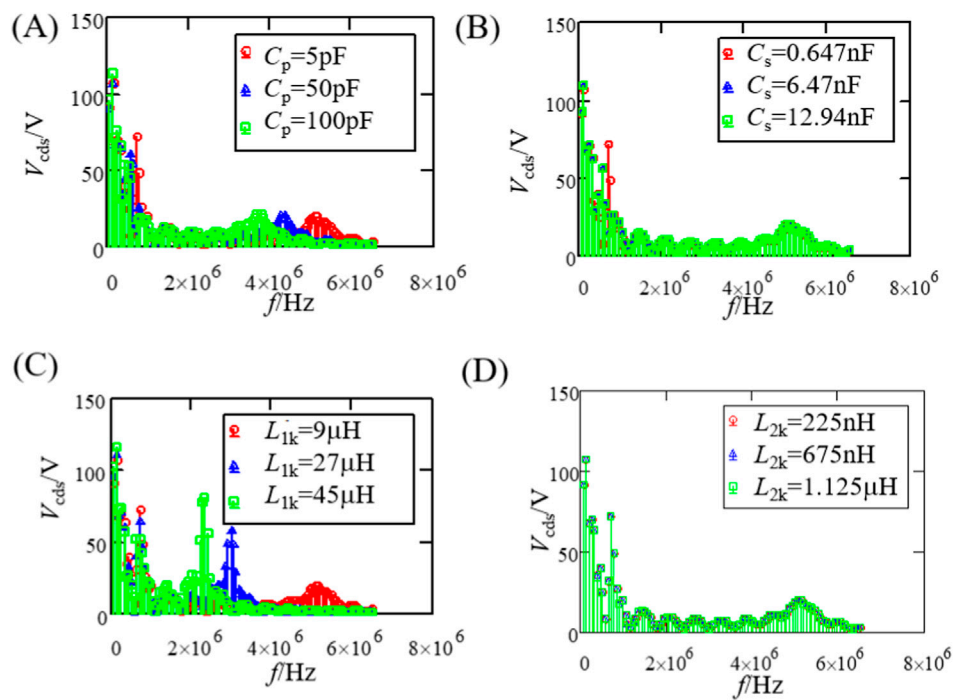


FIGURE 12
FFT results of the MOSFET voltage V_{cds} under different parasitic parameters. (A) Different values of C_p . (B) Different values of C_s . (C) Different values of L_{1k} . (D) Different values of L_{2k} .

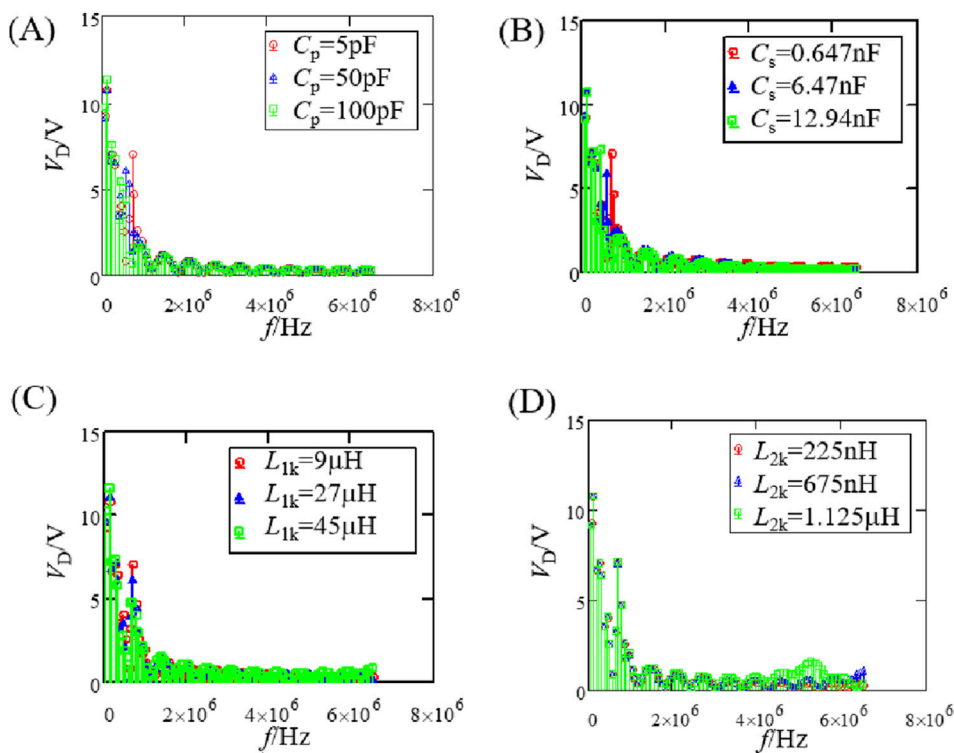
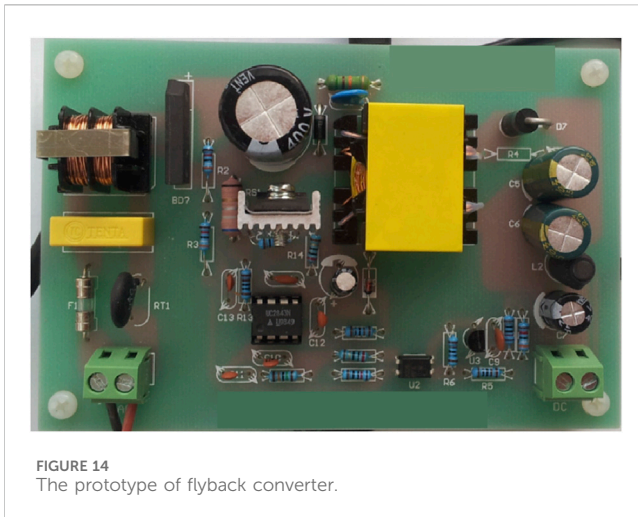


FIGURE 13
FFT results of the diode voltage V_D under different parasitic parameters. (A) Different values of C_p . (B) Different values of C_s . (C) Different values of L_{1k} . (D) Different values of L_{2k} .



frequency common-mode noise is reduced. However, the impact on low-frequency common-mode noise is relatively minor.

5 Experimental verification

The main purpose of this experiment is to verify two aspects, one is to confirm the accuracy of the theoretically constructed switching waveforms, and the other is to validate the effectiveness of analyzing the impact of distributed parameters on EMI noise based on these theoretically constructed switching waveforms. The flyback converter prototype used in this chapter is shown in

Figure 14, and its circuit distributed parameter data as shown in Table 1.

As shown in Figure 15, it includes comparison diagrams between the theoretically calculated main switching waveforms and the measured waveforms. It can be seen that the theoretically calculated switching waveforms accurately reflect the characteristics of each switching waveform during every switching process. There are some differences in peak amplitudes, mainly because there are errors between the values of circuit parameters used in the theoretical analysis and their actual values. However, this does not hinder the use of theoretical switching waveforms for analyzing the impact of flyback distributed parameters on its EMI noise. It should be noted that the time corresponding to each stage of these theoretically calculated waveforms is given based on the measured values, rather than calculated according to the formulas in Section 3–5. This is because Equations 21, 22 ignore the effect of winding resistance, and due to errors in the values of parameters such as leakage inductance and magnetizing inductance, it is impossible to achieve the same calculation results for each stage time as the measured values. To facilitate the comparison of theoretical switching waveforms and measured waveforms, this paper uses the actual measured values of each duration in the theoretical analysis.

To verify the correctness of using the constructed switching waveforms to analyze the impact of distributed parameters on EMI noise, this paper measured the differential and common-mode noise for different values of primary winding capacitance C_p and secondary winding capacitance C_s , as shown in Figure 16. It can be concluded that as C_p increases, its impact on low-frequency differential and common-mode noise is relatively minor, and the frequency corresponding to the peaks of

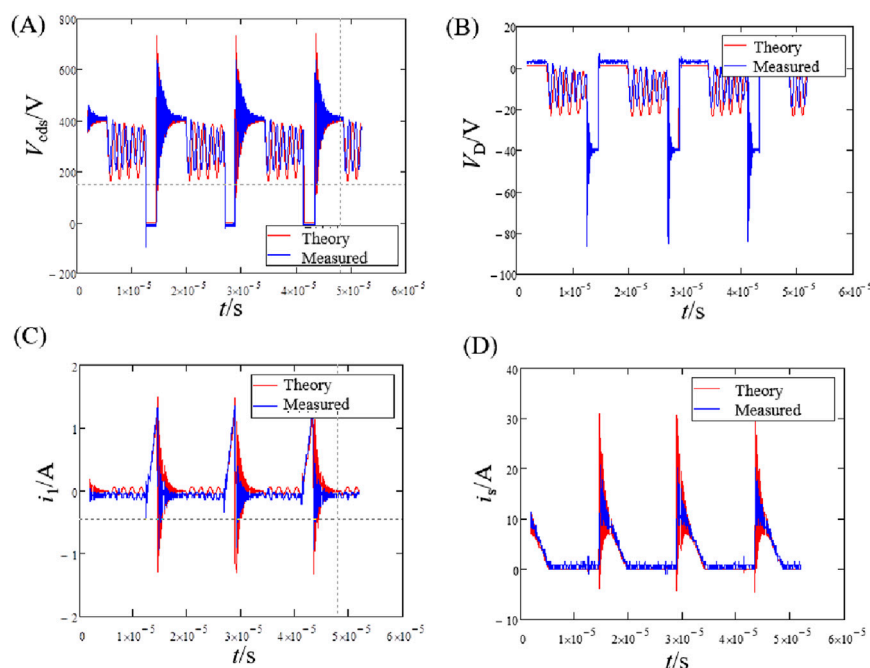


FIGURE 15
Comparisons of theoretical waveforms and measured waveforms. (A) Waveform of V_{cds} . (B) Waveform of V_D . (C) Waveform of i_l . (D) Waveform of i_g .

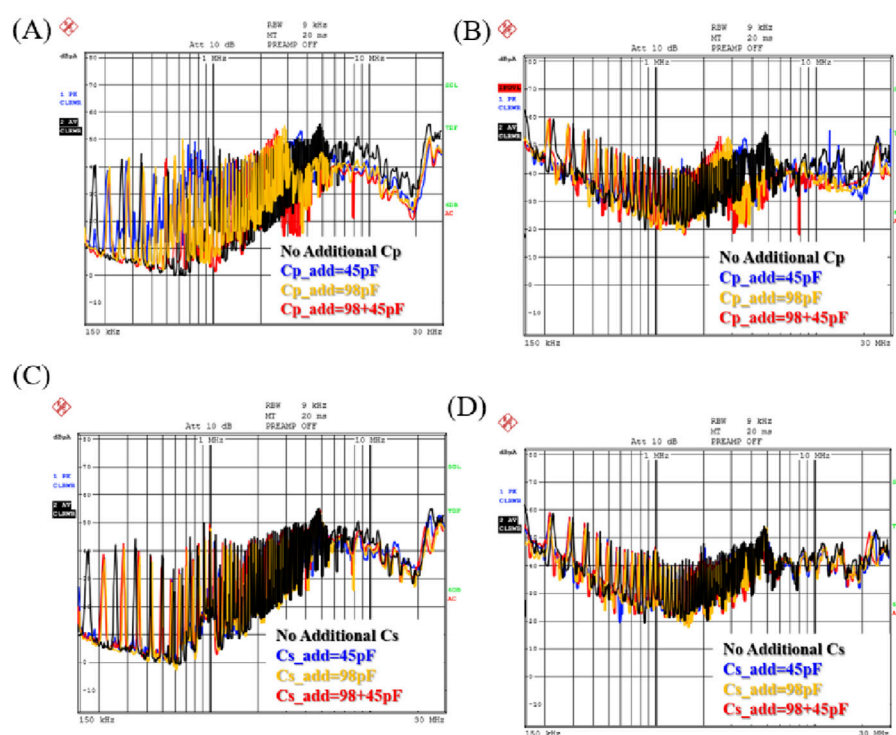


FIGURE 16
Experimental results of parasitic parameters' effect on DM and CM noise. (A) The effect of C_p on CM noise. (B) The effect of C_p on DM noise. (C) The effect of C_s on CM noise. (D) The effect of C_s on DM noise.

differential and common-mode noise at mid-to-high frequency decrease. In contrast, changes in C_s have essentially no effect on the differential and common-mode noise of the entire converter. This conclusion is consistent with the results of theoretical analysis, further validating the effectiveness of the proposed theoretical method for time-domain analysis of the impact of flyback distributed parameters on EMI noise.

6 Conclusion

This paper constructs theoretical time-domain waveforms for the flyback circuit considering transformer distributed parameters, analyzes the impact of transformer distributed parameters on voltage and current waveforms, as well as their effect on EMI noise. Experimental validation confirms the correctness and effectiveness of the established model. The following conclusions can be drawn:

- (1) The constructed theoretical time-domain waveforms match the measured results, accurately expressing the details of the entire cycle's voltage and current waveforms, such as the oscillation characteristics of each part.
- (2) Analyzing the impact of flyback circuit distributed parameters on differential and common-mode noise based on the constructed theoretical waveforms aligns with the trends of the measured noise results, which can guide the design of transformer distributed parameters and EMI rectification.

- (3) The distributed parameters of the primary circuit in the flyback circuit have a larger impact on its EMI noise, while those of the secondary circuit have a smaller effect.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

Author contributions

BL: Writing—original draft. RL: Writing—review and editing. JJ: Writing—review and editing. YW: Writing—review and editing. PX: Writing—review and editing.

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Authors BL, JJ, YW, and PX were employed by Ginlong Technologies Co., Ltd.

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Megawatt-level converter grid-forming control technology by adopting MPC for medium voltage distribution network

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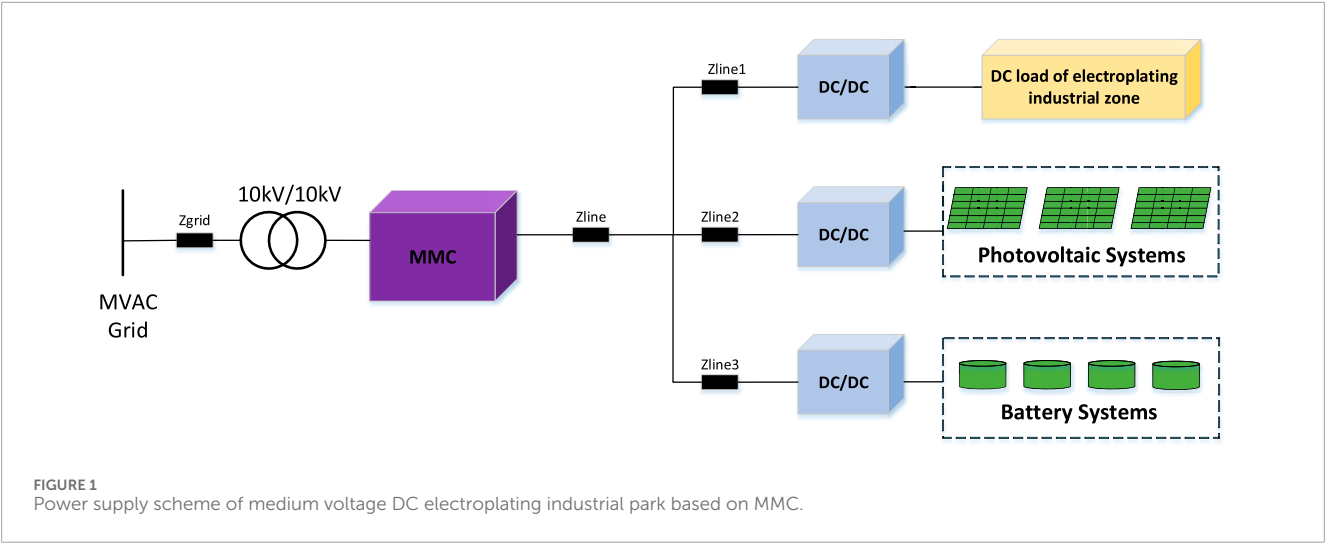
The grid-forming control (GFM) is treated as controlled voltage sources, which can enhance synchronize stability in weak grid case. However, the dynamic response of existing GFM method is limited, and the transient overcurrent in fault is still not solved well, which limits its wide application in distribution network. Drawing from a DC transformation project in an electroplating industrial park in China, this paper proposes an improved GFM control strategy for medium-voltage megawatt converters. The proposal includes the following: firstly, this paper presents an overview of GFM/GFL converters control method; secondly, the establishment of a mathematical model for MMC converters; thirdly, the development of a new MMC grid-forming control technology based on model predictive control, for achieving improved current limitation, dynamic response, and power quality, finally, the verification of these concepts through simulations. This paper provides new insights and strategies for Megawatt-level converter, and reducing carbon emissions in DC industrial parks.

KEYWORDS

DC industrial park, MW-level converter, grid-forming control, model predictive control, carbon reduction, distribution network

1 Introduction

The distribution network is increasingly populated by DC industrial loads, such as those from the electroplating industry, data centers, and energy storage systems, contributing to a significant rise in carbon emissions and environmental pollution. For instance, an electroplating industrial park in Pingyang County, Zhejiang Province, China, has an average annual load of 5 MW, with annual electricity consumption exceeding 100 million kilowatt-hours, 70% of which is DC load. This high energy consumption is accompanied by stringent power quality demands. A critical challenge, therefore, is how to reduce carbon emissions in DC industrial parks, particularly electroplating facilities, through the integration of new energy sources. In 2023, a demonstration project to transform the medium-voltage DC power supply was implemented at Pingyang Electroplating Industrial Park in China. This project utilized two small-capacity VSC converters for ± 10 kV DC energy conversion, leading to significant reductions in energy consumption. However, as the power grid's carbon reduction requirements become more stringent, there is a need to incorporate larger distributed photovoltaic



and energy storage systems. Modular multilevel converter (MMC) technology offers a significant advantage over the C-NPC topology by easily increasing the output voltage and capacity of a single converter. This makes it particularly suitable for applications involving large-scale distributed photovoltaic systems and DC industrial parks. The power supply scheme for a medium-voltage DC electroplating industrial park is shown in Figure 1.

MMC technology is widely utilized in renewable energy, energy storage systems, and HVDC transmission due to its high efficiency. The MMC converters are typically controlled as current sources, which follow the frequency and phase of the grid by phase-locked loops (Pan et al., 2017; Du et al., 2021). This grid synchronization control is known as grid-following control (GFL). The GFL mode based on phase-locked loop control can ensure system stability and power control rapidity under a strong grid, which has many limitations in terms of stability, system voltage, and frequency adjustment, and adapting to a scenario where a high proportion of renewable energy generation units are connected is difficult. When connected to weak grid, the synchronization dynamics of GFL converters is more susceptible to perturbation and then instability (Wang et al., 2018; Hu et al., 2019; Hu et al., 2021).

GFM converters reproduce the behavior of a voltage source behind an impedance, and contributing to the strength of weak power grid by regulating voltage and frequency. Gao et al. (2017) presents an adaptive virtual frequency modulation control strategy that dynamically adjusts the inertia coefficient to improve frequency stability. Zhao et al. (2022) provides a detailed analysis of the virtual synchronous generator's parameter design, focusing on stability and dynamic performance, and outlines the principles for parameter setting. Several GFM control methods are proposed in the literature (Zhong et al., 2014; Rodriguez et al., 2018; Zhang et al., 2010; Matevosyan et al., 2019), comparisons between GFL and GFM converters illustrated in Table 1.

Due to the voltage source behavior of the GFM converters in contrast to GFL converters, the overcurrent protection requires particular attention. Therefore, various current-limiting threshold control methods for GFM converters are reported in the literature, including current threshold limiters, virtual impedance, virtual admittance.

TABLE 1 Comparison between GFL and GFM converters.

	GFL	GFM
PLL	Need	No need
Dynamic performance	fast	slow
Inertia provision	No	Yes
Type of applicable grid	Strong power grid	Weak power grid
FRT	easy	difficult

To control the output current of GFM inverter during faults, paper (Bottrell and Green, 2014) restricts the phase current magnitude with the maximum allowed value through closed-loop current control, authors in (Pirsto et al., 2022) have proposed a state feedback control based cascaded voltage and current loop. When the converter is overloaded, the controller shifts to the current control mode (CCM) from the voltage control mode (VCM) to control the output current. Since the references generated by the voltage controller are unutilized, the voltage source behaviour of the GFM is lost. Also, this necessitates using an anti-windup control for the voltage controller, leading to delays in the recovery period. The current limiting control performance of GFM converters with and without transitioning to GFL with varying grid strengths is compared in (Taul et al., 2020). The comparison have shown that when GFL performed satisfactorily with strong grids, it might lead to unstable behaviour in weak grids because of the stability issues associated with phase locked loop (PLL).

GFM based on virtual impedance control with aim to adjust the impedance $R_v + jX_v$ to limit the phase current magnitude. The virtual impedance with cascaded voltage control proposed in (Lu et al., 2016; Qoria et al., 2019), and without inner-loop control is presented in (Vilathgamuwa et al., 2006; Gouveia et al., 2021). The virtual admittance control method is applied in (Rosso et al., 2020; Rosso et al., 2021) for current limitation, the stabilities of admittance

TABLE 2 Comparisons of existing current-limiting control methods.

References	GFM mode	Enhancement of transient stability	During fault current limiting	Temporary current limiting
Control-mode-switching (Pirsto et al., 2022; Taul et al., 2020)	✗	✓	✓	✗
Current limiter (Bottrell and Green, 2014)	✓	✗	✓	✗
virtual impedance With no inner loop (Vilathgamuwa et al., 2006; Gouveia et al., 2021)	✓	✓	✗	✗
virtual impedance With cascaded voltage control (Lu et al., 2016; Qoria et al., 2019)	✓	✗	✓	✗
virtual admittance With inner current control (Rosso et al., 2020; Rosso et al., 2021)	✓	✓	✓	✗

method and impedance method are compared initially in paper (Huang et al., 2021), resulted that the equivalent circuits of these two methods are basically the same. With the use of conventional cascaded PI control, the bandwidth of the loop is limited owing to sluggish behaviour, and leads to temporary overcurrents or overvoltages when fault is initiated or recovery period. Comparisons of different current limiting methods illustrated in Table 2.

Conventionally, a cascaded dual-loop linear feedback control is deployed in the inner loop of GFM. However, this control scheme PI parameter setting is complex, suffers from a slow transient response, limited control bandwidth, and instantaneous current overlimits. Moreover, it is hard for a linear controller to handle the multi-objective optimization and various system constraints, for example, MMC (Vazquez et al., 2017).

A improved GFM strategy based virtual admittance combined with direct modulation MPC method is proposed in this work. The improved control strategy has simplified structure, fast dynamic response speed and strong current limiting ability.

The main contributions of the work are:

- The proposed scheme combines an virtual admittance reference current and currentlimit objective into the MPC cost functions, which enhanced fault ride-through of GFM-MMC converter.
- Without any transient overcurrents and overvoltages during fault initiation or recovery after fault clear, compared with PI control strategy, MPC method can provide faster active power recovery.
- Verified that the proposed control strategy can run stably under wide SCR, the power quality and dynamic response speed are better than the PI control.

In the parts to follow: Section 2 details the establishment of a mathematical model for MMC converters. The controller design for the MPC-VSG MMC are discussed in Section 3. Section 4 shows verification of these concepts through simulations include current

limitation, dynamic response, and power quality. Finally conclusions of the work are summarised in Section 5.

2 Mathematical model of grid-connect MMC

The MMC consists of a three-phase structure with six bridge arms, where each bridge arm has an inductance L_{arm} , and resistance R_{arm} , with N half-bridge submodules connected in series. The topology of the MMC is illustrated in Figure 2.

As shown in Figure 2, u_{sj} and i_{sj} ($j = a, b, c$) are the phase voltage and phase current of the MMC grid side, respectively. u_{pj} and u_{nj} are the upper and lower bridge arm voltages, respectively. i_{pj} and i_{nj} are the current of the upper and lower bridge arms, respectively. L_g and R_g are the equivalent inductance and resistance of the grid sides, respectively. V_{dc} and I_{dc} are the DC voltage and DC current, respectively. MMC circuit equation is established based on Kirchhoff's voltage law.

$$-\frac{V_{dc}}{2} + u_{nj} + R_{arm}i_{nj} + L_{arm}\frac{di_{nj}}{dt} + R_gi_{sj} + L_g\frac{di_{sj}}{dt} = u_{sj} \quad (1)$$

$$\frac{V_{dc}}{2} - u_{pj} - R_{arm}i_{pj} - L_{arm}\frac{di_{pj}}{dt} + R_gi_{sj} + L_g\frac{di_{sj}}{dt} = u_{sj} \quad (2)$$

Add and subtract Equation 1 and Equation 2, respectively.

$$\frac{di_{sj}}{dt} = -\left(\frac{R_{arm} + 2R_g}{L_{arm} + 2L_g}\right)i_{sj} + \frac{u_{nj} - u_{pj}}{L_{arm} + 2L_g} - \frac{2}{L_{arm} + 2L_g}u_{sj} \quad (3)$$

$$\frac{di_{zj}}{dt} = -\frac{R_{arm}}{L_{arm}}i_{zj} - \frac{u_{nj} + u_{pj}}{2L_{arm}} + \frac{1}{2L_{arm}}V_{dc} \quad (4)$$

where, $i_{zj} = \frac{i_{pj} + i_{nj}}{2}$, $i_{sj} = i_{nj} - i_{pj}$.

$$u_{pj} = \frac{n_{pj}u_{pj}^{\Sigma}}{N} \quad (5)$$

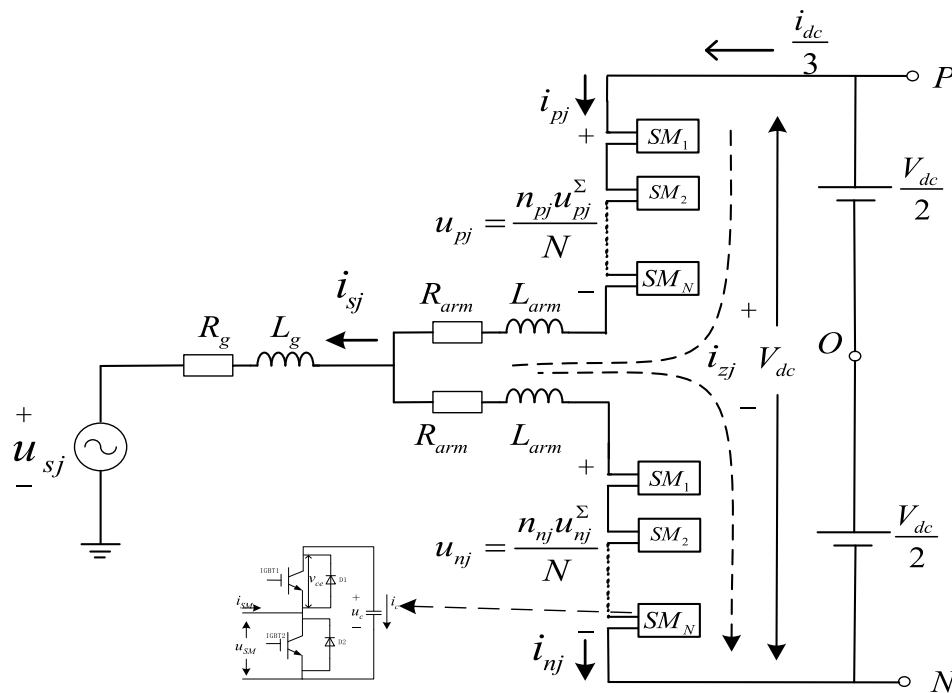


FIGURE 2
Topology of grid-connect MMC.

$$u_{nj} = \frac{n_{nj} u_{nj}^{\Sigma}}{N} \quad (6)$$

where n_{pj} and n_{nj} are the upper and lower bridge arm submodules, respectively. N is the number of submodules put into each phase. u_{pj}^{Σ} and u_{nj}^{Σ} are the sum of upper and lower bridge arm capacitor voltages, respectively.

By Equations 3, 4, the unbalanced current inside MMC i_{zj} can be controlled by the sum of bridge arm voltage ($u_{nj} + u_{pj}$). The AC current i_{sj} of grid side controlled by the difference bridge arm voltage ($u_{nj} - u_{pj}$). This means that the DC circuit and AC circuit of MMC are decoupling from each other (Figure 3). Using the forward Euler to discretize Equations 3, 4, an MMC discrete mathematical model can be obtained, k is the k -th sampling point and T_s is the sampling frequency.

$$\frac{i_{sj}(k+1) - i_{sj}(k)}{T_s} = - \left(\frac{R_{arm} + 2R_g}{L_{arm} + 2L_g} \right) i_{sj}(k+1) + \frac{u_{nj}(k+1) - u_{pj}(k+1)}{L_{arm} + 2L_g} - \frac{2u_{sj}(k+1)}{L_{arm} + 2L_g} \quad (7)$$

$$\frac{i_{zj}(k+1) - i_{zj}(k)}{T_s} = - \frac{R_{arm}}{L_{arm}} i_{zj}(k+1) - \frac{u_{nj}(k+1) + u_{pj}(k+1)}{2L_{arm}} + \frac{1}{2L_{arm}} V_{dc}(k+1) \quad (8)$$

3 MMC grid-forming control method based on model predictive control

3.1 MMC traditional double closed loop PI control

The MMC typically employs a double-closed-loop PI control, as shown in Figure 4. P_{ref} , V_{dcref} , Q_{ref} , U_{smref} , i_{dref} , i_{qref} , u_{abcref} are the references of active power, reactive power, DC voltage and the PCC AC voltage, d-axis current, q-axis current, modulation of arm voltage, respectively; and P , Q , V_{dc} , U_{sm} , i_{sabc} , u_{sd} , u_{sq} , ω_{ref} are the output active power, reactive power, DC voltage, PCC AC voltage, PCC AC current, d-axis voltage, q-axis voltage, angular frequency, respectively; k_{p1} , k_{i1} , k_{p2} , k_{i2} , are out-loop and inner-loop proportional and integral coefficients.

The powers are calculated in the dq-frame as:

$$\begin{cases} P = 1.5(u_{sd}i_{sd} + u_{sq}i_{sq}) \\ Q = 1.5(u_{sq}i_{sd} - u_{sd}i_{sq}) \end{cases} \quad (9)$$

In the dq reference frame, (Equation 9) expressed as:

$$L_{eq} \frac{d}{dt} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} = -R_{eq} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \begin{bmatrix} u_{diffd} \\ u_{diffq} \end{bmatrix} - \begin{bmatrix} u_{sd} \\ u_{sq} \end{bmatrix} + \begin{bmatrix} \omega L_{eq} \\ -\omega L_{eq} \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} \quad (10)$$

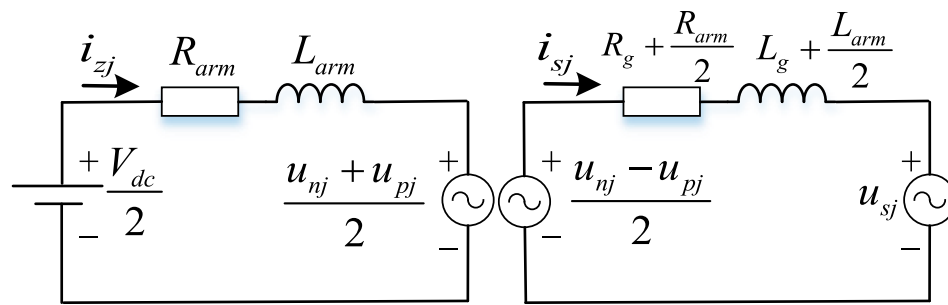


FIGURE 3
MMC AC and DC equivalent circuit.

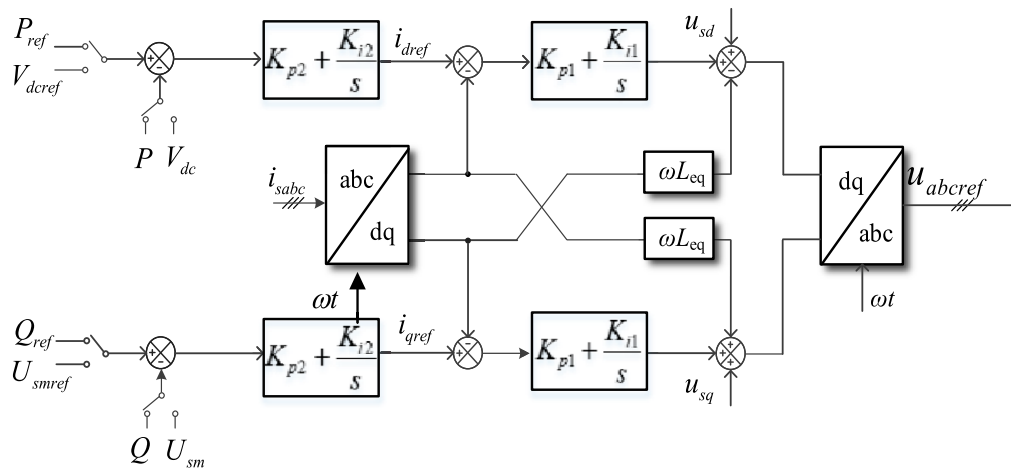


FIGURE 4
Traditional PI double closed loop control.

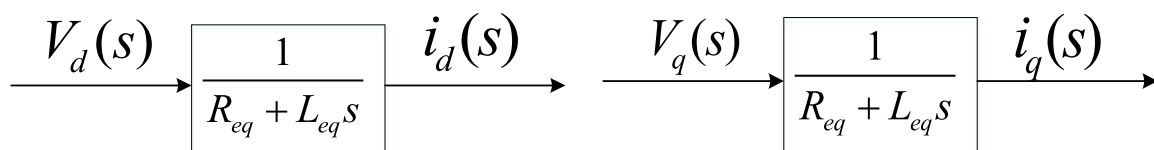


FIGURE 5
The input-output relationship of d-q current.

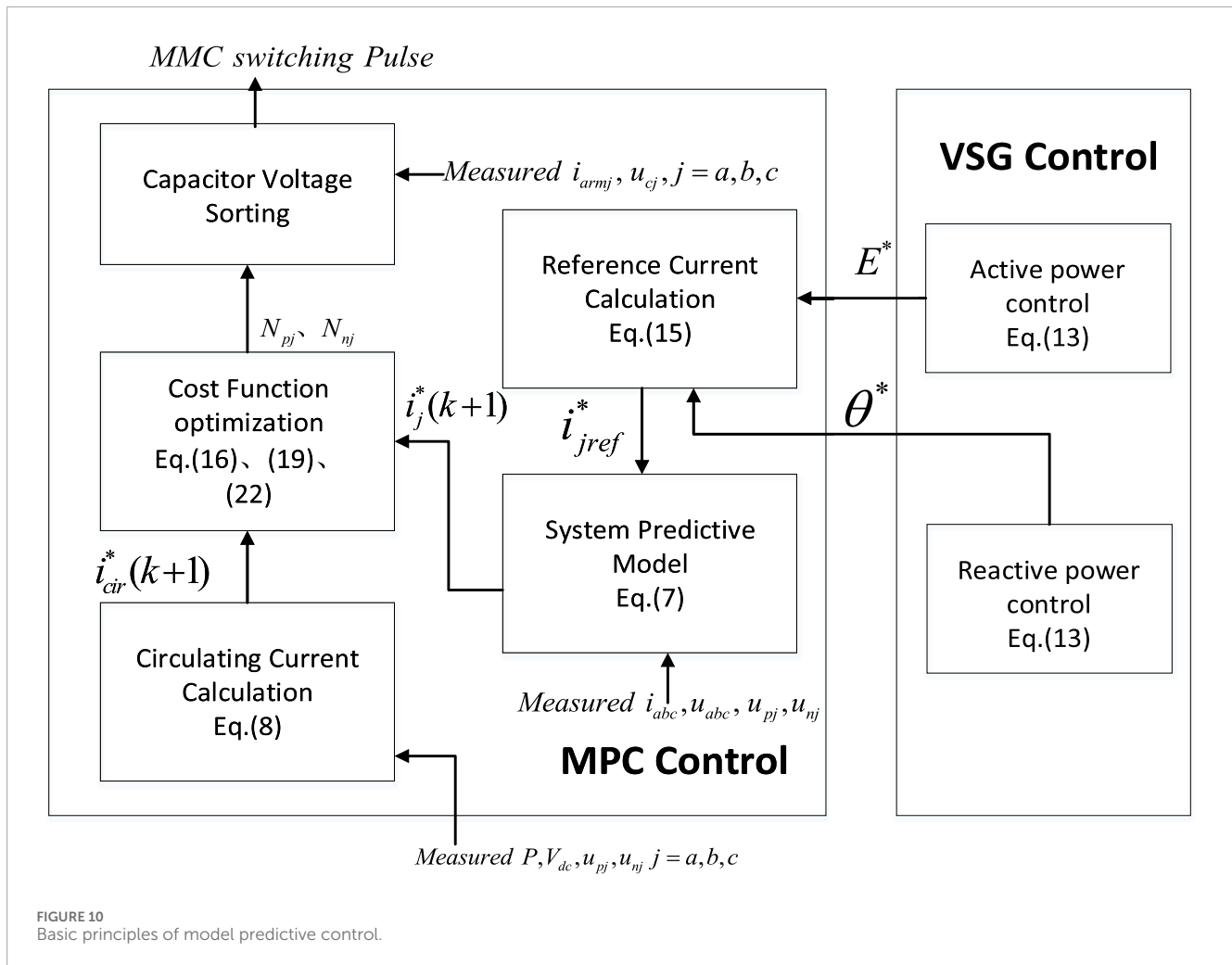
R_{eq} and L_{eq} in (Equation 10) are expressed in $R_{eq} = R_g + \frac{1}{2}R_{arm}$ and $L_{eq} = L_g + \frac{1}{2}L_{arm}$.

In (Equation 10) expressed Laplace domain as:

$$\begin{cases} (R + Ls)i_{sd}(s) = -u_{sd}(s) + u_{diffd}(s) + \omega Li_{sq}(s) \\ (R + Ls)i_{sq}(s) = -u_{sq}(s) + u_{diffq}(s) - \omega Li_{sd}(s) \end{cases} \quad (11)$$

Transfer functions for output variables i_{sd} , i_{sq} and voltages V_d and V_q expressed as:

$$\begin{cases} \frac{i_{sd}(s)}{V_d(s)} = \frac{1}{R_{eq} + L_{eq}s} \\ \frac{i_{sq}(s)}{V_q(s)} = \frac{1}{R_{eq} + L_{eq}s} \end{cases} \quad (12)$$



3.2 Grid-forming virtual synchronization control

Traditional MMC double-PI control typically uses fixed reference values for its targets, making it incapable of actively adjusting to variations in grid voltage and frequency. The core concept of grid-forming control is to modify the converter control mode, enabling it to exhibit rotational inertia and autonomously provide active or reactive power support to the grid. The mathematical equations governing the grid-forming virtual synchronous generator (VSG) control for the MMC are presented as follows.

$$P^* - P = J\omega_0 \frac{d\omega}{dt} + D_p\omega_0(\omega - \omega_0) \quad (13)$$

$$E = E_n + k_q(Q^* - Q) + k_v(U^* - U) \quad (14)$$

$$L_v \frac{di_{jref}}{dt} + R_v i_{jref} = e_{jref} - u_j \quad (15)$$

Equations 13–15 describe the rotor motion equation, the reactive power-voltage control equation, and the electrical equation of the VSG body used for virtual synchronous control.

In these equations, P^* and P are the reference power and actual active power, respectively. J and D_p are the VSG control parameters refer to virtual rotational inertia and damping coefficient respectively, ω_0 and ω refer to the rated velocity and actual electrical angular velocity, respectively, Q^* and Q are the reference power and actual active power, respectively. E_n is the no-load electromotive force in the synchronous generator. R_v and L_v are the virtual resistance and inductance of VSG, respectively. i_{jref} is the three-phase current reference value of VSG output. e_{jref} is the internal potential reference value. u_j is the three-phase voltage of the MMC grid connection point.

The control block diagram for the VSG in the MMC is shown in Figure 6. The stator voltage is determined using P - f control and Q - V control, with virtual impedance R_v and L_v incorporated into the electrical section of the VSG model. The current command value i_{jref} , which satisfies the stator voltage equation, is computed through the virtual impedance loop. The d-q current references, i_{dref} and i_{qref} , are then used as inputs to the current inner loop in Figure 4, completing the simulation of the stator voltage equation in VSG control.

By adjusting the values of J and D_p , different active power response characteristics of the MMC can be achieved. The dynamic response of the output power for the MMC, with different J and D_p coefficients, under VSG control is shown in Figures 7, 8.

TABLE 3 Simulation parameters.

Type	Parameters
Rated power of the MMC/MW	10
Rated voltage of the AC grid/kV	10
Dc side voltage/kV	±10
Transformer leakage/mH	5
Grid side equivalent resistance/Ω	0.01
Number of submodules of the bridge arm	20
Arm inductance/mH	3
Arm resistance/Ω	1
Submodule capacitance value/mF	10
$J/(\text{kg}\cdot\text{m}^2)$	300
$D_p/(\text{N}\cdot\text{m}\cdot\text{s}/\text{rad})$	2000
Reactive power adjustment coefficient kq	0.005
Voltage regulation coefficient kv	1
Control period fs/kHz	10
Number of submodules m for circulation suppression	1

As illustrated in Figures 7, 8, increasing the virtual inertia J results in a larger amplitude and a longer response time to the P/Q reference signals. Conversely, increasing the virtual damping coefficient D_p leads to a smaller amplitude and a shorter response time. In practice, D_p is generally determined by grid standards, and the dynamic performance of the MMC is primarily adjusted by modifying J .

3.3 Grid-forming based on model predictive control

Model predictive control (MPC) is an advanced control method for power electronics, based on predicting the next AC grid-side current by establishing a discrete mathematical model of the converter. It then performs a traversal calculation of all possible switching states of the power electronics, ultimately achieving optimal control within each control period.

Unlike traditional control methods, MPC does not rely on a PI control loop, making it structurally simpler, requiring no complex parameter tuning. MPC offers fast tracking responses, minimal overshoot, and superior dynamic performance (Zheng, 2020). In this paper, MPC is applied to the virtual synchronous control of the MMC, and a corresponding cost function is designed to address circulating currents and DC voltage fluctuations, as illustrated in Figure 9.

The power control outer loop of grid-forming control is implemented as MPC-VSG, where the traditional PI current loop is replaced by the MPC model predictive controller. In this setup, the current reference $i_{j\text{ref}}$ of the VSG output (as shown in Figure 6) serves as the input reference for the model predictive control. The discrete mathematical Equations 7, 8, of the MMC are combined to construct the cost function for MPC.

Depending on disturbances in the AC grid frequency or voltage, or input changes in the active power command (P_{ref} or V_{dc}) or reactive power command (Q_{ref} or U_{sm}), the three-phase current reference $i_{sj\text{ref}}(t + T_s)$ ($j = a, b, c$) is calculated using Equations 13–15.

The first cost function of MPC-VSG is built.

$$J_j = |i_{j\text{ref}}(t + T_s) - i_{sj}(t + T_s)| \quad (16)$$

Combined with (Equation 7), the upper and lower bridge arms voltage difference value $e_j(t + T_s) = u_{nj}(k + 1) - u_{pj}(k + 1)$ corresponding to the minimum J_j can be denoted as $e_j^*(t + T_s)$, whose value range is $\left[\frac{V_{\text{dc}}}{N}\left[-\frac{N}{2}, -\frac{N-1}{2}, \dots, 0, \dots, -\frac{N-1}{2}, \frac{N}{2}\right]\right]$, and V_{dc} is the DC bus voltage. According to the combined Equations 5, 6, the number of upper and lower bridge arms can be calculated, as n_{nj}^* and n_{pj}^* .

Equation 8 highlights the presence of circulating currents within the MMC, which, while circulating only inside the MMC, do not affect the external AC or DC sides. However, these currents contribute to increased power device losses and must be suppressed. In this work, we address this issue by controlling the circulating currents through adjustments to the additional voltage v_{diff} applied simultaneously by Equations 17, 18:

$$\begin{aligned} \frac{i_{sj}(k+1) - i_{sj}(k)}{T_s} = & -\frac{i_{sj}(k+1)R_{\text{arm}}}{L_{\text{arm}}} \\ & -\frac{[u_{nj}(k+1) + v_{\text{diff}}] + [u_{pj}(k+1) + v_{\text{diff}}]}{2L_{\text{arm}}} + \frac{V_{\text{dc}}(k+1)}{2L_{\text{arm}}} \end{aligned} \quad (17)$$

$$v_{\text{diffj}} = [V_{\text{dc}}/N][-m, -(m-1), \dots, 0, m-1, m] \quad (18)$$

The second cost function of MPC-VSG is obtained.

$$J_{\text{diffj}} = |i_{\text{dc}}(t + T_s)/3 - i_{zj}(t + T_s)| \quad (19)$$

where i_{dc} is the DC side current. When J_{diffj} is the lowest value, the corresponding v_{diff} value is calculated and m denoted as $v_{\text{diffj}}^*(t + T_s)$ and m^* . The upper bridge arm submodules number under the MPC-VSG control strategy is obtained by combining cost function (Equations 16, 19).

$$N_{pj} = n_{pj}^* + m^* \quad (20)$$

The lower number under the MPC-VSG control strategy is obtained.

$$N_{nj} = n_{nj}^* + m^* \quad (21)$$

Submodule voltage balancing is achieved using the nearest-level approximation method. By combining Equations 20, 21, the submodule sequence for each arm is determined, and the corresponding control pulses are applied to the MMC.

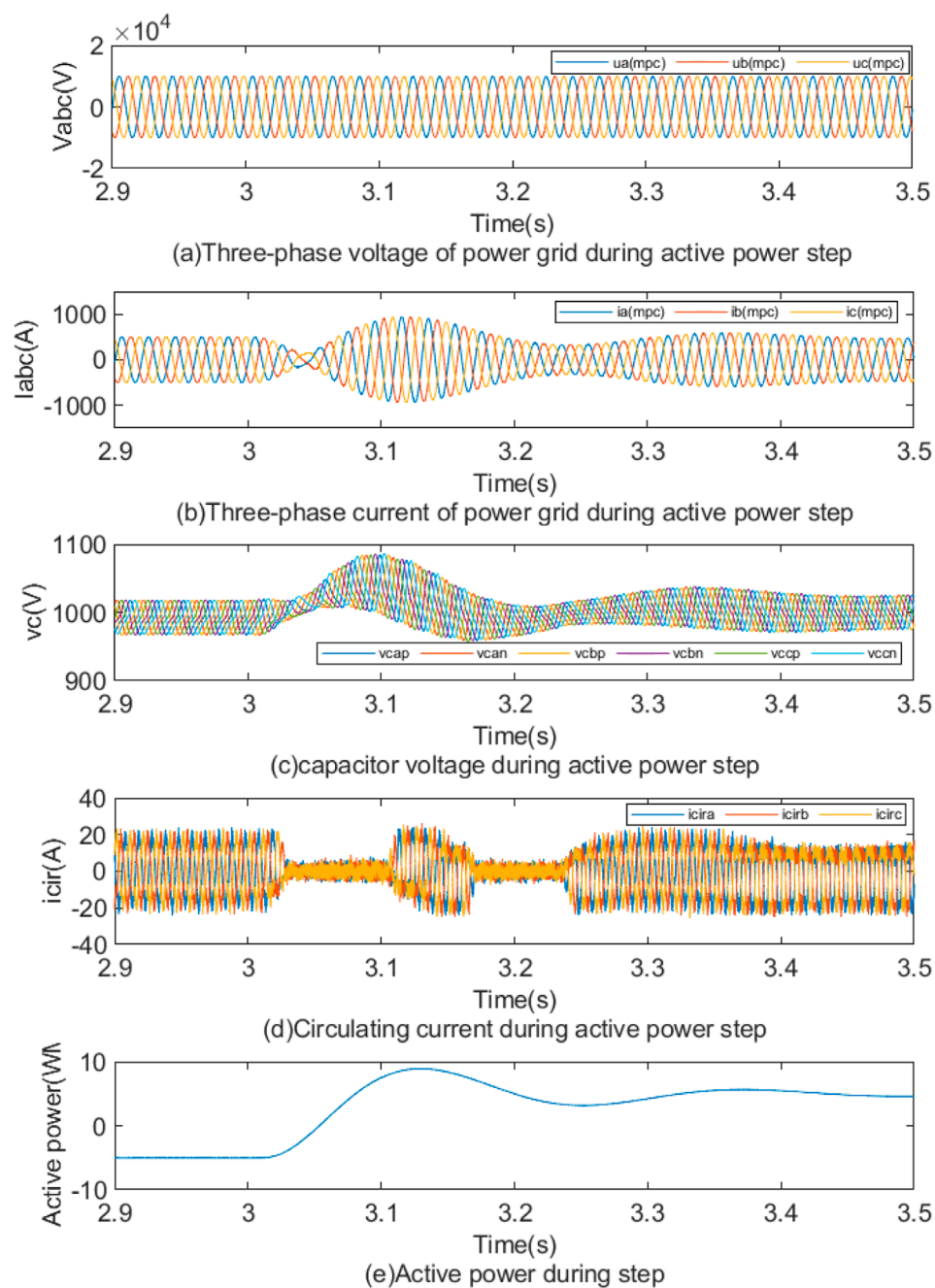


FIGURE 11

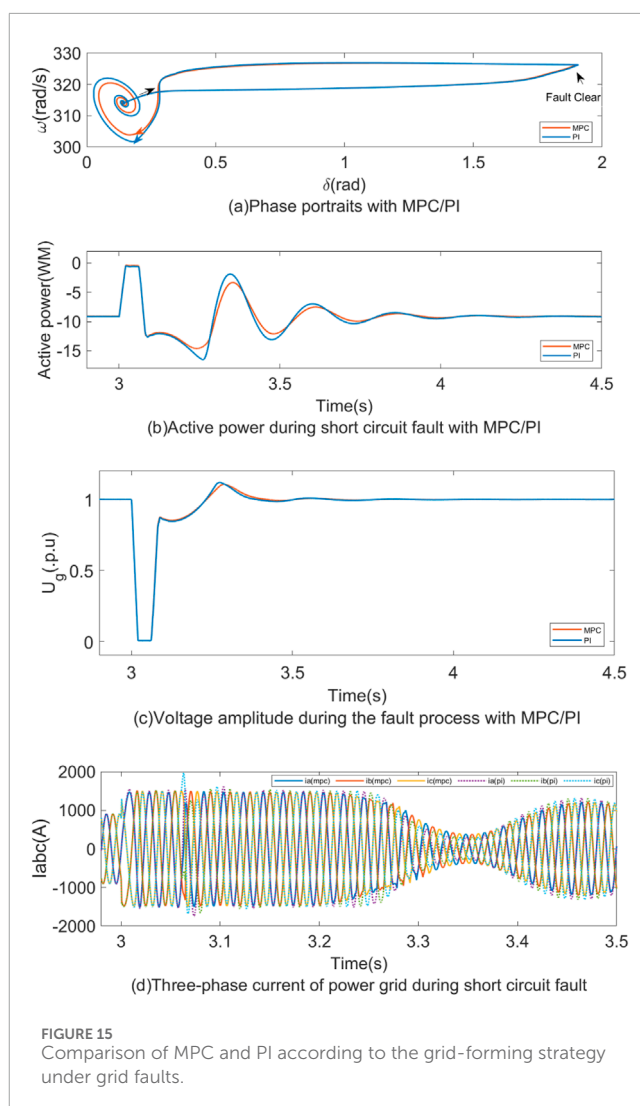
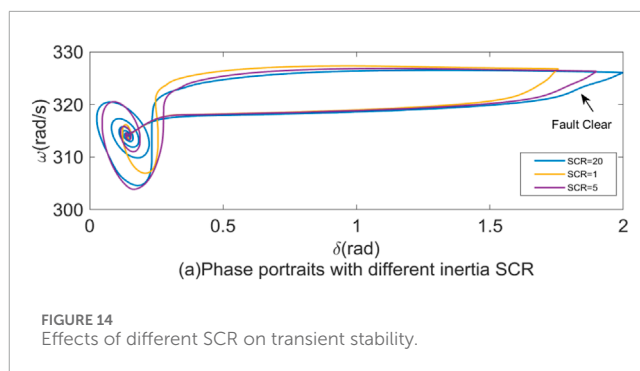
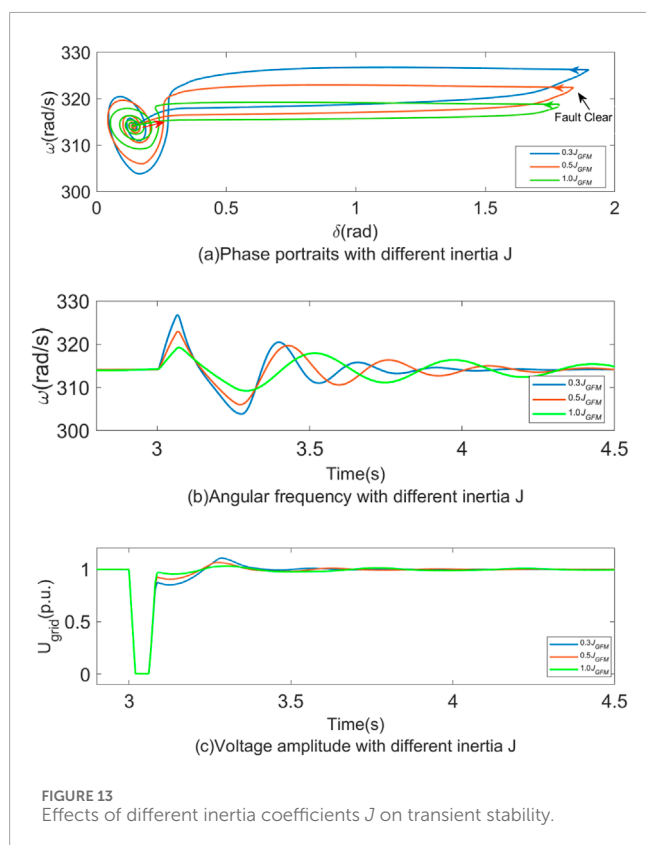
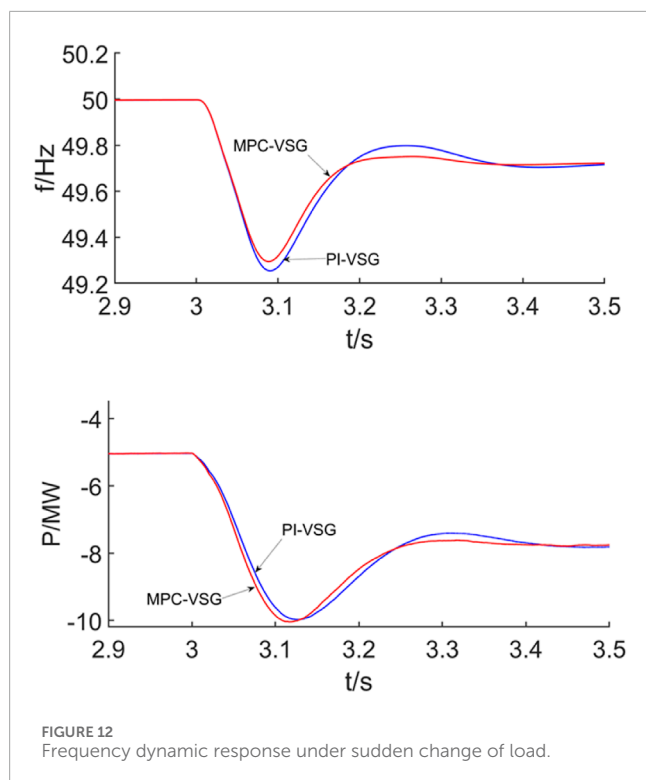
(A) Three-phase voltage of power grid during active power step. (B) Three-phase current of power grid during active power step. (C) capacitor voltage during active power step. (D) Circulating current during active power step. (E) Active power during step.

Three control objectives are incorporated into the proposed Level MPC these are AC current, circulating current, and strict fault current limit. control objectives are achieved through the cost function J outlined in Equation 22.

$$J = (\lambda_1 + \xi) |i_{jref}(t + T_s) - i_{sj}(t + T_s)| + \lambda_2 |i_{dc}(t + T_s)/3 - i_{zj}(t + T_s)| \quad (22)$$

$$\xi = \begin{cases} \infty, & \|i_{sj}\| \geq I_{\max} \\ 0, & \|i_{sj}\| < I_{\max} \end{cases} \quad (23)$$

Within the cost function, λ_1 λ_2 introduces the weighting factor, in order to tracking of reference and regulate the circulating current to its desired value, ξ is additional penalty weighting factor to the fault current strict limit in Equation 24.



3.4 Threshold current controller design of grid-forming

To mitigate transient overcurrents, the proposed method limits the amplitude of the reference current under the d-q framework while maintaining the current phase. This ensures that power supply quality is preserved during fault conditions to the greatest extent possible. The strategy is described as follows:

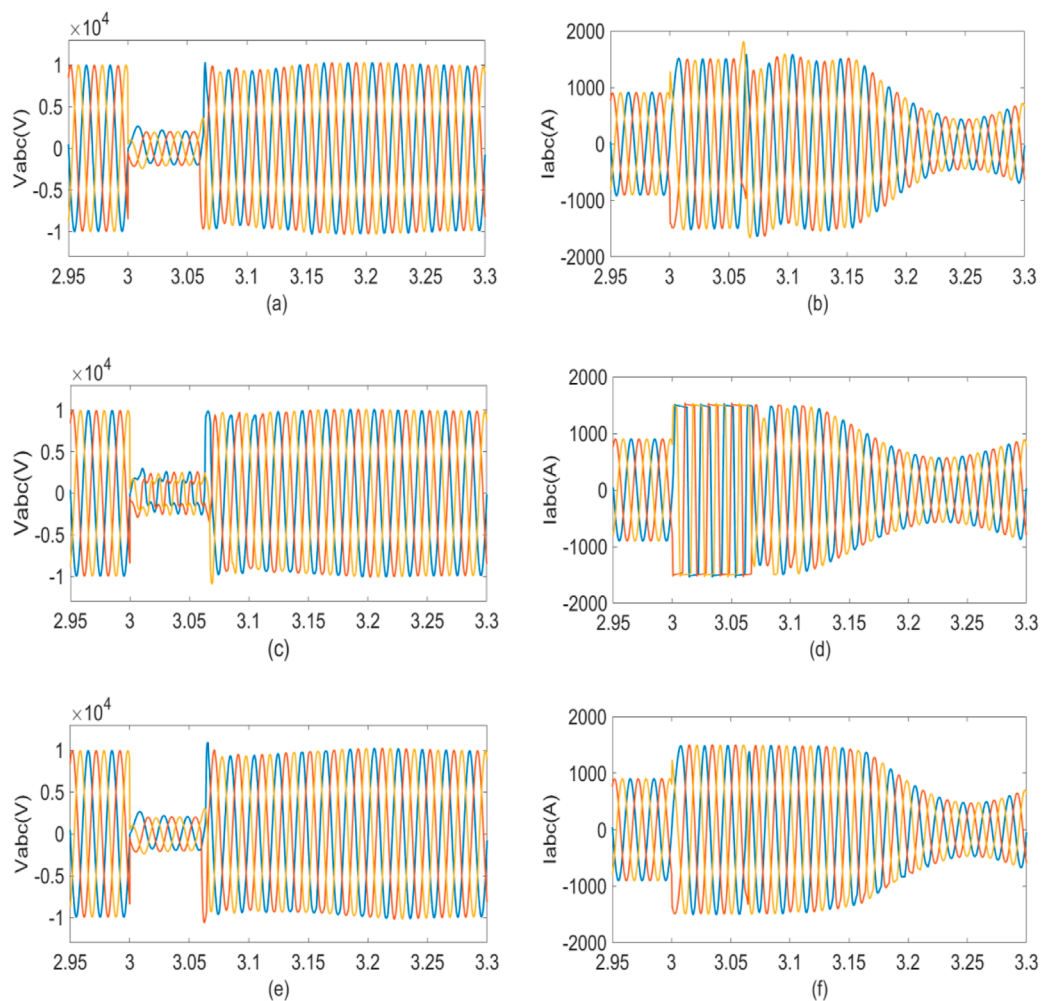


FIGURE 16

Three-phase voltage and current in (A, B) the PI-based strategy, (C, D) the MPC-based strategy in (Matevosyan et al., 2019), and (E, F) the proposed MPC strategy.

where I_{max} denotes the maximum allowable current of the converter. $i_{sd_lim}^*$ denotes the saturated d-q current. i_{sq}^* , i_{sd}^* , $i_{sd_lim}^*$, and $i_{sq_lim}^*$ denote the reference current of axis d and q before and after the limiting respectively.

$$\begin{cases} i_{sd_lim}^* = i_{sd}^* & \sqrt{i_{sd}^{*2} + i_{sq}^{*2}} < I_{max} \\ i_{sd_lim}^* = I_{max} \cos \left[\arctan 2 \left(\frac{i_{sq}^*}{i_{sd}^*} \right) \right] & \sqrt{i_{sd}^{*2} + i_{sq}^{*2}} \geq I_{max} \\ i_{sq_lim}^* = I_{max} \sin \left[\arctan 2 \left(\frac{i_{sq}^*}{i_{sd}^*} \right) \right] & \sqrt{i_{sd}^{*2} + i_{sq}^{*2}} \geq I_{max} \end{cases} \quad (24)$$

The basic principles of model predictive control refer to Figure 10.

4 Simulation verification

To validate the proposed grid-forming control strategy, a grid-forming MMC system was built and tested in Simulink. For the purpose of analysis and comparison, the DC side of the grid-connected MMC is connected to a constant DC voltage source, while

the AC side uses a three-phase programmable voltage source to simulate frequency and voltage fluctuations. The system simulation parameters are provided in Table 3.

4.1 Simulation of dynamic characteristics of MMC

The active power step simulation of the MMC converter is conducted under a grid-forming control strategy. The simulation parameters for the MMC converter and system are provided in Table 3. The number of submodules, denoted as “m,” used to suppress circulating current is set to 1. The time-domain simulation results are presented in Figure 11. Before $t = 3$ s, MMC operates statically at -5 MW, with a circulating current of approximately 20 A. At $t = 3$ s, the active power jumps to 5 MW, with the circulating current remaining around 20 A. During this transition, the capacitor voltage fluctuates by about 6% due to the circulating current. The MMC converter operates stably throughout the step process, fulfilling the specification requirements.

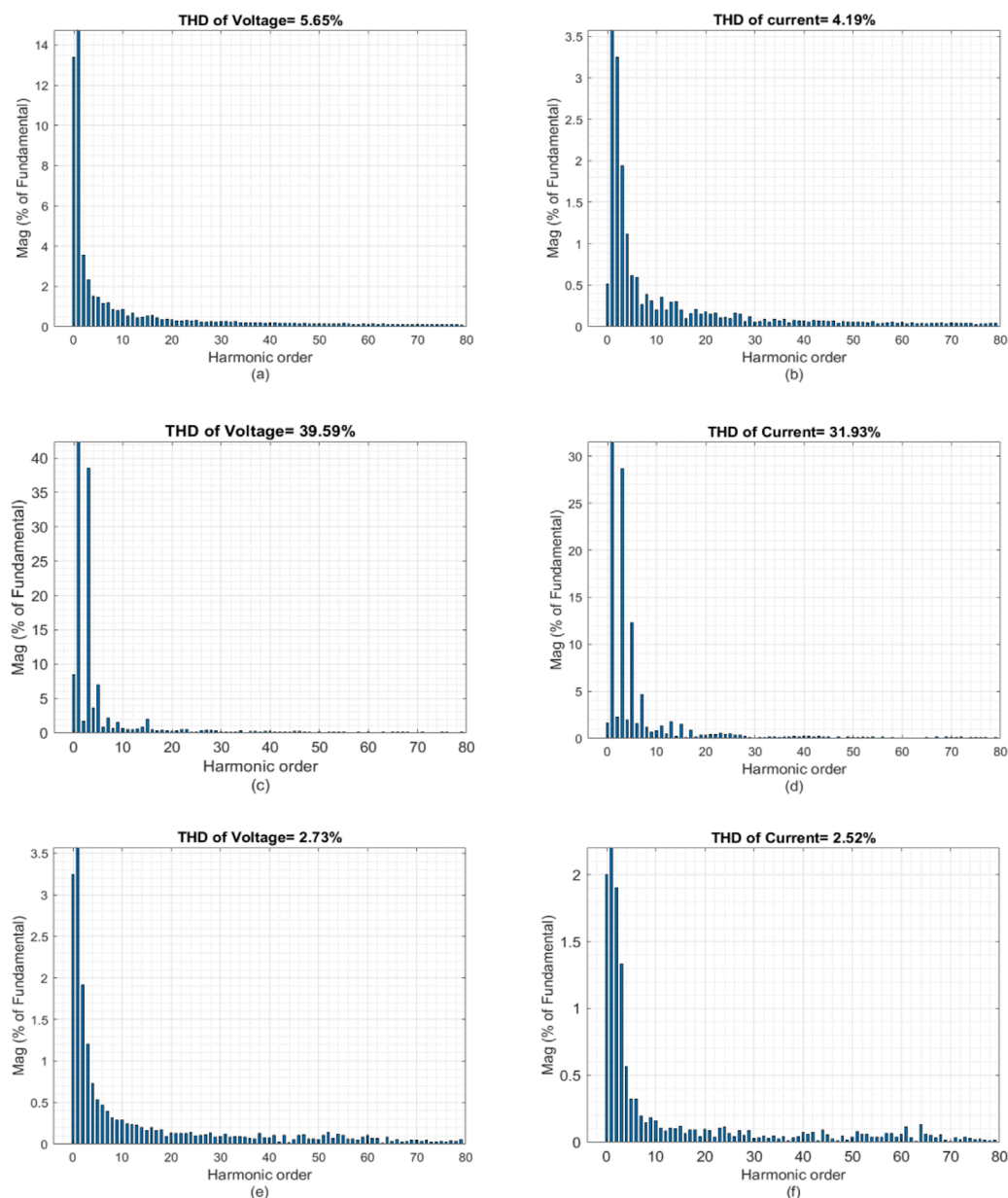


FIGURE 17
THD of voltage and current in (A, B) the PI-based strategy, (C, D) the MPC-based strategy in (Matevosyan et al., 2019), and (E, F) the proposed MPC strategy.

To fully compare the dynamic response characteristics of the inertia support with the proposed control strategy, the simulation experiment under sudden change of load is carried out. During normal operation, the synchronous generator is 20 MVA, the load is 20 MW, and MMC-HVDC provided 5 MW power to the load. At 3 s, the 7 MW load is suddenly put into, with the frequency transient response characteristics under the two controls observed. As shown in Figure 12, at $t = 3$ s, based on PI-VSG control, the grid frequency f drops to 49.25 Hz. However, based on MPC-VSG control, the frequency occurs 49.29 Hz, which reduced by 0.04 Hz.

When the system frequency decreases, both strategies can achieve the rapid response adjustment of active power. Compared

with PI-VSG control, MMC under MPC-VSG control has a faster power tracking speed.

4.2 Grid-forming MMC transient characteristics under grid faults

4.2.1 Effects of different inertia coefficients J and SCR on transient stability

A three-phase short-circuit fault is simulated in the power grid, with the fault location at the converter outlet and a fault resistance of 0.01Ω . The fault occurs at $t = 3$ s and lasts for 60 m, with a

system short-circuit ratio (SCR) of 3. The reference value for the inertia coefficient J is set to 300. The transient fault characteristics are examined for inertia coefficients of 0.3, 0.5, and 1.0 p.u.

Figure 13 illustrates the transient fault characteristics of grid-forming MMC converters under different inertia coefficients J during grid faults. Figure 10A shows the phase portraits of MMC converters during the fault process. It is evident that a smaller inertia coefficient J_{GFM} leads to a larger power angle δ , making the converter more prone to instability. Figures 13B, C demonstrate that increasing the inertia coefficient J_{GFM} enhances the system's ability to suppress frequency ω fluctuations, accelerates voltage recovery, and improves overall system stability.

Figure 14 illustrates the transient fault characteristics of grid-forming MMC converters under different SCR during grid faults. The result shows that when increasing the grid SCR, the power angle δ become larger, that means the stability margin decreases when the grid is stiff.

4.2.2 Comparison of MPC and PI with grid-forming strategy under grid faults

The fault simulation conditions are consistent with those outlined in Section 4.2.1, with the fault current limited to 1.5 p.u. Figure 15 presents the phase diagram and fault current limiting waveform for grid-forming MMC converters under both MPC and PI control strategies. Upon fault clearing, the power angle and active power under the MPC strategy recover to their pre-fault conditions more rapidly than under the PI strategy. As shown in Figure 11C, the grid voltage increases less during fault recovery with the MPC strategy. Furthermore, Figure 11D clearly illustrates that under the PI strategy, the C-phase current reaches 2 kA (2 p.u.), while the current in all three phases under the MPC strategy remains within the threshold.

4.2.3 Power quality assessment

To evaluate the power quality of the output voltage and current under the proposed strategy, the fault simulation conditions are kept consistent with those described in Section 4.2.1, with a fault current limit of 1.5 p. u. and a fault resistance of 0.5 Ω .

The proposed control strategy is compared with the PI strategy and the MPC current limiting strategy from reference (Zheng, 2020). The voltage and current waveforms are shown in Figure 16, and the THD of both voltage and current under the three strategies are compared in Figure 17. As seen in Figures 16E, F, 17E, F, the proposed control strategy maintains excellent power quality during both steady-state and fault conditions, with voltage and current THD of 2.73% and 2.52%, respectively, and accurately limits the current below 1.5 p. u. In contrast, under the PI strategy, current overruns are observed during the fault, with voltage and current THD of 5.65% and 4.19%, respectively. Under the MPC current limiting strategy from reference (Zheng, 2020), the current is strictly limited to less than 1.5 p.u., but the waveform is clipped, resulting in poor power quality with voltage and current THD of 39.59% and 31.93%, respectively.

5 Conclusion

Grid forming converters have become attractive candidates in the high penetration of distributed energy resources grid. This paper first compared existing GFL and GFM strategies, a improved GFM-MMC strategy based virtual admittance combined with direct modulation MPC method is proposed, which obtained faster and strict current-limiting capability, improved dynamic response, and power quality. Finally, the verification of these concepts through simulations. This strategy aims to enhance the stability of the power supply system and improve the fault tolerance of the grid, offering a novel approach to carbon emission reduction in DC industrial parks.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

Author contributions

SL: Methodology, Writing–review and editing. XD: Conceptualization, Methodology, Writing–review and editing. YX: Supervision, Writing–review and editing. YL: Conceptualization, Writing–review and editing. RZ: Supervision, Writing–review and editing.

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Conflict of interest

Authors SL, YX, YL, and RZ were employed by State Grid Ningxia Electric Power Co., Ltd Econ-tech Research Institute. Author XD was employed by Beijing DC T&D Engineering Technology Research Center (NARI China-EPRI Electrical Engineering Co., Ltd.).

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A ground current suppression method for systems with a large number of photovoltaic (PV) inverters

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In large PV plants, a large number of PV inverters are linked together at grid-side to generate more power. However, as the number of PV inverter modules increases, the ground current problem worsens. This can result in a ground current fault, and cause the inverter failed and disconnected from the grid. In this paper, the solutions, including hardware and software, are proposed to suppress the ground current. The hardware solution is to connect filter capacitors back to the DC-bus, directing more ground current to the inverter rather than the grid to reduce the ground current. Additionally, the software solutions, including carrier phase shifting and carrier frequency shifting, are proposed. In the software solutions, the ground current for PV inverters can be cancelled each other, which can alleviate the ground current problem. In the paper, the system architecture is presented first. The concept of connecting the neutral point of the filter capacitors to the DC-bus midpoint is explained. The zero-sequence mathematical model is then built. Moreover, the effect of the zero-sequence voltages of the DC/AC converter and the grid on the zero-sequence current is investigated. Then, the hardware solution is explained. The control architecture is presented, followed by an explanation of how to obtain the three-phase output voltage and zero-sequence components for PV inverters. Two software solutions, including carrier phase shifting and carrier frequency shifting, are detailed. Finally, the experiment is conducted to verify the theory.

KEYWORDS

PV inverters, ground current suppression, zero-sequence, carrier phase shifting, grid-connected

1 Introduction

With the increasing severity of environmental pollution and energy crisis, developing and utilizing renewable energy has become a top priority (Zhang et al., 2023a). As a result, photovoltaic generation has received a lot of attention in recent years, and grid-connected inverters are widely used for energy conversion (Xu et al., 2023). To maximize power generation efficiency in PV systems, MPPT technology is widely used to track maximum power point of PV strings (Ranjan Satpathy et al., 2024a)- (Ranjan Satpathy et al., 2024b). However, when PV modules are blocked to varying degrees, the power curve will have

multiple peaks; thus, numerous improved MPPT algorithms are proposed to improve system power tracking efficiency. At the same time, the topology of grid-connected inverters has direct impact on energy conversion efficiency. Different topology structures are proposed to improve switching losses, EMI suppression, common mode voltage, and so on (Azizi et al., 2025). Grid-connected inverters can be divided into two types based on their topology: isolated and non-isolated. Transformerless non-isolated grid-connected inverters are ideal for photovoltaic grid-connected systems due to their small size, low costs, and simple structure, as well as their high efficiency (Azizi et al., 2025)- (Sharma et al., 2023). As a result, an increasing number of PV grid-connected inverters are adopting non-isolated topologies. However, in the absence of the electrical isolation function, the system can generate a large ground current due to inverter and PV parasitic capacitors connected to the ground. Ground current can increase system losses while also posing safety and electromagnetic compatibility concerns. In recent years, academics have conducted extensive research on ground current suppression technology (Anand et al., 2023)- (Gyawali et al., 2024). The main idea is to create a new freewheeling path between the photovoltaic and AC grid sides. As a result, the common-mode voltage is held constant. The ground current can be reduced when the common-mode voltage remains constant. The following are some examples of typical topology modifications based on this concept.

For single-phase full-bridge topology, there are numerous modifications. H5 topology is one of them (Xiao et al., 2011). The H5 topology, as shown in Figure 1a, adds an additional switch S_{DC} in series at the DC input. The additional switch allows for the decoupling of the AC and DC sides, which suppresses

ground current. However, because S_{DC} is always operated at high frequencies, it withstands high switching losses. There is another H5 variant (Tofigh Azary et al., 2018), which is shown in Figure 1b. In comparison to H5 topology, the switch S_{DC} is divided into two (S_{DC1} and S_{DC2}). As a result, the original S_{DC} loss can be shared by S_{DC1} and S_{DC2} . This can make heat dissipation design less difficult. Another typical topology modification is Highly Efficient Reliable Inverter Concept (HERIC) (Gonzalez et al., 2007)- (Xiao et al., 2014). On the AC side, the HERIC topology adds a freewheeling circuit consisting of two switches (S_{ac1} , S_{ac2}), as shown in Figure 1c. In comparison to H5, the number of conducting devices is always two, resulting in a lower conduction loss. As a result, the device loss distribution is more balanced, potentially extending the device lifetime. The main disadvantage is that the costs are relatively high. Figure 1d depicts the topology of H6 (Ji et al., 2013)- (Liu et al., 2017). Based on the traditional full-bridge topology, this topology adds switches S_5 and S_6 as well as diodes D_1 and D_2 . These additional switches provide a freewheeling path. Then, the DC side can be disconnected from the grid to suppress common-mode voltage. The benefit of this topology is that the voltage stress of the devices is reduced to 1/2 of the DC-bus voltage, effectively reducing switching loss. The main disadvantage remains the additional costs.

For the modulation, unipolar, bipolar, and frequency doubling unipolar are typically used in single-phase full-bridge topology (Mohon et al., 1995)- (Holmes and Lipo, 2003). The common-mode voltage for unipolar and frequency doubling unipolar modulations varies between 0 and V_{dc} . Contrarily, with bipolar modulation, the common-mode voltage is a constant of $V_{dc}/2$ and the ground current is 0. Therefore, the bipolar modulation is frequently

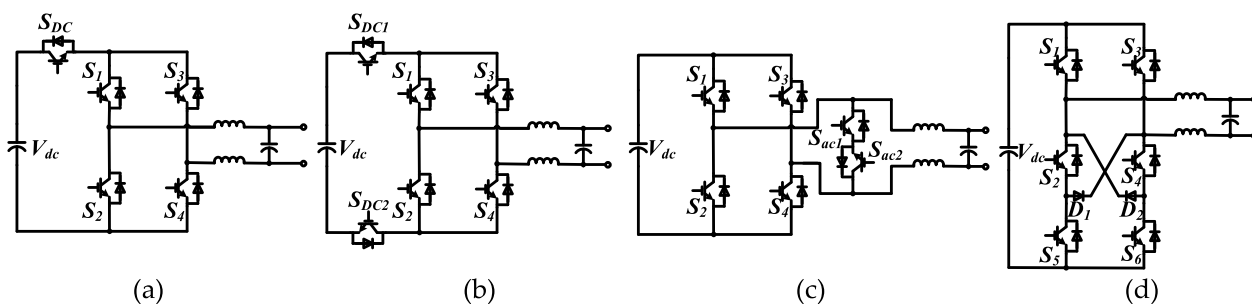


FIGURE 1

Full-bridge topology modifications. (a) H5 topology; (b) H5 variant; (c) HERIC topology; (d) H6 topology.

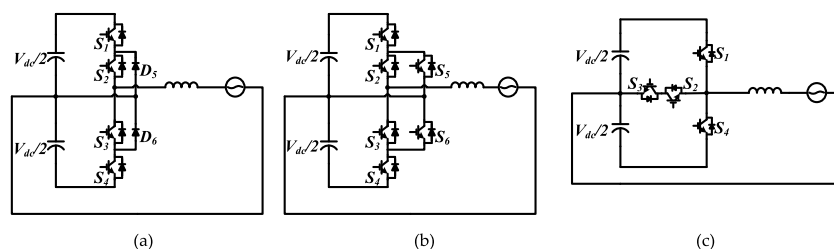


FIGURE 2

Half-bridge topology, (a) NPC topology; (b) ANPC topology; (c) T-type.

employed in single-phase full-bridge topology in order to suppress ground current.

For half-bridge topologies or three phase topologies with three half-bridges, a common method to suppress ground current is to connect the grid to the DC-bus's midpoint. Because the DC-bus capacitors are relatively large, they can provide a low impedance path for the ground current. This allows the majority of the ground current to flow into the DC-bus. As a result, the ground current flowing into the earth can be greatly reduced. In high-power applications, multi-level inverters are frequently used to improve output waveform quality, reduce switching stress, and lower electromagnetic interference levels (Azizi et al., 2025). For typical 1500V PV systems, 3-level topologies are widely adopted. The three-level inverters differ from traditional two-level inverters in that it has an additional intermediate voltage level known as zero level. Figure 2a presents the neutral-point clamped (NPC) topology. The zero level is connected to the midpoint of the DC bus via a diode, improving output voltage control and system power quality (Nabae et al., 1981)- (Barzegarkhoo et al., 2023). It is known that the outer switches (S_1 and S_4) operate at a high frequency, resulting in high switching losses. In contrast, the inner switches (S_2 and S_3) operate at a low switching frequency, with low switching losses. As a result, in long-term operation, the loss distribution is uneven, which affects device reliability. In order to solve the problem of uneven distribution of device losses, Active NPC (ANPC) topology is proposed, as shown in Figure 2b (Barzegarkhoo et al., 2023). Furthermore, T-type 3-level are also widely adopted as shown in Figure 2c. This topology features higher efficiency in medium switching frequency (Zhang et al., 2023c)- (Zhang et al., 2024). Although 3-level topologies described above are widely used in PV systems, they have limitations in terms of ground current suppression (Barzegarkhoo et al., 2023). It is worth noting that this ground current suppression method requires grid neutral-point accessibility. A novel 3-phase H8 architecture is proposed by adding two additional switches in the DC path that allow decoupling the AC side from the DC side during freewheeling phases (Noroozi et al., 2017)- (Concari et al., 2017). Furthermore, optimized modulation strategies are explained.

In practice, 3-wire systems are widely used for 3-phase PV systems. This means that the grid neutral-point cannot be accessed and thus cannot be connected to the dc-bus. In this case, it is common ground current suppression practice to increase the inverter's capacitors to ground. However, this comes at a higher cost, and the suppression effect is not perfect. Furthermore, in large PV plants, a large number of PV inverters are linked together at the grid-side to generate more power. As the number of PV inverter modules increases, the ground current problem worsens. This can easily result in a ground current fault, and cause the inverter system failed and disconnected from the grid. Therefore, this paper will propose solutions for reducing ground current in 3-phase systems with a large number of PV inverter modules. The hardware solution is to connect filter capacitors back to the DC-bus, directing more ground current to the inverter rather than the grid to reduce the ground current. The software solutions include carrier phase shifting and carrier frequency shifting. The ground current for the PV inverter modules can be cancelled each other in software solutions, which can alleviate the ground current problem.

The following are the main contributions of this paper.

- 1) The concept of connecting neutral point of filter capacitors to DC-bus midpoint to suppress ground current is introduced. The effects of the DC/AC converter's zero-sequence voltage and the grid's zero-sequence voltage on zero-sequence current are investigated.
- 2) A hardware solution is proposed to suppress ground current. When the zero-sequence current is primarily caused by the DC/AC converter's zero-sequence voltage, increasing the value of capacitors connected to DC-bus can reduce ground current. When the zero-sequence voltage is primarily from the grid side, increasing the value of capacitors connected to the grid neutral point can reduce ground current.
- 3) Two software solutions, carrier phase shifting and carrier frequency shifting, are proposed to suppress ground current, particularly in systems with a large number of PV inverter modules. The concept and implementation for two proposed software solutions are detailed.

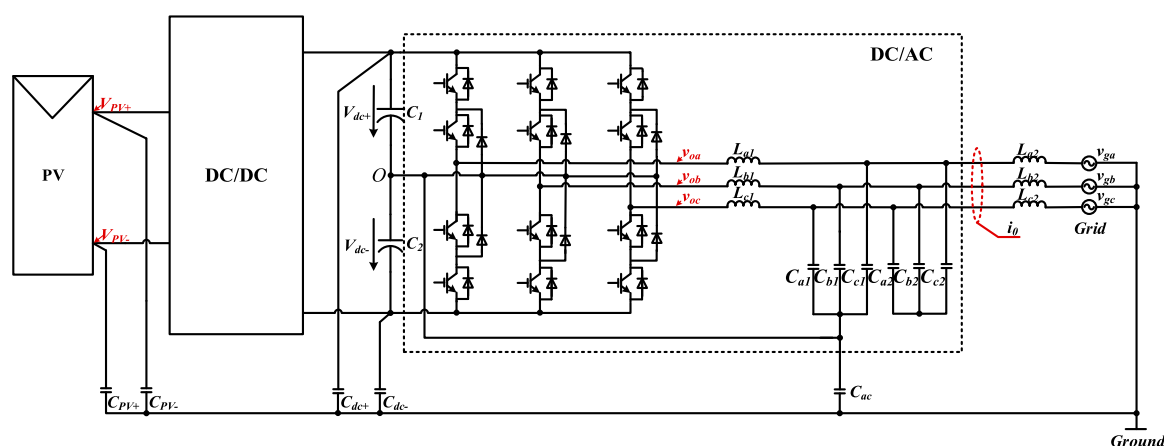


FIGURE 3
System architecture.

It is worth mentioning that the proposed hardware solution simply connects filter capacitors to the dc-bus or the grid neutral point to suppress ground current, which is significantly less expensive than adding additional switching devices. Furthermore,

the software solution, which is intended for multi-inverter parallel scenarios, uses phase shifting and frequency staggering of the carrier to effectively suppress ground currents without incurring additional hardware costs.

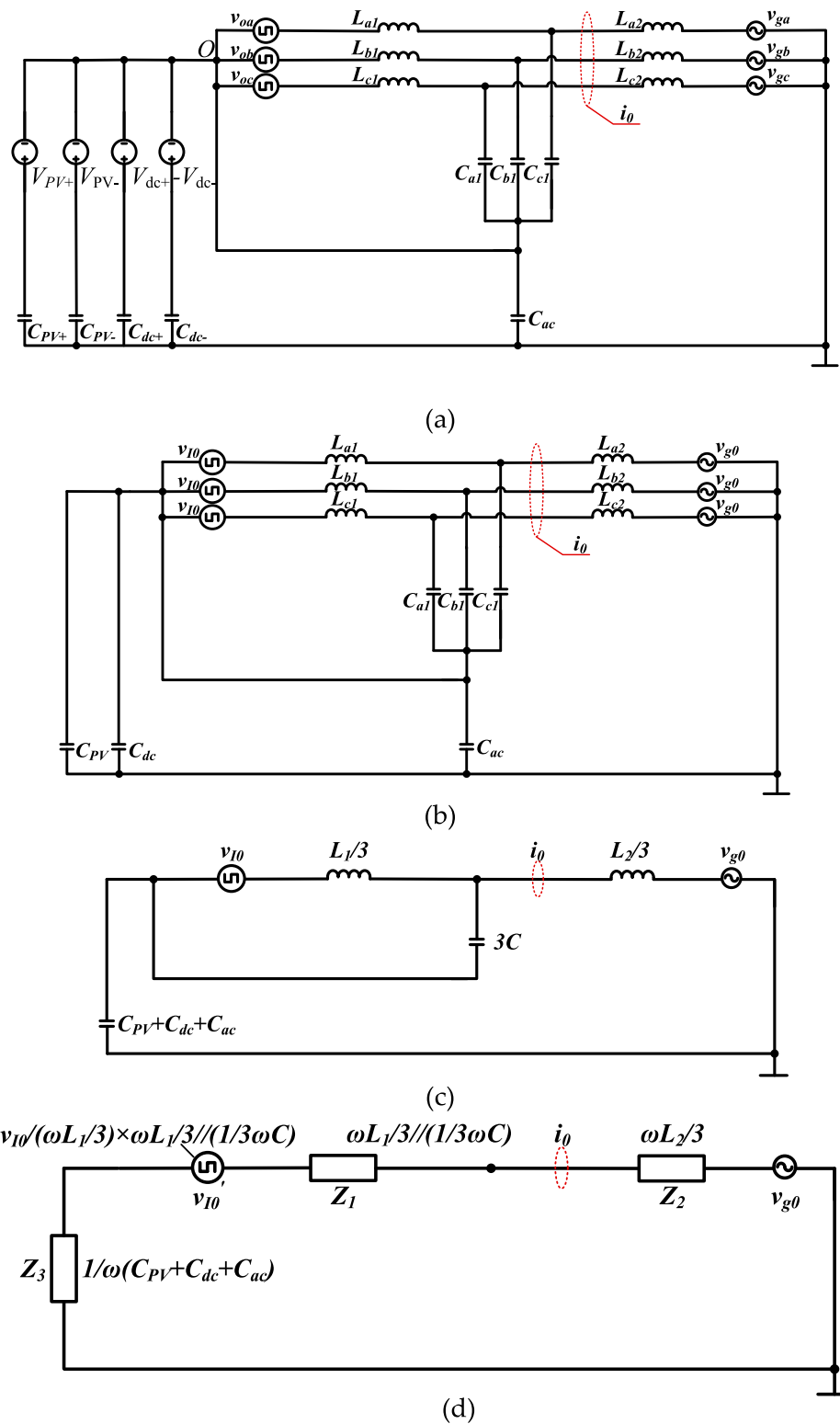


FIGURE 4 Zero-sequence model, (a) equivalent model; (b) simplified model; (c) further simplified model; (d) impedance model.

TABLE 1 System parameters.

Items	Values
Inverter-side inductor L_1	497×10^{-6} H with parasitic resistance 0.02Ω
Grid-side inductor L_2	31.5×10^{-6} H with parasitic resistance 0.02Ω
Filter capacitor C	2.2×10^{-6} F
PV to ground capacitor C_{PV}	65.8×10^{-9} F
DC-bus to ground capacitor C_{dc}	14.1×10^{-9} F
AC side to ground capacitor C_{ac}	33×10^{-9} F

2 System architecture and zero-sequence model

The system architecture is presented in Figure 3. PV feeds the grid via DC/AC converters. The DC/AC converter uses 3-level NPC topology, but not limited to it. In Figure 3, there are three ground capacitors. C_{PV+} and C_{PV-} represent the capacitors of the PV to the ground; C_{dc+} and C_{dc-} represent the capacitors of the DC-bus to the ground; and C_{ac} represents the capacitor of the output filter to the ground. The output zero-sequence current i_0 is measured, which represents the system's ground current.

In the architecture, the output filter capacitors are divided into two groups. C_{a1} , C_{b1} and C_{c1} are in one group, while C_{a2} , C_{b2} and C_{c2} are in the other group. The distinction is whether the neutral

point of the filter capacitors is connected to the DC-bus midpoint. By connecting the filter capacitors (C_{a1} , C_{b1} and C_{c1}) to the DC-bus, a portion of the zero-sequence current from the inverter-side can flow into the DC-bus, which can reduce the ground current flowing into the grid-side.

Assume that the DC/AC converter's output voltages with respect to the DC-bus neutral point O are v_{oa} , v_{ob} , and v_{oc} . Furthermore, the PV positive and negative with respect to the DC-bus neutral point O are V_{PV+} and V_{PV-} , respectively. The equivalent mathematical model is obtained and illustrated in Figure 4a.

Because ground current is proportional to zero-sequence voltage, the zero-sequence equivalent circuit is investigated. Assume the DC/AC converter's output zero-sequence voltage is v_{10} and the grid's zero-sequence voltage is v_{g0} . Furthermore, assuming that $V_{dc+} = V_{dc-} = V_{dc}/2$, $V_{PV+} = -V_{PV-}$, $C_{PV+} = C_{PV-} = C_{PV}/2$, and $C_{dc+} = C_{dc-} = C_{dc}/2$, Figure 4a can be simplified to Figure 4b.

Due to the symmetry of the parameters $L_{a1} = L_{b1} = L_{c1} = L_1$, $L_{a2} = L_{b2} = L_{c2} = L_2$ and $C_{a1} = C_{b1} = C_{c1} = C$, Figure 4b can be simplified to Figure 4c.

The circuit can be further simplified from Figure 4c, and the corresponding impedance model can be obtained, which is shown in Figure 4d.

3 Zero-sequence current analysis

Based on Figure 4, the following will investigate the effect of the DC/AC converter's zero-sequence voltage v_{10} and the grid's zero-sequence voltage v_{g0} on zero-sequence current i_0 .

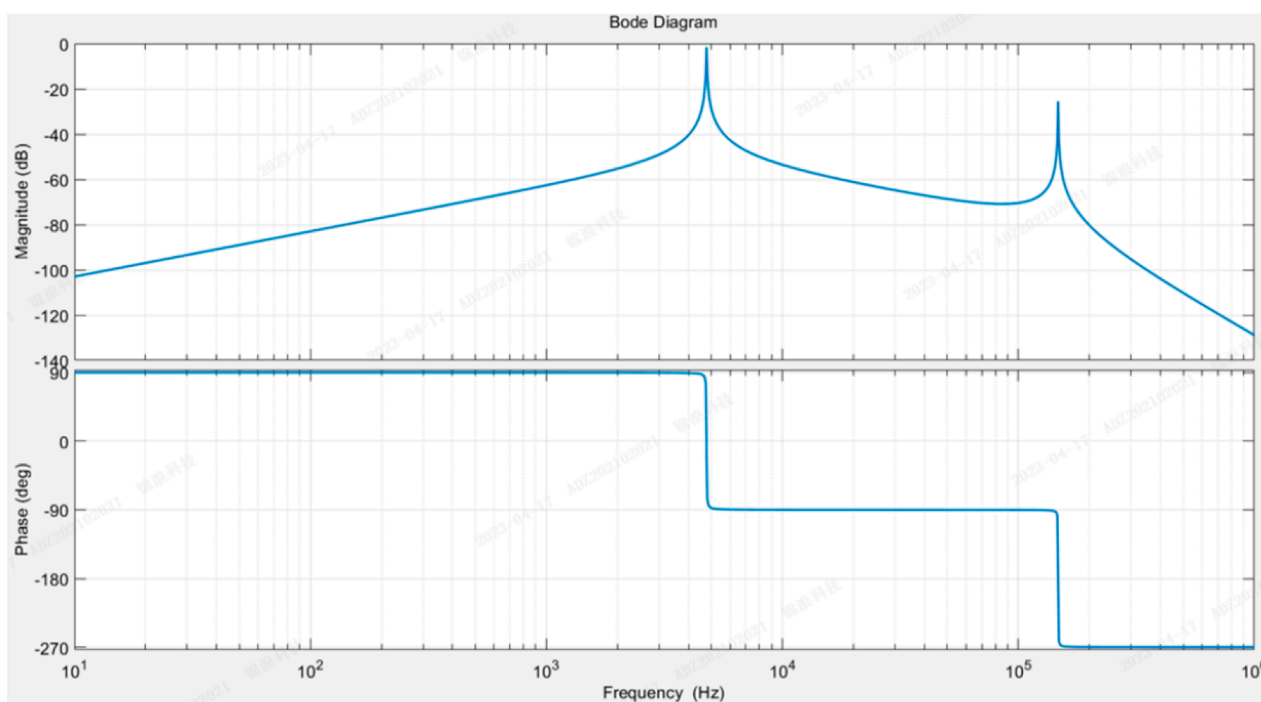
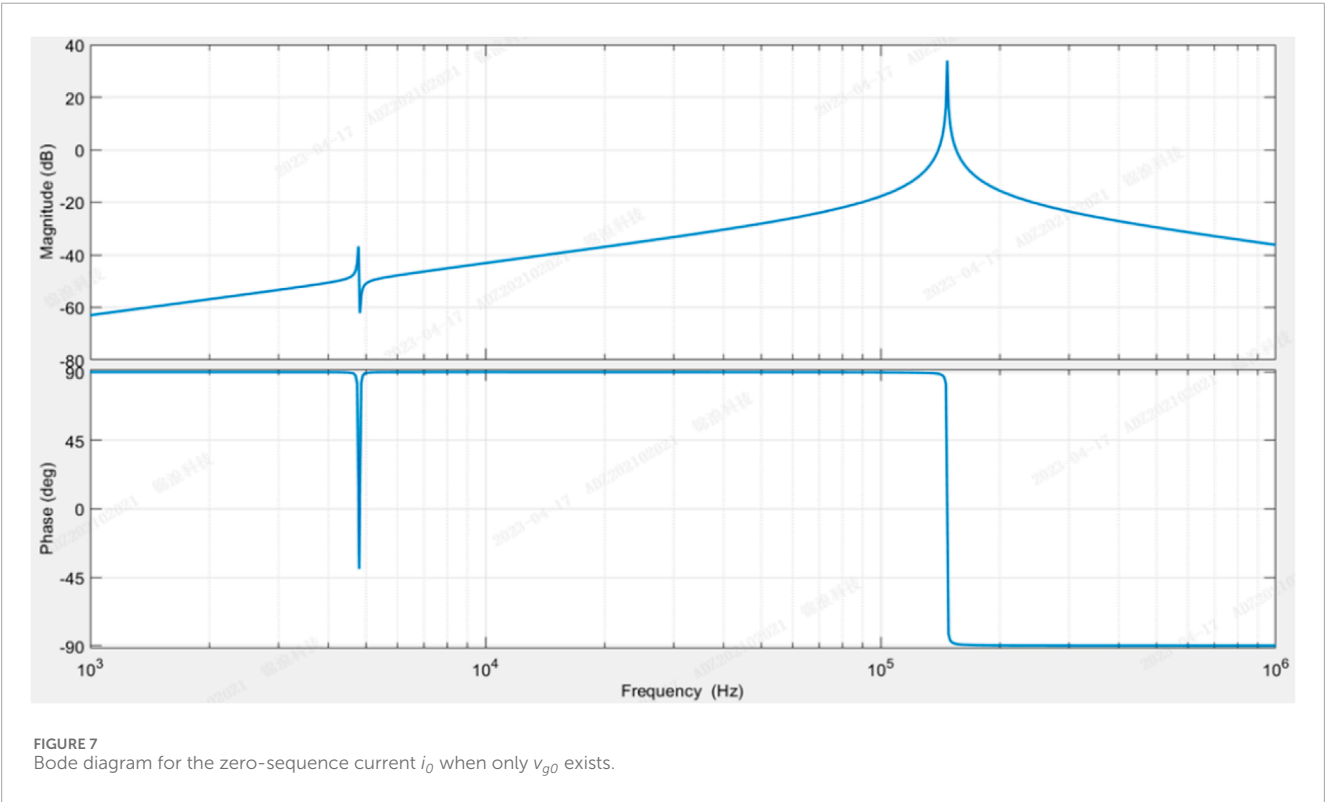
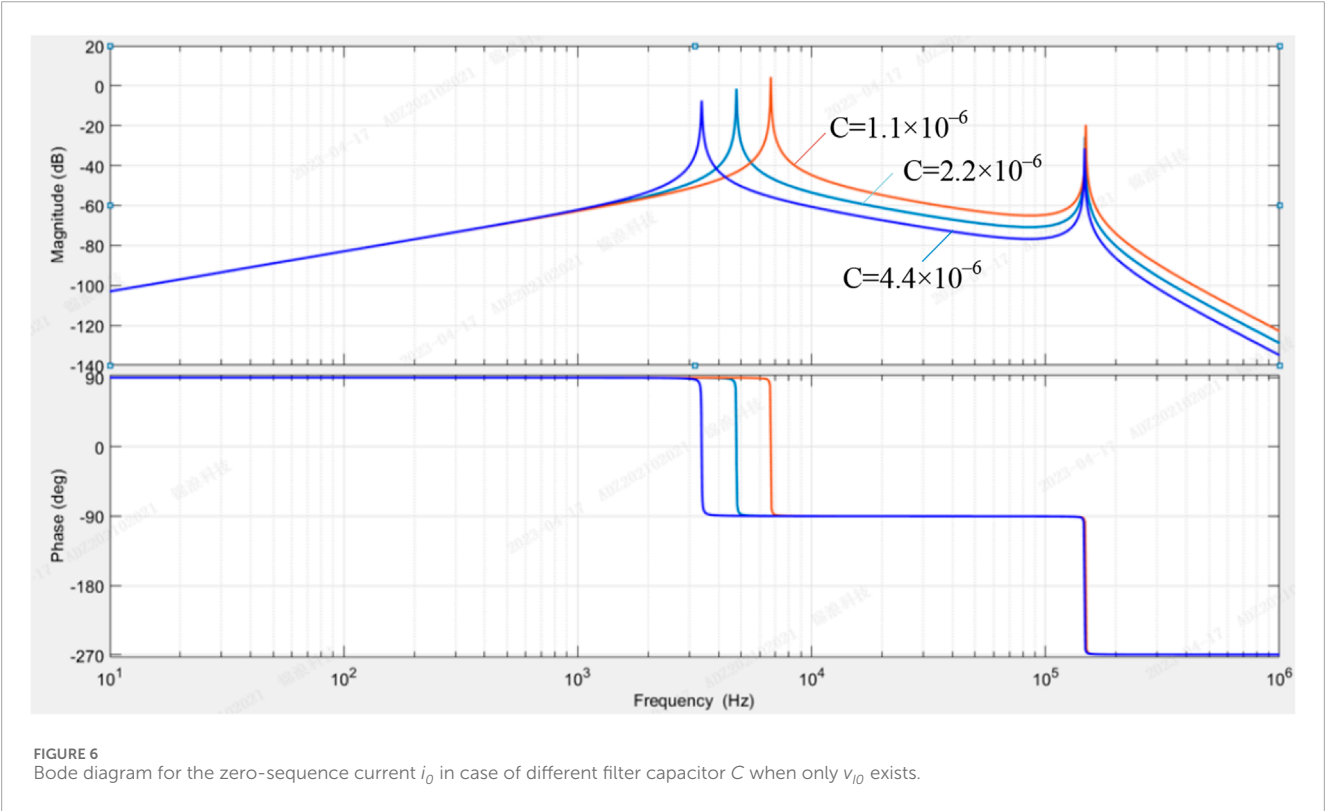


FIGURE 5 Bode diagram for the zero-sequence current i_0 when only v_{10} exists.



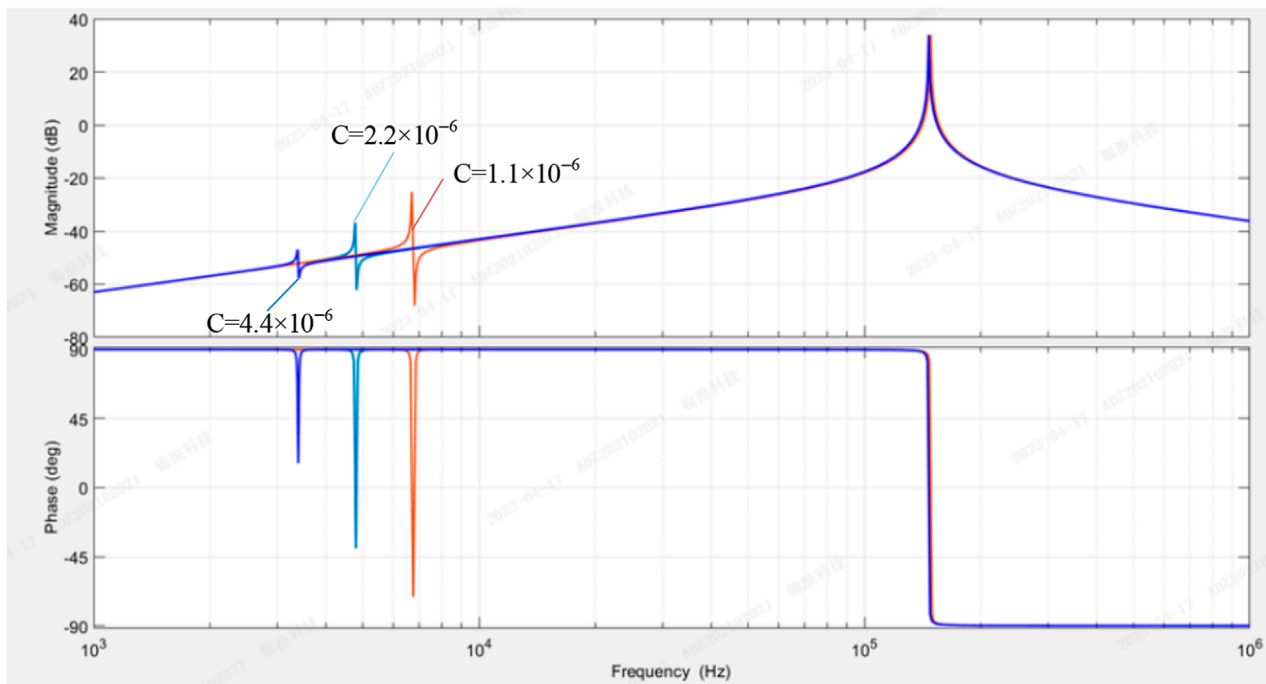


FIGURE 8
Bode diagram for the zero-sequence current i_0 in case of different filter capacitor C when only v_{g0} exists.

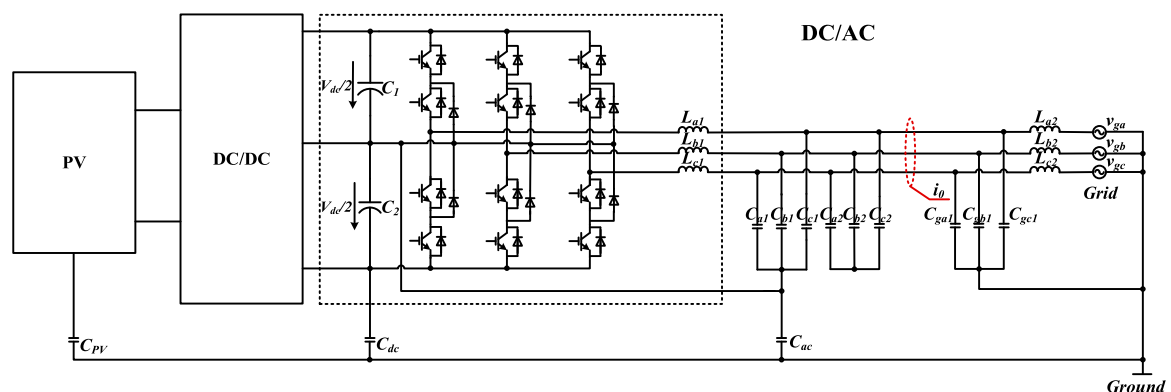


FIGURE 9
Hardware solution: system architecture variant I.

3.1 Effect of DC/AC converter's zero-sequence voltage on zero-sequence current

First, by setting the grid zero-sequence voltage v_{g0} to zero, examine the effect of the DC/AC converter's zero-sequence voltage v_{l0} on the zero-sequence current i_0 . From Figure 4d, the expressions for v'_{l0} , Z_1 , Z_2 and Z_3 are as follows shown in Equations 1–4.

$$v'_{l0} = \frac{v_{l0}}{\frac{\omega L_1}{3}} \left(\frac{\omega L_1}{3} // \frac{1}{3\omega C} \right) \quad (1)$$

$$Z_1 = \frac{\omega L_1}{3} // \frac{1}{3\omega C} \quad (2)$$

$$Z_2 = \frac{1}{3\omega C} \quad (3)$$

$$Z_3 = \frac{1}{\omega(C_{pv} + C_{dc} + C_{ac})} \quad (4)$$

From Figure 4d, the zero-sequence current i_0 can be obtained as well. Table 1 shows the system parameters. With the parameters and Table 1, the zero-sequence current i_0 can be calculated.

Figure 5 shows bode diagram for the zero-sequence current i_0 when only v_{l0} exists and $v_{g0} = 0$. From Figure 5, there are two resonant frequencies. One is the resonance of L_1 and C with a resonant frequency of around 5 kHz. The second is the

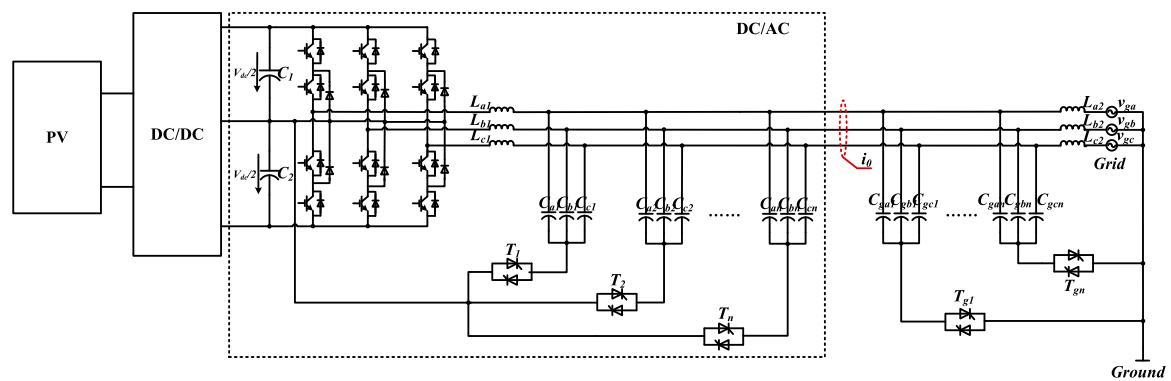


FIGURE 10
Hardware solution: system architecture variant II.

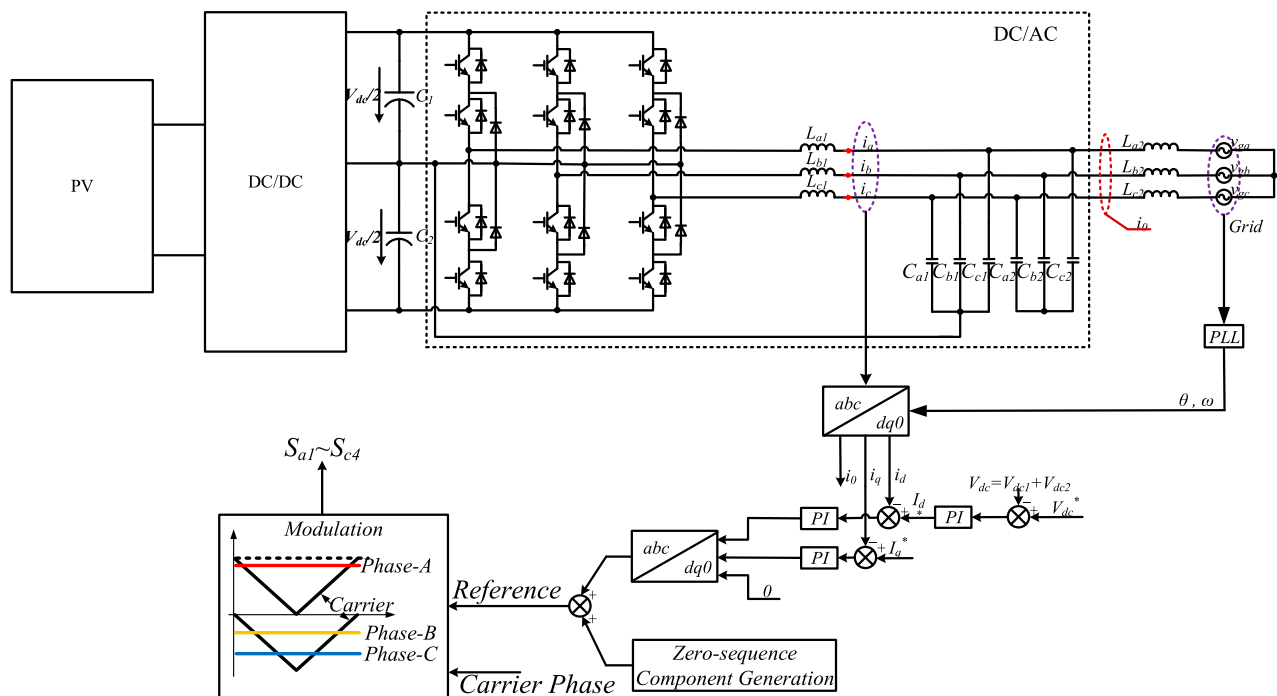


FIGURE 11
Control architecture.

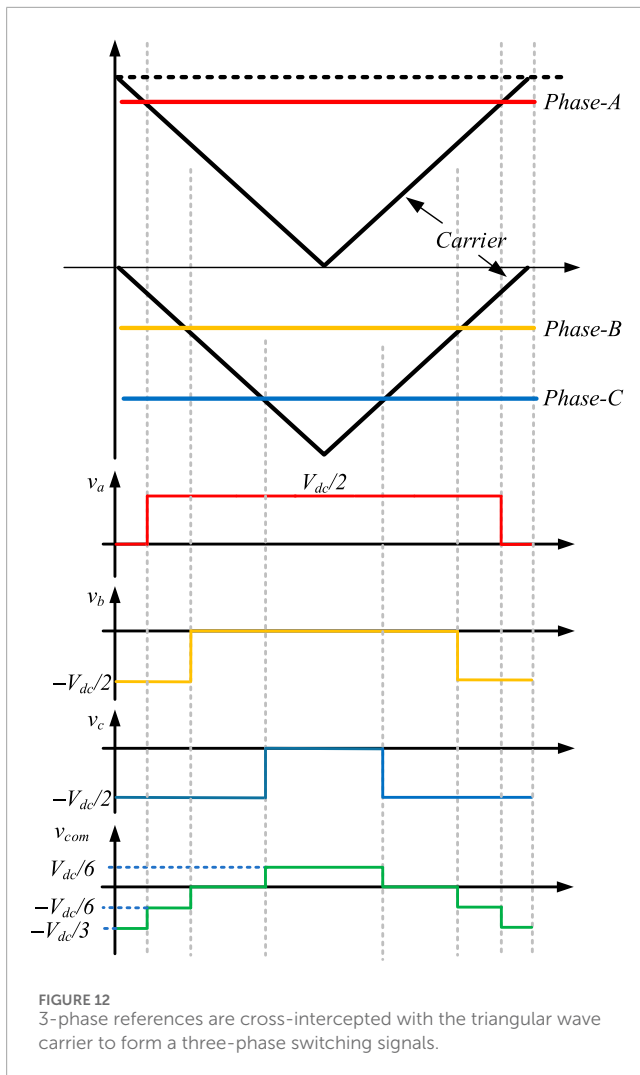
resonance of L_2 and C_{PV} , C_{dc} and C_{ac} , with a resonant frequency of around 120 kHz.

Figure 6 presents bode diagram for the zero-sequence current i_0 in case of different filter capacitors C . From it, it can be seen that the resonant frequency shifts to the left as the value of the capacitor connected to the DC-bus increases. As a result, the zero-sequence current suppression in the high frequency band is improved. Specifically, when the capacitor connected to the DC-bus is doubled, the zero-sequence current suppression in the range of 10 kHz–100 kHz is increased by approximately 7 dB. Furthermore, taking 10 kHz as an example, when C is 1.1 μF , the

magnitude of i_0 is -45 dB. When C is 2.2 μF , the magnitude of i_0 is -53 dB, and when C is 4.4 μF , the magnitude of i_0 is further suppressed to -60 dB.

3.2 Effect of the grid's zero-sequence voltage on zero-sequence current

Similarly, by setting the DC/AC converter's zero-sequence voltage v_{l0} to zero, examine the effect of the zero-sequence grid voltage v_{g0} on the zero-sequence current i_0 . From Figure 4c, the

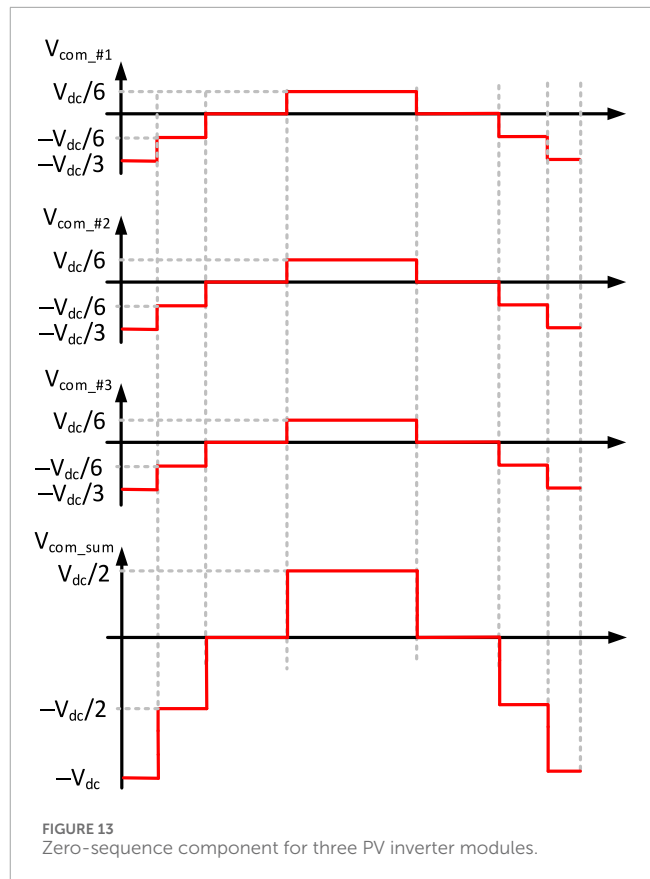


zero-sequence current i_0 can be obtained as well.

$$i_0 = \frac{v_{g0}}{Z_1 + Z_2 + Z_3} \quad (5)$$

With the parameters in Table 1 and Equation 5, the zero-sequence current i_0 can be calculated. Figure 7 shows bode diagram for zero-sequence current i_0 when only v_{g0} exists and $v_{l0} = 0$. From Figure 7, there are two resonant frequencies. The first one is the resonance of L_1 and C with a resonant frequency of around 5 kHz. The second one is the resonance of L_2 and C_{PV} , C_{dc} and C_{ac} , with a resonant frequency of around 120 kHz. Furthermore, compared with the result in Figure 5, of these two resonators, the second resonator (~ 120 K) predominates in this case.

Figure 8 presents bode diagram for the zero-sequence current i_0 in case of different filter capacitors C . From it, the first resonant frequency shifts to the left as the capacitor connected to the DC-bus C increases. Generally, there are no changes on the effect of zero-sequence current suppression. This is because, the zero-sequence current is determined by the loop impedance, and the impedance of the loop is determined primarily by the capacitors of C_{PV} , C_{dc} and C_{ac} , rather than the capacitor connected to DC-bus C .



4 Hardware solution for ground current suppression

According to the above analysis, when the zero-sequence current is caused by the DC/AC converter's zero-sequence voltage, increasing capacitance connected to the DC-bus can greatly improve zero-sequence current suppression. When the zero-sequence current is caused by the zero-sequence grid voltage, increasing capacitance connected to DC-bus C has no effect on the zero-sequence current suppression. Therefore, when the zero-sequence voltage comes primarily from the converter side rather than the grid side, one simple way to reduce ground current is to increase the capacitance of filter capacitors connected to the DC-bus.

When the zero-sequence voltage is primarily from the grid side rather than the inverter side, installing capacitors connected to the grid neutral point is a simple way to reduce ground current. Figure 9 depicts an alternative system architecture in which the capacitors $C_{ga1} \sim C_{gc1}$ are added and connected to the grid neutral point. The capacitors $C_{ga1} \sim C_{gc1}$ can suppress the ground current caused by grid-side zero-sequence components.

It is worth noting that as the capacitance of $C_{a1} \sim C_{c1}$ increases, so does the current flowing into the DC-bus, resulting in greater DC-bus fluctuation. As a result, we need make a compromise in selecting the capacitance connected to the DC-bus in order to achieve a balance between the DC-bus fluctuation and ground current. Figure 10 depicts another alternative system architecture. All filter capacitors are connected to the DC-bus midpoint or the grid neutral

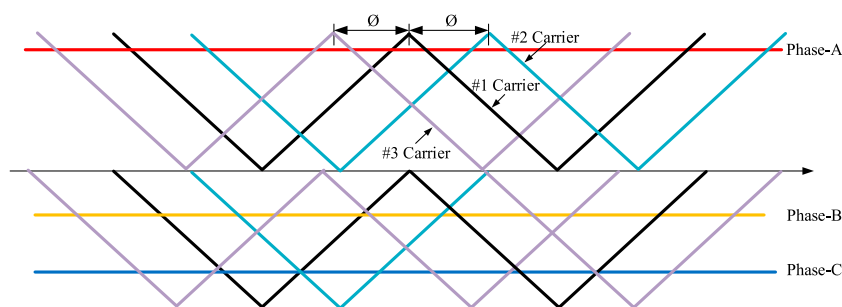


FIGURE 14
Carrier phase shifting method.

via a triode for alternating current (TRIAC). As shown in Figure 10, the filter capacitors (C_{a1} , C_{b1} , C_{c1}) are connected to the DC-bus midpoint via T_I , and the filter capacitors (C_{ga1} , C_{gb1} , C_{gc1}) are connected to the grid neutral-point via T_{g1} . As a result, the system has more degrees of freedom and can freely select which group of the capacitors to connect back to the midpoint of the DC-bus based on the value of the system ground current. On one hand, the ground current can be suppressed in this manner to meet the requirement; on the other hand, the impact on the DC-bus is minimal.

From the above analysis, the proposed methods for suppressing ground current are highly scalable. This approach is not limited by the inverter's power level or the number of parallel inverters, making it particularly useful for high-power applications. The most difficult challenge is dealing with various scenarios in which ground current is caused by either the inverter or the grid. As a result, modified solutions are also proposed in the paper.

5 Software solution for ground current suppression

5.1 Control architecture

Figure 11 depicts the control architecture for the DC/AC converter in PV string systems. The outer loop is a DC-bus voltage loop, while the inner loop is a current loop. The three phases are transformed into d - q frame, and the control method is the same as the conventional one. Then, the three-phase references generated are fed into the modulation unit and cross-intercepted with the triangular wave carrier to form three-phase switching signals.

The concept of the modulation unit is presented in Figure 12. The three-phase reference signals and triangle wave carrier are cross-sectioned, and then three-phase switching signals are obtained. As a result, three-phase output voltages (v_a , v_b and v_c) are generated, and the zero-sequence component of three-phase output voltages v_{com} can be calculated using (6). The three-phase output voltages and the corresponding zero-sequence component are shown in Figure 12 as well.

$$v_{com} = \frac{v_a + v_b + v_c}{3} \quad (6)$$

The three-phase output voltage and zero-sequence components for any PV inverters can be obtained using the same method. In a

large PV plant, many PV inverters are linked together at the grid-side to generate more power. Taking three PV inverter modules as an example, assuming that the voltage reference of each module is the same, the three-phase output voltages and the corresponding zero-sequence voltage components are shown in Figure 13. Using Equation 7, the sum of zero-sequence voltages for three PV inverter modules v_{com_sum} can be obtained, which is shown in Figure 13 as well. From Figure 13, for three PV inverter system, the maximum step of the zero-sequence voltage is $V_{dc}/2$ and the range of the change is $-V_{dc}$ to $V_{dc}/2$. Also, it can be seen that as the size of the PV inverter modules increases, so do the sum of the corresponding zero-sequence voltages and the corresponding ground current. Therefore, as the number of PV inverter modules increases, the ground current may exceed the standard. This can result in a ground current fault, which can cause the inverter system failed and disconnected from the grid.

$$v_{com_sum} = v_{com_\#1} + v_{com_ \#2} + v_{com_ \#3} \quad (7)$$

From the above analysis, it is necessary to suppress ground current, especially for the system with high number of PV inverter modules. The following are two software solutions proposed.

5.2 Software solution I: carrier phase shifting

To suppress ground current, a triangulation wave carrier phase shifting method is proposed in this section. As shown in Figure 14, the triangle wave carrier for different inverter modules is shifted by a degree of Θ . Taking three of inverter modules as an example, when compared to module #1, the triangle wave carrier for module #2 lags by Θ , while the triangle wave carrier for module #3 is ahead by Θ .

Figure 15 depicts the intersection of the reference signals and the phase-shifted triangle wave carriers for three inverter modules, which can result in three-phase switching signals and three-phase output voltages (v_a , v_b , v_c). Due to the fact that the triangular wave carriers for three inverter modules are phase shifted by a degree of Θ . As a result, the output voltages of three inverter modules are phase shifted by the degree of Θ , which is shown in Figure 15.

The corresponding zero-sequence voltage component is obtained from the three-phase output voltages. Figure 16 depicts the proposed method's zero-sequence voltage component. It can be seen

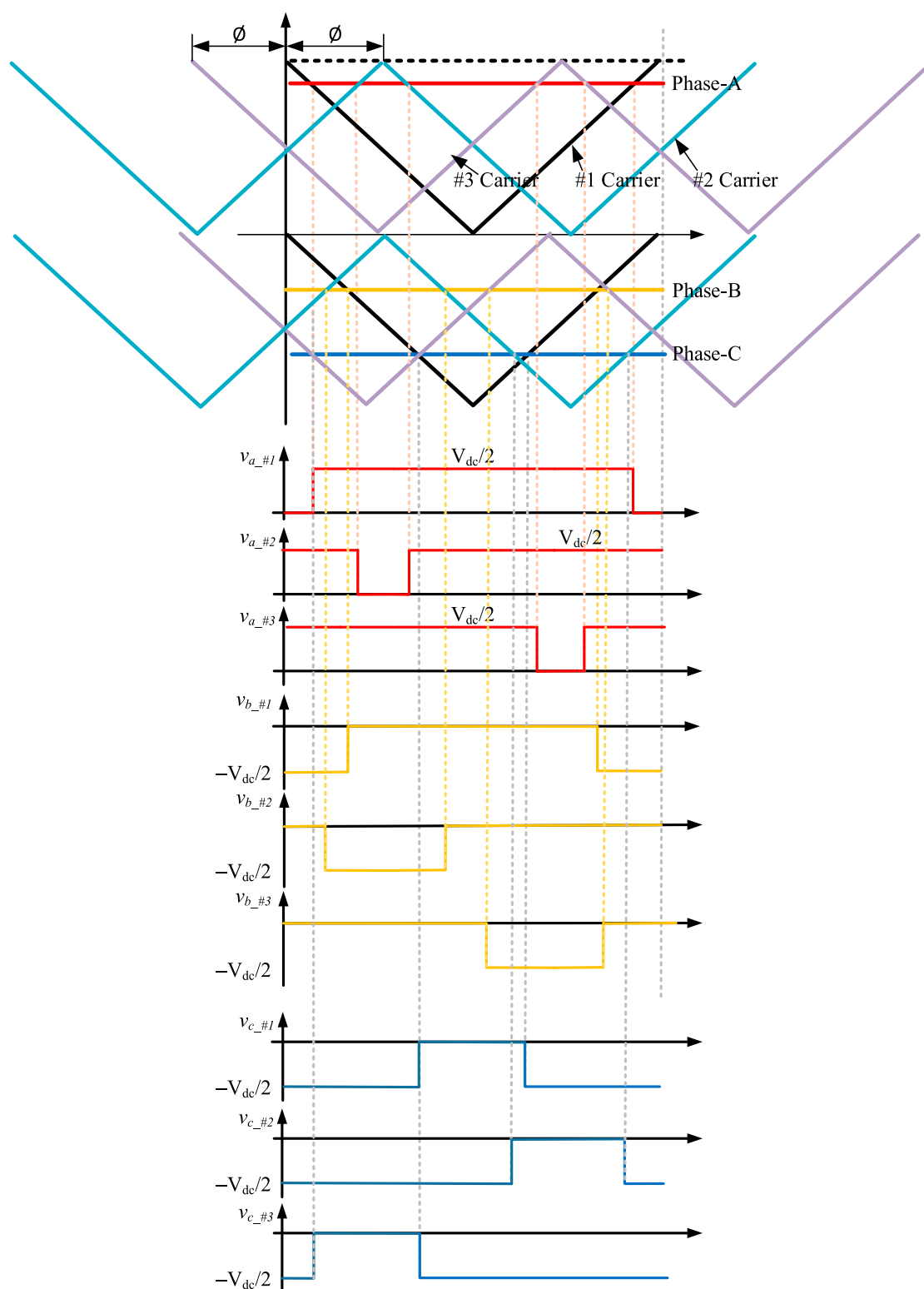
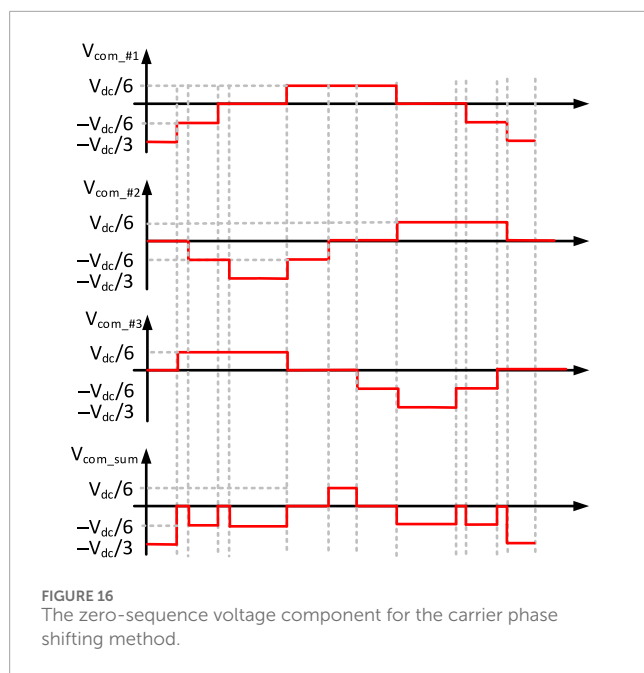


FIGURE 15

The intersection of the reference signals and the phase-shifted triangle wave carriers for three inverter modules.

that the three modules' zero-sequence voltage has a phase shift of ϕ as well. As illustrated in Figure 16, the total zero-sequence voltage of the three modules v_{com_sum} is with a maximum step of $V_{dc}/3$ and

a variation range of $-V_{dc}/3$ to $V_{dc}/6$. For the conventional method, the maximum step is $V_{dc}/2$ and the range of change is $-V_{dc}$ to $V_{dc}/2$. Therefore, compared to the traditional method, the maximum



step and variation range of the zero-sequence voltage are greatly improved by the proposed triangular wave phase shifting. As a result, the proposed method can effectively improve the problem of ground current as the number of PV inverter modules increases. Therefore, the proposed method can effectively suppress the ground current to avoid exceeding the limit.

5.3 Software solution II: carrier frequency shifting

In this section, another method with triangulation wave carrier frequency shifting is proposed to suppress ground current. As shown in Figure 17, the triangle wave carrier frequencies for different inverter modules are shifted by Δf . Taking three of inverter modules as an example, the triangle wave carrier frequency for module #1 is f_1 . When compared to module #1, the triangle wave carrier frequency for module #2 is reduced by Δf , while the triangle wave carrier for module #3 is increased by Δf .

Following the same method above, the reference signals and the triangle wave carriers are intersected for three inverter modules. Then, three-phase switching signals and three-phase output voltages (v_a , v_b , v_c) are obtained. With the three-phase output voltage, the corresponding zero-sequence voltage component is obtained, which is shown in Figure 18. As illustrated in Figure 18, for the zero-sequence voltage sum v_{com_sum} , the step is roughly $V_{dc}/6$ and the range is roughly between $-V_{dc}/3$ to $V_{dc}/6$. Therefore, compared to the traditional method, the maximum step and variation range of the zero-sequence voltage are improved with the proposed triangular wave frequency shifting. As a result, the proposed method can effectively suppress ground current as the number of PV inverter modules increases.

Figure 19 depicts an implementation method, which includes the addition of a carrier synchronization cable and a communication

cable such as CAN. All PV inverter modules can realize triangulation wave carrier synchronization via the carrier synchronization cable. The carrier phase shift angles or frequencies can be transmitted via the communication cable. Normally, there is one inverter module called the master unit, which can calculate all phases or frequencies for all the inverter modules. The rest of the inverter modules are referred to slave units and they receive the phase or frequency information transmitted by the communication cable. Module #2, for example, receives the phase shift angle $\Delta\theta_2$ transmitted through the communication cable and then implements carrier phase shifting.

From above, only the communication process for the master inverter to send phase shift angles to the slave inverters is included. This process can be completed before the inverter is started, and there are almost no additional requirements for the controller's processing capability, making it extremely simple to deploy.

The software solution specifically proposes a method for carrier frequency shifting and phase shifting for multi-inverter systems that has shown excellent performance in inverter parallel operation and is thus appropriate for high-power applications. The greatest challenge is achieving phase angle synchronization. Nonetheless, relevant papers in detail will be published in the future.

Using TI's TMS320F28374S as the main control chip, the phase-shifting and frequency-shifting ground current suppression method described in this paper is integrated into the existing control algorithm. TMS320F28374S's main frequency is 200MHz, and its Read-Only Memory (ROM) resources are allocated as detailed below. The TMS320F28374S DSP chip has 64 KB flash, 1 KB One Time Programmable (OTP) ROM, and 4 KB Boot ROM. The address for Flash ranges from 0x3E8000 to 0x3F8000. The original code only takes about 32 KB. Because the proposed method involves phase or frequency shifting for the carriers, the extra flash space increases by only a few bytes. As a result, there is no risk of using up too much code space.

The system's operating process is described below. There is no operating system, and a single interrupt is used. As shown in Figure 20, the entire process consists of three parts: initialization, the main loop, and the interrupt service routine (ISR). In the Initialization section, RAM is initialized using assembly with CodeStartBranch.asm. In addition, the chip peripherals, interrupts, and global variables are initialized. The main loop section implements communication with the ARM, sub-DSP, and host computer. Other tasks include digital IO processing, start-stop-restart logic, fault handling, and operating mode switching. In the ISR part, DMA triggers AD sampling at a sampling frequency of 160 kHz. The DMA interrupt is triggered every eight samples, giving the interrupt frequency of 20 kHz. The ISR performs sampling conversion and filtering, loop control, phase locking, logic processing, and a small amount of communication handling.

In the implementation of the proposed method, only the communication process for the master inverter to send phase shift angles to the slave inverters is included. This process can be completed before the inverter is started, there will be no impact on the ISR part and the code can be implemented in the main loop. Therefore, there are almost no additional requirements for the controller's processing capability, making it extremely simple to deploy.

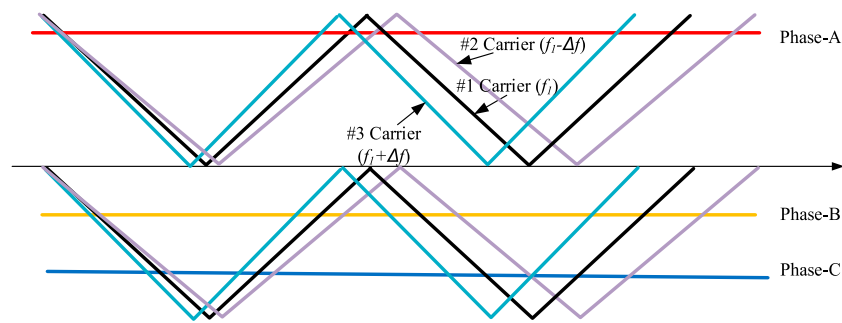


FIGURE 17
Carrier frequency shifting method.

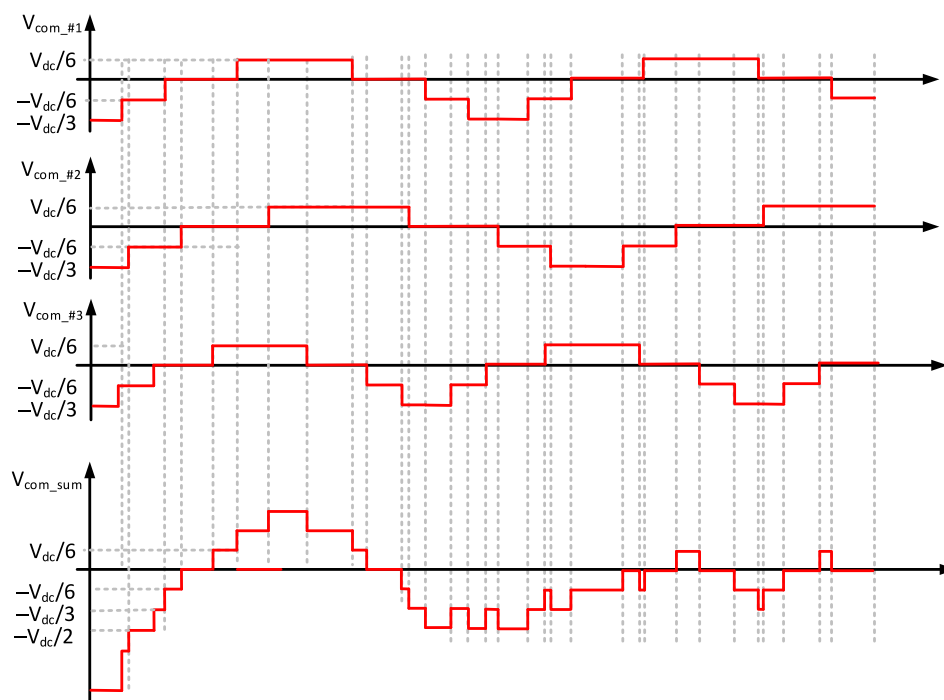


FIGURE 18
The zero-sequence voltage component for the carrier phase shifting method.

6 Experimental verifications

The experiment is carried out in a PV plant with 20 inverter modules, each of which has a rated power of 25 kW. In addition, the system parameters are presented above in Table 2.

Figure 21 shows the zero-sequence current i_0 , inductor current i_a , zero-sequence current after removing high-frequency component and PV negative to ground voltage V_{PV_G} . Although the grid current i_a is of good quality, the zero-sequence current i_0 contains a large number of high-frequency components and the peak-to-peak value is around 10A. Because the ground current has high-frequency components, the voltage to the ground also has high-frequency components. In Figure 21, the PV to ground voltage V_{PV_G} has a significant high-frequency component which is around

100V, as labeled in pink color. After the high-frequency components are removed, the low-frequency variation of the zero-sequence current is large as shown by the green line. The ground-current fault is frequently triggered in this case, resulting in the system being disconnected from the grid. Therefore, as the number of PV inverter modules increases, the ground current problem worsens.

Figure 22 depicts the zero-sequence current i_0 , inductor current i_a , zero-sequence current after removing high-frequency component and PV negative to ground voltage V_{PV_G} after increasing the filter capacitor C from 2.2μF to 4.4 μF. Because the high-frequency components are reduced, the quality of the grid current i_a improves. Furthermore, the high-frequency components of the zero-sequence current i_0 are greatly reduced and the peak-to-peak value is around 0.5A. Because the high-frequency components

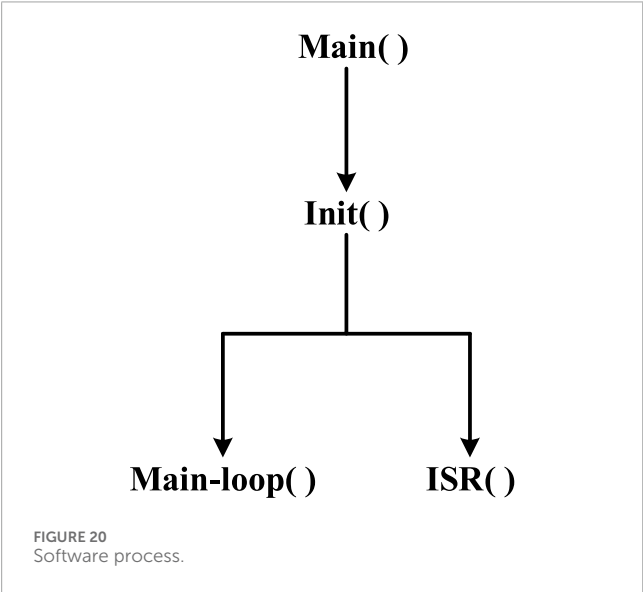
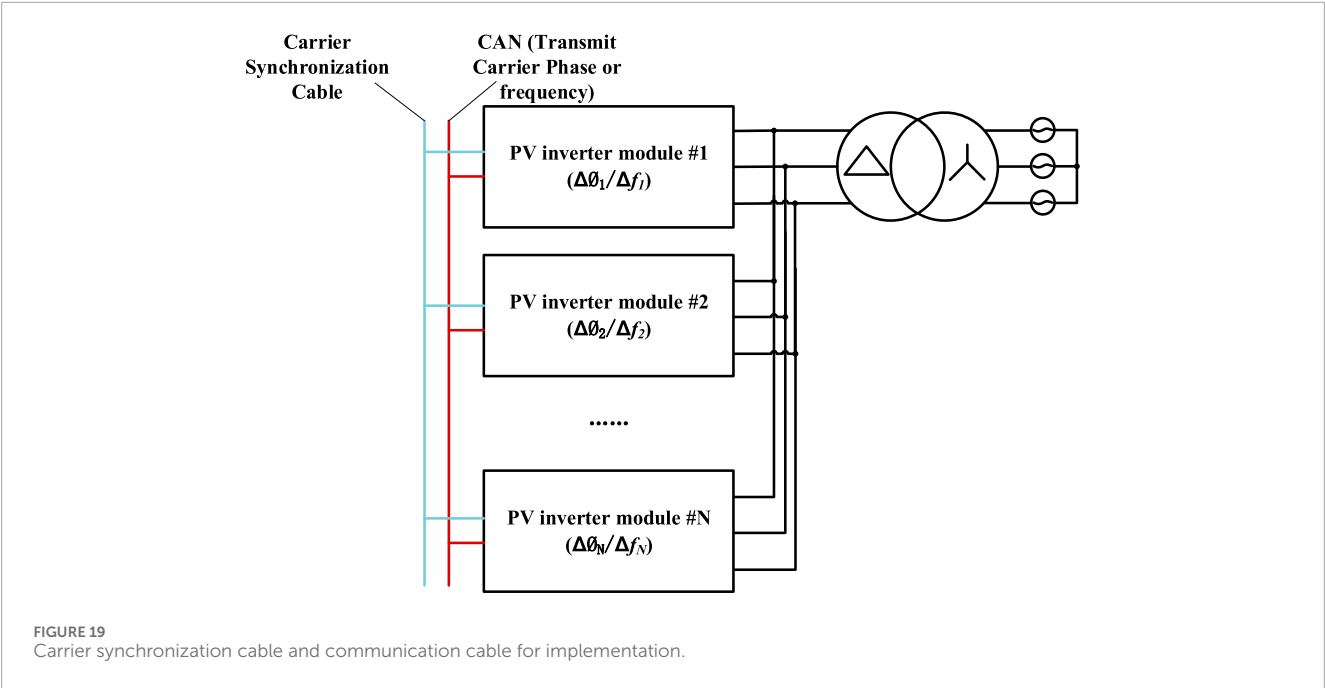


TABLE 2 System Parameters for the experimental setup.

Items	Values
System power	25 kW
DC-bus voltage V_{dc}	350V
Switching frequency	18 kHz
Power switches for NPC inverter	650V/100A
Inverter-side inductor L_1	497 μ H
Grid-side inductor L_2	31.5 μ H
Filter capacitor C	2.2 μ F
PV to ground capacitor C_{pv}	65.8 nF
DC-bus to ground capacitor C_{dc}	14.1 nF
AC side to ground capacitor C_{ac}	33 nF

of the ground current are reduced, the high-frequency components of the ground voltage are suppressed as well. The high-frequency components of PV to ground voltage V_{PV_G} , for example, are significantly reduced which is around 10V. The low-frequency variation of the zero-sequence current is reduced as well, as shown by the green line. Therefore, the method of increasing the capacitance of filter capacitors connected to the DC-bus is effective. The ground-current fault can be avoided, and the ground current problem can be solved.

Figure 23 shows the zero-sequence current i_0 , inductor current i_a , zero-sequence current after removing high-frequency component and PV negative to ground voltage V_{PV_G} when the carrier frequency shifting software solution is applied. The carrier

frequencies are shifted by 1 kHz with respect to each other for every four inverter modules, making them 16 kHz, 17 kHz, 18 kHz, and 19 kHz, respectively. When compared to the original method in Figure 21, the quality of the grid current i_a improves and the high-frequency components are reduced. The high-frequency components of the zero-sequence current i_0 are also reduced and the peak-to-peak value is around 7A. The high-frequency components of the ground voltage are also suppressed. In Figure 23, the high-frequency components of PV to ground voltage V_{PV_G} are reduced which is around 60V. As shown by the green line, the low-frequency variation of the zero-sequence current is also reduced. Therefore,

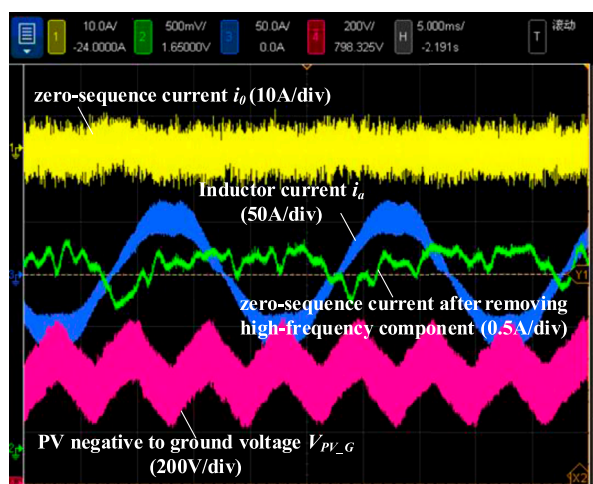


FIGURE 21
Zero-sequence current i_0 , inductor current i_a , zero-sequence current after removing high-frequency component and PV negative to ground voltage.

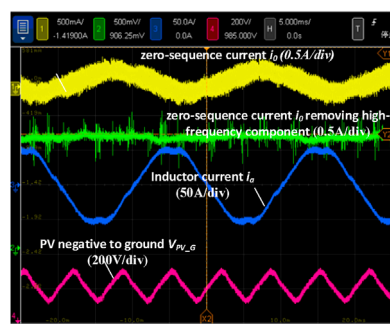


FIGURE 22
After increasing the filter capacitor from 2.2 μF to 4.4 μF , zero-sequence current i_0 , inductor current i_a , zero-sequence current after removing high-frequency component and PV negative to ground voltage with hardware solution.

the carrier frequency shifting method is effective to some extent. The ground-current faults can be reduced, and the ground current problem can be alleviated.

In response to grid voltage fluctuations, the inverter uses dynamic reactive power regulation technology to quickly and precisely control reactive power. By adjusting the inductive or capacitive reactive power output, it actively supports or mitigates grid voltage fluctuations.

As shown in Figure 24a, when the grid is normal, the Root Mean Square (RMS) value of the grid voltage is 230V. At t_1 , the grid voltage is increased to 260V, and the inverter implements high-voltage ride through and outputs reactive power. At t_2 , the grid voltage returns to normal, and the inverter's high voltage ride through is completed. As shown in Figure 24b, at t_1 , the grid voltage is decreased to 191V, and the inverter implements

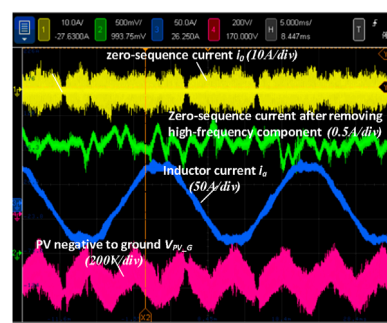


FIGURE 23
Zero-sequence current i_0 , inductor current i_a , zero-sequence current after removing high-frequency component and PV negative to ground voltage when the software solution of carrier frequency shifting is applied.

low-voltage ride through and outputs reactive power. At t_2 , the grid voltage returns to normal, and the inverter's low voltage ride through is completed. In the event of a voltage imbalance, the positive and negative sequence separation method is used to divide the three-phase unbalanced voltage into positive and negative sequence voltages. By suppressing the negative sequence voltage, normal output is achieved. As shown in Figure 25b, the grid voltage is unbalanced, where the grid voltages of three phases are 279.3V, 230V and 179.5V. With the positive and negative sequence separation method applied, the output power can be maintained as steady.

Therefore, the ground current suppression method described in this paper is unaffected by grid fluctuations or unbalanced loads. Furthermore, it does not worsen grid fluctuations or inverter output imbalances. As a result, this aspect of knowledge is not discussed further in this paper.

7 Cost assessment

The ground current suppression scheme discussed in this article has nothing to do with PV modules, so there is no need for them to be replaced and, as a result, the associated costs are avoided.

The phase-shifting and frequency-shifting methods described in this paper have minimal impact on the main controller's storage space and processing time. Therefore, there is no need to upgrade the main control chip or add an external memory chip. It is only necessary to upgrade the code to effectively suppress the problem of excessive ground current. As a result, the controllers will not incur any additional costs.

For multi-inverter communication, it can be accomplished using previously established communication lines, such as 485 communication lines, eliminating the need for additional costs.

The hardware solution's main cost is for additional filter capacitors. However, it is significantly less expensive than adding switch transistors in other alternatives. This paper's proposal to add capacitors between the inverter output neutral point and the dc-bus midpoint will result in a slight cost increase. A filter capacitor

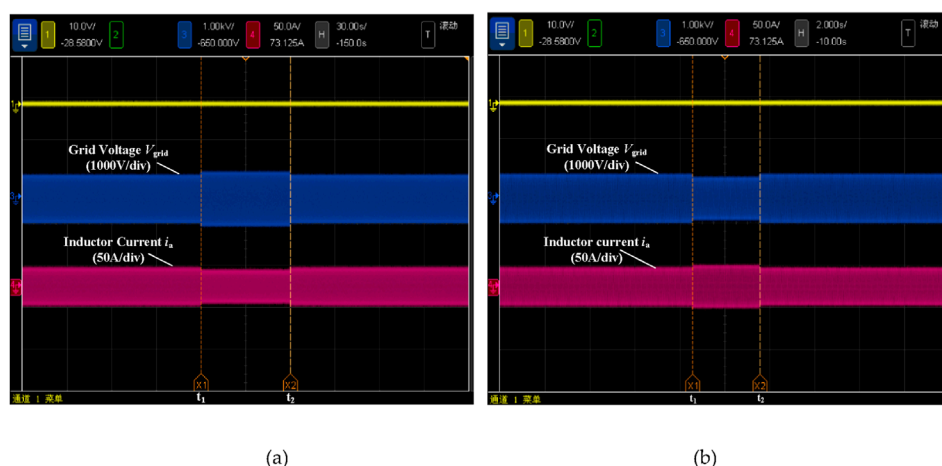


FIGURE 24

Grid voltage high voltage or low voltage ride through, (a) high voltage ride through, (b) low voltage ride through. Regarding grid frequency fluctuations, the inverter employs rapid frequency support technology to quickly adjust active power, thereby suppressing frequency fluctuations and maintaining grid frequency stability. As illustrated in Figure 25a, when the grid frequency is higher than normal, the inverter limits the output power and reduces the active power injected into the grid. Similarly, when the grid frequency is lower than normal, the inverter increases the output power, which supports the grid by injecting additional active power.

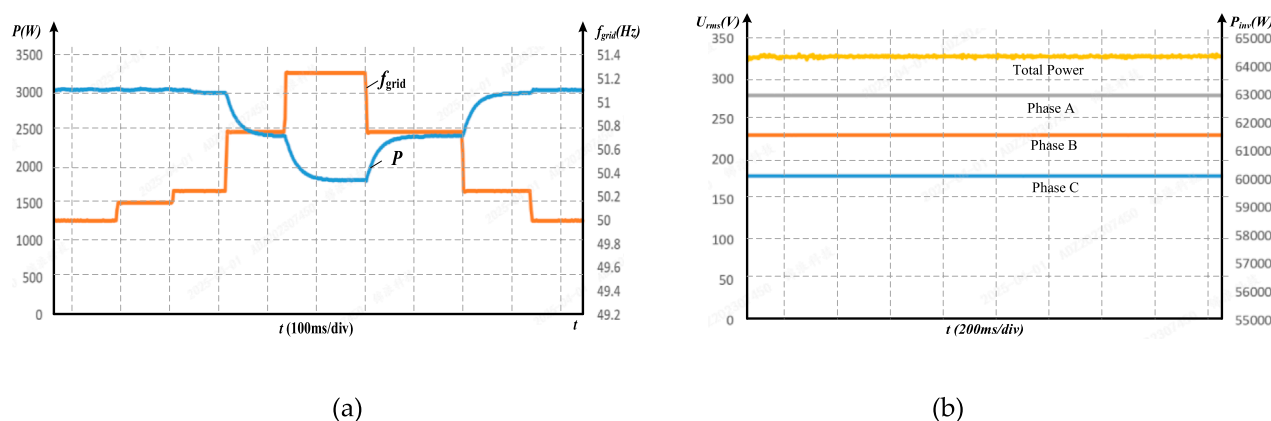


FIGURE 25

Grid frequency fluctuation and voltage imbalance, (a) grid frequency fluctuations, (b) grid voltage imbalance.

(2.2 μ F/450V) costs about ¥0.2. Adding six filter capacitors to a 350 kW PV inverter costs only ¥1.2. Compared to the total cost of ¥30,000, the additional costs are minimal. Traditional methods of adding switching devices and diodes to a dc/ac circuit typically cost more than ¥60. As a result, the economics of the proposed scheme are favorable.

8 Conclusion

In large PV plants, a large number of PV inverters are linked together at the grid-side to generate more power. However, this worsens the ground current problem, which can result in a ground current fault, and cause the inverter system failed and disconnected

from the grid. In this paper, the solutions, including hardware and software, are proposed to suppress the ground current. The hardware solution is to increase the capacitance of filter capacitors connected to the DC-bus, directing more ground current to the inverter rather than the grid to reduce the ground current. Additionally, the software solutions, including carrier phase shifting and carrier frequency shifting, are proposed to suppress ground current. In the software solutions, the ground current for the PV inverters can be cancelled each other, which can alleviate the ground current problem.

First, the system architecture is presented. The concept of connecting the neutral point of the filter capacitors to the DC-bus midpoint is explained. The zero-sequence mathematical model is then built.

The effect of the zero-sequence voltages of the DC/AC converter and the grid on the zero-sequence current is investigated. According to the analysis, when the zero-sequence current is primarily caused by the zero-sequence voltage of the DC/AC converter, increasing the value of the capacitors connected to the DC-bus can greatly improve the effect of the zero-sequence current suppression.

The control architecture is presented. The method to obtain the three-phase output voltage and zero-sequence components for PV inverters is explained. From the analysis, as the number of PV inverter modules increases, the ground current worsens.

Two software solutions, carrier phase shifting and carrier frequency shifting, are proposed to suppress ground current, particularly in systems with a large number of PV inverter modules. The concept and implementation for two proposed software solutions are detailed. From the analysis, the proposed two methods improve the maximum step and variation range of the zero-sequence voltage over the traditional method.

Finally, the experiment is conducted in a PV plant equipped with 20 inverter modules. According to the results of the experiments, the hardware and software methods are both effective. The ground-current fault and the ground current problem can both be avoided with the proposed method.

In the future, as renewable energy becomes more popular, more and more grid-connected inverters will be applied. This proposed method can fundamentally solve the ground current problem, especially in large PV plants equipped a large number of PV inverter modules. The most difficult aspect of the hardware solution is dealing with various scenarios in which leakage current is caused by either the inverter or the grid. As a result, the additional filter capacitors could be installed on either the inverter or grid side. Future work will focus on more practical aspects of this issue. The software solution proposes a method for carrier phase shifting in a multi-inverter system. The most difficult challenge is achieving phase angle synchronization. We did not go into great detail about this topic because of the limited space in this article. Nonetheless, relevant papers will appear in the future.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

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WZ: Methodology, Writing – original draft. DL: Supervision, Writing – review and editing. YW: Conceptualization, Writing – original draft. PX: Software, Writing – review and editing. BL: Validation, Writing – original draft.

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A coordinated power quality improvement control strategy for AC/DC hybrid distribution networks based on three-phase four-leg flexible interconnection converter

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In this paper, A coordinated power quality improvement control strategy for three-phase four-Leg flexible interconnection converter (3P4L-FIC) based AC/DC hybrid distribution networks is proposed, which can address three-phase unbalance issues in AC area and double frequency ripple issue in DC area simultaneously without extra components. Firstly, for the problems of three-phase load unbalance in the AC station area, the split-phase power decoupling control based power quality optimization is delivered for 3P4L-FIC based hybrid distribution networks. Secondly, the doubling frequency pulsation mechanism of DC bus voltage during three-phase unbalanced compensation conditions is analyzed in detail, and a DC voltage doubling frequency ripple migration scheme based on hybrid proportional-integral and resonant controller (PI-R) is proposed. Finally, the 3P4L-FIC based low-voltage multi-distribution area simulation model constructed by Matlab/Simulink is verified, and the results show that the three-phase imbalance of the low-voltage AC line is completely compensated, and the double-frequency ripple of the DC bus voltage is greatly reduced, which realizes the synergistic enhancement of the power quality of the AC-DC distribution area.

KEYWORDS

power quality improvement, three-phase four-bridge-arm converter, unbalance compensation, double-frequency ripple, flexible interconnection

1 Introduction

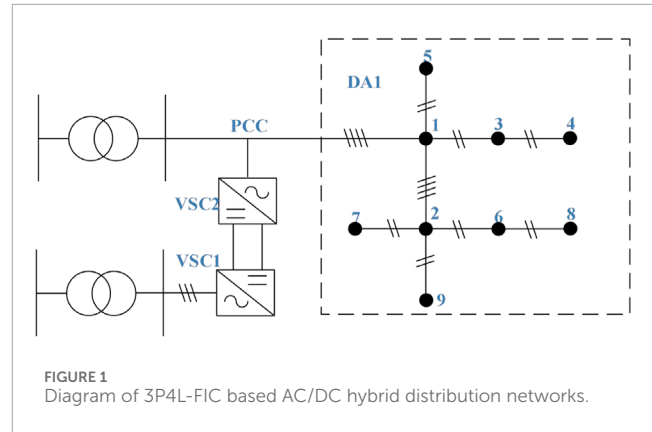
With the advancement of the “dual-carbon” goals, distributed renewable energy and electrified loads are rapidly integrated into low-voltage distribution networks, leading to prominent issues such as three-phase imbalance and voltage violations (Xu et al., 2022; Mocci et al., 2025; Tang et al., 2023). The flexible interconnection system for distribution areas employs power electronic conversion devices to establish flexible DC channels between and within distribution areas (Shi et al., 2023; Liu et al., 2023; Sheng et al., 2024). This system enables power quality management in AC distribution areas, dynamic capacity expansion of distribution areas, and efficient/reliable

accommodation of large-scale high-penetration distributed generation and high-proportion electrified terminal loads (Xiao et al., 2024; Si et al., 2024; Cai et al., 2023; Zheng et al., 2024; Wang et al., 2008).

Compared with conventional three-leg four-wire flexible interconnection converters, three-phase four-leg (3P4L) converters offer advantages including higher DC voltage utilization, reduced requirements for DC-side input capacitance, and effective control of zero-sequence components. These merits have gradually made them a research hotspot for flexible interconnection converters in distribution areas. Shao and Ma (2008), Lu et al. (2020) implemented individual phase voltage control based on established mathematical models of three-phase four-leg inverters, but their control loops lack complete decoupling, resulting in suboptimal performance. Sun et al. (2020), Zhang et al. (2017), Tan et al. (2019) proposed sequence decomposition strategies combined with feedforward methods to address asymmetric output voltage issues through sequence-decoupled control, though grid harmonics were not considered. Yan et al. (2018), Nguyen and Choi (2018) utilized repetitive controllers to suppress harmonic components and improve grid current quality in three-phase four-leg grid-connected inverters, but their control structures and parameter tuning remain relatively complex.

However, the previous studies effectively address unbalanced loads and harmonic suppression on the AC side but fail to consider that double-frequency instantaneous power fluctuations caused by negative-sequence and zero-sequence components under three-phase AC imbalance conditions propagate to the DC side, inducing double-frequency disturbances. Current solutions for double-frequency power decoupling mainly include passive and active power decoupling methods. Passive power decoupling employs large parallel capacitors (Zeng et al., 2018) or LC resonant circuits (Wang et al., 2024) to absorb double frequency power but suffers from high implementation costs and limited effectiveness. Active power decoupling utilizes switching devices to exchange power with decoupling capacitors for pulsating power compensation (Tao et al., 2023; Huang et al., 2023; Zhang et al., 2023; Luo et al., 2023; Fan et al., 2023; Xu et al., 2020). While enhanced switching configurations improve secondary ripple suppression, they simultaneously escalate system power losses. Tan et al. (2025) proposes a control strategy for three-phase four-leg PWM rectifier under unbalanced grid voltage conditions, which aims at achieving not only no second harmonic in dc output voltage, but also unity power factor operation for each phase. Zhou and Lu (2025), Bhowmick et al. (2025) used an active power decoupling (APD) or boost converter circuit to suppress the double-frequency disturbances in DC bus, while the additional APD circuit will increase the system cost.

To address these limitations, this paper proposes a coordinated AC-DC distribution area power quality optimization strategy based on split-phase power decoupling and resonant control, comprehensively considering three-phase imbalance, voltage violations in AC distribution areas, and double-frequency ripple issues in DC distribution areas. The strategy is implemented within the three-phase four-leg flexible interconnected converter (3P4L-FIC) based distribution area architecture shown in Figure 1. Specifically, a power decoupling control strategy based on positive-negative-zero sequence separation is introduced at the inverter side (VSC2) to address the three-phase load unbalance issues.



Simultaneously, a resonant-based double-frequency current suppression strategy is implemented at the flexible interconnection rectifier (VSC1) side to mitigate double-frequency voltage ripples in DC distribution areas. The proposed strategy can address three-phase unbalance issues in AC area and double frequency ripple issue in DC area simultaneously without extra components.

The rest of this study is structured thus. Section 2 presents the mathematical model of 3P4L Converters. Section 3 presents a mechanism of double-frequency DC voltage ripples under unbalanced operating conditions. Section 4 proposes a coordinated power quality improvement strategy for AC-DC distribution areas. Section 5 gives simulation validation and analysis, and Section 6 summarizes the key conclusions drawn from our study.

2 Mathematical model of three-phase four-leg converters

Conventional three-phase three-leg converters exhibit interphase coupling in output power, thereby lacking the capability for independent three-phase power regulation. In contrast, the 3P4L-FIC introduces a zero-sequence leg to the traditional three-leg topology (as illustrated in Figure 2), enabling each port to generate independent voltages, which structural enhancement facilitates decoupled control of unbalanced three-phase power.

Using the current through the filter inductor L_f of the three-phase four-leg inverter $\langle i_f \rangle_{T_s}$, the voltage across the capacitor $\langle u_f \rangle_{T_s}$, and the current $\langle i_g \rangle_{T_s}$ through the load-side filter inductor L_g as state variables, and defining the input voltage $\langle u_{dc} \rangle_{T_s}$, input current $\langle i_p \rangle_{T_s}$, and output voltage $\langle u_{kn} \rangle_{T_s}$ as output variables ($k = a, b, c$), the state-space equations can be formulated as follows

$$\frac{d}{dt} \begin{bmatrix} \langle i_{af} \rangle_{T_s} \\ \langle i_{bf} \rangle_{T_s} \\ \langle i_{cf} \rangle_{T_s} \end{bmatrix} + \frac{R}{L_f} \begin{bmatrix} \langle i_{af} \rangle_{T_s} \\ \langle i_{bf} \rangle_{T_s} \\ \langle i_{cf} \rangle_{T_s} \end{bmatrix} = \frac{1}{L_f} \begin{bmatrix} d_{af} \langle V_{dc} \rangle_{T_s} \\ d_{bf} \langle V_{dc} \rangle_{T_s} \\ d_{cf} \langle V_{dc} \rangle_{T_s} \end{bmatrix} - \frac{1}{L_f} \begin{bmatrix} \langle u_{af} \rangle_{T_s} \\ \langle u_{bf} \rangle_{T_s} \\ \langle u_{cf} \rangle_{T_s} \end{bmatrix} + \frac{L_n}{L_f} \frac{d \langle i_n \rangle_{T_s}}{dt} \quad (1)$$

$$\frac{d}{dt} \begin{bmatrix} \langle u_{af} \rangle_{T_s} \\ \langle u_{bf} \rangle_{T_s} \\ \langle u_{cf} \rangle_{T_s} \end{bmatrix} = \frac{1}{C_f} \begin{bmatrix} \langle i_{af} \rangle_{T_s} \\ \langle i_{bf} \rangle_{T_s} \\ \langle i_{cf} \rangle_{T_s} \end{bmatrix} - \frac{1}{C_f} \begin{bmatrix} \langle i_{ag} \rangle_{T_s} \\ \langle i_{bg} \rangle_{T_s} \\ \langle i_{cg} \rangle_{T_s} \end{bmatrix} \quad (2)$$

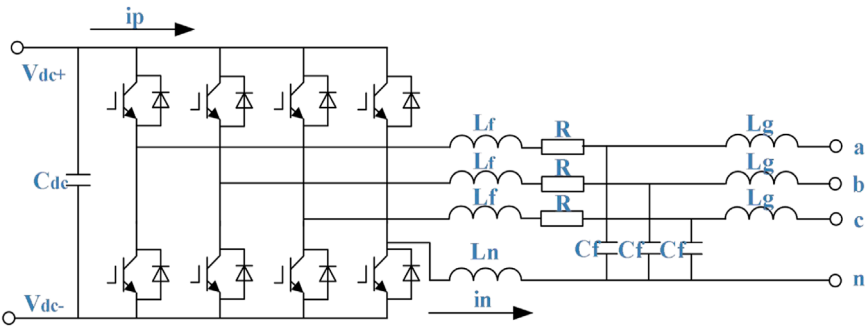


FIGURE 2
Topology architecture of three-phase four-leg converter.

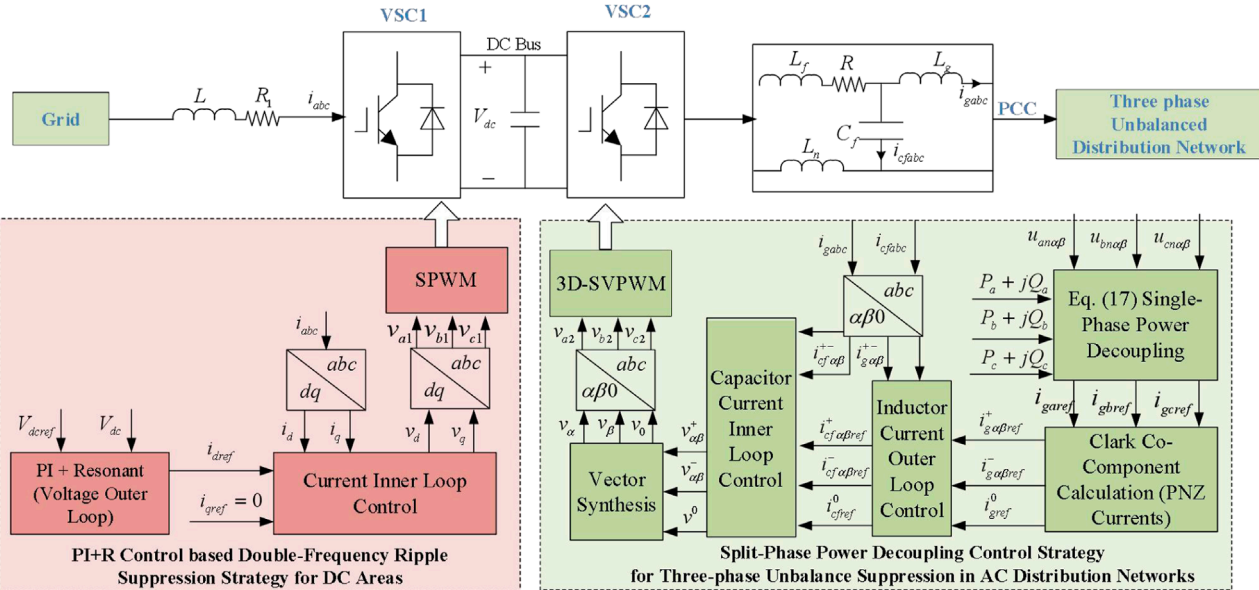


FIGURE 3
Overall diagram of the proposed coordinated power quality improvement strategy for 3P4L-FIC based AC-DC distribution areas.

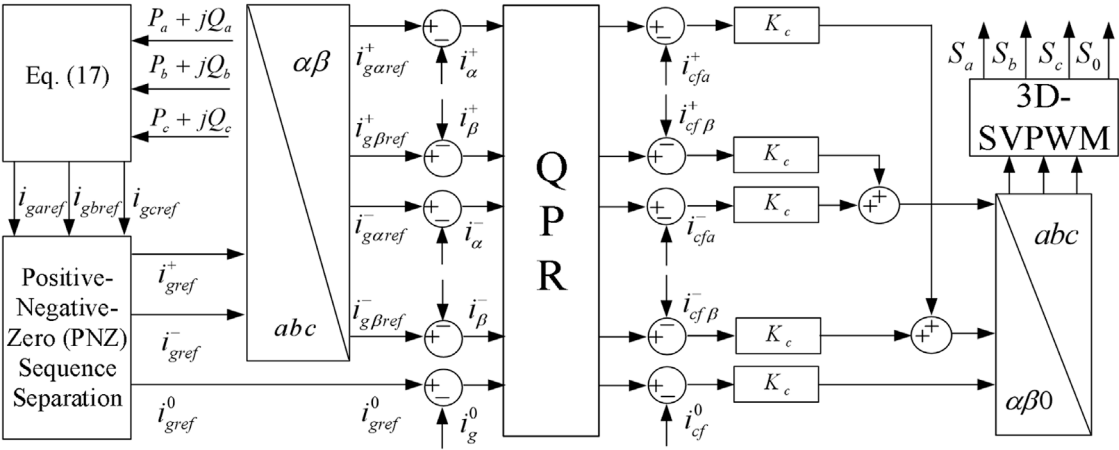


FIGURE 4
The overall control block diagram of the proposed split-phase power decoupling strategy in VSC2.

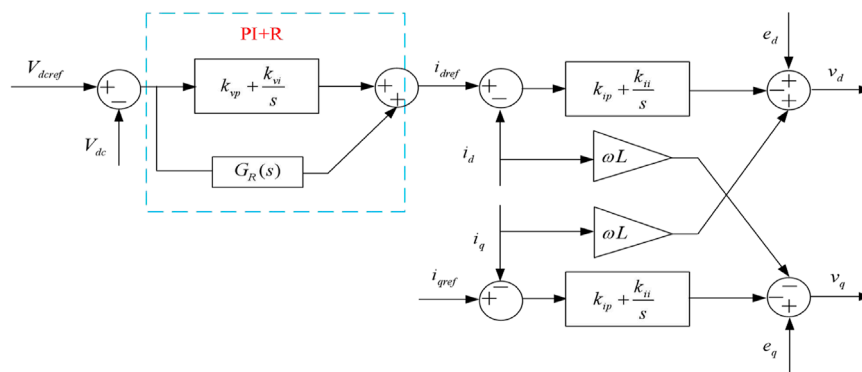


FIGURE 5

The overall control block diagram of PI-R control based double-frequency ripple suppression strategy in VSC1.

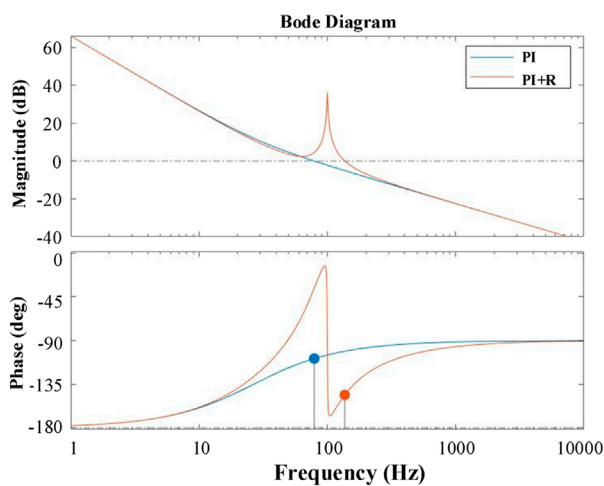


FIGURE 6

Comparison bode diagram between PI and PI-R outer loop.

TABLE 1 System parameters.

Parameters	Value/Unit	Parameters	Value/Unit
L	7e-3/H	R	1e-2/Ω
R _l	1e-2/Ω	C _f	5e-6/F
C _{dc}	3.8e-3/F	L _g	1e-3/H
L _f	5e-3/H	L _n	7e-4/H

TABLE 2 Control parameters.

Parameters	Value	Parameters	Value
K _p	1	K _{rr}	12
ω _{cr}	5	ω _{0r}	314

$$\frac{d}{dt} \begin{bmatrix} \langle i_{ag} \rangle_{T_s} \\ \langle i_{bg} \rangle_{T_s} \\ \langle i_{cg} \rangle_{T_s} \end{bmatrix} = \frac{1}{L_g} \begin{bmatrix} \langle u_{af} \rangle_{T_s} \\ \langle u_{bf} \rangle_{T_s} \\ \langle u_{cf} \rangle_{T_s} \end{bmatrix} - \frac{1}{L_g} \begin{bmatrix} \langle u_{an} \rangle_{T_s} \\ \langle u_{bn} \rangle_{T_s} \\ \langle u_{cn} \rangle_{T_s} \end{bmatrix} \quad (3)$$

Where i_{kf} and i_{kg} represent the currents flowing through inductors L_f and L_g in phase k , respectively. u_{kf} and u_{kn} denote the voltage across capacitor C_f and the converter output voltage in phase k , respectively. d_{kf} ($k = a, b, c$) is the phase duty cycle, derived by averaging the switching functions of the bridge-arm switches over a switching period and filtering out switching-frequency harmonic components. i_n corresponds to the current flowing through inductor L_n .

From Equations 1–3, it is evident that if the neutral inductor and its equivalent resistance are neglected, the three-phase four-leg inverter system reduces to three independent single-phase full-bridge inverter systems. When the neutral inductor is considered, however, mutual coupling arises between the bridge legs, preventing the system from being treated as three decoupled inverter units. This observation underscores the inherent difficulty in achieving decoupled control of the three-phase four-leg inverter in the three-phase stationary coordinate frame.

Translating the fundamental sinusoidal components from the three-phase stationary coordinate frame into a two-phase stationary coordinate frame, followed by Laplace transformation, yields the system's frequency-domain control model, expressed as

$$s \begin{bmatrix} \langle i_{af} \rangle_{T_s} \\ \langle i_{bf} \rangle_{T_s} \\ \langle i_{cf} \rangle_{T_s} \end{bmatrix} + \frac{R}{L_f} \begin{bmatrix} \langle i_{af} \rangle_{T_s} \\ \langle i_{bf} \rangle_{T_s} \\ \langle i_{cf} \rangle_{T_s} \end{bmatrix} = \frac{1}{L_f} \begin{bmatrix} d_{af} \langle V_{dc} \rangle_{T_s} \\ d_{bf} \langle V_{dc} \rangle_{T_s} \\ d_{cf} \langle V_{dc} \rangle_{T_s} \end{bmatrix} - \frac{1}{L_f} \begin{bmatrix} \langle u_{af} \rangle_{T_s} \\ \langle u_{bf} \rangle_{T_s} \\ \langle u_{cf} \rangle_{T_s} \end{bmatrix} + \frac{3L_n s}{L_f} \langle i_0 \rangle_{T_s} \quad (4)$$

$$s \begin{bmatrix} \langle u_{af} \rangle_{T_s} \\ \langle u_{bf} \rangle_{T_s} \\ \langle u_{cf} \rangle_{T_s} \end{bmatrix} = \frac{1}{C_f} \begin{bmatrix} \langle i_{af} \rangle_{T_s} \\ \langle i_{bf} \rangle_{T_s} \\ \langle i_{cf} \rangle_{T_s} \end{bmatrix} - \frac{1}{C_f} \begin{bmatrix} \langle i_{ag} \rangle_{T_s} \\ \langle i_{bg} \rangle_{T_s} \\ \langle i_{cg} \rangle_{T_s} \end{bmatrix} \quad (5)$$

$$s \begin{bmatrix} \langle i_{ag} \rangle_{T_s} \\ \langle i_{bg} \rangle_{T_s} \\ \langle i_{cg} \rangle_{T_s} \end{bmatrix} = \frac{1}{L_g} \begin{bmatrix} \langle u_{af} \rangle_{T_s} \\ \langle u_{bf} \rangle_{T_s} \\ \langle u_{cf} \rangle_{T_s} \end{bmatrix} - \frac{1}{L_g} \begin{bmatrix} \langle u_{an} \rangle_{T_s} \\ \langle u_{bn} \rangle_{T_s} \\ \langle u_{cn} \rangle_{T_s} \end{bmatrix} \quad (6)$$

TABLE 3 Three phase load and photovoltaic access data for distributor area I (Scenario 1).

Nodes	Load/kW			PV/kW		
	Phase A	Phase B	Phase C	Phase A	Phase B	Phase C
1	1.22	1.22	1.22	7.3	3.65	0
2	1.22	1.22	1.22	0	3.65	1.825
3	0	1.22	0	0	1.825	0
4	0	1.22	0	0	1.825	0
5	4.88	0	0	7.3	0	0
6	0	0	1.22	0	0	1.825
7	1.22	0	0	1.825	0	0
8	0	0	1.22	0	0	0
9	0	1.22	0	0	1.825	0

TABLE 4 Three phase load and photovoltaic access data for distribution area I (Scenario 2).

Nodes	Load (kW)			PV(kW)		
	Phase A	Phase B	Phase C	Phase A	Phase B	Phase C
1	3.319	3.319	3.319	0.409	0.103	0
2	3.319	3.319	3.319	0	0.103	0.103
3	0	3.319	0	0	0.103	0
4	0	3.319	0	0	0.103	0
5	13.28	0	0	0.409	0	0
6	0	0	3.319	0	0	0.103
7	3.319	0	0	0.103	0	0
8	0	0	3.319	0	0	0
9	0	3.319	0	0	0.103	0

It can be seen from the Equations 4–6 that under small-signal perturbations around the quiescent operating point, since the DC voltage V_{dc} is regulated under constant-voltage control by Distribution Area II (DA II), the perturbation in V_{dc} is neglected during small-signal analysis of the inverter. Consequently, by separating the DC components and eliminating second-order differential terms at the quiescent operating point, the small-signal model in the rotating coordinate frame is derived as

$$\left(s + \frac{R}{L_f}\right) \begin{bmatrix} \hat{i}_{\alpha f} \\ \hat{i}_{\beta f} \\ \hat{i}_{0f} \end{bmatrix} = \frac{1}{L_f} \begin{bmatrix} d_{\alpha f} V_{dc} \\ d_{\beta f} V_{dc} \\ d_{0f} V_{dc} \end{bmatrix} - \frac{1}{L_f} \begin{bmatrix} \hat{u}_{\alpha f} \\ \hat{u}_{\beta f} \\ \hat{u}_{0f} \end{bmatrix} + \frac{3L_n s}{L_f} \hat{i}_0 \quad (7)$$

$$s \begin{bmatrix} \hat{u}_{\alpha f} \\ \hat{u}_{\beta f} \\ \hat{u}_{0f} \end{bmatrix} = \frac{1}{C_f} \begin{bmatrix} \hat{i}_{\alpha f} \\ \hat{i}_{\beta f} \\ \hat{i}_{0f} \end{bmatrix} - \frac{1}{C_f} \begin{bmatrix} \hat{i}_{\alpha g} \\ \hat{i}_{\beta g} \\ \hat{i}_{0g} \end{bmatrix} \quad (8)$$

$$s \begin{bmatrix} \hat{i}_{\alpha f} \\ \hat{i}_{\beta f} \\ \hat{i}_{0f} \end{bmatrix} = \frac{1}{L_g} \begin{bmatrix} \hat{u}_{\alpha f} \\ \hat{u}_{\beta f} \\ \hat{u}_{0f} \end{bmatrix} - \frac{1}{L_g} \begin{bmatrix} \hat{u}_{\alpha n} \\ \hat{u}_{\beta n} \\ \hat{u}_{0n} \end{bmatrix} \quad (9)$$

From Equations 7–9, it can be observed that the mathematical model of the three-phase four-leg converter is mutually decoupled in the α axis, β axis and 0 -axis, which facilitates controller design and simplifies the implementation of three-phase decoupled control. Furthermore, the mathematical model of the $\alpha\beta$ -axis is identical to

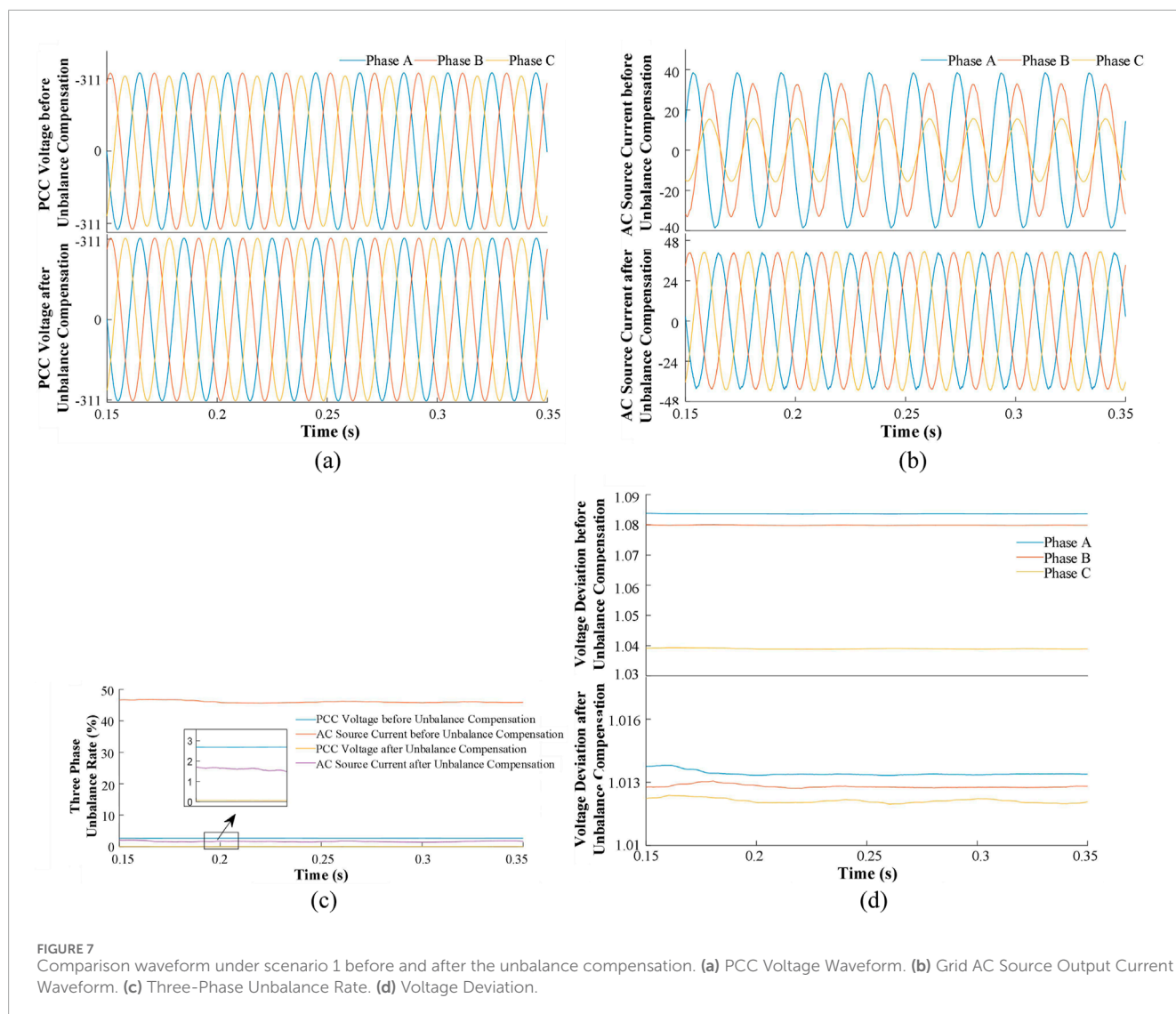


FIGURE 7

Comparison waveform under scenario 1 before and after the unbalance compensation. (a) PCC Voltage Waveform. (b) Grid AC Source Output Current Waveform. (c) Three-Phase Unbalance Rate. (d) Voltage Deviation.

that of a conventional three-phase three-leg converter, while the 0 -axis mathematical model incorporates coupling with three times the zero-sequence inductance L_n .

3 Mechanism of DC double frequency voltage ripple under unbalanced operating conditions

The three-phase four-leg inverter is directly connected to the low-voltage distribution network via an *LCL* filter. In the three-phase four-wire system, due to the inclusion of zero-sequence voltage and current components, the system's instantaneous real power p , instantaneous imaginary power q , and instantaneous zero-sequence power p_0 are defined (Akagi et al., 2007) as follows (with coordinate transformation applied for constant-power transformation)

$$\begin{bmatrix} p_0 \\ p \\ q \end{bmatrix} = \begin{bmatrix} v_0 & 0 & 0 \\ 0 & v_\alpha & v_\beta \\ 0 & v_\beta & -v_\alpha \end{bmatrix} \begin{bmatrix} i_0 \\ i_\alpha \\ i_\beta \end{bmatrix} \quad (10)$$

It can be seen from Equation 10 that the zero-sequence instantaneous power in the system is independent of the positive- and negative-sequence instantaneous powers. According to (Dong et al., 2020; Gao et al., 2016), the instantaneous complex power output from the converter's AC side can be expressed as (excluding the zero-sequence component).

$$\begin{cases} p = P_0 + P_{c2} \cos(2\omega t) + P_{s2} \sin(2\omega t) \\ q = Q_0 + Q_{c2} \cos(2\omega t) + Q_{s2} \sin(2\omega t) \end{cases} \quad (11)$$

In Equation 11, P_0 and Q_0 represent the DC components of the instantaneous active and reactive power, respectively. P_{c2} and P_{s2}

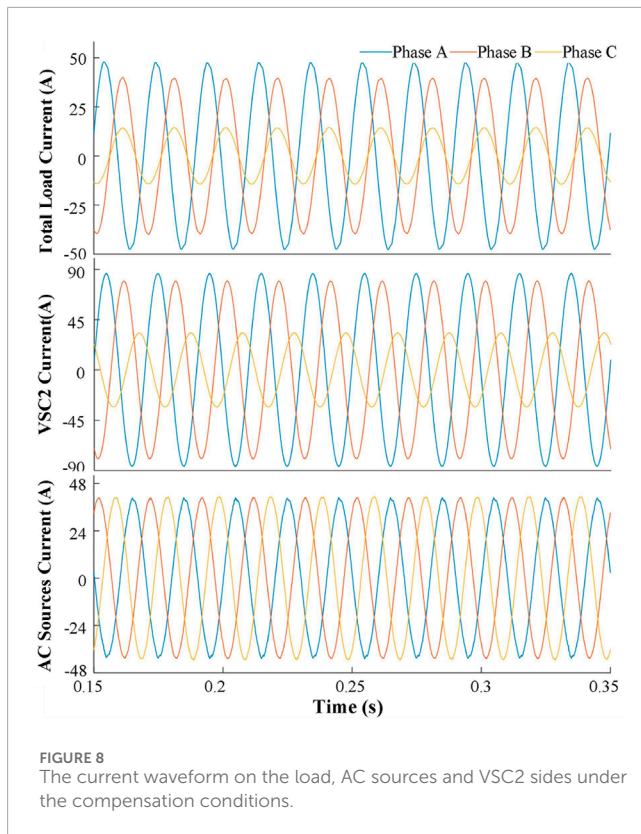


FIGURE 8
The current waveform on the load, AC sources and VSC2 sides under the compensation conditions.

denote the amplitudes of the double-frequency AC components in the instantaneous active power, while Q_{c2} and Q_{s2} correspond to those in the instantaneous reactive power.

The zero-sequence voltage and current components are defined as

$$\begin{cases} u_k^0 = U_0 \cos(\omega t) \\ i_k^0 = I_0 \cos(\omega t + \varphi) \end{cases} \quad (k = a, b, c) \quad (12)$$

In the Equation 12, U_0 and I_0 denote the magnitudes of the zero-sequence voltage and current vectors, respectively. φ represents the phase angle between the zero-sequence voltage and current vectors, and ω is the rotational angular velocity. Therefore, the instantaneous zero-sequence power can be expressed as

$$\begin{cases} p_0 = 3U_0I_0 \cos \varphi - 3U_0I_0 \cos(2\omega t + \varphi) \\ q_0 = U_0I_0 \sin(2\omega t + \varphi) \end{cases} \quad (13)$$

In the equations, p_0 denotes the zero-sequence instantaneous active power, and q_0 represents the zero-sequence instantaneous reactive power. From Equation 13, it is evident that the zero-sequence power itself cannot produce constant power. Therefore, there is no method to eliminate the oscillatory components while retaining only the DC component.

By integrating the above analysis, the total instantaneous power of the three-phase four-wire system is expressed as

$$\begin{cases} p = P_0 + P_{c2} \cos(2\omega t) + P_{s2} \sin(2\omega t) + p_0 \\ q = Q_0 + Q_{c2} \cos(2\omega t) + Q_{s2} \sin(2\omega t) + q_0 \end{cases} \quad (14)$$

From Equation 14, under unbalanced operating conditions, the instantaneous power fluctuations at the second harmonic frequency generated by the negative-sequence and zero-sequence components will penetrate into the DC side, causing second harmonic frequency disturbances in the DC network. This seriously affects the reliability and stability of the system.

4 Coordinated power quality improvement strategy for 3P4L-FIC based AC-DC distribution areas

Figure 3 shows the overall block diagram of the proposed coordinated power quality improved control strategy for 3P4L-FIC distribution areas, which consists of two part. One is the split phase power decoupling control strategy for three phase unbalance compensation of the AC distribution network, which is based on the positive, negative, and zero sequence separation control on the inverter side (VSC2) of the 3P4L-FIC. The other one is hybrid proportional-integral and resonant controller (PI-R) based double-frequency voltage ripple suppression strategy for DC-bus, which is implemented on the rectifier side (VSC1) of the 3P4L-FIC. The proposed coordinated strategy can address three-phase unbalance issues in AC area and double frequency ripple issue in DC area simultaneously without extra components.

4.1 Split-phase power decoupling control based three-phase unbalance compensation strategy for AC distribution areas

Based on the mathematical model of the 3P4L-FIC established in Section 2 under the $\alpha\beta 0$ coordinate frame, and combined with the single-phase power decoupling calculation in Equation 17, the reference command current for split-phase power decoupling is derived. This current is then transformed into positive-, negative-, and zero-sequence reference command currents via Clark transformation, thereby achieving the split-phase power decoupling control strategy. The overall control block diagram of the split-phase power decoupling for VSC2 is shown in Figure 4.

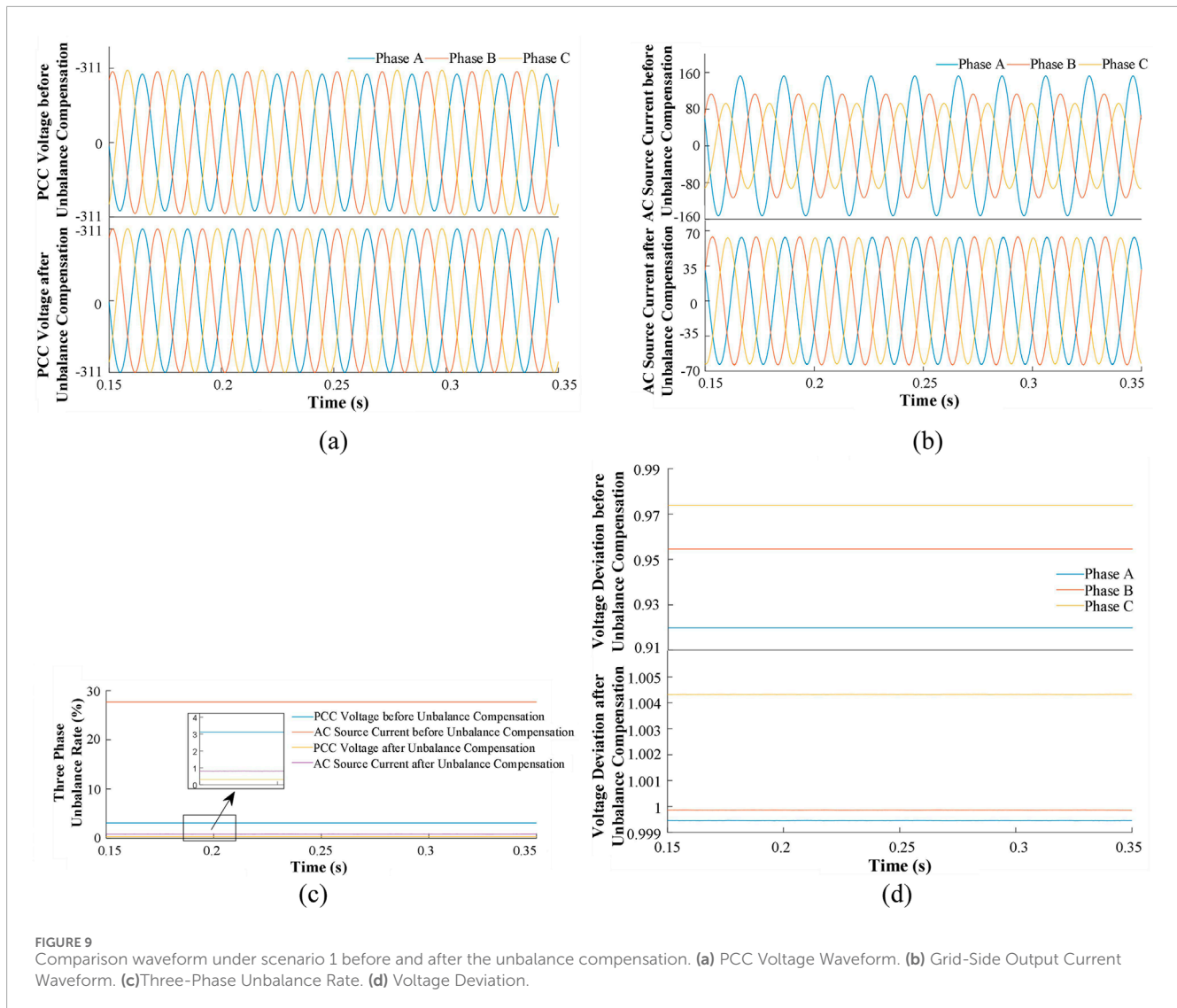
In the diagram, K_r is the proportional coefficient, and QPR denotes the Quasi-Proportional Resonant (QPR) controller, whose transfer function is expressed as:

$$G_{qpr}(s) = k_{pr} + \frac{2k_{rr}\omega_{cr}s}{s^2 + 2\omega_{cr}s + \omega_0^2} \quad (15)$$

In Equation 15, k_{pr} is the proportional coefficient, k_{rr} is the gain coefficient, ω_{cr} is the bandwidth coefficient, ω_0 is the resonant frequency.

Based on the generalized instantaneous reactive power theory for single-phase systems (Akagi et al., 2007), the $\alpha\beta$ two-phase signals are constructed to obtain the single-phase instantaneous active power and reactive power.

$$\begin{bmatrix} p \\ q \end{bmatrix} = \frac{1}{2} \begin{bmatrix} u_\alpha & u_\beta \\ u_\beta & -u_\alpha \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (16)$$



In Equation 16 u_α , u_β , i_α and i_β represent the $\alpha\beta$ -axis voltage and current phasors at the PCC, constructed via SOGI, for single-phase voltage and current, respectively.

Due to the advantages of the Second-Order Generalized Integrator (SOGI), such as harmonic filtering, the SOGI (Wei et al., 2016) is adopted to construct the $\alpha\beta$ -axis signals for u_{an} , u_{bn} , and u_{cn} , i.e., $u_{an\alpha}$, $u_{an\beta}$, $u_{bn\alpha}$, $u_{bn\beta}$, $u_{cn\alpha}$ and $u_{cn\beta}$. Therefore, based on the generalized instantaneous power theory for single-phase systems, the reference current commands for single-phase P/Q control are derived as

$$i_{kref} = 2 * \frac{u_{kna}p + u_{kn\beta}q}{u_{kna}^2 + u_{kn\beta}^2} (k = a, b, c) \quad (17)$$

4.2 PI-R control based double-frequency ripple suppression strategy for DC areas

The emergence of double-line-frequency voltage ripple components on the DC side, combined with the insufficient

gain of conventional rectifier control loops at double-line-frequency, results in the inability of standard rectifier control systems to maintain stable DC voltage output under unbalanced load conditions. According to the Internal Model Principle, to achieve zero steady-state error tracking of the double-line-frequency pulsating components in the output voltage caused by load imbalance, the controller must incorporate sinusoidal internal models at the corresponding frequencies. This ensures that the open-loop magnitude-frequency characteristic curve of the system exhibits infinite gain at double-line-frequency (Zou et al., 2014; Ortega et al., 2021).

To address the double-line-frequency disturbances in the DC bus caused by three-phase imbalance in the AC distribution area, this paper employs a damped resonant controller to suppress the double-line-frequency pulsating components within the control loop. A voltage outer loop PI-R control strategy is proposed, with its expression given in Equation 17. The overall control block diagram of the VSC1 ripple suppression strategy is illustrated in Figure 5.

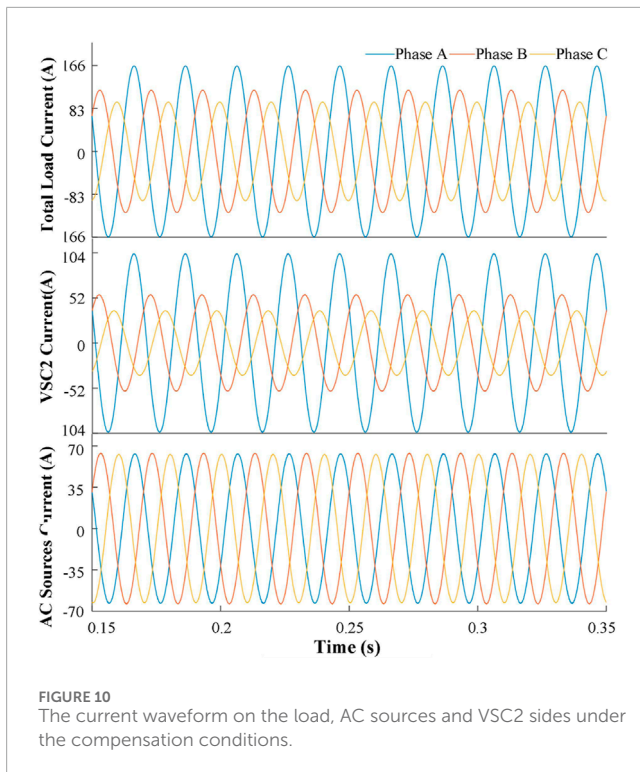


FIGURE 10
The current waveform on the load, AC sources and VSC2 sides under the compensation conditions.

$$G_{pir}(s) = k_p + \frac{k_i}{s} + \frac{2k_r\omega_c s}{s^2 + 2\omega_c s + \omega_0^2} \quad (18)$$

In Equation 18, k_p is the proportional coefficient, k_i is the integral coefficient, k_r is the gain coefficient, ω_c is the bandwidth coefficient, ω_0 is the resonant frequency.

Figure 6 shows the comparison of voltage loop frequency response characteristics between the conventional voltage outer loop and the inverter with the introduced resonant controller. The parameters of the PI-R controller are set as follows: proportional coefficient $k_p = 1.787$, integral coefficient $k_i = 300$, gain coefficient $k_r = 150$, bandwidth coefficient $\omega_c = \pi$, and resonant frequency is $2\pi \cdot 2f$. From the Bode plot in Figure 6, it can be observed that after introducing the proposed resonant controller for double-frequency ripple suppression, the magnitude and phase frequency curves overlap with those of the traditional single-PI voltage outer loop in the low-frequency region. However, at twice the line frequency (100 Hz) of the inverter output, the PI + R control significantly increases the magnitude gain in the Bode plot, effectively enhancing the voltage loop's control over the double-frequency AC components in the output voltage caused by unbalanced loads. Additionally, the system achieves a cutoff frequency of 137 Hz and a phase margin of 34° , which satisfies the stability requirements.

5 Simulation verification and analysis

A simulation model of the dual-terminal low-voltage flexible interconnected grid, as shown in Figure 1, was developed on the MATLAB platform. The converter system parameters and control parameters are summarized in Tables 1, 2. Based on measurement data from a demonstration construction project in a city in

Guizhou Province, this study analyzes whether the distribution area experiences operational issues such as three-phase imbalance and voltage deviation under typical conditions, with the power factor set to 0.9. The operational data for Distribution Area I are as follows:

Scenario 1: Data from 12:00 noon are selected as a typical case of high photovoltaic (PV) penetration and low load proportion, three phase load and photovoltaic access data for substation area I are summarized in Table 3.

Scenario 2: Data from 18:00 (6:00 p.m.) are selected as a typical case of high load proportion and low PV penetration. The three phase load and photovoltaic access data for substation area I are summarized in Table 4.

5.1 Verification of power quality optimization strategy for AC distribution areas

Scenario 1: Figures 7a–c illustrate the PCC voltage waveforms, source-side output current waveforms, and three-phase unbalance rates before and after implementing the proposed three-phase unbalance compensation control strategy. When the compensation strategy is blocked, the distribution area operated independently, with Distribution Area I bearing the full three-phase unbalanced load. The three-phase unbalance rate of its output currents entirely depended on the load level, resulting in a PCC voltage unbalance rate of approximately 3% and a source-side output current unbalance rate as high as 47%, exceeding the 2% three-phase unbalance limit. This severely compromised the stable operation of the distribution area. After enable the compensation strategy, the PCC voltage unbalance rate was reduced to nearly 0%, and the source-side current unbalance rate decreased to around 1.7%, significantly mitigating the three-phase imbalance and complying with the system's unbalance limit requirements.

Figure 7d shows the voltage deviation value before and after unbalance compensation. During the high PV penetration and low load proportion period (midday), the PV generation exceeded the load demand, causing reverse power flow and overvoltage at the PCC. As shown in Figure 7d, the voltage deviations without compensation for Phases A, B, and C were 1.084 pu, 1.08 pu, and 1.039 pu, respectively. The deviations for Phases A and B exceeded the upper limit of 1.07 pu, risking insulation damage to equipment. When the compensation strategy is enabled, Distribution Area I transferred a portion of PV generation to Distribution Area II, reducing its voltage deviation. The PCC phase voltage deviations stabilized around 1.013 pu, closer to the nominal value of 1 pu, thereby enhancing equipment safety.

Figure 8 shows the current waveforms at the source-side, load-side, and inverter-side under the compensation conditions. As demonstrated in Figure 8, after implementing the compensation control strategy, the inverter-side output current compensates for the unbalanced load-side output current, thereby reducing the three-phase unbalance rate of the source-side output current and mitigating its adverse impact on system stability.

Scenario 2: Figures 9a–c illustrate the PCC voltage waveforms, source-side output current waveforms, and three-phase unbalance rates before and after implementing the proposed three-phase unbalance compensation control strategy. As shown in Figures 9a–c,

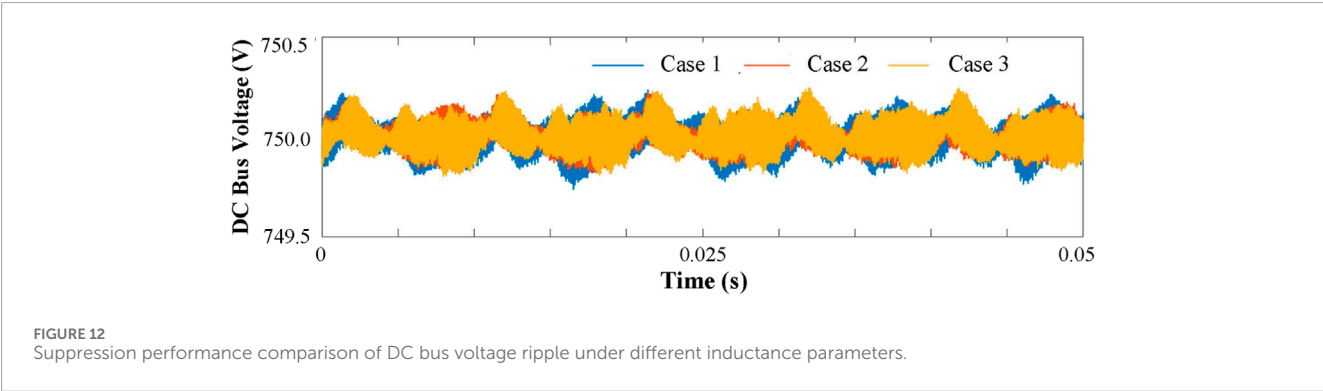
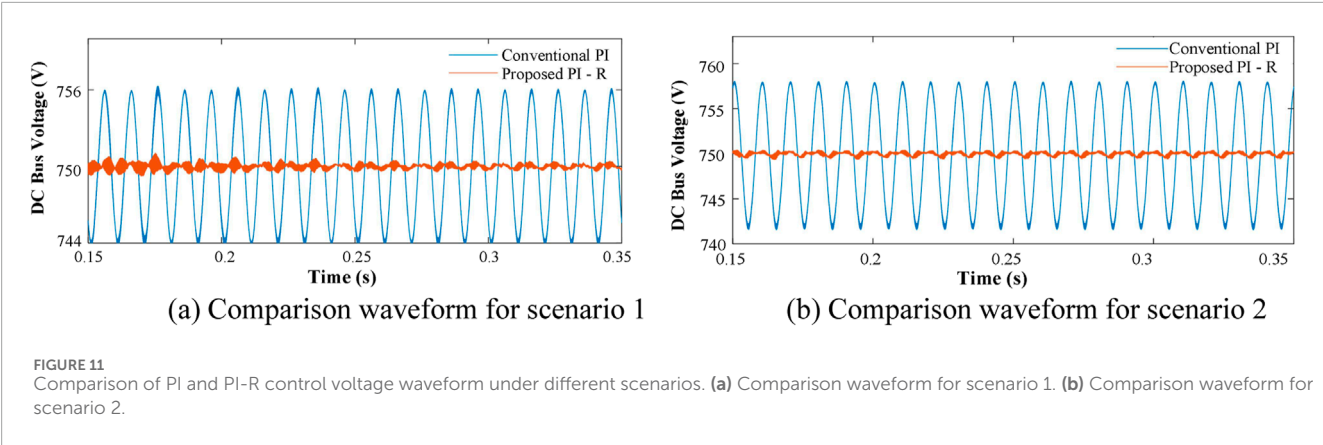


TABLE 5 The performance comparison for conventional and proposed strategy.

Scenario 1		Scenario 2	
Three-Phase Unbalance Rate	Before → After	Three-Phase Unbalance Rate	Before → After
PCC Voltage Unbalance Rate (%)	3%→0%	PCC Voltage Unbalance Rate (%)	3.1%→0.1%
Source-Side Output Current Unbalance Rate	47%→1.7%	Source-Side Output Current Unbalance Rate	27%→0.9%
Voltage Deviation Analysis	—	Voltage Deviation Analysis	—
Phase A	1.084→1.014	Phase A	0.92→0.9995
Phase B	1.08→1.013	Phase B	0.956→0.9998
Phase C	1.039→1.012	Phase C	0.974→1.0042
Double-Frequency Ripple Suppression Effect: Conventional PI Control: 12 V → PI-R Control: ~2 V		Double-Frequency Ripple Suppression Effect: Conventional PI Control: 14 V → PI-R Control: ~1 V	

under Scenario two operating conditions, the PCC voltage unbalance rate is approximately 3.1%, and the source-side output current unbalance rate reaches 27%, both exceeding the three-phase unbalance limit. After enable the compensation strategy, the PCC voltage unbalance rate is reduced to around 0.1%, and the source-side current unbalance rate decreases to 0.9%, effectively mitigating the three-phase imbalance and meeting the system's unbalance limit requirements. Figure 9d shows the voltage deviation waveforms before and after the unbalance compensation. During the low PV penetration and high load proportion period (evening),

reduced PV generation combined with increased load demand leads to under voltage at the PCC. As seen in Figure 9d, the voltage deviations without compensation for Phases A, B, and C are 0.92 pu, 0.956 pu, and 0.974 pu, respectively. The deviations for Phases A and B fall below the lower voltage deviation limit (0.97 pu). When the compensation strategy is enabled, the PCC phase voltage deviations stabilize near 1 pu, ensuring safer operation of the distribution area. Figure 10 shows the current waveforms on the source side, the load side, and the inverter side under the compensation conditions. It can be seen that, after implementing

the compensation control strategy, the output current on the inverter side compensates for the unbalanced output current on the load side, achieving three-phase balance on the source side.

5.2 Verification of the double-frequency ripple suppression strategy

Figure 11a,b show the double-line-frequency waveforms of the DC bus voltage under Scenarios one and 2, respectively, comparing traditional PI control with the PI-R control. As shown in Figures 11, 12, the conventional PI voltage outer loop fails to suppress the double-line-frequency pulsating components in the DC voltage. In contrast, the PI-R control effectively mitigates the double-frequency components by introducing a resonant controller into the traditional control loop, while maintaining excellent steady-state characteristics.

To validate the proposed strategy sensitive to the change of the inductance value, the suppression performance of DC bus voltage ripple under different inductance parameters are carried out. The results are plotted in Figure 12, in which three cases are set. All the inductance values both in inverter and rectifier side in Case 1 are reduced by 50%, and increase 50% in Case 3, and Case 2 keep the same value as Table 1. It can be seen that under different inductance variations, the double-frequency peak-to-peak ripple in dc bus voltage can be suppressed by the proposed method within the range of ± 1 V, which fully demonstrates that the proposed strategy sensitive to the change of the inductance value.

Table 5 summarizes the performance comparison between the proposed control strategy and the traditional method under two scenarios. The results show that the proposed split-phase power decoupling method can reduce the power unbalance degree at the AC sources to less than 2%, which can ensure the safe and balanced operation of the AC transformer. Moreover, the proposed method can compensate for the voltage unbalance degree at the PCC point and simultaneously solve the voltage deviation problem. The proposed double-frequency ripple suppression method can effectively suppress the second - harmonic voltage, reducing the peak-peak DC voltage ripple pulsation to around 1–2 V at the minimum. The results fully demonstrate the effectiveness of the proposed control strategy.

6 Conclusion

This paper proposes a coordinated power quality improvement strategy for 3P4L-FCI based AC-DC distribution areas. To address issues such as three-phase imbalance and voltage limit violations in AC distribution areas, a split-phase power decoupling control strategy utilizing the three-phase four-leg converter is developed, effectively reducing the PCC voltage unbalance rate, source-side output current unbalance rate, and voltage deviations under two extreme operating conditions. Furthermore, by analyzing the double-line-frequency pulsation mechanism of the DC bus voltage under unbalanced AC distribution conditions, a double-frequency ripple suppression scheme based on a hybrid proportional-integral and resonant controller (PI-R) controller is proposed, which significantly mitigates the double-frequency components in the DC

bus voltage. Finally, a MATLAB/Simulink-based simulation model of 3P4L-FCI based AC-DC distribution areas is established. The results validate the effectiveness of the proposed control strategies.

Data availability statement

The original contributions presented in the study are included in the article/supplementary material, further inquiries can be directed to the corresponding author.

Author contributions

QF: Conceptualization, Writing – original draft. JL: Conceptualization, Writing – review and editing. JZ: Conceptualization, Methodology, Writing – review and editing. YX: Funding acquisition, Supervision, Writing – review and editing. QH: Data curation, Investigation, Writing – review and editing. YG: Investigation, Validation, Writing – review and editing.

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Conflict of interest

Authors QF, JZ, YX, and YG were employed by Electric Power Research Institute of Guizhou Power Grid Co. Ltd. Authors JL and QH were employed by XJ Electric Co. Ltd.

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Generative AI statement

The authors declare that no Generative AI was used in the creation of this manuscript.

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Unit power rating balancing for differential-power-processing-based distributed photovoltaic systems

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Differential power processing (DPP) architectures are effective solutions for photovoltaic (PV) systems experiencing uncertainties in environmental conditions, such as non-uniform irradiation, which can significantly degrade electricity production efficiency and operating safety. Among the different architectures available, the PV-to-bus DPP configuration has shown excellent performances for control flexibility and galvanic isolation capability. However, recent control schemes for PV-to-bus architectures offer lower-than-desirable efficiency enhancements, such as those involving implementation complexities and unit power distributions. Hence, a unit power rating balancing (UPRB) scheme is proposed in this work to reconfigure the distribution of differential power among the unit converters with the aim of ensuring submodule-level optimization and enhanced performance of the entire PV system. The proposed UPRB scheme integrates perturbation-and-observation-based maximum power-point tracking units to maximize the energy harvested from PV modules, and the unit balance-point tracking unit is employed to determine the optimal string current reference directly for mitigating the unbalanced differential power in the DPP units. By suppressing the maximum processed power in each DPP unit, the capital cost and system size can be reduced. Simulation and experimental studies were conducted, whose evaluation results support the applicability of the proposed control scheme.

KEYWORDS

differential power processing, photovoltaic system, energy harvesting, unit power rating balancing, efficiency

1 Introduction

Recently, solar power generation programs have gained widespread acceptance globally given the advancements in photovoltaic (PV) technology (Blaabjerg et al., 2023). In typical solar farms and distributed PV systems, the PV panels are first connected in series to enhance the voltage for achieving the required value and then connected in parallel for construction of the PV generation grid (Zhu et al., 2024b). However, the uncertainties associated with environmental conditions, such as partial shading, dust deposition, and cell aging, can easily cause series string mismatches between the panels and reduce the total power generated by the distributed PV system (Aifan et al., 2024; Zhu et al., 2024a). Typically, bypass diodes are used to protect a PV substring with 20–24 cells. However,

the power generated still exhibits a significant reduction trend under various mismatch conditions owing to the occurrence of multiple maximum power points (MPPs) on the power–voltage curve (Nunes et al., 2022; Ali et al., 2023; Zhu et al., 2022). To mitigate the power losses caused by multiple MPP scenarios, various advanced maximum power-point tracking (MPPT) solutions have been proposed for partially shaded PV systems, such as the reference-voltage-line-aided method (Li et al., 2022), particle swarm optimization algorithm (Refaat et al., 2023), peafowl optimization algorithm (Li et al., 2021), and machine-learning-based method (Yilmaz et al., 2023). However, these methods have mostly increased the implementation complexity and computational requirements of the control units. Furthermore, a portion of the available power from the PV string is dissipated by implementing such MPPT methods, which reduces the overall energy efficiency of the system (Chu et al., 2024).

To address these issues, new architectures like full power processing (FPP) and differential power processing (DPP) have been proposed to enhance power generation under PV mismatch conditions. As shown in Figure 1, DC power optimizers (DCPOs) and microinverters are the widely adopted power interfaces in the FPP architecture for distributed PV systems (Celik et al., 2018; Ramli and Salam, 2019). Here, each DCPO or microinverter is connected in series and includes an MPPT unit to maximize the power extracted from the corresponding PV module, thereby guaranteeing the efficiency of the system. However, a certain level of power loss is unavoidable in this architecture as the PV power production is fully processed, degrading both the energy production and system reliability (Shenoy et al., 2013). Compared to FPP, the DPP architecture reduces power losses by processing only a portion of the total PV power through the DPP converter units while the main power is fed to the centralized converter directly (Jeong et al., 2019). This means that only a fraction of the total power is processed by the DPP converter and that the power rating specified for each DPP converter is relatively low; further, this architecture requires a small-sized and low-cost design (Amaral da Luz et al., 2023). Notably, under consideration of similar converter efficiencies, the total power loss of the DPP architecture is lower than that of the DCPO on average. Meanwhile, if a low-level or no-mismatch condition occurs, none of the power passes through the DPP converter, which further improves the power transmission efficiency. A comparison between FPP and DPP showed that the system could achieve higher efficiency under the DPP architecture owing to contribution from the fraction of output power processing capability (Olalla et al., 2015).

The DPP architecture can be broadly divided into two types, namely PV element to PV element (PV-PV) architecture and PV element to bus architectures, according to the port connection conditions of the DPP converter. For the PV-PV architecture shown in Figure 2a, the number of DPP converter units is always one less than the number of PV modules; here, the proposed topologies include switched-capacitor converter (SCC) (Uno et al., 2022) and switched-inductor converter (SLC) (Shenoy et al., 2013). The SCC system is based on the principle of voltage equalization among all submodules but cannot achieve an accurate MPP for each submodule when a mismatch occurs (Dong et al., 2019). The synchronous bidirectional buck-boost topology is used as the DPP converter in the SLC configuration, which provides bidirectional

transition of power flow between the relative submodules. However, the PV-PV architecture requires connections between adjacent PV elements, which increases the system complexity and necessitates coordinated control. The PV element to bus DPP architecture topologies can be divided into PV element to isolated port (PV-IP) and PV element to non-isolated port (PV-NIP) bus architectures, as demonstrated in Figures 2b–d. In the PV-IP architecture illustrated in Figure 2b, the secondary sides of the DPP converters are parallel, establishing an isolated port bus; accordingly, selection of the secondary-side voltage of each DPP converter unit can be decoupled from the PV voltage. The PV-IP DPP architecture provides high flexibility of hardware design and reliable voltage equalization capability without an additional communication network for the DPP and central converters (Chu et al., 2017). In the type I PV-NIP bus architecture shown in Figure 2c, the secondary sides of the DPP converters are connected in parallel and share the same input voltage as the central converter; the switching components on the secondary side experience higher voltage stresses than those on the primary side as they share the same voltage (i.e., the sum of the PV module output voltage), which increases the power loss and implementation cost. Moreover, both the submodule and central converter require MPPT control and parameter sensors (Khan and Xiao, 2017). In this case, the string current cannot be regulated, and the power rating limitation of each DPP converter becomes invalid. Furthermore, given the high voltage stresses on the secondary side of the PV-NIP architecture, the system flexibility is further increased and power loss is reduced. By utilizing this topology, the differential power processed by the DPP and string converters increases the system power loss (Chu et al., 2022). To address these limitations, the type II PV-NIP bus architecture implementation is shown in Figure 2d; the secondary sides in this topology are directly connected to the bus, and the central converter can control the string current directly. Thus, the currents in both the DPP and central converters can be decoupled to improve and the reliability and control flexibility of the system. The important features of different control schemes for the PV-NIP DPP architectures are compared in Table 1.

To achieve balanced distribution of the differential power and simple extraction of the maximum power, a unit power rating balancing (UPRB) scheme with DPP-converter-unit-level-independent MPPT control is proposed herein for the PV-NIP DPP architectures. To ensure bidirectional power transfer capability, various bidirectional DC-DC converters may be employed as the DPP units, e.g., dual active bridge (DAB) converters (Wang et al., 2021; Bu et al., 2022; Shi et al., 2022; Feng et al., 2025) and quasi-Z-source-based isolated bidirectional DC-DC converter (Kafle et al., 2019). Compared to the aforementioned bidirectional solutions, the flyback topology outperforms in simple configurations and requires fewer power switches, making it cost-effective and suitable for low-voltage applications. Accordingly, the bidirectional flyback topology is employed as the DPP converter and implemented in the PV-NIP bus DPP architecture, where the voltage gain of the flyback converter is easily adjusted by regulating the turns ratio of the transformer and matching the required voltage level of the bus. The proposed UPRB is an integrated perturbation and observation (P&O)-based MPPT controller in each DPP converter that maximizes the energy harvested from the PV modules; further, a unit balance-point tracking (UBPT) unit

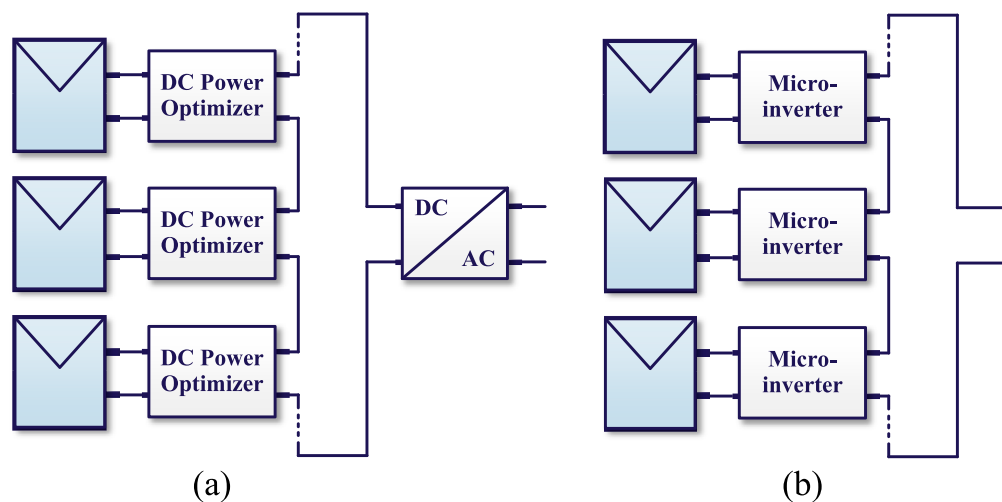


FIGURE 1
Typical full power processing architectures for distributed photovoltaic (PV) systems: (a) DC power optimizer; (b) microinverter.

is employed to determine the optimal string current reference adaptively to mitigate the unbalanced differential power in the DPP units. Notably, the MPPT control at the DPP converter unit level is decoupled from the string-level current optimization. The implementation of the widely adopted P&O-based MPPT method herein aims to reduce control complexity. For better maximum energy yield, the P&O method may be replaced with any other advanced MPPT method. Both simulations and experiments were carried out in this work to validate the effectiveness of the proposed UPRB scheme. The remainder of this paper is organized as follows. Section 2 presents an analysis of the theoretical differential power distributions in the PV-NIP bus DPP architectures. The technical details of the proposed UPRB scheme are introduced in Section 3. The simulations and experimental validations are presented in Section 4 and Section 5, respectively, and the conclusions of the study are summarized in Section 6.

2 Theoretical power distribution in the PV-NIP DPP architecture

Compared to other DPP architectures, the type II PV-NIP bus DPP architecture offers better system efficiency and control flexibility; here, the DPP converter only transfers energy between its relative series-connected elements and the entire string, which offers the benefits of modularity and independent converter operating state. Accordingly, the type II PV-NIP bus DPP architecture shown in Figure 2d is employed in this work.

2.1 Generalized power distribution model

In a PV-NIP bus architecture with bidirectional DPP converter units, the direction of differential current flow according to

Kirchhoff's current law (KCL) depends on the differences between the string current and currents in the corresponding PV elements. As demonstrated in Figure 3a, when the current of the i^{th} PV element $I_{pv,i}$ is greater than the string current I_{string} ($I_{pv,i} > I_{\text{string}}$), the DPP converter will remove the current from the PV element and feed the bus for approaching the MPP of the relative element, thereby creating the differential current $I_{dpp,i} = I_{pv,i} - I_{\text{string}}$ such that the power flows into the DPP converter from the PV element. When the PV element current is equal to the string current ($I_{pv,i} = I_{\text{string}}$), as shown in Figure 3b, no power will be injected or removed between the PV element and PV string; this means that the differential current $I_{dpp,i}$ is equal to zero. When the PV element current is less than the string current ($I_{pv,i} < I_{\text{string}}$), the power in the string will feed the PV element via injected differential current from the bus through the DPP converter unit to achieve the MPP, as demonstrated in Figure 3c; here, the differential current flow through the i^{th} DPP converter unit $I_{dpp,i} = I_{pv,i} - I_{\text{string}}$.

Accordingly, the system can be configured using a mathematical model to explain the relationships between the PV element current, string current, and differential current through basic ascension of each PV element with the operating current $I_{pv,i}$ and voltage $V_{pv,i}$ as well as a known string current I_{string} . According to KCL, the differential current of the i^{th} DPP converter unit can be expressed as $I_{dpp,i} = I_{pv,i} - I_{\text{string}}$ and given in matrix form as Equation 1:

$$\begin{bmatrix} 1 & -1 & 0 & \cdots & 0 \\ 0 & 1 & -1 & \ddots & \vdots \\ 0 & \ddots & \ddots & \ddots & 0 \\ \vdots & \ddots & 0 & 1 & -1 \\ 0 & \cdots & 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} I_{dpp,1} \\ I_{dpp,2} \\ \vdots \\ I_{dpp,n-1} \\ I_{dpp,n} \end{bmatrix} = \begin{bmatrix} I_{pv,2} - I_{pv,1} \\ I_{pv,3} - I_{pv,2} \\ \vdots \\ I_{pv,n} - I_{pv,n-1} \\ I_{\text{string}} - I_{pv,n} \end{bmatrix}. \quad (1)$$

Here, the conditions of the i^{th} PV element are known but the operating condition of the central converter is unknown.

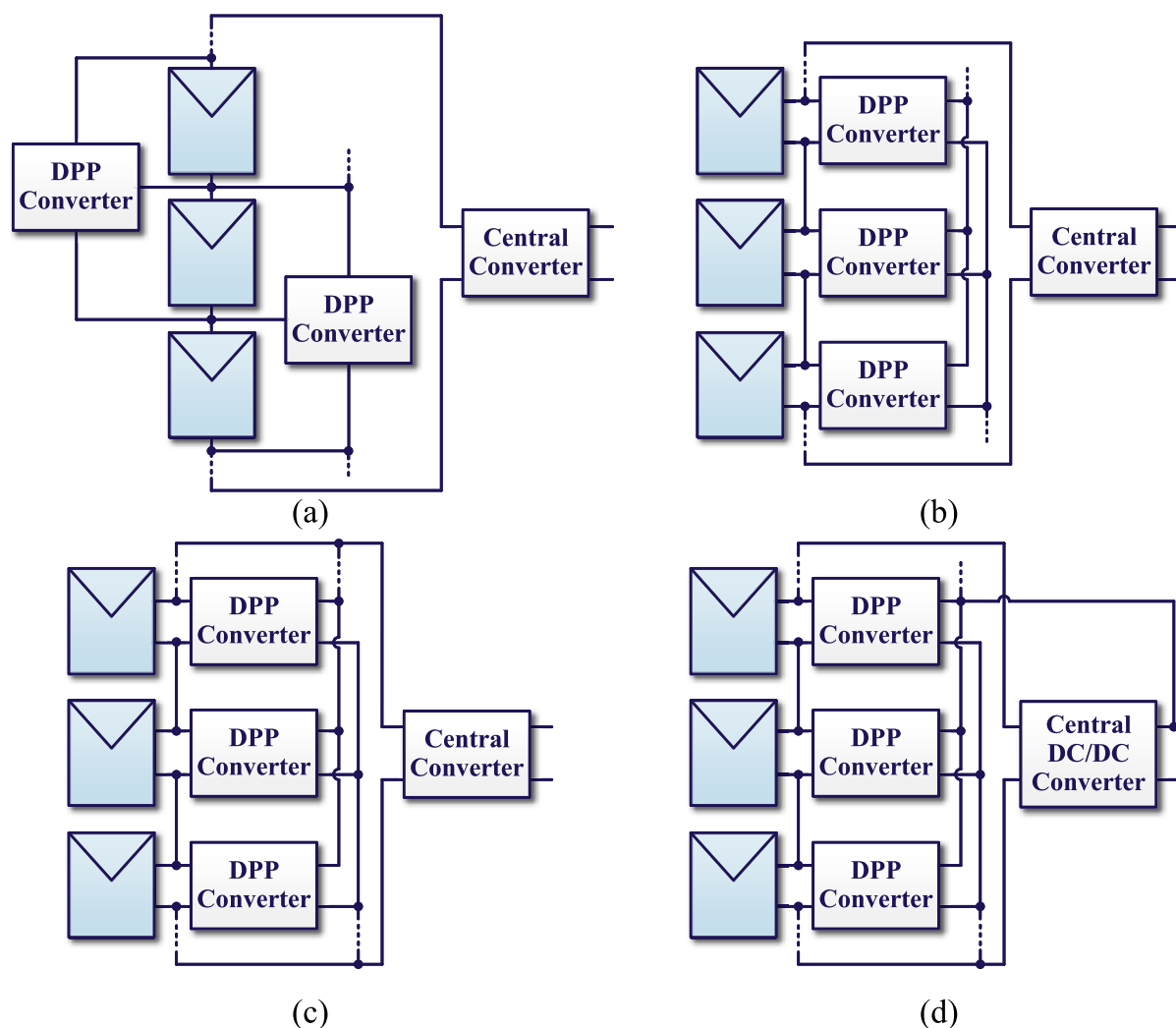


FIGURE 2 Typical differential power processing (DPP) architectures for distributed PV systems: (a) PV element to PV element architecture; (b) PV element to isolated port bus architecture; (c) PV element to non-isolated port (PV-NIP) bus architecture type I; (d) PV-NIP bus architecture type II.

TABLE 1 Key features of different control schemes for the PV-NIP DPP architectures.

Reference	Objective	MPPT	String control	Complexity	Tracking speed	Oscillations
Jeon et al. (2017)	Total minimum	P&O	P&O	Moderate	Low	High
Jeon and Park (2019)	Unit balance	P&O	P&O	Moderate	Low	High
Chu et al. (2019)	Total minimum	Modified P&O	Direct	Moderate	High	Moderate
Chu et al. (2020)	Adaptive	Direct	Direct	Simple	High	Low
Zhu et al. (2022)	Unit balance	Modified P&O	Direct	High	Moderate	Low
This work	Unit balance	P&O	Direct	Moderate	Moderate	Moderate

The central converter determines the conditions of the string current I_{string} . Hence, the condition of the i^{th} DPP converter unit is unknown, which offers a range of control freedom for this codetermined system.

The ideal differential power flowing through the i^{th} DPP converter unit $P_{\text{dpp},i}$ is expressed as in Equation 2:

$$P_{\text{dpp},i} = V_{\text{pv},i} |I_{\text{dpp},i}| = V_{\text{pv},i} |I_{\text{pv},i} - I_{\text{string}}|. \quad (2)$$

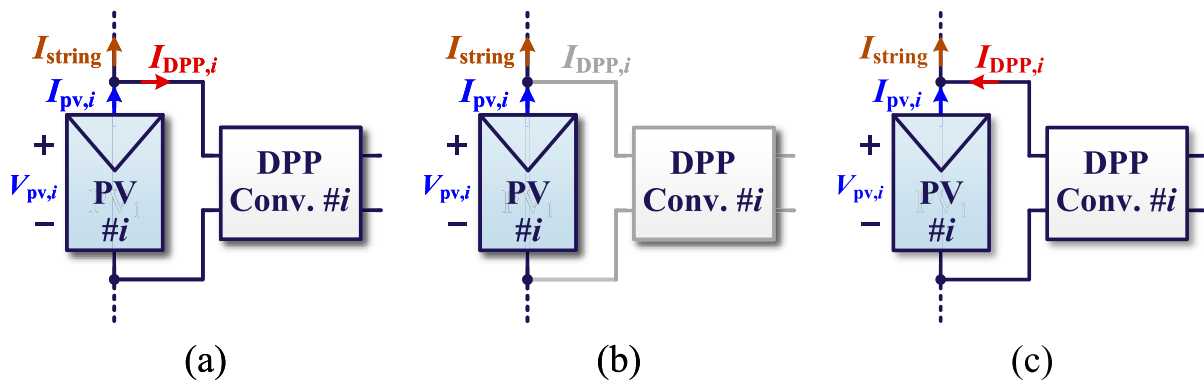


FIGURE 3

Differential current paths of the i^{th} DPP converter unit in the type II PV-NIP bus DPP architecture: (a) differential current path when $I_{pv,j} > I_{\text{string}}$; (b) differential current path when $I_{pv,j} = I_{\text{string}}$; (c) differential power path when $I_{pv,j} < I_{\text{string}}$. Here, $I_{pv,j}$ is the current of the i^{th} PV element, $V_{pv,j}$ is the voltage of the i^{th} PV element, $I_{dpp,i}$ is the differential current flowing through the i^{th} DPP converter unit, and I_{string} is the current of the PV string.

Then, the ideal total differential power P_{DPP} of a DPP system with n PV elements and n DPP converter units can be expressed as in Equation 3:

$$P_{\text{DPP}} = \sum_{i=1}^n P_{\text{dpp},i} = \sum_{i=1}^n V_{pv,i} |I_{\text{dpp},i}| \quad (3)$$

$$= V_{pv,1} |I_{pv,1} - I_{\text{string}}| + V_{pv,2} |I_{pv,2} - I_{\text{string}}| + \dots + V_{pv,n} |I_{pv,n} - I_{\text{string}}|$$

Considering the power rating, P_{DPP} can be regulated to a specific value by changing the string current, as mentioned before. In reality, the voltage variation is far lower than the current changes at different irradiation levels. Therefore, the power rating of each DPP converter is affected by the string current resulting from different DPP currents as well as the approximate PV voltage. In a system with n PV elements, assuming that the PV voltage and current match the conditions $V_{pv,1} \cong V_{pv,2} \cong \dots \cong V_{pv,n}$ and $I_{pv,1} < I_{pv,2} < \dots < I_{pv,n}$, a maximum differential power rating $P_{\text{dpp,max}}$ will exist among the DPP converter units whose particular values would follow the string current changes.

When the string current I_{string} is less than the minimum PV current $I_{pv,1}$ ($I_{\text{string}} < I_{pv,1}$), the unit processing the greatest power under this assumption will be the i^{th} DPP converter whose power rating is expressed by Equation 4:

$$P_{\text{dpp,max}} = V_{pv,n} (I_{pv,n} - I_{\text{string}}), \text{ when } I_{\text{string}} < I_{pv,1}. \quad (4)$$

As the string current increases, in the region of between $I_{pv,1}$ and average of $I_{pv,1}$ and $I_{pv,n}$ ($I_{pv,1} < I_{\text{string}} < \frac{I_{pv,1} + I_{pv,n}}{2}$), the largest differential power rating still occurs at the i^{th} DPP converter whose value is given by Equation 5:

$$P_{\text{dpp,max}} = V_{pv,n} (I_{pv,n} - I_{\text{string}}), \text{ when } I_{pv,1} < I_{\text{string}} < \frac{I_{pv,1} + I_{pv,n}}{2}. \quad (5)$$

With continued increase of the string current, the DPP converter with the most stressed differential power rating will change to the first converter unit whose string current is in the range of average

of $I_{pv,1}$ and $I_{pv,n}$ to less than $I_{pv,n}$, namely $\frac{I_{pv,1} + I_{pv,n}}{2} < I_{\text{string}} < I_{pv,n}$, as given by Equation 6:

$$P_{\text{dpp,max}} = V_{pv,1} (I_{pv,1} - I_{\text{string}}), \text{ when } \frac{I_{pv,1} + I_{pv,n}}{2} < I_{\text{string}} < I_{pv,n}. \quad (6)$$

If the string current increases beyond $I_{pv,n}$ ($I_{\text{string}} > I_{pv,n}$), the first DPP converter unit sustains the largest differential power rating as expressed by Equation 7:

$$P_{\text{dpp,max}} = V_{pv,1} (I_{pv,1} - I_{\text{string}}), \text{ when } I_{\text{string}} > I_{pv,n}. \quad (7)$$

In summary, for a lower string current value less than the average of $I_{pv,1}$ and $I_{pv,n}$, the lower differential power will be processed such that the largest differential power stressing the n^{th} DPP converter unit can be released. Thus, the maximum differential power can be balanced among the n DPP converter units when I_{string} is equal to $\frac{I_{pv,1} + I_{pv,n}}{2}$. Hence, the expressions for $P_{\text{dpp,max}}$ in a PV-NIP DPP system with n PV elements and n DPP converter units can be revised as Equations 8a–c under the varied string current:

$$P_{\text{dpp,max}} = V_{pv,n} (I_{pv,n} - I_{\text{string}}), \text{ when } I_{\text{string}} < \frac{I_{pv,1} + I_{pv,n}}{2} \quad (8a)$$

$$P_{\text{dpp,max}} = V_{pv,1} (I_{pv,1} - I_{\text{string}}) = V_{pv,n} (I_{pv,n} - I_{\text{string}}), \text{ when } I_{\text{string}} = \frac{I_{pv,1} + I_{pv,n}}{2} \quad (8b)$$

$$P_{\text{dpp,max}} = V_{pv,1} (I_{pv,1} - I_{\text{string}}), \text{ when } I_{\text{string}} > \frac{I_{pv,1} + I_{pv,n}}{2} \quad (8c)$$

2.2 Power distribution in the example DPP system

In PV-NIP DPP architectures with odd or even numbers of PV elements and DPP converter units, as shown in Figures 4, 6,

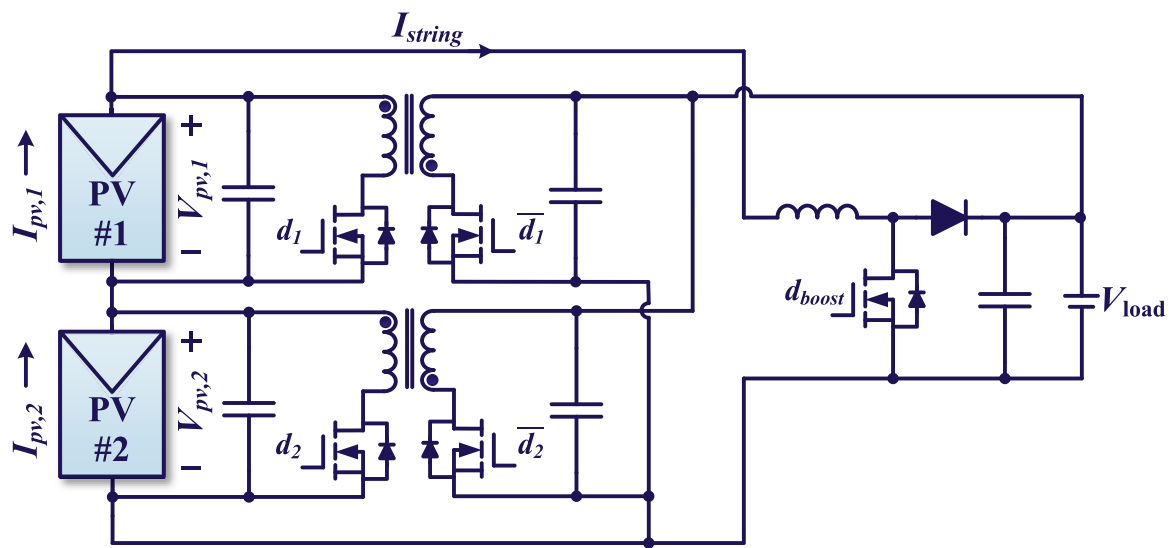


FIGURE 4
PV-NIP bus DPP architecture with even numbers (two) of PV elements and bidirectional flyback-based DPP converter units.

respectively, the differential power stress on each DPP converter can be calculated using the above approach. Assuming that there is no power loss during power transfer and that each PV element operates at its MPP with a very close MPP voltage, the theoretical differential power analysis can be carried out for the example architecture.

2.2.1 Example of a system with two PV elements

According to Equation 3, the total differential power for the DPP architecture with three PV elements shown in Figure 4 is given by Equation 9:

$$P_{DPP} = |P_{dpp,1}| + |P_{dpp,2}| = V_{pv,1} |I_{pv,1} - I_{string}| + V_{pv,2} |I_{pv,2} - I_{string}| \quad (9)$$

Assuming $I_{pv,1} < I_{pv,2}$, the string current I_{string} is swept from zero to a value exceeding $I_{pv,2}$, and the largest differential power among the DPP converters can be expressed as Equations 10a–c according to the different values of I_{string} :

$$P_{dpp,max} = V_{pv,2} (I_{pv,2} - I_{string}), \text{ when } I_{string} < \frac{(I_{pv,1} + I_{pv,2})}{2} \quad (10a)$$

$$P_{dpp,max} = V_{pv,1} (I_{pv,1} - I_{string}) = V_{pv,n} (I_{pv,n} - I_{string}), \text{ when } I_{string} = \frac{(I_{pv,1} + I_{pv,2})}{2} \quad (10b)$$

$$P_{dpp,max} = V_{pv,1} (I_{pv,1} - I_{string}), \text{ when } I_{string} > \frac{(I_{pv,1} + I_{pv,2})}{2} \quad (10c)$$

To illustrate the differences in differential power obtained, a case study was conducted using the MSX-60 PV module, which has an MPP voltage of 17 V and MPP current of 3.5 A under standard test conditions. In the case study, the example is assumed under the mismatching condition, where the two PV elements are operated at unique MPPs with MPP currents of 1.4 A and 3.5 A for PV elements 1 and 2, respectively.

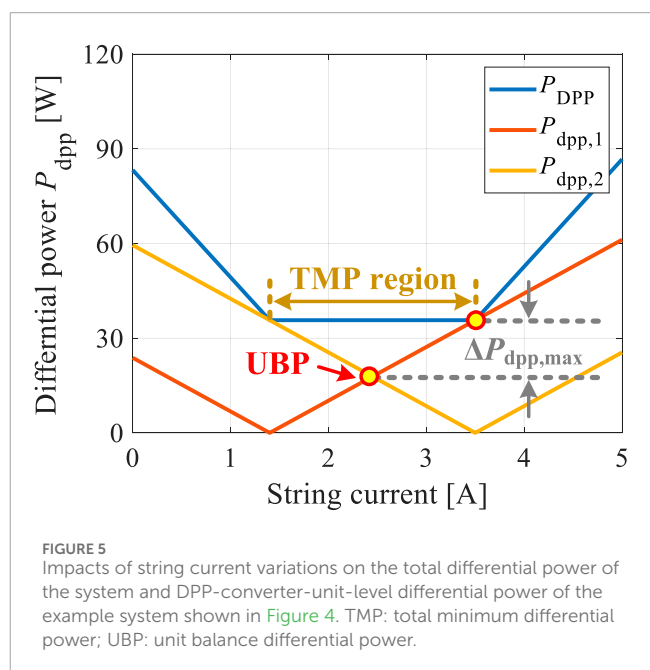


FIGURE 5
Impacts of string current variations on the total differential power of the system and DPP-converter-unit-level differential power of the example system shown in Figure 4. TMP: total minimum differential power; UB: unit balance differential power.

As demonstrated in Figure 5, the impacts of string current variations on the total differential power P_{DPP} and DPP-converter-unit-level differential power values ($P_{dpp,1}$ and $P_{dpp,2}$) were investigated. The total minimum differential power (TMP) P_{DPP} can be obtained by restricting the string current to the range between 1.4 A and 3.5 A. The extreme power stresses of the two DPP converter units within the region of the TMP are 35.7 W and 0 W. Thus, the differential power stresses on the DPP converter units can be balanced using an operating string current of 2.45 A, which may be considered the unit balance differential power (UBP) point and expressed using Equations 10b.

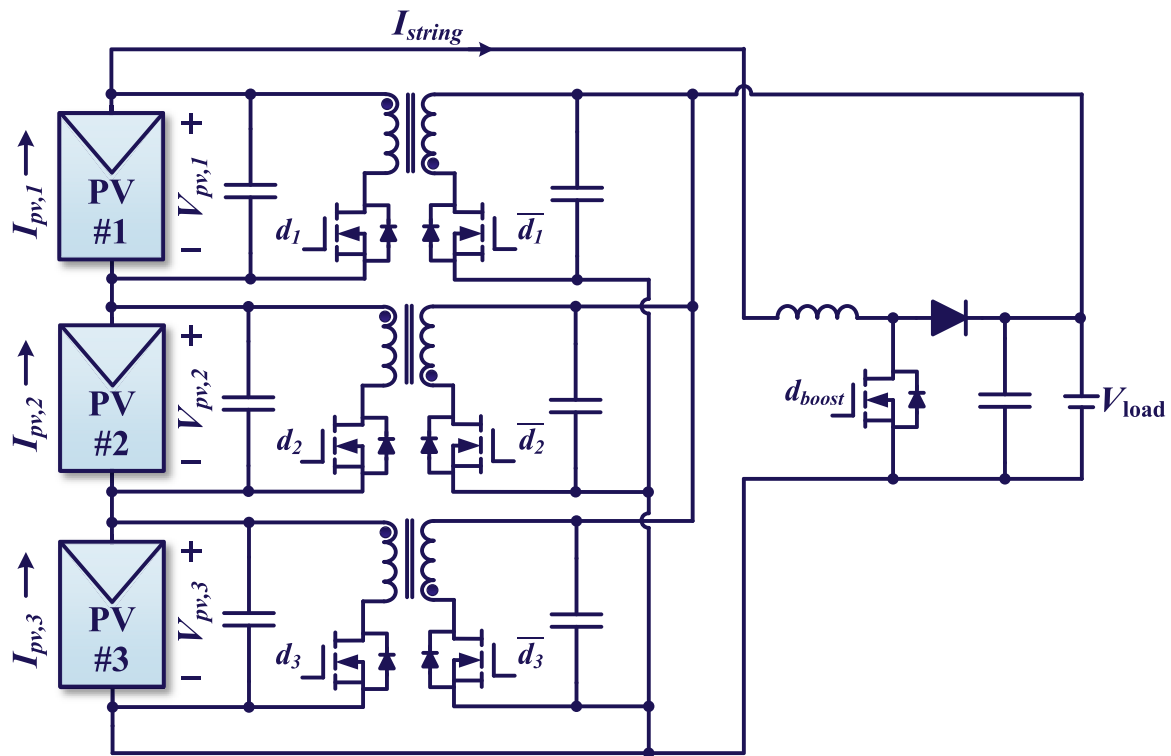


FIGURE 6
PV-NIP bus DPP architecture with odd numbers (three) of PV elements and bidirectional flyback-based DPP converter units.

2.2.2 Example of a system with three PV elements

The total differential power for a DPP architecture with three PV elements shown in Figure 6 is given by Equation 11:

$$P_{DPP} = |P_{dpp,1}| + |P_{dpp,2}| + |P_{dpp,3}|$$

$$= V_{pv,1} |I_{pv,1} - I_{string}| + V_{pv,2} |I_{pv,2} - I_{string}| + V_{pv,3} |I_{pv,3} - I_{string}| \quad (11)$$

Assuming $I_{pv,1} < I_{pv,2} < I_{pv,3}$, the string current I_{string} is swept from zero to a value exceeding $I_{pv,3}$, and the largest differential power among the DPP converters can be expressed by Equations 12a–c according to the different values of I_{string} :

$$P_{dpp,max} = V_{pv,3} (I_{pv,3} - I_{string}), \text{ when } I_{string} < \frac{(I_{pv,1} + I_{pv,3})}{2} \quad (12a)$$

$$P_{dpp,max} = V_{pv,1} (I_{pv,1} - I_{string}) = V_{pv,n} (I_{pv,n} - I_{string}),$$

$$\text{when } I_{string} = \frac{(I_{pv,1} + I_{pv,3})}{2} \quad (12b)$$

$$P_{dpp,max} = V_{pv,1} (I_{pv,1} - I_{string}), \text{ when } I_{string} > \frac{(I_{pv,1} + I_{pv,3})}{2} \quad (12c)$$

In this case, the example mismatches of the MPP currents in the three-element DPP system are 0.7 A, 2.8 A, and 3.5 A for PV elements 1, 2, and 3, respectively. As demonstrated in Figure 7, the impacts of string current variations on the total differential power P_{DPP} and DPP-converter-unit-level differential power values ($P_{dpp,1}$,

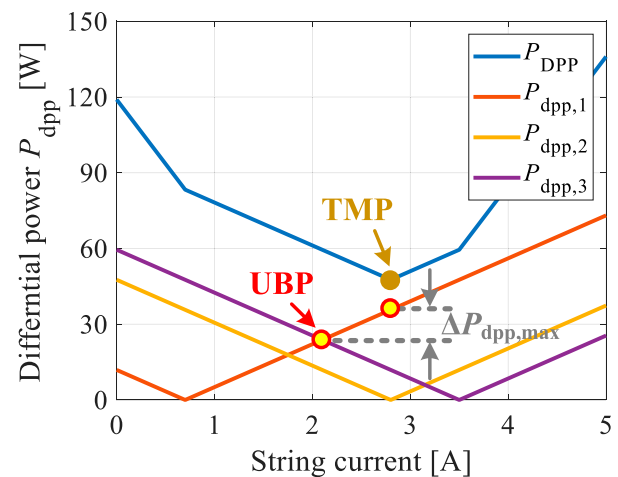


FIGURE 7
Impacts of string current variations on the total differential power of the system and DPP-converter-unit-level differential power of the example system shown in Figure 6. TMP: total minimum differential power; UBP: unit balance differential power.

$P_{dpp,2}$, and $P_{dpp,3}$) were analyzed. Unlike Figure 5, the TMP P_{DPP} was obtained only at the string current of 2.8 A, its maximum unit-level differential power was 35.7 W. Thus, the maximum differential power stresses at the DPP converter unit level can be balanced using a string current of 2.8 A, which is determined using Equation 12b.

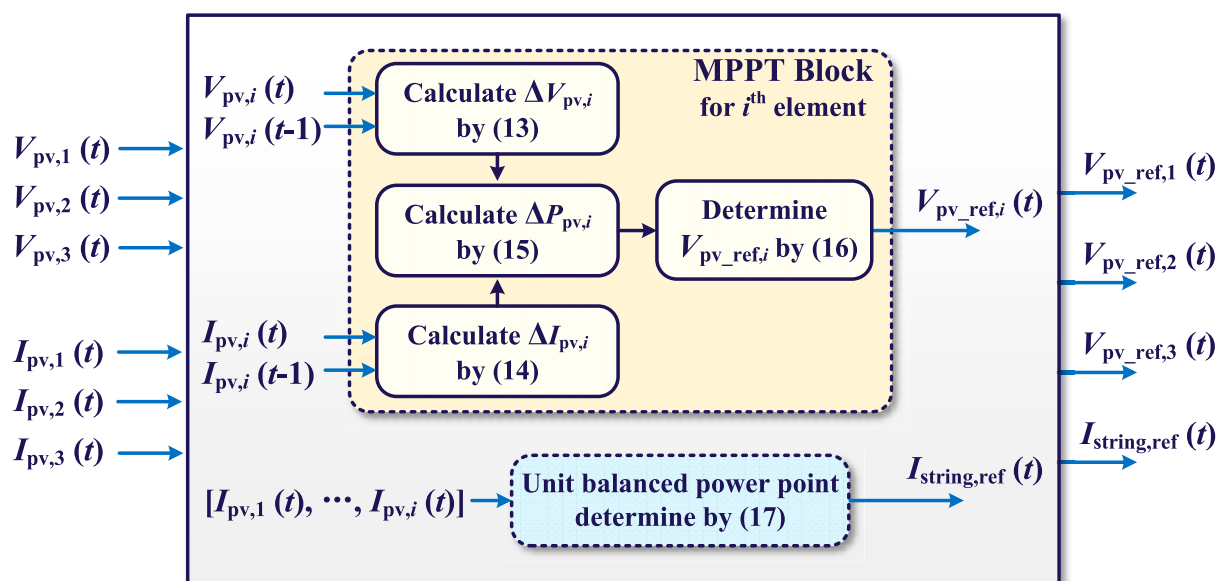


FIGURE 8
Control diagram of the proposed unit power rating balancing (UPRB) scheme.

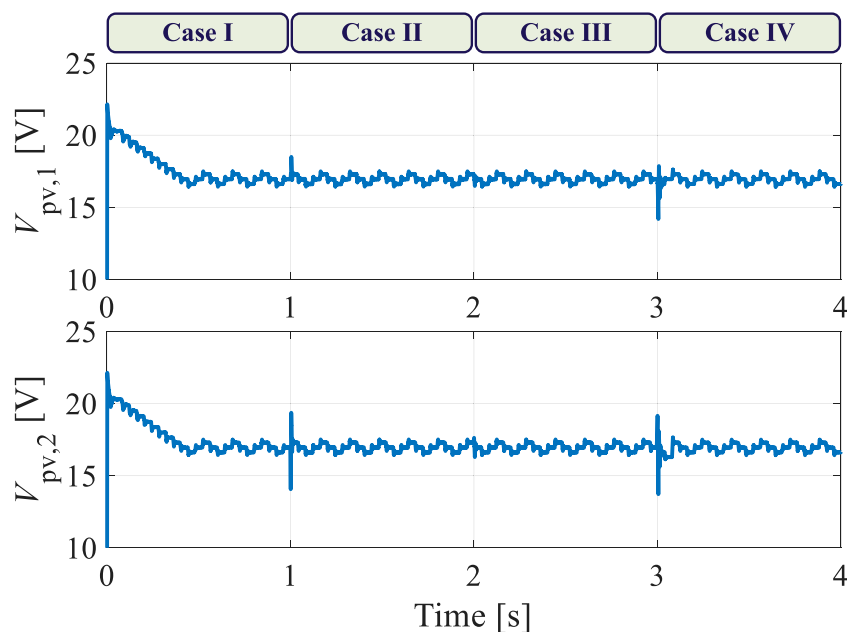


FIGURE 9
Simulation results of the PV element voltages under various shading scenarios.

In summary, the tracking for the TMP may induce a significant unbalancing in the unit differential power distribution in the PV-NIP DPP system. It should be noted that UBP distribution can be achieved by regulating the string current. In a DPP system with even number of elements, the tracking for UBP could also ensure minimization of the total differential power in some cases.

3 Proposed UPRB scheme

Notably, the mitigation of unbalanced differential power can reduce uneven thermal distribution as well as the aging effect in relative converters, improving the lifetime of the DPP system (Zhu et al., 2022). In accordance with the above analysis, the unit-level differential power can be optimized sufficiently with minimal

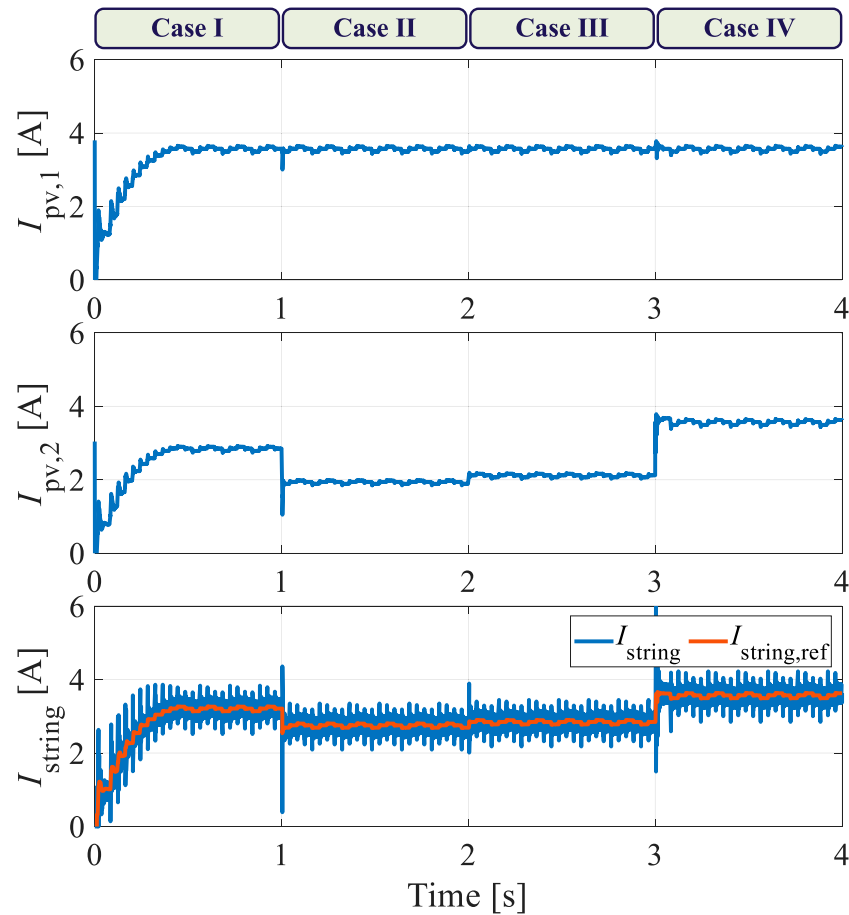


FIGURE 10
Simulation results of the PV element currents and string current under various shading scenarios.

uneven distribution by regulating the string current to the optimal UBP point. Thus, a UPRB control scheme is proposed herein, whose block diagram is illustrated in Figure 8. The proposed scheme requires periodic measurement of the currents and voltages of each of the PV elements.

3.1 MPPT block

To maximize the energy harvested from the PV modules, a P&O-based MPPT controller is integrated into each DPP converter. Based on periodic measurement of the PV voltage of each element, the change in voltage at the most recent sampling interval of the i^{th} element ($\Delta V_{pv,i}(t)$) is given by Equation 13:

$$\Delta V_{pv,i}(t) = V_{pv,i}(t) - V_{pv,i}(t-1) \quad (13)$$

where $V_{pv,i}(t)$ is the measured voltage of the i^{th} PV element in the present sampling period, and $V_{pv,i}(t-1)$ is the measured voltage of the i^{th} PV element in the previous sampling period.

Similarly, the current change in the i^{th} PV element between any two adjacent sampling intervals can be expressed by Equation 14:

$$\Delta I_{pv,i}(t) = I_{pv,i}(t) - I_{pv,i}(t-1) \quad (14)$$

where $I_{pv,i}(t)$ is the measured current of the i^{th} PV element in the present sampling period, and $I_{pv,i}(t-1)$ is the measured current of the i^{th} PV element in the previous sampling period.

Hence, the power change in the i^{th} PV element ($\Delta P_{pv,i}$) can be expressed by Equation 15:

$$\Delta P_{pv,i}(t) = P_{pv,i}(t) - P_{pv,i}(t-1) = \Delta V_{pv,i}(t) \cdot \Delta I_{pv,i}(t). \quad (15)$$

Accordingly, the voltage reference regulation for the i^{th} PV element to ensure maximum power extraction during the subsequent perturbation interval ($V_{pv_ref,i}(t)$) is determined by Equations 16a–d:

$$V_{pv_ref,i}(t) = V_{pv_ref,i}(t) + \Delta V_{\text{step}}, \text{ when } \Delta P_{pv,i}(t) \geq 0 \text{ and } \Delta V_{pv,i}(t) \geq 0 \quad (16a)$$

$$V_{pv_ref,i}(t) = V_{pv_ref,i}(t) - \Delta V_{\text{step}}, \text{ when } \Delta P_{pv,i}(t) \geq 0 \text{ and } \Delta V_{pv,i}(t) < 0 \quad (16b)$$

$$V_{pv_ref,i}(t) = V_{pv_ref,i}(t) + \Delta V_{\text{step}}, \text{ when } \Delta P_{pv,i}(t) < 0 \text{ and } \Delta V_{pv,i}(t) < 0 \quad (16c)$$

$$V_{pv_ref,i}(t) = V_{pv_ref,i}(t) - \Delta V_{\text{step}}, \text{ when } \Delta P_{pv,i}(t) < 0 \text{ and } \Delta V_{pv,i}(t) \geq 0 \quad (16d)$$

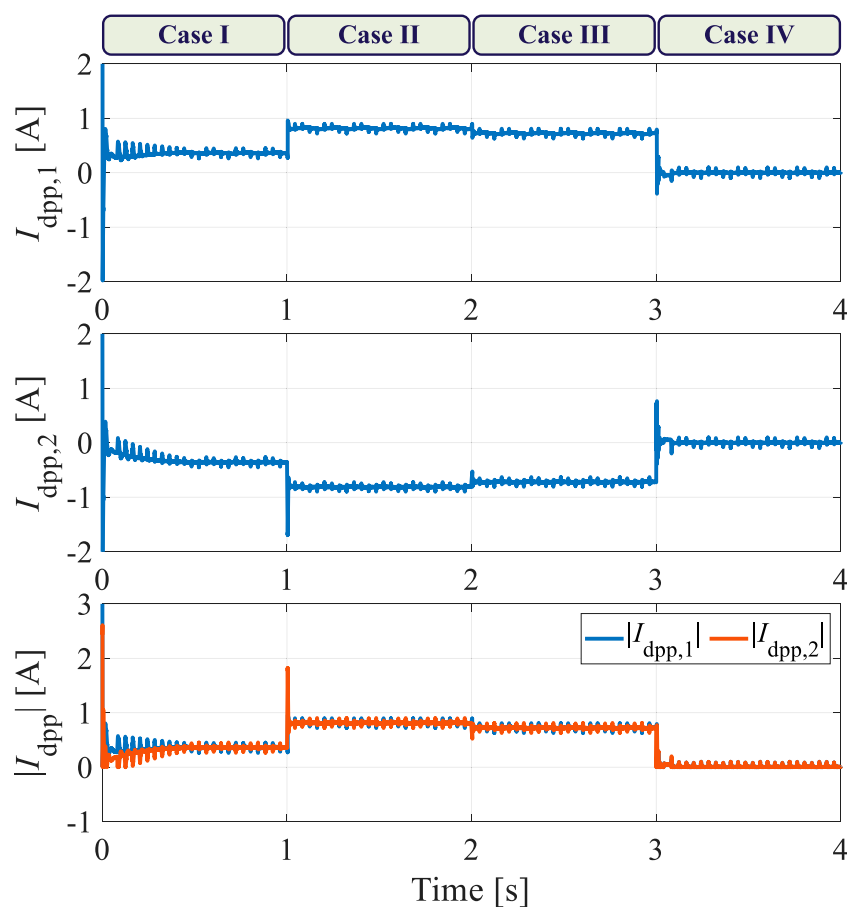


FIGURE 11
Simulation results of the differential currents flowing through the DPP converter units under various shading scenarios.

where ΔV_{step} is the optimal voltage step size for ensuring balanced tradeoff between the steady-state and dynamic performances of the MPPT control (Lv et al., 2024). Then, a proportional integral (PI) controller can be employed to determine the appropriate duty cycle of the bidirectional DPP converter unit.

3.2 UBPT block

To mitigate the unbalanced differential power distribution among the DPP converter units, a differential UBPT unit is employed to determine the optimal string current reference adaptively. Based on the power rating analysis above, power balance is achieved at the midpoint of the PV string. To mitigate the oscillations induced by the P&O mechanism for string current reference optimization, the string current reference $I_{\text{string,ref}}(t)$ is directly determined by Equation 17:

$$I_{\text{string,ref}}(t) = \frac{\max[I_{pv,1}(t) \quad I_{pv,2}(t) \quad \dots \quad I_{pv,n}(t)] + \min[I_{pv,1}(t) \quad I_{pv,2}(t) \quad \dots \quad I_{pv,n}(t)]}{2} \quad (17)$$

To regulate the string current with fast response and high accuracy, a PI controller is employed to control the string current based on the reference calculated in Equation 17. If the optimization

objective of the string current reference is with respect to the TMP, $I_{\text{string,ref}}(t)$ can be expressed by Equation 18:

$$I_{\text{string,ref}}(t) = \arg \min_{I_{\text{string,ref}}(t) \in [I_{pv,1}(t), \dots, I_{pv,n}(t)]} \left[\sum_{i=1}^n V_{pv,i}(t) I_{pv,i}(t) \right] \quad (18)$$

Hence, in a PV-NIP DPP system with even number of PV elements, the TMP is obtained by tracking the UBP, as shown in Figure 5 for the example with two PV elements.

4 Simulations

The proposed control strategy was simulated in PSIM using the example system setup shown in Figure 4. The MSX-60 PV module with an MPP voltage of 17 V and MPP current of 3.5 A was used as the PV source under standard test conditions. Bidirectional flyback converters with a switching frequency of 50 kHz were employed as the DPP converter units. The perturbation interval for the unit-level P&O-based MPPT control was set to 0.04 s.

The irradiation was changed every 1 s to simulate unpredictable environmental conditions for evaluating the proposed control scheme. During the first second (Case I), the irradiation on PV elements 1 and 2 were 1,000 and 800 W/m², respectively. During

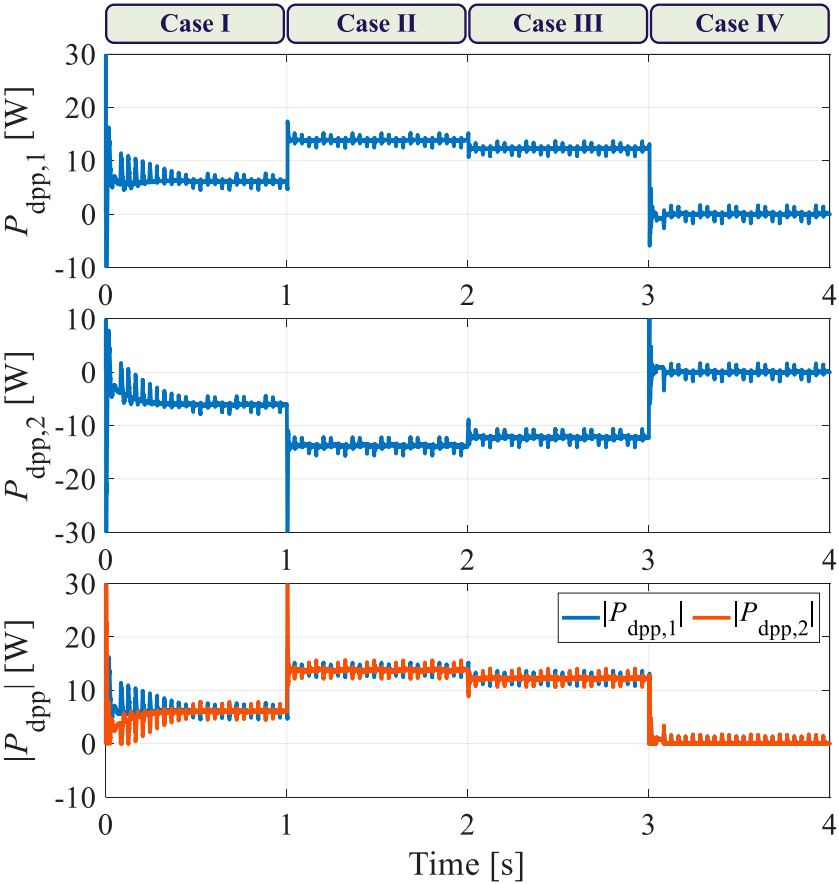


FIGURE 12
Simulation results of the differential power values processed by the DPP converter units under various shading scenarios.

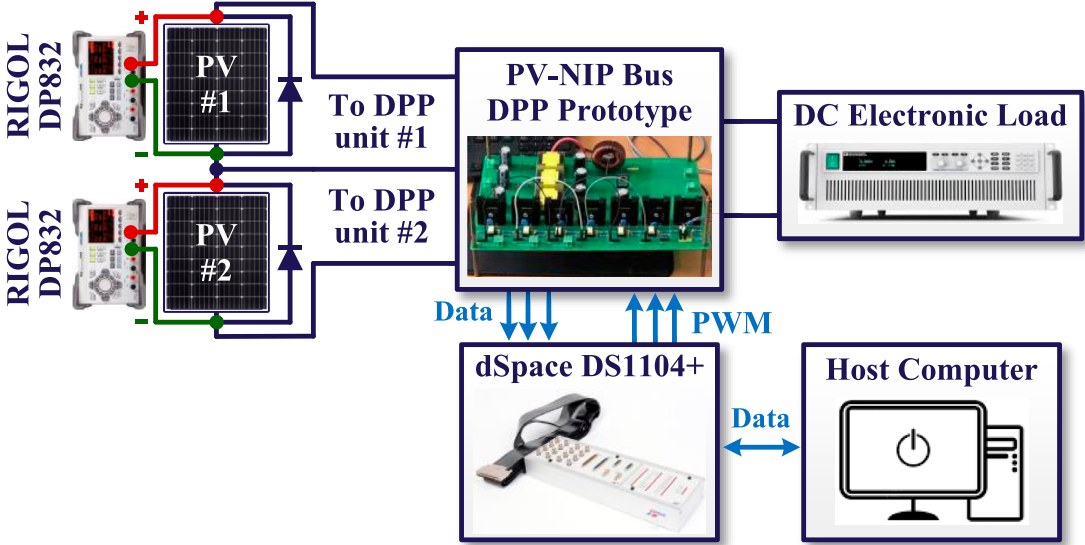
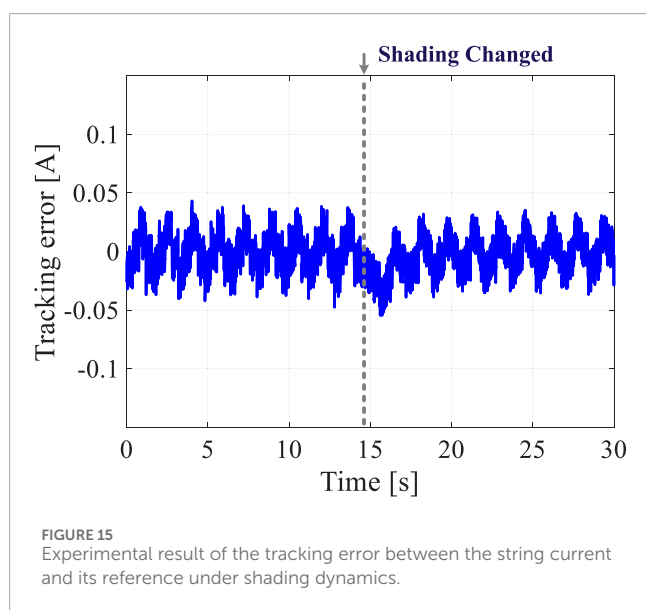
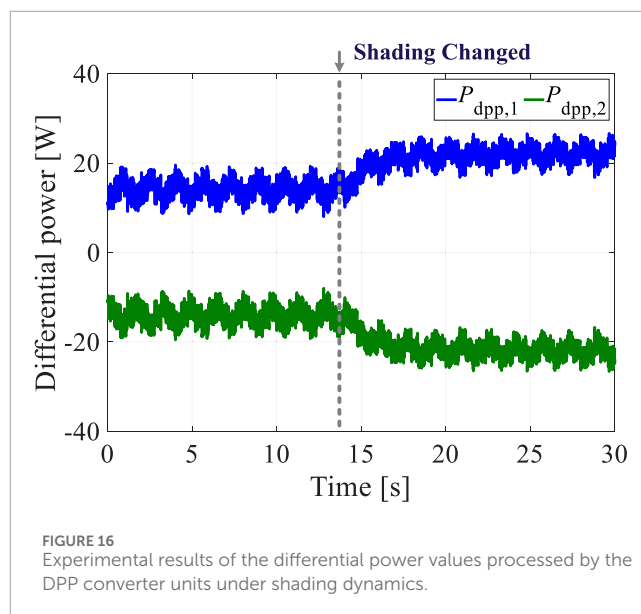
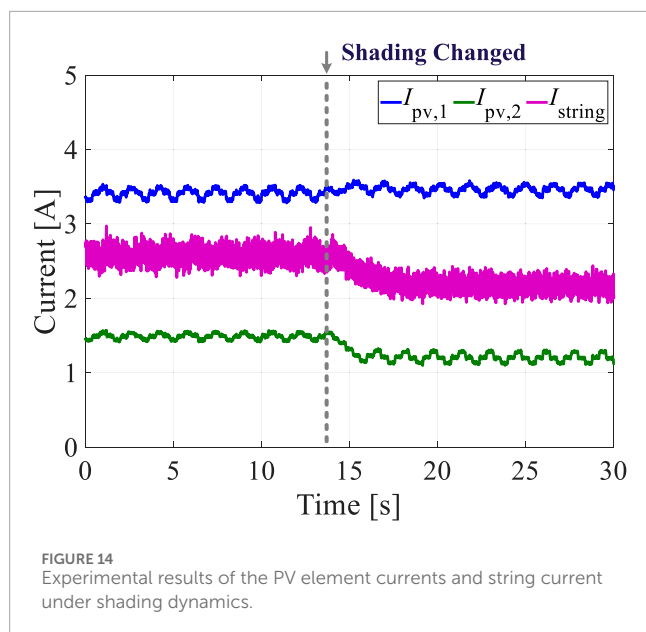


FIGURE 13
Diagram of the PV-NIP DPP experimental system.



the next second (Case II), the irradiation on PV elements 1 and 2 were 1,000 and 550 W/m^2 , respectively. During the third second (Case III), the irradiation on PV elements 1 and 2 were 1,000 and 600 W/m^2 , respectively. During the last second (Case IV), the irradiation on PV elements 1 and 2 were 1,000 and 1,000 W/m^2 , respectively. Figure 9 shows the output voltages satisfying the tracking for the MPPs of the two PV elements. Figure 10 shows the changing output currents of the two PV elements that match the corresponding MPPs during the stepped irradiation changes. The string current and its reference value are also shown in Figure 10. The current reference changes with irradiance and is equal to the average of the maximum and minimum measured PV currents. Furthermore, the reference changes immediately balance the unbalanced DPP currents upon changes in irradiation, as shown in Figure 11; the sign of the signal is repositioned in the direction of the differential current, which is

positive when flowing from the PV element to the DPP converter and negative when flowing from the DPP converter to the relative PV element. Figure 12 shows the differential power distributions and net differential power values of the two DPP converter units. The proposed control scheme aims to achieve relative balance of the DPP power in each converter, especially in units connected to PV elements receiving the maximum and minimum irradiation, which is used to balance the extreme values between the converter units; here, the effectiveness of the proposed UPRB scheme is validated through simulations under shading dynamics.

5 Experimental validations

To verify the effectiveness of the proposed scheme, experiments were carried out with prototypes based on the PV-NIP DPP architecture shown in Figure 4. Considering the characteristics of the proposed control scheme, the power rating of the flyback-based DPP converter unit was designed to be 40 W with a switching frequency of 20 kHz to cover the worst possible range of shading mismatches. The transformer inductance in the flyback DPP unit is 300 μH . The centralized boost converter is rated at 200 W and has a switching frequency of 20 kHz. Here, the programmable DC supply (RIGOL DP832) serves as a controllable current source and is connected in parallel with the MSX-60 PV elements to mimic irradiance changes in the laboratory. The proposed scheme was implemented on the dSPACE DS1104 R&D Controller Board, and the experimental platform is illustrated in Figure 13.

The experimental setup comprises a simplified DPP architecture with two PV elements and two bidirectional flyback DPP converters. The sample time for the algorithm is 0.4 s, and the data are measured using voltage and current probes before being stored in a host computer. To evaluate the proposed control scheme, a step change in irradiation was initiated for PV element 2 at 14 s, while the irradiation on PV element 1 remained unchanged. The current through each PV element during the entire recording time is shown

in Figure 14. When the irradiation on PV element 2 changed, the three-level oscillation observed at steady state was broken, and the current was regulated to the next steady state. During changes in irradiation, the string current changes in response to the changes in the PV elements, and the measured string current is the average of the unit currents. The tracking error between the string current and its reference is shown in Figure 15; here, the string current follows the changes in its reference under shading dynamics, and the tracking error is successfully restricted to the range of -0.05 A to 0.05 A. As shown in Figure 16, the differential power of each DPP converter unit can be balanced using the proposed method. Thus, the effectiveness of the proposed control scheme is validated through experiments under shading dynamics.

6 Conclusion

Unpredictable mismatch conditions, such as partial shading, manufacturing tolerances, thermal gradients, dissimilar aging, or dirt, can result in mismatched output characteristics of series-connected PV elements. DPP architectures are promising candidates for future PV generation systems as they help achieve higher efficiencies by processing only a portion of the full power generated while achieving the individual MPPs. In this work, we analyzed the differential power distributions in a bidirectional flyback-based PV-NIP bus DPP architecture. Then, we propose a UPRB scheme to reconfigure the differential power distributions among the unit converters with the aim of ensuring submodule-level optimization and enhancing the performance of the entire PV system. To achieve the individual MPPs of the PV elements, a P&O-based MPPT control is implemented in each DPP converter unit. To balance the power distribution among all DPP converter units in the system, UBPT control was implemented to optimize the string current. The proposed control scheme was verified using both simulations and experiments under shading dynamics and shown to reliably balance the differential power distribution.

Data availability statement

The raw data supporting the conclusions of this article will be made available by the authors without undue reservation.

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Author contributions

JN: data curation, formal analysis, funding acquisition, methodology, validation, writing – original draft, and writing – review and editing. ZZ: formal analysis, methodology, software, and writing – review and editing. YZ: formal analysis, funding acquisition, methodology, validation, writing – original draft, and writing – review and editing.

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Conflict of interest

Authors JN and ZZ were employed by PowerChina Huadong Engineering Cooperation Limited.

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A review of dynamic voltage support for power grids with large-scale penetration of renewable generation

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Dynamic voltage stability analysis and control of power systems with large-scale penetration of renewable energy have gained significant attention in relevant fields. This article provides the latest development of dynamic voltage support technology under such a scenario, covering the following four main aspects. Firstly, the mechanism and challenges of dynamic voltage analysis with the development of large-scale renewable energy are described, and the state-of-the-art status and problems of dynamic voltage support are discussed. Second, the aggregation modeling method suitable for dynamic voltage support of both photovoltaic (PV) and wind farms is analyzed, considering the dynamic voltage response characteristics and control technologies. Then, the innovation of multi-types reactive power sources coordinated control based on autonomous decentralized control and its realization are discussed. Finally, the key problems of dynamic voltage support technology to be solved in the renewable energy-penetrated power grids are described, and the future research and development work is outlined.

KEYWORDS

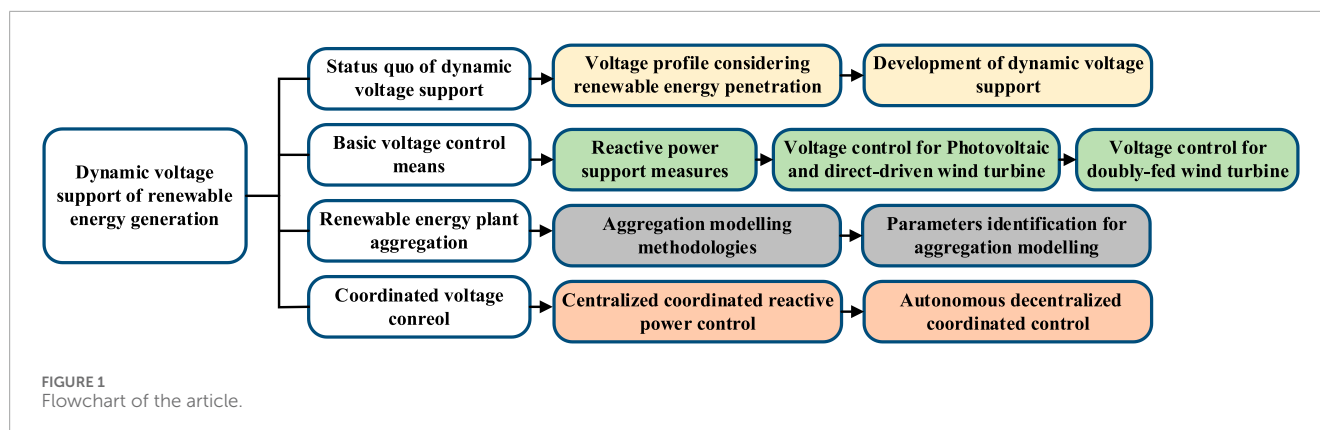
dynamic voltage stability, large-scale penetration of renewable energy, dynamic voltage support, multi-type reactive power sources, coordinated control

1 Introduction

The large-scale penetration of renewable energy is deeply changing the operation characteristics of the traditional power grids. Renewable generations are characterized by stochasticity, decentralization, low inertia, and low immunity (Wu et al., 2023a; Zhang et al., 2023; Zhang et al., 2024; Huang and Xu, 2024). Both the steady-state operation and dynamic process of the grids, as well as the frequency and voltage characteristics, will change to significantly with the high penetration of renewable resources (Huang et al., 2024; Shang et al., 2022).

The frequency and power angle problems essentially result from the imbalance of the active power, which is determined by the frequency support of renewable-penetrated power grid as discussed in Reference (Chen et al., 2024). Thus, this article focuses on the dynamic voltage problem and the relevant control technologies of the renewable generations.

The emergence of dynamic voltage problem in large-scale renewable-penetrated power systems is brought about by the following intrinsic properties: 1) Fluctuating output power of renewable energy units will exacerbate the power imbalance of the system, thereby causing voltage fluctuations; 2) The transmission and consumption of large-scale renewable generations are usually based on high-voltage direct current (HVDC)



devices, which may face challenges such as power regulation, commutation failure, and unipolar/bipolar blockages, leading to active/reactive power variations and aggravating dynamic voltage issues; 3) With increasing renewable penetration, the grid strength dominated by conventional synchronous generators weakens, making the system more prone to dynamic voltage fluctuations. In practice, voltage problems caused by high-proportion renewable-penetrated grids have occurred in power supply centers (e.g., Xinjiang, Qinghai, Ningxia Province) and load centers (e.g., Shandong, Jiangsu Province) in China (Chow, 2023; He et al., 2021). While multiple factors contribute to grid stability, the low voltage support capabilities of some wind and PV generation systems have played a significant role in recent notable grid stability disturbances, such as the “9.28” South Australia blackout in 2016 (Australian Energy Market Operator, 2016) and the “8.9” United Kingdom blackout in 2019 (National Grid ESO, 2019).

On the other hand, renewable energy generations provide foundational support for grid regulation and control, despite their operational challenges. The operation of PV and wind generation systems is enabled by multiple types of converters, which offer advantages such as flexible control, rapid response, and decentralized deployment. Consequently, dynamic voltage control at the point of common coupling (PCC) can be achieved through well-designed control strategies.

Therefore, as shown in Figure 1, this paper first discusses the mechanism and technologies of dynamic voltage support for the renewable energy power generations in Section 1. Second, the basic control means of voltage profile are summarized and analyzed in Section 2. Then, the aggregated modelling method used for dynamic voltage support of PV and wind power plants is explored, followed by a description of decentralized coordinated control of multiple reactive sources. Finally, the paper concludes by pointing out the key issues that need to be further addressed for large-scale renewable-penetrated power grids.

2 State of art in dynamic voltage support of renewable energy power plant

Voltage problem is one of the basic problems in power system operation and control, which can be divided into steady-state and

dynamic stage. In both stages, the voltage stability and voltage quantity are both significant. It should be noted that the voltage problem has great diversity in different time-scale in terms of description and solution.

Table 1 summarizes the voltage problem from the perspective of time-domain. The dynamic voltage problem is usually discussed in terms of root mean square (RMS), which refers to the characteristics of voltage changes during the transient process on time-scales of milliseconds to minutes, including voltage dips and rises, voltage variations in Table 1, all of which are dynamic voltage phenomenon (Hatziaargyriou et al., 2021). These dynamic voltage phenomena may be caused by fluctuations in loads and power resources during the normal operation of the grid, or due to various types of faults. Dynamic voltage process and voltage stability are two closely related concepts: On the one hand, voltage stability is embodied through the process of dynamic voltage evolution; On the other hand, the control of voltage dynamics is the basic means to realize voltage stability control (Blaabjerg et al., 2023). Dynamic voltage support is defined to reduce voltage fluctuations and maintain a stable voltage level by providing “timely” and “appropriate” dynamic reactive power.

From the perspective of frequency domain, after the renewable energy generation is connected to the grid through the converters, the nonlinear behavior of the converter leads to harmonics in the voltage dynamics. At the same time, as a large number of renewable energy generations are connected, the power grid presents a smaller short-circuit ratio and damping ratio. Thus, the renewable energy plants and the power grid is prone to oscillation, the frequency of which ranges in tens to thousands of Hertz (Hatziaargyriou et al., 2021), bringing about voltage fluctuations and stability problems. These issues have been systematically discussed in other references and will not be analyzed in depth in this article.

From the early stage of power systems, researchers have analyzed and studied the voltage problem and proposed systematic solutions. However, the large-scale penetration of renewable energy resources represented by PV and wind power has brought a series of essential changes to the operation and control of the grid, and the attributes of the voltage problem and its control strategy have also presented new features at multiple levels (Huang et al., 2024). During the decades of development of PV and wind power, the grid-connected technology has been continuously developed and progressed (He et al., 2021). Before 2010, the core of the technology lay in the maximum power

TABLE 1 Description of voltage problems based on time-scale.

Voltage phenomena	Time scale	Description	Solution
Voltage pulse/Voltage transient	<10 ms	Instantaneous value	Suppression from source
Temporary drop/rise	10~500 ms	Instantaneous/valid value	Dynamic voltage regulation (DVR)
Voltage fluctuation/flicker	10 ms~1 min	Valid value (rapid change)	Static var compensation (SVC)
Undervoltage/overvoltage	>1 min	Valid value (differential equation)	Source and load adjustment
Voltage deviation	0.5~10 min	Valid value (algebraic equation)	Capacitors/reactance switching

tracking, which operates according to the fixed power factor, and is manifested as “unsupported and low disturbance” (Wu et al., 2023b). After 2010, with the increasing proportion of renewable energy capacity, these units have had a significant impact on the traditional power grids, and the voltage control strategy has also presented new characteristics at various levels.

With the more and more prominent impact of renewables on the power system operation, the industry has begun to carry out research and application of “grid-friendly” units, generally realizing reactive power regulation, low-voltage ride-through (LVRT) (Liu et al., 2014; Gong et al., 2023; Garcia and Santana, 2024; Martínez-Treviño et al., 2021; Fang and Zhang, 2022), and reactive power control in steady-state operation at the plant level (Cheng et al., 2023).

According to the national standards for energy industry, the requirement of converters used for connection of PV and wind power in normal operating conditions are as follows: When the output active power of inverter is greater than 50% of its rated power, the power factor should not be less than 0.98 (forward or lagging); The output active power at 20%~50% of the rated power, the power factor should not be less than 0.95 (forward or lagging) (Fang and Zhang, 2022). It can be seen that the converter can provide 20%–30% of the rated capacity of reactive power under normal operation. For converters with grid support capability, 48% of the rated capacity of the reactive power can be dynamically supplied (Naz et al., 2021). In fact, the output reactive power can play an important role in voltage control.

The reactive power control modes can be adopted in different ways according to the needs of grid operation (Jiao et al., 2022), including constant voltage control, Q-V sag control, constant power factor control and constant reactive power control, etc. The converter is capable of accepting commands from the control system to output reactive power and switching between various control modes online (Tamboli and Jadhav, 2018).

2.1 Steady-state voltage control at renewable energy plant and the correlation with dynamic voltage support

For the reactive power/voltage control in the steady-state operation of the power grid, PV and wind farms have formed a mature control technology (Keskes et al., 2021; Chen et al., 2023; Djalal et al., 2023; Pal et al., 2023; Khan et al., 2021;

Rajeev and Agarwal, 2020; Wang et al., 2021; Richard et al., 2020; Liu et al., 2023). So far, PV inverters, doubly-fed wind turbines (DFIG) and direct-driven wind turbines have generally adopted vector control technology, which realizes the decoupling control of active and reactive power, and lays the foundation for the reactive power support of PV and wind power plants to the power grid (Hadavi et al., 2022; Li et al., 2020; Qi et al., 2020; Fang et al., 2021; Ikram et al., 2023; Hadavi et al., 2022; Gui et al., 2024; Bhyri et al., 2024).

The automatic voltage control (AVC) system set up in renewable energy plant can maintain the voltage at the PCC within the required range by coordinating and controlling all the local reactive power resources using an optimization algorithm. The AVC system control strategy of a renewable energy plant is mainly carried out by coordinating the distribution of reactive power output from each reactive power resource in the plant. Reactive power allocation among units mainly includes: Equal power factor allocation, equal margin allocation, capacity sag allocation and optimal allocation. Optimized allocation is carried out by establishing an objective function and applying an optimization algorithm to calculate the reactive power output of each unit with reactive power regulation capability in the plant. The optimization objectives can be various, such as minimizing active losses, minimizing voltage fluctuations at the PCC of the wind farm, and minimizing operating costs, etc. The commonly used optimization algorithms are traditional optimization methods, such as interior point method and intelligent algorithms such as genetic algorithm and particle swarm algorithm.

The steady state reactive power regulation of renewable energy plants has been generally realized worldwide (Xu et al., 2021; Ahn et al., 2014; Ahn et al., 2013; Wu and Wang, 2021; Zhang et al., 2021). Wind farms from GE could realize reactive power/voltage control with a delay time of 100 ms and a response time of 400 ms in 2004 (Liu et al., 2014). In China, the research of reactive power/voltage control strategies and the development of devices for PV and wind farms started around 2006 (Zhao et al., 2023). Nanrui, Xuji and other electrical equipment companies have developed advanced AVC systems, which have been successfully applied to the voltage control of renewable energy plants (Liu and Wang, 2021).

The relevant departments of China have jointly launched a specification for reactive power configuration for wind power plants (Xia et al., 2016), which sets out clear requirements for the configuration and control of reactive power. Although the steady-state voltage problem is very different from the dynamic

voltage problem in terms of time-scale, measurement indicators and control measures, the two are also closely related.

On the one hand, PV and wind power converters have fast reactive power regulation capability, and various reactive power compensation devices configured in renewable energy plants, such as static reactive power compensator (SVC) and static var generator (SVG) are capable of achieving dynamic voltage regulation (Palanimuthu et al., 2023). These devices involved in steady-state voltage regulation are also the basis of the equipment to achieve dynamic voltage support. On the other hand, the steady-state operating state of the converter determines the static operating point of reactive power regulation, which is related to the range and margin of dynamic reactive power regulation (Ibrahim et al., 2022).

The steady-state reactive power control process should also enable the renewable power plants to have sufficient reactive power reserve capacity to cope with system faults and provide effective voltage support during the transient process. The following section focuses on the dynamic voltage support of PV and wind farms and discusses the dynamic voltage support of different types of standalone devices, the aggregation response characteristics at the plant level and the coordinated control strategies between different reactive power resources.

2.2 The research of the dynamic voltage support at renewable energy plant

With the development and implementation of the renewable-penetrated power systems, when the proportion of renewables is greatly increased to become the main body of power systems or even the main body of power quantity, the dynamic voltage support from the renewable units has become critical, and has become one of the common concerns of the power system (Tong et al., 2015).

Active support based on auxiliary devices, including crowbar protection, is commonly applied in DFIGs to limit over currents in the rotor, thereby protecting the converters (Naderi et al., 2019). However, activating crowbar protection places the DFIG in asynchronous operation, consuming significant reactive power, and deteriorating the wind farm voltage, which is detrimental to voltage recovery in wind power grid-connected systems (Firouzi and Gharehpetian, 2018). Utilizing a DC-link chopper can effectively suppress internal over voltages within wind turbine generators (WTGs) and enhance the wind turbine's voltage disturbance rejection capability (Dong et al., 2021). Implementing fault current limiters (FCLs) is an effective solution for addressing fault current exceeding limits in wind power systems. This device can significantly improve the WTG's LVRT capability, providing strong support for the application of DFIG transient voltage active support technologies (Ahmidi et al., 2012). Current research indicates that applying FCLs during grid faults has a positive effect on maintaining the WTG's reactive power voltage support capability.

The Dynamic Voltage Restorer (DVR) is an effective device for mitigating grid voltage sags, enabling wind power transient voltage active support, and enhancing the LVRT capability of WTGs and wind farms. In practical applications, the DVR is typically connected in series between the WTG and the grid via an injection transformer. It serves to isolate grid faults, rapidly restore voltage, and improve the WTG's LVRT performance (Torkaman and Keyhani, 2018). Its

topology and typical control scheme are illustrated in Figure 2. Literature (Chen et al., 2018) demonstrates that a DVR can maintain the wind turbine's terminal voltage under fault conditions and achieve voltage sag compensation for the wind power grid-connected system. The combination of FCLs and DVRs, as explored in (Flannery and Venkataramanan, 2008), effectively enhances the WTG's immunity to disturbances and increases the unit's voltage support capability during faults. Integrating a DVR with an energy storage system (ESS), as presented in (Kanjia et al., 2014), can effectively suppress wind power fluctuations and compensate for grid voltage disturbances.

Furthermore, some scholars have proposed modifying the topology of wind turbine generators to improve their fault ride-through performance and active voltage support capabilities. Literature (Ambati et al., 2015) proposes a WTG topology based on series converter compensation of grid voltage to support the voltage at the wind power grid-connection point under fault conditions. Through the improvement of the DFIG grid-side converter structure, wind farms can meet the reactive power needs of the grid-connected system and eliminate the influence of grid voltage changes on the wind power system operating voltage, as demonstrated in (Zhu et al., 2017). Active transient voltage support technology of DFIGs based on auxiliary devices relies on the correct and effective operation of external devices. This technology can not only improve the LVRT performance of wind turbine generators, but also enable the wind farm to inject reactive power into the grid-connected system during grid faults, realizing active support for the grid transient voltage support of the wind farm.

Active support methods based on wind power's own control strategy are important measures to maintain voltage stability of new power systems with wind power as the core power source. Transient voltage active support technology based on wind power's own control strategy can realize the safe grid connection operation of wind power from the control perspective and make full use of the available reactive power capacity of the unit to implement voltage support for the grid-connected system under fault conditions. Low voltage ride-through control technologies, such as rotor current control (Zhou and Blaabjerg, 2018), de-excitation control (Xie et al., 2013), and improved speed control (Li et al., 2018), can improve the wind turbine's operating control to achieve the purpose of suppressing the unit's overcurrent and overvoltage. This type of control can effectively protect operating equipment, improve and enhance the transient operating characteristics and LVRT capability of wind turbines, and lay the foundation for the implementation of active support technology in DFIG wind farms. Transient voltage active support technology based on the WTG's own control strategy can actively provide reactive power to the grid and assist in restoring system voltage during grid voltage dips by adjusting the control structure, improving the control strategy, and optimizing the control algorithm. In terms of adjusting the control structure, current research mostly adopts the method of adding control links to achieve active support. In terms of optimizing control algorithms, multiple types of optimization methods are used to improve the control performance of the unit. Based on the coefficient of variation method, literature (Huang et al., 2019) proposes a multi-machine reactive power support allocation method and reactive power control method considering the operating differences within the wind farm. A typical control schematic diagram of transient voltage

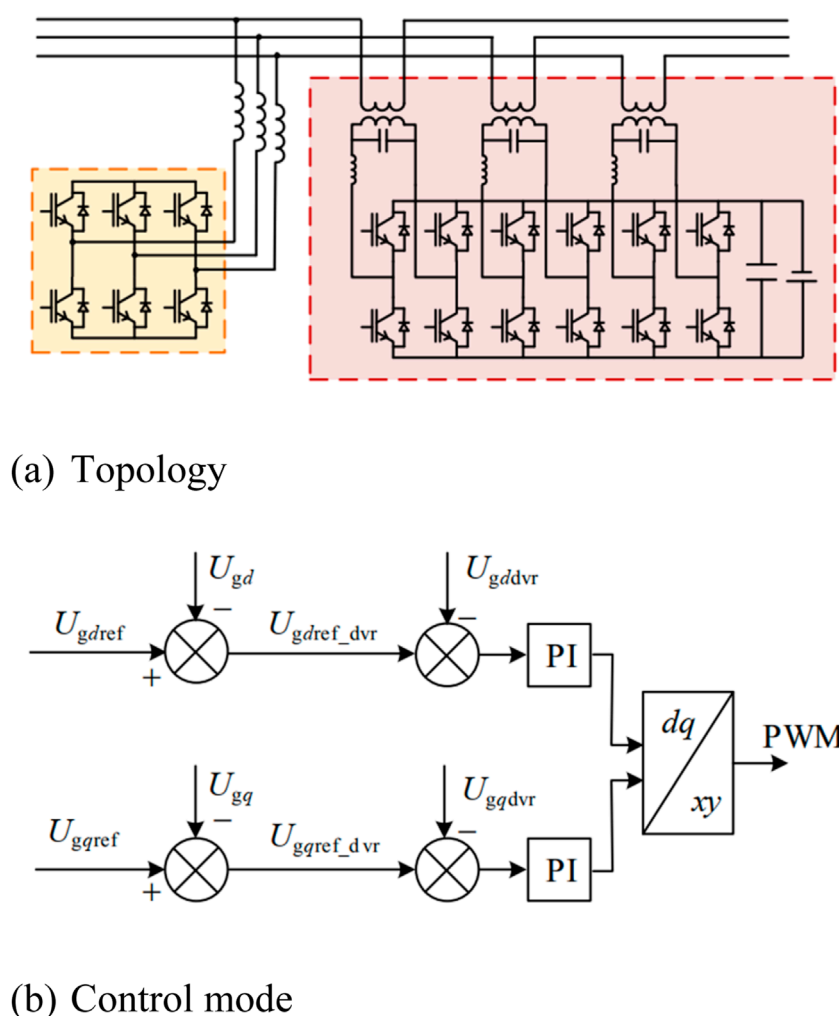


FIGURE 2
Topology and typical control of dynamic voltage restorer. (a) Topology. (b) Control mode.

active support technology based on the unit's own control strategy is shown in Figure 3.

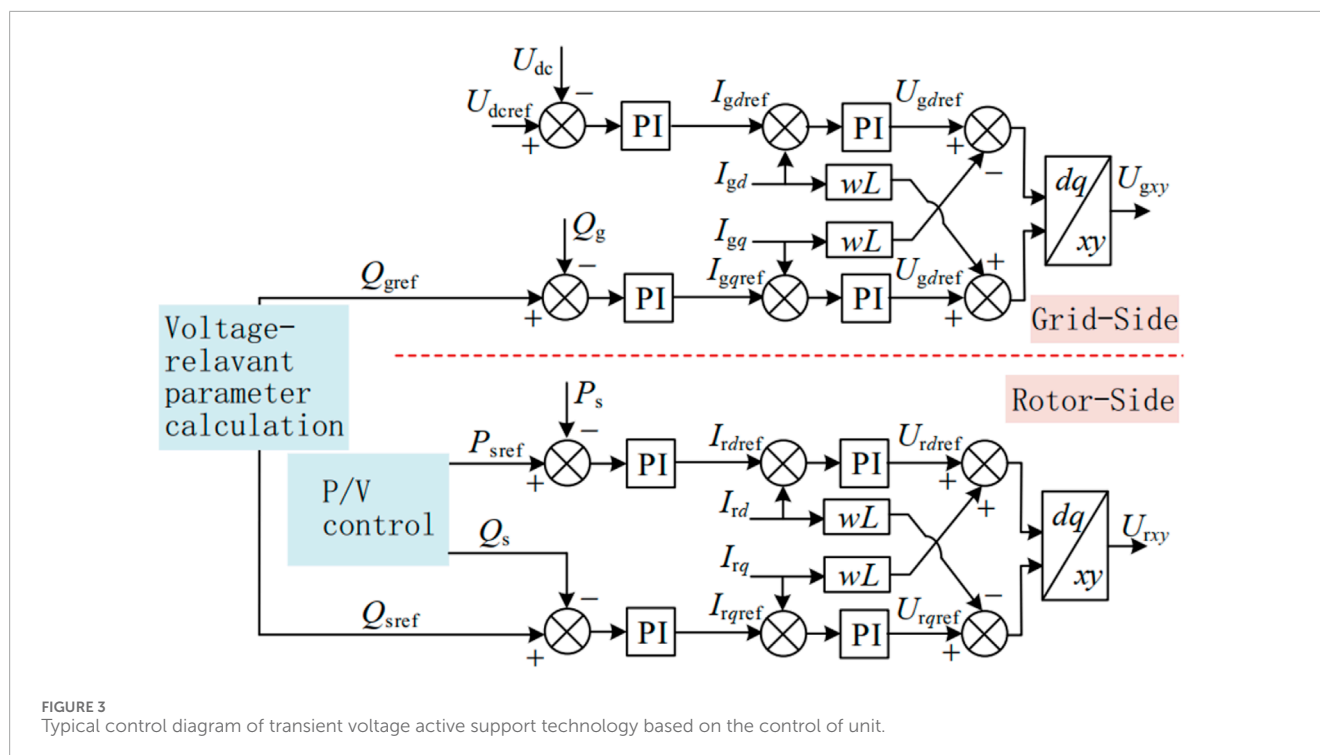
The role of PV and wind turbines in supporting the dynamic voltage of the power grids starts with the importance, development and application of low/high voltage ride-through (LVRT/HVRT) technologies of the units (Du et al., 2021). The LVRT standard specifies the fault ride-through capability for magnitude and time durations of different voltage dips as well as the response time of dynamic reactive current and the relationship between different voltage dips and the dynamic reactive current provided (Xia et al., 2016).

The application of these technologies and the implementation of the standards have improved the support of renewable units for dynamic voltage to a certain extent (Joseph et al., 2020; Lee et al., 2021). However, up to now, the LVRT and HVRT technologies are still limited to the fault ride-through of single units, and the research hotspots mainly focus on the negative sequence current control under asymmetrical fault at the level of single units, high voltage ride-through, and continuous ride-through realization.

Virtual synchronous generator (VSG) technology can improve the voltage stability in the weak grid (Mohiuddin and Qi, 2022), but due to the overload capacity limitations, low-voltage process is usually switched to phase-locked synchronous current source control mode, which cannot present a real characteristic of dynamic voltage source (Alonso et al., 2022).

In recent years, some enterprises and research institutions in China have carried out new support technologies: Nanrui has piloted fast power regulation technology for PV plants in Tibet, Inner Mongolia, Gansu, Yunnan and other places since 2018, adopting GOOSE communication protocol to greatly reduce communication delay. Therefore, the response time of active and reactive power control can be reduced to 30 ms in non-fault conditions (Chen et al., 2023).

In conjunction with the actual situation in China, China Electric Power Research Institute has researched the technology to improve the dynamic stability and consumption capacity of PV and wind power plants through the dynamic control strategy and parameter optimization, which has achieved good performance (Aboshady et al., 2023). However, these techniques



focus on the initiative regulation of PV and wind power plants, emphasizing the role of centralized control.

The above analyses show that the problem of voltage support for PV and wind farms in steady state and small disturbance scenarios for a high proportion renewable penetrated power grids has been basically solved. However, the support for dynamic processes considering large disturbances is still limited to fault ride-through at the single-unit level.

In the future, renewable energy will play a dominant role in the security and stability of the power system, and there is an urgent need to expand the support scenario of PV and wind power from steady state and small disturbances to dynamic and large disturbances, and to expand the support mode from single-unit response to single-unit-plant-grid cooperative.

3 Basic means of dynamic voltage control

In response to the fluctuation of voltage, renewable energy plants need to compensate enough reactive power to ensure the stability of voltage (Qiao et al., 2009). At present, China has put forward relevant specifications for the reactive power compensation capability of renewable energy plants, requiring wind farms and PV plants to be equipped with reactive power compensation devices, and requiring wind turbines to be equipped with reactive power control systems and PV inverters to have reactive power compensation capability (Tong et al., 2015).

This section first introduces typical dynamic reactive power regulation measures, and then introduces voltage control techniques for PV inverters and wind turbine converters. Relevant

reactive power control characteristics at the plant level will be discussed in Section 4.

3.1 Typical dynamic reactive power regulation measures

Typical dynamic reactive power regulation means include synchronous condenser, SVC, SVG, and direct current reactive power modulation, etc. (Qi et al., 2020). The advantages and disadvantages are listed in Table 2. Among them, the most common mechanical rotating reactive power compensation device: the synchronous condenser, is the earliest equipment applied to power system for dynamic reactive power regulation and voltage control. The synchronous condenser can be equivalently regarded as a synchronous generator with zero active power output. During normal operation, it works at synchronous speed and regulates the voltage by over-excitation or under-excitation according to the demand of the power grid (Richard et al., 2020; Liu et al., 2023; Hadavi et al., 2022; Li et al., 2020).

Because the synchronous condenser is a rotating device, the response process requires hundreds of milliseconds, which leads to high operation and maintenance cost, and necessitates investment in supporting starting and protection equipment. Thus, it was once regarded as an obsolete technology (Gui et al., 2024). However, in recent years, with a high proportion of renewable energy resources integrated into the power grid, new requirements on the inertia and dynamic voltage support of the system, synchronous condenser has gained favor again with a certain degree of rotational inertia support and characteristics of the dynamic reactive power compensation.

Static reactive power compensator (SVC) is a kind of static reactive power compensator that uses thyristors as the switching

TABLE 2 Comparison of typical dynamic reactive power regulation measures.

Measure	Advantages	Disadvantages
Synchronous condenser	1. Overloaded capacity 2. Rotation inertia 3. Dynamic reactive power supply 4. Suitable for variable scenario	1. Rotating device 2. High cost
SVC	1. Low cost 2. Steady reactive power 3. Large capacity of capacitor	1. Limited by voltage profile 2. Resonance risk
SVG	1. Low cost 2. Flexible control mode 3. Fast response speed	1. Failed in imbalanced fault 2. Small capacity of capacitor
HVDC	1. Flexible control mode 2. Fast response	1. Complex structure 2. Limited to specified scenario 3. High cost

device composed of inductors, fixed or variable capacitors in parallel. The main SVC devices are thyristor switched capacitor (TSC), and thyristor controlled reactor (TCR)+fixed capacitor (FC), etc. A TSC can achieve zero voltage input and zero current withdrawal of reactive power compensation. A TCR + FC can continuously regulate the reactive power between output and absorption, so that the voltage can be quickly and dynamically regulated, with a response time between 50 and 100 ms.

Static reactive generators (SVG) are based on fully controlled devices that constitute the voltage-type or current-type grid-connected inverters, which can regulate reactive power quickly, stably and continuously, among these, voltage-source three-phase bridge inverters are the most widely used devices. SVG has two reactive output modes: Fixed reactive power control or specified voltage control of the PCC.

Compared with SVC, which requires large capacity capacitors or reactors, SVG only requires small capacity capacitors on the DC side and can both supply and absorb reactive power (Li et al., 2022). Moreover, due to the limitation of impedance of inductance and capacitance, the maximum current compensated to the grid by a traditional SVC decreases with the voltage reduction, while the SVG can adjust the AC voltage amplitude and phase by switching the device, and its maximum current compensation is not affected by the voltage magnitude (Xia et al., 2016).

In addition, SVG has a fast response speed, the response time from zero reactive power output to maximum reactive power output is about 20 ms (Ibrahim et al., 2022), which meets the requirement of national standard “Reactive Power Compensation Response Speed Less Than 30 ms”, and the harmonics are small, so it is gradually replacing SVC as the mainstream reactive power compensation device.

Conventional LCC-type HVDC transmission can be used to regulate the voltage of AC power grid by reactive power modulation (Lee et al., 2021). Especially when overvoltage occurs in the grid, the overvoltage can be suppressed by increasing the trigger angle and increasing the reactive power consumption.

In addition, in recent years, the rapid development and application of voltage-sourced converter (VSC) HVDC

transmission technology, large-scale energy storage technology, etc., rely on the SVC, due to its active-reactive decoupling characteristics (Mohiuddin and Qi, 2022). The principle is similar to that of PV inverters and will be specifically discussed in the next Section.

3.2 Dynamic voltage control methods for PV and direct-driven wind power

Both PV and direct-driven wind farms use VSC-type converters with similar reactive voltage response characteristics among all the renewable energy plants. The control modes of VSC converters can be divided into grid-following (GFL) control and grid-forming (GFM) control (Zhang et al., 2021). The GFL requires a phase-locked loop to phase-lock the voltage at the PCC and to achieve synchronous grid connection (Zhao et al., 2023); the GFM actively constructs the grid voltage and frequency according to the output power of the converter.

Renewable energy plants in China require a certain period of time and a certain amount of reactive power support when a high/low voltage fault occurs in the system, and the grid-connected inverter switches the control mode and adopts an open-loop control strategy with a single current loop. When a high-voltage fault occurs in the system, the grid-connected inverter needs to switch the control mode without taking off the grid for a certain period of time, however, in contrast to LV faults, HVFT ability requires the grid-connected inverter to absorb reactive power, and at the same time, an increase of the voltage reference in the DC-side is usually used to stabilize the fluctuations in the DC-side voltage during the HVFT period (Palanimuthu et al., 2023).

In renewable energy plants, the generation unit usually executes after receiving the command issued for the plant. At this time, the fastest reported reactive power response time can already be comparable to SVG, which is about 23–27 ms (Lee et al., 2021). When high/low voltage faults occur, the voltage detection delay will be increased before the command is issued. The command issued at this time is usually difficult to meet the reactive power support

demand during the fault period, so to improve the dynamic reactive power support capacity of the renewable energy plant, the high/low voltage control needs to be further studied.

Because of the stability problem in the weak power network, the converter should adopt the control mode in the grid with small short-circuit ratio and low system inertia (Alonso et al., 2022). At present, some scholars refer to the virtual synchronous machine control and have applied it to the PV and wind power to improve the stability of the system in low inertia and weak power grids (Aboshady et al., 2023).

In addition, commonly used GFM control also includes sag control and virtual oscillator control (Ghosh et al., 2023). The sag control can realize global power distribution without communication by simulating the sag curves of active power-frequency and reactive power/voltage of the synchronous generator (Camal et al., 2023). However, due to the influence of line impedance, accurate power distribution is difficult to achieve (Wang et al., 2024).

Virtual oscillator control obtains a sinusoidal voltage to the oscillator, which is used as a reference voltage to control the output of the inverter (Ghosh et al., 2023). Because the converter external characteristics are represented by the physical characteristics of the oscillator, the synchronization between different converters and power distribution can be achieved through the resonance principle of the oscillator (Camal et al., 2023).

It is important to note that VSC converters usually use insulated gate bipolar transistor (IGBT) fully controlled switching devices to achieve commutation, and the IGBT's current flow capability has limitations (Palanimuthu et al., 2023). Therefore, when the system has a serious fault, even if the VSC converter uses a GFM control strategy, the VSC converter may degenerate into a fixed power control because of current limiting.

3.3 Reactive-voltage control methods for doubly-fed wind turbines

DFIG is currently the mainstream model for wind power generation. The stator of DFIG is directly connected to the grid-connected transformer; the rotor is connected to the grid-connected transformer through the rotor side converter (RSC) and the grid-side converter (GSC) (Chang et al., 2020). Controlled by RSC, the stator side outputs the maximum active power and inductive reactive power to maintain the voltage. GSC controls the rotor power output, similar to PV converters. When there is a transient voltage drop, the unit provides reactive power to the system to help restore the grid voltage.

The maximum reactive power output capacity of DFIG can reach more than 30% of the rated capacity. Based on the existing control strategy, the dynamic reactive response speed of the DFIG is about 60 ms (Bhyri et al., 2024). Compared with PV units, wind turbines are more prone to large-scale off-grid due to high voltage during voltage recovery. At present, some countries have required wind turbines to have HVRT ability. Literature (Du et al., 2021) proposes a method of realizing the HVRT through the collaborative control of the main control system and the converter of wind turbine. However, the existing research has not considered the influence of

overvoltage on the electromagnetic characteristics of the unit, and HVRT technology must be studied further.

In addition, a virtual synchronous control strategy applied to DFIG has been proposed to simulate the rotor equation of synchronous generators, including active power control, reactive power control, damping control, etc. Virtual impedance control of DFIG is applied to RSC, while GSC still adopts voltage-oriented vector control based on phase-locked synchronization. However, when a voltage drop occurs in the power grid, virtual synchronous control will cause the RSC to be impacted by strong electromagnetic stress and rotor overcurrent, and the unit will face the risk of off-grid.

4 Aggregation characteristics of dynamic reactive power support for renewable power units

4.1 The requirement of aggregation modelling

The PV and wind turbine units are dispersed with random and fluctuating output (Camal et al., 2023). Thus, their active/reactive output characteristics and voltage response characteristics are mainly determined by the control mode of power electronic devices. Existing research on renewable energy units has provided a mature summary for response characteristics at equipment level (Wang et al., 2024).

However, as the proportion of PV and wind units connected to the system increasing year by year, the role of renewables is gradually changing from auxiliary power supply to the dominant power supply of the power system (Li et al., 2022). Therefore, it is inevitable to set the maximum power tracking as the basic control objective of the controlled current source characteristics to support the voltage and frequency stability of the power system (Wang et al., 2016). The dynamic voltage control of power plants has become an important technical bottleneck restricting the development of renewable energy.

Based on the characteristics of voltage response at the PCC, it is difficult to achieve the analysis of dynamic characteristics through the model of the single unit when designing the dynamic voltage regulation strategy (Ibrahim et al., 2022). In order to accurately grasp the voltage response characteristics of renewable energy plants and fully reflect the dynamic reactive power support capability of PV and wind power plants under different control strategies and topologies, it is urgent to explore and establish of a dynamic voltage support research direction. The research on aggregation model for dynamic reactive power support should be able to fully reflect the autonomous decentralized response characteristics of the units under transient disturbances on the one hand, and characterize the overall effect and multi-time-scale dynamics of voltage support for the PCC in the plants on the other hand.

Providing an accurate analytical model for the dynamic voltage stability analysis of the power plants through an aggregation modelling is a prerequisite for the subsequent analysis of dynamic reactive power support and the design of reactive power compensation strategies, and is also the theoretical basis for the

research of the coordinated control of renewable energy units and dynamic reactive power compensation equipment at the plant level.

4.2 Aggregation modelling methodologies

In view of the above modelling requirements, the aggregation modelling should take into account the external natural resources conditions and focus on the response characteristics of the aggregation model to the power system voltage fluctuation. However, there is still little literature on the aggregation modeling of dynamic voltage response characteristics at plant level, and the commonly used aggregation modelling methods for renewable energy plants mainly include mode equivalence method, Singular perturbation method and dynamic aggregation method.

According to the aggregation results, the commonly used aggregation modelling methods include aggregated single-machine modelling and grouped multi-machine aggregation modelling (Shang et al., 2022), which can be flexibly selected according to the analysis scenarios such as the size of the plants and the flexibility of the operating state. This type of modelling method essentially provides a method of simplification and downgrading model. In order to facilitate the design of control parameters, it is usually necessary to retain the mapping relationship between the overall characteristics of the plants and the parameters. Then, the parameter estimation or parameter aggregation is used to group the system to one unit.

The detailed description of the dynamic characteristics of renewable energy resources are still rarely taken into account in the existing aggregated modeling literatures (Camal et al., 2023). In fact, the difference of terminal voltage at different locations in actual operation will directly affect the working mode of the units, and whether they will enter the fault ride through mode.

In view of the research objectives of dynamic characteristics of voltage response, the clustering should consider the relevant indicators including the selection of natural resources, output characteristics, operating environment, the critical value of HVRT/LVRT and control parameters (Palanimuthu et al., 2023), so as to characterize the dynamic voltage response characteristics of the unit and reflect the influence of location dispersion. The hierarchical clustering method is not affected by the initial setting value and can avoid the local optimal characteristics, obtaining a widely application in power systems.

To meet the requirements of dynamic voltage analysis and control, the trade-off between accuracy and complexity should be balanced in modeling (Xia et al., 2016). The applicability of the model should be evaluated through errors and characteristic differences.

4.3 The parameter identification of aggregation model

In order to improve the dynamic equivalence performance of the aggregation model, it is necessary to analyze the factors causing the error of the model, considering the influence of the LVRT strategy on the external characteristics under typical operating conditions and

the quantitative relationship between the capacity of the plant and the error of the aggregation model (Martínez-Treviño et al., 2021).

For the estimation of aggregation parameters, capacity weighting method and equal power loss method have been well applied (Qiao et al., 2009). The equivalent aggregation model of the plant utilizing the equivalent admittance method has been evaluated time-consuming and heavy calculation burden.

The parameters of the aggregated model established by the state-space model method can achieve a better parameter fitting in the optimized equations, and can guarantee the accuracy of the model (Wang et al., 2022). However, all the state-space equations of large wind farms or PV plants in the optimization solutions will lead to high computational burden, and the efficiency of parameter identification needs to be improved.

From the perspective of engineering application, the parameter identification method based on phasor measurement unit (PMU) can further improve the model accuracy. By comparing with actual engineering operation data, PMU-based methods can guarantee the reliable parameter identification results and can be widely used in the parameter identification of aggregate models (Qiao et al., 2009).

5 Coordinated dynamic reactive power control for renewable energy plants

5.1 Basic problems confronted with coordinated control of multi-type reactive power

The output characteristics of renewable units are completely different due to the different device attributes and control strategies of the multi-type reactive power resources in the plants (Asadollah et al., 2020). From the perspective of topology, DFIG is connected to the grid through the multiple energy paths of generators and converters and full-power direct-driven wind turbines are connected through VSC converters (Zhang et al., 2021). PV inverters, whether centralized or serial, are connected to the grid through VSC converters. Different types of renewable energy devices and different connection methods lead to great differences in the reactive power regulation characteristics of these renewable energy units. Therefore, the voltage regulation cannot be achieved by controlling a single reactive power resource.

In the context of high-proportion of renewable energy penetration, especially under weak grid and AC/DC fault conditions, the grids need more reactive power to support the dynamic voltage profile of renewable energy plants, and improper coordination and control may cause reactive power circulation and reactive power oscillation. This puts forward high requirements for the coordinated control of multi-type reactive power sources.

The existing HVRT/LVRT strategies cannot match the demand of voltage support at the PCC, and the control mode mainly relies on the traditional automatic voltage control (AVC) (Nguyen et al., 2021), which uses centralized control as the adjustment method, and its reactive power adjustment time is on the minute level, thereby failing to meet the multi-time scale control demand of voltage support of renewable energy plants.

TABLE 3 Comparison of plant-level reactive power compensation equipment.

Equipment	Response speed	Dynamic voltage support ability	Cost
Condenser	Middle	Strong	Middle
SVG/STATCOM	Fast	Middle	Low
Fixed capacitor	Slow	Weak	Low
Synchronous generator	Middle	Strong	High

According to the current national standard in China, PV and wind turbine inverter is an important reactive power resources compared to SVG and other devices (Lu et al., 2023). How to coordinate and utilize the reactive power compensation capability of renewable energy equipment and solve the problem of insufficient dynamic control capability is the basic challenge of dynamic voltage control of renewable energy plants.

When designing the control strategy, it is necessary to coordinate the response time and capacity of the various types of reactive power compensation equipment according to the different response characteristics of reactive power compensation devices in actual engineering operation. Otherwise, the compensation equipment may be taken off the grid earlier than the renewable energy units during the fault due to the electrical limitation of the equipment (Wang et al., 2013).

As synchronous rotation equipment, generators and condensers can also provide short circuit capacity for the system and dynamic voltage support through excitation (Hadavi et al., 2022). At the same time, the operating characteristics of condensers meet the needs of dynamic reactive power in HVDC transmission and the dynamic process of large-scale renewable energy plants. Compared with SVC, STATCOM and other reactive power compensation devices, the new-type condensers have stronger dynamic reactive power support and dynamic voltage regulation capabilities, which is more suitable for practical application (Ge et al., 2022).

As shown in Table 3, different types of reactive power resources considering the characteristics, response time scale, support capacity, and cost, etc., are summarized and compared.

5.2 Centralized control

The existing technologies that widely used in the industry includes AVC, dynamic automatic voltage control (DAVC) (Lu et al., 2023), whose main control objectives are to optimize voltage deviation, improve static voltage stability and reactive power margin. The static voltage control system of the power grid is mainly designed based on the concept of “voltage control zone” and “central bus voltage”, which has been widely used. However, to meet the voltage regulation requirements of renewable penetrated power systems, it is still necessary to further study dynamic

voltage regulation technology to overcome the problems as follows: 1) AVC systems need to adjust each other among multi-level dispatching centers; 2) The overall voltage control quality is low; 3) The dynamic voltage support ability is insufficient. Literature (Wang et al., 2013) put forward the collaborative voltage control technology of network-provincial-ground multi-level control center, which realized the cooperation among multi-level AVC systems through the interaction of a few key variables. Literature (Ge et al., 2022) proposed a preventive control method aiming at the safety and stability of power grid, where representative serious faults were selected from fully anticipated fault sets and dynamic voltage partitions were formed correspondingly to achieve global optimal control through coordination. Literature (Yang et al., 2019) improved the dynamic support performance of wind farms on voltage through the coordinated control of AVC and wind turbines designed with constant voltage control mode.

When the voltage is disturbed during the fault, the reactive power response of the renewable energy units is completely determined by its own control system design, thus, it is impossible to realize the cooperative control of each unit by means of communication in this time range since the duration of the disturbance is short. Therefore, the characteristic of unit autonomous response determines the fluctuation of voltage profile and the distribution of reactive power. The control parameters and operation mode are the key to the response characteristics of the voltage support, and this characteristic determines whether it will cause cascading failures or expand the accident scope.

The existing control strategy is mainly based on the idea of control mode switching to achieve reactive power voltage control in the fault stage. Thus, the smooth switching control methods is essential for ensuring the stable operation of the control system. After the fault is cleared, where the voltage recovery stage comes, the purpose of collaborative control of renewable energy units should give priority to how to adjust the reactive power distribution of each unit, output power limiting and the influence of active and reactive power coupling on system stability, so as to provide the reactive power to the system to help the voltage recovery. Literature (Heming et al., 2013) discussed the mechanism of the influence of active and reactive power coupling on voltage stability in detail. Literature (Wang et al., 2020) discussed the mechanism of overvoltage and put forward suggestions for optimization of wind turbine to effectively suppress the system overvoltage level during voltage recovery process.

The above dynamic voltage regulation method based on the preset strategy is convenient to implement and easy to operate, and it is an important development direction of the preventive voltage control of the power plants. However, there is still much room for improvement in the optimization dynamic process and the collaborative control of voltage recovery after failure. How to coordinate the reactive power resource at plant level and make full use of reactive power compensation capability of the renewable energy units to reshape the dynamic response characteristics of reactive power on multiple time scales and effectively improve the dynamic voltage support capability of the plant remains to be explored.

5.3 Autonomous decentralized control

In view of the shortcomings of the existing centralized control at the plant level, it is urgent to explore the autonomous decentralized control of multiple reactive power resources at the plant, which is different from the centralized control, so as to realize the full coverage of the process of voltage control from steady state to dynamics, and the control mode change from passive response of single-unit to active support of the plant (Mudaliyar et al., 2020).

Decentralized response-aggregate support is the basic goal of multi-reactive power resource autonomous decentralized control (Derakhshan et al., 2023). Autonomous control uses the active response characteristic of the reactive power resource unit. When detecting the voltage fluctuation, it operates autonomously according to the preset control strategy, and actively adjusts the reactive power output in response to the voltage fluctuation to achieve the control purpose using decentralized response from the device (Samende et al., 2021).

Aggregate support is a basic requirement for dynamic reactive power response (Sun et al., 2020). Based on the obtained global information, multi-objective optimization can be carried out. The aggregate model of the plant discussed in Section 4 is used to analyze the reactive power adjustment margin and the operating state of the operating point, then the dynamic reactive power compensation demand of the plant can be calculated (Xu et al., 2020).

According to the operation mode of power grids and the characteristics of faults, the occurrence, development and recovery process of voltage can be predicted. Then, the distribution of reactive power and control parameters are sent to the units according to refresh cycle, so as to achieve the purpose of aggregate support by dispersed response of the plants.

In device level, the control objects include reactive power compensation devices such as renewable energy units, SVC and SVG, which are mainly responsible for voltage profile detection and uploading, operation mode switching, and online adjustment of control parameters. Based on the above ideas of autonomous decentralized control, the global information of the plant is obtained and the reactive power margin, voltage support capacity and reactive power demand are analyzed first based on the aggregate model, then, multi-objective including network loss and voltage stability indicators are optimized. Based on the optimization results, the reasonable operating point and control parameters of each unit at the current operating point can be given, and the parameters are sent to each unit for adjustment.

When the large disturbance occurs, each unit realizes the quick support of the dynamic reactive power based on the operating point and control parameters in an autonomous control mode. After the disturbance, the unit continues to dynamically adjust the operating parameters according to the received instructions.

Compared with the traditional centralized control mode, the autonomous decentralized control mode has a more flexible adjustment mode, and can improve the collaborative support ability of dynamic voltage. The upper layer control mainly includes (Sun et al., 2020): 1) Dynamic optimization based on the unit operating point and power network operating state obtained by global information, reactive power distribution determination and reactive power compensation margin reservation, which is required in advance to cope with possible

disturbances; 2) Adjustment of action parameters and limits, such as whether to participate in voltage regulation, entry conditions, withdrawal conditions, etc.; 3) The adjustment of sag parameters and control parameters is based on the operating state of the power grid, such as the adjustment of short-circuit ratio. The optimization algorithm based on global information can effectively avoid the local optimal solution and make full use of the reactive power regulation ability of the plant.

The lower layer is the equipment layer, which is mainly responsible for: 1) Communicating with the upper layer and providing the basic operation information of the equipment; 2) Implement upper-layer delivery control policies; 3) Sent adjustment instructions of the control parameters delivered by the upper layer during steady state operation. The architecture of the proposed autonomous decentralized voltage control is illustrated in Figure 4.

The hierarchical coordination can ensure the continuous of dynamics, and the rapid autonomous reactive power response of a single unit can be realized based on the preset operating mode and control parameters (Derakhshan et al., 2023). After the fault is cleared, the adjustment of the single unit based on the global information optimization results helps the system voltage recovery, so that the plant switches to the steady-state operation mode, thereby improving the dynamic reactive power adjustment capability of the renewable energy plant and the voltage support capability of the PCC.

5.4 Future development trend

Although the voltage stability challenges of power systems with large-scale renewable energy have received significant attention, the research and implementation of dynamic voltage support are still in their early stages. This paper discusses the main achievements in voltage support, progressing from individual equipment to plant-level and coordinated control strategies, as presented in the first four sections.

However, due to the inherent complexity of power systems, the stochastic nature of disturbances, and the spatio-temporal multi-scale characteristics of the system response, several significant technical challenges remain to be explored:

- 1) The 10 ms Problem after Disturbance Clearing: As indicated by the analyses presented in the preceding sections, the response times of various reactive power resources differ considerably. The dynamic response time of PV converters and direct-drive wind turbines is typically greater than 20 ms, while that of DFIGs is approximately 60 ms (Wu and Zhang, 2022). The response time of static VAR generators (SVGs) is also more than 20 ms, and that of static VAR compensators (SVCs) is around 60 ms, whereas the response time of synchronous condensers exceeds 100 ms (Z. Rafiee et al., 2022). Consequently, within the initial 10 ms following a voltage transient, insufficient reactive power is typically provided by existing voltage support devices. Therefore, how to effectively leverage renewable energy converters and SVG devices to provide dynamic voltage support represents a crucial area for further investigation.
- 2) Equivalence of Deterministic Controllable Reactive Power Resources: As discussed previously, the dynamic reactive

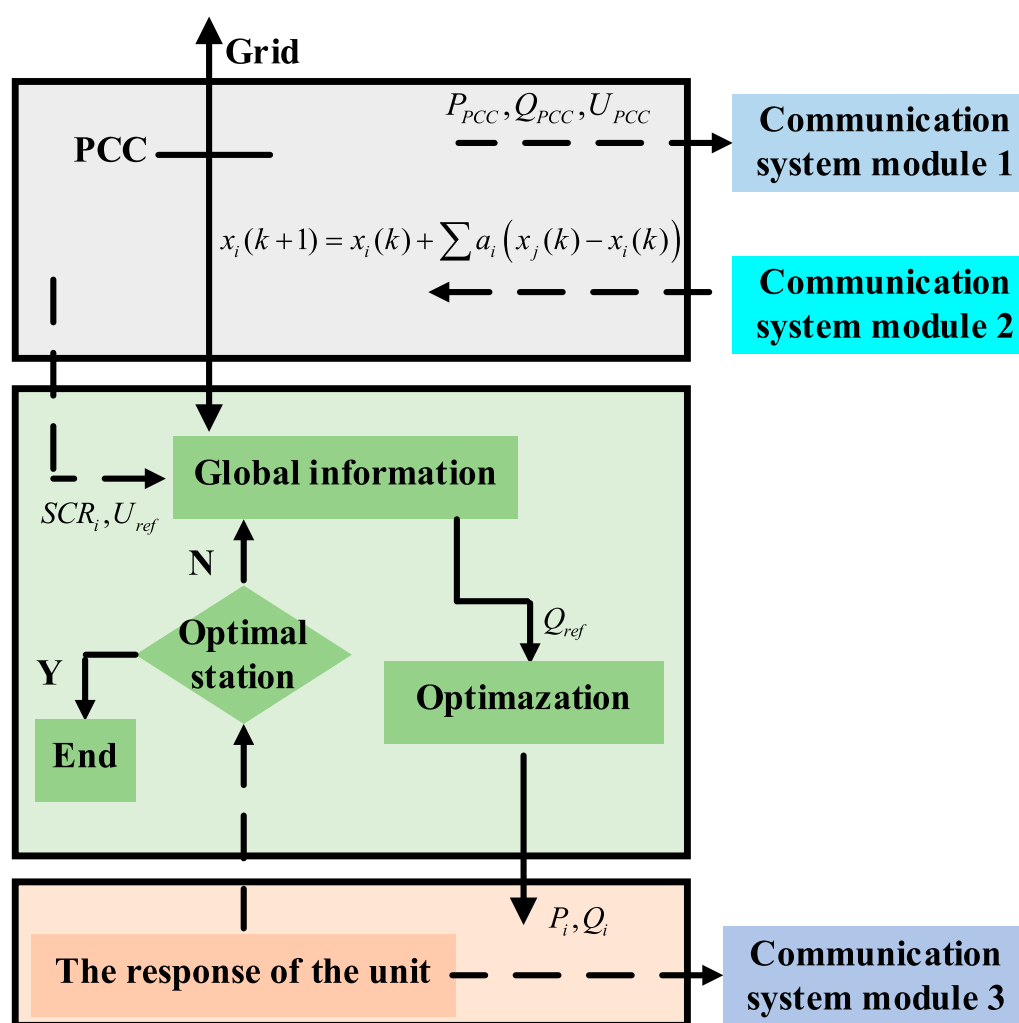


FIGURE 4
The architecture of the autonomous decentralized voltage control.

power compensation performance of PV and wind power converters within renewable energy plants is highly dependent on factors such as device type, operating state, and control strategy (Zhao et al., 2023), resulting in considerable randomness in the overall compensation effect. This inherent variability introduces significant uncertainty into the dynamic voltage support capabilities (Ghosh et al., 2023). An ideal approach involves organically integrating reactive power compensation equipment at the plant level, such as SVGs and synchronous condensers, with the distributed reactive power compensation provided by PV and wind power units, and achieving equivalent deterministic controllable reactive power through coordinated control strategies. Key research areas include accurately characterizing the reactive power compensation characteristics of renewable energy units and coordinating the capacity and response times of various compensation devices.

- 3) HVRT/LVRT Technology and Control Delay: Historically, HVRT and LVRT strategies have often been discussed in isolation. However, low-voltage and high-voltage faults can

occur in rapid succession, with alternating intervals potentially spanning tens of milliseconds. Given that the typical response time of reactive power sources ranges from 20 ms to 30 ms, poorly coordinated control strategies can exacerbate voltage fluctuations, potentially leading to voltage instability [59]. Therefore, future research should focus on addressing the phenomenon of alternating low/high voltage conditions in renewable energy systems and enhancing the response speed of various reactive power resources.

- 4) Autonomous Decentralized Control for Dynamic Voltage Support in Renewable Energy Plants: This control technology necessitates the rapid acquisition and dynamic updating of global information within the plants, the application of multi-objective optimization algorithms incorporating global information, and the development of device-level cooperative control strategies. While global information from renewable energy units is obtained at the substation level, the impact of communication delays and the frequency of information exchange remains unclear. Furthermore, in terms of multi-objective optimization algorithms, establishing a

comprehensive objective function that effectively combines system stability with dynamic reactive power capabilities presents a valuable area for future research.

6 Conclusion

Focusing on dynamic voltage control in power systems with high penetration of renewable energy, this paper *examines* the research focus of reactive power/voltage regulation technologies, from renewable energy converters to auxiliary equipment. The development of dynamic voltage support technology within renewable energy plants is also *reviewed*, and the basic objectives and requirements of dynamic voltage support are defined. The basic idea of *an* autonomous decentralized control approach for multi-reactive power resources is proposed for dynamic voltage support.

According to the dynamic voltage support requirements of renewable energy plants, the article *identifies* the direction of further research work, *spanning* four key aspects. *Given* the increasing proportion of renewable energy in power systems, dynamic voltage support is a *critical* endeavor with strong engineering application potential. From the theoretical analysis of dynamic voltage support to its practical application, it is essential to integrate the advancements in renewable energy converters, reactive power compensation equipment, and the information systems of power plants to achieve effective dynamic voltage support and ensure the stability of power systems.

Author contributions

SG: Conceptualization, Data curation, Investigation, Methodology, Writing – review and editing. HZ: Conceptualization, Funding acquisition, Writing – original draft. XM: Investigation,

Resources, Software, Writing – original draft. YZ: Formal Analysis, Investigation, Visualization, Writing – review and editing.

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Conflict of interest

Authors SG, HZ, XM, and YZ were employed by State Grid Jilin Electric Power Co. Ltd.

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Generative AI statement

The author(s) declare that no Generative AI was used in the creation of this manuscript.

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Exploring the performance of GaN trench CAVETs from cryogenic to elevated temperatures

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Fabricated GaN trench current aperture vertical electron transistors (CAVETs) were characterized across a wide temperature range for the first time, including *in situ* cryogenic measurements down to 10 K and *ex situ* thermal shock testing at elevated temperatures of 773 K and 1073 K. The device featured a highly conductive AlGaIn/GaN channel regrown on p-GaN following trench etching. As the temperature decreased, the field-effect mobility in the regrown two-dimensional electron gas (2DEG) channel increased from 1886 cm²/(V·s) at 296 K to 3577 cm²/(V·s) at 10 K. The device maintained a stable threshold voltage (V_{TH}). The subthreshold slope (SS) decreased from 98.32 mV/dec to 51.31 mV/dec, and the I_{ON}/I_{OFF} ratio increased from 3×10^9 to 9×10^{10} over the same temperature range. The specific on-state resistance ($R_{ON,sp}$) decreased from 1.02 mΩ cm² at 296 K to 0.586 mΩ cm² at 10 K. Furthermore, 1-min thermal shock testing was conducted as a preliminary method to assess the resilience of trench CAVET at elevated temperatures. The device maintained field effect transistor (FET) functionality after exposure to 773 K, albeit with reduced current. Testing at 1073 K resulted in more significant performance degradation, including a sharp increase in $R_{ON,sp}$ and failure to achieve pinch-off due to a pronounced surge in gate leakage.

KEYWORDS

gallium nitride, current aperture vertical electron transistor, vertical transistor, cryogenic electronics, high-temperature electronics, extreme temperature applications

1 Introduction

Power electronics operating in extreme temperature environments have garnered increasing attention for applications such as space, healthcare, and transport systems. Under high-temperature conditions, power devices must retain functionality despite potential performance degradation. Conversely, at cryogenic temperatures, devices are expected to exhibit enhanced electrical performance. Gallium nitride (GaN) is considered a promising candidate for reliable operation across these extreme thermal regimes, owing to its wide bandgap of 3.4 eV and the presence of thermally stable two-dimensional electron gas (2DEG).

As temperature decreases from room temperature (RT), conventional silicon (Si) power transistors benefit from reduced scattering effects, leading to higher carrier mobility, lower on-resistance (R_{ON}), and improved power density with reduced losses (Rajashékara and

Akin, 2013). However, Si devices suffer from intrinsic carrier freeze-out below ~ 100 K, causing a negative temperature coefficient of R_{on} (i.e., resistance increases as temperature decreases) (Ahmad, 1987). Silicon carbide (SiC) devices face additional challenges due to interface state trapping, resulting in degraded current conduction and increased R_{on} (Chen et al., 2013). Unlike Si and SiC power transistors, which rely on impurity doping to enable conduction, GaN CAVETs benefit from a 2DEG channel formed by polarization effects even without intentional doping. This intrinsic mechanism enables high current conductivity, low R_{on} , and fast switching capability at cryogenic temperatures (Nela et al., 2021). As a result, GaN CAVETs are well-suited for cryogenic power electronics applications, such as onboard electronics for space applications and image sensors, where systems operate directly at low ambient temperatures. In such environments, maintaining conventional power devices at room temperature would require bulky and inefficient thermal insulation. Cryogenically compatible electronics, by contrast, offer a more compact and efficient solution (Gui et al., 2020). Therefore, understanding the behavior of power devices at cryogenic temperatures is critical for advancing performance in emerging cryogenically cooled power electronics applications.

High-temperature operations are found in applications such as deep-well drilling, automotive systems, and spacecraft. In particular, some systems must withstand brief exposure to extreme thermal events, where temperatures rise sharply for short durations. Under such high-temperature conditions, solid-state switching devices, which serve as critical components in power electronics converters, must sustain functionality despite potential performance degradation. The 2DEG channel formed at an AlGaIn/GaN heterojunction has demonstrated excellent thermal stability, supporting reliable operation at elevated temperatures (Yuan, 2022; Hassan et al., 2018). Lateral GaN high electron mobility transistors (HEMTs), leveraging a 2DEG channel along with the high critical electric field, demonstrate high power density and efficiency, making them commercially viable for medium-power and high-frequency applications (1–10 kW) (Chowdhury and Mishra, 2013; Ji et al., 2016). Nevertheless, vertical device architectures are preferred for power electronics due to superior electric field management and more efficient chip area utilization. Among GaN vertical transistors (Ji, 2017; Oka et al., 2015; Jeong et al., 2023; Zhang, 2018; Liu, 2020), the current aperture vertical electron transistor (CAVET) (Chowdhury et al., 2008; Wen, 2024; Chowdhury et al., 2012; Ji et al., 2018a) stands out as the only structure that contains a high-conductivity 2DEG channel for current transport, combining a vertical p–n junction for off-state voltage blocking (Wen et al., 2024a). For low-to-medium voltage applications, where the channel conductivity significantly influences the total R_{on} , GaN CAVETs offer a compelling advantage in power device figure of merit ($BV^2/R_{on,sp}$) due to their inherently high channel conductivity (Wen et al., 2024b). Several studies have explored GaN trench CAVET structures (Ji et al., 2018a; Wen et al., 2024c; Shibata, 2016), in which a regrown AlGaIn/GaN heterostructure is formed on a selective-area p-GaN current-blocking layer (CBL) by trench etching.

Despite successful demonstrations, many fundamental properties of GaN CAVETs, including their temperature-dependent behavior, remain insufficiently explored. Prior studies have primarily focused on device performance from RT up to moderately elevated temperatures (<473 K) (Shibata, 2016; Döring, 2024). In our previous work, we

reported *in situ* DC and switching characterization of GaN trench CAVETs operating at temperatures up to 573 K (Wen et al., 2024c). Building on that foundation, this work further extends the temperature characterization range by presenting the first *in situ* electrical characterization of GaN trench CAVETs from 296 K down to cryogenic temperatures as low as 10 K. Additionally, for the first time, the of CAVETs are subjected to 1-minute thermal shock tests at 773 K and 1073 K to *ex-situ* access thermal survivability under high-temperature conditions. The device retained proper transistor functionality after exposure to 773 K but exhibited pinch-off failure at 1073 K. Evaluating GaN CAVETs across this broad temperature range enables an assessment of their suitability for power electronics intended for harsh and thermally demanding environments. This work provides preliminary insights into the potential of GaN CAVETs to serve as robust candidates for future power electronics applications operating under extreme temperature conditions.

2 Materials and methods

The GaN trench CAVET was fabricated on a bulk GaN substrate, as illustrated in Figure 1. We implemented an etch-then-regrowth approach to form a regrown AlGaIn/GaN heterostructure by metal-organic chemical vapor deposition (MOCVD), with the channel extending along slanted trench sidewalls. The 2DEG conductivity is strongly influenced by the etch-then-regrowth process, presenting two key challenges (Ji et al., 2018b). First, Mg inevitably out-diffuses from p-GaN CBL into the channel during high-temperature MOCVD regrowth, which can significantly compensate the 2DEG charge and degrade output current (Xing et al., 2003). To address this, we insert a thin low-temperature GaN (LT-GaN) layer, named the Mg stop layer (MSL), between the p-GaN CBL and the regrown channel, effectively suppressing Mg redistribution (Wen et al., 2023). Second, the etch-then-regrowth process can easily introduce interface issues (Fu et al., 2019; Fu et al., 2021), particularly Si impurity accumulation at the regrowth interface (Fu et al., 2021). The unmodulated impurities can increase off-state leakage through a parasitic conduction path (Li et al., 2018). To mitigate interfacial Si contamination, we perform a chemical cleaning process using ultraviolet-ozone (UVO) and hydrofluoric (HF) acid treatment before MOCVD regrowth (Noshin et al., 2022).

The fabrication process of the trench CAVET is summarized in Figure 2. The structure was grown by MOCVD on a ~ 400 - μm GaN substrate with n + doping about $1 \times 10^{18} \text{ cm}^{-3}$, followed by a 3- μm n-GaN drift layer with a Si doping density of $\sim 6 \times 10^{16} \text{ cm}^{-3}$. A 300-nm p-GaN layer was grown as the CBL with Mg doping density of $\sim 1 \times 10^{19} \text{ cm}^{-3}$. On top of the p-GaN, another 100 nm of LT-GaN was grown at 1023 K to prevent Mg out-diffusion from p-GaN into the subsequent regrown layers. Trenches were etched to ~ 500 nm deep using inductively coupled plasma reactive ion etching (ICP-RIE) to create the aperture region. After tetramethylammonium hydroxide (TMAH) wet etching to smooth the trench sidewalls (Kodama et al., 2008), UVO cleaning and HF treatment were performed immediately before the regrowth to reduce the Si concentration accumulating at the regrowth interface. The MOCVD regrowth contained 140 nm unintentionally doped (UID) GaN and 30 nm $\text{Al}_{0.3}\text{Ga}_{0.7}\text{N}$. Then, after device isolation

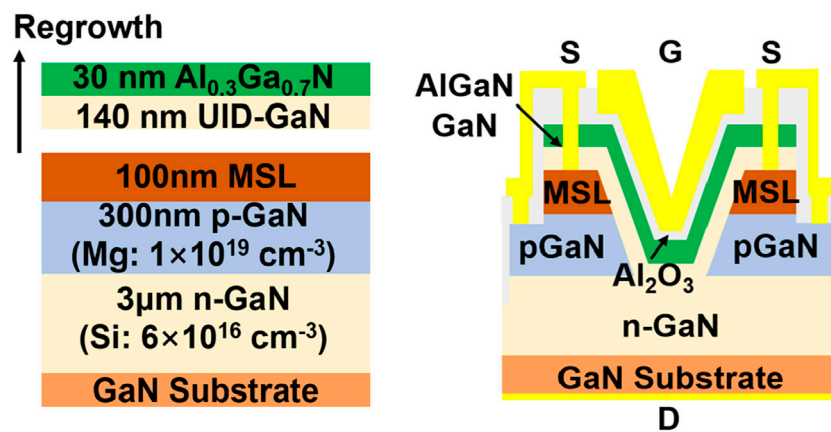


FIGURE 1
Schematic of trench CAVET fabricated on a GaN substrate.

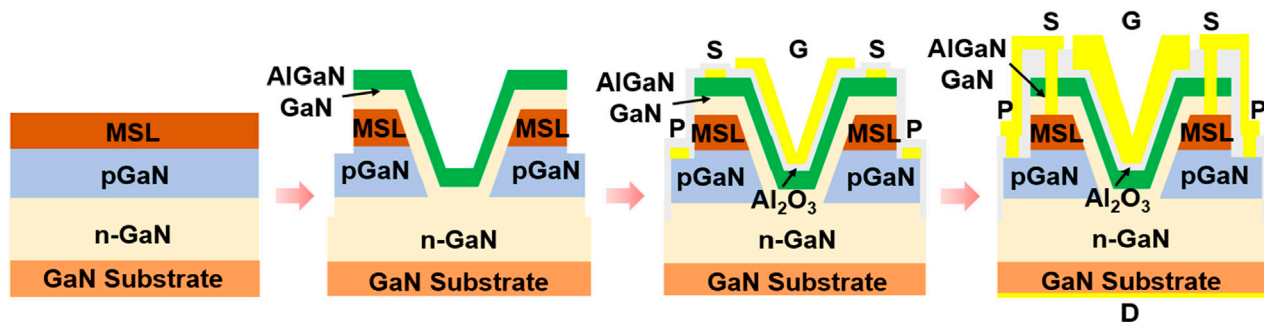


FIGURE 2
Schematic process flow of the GaN trench CAVET.

by dry etching, the p-GaN CBL was activated by rapid thermal annealing (RTA). The Ti/Al/Ni/Au metal stack was deposited on AlGaIn with lift-off, followed by RTA post-annealing at 1073 K to form an ohmic source contact. Ni/Au p-GaN contact was deposited on the exposed p-GaN surface. A 15 nm layer of Al_2O_3 was deposited by plasma atomic layer deposition (ALD) as a gate dielectric with post-deposition annealing (PDA) in N_2 . Ni/Au formed the gate contact covering the trench-shaped channel. A 170 nm SiO_2 was deposited as a passivation layer using plasma enhanced chemical vapor deposition (PECVD), followed by via opening. Ti/Au formed the gate and source pads and connected the p-GaN contact to the source. A Ti/Au drain electrode was placed on the backside.

3 Experimental results

3.1 On the cryogenic characterization of GaN trench CAVET

CAVETs are unique among vertical GaN devices in that they employ a polarization-induced 2DEG channel, which enables a temperature-insensitive 2DEG channel without relying on

impurity doping. This intrinsic conduction mechanism makes them highly attractive for cryogenic power electronics. Notably, in trench CAVETs, the AlGaIn/GaN heterostructure that forms the 2DEG channel is *ex situ* regrown on a trench-shaped surface, with partial overlap on the p-GaN CBL. This geometry introduces challenges in the MOCVD regrowth process, particularly in achieving high-quality heterointerfaces and maintaining excellent 2DEG conductivity. Given that, we focused on characterizing the trench CAVET across a wide temperature range. The fabricated wafer was measured using a cryogenic probe station equipped with liquid helium for temperature control, enabling electrical characterization down to approximately 10 K. We extracted the 2DEG charge density through capacitance–voltage (C–V) measurements at 1 MHz, as shown in Figure 3b. The C–V measurements were performed on a circular Schottky diode fabricated on the CAVET wafer (Figure 3a). The 2DEG charge density (n_s) determined from C–V was $7.36 \times 10^{12} \text{ cm}^{-2}$ at 296 K and $6.83 \times 10^{12} \text{ cm}^{-2}$ at 10 K, indicating an almost negligible variation with temperature. The measured capacitance in the accumulation region is nearly flat, which shows the capacitance of the AlGaIn barrier layer (C_{AlGaIn}). As the voltage bias becomes more negative, electrons in the 2DEG channel at the AlGaIn/GaN heterojunction

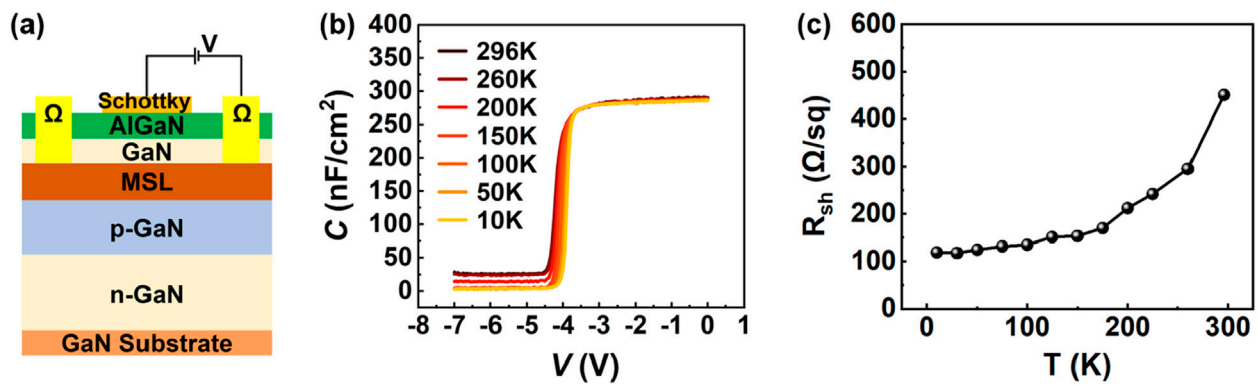


FIGURE 3 C-V characteristics and TLM of the regrown AlGaIn/GaN heterostructure. (a) Circular Schottky diode for C-V measurements. (b) The C-V plot at different temperatures from 296 K to 10 K. (c) Temperature-dependent R_{sh} from TLM.

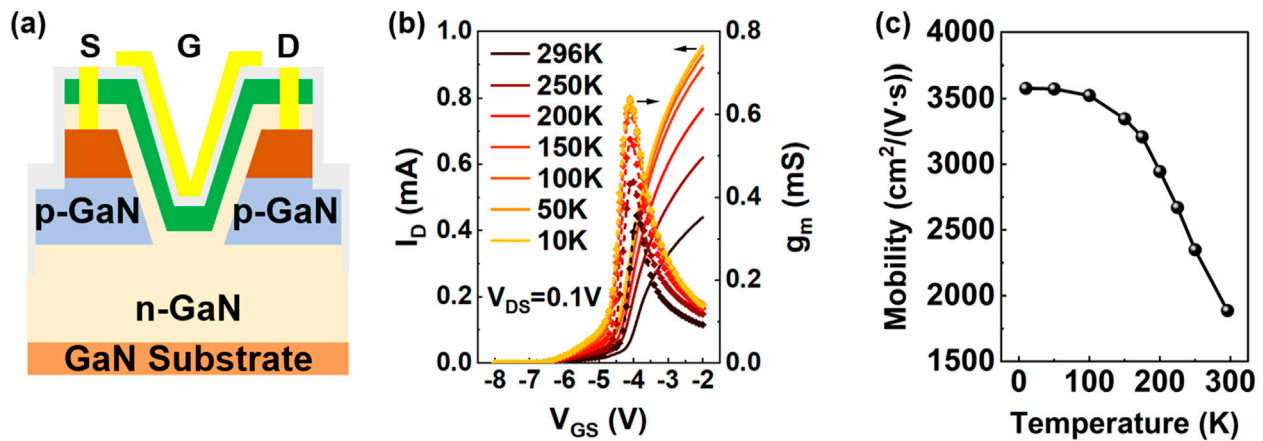


FIGURE 4 (a) Lateral trench structure on the GaN substrate, involving identical epitaxial layers to those of the CAVET. (b) Temperature-dependent I_D - V_{GS} with corresponding transconductance. (c) Extracted field-effect mobility as a function of temperature.

begin to gradually deplete under gate control, resulting in an increasing depletion width in the GaN layer. The total measured capacitance consists of the C_{AlGaIn} and the GaN depletion capacitance ($C_{depletion}$) in series, as shown in Equation (1):

$$\frac{1}{C} = \frac{1}{C_{AlGaIn}} + \frac{1}{C_{depletion}} \quad (1)$$

In the transition region where the 2DEG is almost depleted, the C-V curve slope becomes steeper as the temperature decreases from 296 K due to stronger electron confinement at the AlGaIn/GaN interface at cryogenic temperatures (Liang et al., 2022; Nicollian and Brews, 2002; Miczek et al., 2008). When the gate bias falls below the threshold voltage, the 2DEG becomes fully depleted, and the C-V curve reaches a lower plateau. This region corresponds to the depletion capacitance of the UID-GaN channel and the MSL beneath the heterojunction. At cryogenic temperatures, the ionization rate of residual donors in GaN decreases, resulting in a wider depletion region and thus a lower capacitance.

The temperature dependence of the channel sheet resistance (R_{sh}) was extracted from the transfer length method (TLM), as presented in Figure 3c. R_{sh} decreased from 450.5 Ω/sq at 296 K to 118.9 Ω/sq at 10 K, which was primarily attributed to the improvement in 2DEG carrier mobility due to the reduction of scattering effects at lower temperatures (Zanato et al., 2004).

The transconductance ($g_m = \partial I_D / \partial V_{GS}$) extracted from I_D - V_{GS} at $V_{DS} = 0.1$ V across different temperatures is shown in Figure 4. Measurements were conducted on a lateral trench channel structure with identical epitaxial layers to those of the CAVET, as illustrated in Figure 4a. The field-effect mobility (μ_{FE}) of the regrown AlGaIn/GaN channel was calculated using (Equation 2) the following equation (Wei et al., 2015):

$$\mu_{FE} = \frac{L_G \cdot g_m}{W_G \cdot C_g \cdot V_{DS}} \quad (2)$$

where L_G/W_G is the gate length-to-width ratio, and C_g is the capacitance between the gate and channel. The extracted mobility

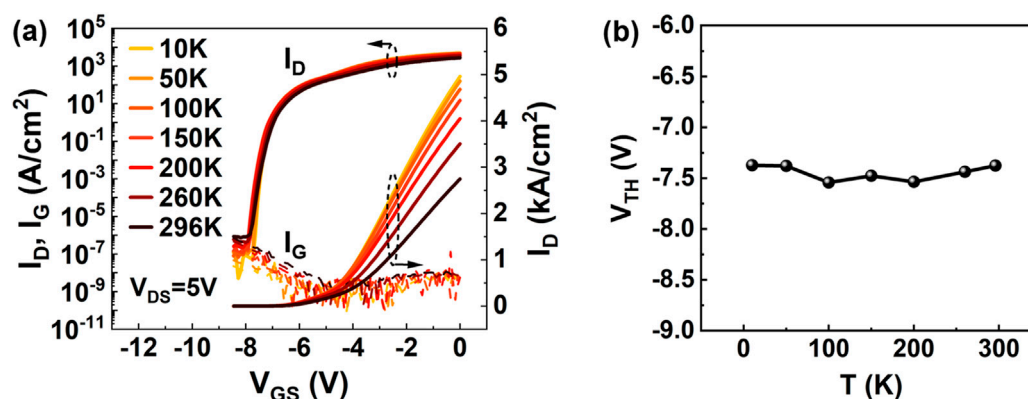


FIGURE 5 (a) Temperature-dependent I_D - V_{GS} transfer characteristics of the fabricated trench CAVET with as-grown MSL. (b) The extracted V_{TH} across temperatures from 296 K to 10 K.

from the peak g_m , as shown in Figure 4c, increased from 1886 $\text{cm}^2/(\text{V}\cdot\text{s})$ at 296 K to 3577 $\text{cm}^2/(\text{V}\cdot\text{s})$ at 10 K. In the relatively high-temperature range, polar optical (LO) phonon scattering is the dominant mobility-limiting mechanism for the 2DEG (Zanato et al., 2004; Kaasbjerg et al., 2013). As the temperature decreases from 296 K, the mobility limited by LO phonon scattering increases exponentially (Lisesivdin et al., 2010; Karmakar et al., 2023). Below approximately 100 K, the mobility gradually saturates and exhibits weak temperature dependence, consistent with prior reports (Zanato et al., 2004; Gökden, 2003). In this deep cryogenic regime, phonon-related scattering mechanisms are largely suppressed, and the remaining mobility limitation can be primarily attributed to interface roughness and dislocation scattering (Nedwell, 1999).

Figure 5a illustrates the temperature-dependent I_D - V_{GS} transfer characteristics of the trench CAVET in both logarithmic and linear scales. At lower temperatures, the off-state leakage current decreased due to the suppression of thermally generated carriers within the material (Kizilyalli and Aktas, 2015; Chen et al., 2024). Coupled with the increased on-state current at lower temperatures, the I_{on}/I_{off} ratio improved significantly from 3×10^9 at 296 K to 9×10^{10} at 10 K. The off-state leakage current decreased at lower temperatures, consistent with the reduced capacitance values observed in the lower plateau of the C-V curves in Figure 3b at cryogenic temperatures. The extracted subthreshold slope (SS) improved from 98.32 mV/dec at 296 K to 51.31 mV/dec at 10 K. Figure 5b shows the threshold voltage (V_{TH}) defined at 1 $\mu\text{A}/\text{mm}$. No significant variation in V_{TH} was observed across the temperature range, showing an average value of -7.45 V and a standard deviation of 0.069 V. The 2DEG at the AlGaIn/GaN interface is primarily induced by polarization fields rather than doping, and the degenerate electron population in the 2DEG channel is nearly temperature-independent (Smorchkova et al., 1999). Because the n_s strongly influences V_{TH} (Huque et al., 2009), the observed V_{TH} stability of CAVET is consistent with the fact that n_s remained almost constant across temperatures.

Figures 6a,b compare the I_D - V_{DS} output characteristics of the fabricated trench CAVET measured at 296 K and 10 K with the gate voltage swept from 0 V in steps of -1 V. At 296 K, the specific on-

resistance ($R_{on,sp}$) was 1.02 $\text{m}\Omega \text{ cm}^2$, and the maximum drain current density ($I_{D,max}$) at $V_{DS} = 10$ V and $V_{GS} = 0$ V was 3.66 kA/cm^2 . At 10 K, $R_{on,sp}$ decreased to 0.586 $\text{m}\Omega \text{ cm}^2$, and $I_{D,max}$ increased to 5.82 kA/cm^2 . The normalized on-resistance at each temperature, defined as $R_{on,sp}(T)/R_{on,sp}(296 \text{ K})$, is plotted in Figure 6c alongside data from a lateral trench HEMT (Figure 4a). As the temperature decreased from 296 K to 10 K, both devices exhibited a similar decreasing trend in $R_{on,sp}$, which continuously decreased as the temperature dropped from 296 K and gradually saturated below 100 K. At 10 K, the normalized $R_{on,sp}$ of the lateral trench HEMT was 0.541, while that of the vertical CAVET was slightly higher at 0.576. The slightly less pronounced reduction in $R_{on,sp}$ for the vertical CAVET, compared to the lateral counterpart, indicates additional contributions from drift resistance (R_{drift}) and substrate resistance (R_{sub}). The $R_{on,sp}$ in a CAVET primarily comprises the channel resistance (R_{ch}), R_{drift} , and R_{sub} (Chowdhury, 2019). In the fabricated trench CAVET with a 3- μm drift layer, R_{ch} accounted for approximately 70% of $R_{on,sp}$ at RT, and its temperature dependence was primarily governed by channel mobility, as the 2DEG charge density remained nearly constant. At cryogenic temperatures, the bulk electron mobility is primarily constrained by ionized impurity scattering, while the 2DEG channel experiences a more pronounced mobility enhancement (Lisesivdin et al., 2010; Shur et al., 1996).

3.2 On the high-temperature survivability of GaN trench CAVET

Thermal shock testing serves as an effective preliminary method to assess the resilience of devices under extreme temperature fluctuations and rapid thermal cycling. As an initial effort to explore the impact of elevated temperatures (≥ 773 K) on the fabricated GaN trench CAVETs, two samples were subjected to rapid thermal annealing (RTA) in a nitrogen ambient atmosphere for 1 minute at 773 K (500 $^{\circ}\text{C}$) and 1073 K (800 $^{\circ}\text{C}$), respectively.

One-minute 773 K thermal shock tests were performed three times on the trench CAVET. The post-annealing I-V characteristics, including both transfer and output curves, are presented in Figure 7

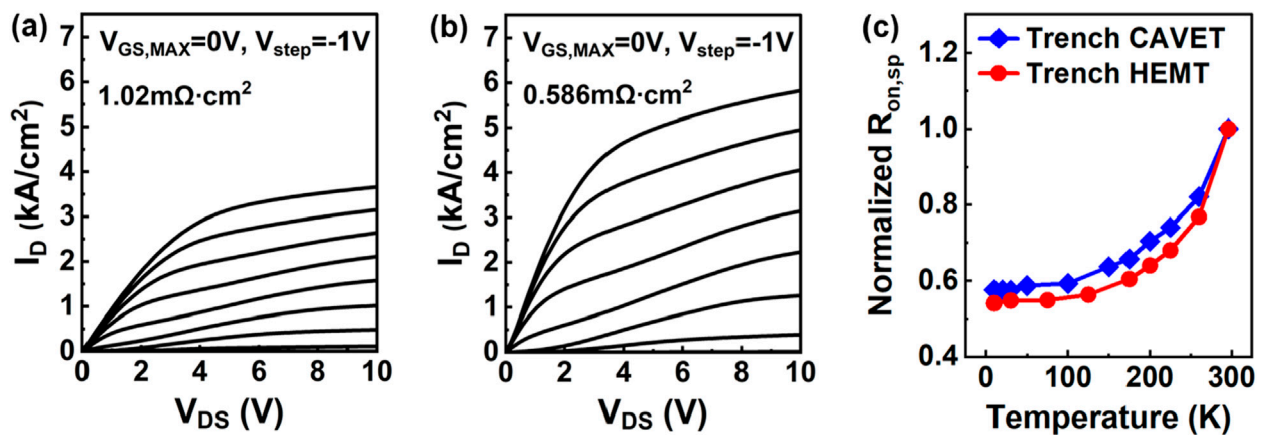


FIGURE 6 I_D - V_{DS} output characteristics of the fabricated trench CAVET at (a) 296 K and (b) 10 K. (c) Temperature-dependent normalized $R_{on,sp}$ of the fabricated trench CAVET and trench HEMT.

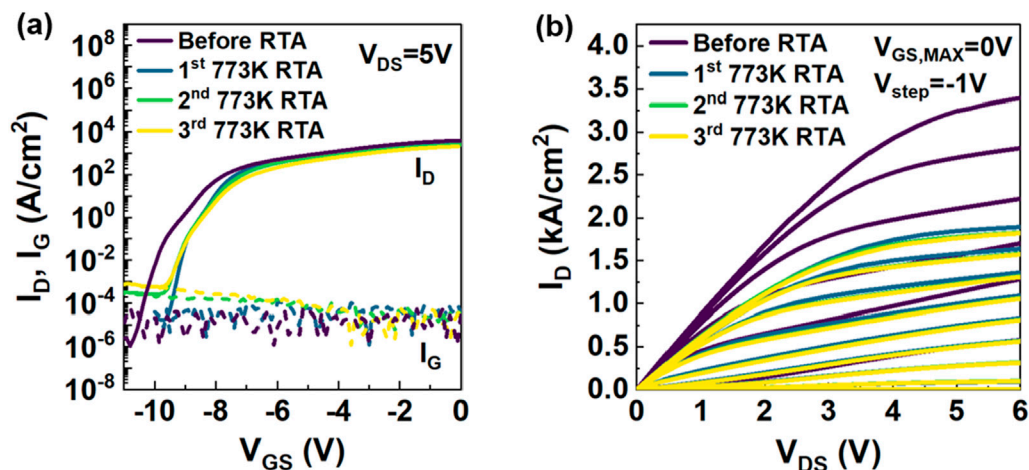


FIGURE 7 (a) I_D - V_{GS} transfer characteristics and (b) I_D - V_{DS} output characteristics trench CAVET after repeated 773 K thermal shock testing.

to explore the impact of repeated thermal shock tests on device transport properties. The devices retained typical FET behavior throughout the tests. The transfer characteristics in Figure 7a exhibit a positive V_{TH} shift from -9.9 V before the thermal shock to -9.2 V after the first run, with no further obvious change observed after the second and third runs. The observed rightward shift in V_{TH} after thermal shock tests may be attributed to annealing effects in the ALD-grown Al_2O_3 gate dielectric layer during thermal shock tests. Specifically, such annealing may reduce the trap state density either at the oxide/AlGaN interface or within the oxide layer, resulting in a positive shift in V_{TH} (Nakazawa et al., 2019; Gupta et al., 2019). Additionally, gradual gate degradation was observed after the repeated thermal shock tests, as reflected by an increase in off-state gate leakage current from 10^{-5} A/cm^2 to 10^{-3} A/cm^2 after three cycles. The output curves in Figure 7b show that the $R_{on,sp}$ increased to $1.58 \text{ m}\Omega \text{ cm}^2$ after the first test and $1.60 \text{ m}\Omega \text{ cm}^2$ after the third thermal test. The output

current density at $V_{GS} = 0 \text{ V}$ and $V_{DS} = 6 \text{ V}$ decreased to 1892 A/cm^2 after the first thermal test and 1819 A/cm^2 after the third run.

After exposure to 1073 K, the devices exhibited more severe performance degradation. As shown in Figure 8a, proper pinch-off behavior was no longer observed due to a substantial increase in gate leakage current. Extracted from the output characteristics in Figure 8b, the $R_{on,sp}$ increased significantly to $24.2 \text{ m}\Omega \text{ cm}^2$, and the output current density dropped to 170 A/cm^2 . Compared to the 773 K treatment, visible metal degradation and peel-off were observed at 1073 K, as shown in Figure 9, along with a distinct color change in the gate metal, indicating possible metal alloying contributing to the increased leakage.

These results represent the first demonstration of the thermal survivability of GaN trench CAVETs under high-temperature shocks and provide a foundation for future studies exploring their reliability under combined high-temperature and bias-stress

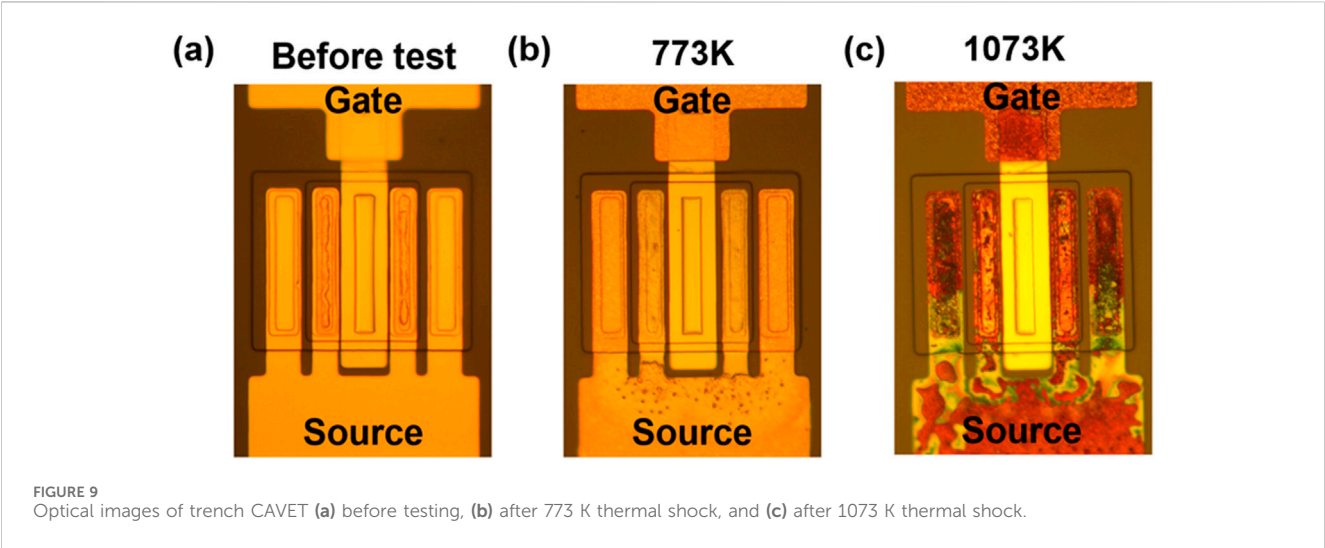
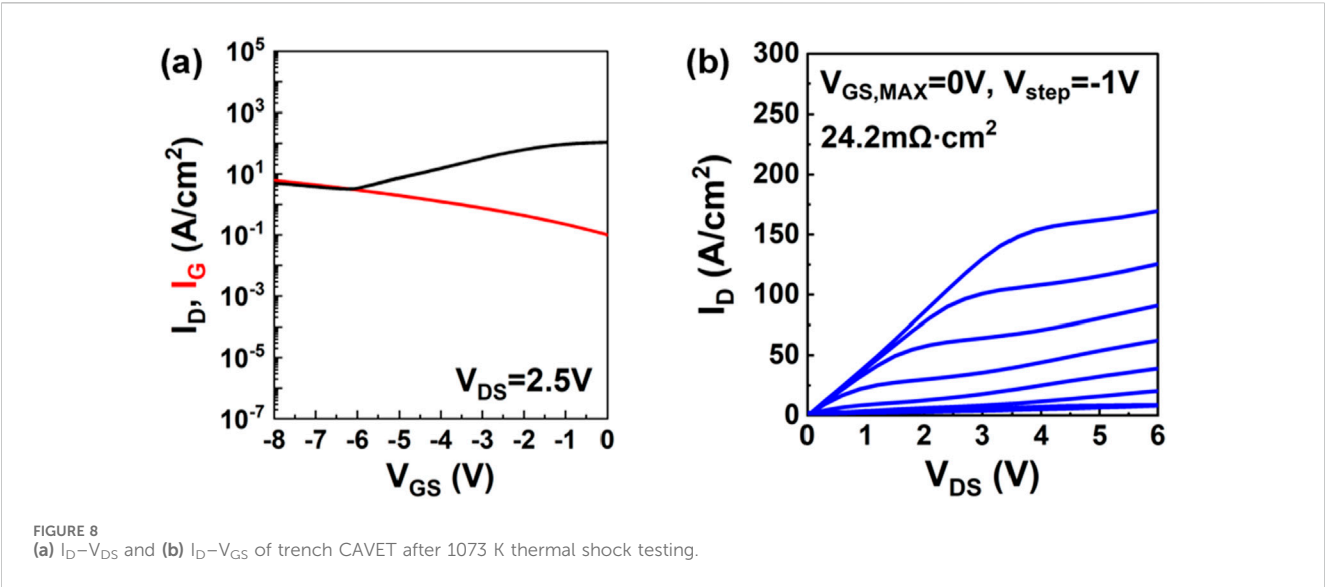


TABLE 1 A summary of the GaN trench CAVET performance in this work and in Wen et al. (2024c) at different temperatures, including both *in situ* measurements (i.e., devices measured at the specified temperature) and *ex-situ* measurements (i.e., devices measured after undergoing 1-min thermal shock testing). *Note that the minimum current measurement resolution of the source measurement unit (SMU) used for gate current measurement in *ex-situ* characterization is approximately 10^{-5} A/cm². Currents below this threshold may not be accurately detected.

Measurement conditions	Temperature	$R_{on,sp}$	I_D at $V_{GS} = 0$ V and $V_{DS} = 6$ V	Off-state I_G
<i>In situ</i>	10 K	0.59 mΩ cm ²	5191 A/cm ²	$\sim 2 \times 10^{-8}$ A/cm ²
<i>In situ</i>	296 K (RT)	1.0 mΩ cm ²	3314 A/cm ²	$\sim 4 \times 10^{-7}$ A/cm ²
<i>In situ</i>	573 K (Wen et al., 2024c)	2.3 mΩ cm ²	1760 A/cm ²	$\sim 4 \times 10^{-4}$ A/cm ²
<i>Ex-situ</i>	773 K	1.6 mΩ cm ²	1892 A/cm ²	$\sim 2 \times 10^{-5}$ A/cm ² *
<i>Ex-situ</i>	1073 K	24 mΩ cm ²	170 A/cm ²	~ 10 A/cm ²

conditions. The electrical characteristics of the GaN trench CAVETs from this work and our prior study (Wen et al., 2024c) are summarized in Table 1, including both *in situ* measurements (i.e., devices measured at the specified temperature) and *ex-situ* measurements (i.e., devices measured after undergoing 1-min thermal shock testing).

4 Conclusion

In summary, this study presents the first investigation of the device characterization of GaN trench CAVETs over a broad temperature range. For cryogenic characterization, the 2DEG charge density was $7.36 \times 10^{12} \text{ cm}^{-2}$ obtained from the regrown AlGaIn/GaN channel with a slight 7% reduction from 296 K to 10 K. The V_{TH} of CAVET exhibited excellent thermal stability. The field-effect mobility increased from $1886 \text{ cm}^2/(\text{V}\cdot\text{s})$ at 296 K to $3577 \text{ cm}^2/(\text{V}\cdot\text{s})$ at 10 K, contributing to improved device conductivity. Specifically, $I_{\text{D,max}}$ increased from 3.66 kA/cm^2 to 5.82 kA/cm^2 , and $R_{\text{on,sp}}$ decreased from $1.06 \text{ m}\Omega \text{ cm}^2$ to $0.586 \text{ m}\Omega \text{ cm}^2$. Additionally, we performed 1-min thermal shock testing to preliminarily evaluate the survivability of trench CAVETs under high-temperature conditions. The device maintained normal functionality after 773 K testing with approximately 50% current reduction. Exposure to 1073 K caused more significant degradation in both $R_{\text{on,sp}}$ and gate control. These findings demonstrate the potential of GaN trench CAVETs for applications in extreme temperature environments.

Data availability statement

The raw data supporting the conclusions of this article will be made available by the authors, without undue reservation.

Author contributions

XW: Conceptualization, Data curation, Investigation, Methodology, Visualization, Writing – original draft, Writing – review and editing. KL: Conceptualization, Investigation, Methodology, Writing – original draft, Writing – review and editing. HK: Conceptualization, Investigation, Methodology, Resources, Validation, Visualization, Writing – original draft, Writing – review and editing. MN: Conceptualization, Investigation, Methodology, Writing – original draft, Writing – review and editing. CM: Conceptualization, Investigation, Methodology, Writing – original draft, Writing – review and editing. SC: Conceptualization, Funding

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- ## Conflict of interest
- The authors declare that the research was conducted in the absence of any commercial or financial relationships that could be construed as a potential conflict of interest.
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