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Super-Twisting Sliding Mode Control for Single-Stage Isolated Bidirectional V2G Chargers

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How To Cite: Luo, W.; Jin, Z.; Zhu, Z.; et al. Super-Twisting Sliding Mode Control for Single-Stage Isolated Bidirectional V2G Chargers. *Power Electronics Research and Applications Transactions* **2026**. <https://doi.org/10.53941/perat.2026.100009>

Received: 14 September 2026

Revised: 21 September 2026

Accepted: 22 September 2026

Published: 29 September 2026

Abstract: Vehicle-to-grid (V2G) technology plays a critical role in modern power systems by utilizing electric vehicles as distributed energy storage to smooth power fluctuations and achieve peak-shaving and valley-filling. As a vital power electronics interface, single-stage isolated bidirectional AC/DC converters offer significant advantages in high power density and electrical isolation. However, conventional control methods often suffer from sluggish dynamic response and insufficient robustness when subjected to parameter uncertainties and external disturbances. To address these issues, this paper proposes an improved super-twisting sliding mode control (STA) strategy integrated with a low-pass filter. The proposed strategy effectively suppresses inherent chattering while retaining the strong robustness characteristic of traditional sliding mode control. Simulation results demonstrate that the control strategy maintains superior dynamic performance and robustness under external noise interference and variations in battery internal resistance, offering a highly effective solution for high-performance power conversion in V2G systems.

Keywords: vehicle-to-grid; single-stage isolated bidirectional AC/DC converter; super-twisting sliding mode control; chattering suppression

1. Introduction

Vehicle-to-grid (V2G) technology integrates electric vehicles (EVs) into the power grid as distributed energy storage units. By storing excess power during off-peak hours and discharging energy back during peak demand, V2G achieves effective peak-shaving and valley-filling. This capability significantly mitigates grid load fluctuations and dampens the operational impact caused by the intermittent nature of renewable energy integration [1,2]. With the rapid increase in EV market penetration [3], developing robust V2G applications has become crucial for building safe, flexible, and sustainable modern power systems. Figure 1 shows the schematic diagram of a typical V2G system.

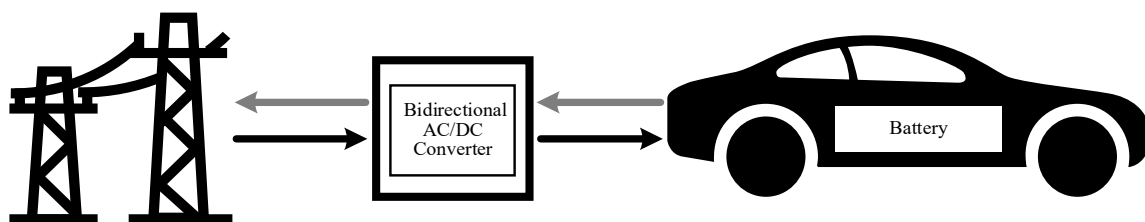


Figure 1. Schematic diagram of the V2G system framework.



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As the key hardware interface in V2G systems, the bidirectional AC/DC converter governs power exchange between the grid and EV batteries. Topology selection generally balances system efficiency and safety: while non-isolated or two-stage architectures (Figure 2a) are common, single-stage isolated bidirectional AC/DC converters (Figure 2b) have gained widespread attention in high-power applications due to their inherent galvanic isolation, high power density, and reduced conversion stages [4,5]. Consequently, investigating advanced control strategies for this single-stage isolated converter topology holds important theoretical and practical value for enabling high-performance V2G deployment.

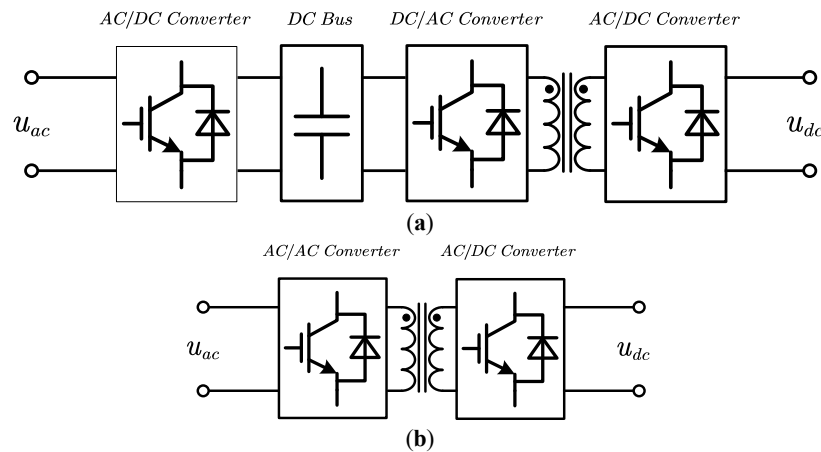


Figure 2. Bidirectional AC/DC converter: (a) two-stage isolated AC/DC converter, (b) single-stage isolated AC/DC converter.

Single-stage isolated bidirectional AC/DC converters, derived from the classical dual active bridge (DAB) architecture, are generally classified into three topological categories: quasi-single-stage, single-stage half-bridge, and single-stage full-bridge topologies [6]. The quasi-single-stage bidirectional AC/DC converter consists of a front-stage bidirectional AC/DC converter and a rear-stage isolated bidirectional DC/DC converter. Similar to conventional two-stage isolated AC/DC converters, its primary distinction lies in replacing the bulky DC-link electrolytic capacitor with a smaller small-capacity capacitor [7]. Consequently, the intermediate DC-bus voltage exhibits a unipolar double-frequency pulsating waveform rather than a steady DC voltage.

Compared with the quasi-single-stage configuration, the single-stage half-bridge bidirectional AC/DC converter topology eliminates the intermediate DC link completely and reduces the semiconductor switch count. However, it requires a high-frequency center-tapped transformer, which increases transformer design complexity and halves primary winding utilization, thereby limiting the improvement of power density [8]. To overcome these limitations, the single-stage full-bridge bidirectional AC/DC converter topology offers a simpler transformer design, reduced voltage stress on switching devices, and superior system reliability [9]. Furthermore, it can be extended to three-phase operation simply by incorporating two additional bidirectional switches [10], albeit at the cost of increased control complexity due to the higher device count. Given its balanced trade-off between power density, reliability, and modularity, the single-stage full-bridge topology is selected as the primary research object of this paper.

Regarding control strategies for isolated single-stage bidirectional AC/DC matrix converters, dq current control is widely adopted [11,12]. In single-phase systems, this strategy typically involves phase-shifting the grid-side current by 90° to construct orthogonal signals in the stationary reference frame, enabling dq decoupling. However, this scheme suffers from incomplete decoupling, and incorporating traditional PI controllers within certain loops compromises dynamic performance and robustness. Alternatively, a Smith predictor has been applied for DC-side current control to bypass the single-phase decoupling issue and achieve good dynamic response [13]. Nevertheless, this model-dependent approach exhibits limited robustness when system parameter uncertainties occur. Observer-based and passivity-based control methods have also been investigated to enhance the robustness of power converters under challenging load conditions [14].

To overcome these limitations, this paper proposes a super-twisting sliding mode control strategy. Compared with conventional PI control, sliding mode control inherently offers stronger robustness and faster dynamic response, and has been successfully applied to motor control and various power electronic topologies [15–20]. For instance, a terminal sliding mode control was applied to a dual active bridge DC/DC converter to improve system response speed and steady-state accuracy [21,22]. In [23], a sliding-mode PID controller was introduced for DC/DC converters, yielding superior dynamic and static performance over conventional SMC. Furthermore, a

global non-singular fast terminal sliding mode controller was adopted in [24] for an interleaved boost converter, effectively ensuring global stability while mitigating chattering.

2. Operating Principle of the Topology

The topology of the single-stage isolated bidirectional AC/DC converter studied in this paper is illustrated in Figure 3. The converter primarily consists of an LC filter, a single-phase matrix converter, a high-frequency transformer (HFT), and a secondary-side full-bridge converter. Specifically, each bidirectional switch in the matrix converter is configured by two anti-series-connected IGBTs. By configuring appropriate switching sequences, these bidirectional switches achieve four-quadrant operation with bidirectional current conduction and bidirectional voltage blocking capabilities. Compared with conventional two-stage bidirectional AC/DC architectures, the proposed single-stage topology eliminates the front-stage rectifier and bulky DC-link electrolytic capacitors, thereby significantly improving system power density and efficiency.

In Figure 3, L_f and C_f represent the grid-side filter inductance and capacitance, respectively; L_σ denotes the equivalent leakage inductance of the transformer; C_b is the battery-side filtering capacitor; and R_b represents the internal resistance of the battery.

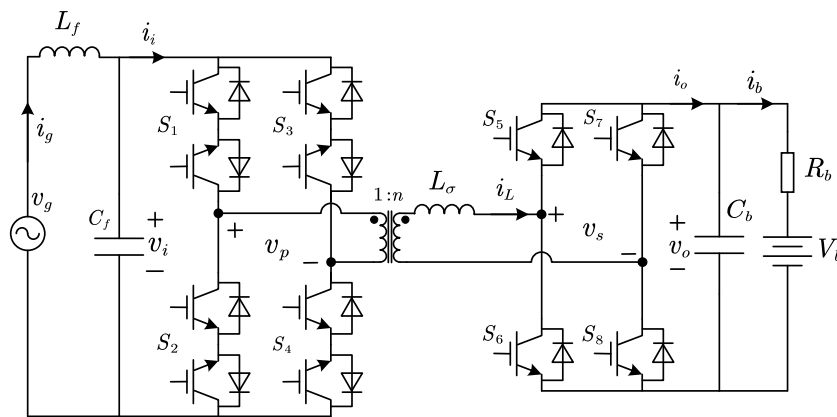


Figure 3. Single-stage isolated bidirectional AC/DC converter.

The operating principle of this converter is similar to that of a Dual Active Bridge (DAB) converter. The single-phase matrix converter and the secondary full-bridge converter modulate the grid-side AC voltage v_g and the DC battery voltage V_b into high-frequency square-wave voltages, v_p and v_s , respectively. These two high-frequency AC voltages are coupled through the high-frequency transformer with a turns ratio of n . By adjusting the phase-shift angle between v_p and v_s across the leakage inductor L_σ , both the magnitude and direction of the power transfer can be precisely regulated.

The operating waveforms of the high-frequency transformer primary/secondary AC voltages, as well as the voltage across and current through the leakage inductor over a single switching cycle, are illustrated in Figure 4.

Neglecting commutation procedures and dead-time constraints, the two bidirectional switches on the same leg of the matrix converter are operated in a complementary manner, while the diagonally opposite switches share identical gating signals, thereby modulating the grid-side AC voltage into a two-level high-frequency square wave. On the DC side, the upper and lower switches of the same inverter bridge arm are driven complementarily, generating a three-level high-frequency square wave with a pulse duty cycle of $d(t)$. To guarantee a sinusoidal grid-side current, the pulse width $d(t)$ is dynamically modulated in proportion to the sinusoidal variation of the grid voltage.

Considering its physical definition, the duty cycle d must be strictly bounded within $[0, 1]$, which is expressed as:

$$d = \frac{|nv_g|}{V_b}. \quad (1)$$

To simplify the mathematical derivation, the normalized phase-shift ratio δ is defined as follows:

$$\delta = \frac{\Delta T}{T_s/4}, -1 \leq \delta \leq 1. \quad (2)$$

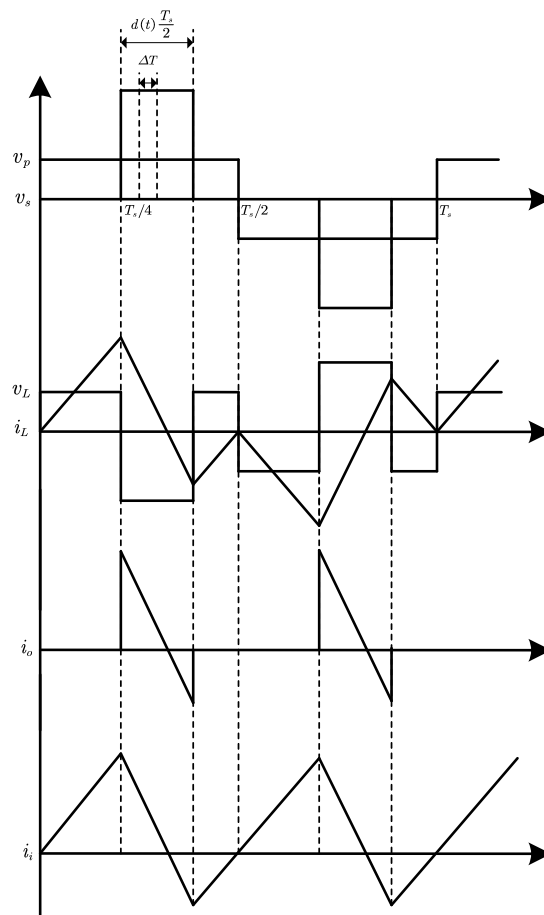


Figure 4. Modulation Scheme of the proposed converter.

Next, the average input current over a single switching cycle T_s is calculated. Since the switching frequency f_s is significantly higher than the cutoff frequency of the grid-side LC filter, the grid-side current i_g can be well approximated by the switching-cycle average of the input current i_{in} , which is expressed as:

$$\begin{aligned}
 i_g(t) &= \langle i_i \rangle_{T_s} = \frac{1}{T_s} \left(\int_0^{T_s/2} i_L(t) dt + \int_{T_s/2}^{T_s} -i_L(t) dt \right) \\
 &= \frac{nV_b \delta T_s}{4L_\sigma} d.
 \end{aligned} \quad (3)$$

As indicated by (1) and (3), neglecting the phase shift introduced by the LC filter, the grid-side current is strictly in phase with the grid voltage, thereby achieving power factor operation. Furthermore, the direction of power flow is uniquely determined by the sign of the normalized phase-shift ratio D : when $\delta > 0$, power flows from the AC side to the DC side; conversely, when $\delta < 0$, power is fed back from the DC side to the AC side.

Similarly, the DC-side battery current i_b over a switching cycle can be derived as follows:

$$i_b(t) = \langle i_o \rangle_{T_s} = \frac{n^2 V_g^2 T_s \delta}{4L_\sigma V_b}. \quad (4)$$

As can be seen from (4), the DC-side current i_b contains a double-frequency AC ripple component and is entirely governed by the normalized phase-shift ratio δ .

3. Design of Controller

To manage the battery charging and discharging processes, precise control of the output current delivered to or from the batteries is essential. As analyzed in the previous section, the output current of the bidirectional charger is directly regulated by the phase-shift ratio δ , which modulates the switching signals for both the single-phase matrix converter and the full-bridge inverter. This configuration operates as a single-input single-output control

system, where δ serves as the primary degree of freedom for battery current regulation. Furthermore, the modulation scheme effectively decouples the AC grid input from the DC battery output, allowing the charger output to be modeled as a controlled current source.

To extract the fundamental DC component of the battery current, a notch filter is employed for signal processing. As a specialized band-stop filter, the notch filter provides sharp attenuation at a specific target frequency (i.e., the double-frequency ripple component) while leaving other frequency components virtually unaffected. The transfer function of the notch filter is expressed as follows:

$$G_n(s) = \frac{s^2 + \omega^2}{s^2 + 2\zeta\omega s + \omega^2}, \quad (5)$$

where ω denotes the notch frequency and ζ represents the damping ratio that controls the notch depth and bandwidth. To effectively eliminate the double-frequency ripple component (100 Hz, corresponding to a 50 Hz grid), the parameters are selected as $\omega = 200\pi$ rad/s (i.e., 100 Hz) and $\zeta = 0.55$. The Bode plot of the designed notch filter is illustrated in Figure 5. As observed from the Bode magnitude response, the notch filter provides substantial attenuation at 100 Hz while exerting minimal phase and gain distortion on harmonics at other frequencies. Consequently, the double-frequency AC ripple in the measured current can be effectively removed, yielding a clean DC feedback signal for the controller.

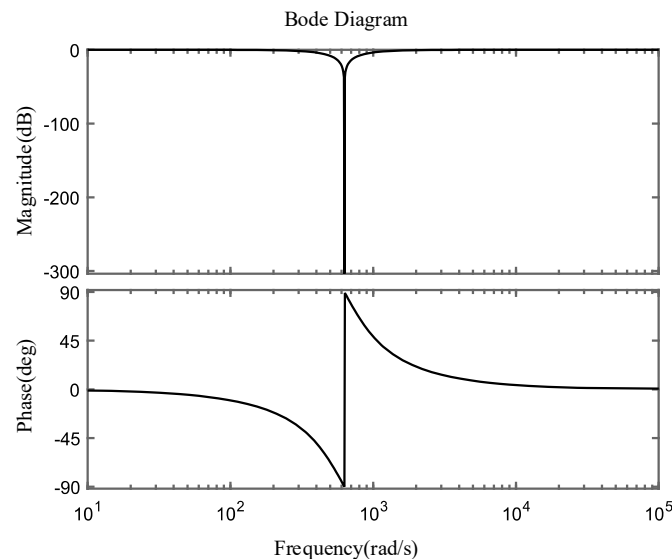


Figure 5. Bode diagram of a notch filter.

Since the battery-side capacitor C_b effectively filters out high-frequency switching harmonics from the input current, the high-frequency components in the converter output current can be neglected. Meanwhile, owing to the notch filter, only the fundamental DC component of the control variables is taken into account while the double-frequency AC ripple is ignored. Consequently, the simplified equivalent model of the converter is established, as illustrated in Figure 6.

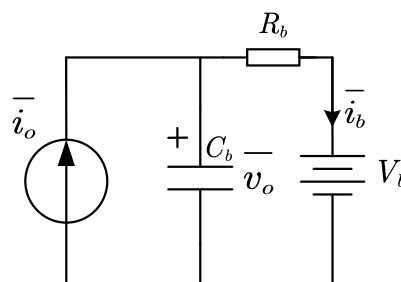


Figure 6. Simplified equivalent model.

In Figure 6, the controlled current source $\bar{i}_o$ is governed by the phase-shift ratio δ , and its mathematical expression is given as follows:

$$\bar{i}_o = \frac{n^2 V_m^2 T_s}{8 L_\sigma V_b} \delta. \quad (6)$$

Based on the simplified equivalent model in Figure 6, the mathematical model of the converter is derived as follows:

$$\bar{i}_b = \bar{i}_o - C_b \frac{d\bar{v}_o}{dt}, \quad (7)$$

$$\bar{v}_o = V_b + \bar{i}_b R_b. \quad (8)$$

By substituting (6) and (8) into (7), the dynamic formulation governing the relationship between the battery-side DC current component $\bar{i}_b$ and the control input δ is obtained as follows:

$$\frac{d\bar{i}_b}{dt} = -\frac{1}{C_b R_b} \bar{i}_b + \frac{K}{C_b R_b} \delta, \quad (9)$$

where
$$K = \frac{n^2 V_m^2 T_s}{8 L_\sigma V_b}.$$

For the STA, the closed-loop performance relies directly on the appropriate construction of the sliding surface, which dictates the trajectory convergence and steady-state precision of system states in the sliding phase. The sliding surface is defined based on the tracking error as follows:

$$s = e. \quad (10)$$

By taking the time derivative of the sliding surface $s(t)$ and setting $\dot{s}(t) = 0$, the equivalent control law δ_{eq} is derived as follows:

$$\delta_{eq} = \frac{1}{K} \bar{i}_b. \quad (11)$$

Incorporating the super-twisting sliding mode control algorithm, the output phase-shift ratio δ is composed of two distinct parts:

$$\delta = \delta_{eq} + \delta_{st}. \quad (12)$$

$$\delta_{st} = \lambda |s|^{\frac{1}{2}} \text{sgn}(s) + \zeta \int_0^t \text{sgn}(s) dt. \quad (13)$$

The super-twisting sliding mode control generates a continuous and smooth control output by shifting the discontinuous sign function of traditional sliding mode into an integral operator. Concurrently, the fractional-power proportional term in the control law guarantees rapid state convergence. However, as the sliding variable s approaches zero, the equivalent gain $\lambda/|s|^{0.5}$ of this proportional term escalates toward infinity, which significantly amplifies high-frequency noise and measurement disturbances in practical circuits. Furthermore, the discrete-time implementation of digital controllers leads to repeated zero-crossings of system states around the sliding surface, causing high-frequency chattering. To tackle these issues, a low-pass filter-based super-twisting sliding mode control strategy is proposed in this paper, effectively mitigating chattering and noise sensitivity while retaining superior dynamic performance.

The transfer function of the introduced low-pass filter is expressed as follows:

$$G_{LPF}(s) = \frac{1}{Ts + 1}, \quad (14)$$

where T denotes the time constant of the first-order low-pass filter.

To prevent introducing extra phase lag caused by directly cascading a low-pass filter into the control loop, the low-pass filter is integrated into the augmented plant model. Specifically, the dynamic mathematical model of the original converter is cascaded with a first-order low-pass filter at the output, constructing an augmented equivalent system. Based on this augmented structure, the super-twisting sliding mode controller is synthesized.

Consequently, the dynamic relationship between the output of the super-twisting controller and the actual phase-shift ratio δ applied to the system is expressed as follows:

$$T \frac{d\delta}{dt} + \delta = u. \quad (15)$$

By substituting (15) into (9), the system model combining the converter dynamics and the low-pass filter is obtained as follows:

$$\frac{d^2 \bar{i}_b}{dt^2} = -\frac{T + C_b R_b}{TC_b R_b} \frac{d\bar{i}_b}{dt} - \frac{1}{TC_b R_b} \bar{i}_b + \frac{K}{TC_b R_b} u. \quad (16)$$

Based on the new system model, the sliding surface is redesigned as follows:

$$s = e + a\dot{e}, \quad (17)$$

where $a > 0$.

By setting the time derivative of the redesigned sliding surface to zero ($\dot{s}(t) = 0$), the equivalent control law for the super-twisting sliding mode controller is derived as follows:

$$u_{eq} = \left(\frac{T}{K} + \frac{(a-T)C_b R_b}{aK} \right) \frac{d\bar{i}_b}{dt} + \frac{1}{K} \bar{i}_b. \quad (18)$$

As observed, the derived equivalent control law explicitly contains the internal resistance of the battery R_b , which is challenging to measure directly in real-time applications. To reduce the controller's sensitivity to internal system parameters and enhance its robustness, the parameter can be set as $a = T$, thereby eliminating the terms dependent on the battery internal resistance R_b .

Consequently, the overall control law of the proposed LPF-STC scheme is synthesized as follows:

$$\begin{cases} u = u_{eq} + u_{st}, \\ u_{eq} = \frac{T}{K} \frac{d\bar{i}_b}{dt} + \frac{1}{K} \bar{i}_b, \\ u_{sw} = \lambda |s|^{\frac{1}{2}} \text{sgn}(s) + \zeta \int_0^t \text{sgn}(s) dt. \end{cases} \quad (19)$$

The overall control block diagram of the proposed LPF-STC scheme for the bidirectional V2G converter is illustrated in Figure 7.

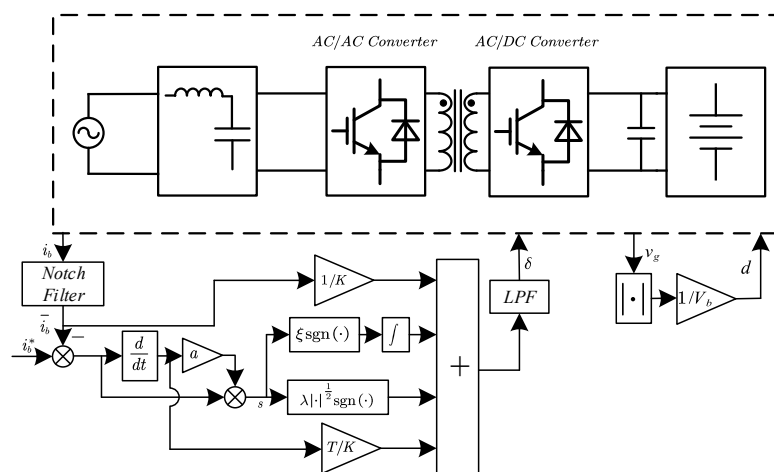


Figure 7. Overall control block diagram.

4. Simulation Results

This section presents the simulation results of the bidirectional converter and evaluates the effectiveness of the proposed control strategy. To demonstrate its superiority, a performance comparison between the proposed

LPF-STA scheme and the conventional PI controller is conducted under various operating conditions. The simulation parameters are shown in Table 1.

Table 1. Simulation Parameters.

Parameter	Value
Grid-side Voltage (RMS)	220 V
Battery Voltage	800 V
Switching Frequency	50 kHz
Transformer Turns Ratio	1
Leakage Inductor	50 μ H
Filter Inductor	1 mH
Filter Capacitor	4 μ F
Battery-side Output Capacitor	940 μ F

To verify the dynamic performance and bidirectional power flow capability of the proposed control strategy, a step-response simulation is conducted under varying current reference commands, as illustrated in Figure 8. The simulation is initialized at $t=0$ s with a current command $i_{bref}=3$ A in the battery charging mode. At $t=0.1$ s, the reference undergoes a step change to -3 A, causing the converter to seamlessly switch into the V2G discharge mode. Subsequently, the discharge current reference is stepped to -1 A at $t=0.2$ s. Finally, at $t=0.3$ s, a reference step back to 2 A is applied to restore the battery charging mode.

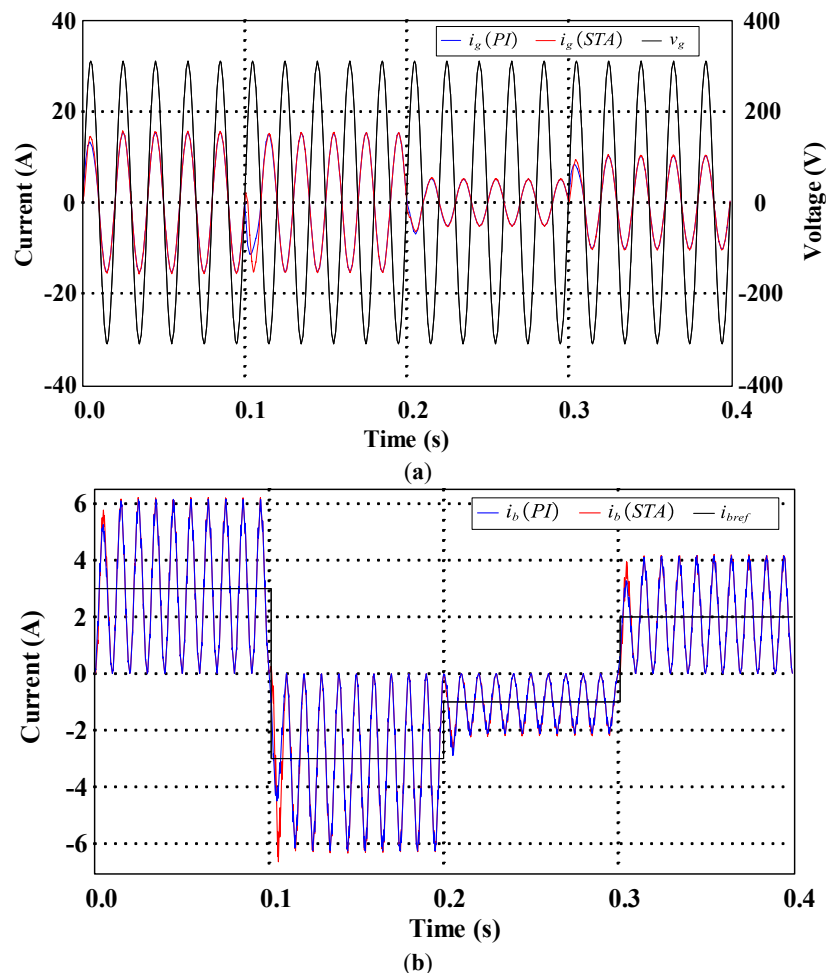


Figure 8. Simulation comparison between the benchmark PI controller and the proposed LPF-STA: (a) Grid-side voltage and current waveforms. (b) Battery-side current tracking response.

To clearly evaluate and compare the dynamic response of both control schemes without the interference of double-frequency AC ripples, the extracted fundamental DC component of the battery current $\bar{i}_b$ is depicted in Figure 9.

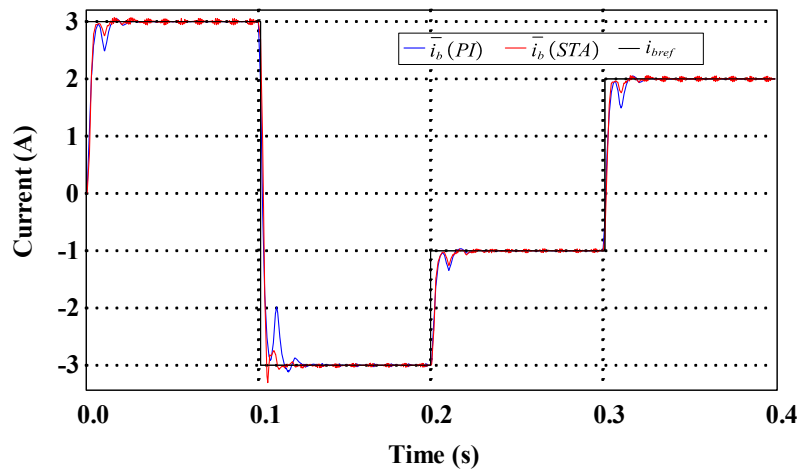


Figure 9. Comparison of battery average current tracking under step reference changes.

The average battery current profiles under both the conventional PI controller and the proposed LPF-STA strategy are presented in Figure 9. It can be observed that while both control methods effectively track the current reference commands, the proposed strategy significantly enhances the dynamic responsiveness of the system. Specifically, when subjected to large reference step variations, the conventional PI controller exhibits noticeable transient oscillations and extended recovery periods. In contrast, the proposed LPF-STA scheme substantially reduces both the rise time and the settling time, achieving superior transient tracking performance without significant overshoot.

To evaluate the parameter robustness of the proposed control scheme against variations in internal battery impedance, simulations were performed under three distinct battery equivalent internal resistances: $R_b = 0.01\Omega$, 0.05Ω , and 0.1Ω . As derived in Section 2, by setting the parameter $a = T$, the proposed LPF-STA control law theoretically eliminates its explicit dependence on R_b . Figure 10 depicts the dynamic response of the battery average current $\bar{i}_b$ under these varying resistance conditions across dynamic step reference changes and power flow reversals.

As observed in Figure 10, the current tracking trajectories for all three R_b values are virtually identical, exhibiting highly consistent dynamic transients, negligible overshoot, and rapid settling times. This confirms that variations in R_b do not degrade the control performance, thereby validating the strong parameter immunity and dynamic robustness of the proposed controller in practical V2G applications.

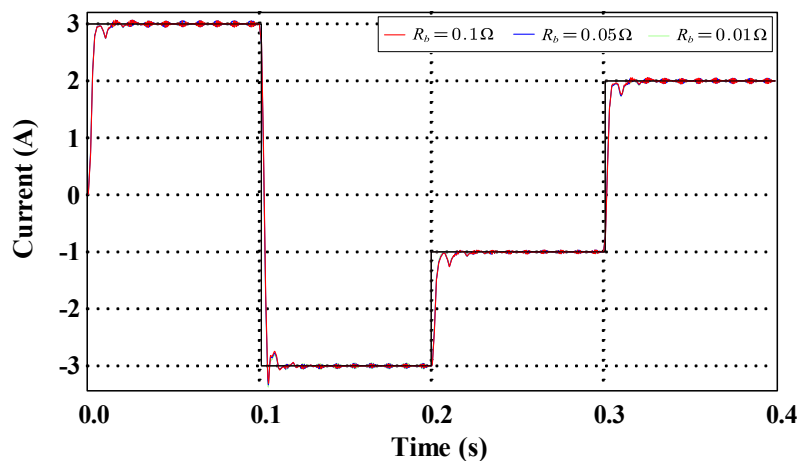


Figure 10. Dynamic response of the battery average current $\bar{i}_b$ under different battery internal resistances.

5. Experimental Results

To further validate the real-time control capability and practical feasibility of the proposed LPF-STA algorithm, a hardware-in-the-loop (HIL) experimental platform was constructed. The power stage of the bidirectional V2G converter, along with the battery model and grid interface, is emulated in real time using a PLECS RT-Box CE target. The proposed control strategy is programmed and executed on a TI TMS320F28379D

dual-core Digital Signal Processor (DSP). Physical analog and digital signals are exchanged between the RT-Box CE and the DSP board through direct PWM and ADC hardware interconnections. The overall structure of the HIL test bench is illustrated in Figure 11.

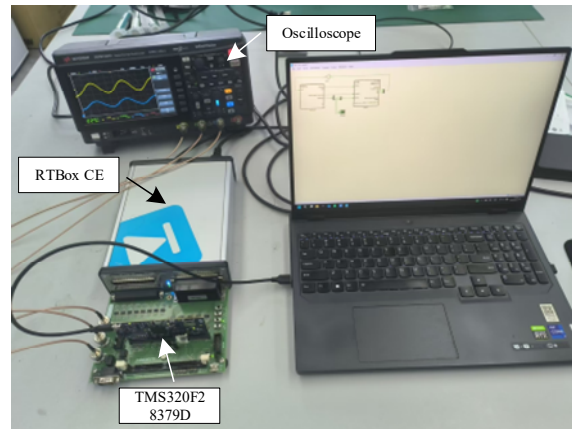


Figure 11. Physical setup of the hardware-in-the-loop (HIL) experimental platform.

Figure 12a,b illustrate the grid-side voltage/current waveforms and the battery current profile, respectively, during a step transition of the current reference from 3 A to −3 A. Conversely, Figure 13a,b show the corresponding transient waveforms when the current reference steps back from −3 A to 3 A. The HIL experimental results validate that the proposed algorithm achieves superior dynamic performance alongside high steady-state tracking precision, thereby demonstrating its practical feasibility for digital controller implementation.

Analyzing the HIL experimental results in Figures 12 and 13, it is clear that the grid current i_g maintains a clean sinusoidal waveform with low harmonic distortion during steady-state operation. In Figure 12a, when operating in the battery charging mode ($i_{bref} = 3$ A), the grid current i_g is strictly in phase with the grid voltage v_g , imposing a near-unity power factor at the AC interface. Conversely, in Figure 13a, under the Vehicle-to-Grid (V2G) discharging mode ($i_{bref} = -3$ A), i_g smoothly shifts by 180° relative to v_g , exhibiting a clean anti-phase operation and enabling seamless reverse active power transfer back to the grid.

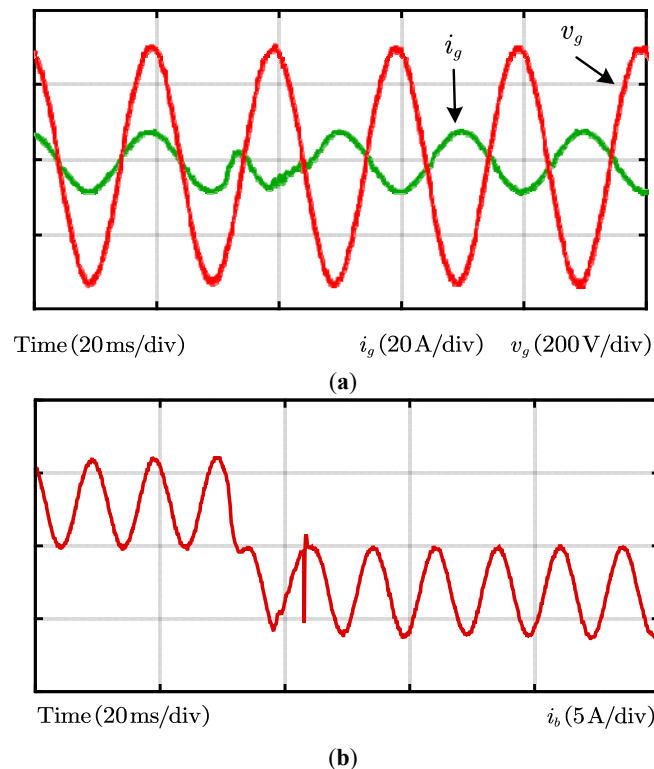


Figure 12. reference step from 3 A to −3 A: (a) Grid-side voltage v_g and current i_g waveforms. (b) Battery current i_b waveform.

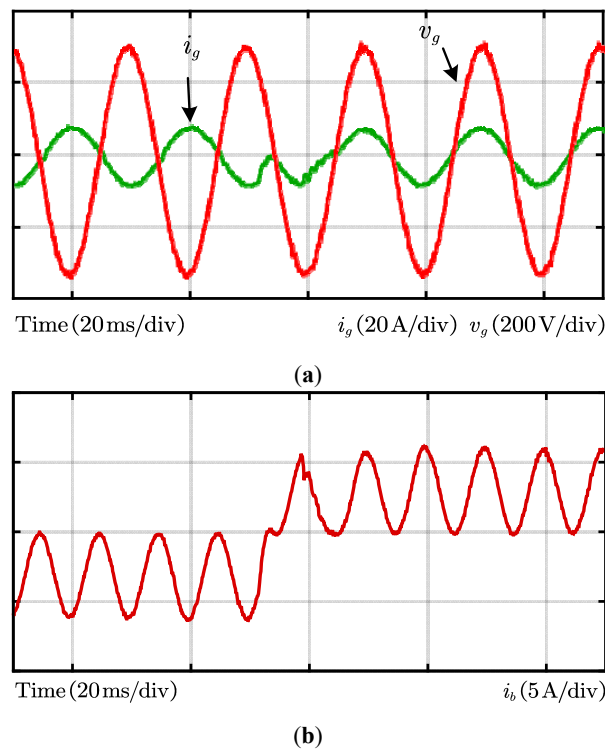


Figure 13. Transient waveforms for a battery-current reference step from -3 A to 3 A: (a) Grid-side voltage v_g and current i_g ; (b) Battery current i_b .

Furthermore, as depicted in Figures 12b and 13b, the proposed LPF-STA scheme handles the bidirectional step reference transitions within less than half a grid cycle, suppressing transient voltage/current spikes effectively. The results confirm that the grid-side power factor remains virtually close to 1 in both operational modes. Therefore, the proposed control topology successfully guarantees high-quality power exchange and robust bidirectional power flow control for practical V2G applications.

6. Conclusions

In this paper, an LPF-STA strategy was proposed for bidirectional V2G converters to address chatter mitigation, dynamic speed, and parameter sensitivity challenges. By incorporating an LPF into the sliding surface, high-frequency chattering is effectively suppressed without compromising the fast transient response inherent to sliding mode control. Furthermore, by enforcing the design parameter condition $\alpha = T$, the explicit dependence of the equivalent control law on the battery internal resistance R_b was eliminated, conferring strong parameter immunity against internal impedance variations. Comparative simulation results demonstrated that the proposed control scheme significantly outperforms conventional PI controllers, achieving shorter rise and settling times and seamless power flow reversals with unit power factor. Finally, the execution and practical feasibility of the algorithm were validated via a HIL experimental platform utilizing a PLECS RT-Box CE and a TMS320F28379D DSP board.

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Wensheng Luo (Member, IEEE) received the B.S. degree in mechanical design, manufacturing and automation from Harbin Engineering University, China, in 2007; the M.S. degree in mechatronic engineering, and the Ph.D. degree in control science and engineering from Harbin Institute of Technology, China, in 2009 and 2019 respectively; the Ph.D. degree in automation, electronics and telecommunications engineering from the University of Seville, Spain, in 2019. She is currently an associate professor at the School of Electrical Engineering and Automation, Harbin Institute of Technology, China. Her research interests include observer-based control, sliding mode control, model predictive control, and their application in power converters.

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Author Contributions

W.L.: conceptualization, methodology, investigation; Z.J.: investigation, writing—draft preparation; Z.Z.: data curation, investigation; B.L.: writing—reviewing and editing; L.G.F.: software, validation; D.X.: supervision. All authors have read and agreed to the published version of the manuscript.

Funding

This work was supported in part by the National Natural Science Foundation of China, under grant 62673166, W2512030, the National Key Laboratory of Laser Spatial Information Foundation, under grant LSI2026WDZC14.

Institutional Review Board Statement

Not applicable.

Informed Consent Statement

Not applicable.

Data Availability Statement

All data supporting the findings of this study are included in the article.

Conflicts of Interest

The authors declare no conflict of interest.

Use of AI and AI-Assisted Technologies

No AI tools were utilized for this paper.

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