Robust and Unity Input Power Factor Control Scheme for Electric Vehicle Battery Charger

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Abstract

This study develops a digital control scheme with power factor correction for a front-end converter in an electric vehicle battery charger. The front-end converter acts as the boost-type switching-mode rectifier. The converter assumes the two roles of the battery charger, which include power factor control and robust charging performance. The proposed control scheme consists of a charging control algorithm and a grid current control algorithm. The scheme aims to obtain unity input power factor and robust performance. Based on the linear average model of the converter, a constant-current constant-voltage charging control algorithm that passes through only one proportional-integral controller and a current feed-forward path is proposed. In the current control algorithm, we utilized a second band pass filter, a single-phase phase-locked loop technique, and a duty-ratio feed-forward term to control the grid current to be in phase with the grid voltage and achieve pure sinusoidal waveform. Simulations and experiments were conducted to verify the effectiveness of the proposed control scheme, both simulations and experiments.

Key words: Electric vehicles (EVs), Battery charger, Power factor correction (PFC) boost converter, Feed-forward control

1. Introduction

As a feasible solution for the severe global environmental pollution and energy depletion problems, electric vehicles (EVs) have become an attractive alternative to replace the conventional internal combustion engine vehicles[1,2]. Electric-based vehicles, which are powered by a battery storage system, are getting attention because their driving range can be extended by means of recharging the battery from the electric grid through a charger system at a public charging station or at home[3]. Based on the state of the art of EV technologies, it is estimated that by the year 2050, at least half of all vehicles on the road will be electric[4]. Therefore, in the near future, EVs will grow to be a prominent load on the electric utility distribution system. The impacts of EVs on distribution systems and unity interfaces were reviewed in [5] and [6], which include the power quality concern, the security problem, the load demand increasing, and the infrastructure requirement. In order to reduce these impacts, the EV battery charger should operate at unity input power factor and inject zero current harmonics into the grid.

The other requirement in the EV battery charging application is the charger must supply an appropriate current for the battery[7]. Depending on kind of the battery chemistry, there are particular charging methods for each battery. For example, Nickel batteries are charged by a constant current charging method, whereas Li-on batteries require constant current-constant voltage (CC-CV) charging strategy.
The other charging methods including the sinusoidal-ripple-current charging strategy and fast charge method were introduced in [9] and [10]. Based on these charging methods, the battery charging power varies because the battery voltage and current are functions of the charging time and the state of charge of the battery. In other words, the battery charger is a tracking application, in which there may be no single nominal operation point during the charging process\[31\]. Therefore, the battery charger must provide sufficiently stable well-characterized performance and a good dynamic response.

An EV battery charger generally can be divided into the power component and the control side, which are overviewed and evaluated in [12] by taking into account the power rating, required components and space, cost, and charging time. Among the feasible charger configurations, the inductive contactless topology shown in Fig. 1 is generally used due to its several advantages such as safety, reliability, and user convenience.

In Fig. 1, the charging system contains off-board and on-board parts in which power is transferred through a high-frequency inductive interface to achieve a maximum safe charging process. In the off-board infrastructure, the primary side DC/AC inverter generates the high frequency and dc-pulse voltage with a fixed duty cycle for the primary transducer component so as to enhance the system efficiency\[31,34\]. On the other hand, the on-board part containing a full-bridge diode AC/DC rectifier converts the high frequency output of the secondary transducer into a dc voltage to charge the battery.

The power factor correction (PFC) boost converter in the battery charger not only regulates the grid current to obtain the desired input power factor, but also controls the charging process by regulating the dc link voltage. The PFC boost converter is usually controlled by a voltage controller in the outer loop and a current controller in the inner loop\[19\]. The voltage controller regulates the dc output voltage, whereas the current controller adjusts the grid current to be in phase with the grid voltage waveform for the unity input power factor (PF) purpose. To enhance the converter performances, both academic researchers and industrial developers have proposed many control methods. For example, in [16], a notch-filter was used to improve the control dynamic response and to reduce the grid current distortion. Authors in [17] applied the washout filter to stabilize the boost PFC converters by suppressing the period-doubling bifurcation at the line frequency. In addition, an adaptive control strategy, a grid current sensorless control, a robust control algorithm, and feed-forward technique are introduced respectively in [18], [20], [20] and [21]. However, these control techniques consider either the inner control loop or the outer control loop at a single operating point so that it is hard to obtain simultaneously both requirements in EV battery charger including the unity power factor and robust charging performance.

In this paper, we develop a control scheme for the PFC boost converter in the battery charging system by taking into account both the input power factor and the output charger performance. The proposed control scheme is composed of two parts to achieve the unity input power factor and robust charging performance. The first part located in the outer is a charging control algorithm that manages the charging voltage and charging current to satisfy the CC-CV charging method. Compared with the conventional charging control scheme, the proposed one requires only one proportional-integral (PI) controller; it is able to simplify the implementation of the control system. In addition, with a current feed-forward path, the charging power variation is totally canceled out, so that the charging system can operate robustly in all operating points. Meanwhile, the second part located in the inner is a grid current control algorithm aiming to regulate the grid current in phase with the grid voltage. In this part, we apply a single-phase phase-locked loop (PLL) and a second band-pass filter (BPF) to create a pure sinusoidal grid current, leading to a harmonic-free grid current regardless of the grid voltage condition. Furthermore, a duty ratio feed-forward is added to eliminate the phase displacement between the grid voltage and grid current so that the charger obtains the unity input power factor.

The paper is organized as follows. In Section 2, the PFC boost converter is modeled in both the rectified
grid voltage period and the switching time interval aiming to develop the system control scheme. Next, the charging control and current control algorithms are described in Section 3 and Section 4, respectively. Section 5 shows the simulation and experimental results in order to verify the proposed control scheme values. Finally, in Section 6, the paper is concluded.

2. PFC Boost Converter Model

Fig. 2 shows the block diagram of the PFC boost converter in the EV battery charger, which is composed of a full bridge diode rectifier, a step-up dc chopper with energy storage elements, and an electromagnetic interference (EMI) filter. It is assumed that the converter operates with a unity input power factor, and the grid voltage \( v_G(t) \) and the grid current \( i_C(t) \) are defined as

\[
v_G(t) = V_G \sin(\omega t); \quad i_C(t) = I_C \sin(\omega t) \tag{1}
\]

where \( V_G \) is the grid peak voltage, \( I_C(t) \) is the grid peak current, which is a time varying value, and \( \omega \) is the grid frequency, so the period of the rectified grid voltage becomes \( T_R = \pi/\omega \). If the load voltage and the load current of the converter are denoted as \( v_B(t) \) and \( i_B(t) \), respectively, applying the power conservation law to the converter yields

\[
v_C(t) i_C(t) = \frac{1}{2} \frac{d}{dt} [L i_R^2(t)] + \frac{1}{2} \frac{d}{dt} [C v_B^2(t)] - v_B(t) i_B(t) \tag{2}
\]

where \( i_R(t) \) is the rectified grid current through the inductor \( L \), or \( i_R(t) = |i_C(t)| \). Substituting (1) into (2), then

\[
v_C(t) i_C^2(\omega t) = \frac{1}{2} \frac{d}{dt} [L i_R^2(\omega t)] + \frac{1}{2} \frac{d}{dt} [C v_B^2(\omega t)] + v_B(t) i_B(t) \tag{3}
\]

By averaging all terms within (3) over a period \( T_R \)

Fig. 2. PFC boost converter configuration.

\[ I_G(s) \]

\[ \frac{2P(s)}{C} \]

\[ Y(s) \]

Fig. 3. Transfer function representation of averaged model of PFC boost converter.

[23], the converter model in the time domain can be obtained as expressed in (4):

\[
dy(t) \over dt = \frac{V_G I_C(t) - 2p(t)}{C} \tag{4}
\]

where \( y(t) \) and \( p(t) \) stand for the averaged values of the squared load voltage and the output power for one \( T_R \) respectively, and they are defined as follows:

\[
y(t) = \frac{1}{T_R} \int_0^{T_R} v_B^2(\tau) d\tau \tag{5}
\]

\[
p(t) = \frac{1}{T_R} \int_0^{T_R} v_B(\tau) i_B(\tau) d\tau \tag{6}
\]

From (4), the transfer function in the frequency domain can be expressed in (7), and the block diagram of the PFC boost converter model is then illustrated in Fig. 3.

\[
Y(s) = \frac{V_G I_C(s) - 2P(s)}{Cs} \tag{7}
\]

Thanks to the averaged-linear model in (7), the linear control theory can be easily applied to control the PFC boost converter.

In order to control the grid current, the converter behavior is analyzed during a switching period \( T_S \). Assuming that the switch \( M \) is turned on during the interval \( t_{ON} \) then the change in the inductor current denoted as \( \Delta i_{ON} \) during the switch on is

\[
\Delta i_{ON} = \frac{1}{L} \int_0^{t_{ON}} |v_C(\tau)| d\tau \tag{8}
\]

Meanwhile, the change in the inductor current, \( \Delta i_{OFF} \), during the switch off is

\[
\Delta i_{OFF} = \frac{1}{L} \int_0^{t_{OFF}} \{v_C(\tau) - v_B(\tau)\} d\tau \tag{9}
\]
Fig. 4. The overall control scheme for PFC boost converter in EV battery chargers.

Considering that the converter is operating at the boundary of continuous conduction mode (CCM) and discontinuous conduction mode (DCM), the magnitudes of the current changes in (8) and (9) are the same but opposite directions, i.e.,

$$\Delta i_{ON} + \Delta i_{OFF} = 0$$  \hspace{1cm} (10)

In addition, the grid and load voltages are assumed to be constant during the short interval $T_g$ then the duty ratio $d(t)$ of the converter switch can be obtained from (8)–(10) as

$$d(t) = \frac{i_{ON}}{T_s} = 1 - \frac{v_G(t)}{v_p(t)}$$  \hspace{1cm} (11)

3. The Proposed Charging Control Algorithm

Fig. 4 illustrates the overall control scheme for the charging system, which is a digital type and is composed of two cascading blocks that are called the charging control algorithm and the current control algorithm. In order to manage the battery charging current and voltage, the charging control algorithm determines the grid peak current $I_G^p$ from the information such as the instantaneous battery voltage and current, the charging voltage and charging current ratings ($V_M$ and $I_M$), and the grid peak voltage, $V_G$. Meanwhile, the current control algorithm, which will be discussed meticulously in the next Section, regulates the switch duty cycle in order to achieve a desired grid current waveform in terms of the magnitude and phase.

In Fig. 5(a), a popular charging method called the CC-CV method is shown [22]. It is usually used to charge lithium-ion, lead-acid, or other batteries that are vulnerable to damage if their voltage exceeds the voltage limit, $V_{AB}$. In CC mode, the charging current is set at the rated quantity $I_M$ in order to reduce the charging time. When the battery voltage reaches $V_{AB}$ the charging process is shifted to the CV mode. During the CV mode, the charging current decreases to a predetermined cut-off current level, $I_{Cut}$, which indicates that the battery is fully charged. To implement the CC-CV charging method, a control algorithm including two cascaded PI controllers as shown in Fig. 5(b) is usually adopted [11], [14], and [26]. The PI-based voltage regulator located in the outer loop regulates the battery charging voltage, whereas the current controller located in the inner loop controls the battery charging current. Even though the control scheme in Fig. 5 is able to manage the CC-CV charging method, it still poses many drawbacks, such as the complexity and difficulty in the design of the controller gains.

In order to overcome the disadvantages of the conventional charging control method, we propose an advanced charging control algorithm as shown in Fig. 6. In the proposed control method, the charging current and charging voltage are regulated individually.
The CC or CV mode is selected based on the battery voltage $v_p(t)$ and battery charging current $i_p(t)$ by using the modes selector. During the CC mode, the battery current is controlled to be constant while the battery voltage is used only to supervise the charging process. During the CV mode, the battery voltage is controlled until the battery current reaches the cut-off level $I_{ov}$. Because the CC mode and CV mode operate independently, the charging algorithm for the CC and CV modes can be implemented more accurately and effectively. Furthermore, the proposed charging control algorithm requires only one PI regulator. This makes the control system simpler, so it is easy to design and implement.

### 3.1 Constant Voltage Control Mode

In CV mode, the battery voltage is controlled with the maximum voltage $V_{ah}$ to provide the grid current reference $I_{ah}^*$ for the inner current control algorithm. In this paper, the converter is modeled over one rectified grid period $T_{gr}$ as described in (7), by considering not the battery voltage, but its squared value in order to design the controller effectively. Therefore, the charging control algorithm takes into account the squared value, and the reference values, $y_0$ and $x_0$ in Fig. 6, are defined as follows:

$$y_0 = V_{ah}^2; \quad x_0 = I_{ah}^2$$

(12)

The complete charging system in the CV mode can be shown schematically in Fig. 7. By comparing the squared battery voltage and its reference, the desired grid current magnitude is created through a PI controller and a current feed-forward term $(2P(s)/V_G)$:

$$I_G^*(s) = k_p[y_0(s) - Y(s)] + k_i \frac{Y(s) - Y(s)}{s} + \frac{2P(s)}{V_G}$$

(13)

where $k_p$ and $k_i$ are the proportional and integral gains of the PI controller, and $Y_0(s) = y_0/s$. With the assumption that the grid current controller achieves its desired reference, i.e., $I_G^*(t) = I_G^*(t)$, then substituting (13) into the converter model defined in (7) yields the squared output voltage, which is expressed as:

$$Y(s) = \frac{V_G(k_p^2 + k_i)}{(Cs^2 + V_Gk_p^2 + V_Gk_i)s}y_0$$

(14)

Based on the final value theorem, we can prove that the squared output voltage in (14) reaches its reference in the steady state:

$$\lim_{t \to \infty}(s \cdot Y(s)) = y_0$$

(15)

In addition, the squared output voltage indicated in (14) is independent of the load power. This means that the system can be stable regardless of any load variation, which is a crucial requirement in the control of a tracking application such as a battery charger.

In order to determine the PI controller gains, $k_p$ and $k_i$, (14) is transformed into the time domain and the results are described by (16)–(19):

$$y(t) = y_0 \left[1 - e^{-\frac{\omega_n}{\sqrt{1-\xi^2}}} \sin[(\omega_n - \xi) t + \theta] + \frac{2e^{-\frac{\omega_n}{\sqrt{1-\xi^2}}} \sin((\omega_n + \xi) t)]}{\sqrt{1-\xi^2}}\right]$$

(16)

$$\omega_n = \sqrt{\frac{V_Gk_i}{C}}$$

(17)

$$\xi = \frac{k_p}{2} \sqrt{\frac{V_G}{Ck_i}}$$

(18)

$$\theta = \arccos(\xi).$$

(19)

From (17) and (18), the natural frequency is dependent only on the integral gain $k_i$, while the damping ratio $\xi$ depends upon both gains of the PI controller. To make the system viable, the damping ratio needs to be less than unity, and then the PI controller gains must satisfy the following relationship:

Fig. 7. The charging system diagram under CV mode.
In Fig. 8, the waveforms of the squared output voltage corresponding to some specified damping ratio and the natural frequency values are plotted. It is observed that increasing the natural frequency results in a reduction of the setting time or an enhancement of the system response. In addition, with a higher damping ratio, the overshoot of the output voltages becomes lower. However, the increases of both the natural frequency and damping ratio cause a higher steady state error and can make the system more sensitive to noise under practical circumstances.

3.2 Constant Voltage Control Mode

During constant current mode, the charging current is kept at the rating \( I_S \) while the battery voltage increases until it reaches the limit \( V_S \). Actually, the voltage increment is not linear in terms of time, but the charging time in this mode takes several minutes. Thus, it is acceptable to assume that the battery voltage is constant during one rectified grid period \( T_g \), and a ratio between the battery voltage and the battery current, \( R_B \), which is defined in (21), is kept constant during one rectified grid period.

\[
R_B = \frac{v_B(t)}{i_B(t)}. \tag{21}
\]

For current control, the converter model shown in (7) can be modified in terms of the squared charging current, \( X(s) \):

\[
X(s) = \frac{V_G J_G - 2P(s)}{CR_B^2 s}. \tag{22}
\]

\[
\frac{k_p^2}{k_I} < \frac{4C}{V_G}. \tag{20}
\]

Fig. 8. Time response of the squared output voltage.

Similar to the CV mode, the feed-forward and PI controller are also used to generate the grid current reference in Fig. 6, and the squared charging current in the CC mode is derived as follows:

\[
X(s) = \frac{V_G \left( k_p \frac{k_p}{R_B^2 s + k_I} \right) x_0}{\left( C s^2 + V_G \frac{k_p}{R_B^2} s + V_G \frac{k_I}{R_B^2} \right) s}. \tag{23}
\]

Comparing the squared charging voltage in (14) with the squared charging current in (23), the PI controller used in CV mode can be applied to CC mode if the PI gains are increased by \( R_B^2 \) times. In other words, by multiplying the difference between the squared charging current and its reference with the coefficient \( k \) (where \( k = R_B^2 \)) as shown in Fig. 6, the PI controller defined in CV mode can be used in CC mode without affecting the charging system performance. In order to determine \( R_B \), a second-order low-pass filter with a cut-off frequency less than the double grid frequency is utilized so as to obtain the averaged value of the battery charging current and the battery voltage.

4. Grid Current Control Algorithm

The grid current control algorithm aims to regulate the grid current to obtain the desired magnitude \( I_g^* \) and to be in phase with the grid voltage. In Fig. 9, the phase and magnitude of the grid voltage fundamental component are detected through a second BPF and a single PLL block. Since the grid current is controlled in phase with only the fundamental component of the grid voltage, the grid current can be a pure sinusoidal waveform under any grid voltage condition. In order to detect the fundamental
The component of the grid voltage without the phase shift, the center frequency of the BPF is selected to be the grid frequency, and the filter transfer function is defined as follows:

$$H(s) = \frac{bs}{s^2 + bs + \omega_c^2}$$  \hspace{1cm} (24)

where \(\omega_c\) is the center frequency of the filter \((\omega_c = \omega \text{ rad/s})\), and \(b\) is the bandwidth of the filter and is chosen to be one-tenth of the center frequency \((b = \omega / 10 \text{ rad/s})\). Subsequently, the phase of the fundamental voltage is obtained by using the single PLL technique depicted in Fig. 10, which emulates the three-phase PLL method. A virtual \(\pi/2\) phase-lag voltage \((v_\beta)\) compared with the fundamental component grid voltage \((v_F)\) is obtained by using a first-order 90-degree-phase-shift function (PSF) that is defined as

$$P(s) = \frac{s - \omega}{s + \omega}$$  \hspace{1cm} (25)

At the grid frequency \(\omega\), the output signal of this function lags at an angle of \(\pi/2\) and has an equal magnitude in comparison with its input. Afterwards, \(v_\alpha\) and \(v_\beta\) are changed to the direct quantities \(v_d\) and \(v_q\) by using the Park Transform:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} \cos \theta & -\sin \theta \\ \sin \theta & \cos \theta \end{bmatrix} \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix}$$  \hspace{1cm} (26)

When the angle \(\theta\) in (26) equals the phase of \(v_\alpha\) (i.e., the phase of \(v_F\)), \(v_d\) becomes zero and \(v_q\) has the same value as the peak value of \(v_F\); the \(v_d\) component is kept at zero by the PI controller as shown in Fig. 10.

In order to improve the current control performance, the duty-ratio feed-forward method in [21] is applied by using the duty ratio defined in (11). Actually, if the control scheme is implemented through an analog circuit, its bandwidth will be high enough to provide good performance. On the other hand, the digital control implemented by the DSP has a limitation in terms of the bandwidth due to the limited sampling time and switching frequency. Therefore, in the grid current control algorithm, the feed-forward term is indispensable in ensuring that the grid current waveform is identical to the grid voltage.

### 5. Simulation and Experimental Results

In order to verify the effectiveness of the proposed control scheme, a simulation and experiment are carried out. The prototype 2kW PFC boost converter with the parameters in Table 1 is set up in a laboratory and managed by a high performance DSP (TMS320F28335 from Texas Instruments). In both the simulation and experiment, an 110Vrms/60Hz ac grid is utilized to supply power for the system. The charging control algorithm is designed to achieve a natural frequency of \(\omega_n = 250 \text{ rad/s}\) and a damping
ratio of $\xi = 0.707$. From (17) and (18), the PI gains are obtained to be $k_p = 1.89$ and $k_i = 0.011$. In the grid current control, the duty-ratio feed-forward component significantly improves the steady-state response of the grid current waveform; and the PI gains, which decides the transient response, are selected to be $k_p = 0.078$ and $k_i = 15.4$.

In the simulation, the performance of the proposed control scheme is evaluated in both CC and CV modes with a 40Ω resistive load. Fig. 11 shows the simulation results in CC mode, where the output current is regulated at 5A from the starting point, and the current reference is changed to 8A at the time $t=0.5s$. Based on the five-percent error criterion, it is observed that the system performs with 20ms ($2.4T_\omega$) of rising time, 30ms ($3.6T_\phi$) of setting time, and 3.5% overshoot. During the transient state, THD is maintained at less than 8.54%, and the power factor is higher than 0.958. In steady state, the load current tracks the reference perfectly and the grid current shows not only a pure sinusoidal waveform (THD=1.83%), but also a near unity input power factor (PF=0.99985). Compared with the conventional control methods in [6] and [9], in which the grid current THD is in the range of 3.36% to 5.26%, the proposed control strategy significantly enhances the grid current waveform.

Fig. 12 shows the system performance under CV control mode. At 0.7 seconds, the load voltage reference is increased from 200V to 300V. As we can see, the system performance is as good as that of the CC control mode.

The platform setup to experimentally verify the proposed control scheme is shown in Fig. 13(a). In order to correctly detect the magnitude and phase of the fundamental grid voltage $v_F$, we use the BPF and PLL as shown in Fig. 9. In the experiment, the 110Vrms/60Hz ac source distorted with THD=9% is used because it is hard to obtain a pure sinusoidal voltage source in the laboratory. Fig. 13(b) shows that thanks to the BPF and PLL, the fundamental grid voltage $v_F$ is detected perfectly regardless of the distorted grid voltage.

In order to investigate the contribution of the BPF and PLL, the performances with and without the BPF and PLL blocks are plotted in Fig. 14 under the
several different distorted grid voltage conditions with the aid of a Programmable AC Power Source (Chroma 61704). In Fig. 14, we can see that the grid current maintains an almost sinusoidal waveform with low THD (around 2%) independent of the grid voltage conditions (the solid line). Meanwhile, in the case without the BPF and PLL, the grid current is distorted with the THD proportional to that of the grid voltage (the dashed line). According to the IEC-100-3-2 standard\(^\text{29}\), the maximum THD level of the grid current is 6.4%. Therefore, the conventional grid current control algorithm cannot comply with the standard when the grid voltage distortion is higher than 3.5%. On the contrary, the proposed control scheme is able to satisfy the standard under any grid
voltage condition, which verifies the value of the proposed method.

Fig. 15 shows the experimental performance of the CC mode with the load current of 5A for a 1.3kW resistive load. Even though the grid voltage is highly distorted, the grid current is almost a sinusoidal waveform (THD=2.01%) and matches the phase of the grid voltage (PF=0.9994). In addition, we can see that the load current and load power are stable, and the load current tracks its command (5A) successfully.

In order to evaluate the grid current quality, its FFT is analyzed in Fig. 16. Even though the grid current also contains third and fifth order components, their magnitudes are very small, with 1.12% and 1.67% magnitude, respectively, as compared to the fundamental component.

In CV mode, the proposed control scheme aims to regulate the output voltage at 200V dc to supply 1.4kW power for a resistive load. In CV mode, the source voltage is also distorted with THD=6.2%. Fig. 17 shows the steady state performances in CV mode. It is recognized that the output voltage is kept stable at 200V and the grid current is almost sinusoidal as in the CC mode. Moreover, in order to validate the dynamic performance of the charging system, the output load power is alternated between 1kW and 1.4kW, and the results are shown in Fig. 18. We can see that the grid current is always maintained as a stable sinusoidal waveform in spite of the load power variation. It is evident that the proposed control scheme has good dynamic performance and robustness against load variations.

6. Conclusions

In this paper, an effective control scheme with unity input power factor and robust performance for an EV charging system has been presented. The battery charger adopts a PFC boost converter not only for regulating the input power factor, but also for managing the battery charging current and voltage. The proposed charging control algorithm obtains good dynamic performance although it requires only one PI regulator. In the grid current control strategy, the input current is controlled to be in phase with the fundamental component of the grid voltage, so that the charging system injects zero harmonics into the grid and achieves a unity input power factor regardless of the grid voltage distortion. The efficiencies of the proposed control scheme have been proven through simulations and experiments.

The proposed control scheme can be utilized in charging systems that adopt other front-end PFC converters, such as the PFC buck-based topology or the PFC symmetrical bridgeless boost rectifier-based configuration.

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