Reducing Capacity of Smart Charger for Electric Vehicles on Single-Phase Three-Wire Distribution Feeders with Reactive Power Control

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This paper proposes a new control algorithm to reduce the capacity of the previously proposed smart charger for electric vehicles (EVs) on single-phase three-wire distribution feeders using reactive power control on the source side. The basic principle of the proposed reactive power control algorithm is discussed in detail. It is shown that controlling the reactive power on the source side reduces the capacity of the previously proposed smart charger. A digital computer simulation is carried out to confirm the validity of the proposed control algorithm using PSIM software. A prototype experimental model is constructed and tested. The experimental results demonstrate that balanced source currents with a power factor of 0.9, which conforms to Japanese regulations, are obtained on the secondary side of the pole-mounted distribution transformer during both the battery charging and discharging operations in EVs reducing the capacity of the smart charger by 31% in comparison to that of the smart charger with the previously proposed control algorithm.

Keywords: single-phase three-wire distribution system, smart charger, three-leg inverter, constant dc-capacitor voltage control, single-phase d - q transformation

1. Introduction

Smart meters are now installed for household consumers in Tokyo and Kansai electric power companies' areas, but not for commercial consumers yet. The smart meters can report to the electric power companies the power consumption conditions of each household consumer. It is natural that household electricity prices for consumers with high-quality power consumption should be lower than that for consumers with low-quality power consumption in the near future. Thus, each household consumer will be responsible for improving the power consumption conditions. Single-phase three-wire distribution feeders are used for household consumers in Japan. Load conditions for household consumers are always unbalanced. These unbalanced conditions cause unbalanced voltages on the secondary side of pole-mounted distribution transformers. The secondary-side source currents of the pole-mounted distribution transformers are also unbalanced. It is well known that these unbalanced source-current conditions increase losses in pole-mounted distribution transformers. Power companies compensate for these losses, although the losses are caused by household consumers. There

are a large number of pole-mounted distribution transformers in Japan, so the total economic losses are great. It is also natural that household consumers should be responsible for compensating the losses in pole-mounted distribution transformers in the near future.

On the other hand, electric vehicles (EVs) are now commercially available. The Mitsubishi i-MiEV is a five-door hatchback electric kei car, and the Nissan LEAF is a medium-size five-door hatchback electric car. The lithium ion batteries equipped in the i-MiEV can store 16 kWh, whereas in the LEAF, the maximum stored power is 24 kWh. These EVs are highly mobile owing to the stored electric power. With this mobility, an interesting concept, in which the stored power is used as a peak power shaver, has been proposed^{(1)–(3)}. To achieve this, a PWM rectifier with a bidirectional dc-dc converter is used as a charger. The present authors, thus, have proposed a smart charger for EVs with a power quality compensator on single-phase three-wire distribution feeders⁽⁴⁾. The proposed smart charger consists of four-leg IG-BTs. The first and second legs are connected to Feeder1 and Feeder2, respectively. The third leg is connected to the neutral line. This is an advantage of the previously proposed smart charger in comparison to the charger in Ref. (3). These three legs are used as a single-phase PWM rectifier, which converts power from ac to dc during the battery charging operation or from dc to ac during the battery discharging operation. This PWM rectifier can compensate for reactive and unbalanced active currents on single-phase three-wire distribution feeders by connecting the third leg to the neutral line. Simulation and experimental results have demonstrated that

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balanced source currents with a power factor of unity are obtained on the source side of single-phase three-wire distribution feeders during both the battery charging and discharging operations in EVs. When EVs are not connected to the proposed charger, the power quality is also improved on the secondary side of the pole-mounted distribution transformer.

For practical use, lower-priced chargers are needed. Reducing the capacity of the three-leg PWM rectifier, which performs as a smart charger, is effective to offer the smart charger as a commercial product. A source-side power factor of 0.9 is acceptable for home appliances in Japan according to the regulations⁽⁵⁾. Thus, controlling the reactive power on the source side is effective in reducing the capacity of the proposed smart charger. The present authors have proposed a method for reducing the required capacity of a three-leg structured PWM rectifier with reactive power control⁽⁶⁾. The correlation and cross-correlation coefficients between the source voltage and load currents are applied to define the fundamental active-reactive current in single-phase circuits. Going back by $T_S/4$ in the detected source voltage, the orthogonal component to the source voltage is generated, where T_S is the time period of the source voltage. The cross-correlation coefficient between the source voltage and the load current is calculated with this orthogonal component, and the reactive current of the load current is calculated with this cross-correlation coefficient. The proposed control algorithm with the correlation and cross-correlation coefficients requires many calculation steps. For practical use, a simple reactive-power control method is also required.

In this paper, we propose a new control algorithm for the previously proposed smart charger to reduce the capacity of the three-leg PWM rectifier, which performs as the smart charger, with reactive power control on the source side. A single-phase PLL in single-phase d - q coordinates is used to detect the phase angle of the source voltage. The in-phase and orthogonal components to the source voltage can be easily generated with the detected phase angle. Thus, the authors offer a simpler reactive-power control algorithm than that of Ref. (6). The basic principle of the proposed control algorithm is discussed in detail and then confirmed by a digital computer simulation using PSIM software. A prototype experimental model is also constructed and tested. The experimental results demonstrate that balanced source currents with a power factor of 0.9, which is acceptable for Japanese home appliances, are achieved during both the battery charging and discharging operations in EVs. This reduces the capacity of the smart charger by 31% in comparison to that of the smart charger employing the previously proposed control algorithm.

Note that the charging or discharging power in both the simulation and experimental results is the same as that in Ref. (4). Thus, reducing the required capacity of the smart charger does not affect the battery charging and discharging capabilities.

2. Reduction in Capacity of Smart Charger with Reactive Power Control

Figure 1 shows a power circuit diagram of the previously proposed smart charger⁽⁴⁾. The rating of the pole-mounted distribution transformer is 6600 V_{rms}, 5.0 kVA, and 60 Hz on

the primary side, and 105 V_{rms} and 24 A_{rms} on the secondary side. The proposed smart charger consists of four-leg IGBTs. Three legs are used for a PWM rectifier, which converts power from ac to dc during the battery charging operation or from dc to ac during the battery discharging operation. One leg is connected to Feeder1, the second leg is connected to Feeder2, and the third leg is connected to the neutral line. This PWM rectifier can compensate for reactive and unbalanced active currents on single-phase three-wire distribution feeders by connecting the third leg to the neutral line. The fourth leg is used as a bidirectional dc-dc converter for both the battery charging and discharging operations in EVs.

Figure 2 shows a control circuit diagram of the newly proposed control algorithm for Fig. 1 with reactive power control on the source side. The basic principle of the proposed control algorithm is discussed here. Figure 3 shows a phasor diagram of the source voltage v_L , source current i_S , and load current i_L . $\hat{V}_L$ is the source voltage phasor, $\hat{I}_L$ is the load current phasor, and $\hat{I}_S$ is the source current phasor. $\hat{I}_{Lp}$ is the active current phasor for $\hat{I}_L$, and $\hat{I}_{Lq}$ is the reactive current phasor for $\hat{I}_L$. When balanced source currents with a power factor of unity are achieved, the smart charger supplies $\hat{I}'_{SC}$, which can perfectly compensate for the reactive current $\hat{I}_{Lq}$. This perfect reactive current compensation requires a large-capacity smart charger. According to Japanese regulations⁽⁵⁾, a power factor of 0.9 is acceptable for home appliances. The source-side power factor is 0.9 if only $K\hat{I}_{Lq}$ is compensated, where $K = (I_{Lp} + I_{Sbat}) \tan(\cos^{-1} 0.9) / I_{Lq}$. Then, the charger currents fed by the smart charger are $\hat{I}_{SC}$, where balanced source currents with a power factor of 0.9 are obtained. Comparing $\hat{I}'_{SC}$ with $\hat{I}_{SC}$, it can be seen that controlling the fundamental reactive currents on the source side reduces the amplitudes of the charger currents of the smart charger. Thus, we can reduce the capacity of the smart charger. The required capacity of the smart charger with the proposed reactive power control will be shown in simulation and experimental results.

Let the primary-side voltage v_S , secondary-side voltages v_{L1} and v_{L2} , and load currents i_{L1} and i_{L2} be

$$\begin{aligned} v_S &= \sqrt{2}V_S \cos \omega_S t, \\ v_{L1} &= v_{L2} = \sqrt{2}V_L \cos \omega_S t, \\ i_{L1} &= \sqrt{2}I_{L1} \cos(\omega_S t - \phi_{L1}), \\ i_{L2} &= \sqrt{2}I_{L2} \cos(\omega_S t - \phi_{L2}). \end{aligned} \quad \dots\dots\dots (1)$$

The source voltage v_S is detected, and then the phase angle $\theta_S = \omega_S t$ is generated by the single-phase PLL in single-phase d - q coordinates⁽⁷⁾⁽⁸⁾. $\sqrt{2} \cos \omega_S t$, which is in phase with v_S , and $\sqrt{2} \sin \omega_S t$, which is orthogonal to v_S , are generated with θ_S . The generated $\sqrt{2} \cos \omega_S t$ is multiplied by the load currents i_{L1} and i_{L2} . Then,

$$\begin{aligned} &\frac{1}{2} \{ (i_{L1} + i_{L2}) \cdot \sqrt{2} \cos \omega_S t \} \\ &= \frac{1}{2} (I_{L1} \cos \phi_{L1} + I_{L2} \cos \phi_{L2}) \\ &\quad + \frac{1}{2} \{ I_{L1} \cos(2\omega_S t - \phi_{L1}) + I_{L2} \cos(2\omega_S t - \phi_{L2}) \}. \end{aligned} \quad \dots\dots\dots (2)$$

$2\omega_S$ components are eliminated using a moving-average

Fig. 1. Power circuit diagram of proposed smart charger for electric vehicles

Fig. 2. New control circuit diagram of smart charger with reactive power control

Fig. 3. Phasor diagram of source voltage, source current, and load current.

low-pass filter (LPF). The transfer function of the moving-average LPF is expressed as

$$H(z) = \frac{1}{N} \sum_{n=0}^{N-1} z^{-n}, \dots \dots \dots (3)$$

where N is the number of samples. The dc component I_{Lp} is given as

$$I_{Lp} = \frac{1}{2} \overline{\{(i_{L1} + i_{L2}) \cdot \sqrt{2} \cos \omega_s t\}} \\ = \frac{1}{2} (I_{L1} \cos \phi_{L1} + I_{L2} \cos \phi_{L2}). \dots \dots \dots (4)$$

The dc component I_{Lp} corresponds to the mean active component for i_{L1} and i_{L2} . Thus, the mean active current i_{Lp} for each load is expressed as

$$i_{Lp} = \sqrt{2} I_{Lp} \cos \omega_s t. \dots \dots \dots (5)$$

The mean reactive component I_{Lq} of i_{L1} and i_{L2} with the generated $\sqrt{2} \sin \omega_s t$ is also given by

$$I_{Lq} = \frac{1}{2} \overline{\{(i_{L1} + i_{L2}) \cdot \sqrt{2} \sin \omega_s t\}} \\ = \frac{1}{2} (I_{L1} \sin \phi_{L1} + I_{L2} \sin \phi_{L2}). \dots \dots \dots (6)$$

The mean reactive current i_{Lq} of i_{L1} and i_{L2} is expressed as

$$i_{Lq} = \sqrt{2} I_{Lq} \sin \omega_s t. \dots \dots \dots (7)$$

Let the dc-side power that is from or to the battery be P_{bat} . P_{bat} is expressed as

$$P_{bat} = V_{Cf2} \cdot I_{LS2}, \dots \dots \dots (8)$$

where V_{Cf2} is the mean value of v_{Cf2} and I_{LS2} is the mean value of i_{LS2} . The equivalent RMS value I_{Sbat} on the source side, which is calculated from P_{bat} , is given by

$$I_{Sbat} = \frac{P_{bat}}{V_L}. \dots \dots \dots (9)$$

The total active current i_{Sbat} , which is given by P_{bat} , is expressed as

$$i_{Sbat} = \sqrt{2} \frac{P_{bat}}{V_L} \cos \omega_s t \\ = \sqrt{2} I_{Sbat} \cos \omega_s t, \dots \dots \dots (10)$$

Thus, the desired active current for each feeder on the source side is given by

$$\begin{aligned} i_{Sp} &= i_{Lp} + \frac{i_{Sbat}}{2} \\ &= \sqrt{2}(I_{Lp} + \frac{I_{Sbat}}{2}) \cos \omega_s t. \end{aligned} \quad \dots\dots\dots (11)$$

In Fig. 2, the sum of i_{L1} , i_{L2} and i_{Sbat} is calculated. Then, the generated sum is multiplied by the gain of 0.5. A moving-average LPF is used to calculate the active component I_{Sp} , which is $I_{Lp} + I_{Sbat}/2$. The required reactive component Ki_{Sq} is calculated with the power-factor-control gain of K and the i_{Sq} , which multiplies I_{Sp} by $\sqrt{2} \sin \omega_s t$. Thus, the desired source current i_s^* in both feeders, Feeder1 and Feeder2, is expressed as

$$i_s^* = \sqrt{2}I_{Sp}\{\cos \omega_s t + \tan(\cos^{-1} 0.9) \sin \omega_s t\}. \quad \dots\dots\dots (12)$$

The dc-capacitor voltage v_{DC} of the single-phase three-leg PWM rectifier is detected, and then the difference Δv_{DC} between the reference value V_{DC}^* and the detected dc-capacitor voltage v_{DC} is amplified by a PI controller. The reference active component i_{DC}^* to control the dc-capacitor voltage v_{DC} is obtained by multiplying I_p by $\sqrt{2} \cos \omega_s t$, which is calculated from θ_s . The reference values i_1^* , i_2^* , and i_3^* for the smart charger are thus expressed as

$$\begin{aligned} i_1^* &= i_{L1} - i_s^* - 0.5i_{DC}^*, \\ i_2^* &= -i_{L2} + i_s^* + 0.5i_{DC}^*, \\ i_3^* &= -(i_1^* + i_2^*). \end{aligned} \quad \dots\dots\dots (13)$$

During the battery discharging operation of the circuit in Fig. 1, the power discharged from the batteries is automatically injected into the utility grid by constant dc-capacitor voltage control. During the battery charging operation, the power charged to the batteries is also automatically provided by the utility grid with constant dc-capacitor voltage control. In simulation and experimental results, which will be shown later, the power charged to and that discharged from the batteries are the same as those in Ref. (4). The authors claim that reducing the capacity of the smart charger with reactive power control does not affect the battery charging and discharging operations.

It is well known that a steady-state error remains when current feedback control based on the sine-triangle intercept technique with a PI controller is used. To avoid this error, an interesting current feedback control scheme in d - q coordinates for single-phase circuits has been proposed⁽⁹⁾. Figure 4 shows a detailed block diagram of the single-phase current feedback control in d - q coordinates in Fig. 2. The reference value i_1^* is delayed by $T_s/4$ through the $T_s/4$ delay block. This reference value corresponds to the α -phase reference value, and the reference value delayed by $T_s/4$, $i_{1\beta}^*$, corresponds to the β -phase reference value. The detected output current i_1 of the first leg is detected. This detected current corresponds to the α -phase current i_1 . i_1 is also delayed by $T_s/4$. This delayed component corresponds to the β -phase

Fig. 4. Control circuit diagram of single-phase current feedback in d - q coordinates

current $i_{1\beta}$. The differences between the reference values i_{d1}^* and i_{q1}^* and the detected values i_{d1} and i_{q1} are amplified by the PI controllers in d - q coordinates. The amplified differences are retransformed to α - β coordinates, and the sine-triangle intercept technique is used to control the output currents i_1 and i_2 Profs. Saitou and Shimizu proposed a novel current feedback control method in d - q coordinates for single-phase circuits, in which a Hilbert transform is used⁽¹⁰⁾⁽¹¹⁾. In their method, the Hilbert transform is implemented by numerical differentiation, which may result in instability in experimental systems. Thus, the feedback control in d - q coordinates for single-phase circuits with a $T_s/4$ delay block is more feasible.

3. Required Capacity of DC Capacitor

From Ref. (13), the battery current ripple should be less than 5%. The required capacity of the dc capacitor C_{DC} is discussed to ensure that the battery current ripple is less than 5% in the steady state. Here, the transfer function of the bidirectional dc-dc converter is derived to discuss the battery current ripple. It is well known that the transfer function of the bidirectional dc-dc converter can be derived using the state-space averaging method. Figure 5 shows an analysis circuit diagram of the bidirectional dc-dc converter. Let the state variables for the bidirectional dc-dc converter be $\mathbf{x} = [i_{LS2} \ v_{DC}]^T$. The state-space averaging equation and the output equation during the battery charging operation are given by

$$\begin{cases} \frac{d\mathbf{x}}{dt} = \bar{\mathbf{A}}\mathbf{x} + \bar{\mathbf{b}}v_{DC} \\ i_{bat} = \bar{\mathbf{c}}\mathbf{x} \end{cases} \quad \dots\dots\dots (14)$$

where

$$\begin{aligned} \bar{\mathbf{A}} &= \begin{bmatrix} 0 & \frac{1}{L_{S2}} \\ -\frac{1}{C_{f2}} & -\frac{1}{rC_{f2}} \end{bmatrix}, & \bar{\mathbf{b}} &= \begin{bmatrix} \frac{D}{L_{S2}} \\ 0 \end{bmatrix}, \\ \bar{\mathbf{c}} &= \begin{bmatrix} 0 & -\frac{1}{r} \end{bmatrix}, \end{aligned}$$

and $D = V_{bat}/V_{DC}$. The transfer function $G_1(s)$ between $I_{bat}(s)$ and $V_{DC}(s)$ during the battery charging operation is expressed as

$$G_1(s) = \frac{I_{bat}(s)}{V_{DC}(s)} = \bar{\mathbf{c}}(s\mathbf{I} - \bar{\mathbf{A}})^{-1}\bar{\mathbf{b}}. \quad \dots\dots\dots (15)$$

Thus, $G_1(s)$ is given by

$$G_1(s) = \frac{D}{rL_{S2}C_{f2}} \frac{1}{s^2 + \frac{s}{rC_{f2}} + \frac{1}{L_{S2}C_{f2}}} \quad \dots\dots\dots (16)$$

Fig. 5. Analysis circuit diagram of the bidirectional dc converter

For the battery discharging operation,

$$\bar{A} = \begin{bmatrix} 0 & \frac{1}{L_{S2}} \\ -\frac{1}{C_{f2}} & -\frac{1}{rC_{f2}} \end{bmatrix}, \quad \bar{b} = \begin{bmatrix} -\frac{1-D}{L_{S2}} \\ 0 \end{bmatrix},$$

$$\bar{c} = \begin{bmatrix} 0 & -\frac{1}{r} \end{bmatrix}.$$

The transfer function $G_2(s)$ during the battery discharging operation is given by

$$G_2(s) = \frac{\frac{1-D}{rL_{S2}C_{f2}}}{s^2 + \frac{s}{rC_{f2}} + \frac{1}{L_{S2}C_{f2}}}. \quad (17)$$

The absolute value $\|G_1(j2\omega_s)\|$ for the $2\omega_s$ component, where ω_s is the angular frequency of the source voltage v_s , is 0.376, while $\|G_2(j2\omega_s)\|$ is 0.0261.

The inverse of the absolute value $\|G_2(j2\omega_s)\|$ is larger than that of $\|G_1(j2\omega_s)\|$ for the $2\omega_s$ component. The allowed ripple ratio of the dc-capacitor voltage v_{DC} is discussed with the inverse of 0.0261. From Ref. (13), the ripple current amplitude of i_{bat} should be less than 0.25 A when the mean value of the battery current $\bar{i}_{bat}$ is 5 A. The allowed fluctuation voltage $\bar{v}_{DC}$ is 9.58 V. In the following simulation and experimental results, the mean value of the dc-capacitor voltage $\bar{v}_{DC}$ is 385 V. The allowed ripple ratio of the dc-capacitor voltage v_{DC} is less than 2.5%. The required capacity of the dc capacitor C_{DC} with a ripple ratio of 2.5% is discussed considering the instantaneous power flows into C_{DC} . The instantaneous power p_C flowing into the three-leg rectifier is given by

$$\begin{aligned} p_C &= -v_{L1} \cdot i_{SC1} + v_{L2} \cdot i_{SC2} \\ &= -P_{bat} - P_{bat} \cos 2\omega_s t \\ &\quad - P_r \sin 2\omega_s t, \end{aligned} \quad (18)$$

where

$$\begin{aligned} P_{bat} &= V_L (I_{L1} \cos \phi_{L1} + I_{L2} \cos \phi_{L2} - 2I_{Sp}), \\ P_r &= V_L (I_{L1} \sin \phi_{L1} + I_{L2} \sin \phi_{L2} - 2KI_{Sp}). \end{aligned}$$

In Eq. (18), the ac component $\tilde{p}_C$ of p_C is written as

$$\tilde{p}_C = -P \cos(2\omega_s t - \phi_p), \quad (19)$$

where

$$P = \sqrt{P_{bat}^2 + P_r^2}, \quad \phi_p = \tan^{-1} \frac{P_r}{P_{bat}}.$$

Figure 6 shows the schematic waveforms for v_{DC} and $\tilde{p}_C$ during the battery charging operation. If v_{DC} is increased, $\tilde{p}_C$ is

 Fig. 6. Schematic waveforms for v_{DC} and $\tilde{p}_C$ during battery charging operation

increased from $t_0 + T_s/4$ to $t_0 + 3T_s/8$, where the initial value t_0 is $\phi_p/2\omega_s$. V_{DC} is the mean value of v_{DC} . The fluctuation ratio of v_{DC} is 2.5% in the steady state; then, the following equation is obtained:

$$\begin{aligned} &\int_{t_0+T_s/4}^{t_0+3T_s/8} \tilde{p}_C dt \\ &= \frac{1}{2} C_{DC} \{V_{DC}^2 - (0.975 V_{DC})^2\}. \end{aligned} \quad (20)$$

From Eq. (20), if v_{DC} is decreased, the following formula is obtained :

$$\begin{aligned} &\int_{t_0+T_s/8}^{t_0+T_s/4} \tilde{p}_C dt \\ &= \frac{1}{2} C_{DC} \{(0.975 V_{DC})^2 - V_{DC}^2\}, \end{aligned} \quad (21)$$

where $\tilde{p}_C$ is decreased from $t_0 + T_s/8$ to $t_0 + T_s/4$. From Eqs. (20) and (21), the fluctuation voltage amplitude of v_{DC} is $0.025V_{DC}$, which is expressed as

$$0.025V_{DC} = \frac{P}{1.975\omega_s C_{DC} V_{DC}}. \quad (22)$$

The RMS value I_0 of the fluctuation current flowing to C_{DC} is given by

$$I_0 = \frac{0.025 V_{DC}}{X_C} = \frac{\sqrt{2}P}{1.975 V_{DC}}, \quad (23)$$

where $X_C = 1/2\omega_s C_{DC}$. From Eq. (20) to Eq. (22), I_0 is determined to be $4.17 A_{rms}$. From Ref. (12), the required capacity of C_{DC} is set to $1200 \mu F$ in the steady state.

The dc-capacitor voltage fluctuations in the transient state are caused by the load variations. These fluctuations also occur after the charging or discharging operation is started or stopped. The dc-capacitor voltage fluctuations caused by the load variations are larger than those caused by the steady state. Thus, the dc-capacitor voltage fluctuations caused by the load variations are considered here. In the constructed experimental model, as shown later, the maximum dc-capacitor voltage v_{DC} is less than 400 V, while the mean value of the

dc-capacitor voltage is 385 V. Only $\Delta V_{DC} = 15$ V is allowed in the transient state when the load conditions are varied. The authors have confirmed by simulation that the fluctuation ΔV_{DC} is less than 15 V in the transient state when the load varies from heavy load to light load with a dc-capacitor C_{DC} of capacity 2200 μ F. Thus, a dc capacitor C_{DC} of capacity 2200 μ F is used in the following simulations and experiments. Typical simulation and experimental results will be shown later.

4. Simulation Results

A digital computer simulation using PSIM software was carried out to confirm the validity and high feasibility of the proposed control algorithm. Table 1 shows the constants for Fig. 1 that are used in the following simulation and experimental verifications. In the simulation results, the battery voltage V_{bat} is 360 V, which is used in the Nissan LEAF. The rating of the distribution transformer is 6600 V_{rms} , 5.0 kVA, and 60 Hz on the primary side, and 105 V_{rms} and 24 A_{rms} on the secondary side. The unbalanced ratio between Feeder1 and Feeder2 is defined by

$$\text{Unbalanced ratio} = \frac{S_1 - S_2}{S_A \times 0.5} \times 100 \text{ [\%]}, \dots \dots (24)$$

where S_1 is the apparent power of Load1 on Feeder1 and S_2 is the apparent power of Load2 on Feeder2. S_A is the total apparent power, which is the sum of the apparent powers S_1 and S_2 on Feeder1 and Feeder2, respectively. According to Japanese guidelines for the domestic use of power, the unbalanced ratio should be less than 40%⁽¹⁴⁾. Thus, an RL load of 1.2 pu with a power factor of 0.8 is connected to Feeder1 and an RL load of 0.8 pu with a power factor of 0.9 is connected to Feeder2, as shown in Fig. 1, so that the unbalanced ratio given by Eq. (24) is 40%. $K_P = 0.3$ and $T_I = 0.02$ s were used in the PI controller for constant dc-capacitor voltage control in both the simulation and experiment. $K_P = 0.05$ and $T_I = 8$ ms were used in the PI controller for current feedback control in d - q coordinates in the single-phase circuits. The reference value V_{DC}^* for the dc-capacitor voltage of the three-leg PWM rectifier was 385 V.

Figure 7 shows the simulation results for Fig. 1, where the proposed smart charger is under the battery charging operation with constant battery current control. Although the load currents i_{L1} and i_{L2} are unbalanced, the source currents i_{S1} and i_{S2} are balanced with a power factor of 0.9. As stated previously, the ripple of the battery current i_{bat} should be less than $\pm 5\%$ ⁽¹³⁾. From the simulation results in Fig. 7, the ripple of the battery current is approximately 3.47%.

The total harmonic distortion (THD) is defined as

$$\text{THD} = \frac{\sqrt{\sum_{n=2}^{\infty} I_n^2}}{I_1} \times 100 \text{ [\%]}, \dots \dots (25)$$

The THDs of i_{S1} and i_{S2} calculated from Fig. 7 are 1.54% and 0.89%, respectively. These values satisfy the guidelines⁽¹⁵⁾.

Figure 8 shows the simulation results for Fig. 1, where the proposed smart charger is under the battery discharging operation with constant battery current control. Although the load currents i_{L1} and i_{L2} are unbalanced, the source currents i_{S1}

Table 1. Circuit Constants for Fig. 1

Item	Symbol	Value
Load1 (1.2 pu, Power factor 0.8)	L_{L1}	5.8 mH
	R_{L1}	2.9 Ω
Load2 (0.8 pu, Power factor 0.9)	L_{L2}	6.3 mH
	R_{L2}	4.9 Ω
Filter inductor	L_{f1}	0.46 mH
Filter capacitor	C_{f1}	10.4 μ F
Switching inductor for three-leg inverter	L_{S1}	1.0 mH
DC capacitor	C_{DC}	2200 μ F
DC-capacitor voltage	V_{DC}^*	385 V
Switching inductor for dc-dc converter	L_{S2}	3.3 mH
Filter capacitor for dc-dc converter	C_{f2}	1000 μ F
Battery current	I_{LS2}^*	5 A
Internal resistance of battery	r	72 m Ω
Switching frequency	f_{sw}	12 kHz
Dead time	T_d	3.5 μ s

Fig. 7. Simulation waveforms for Fig. 1 during battery charging operation

and i_{S2} are balanced with a power factor of 0.9. The ripple of i_{bat} is 3.59%. The THDs of i_{S1} and i_{S2} are 1.63% and 2.84%, respectively. These values also satisfy the guidelines⁽¹⁵⁾.

The required capacity of the three-leg PWM rectifier, which performs as a smart charger, is now discussed. A definition of apparent power is generally used to calculate the required capacity of the PWM rectifier. However, the third leg that is connected to the neutral line is grounded. The definition of the usually used apparent power is not applicable to the three-leg PWM rectifier. This means that the capacity of the third leg is zero. If the third-leg capacity is calculated with the feeder voltage V_L , the required capacity S_C of the three-leg PWM rectifier is expressed as

$$S_C = \frac{V_L \left(\sum_{n=1}^3 I_{SCn} \right)}{2 V_L \cdot I_A} \text{ [pu]}, \dots \dots (26)$$

where I_{SCn} is the RMS value of the output currents of the three-leg PWM rectifier, and I_A is the RMS value of the rating current for each feeder. Thus, the authors propose a current capacity for the PWM rectifier with a three-leg topology. A current capacity A_C for a three-leg PWM rectifier is defined

Fig. 8. Simulation waveforms for Fig. 1 during battery discharging operation

Fig. 9. Simulation waveforms for Fig. 1 with the load variation of Load1 during battery discharging operation

as

$$A_C = \frac{\sum_{n=1}^3 I_{SCn}}{2I_A} \text{ [pu]}. \dots \dots \dots (27)$$

From Figs. 7 and 8, A_C is 0.71 pu. On the other hand, in the simulation results in Ref. (4), the required current capacity of the smart charger is 0.97 pu. We can thus reduce the capacity of the smart charger by 27% using the proposed control algorithm with reactive power control on the source side in comparison to using the previously proposed control algorithm.

Figure 9 shows simulation results for Fig. 1 with the load variation of Load1 during the battery discharging operation. Load2 is kept constant, and Load1 is varied from 1.2 pu to 0.53 pu with an unbalanced ratio of 40%. Before and after Load1 was varied, the source currents i_{S1} and i_{S2} are balanced with a power factor of 0.9 during the battery discharging operation. The dc-capacitor voltage fluctuation ΔV_{DC} is less

Fig. 10. Block diagram of constructed experimental model for smart charger in Fig. 1 with proposed control strategy.

than 15 V. Thus, using a dc capacitor C_{DC} of 2200 μF is sufficient to ensure the battery current ripple, which should be less than 5%.

5. Experimental Results

It is difficult to construct an experimental model with the actual voltage rating of the pole-mounted distribution transformer for the smart charger of Fig. 1 in the laboratory. A reduced-scale experimental model was thus constructed and tested to demonstrate the validity and high applicability of the proposed control method, which involves reactive power control. Figure 10 shows a block diagram of the constructed prototype experimental model. Figure 11 shows the battery model for Fig. 10. The rating of the pole-mounted distribution transformer is 210 V_{rms}, 5.0 kVA, and 60 Hz on the primary side, and 105 V_{rms} and 24 A_{rms} on the secondary side. The charged power is consumed by a 60 Ω resistor R , which is connected in parallel to capacitor C_{F2} , during the battery charging operation. For the battery discharging operation, a dc supply power (Takasago: HX0300-25) with an internal resistor r of 72 m Ω is connected to C_{F2} . The maximum voltage of HX0300-25 is 300 V. In the experimental results of Figs. 13 and 14, V_{bat} was 300 V as shown in Fig. 11(b). The detected primary-side voltage v_S , load currents i_{L1} and i_{L2} , charger currents i_1 and i_2 , and dc-capacitor voltage v_{DC} are fed into a digital signal processor (DSP) (TMS320C6713, 225 MHz) through 12-b A/D converters, where the sampling time T_S is 83.3 μs . In the DSP, the reference values i_1^* , i_2^* , and i_3^* for the three-leg PWM rectifier, which acts as a power quality compensator, are calculated with Eq. (13). The battery voltage v_{Cr2} and battery current i_{LS2} of the bidirectional dc-dc converter are also fed into the DSP through 12-b A/D converters. The feedback control in d - q coordinates is used for the single-phase circuits in Fig. 4. This current feedback control was carried out in the DSP. $K_P = 0.3$ and $T_I = 0.02$ s were used in the PI controller for constant dc-capacitor voltage control in the experiment. $K_P = 0.05$ and $T_I = 8$ ms were used in the PI controller for current feedback control in d - q coordinates for the single-phase circuits.

Figure 12 shows the experimental results for the experimental model in Fig. 10, where the proposed smart charger

Fig. 11. Battery for EV with the experimental setup shown in Fig. 10 during each battery operation

Fig. 12. Experimental waveforms for experimental model in Fig. 10 during battery charging operation

was under a battery charging operation with constant battery current control. Although the load currents i_{L1} and i_{L2} are unbalanced, the source currents i_{S1} and i_{S2} are balanced with a power factor of 0.9 during the charging of the batteries in the EV. The ripple of i_{bat} is 2.38%. The THDs of i_{S1} and i_{S2} calculated from Fig. 10 are 1.68% and 1.66%.

Figure 13 shows the experimental results for the experimental model in Fig. 10, where the proposed smart charger was under a battery discharging operation with constant battery current control. Although the load currents i_{L1} and i_{L2} are unbalanced, the source currents i_{S1} and i_{S2} are balanced with a power factor of 0.9 during the discharging of the batteries in the EV. The ripple of i_{bat} is 3.36%. The THDs of i_{S1} and i_{S2} are 4.50%, and 4.31%, respectively. In Fig. 13, A_C is 0.66 pu. In the experimental results in Ref. (4), the capacity of the smart charger is 0.96 pu. Therefore, controlling the power factor to 0.9 on the source side can reduce the capacity of the smart charger with the newly proposed control algorithm by 31% in comparison to that of the smart charger with the previously proposed control algorithm⁽⁴⁾.

Figure 14 shows experimental results for the experimental model in Fig. 10 with the load variation of Load1 during the battery discharging operation. Load2 is kept constant, and Load1 is varied from 1.2 pu to 0.53 pu with an unbalanced ratio of 40%. Before and after Load1 was varied, the source currents i_{S1} and i_{S2} are balanced with a power factor of 0.9 during the battery discharging operation. The dc-capacitor voltage fluctuation ΔV_{DC} is less than 15 V. Thus, using a dc capacitor C_{DC} of 2200 μF is sufficient to ensure the battery current ripple, which should be less than 5%. These experimental results in Figs. 12, 13, and 14 agree well with the

Fig. 13. Experimental waveforms for the experimental model in Fig. 10 during battery discharging operation

Fig. 14. Experimental waveforms for the experimental model in Fig. 10 with the load variation of Load1 during battery discharging operation

simulation results in Figs. 7, 8, and 9.

Note that the charging or discharging power in both the simulation and experimental results is the same as that in Ref. (4). Thus, reducing the capacity of the smart charger does not affect the battery charging and discharging capabilities.

6. Conclusion

In this paper, we have proposed a new control algorithm for reducing the capacity of the previously proposed smart charger with reactive power control. The smart charger consists of four-leg structured IGBTs. The three-leg structure topology is used for a single-phase PWM rectifier, where two legs are connected to each feeder and the third leg is connected to the neutral line. This PWM rectifier can compensate for reactive and unbalanced active currents on single-phase three-wire distribution feeders. The fourth leg is used as a bidirectional dc-dc converter for battery charging and discharging operations. The authors have shown theoretically

that controlling the reactive power on the source side can reduce the capacity of the previously proposed smart charger. The basic principle of the proposed control algorithm was discussed in detail and was then verified with a digital computer simulation using PSIM software. A reduced-scale experimental model was also constructed and tested. The experimental results demonstrated that balanced source currents with a power factor of 0.9 are achieved on the secondary side of the single-phase three-wire feeders during both battery charging and discharging operations in EVs. This reduces the capacity of the smart charger with the newly proposed control algorithm by 31% in comparison to that of the smart charger with the previously proposed control algorithm. When EVs are not connected to the proposed smart charger, the power quality is also compensated on the secondary side of the pole-mounted distribution transformer. Reducing the capacity of the smart charger does not affect the battery charging and discharging operations as shown in both the simulation and experimental results. Thus, it is concluded that the proposed charger with the new control algorithm is useful for practical applications.

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