Discrete-Time Repetitive Control of Flyback CCM Inverter for PV Power Applications

Sung-Ho Lee, Woo-Jun Cha, Bong-Hwan Kwon, Member, IEEE, and Minsung Kim, Member, IEEE

Abstract—In this paper, a discrete-time repetitive controller (RC) is proposed for flyback inverter operating in continuous conduction mode, which has simple structure, low cost, and high efficiency. Conventional controller results in poor control performance due to the effect of the right-half-plane zero in CCM operation. To achieve the accurate tracking performance and disturbance rejection, the repetitive controller is developed and applied to flyback inverter in CCM operation. In the RC scheme, a low-pass filter is used to allow tracking and rejection of periodic signals within a specified frequency range. A phase lead compensator is also used to compensate for the system delay caused by digital implementation. The stability of the closed-loop system is derived and the zero tracking error is achieved. Numerical simulations validated the proposed control approach, and experimental tests using a 200-W digitally-controlled module integrated converter prototype confirmed its feasibility.

Index Terms—Module-integrated converter, right-half-plane zero, low-pass filter, phase-lead compensation.

I. INTRODUCTION

RECENTLY, an AC photovoltaic (PV) module system has been of interest as a trend for future PV power systems [1, 2]. Unlike conventional centralized PV systems, the AC PV module system provides independent operation for each individual PV module; this design allows all modules to operate at their maximum power point (MPP) and reduces power losses caused by PV module mismatch and partial shading [3, 4]. Moreover, the AC PV module system is more reliable and easier to maintain than are centralized PV systems [5].

In the AC PV module system, a module-integrated converter (MIC) is attached to a single PV panel to ensure maximum power extract and to provide the power to the utility grid. The acceptability of the MIC is evaluated by conversion efficiency, harmonic of output current, cost, and reliability. To meet these requirements, a single-stage flyback inverter topology with an unfolding circuit has been widely used due to its small number of components and its potential for high efficiency and reliability [6, 7].

Operation modes of the flyback inverter can be classified into discontinuous conduction mode (DCM) and continuous conduction mode (CCM). In flyback inverters that operate under DCM [3], [8–10], the transfer function from control input to output current is a constant; thus, the output current control can be controlled easily and simply. However, the flyback inverter in DCM imposes high current stress on devices; it results in decrease of conversion efficiency. In addition, the filter design becomes difficult due to the high current ripple. Compared to the DCM flyback inverters, the CCM flyback inverter has many advantages such as higher efficiency with lower current stress, easier filter design, and lower electromagnetic interference (EMI). However, in a CCM flyback inverter, the transfer function from control input to output current has a right-half-plane (RHP) zero which destabilizes the feedback control loop [11]. Moreover, the RHP zero location depends on the operating point, which varies in a wide range in the case of PV MIC applications. Thus, the controller should accommodate the minimum RHP zero; this requirement causes the degradation of the controller bandwidth and difficulty in controlling the output current [12]. This characteristic limits the use of CCM flyback inverters in practice despite their numerous merits.

To avoid this problem, some studies [11–13] used feedback control of the primary current; this approach is an open-loop control of the output current. In this case, the control approach bypasses the difficulties posed by the RHP zero. However, total harmonic distortion (THD) of the output current is high, because this approach controls the output current indirectly.

The repetitive controller is widely used as an effective solution to track a periodic signal and to eliminate periodic disturbance in dynamic systems [14–16]. In the repetitive control scheme, the present control input is obtained using knowledge obtained in the previous period, and this operation is performed repetitively. Hence, the repetitive controller generates the control input that ensures zero tracking error [19].

In this paper, the feedback of the output current and the use of repetitive control algorithm are proposed for a flyback inverter that operates in CCM. This direct control of the output current can ensure the stability of a closed-loop system more clearly, and does not need the additional sensor to measure the primary current. The conventional controller, which consists of a linear controller and a feedforward controller, results in poor control accuracy due to the RHP zero which limits the available controller bandwidth. To ensure good tracking ability and disturbance rejection, the repetitive controller is
During turn-on subinterval:

\[
\dot{x}(t) = \begin{bmatrix}
0 & -\frac{1}{L_m} & 0 & 0 \\
-\frac{1}{L_m} + \frac{R_{pf}C_{in}}{L_m} & 0 & 0 & 0 \\
0 & 0 & -\frac{R_{c_f} + R_f}{L_f} & \frac{1}{L_f} \\
0 & 0 & \frac{1}{C_f} & 0
\end{bmatrix} x(t) + \begin{bmatrix} 0 \\
\frac{1}{R_{pf}C_{in}} \\
0 \\
0
\end{bmatrix} v(t),
\]

\[
y(t) = [0 \ 0 \ 1 \ 0] x(t),
\]

During turn-off subinterval:

\[
\dot{x}(t) = \begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
\frac{R_{c_f}}{nL_m} & 0 & 0 & 0 \\
0 & -\frac{1}{C_f} & 0 & 0
\end{bmatrix} x(t) + \begin{bmatrix} 0 \\
\frac{1}{R_{pf}C_{in}} \\
0 \\
0
\end{bmatrix} v(t),
\]

\[
y(t) = [0 \ 0 \ 1 \ 0] x(t),
\]

where \(i_Lm\) is the current through the magnetizing inductor, \(v_{C_{in}}\) is voltage across the capacitor, \(i_{L_f}\) is the current through the output inductor, \(v_{C_f}\) is the voltage across the output capacitor. \(n\) is the secondary to primary turns ratio of the transformer.

Combining (1)–(4) using the state-space averaging method, the averaged model can be developed as in [1]

\[
\dot{x}(t) = \begin{bmatrix}
\frac{R_{c_f}}{n^2L_m} & 0 & 0 & -\frac{1}{nL_m} \\
0 & 0 & \frac{R_{c_f}}{nL_m} & 0 \\
\frac{R_{c_f}}{nL_f} & 0 & 0 & -\frac{1}{L_f} \\
0 & -\frac{1}{C_f} & 0 & 0
\end{bmatrix} x(t) + \begin{bmatrix} 0 \\
\frac{1}{R_{pf}C_{in}} \\
0 \\
0
\end{bmatrix} v(t),
\]

\[
y(t) = [0 \ 0 \ 1 \ 0] x(t),
\]

where \(D_c(t)\) is the control duty cycle.

\[D_c(t) = \begin{bmatrix} 0 & 0 \ \ \ 0 \\
\frac{1}{R_{pf}C_{in}} & 0 \\
0 & -\frac{1}{L_f} \\
0 & 0
\end{bmatrix} v(t),
\]

where \(D_c(t)\) is the control duty cycle.

B. Problem formulation

By linearizing the averaged model, the small signal model can be derived as follows:

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where the parameters $a_j$ and $b_j$ for all $j = 1, ..., 4$ are defined in the Appendix A. By analyzing the small-signal model, the system has one RHP zero, two left-half-plane (LHP) zero, and four LHP poles.

In the following section, the controller will be developed in discrete-time domain. To get a better connection between the flyback inverter system and the controller, the small-signal model needs to be represented as the discrete-time transfer function. Using backward difference method with sampling period of $T_s$, it can be described as:

$$ G_{id}(z) = \frac{e_1 z^4 + e_2 z^3 + e_3 z^2 + e_4 z}{z^4 + f_1 z^3 + f_2 z^2 + f_3 z + f_4}, \quad (8) $$

where the parameters $e_j$ and $f_j$ for all $j = 1, ..., 4$ are defined in the Appendix A.

### III. Controller Design

#### A. Conventional controller

The operating condition of the flyback inverter in CCM changes slowly during a grid period compared to a switching period; the flyback inverter can be assumed to operate in quasi-steady state in each instant of the grid period. By assuming quasi steady-state operation and using volt-second balance for the magnetizing inductance $L_m$ over one switching period $T_s$, the nominal duty ratio $D_n(t)$ for flyback inverter in CCM can be obtained as [1]

$$ D_n(t) = \frac{|v_g(t)|}{|v_g(t)| + a V_{po}(t)}. \quad (9) $$

Even though $D_n(t)$ does not directly determine the output current, the use of the nominal duty helps the flyback inverter in CCM to generate the desired output current by alleviating the effect of the disturbances. Thus, it reduces the burden on the state-feedback controller, and so makes the overall control system easily achieve the fast response.

Since $D_n(t)$ has been used to generate the desired output current by alleviating the effect of disturbances such as the grid voltage, a state feedback controller can be utilized to improve the control quality with suppressing the remaining disturbances. The transfer function of a state feedback controller is represented as

$$ C_{fs}(z) = k_p + k_i \frac{T_s}{1 - z^{-1}}. \quad (10) $$

where $k_p$ is the proportional controller gain, $k_i$ is the integral controller gain.

However, in flyback inverter under the operating mode of CCM, the nominal duty plus the state feedback controller cannot satisfy the desired control performance, because the transfer function of the flyback inverter in CCM operation has a RHP zero, which was obtained from small signal model (7). In this conventional control scheme, the RHP zero limits the available controller bandwidth. Moreover, the controller should accommodate the minimum RHP zero; it causes the degradation of the available controller bandwidth and so difficulty in controlling the output current.

Fig. 3 shows the Bode plot of the compensated system when the state-feedback controller is used (blue); when the high-gain state-feedback controller is used (red). The dot in each magnitude Bode plot denotes the gain margin of the compensated system. The dot in each phase Bode plot denotes the phase margin of the compensated system.

#### B. Repetitive controller with low-pass filter

The repetitive controller handles systems that performs the same task repetitively. In repetitive control scheme, the knowledge obtained from the previous trial is used to improve the control input for the next trial. Hence, the control input in each trial is adjusted using the tracking error signals obtained from the previous trial, and theoretically achieves zero tracking error [14, 15]. The concept of repetitive controller is shown in Fig. 4.

In the conventional control scheme, the RHP zero limits the available controller bandwidth. To achieve fast dynamical response to input signal at a specific frequency, the repetitive controller is developed. To allow tracking/rejecting periodic signals within a specified frequency range, a low-pass filter is adopted in the repetitive control scheme. Moreover, to compensate for the system delay which comes from digital implementation, the phase lead compensator is used to the repetitive controller scheme.
The transfer function of the repetitive controller is

$$C_{rc}(z) = \frac{k_r}{1 - z^{-N}G(z)}G_{inv}(z),$$

(11)

where $k_r$ is the repetitive controller gain, $N = f_s/f_g$ with $f_s = 1/T_s$ being the sampling frequency and $f_g$ being the frequency of reference, $Q(z)$ is the low-pass filter, and $G_{inv}(z)$ is the inverse function of the system.

For tracking/rejecting periodic signals within a specified frequency range and easy implementation, $Q(z)$ is chosen as moving average filter with zero phase shift:

$$Q(z) = \sum_{i=0}^{p} \alpha_i z^{-i} + \sum_{i=1}^{\infty} \alpha_i z^{-i},$$

(12)

where $\alpha_0 + 2\sum_{i=1}^{p} \alpha_i = 1$ with $\alpha_i > 0$, $p$ is the number of samples to be used for filtering. Here, a first order filter $Q(z) = \alpha_1 z + \alpha_0 + \alpha_1 z^{-1}$ is used; as $\alpha_0$ becomes larger, the cutoff frequency of $Q(z)$ becomes higher.

Due to the system uncertainty, the system delay, and the unknown disturbance, $G_{inv}(z)$ cannot be precisely implemented. Here, to compensate even the effect of the system delay, $G_{inv}$ are set as

$$G_{inv}(z) = z^{-m}.$$

(13)

where $m$ is the prediction index.

With the complete repetitive controller, the overall control system is shown in Fig. 5. This control scheme consists of three components: the linear feedback control term that forces the closed-loop system to stay within uniform bound; the nominal duty term that generates the desired output current by alleviating the disturbances; the repetitive controller term that enhances the command input response with suppressing the residual disturbances.

Fig. 6 shows the Bode plot of the compensated system when the repetitive controller is used. Used parameters are listed in Table I. The repetitive controller increases the gain around $w_l = 2\pi f_g$, with $l = 0, 1, 2, ..., L$ without decreasing the phase margin. Thus, it enhances the tracking and disturbance rejection performance. The resonant peak and sharp phase drop at a frequency of 8kHz are introduced by capacitor-inductor filter. It also restricts the controller bandwidth. In grid voltage disturbance, fundamental and low-order harmonic components are dominant. To avoid the control at a high-frequency and suppress fundamental and low-order harmonics components, the low-pass filter is adopted in the repetitive control scheme. It disables the learning at high-frequency. The resulting repetitive controller amplifies the system gain only within a low frequency region so that it improves the output current tracking performance while suppressing the grid voltage disturbance at a fundamental and low-order harmonic frequencies.

**C. Stability analysis**

In the repetitive control system shown in Fig. 5, the input-output relationship of the flyback inverter in CCM is described as

$$y(z) = G(z)y_d(z) + S(z)v(z),$$

(14)

where $G(z)y_d(z)$ is the command input response and $S(z)v(z)$ is the disturbance response; $G(z)$ is the transfer function from reference to output and $S(z)$ is the sensitivity function; $G(z)$ and $S(z)$ are represented as

$$G(z) = \frac{y(z)}{y_d(z)} = \frac{[1 - Q(z)z^{-N}(1 - k_r z^m)]G_c(z)}{1 - Q(z)z^{-N}(1 - k_r z^m)G_c(z)},$$

(15)

$$S(z) = \frac{y(z)}{v(z)} = \frac{[1 - Q(z)z^{-N}(1 - k_r z^m)]}{1 - Q(z)z^{-N}(1 - k_r z^m)G_c(z)}.$$

(16)
where \( G_c(z) = \frac{G_{fb}(z)G_{id}(z)}{1 + G_{fb}(z)G_{id}(z)} \). Detailed derivations of (15) and (16) are shown in the Appendix B for the sake of completeness.

Defining \( w(z) = y_{2d}(z) - y(z) \) and from (15) and (16), the relationship between the error \( e(z) = y(z) - y_{2d}(z) \) and the reference \( y_{2d}(z) \) and disturbance \( v(z) \) can be obtained as

\[
e(z) = \frac{(1 - Q(z)z^{-N})(1 - G_c(z))}{1 - Q(z)z^{-N}(1 - k_r z^m G_c(z))},
\]

(17)

The transfer function \( e(z)/w(z) \) can be written as three systems connected in cascade. The term \( 1 - Q(z)z^{-N} \) is a low-pass filter and a time delay, and so it is stable. The term \( 1 - G_c(z) \) which has the same roots as \( G_c(z) \). Thus, it can be stable by designing the linear controller. Finally, the term \( 1 - Q(z)z^{-N}(1 - k_r z^m G_c(z)) \) can be described as a positive-feedback closed-loop system with the term \( Q(z)z^{-N}(1 - k_r z^m G_c(z)) \) in the feedback path. A sufficient condition for the stability of the system with the term \( 1 - Q(z)z^{-N}(1 - k_r z^m G_c(z)) \) is

\[
|Q(z)z^{-N}(1 - k_r z^m G_c(z))| \\
\leq |z^{-N}| \cdot |Q(z)(1 - k_r z^m G_c(z))| \\
\leq |Q(z)(1 - k_r z^m G_c(z))| < 1
\]

\( \forall z = e^{jwT_s}, 0 < w < \frac{\pi}{T_s} \)

(18)

As a result, if the overall control system holds the following stability conditions

1) The closed-loop system \( G_c(z) \) without repetitive controller is stable;
2) \( |Q(z)(1 - k_r z^m G_c(z))| < 1, \quad \forall z = e^{jwT_s}, 0 < w < \frac{\pi}{T_s} \)

then, the system is stable [16, 17]. Moreover, if the frequency of the reference \( y_{2d}(t) \) and disturbance \( v(t) \) approaches to \( w_l = 2\pi f_p \), with \( l = 0, 1, 2, ..., L \) \((L = N/2 \) for even \( N \) and \( L = (N - 1)/2 \) for odd \( N \)), then \( z^{-N} = 1 \). Assuming \(|Q(z)| = 1 \) and \( \angle Q(z) = 0 \), from (15) and (16), we have

\[
\lim_{w_l \to w} |G(z)|| = 1, \quad \forall w_l
\]

(19)

\[
\lim_{w_l \to w} |S(z)|| = 0, \quad \forall w_l
\]

(20)

Equations (19) and (20) indicate that, for any periodic reference and disturbance, a zero steady-state tracking error and disturbance rejection can be guaranteed [19].

### IV. SIMULATION AND EXPERIMENTAL RESULTS

To demonstrate the feasibility of the proposed controller, simulation using a simulator Psim and experiment using the prototype were conducted for the flyback inverter operated in CCM. The flyback inverter in CCM was designed for the following specifications: input voltage \( V_{pv} = 60 \) V, grid voltage \( v_g = 220 \) Vrms, and rated output power \( P_o = 200 \) W. The major parameters for proposed inverter are listed in Table I. The total system configuration is shown in Fig. 7. The flyback inverter in CCM is controlled by a dSPIC35EP512GM604 digital signal controller. The controller is also implemented with other functions such as phase-locked loop (PLL) and over current protection.

The gains of the state feedback controller (10) and repetitive controller (11) are set as follows: the gains of proportional-integral controller \( k_p = 0.1 \) and \( k_i = 0 \), the gain of repetitive controller \( k_r = 0.02 \), the gains of low-pass filter \( \alpha_0 = 0.5 \) and \( \alpha_1 = 0.25 \), the prediction index \( m = 1 \). The repetitive control gain \( k_r \) is chosen to satisfy the convergence condition \(|Q(z)(1 - k_r z^m G_c(z))| < 1 \).

#### A. Simulation results

Fig. 8(a) shows waveforms of the reference and output currents when conventional control scheme with linear controller plus feedforward controller is applied. Linear controller \((k_p = 0.1) \) plus feedforward controller is applied for the first 0.05 s. Afterward, linear controller \((k_p = 0.13) \) plus feedforward controller is applied. The control scheme shows poor tracking ability due to the effect of RHP zero in CCM operation. Fig. 8(b) shows waveforms of the reference and output current when proposed control scheme is applied. As the learning operation of the proposed controller proceeds, the output current gradually tracks the reference current, and
finally the tracking error becomes almost zero.

B. Experimental results

Fig. 9 shows waveforms of the grid voltage $v_g$ and output current $i_L$ at quarter, half, and full loads conditions when the proposed control scheme is applied. The output current is perfectly sinusoidal and has desired power level. The total harmonic distortion (THD) of the output current is measured less than 2.5% under full load condition. Fig. 10 represents measured efficiency according to load conditions. The maximum efficiency is 96.3%.

V. Conclusion

In this paper, a repetitive controller is proposed for flyback inverter operating in CCM, which has simple structure, low cost, and high efficiency. Conventional controller results in poor tracking due to the effect of RHP zero in CCM operation. To achieve a fast dynamical response, a repetitive controller is developed and applied to the flyback inverter in CCM operation. The low pass filter is adopted to allow tracking/rejecting periodic signals within a specified frequency range. The phase lead compensator is used to compensate for the system delay which comes from digital implementation. The stability of closed-loop system is derived and the zero tracking error is achieved. The effect of RHP zero is analyzed. Comparison of the accuracy of linear controller plus feedforward controller and proposed controller in the simulation and experimental results verify the validity of the proposed control scheme.

APPENDIX A

THE VALUES OF SYSTEM PARAMETERS

The system parameters are shown in the following equations as in [1]:

\[
an_1 = -\frac{R_{C_f} I_{L_m}}{n I_f}
\]

\[
an_2 = -\frac{I_{L_m}}{C_{in} D_1 I_{L_f} R_{C_f} R_{pv} V_{C_i n}} \left[ \frac{n (I_{L_m} L_m R_{C_f} - C_{in} D_1 R_{C_f} R_{pv} V_{C_i n})}{C_{in} L_f L_m R_{pv} n^2} \right]
\]

\[
a_2 = \frac{C_{in} D_1 I_{L_f} R_{C_f} R_{pv} - C_{in} D_1 R_{C_f} R_{pv} V_{C_i n}}{C_{in} L_f L_m R_{pv} n^2}
\]
Appendix B
Derivations for $G(z)$ and $S(z)$

Appendix provides the details of derivations for the transfer function from reference to output $G(z)$ and sensitivity function $S(z)$. First, the transfer function from reference to output $G(z)$ is to be derived. From Fig. 5, the following relationship can be obtained

$$y(z) = e(z) \left( 1 + k_r \frac{Q(z)z^{-N}}{1 - Q(z)z^{-N}G_{inv}(z)} \right) C_{fb}(z)G_{id}(z), \quad (37)$$

$$e(z) = y_d(z) - y(z). \quad (38)$$

Substituting (38) to (37) and moving $y(z)$-related term to the left-hand side of (37), we have

$$y(z) \left( 1 + \frac{Q(z)z^{-N}}{1 - Q(z)z^{-N}G_{inv}(z)} \right) C_{fb}(z)G_{id}(z)$$

$$= y_d(z) \left( k_r \frac{Q(z)z^{-N}}{1 - Q(z)z^{-N}G_{inv}(z)} \right) C_{fb}(z)G_{id}(z). \quad (39)$$

Multiplying $1 - Q(z)z^{-N}$ on the both side of (39) and substituting (13) to (39), we obtain

$$y(z)/(1 - Q(z)z^{-N}) + (1 - Q(z)z^{-N} + k_r z^{-m} Q(z)z^{-N}) \cdot C_{fb}(z)G_{id}(z)$$

$$= y_d(z) \left( 1 - Q(z)z^{-N} + k_r z^{-m} Q(z)z^{-N} \right) C_{fb}(z)G_{id}(z). \quad (40)$$

Rearranging (40), we have

$$G(z) = \frac{y(z)}{y_d(z)}$$

$$= \frac{(1 - Q(z)z^{-N})}{1 - Q(z)z^{-N} + k_r z^{-m} Q(z)z^{-N} \cdot C_{fb}(z)G_{id}(z)}$$

$$= \frac{1 - Q(z)z^{-N} \cdot C_{fb}(z)G_{id}(z)}{1 - Q(z)z^{-N} \cdot C_{fb}(z)G_{id}(z) - \left( k_r z^{-m} Q(z)z^{-N} \cdot C_{fb}(z)G_{id}(z) \right)}$$

$$\frac{1}{\left( 1 - k_r z^{-m} C_{fb}(z)G_{id}(z) \right)} \left( \frac{C_{fb}(z)G_{id}(z)}{-C_{fb}(z)G_{id}(z)} \right)$$

$$= \frac{1}{1 - k_r z^{-m} C_{fb}(z)G_{id}(z)}.$$  

Sensitivity function $S(z)$ can be derived in the same manner of the derivation for $G(z)$. From Fig. 5, the following relationship can also be obtained

$$y(z) = e(z) \left( 1 + k_r \frac{Q(z)z^{-N}}{1 - Q(z)z^{-N}G_{inv}(z)} \right) C_{fb}(z)G_{id}(z), \quad (42)$$

$$e(z) = -y(z). \quad (43)$$

Substituting (43) to (42) and moving $y(z)$-related term to the left-hand side of (37), we have
\[ y(z) \left( 1 + \frac{k_r Q(z)z^{-N}}{1 - Q(z)z^{-N}G_{inc}(z)} \right) C_{fb}(z)G_{id}(z) = v(z). \] (44)

Multiplying \( 1 - Q(z)z^{-N} \) on the both side of (44) and substituting (13) to (44), we obtain

\[ y(z)(1 - Q(z)z^{-N} + (1 - Q(z)z^{-N} + k_r Q(z)z^{-N} z^m) \cdot C_{fb}(z)G_{id}(z)) = (1 - Q(z)z^{-N})v(z). \] (45)

Rearranging (45), we have

\[ S(z) = \frac{y(z)}{v(z)} = \frac{(1 - Q(z)z^{-N})}{1 - Q(z)z^{-N} + (1 - Q(z)z^{-N} + k_r Q(z)z^{-N} z^m) C_{fb}(z)G_{id}(z)} \]

\[ = \frac{(1 - Q(z)z^{-N})}{1 - Q(z)z^{-N} (1 + C_{fb}(z)G_{id}(z))} \frac{1 + C_{fb}(z)G_{id}(z) - Q(z)z^{-N} (1 + C_{fb}(z)G_{id}(z)) - k_r z^m C_{fb}(z)G_{id}(z)}{1 + C_{fb}(z)G_{id}(z)} \]

\[ = \frac{(1 - Q(z)z^{-N})}{1 - Q(z)z^{-N} (1 - k_r z^m C_{fb}(z)G_{id}(z))}. \] (46)

Finally, \( G(z) \) and \( S(z) \) can be obtained from (41) and (46).

REFERENCES


Minsung Kim (M’14) was born in Ulsan, Korea, in 1986. He received the B.S. degree in electrical engineering from Pohang University of Science and Technology (POSTECH), Pohang, Korea, in 2004, and the Ph.D. degree in electrical engineering from POSTECH, Pohang, Korea, in 2013. Since 2013, he has been with Future IT Research Laboratory, POSTECH, Pohang, Korea, where he is currently senior researcher. His current research interests include renewable energy system, nonlinear system analysis, and controller design for industrial process.