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A Voltage-Sensorless Controller for Grid Converters

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Abstract—Grid converters need information about the point-of-common-coupling voltage for their control. The voltage is typically measured for current and ac-bus voltage control, and for synchronization. In order to increase reliability by decreasing dependence on this measurement, this paper proposes a multifunctional grid-voltage sensorless controller that can operate either in current- or voltage-control mode. The control system is designed to be robust against variations in grid strength, covering strong and very weak grid conditions. The controller is experimentally tested and its robustness is verified in these conditions. The results show that sensorless ac-bus voltage control can provide grid voltage support.

Index Terms—Estimation, grid-connected converter, grid-supporting control, sensorless control, voltage-source converters.

I. INTRODUCTION

Energy flow through grid-connected power converters, grid-following as well as grid-forming, is enabled by power, current, and voltage control algorithms. Certain quantities that are inputs to these algorithms can be estimated. A possible quantity is the grid voltage [1]–[17]. Its estimation can be based on instantaneous power theory [1], [2], virtual flux models [3]–[5], internal states of a current controller [6], [7], model-based observers [8]–[16] or neural networks [17]. Besides the fundamental-frequency positive-sequence component, negative sequence [4], [5], [10]–[16] or other harmonic components [14]–[16] of the voltage can be estimated.

The voltage estimation is motivated, e.g., by improved immunity to electrical noise and cost savings due to reduced number of sensors [1], [8]. Grid codes and standards for power generating plants typically specify a voltage range when the converter can be connected to the grid at the point of common coupling (PCC). Hence, the grid-voltage measurement may be compulsory. However, even in such a situation, a voltage estimator in parallel with the measurement increases reliability, e.g., in the case of sensor faults or noise. The parallel voltage estimator may also provide cost savings, if it enables the voltage measurement with a lower resolution or lower bandwidth.

The majority of the grid-voltage sensorless control methods have been developed for ac-bus current, power, and dc-bus voltage control, without ability to control ac-bus voltage [1]–[19]. Certain other control methods are inherently grid-voltage sensorless, i.e., the PCC voltage is not used in the control algorithm. One example is [20], where an islanded system and grid-forming control based on voltage and frequency droop is

Fig. 1. Grid-voltage sensorless control system.

considered. Another example is power-synchronization control (PSC) [21]. PSC can operate without the grid-voltage sensors and control the ac-bus voltage at the converter terminals [21] or at the PCC [22] when the converter current magnitude is below its maximum permissible value. However, when the current is in danger of exceeding this limit, e.g., in fault situations, the measured PCC voltage is used to accomplish the current limitation in [21], [22]. Since only a few methods [20], [21] are capable to control ac-bus voltage without grid-voltage sensors, the field is open for further research.

This paper proposes a grid-voltage sensorless controller that can operate either in ac-bus voltage or current (or power) control mode. The main contribution is in the development of the voltage-sensorless voltage-control mode using a cascaded current controller. Compared to [20], [21], the proposed controller is not power-synchronization or droop based. To parametrize the proposed controller, the robust design recently presented in [22], for a voltage- and current-controlled converter with the grid-voltage sensors, is modified to voltage sensorless. The presented robust design enables stable operation in strong as well as weak grids, which is experimentally verified. Grid-voltage support capability of sensorless ac-bus voltage control is also demonstrated.

II. SYSTEM MODEL

Fig. 1 shows the block diagram of a grid-voltage sensorless control system. The system modeling is carried out using complex space vectors that are marked with boldface letters, e.g., the converter current is $\mathbf{i}_c = i_{cd} + j i_{cq}$. To distinguish different reference frames, a vector with superscript s refers to

Fig. 2. Reduced-order grid-voltage estimator.

the stationary $\alpha\beta$ reference frame and a corresponding vector in the synchronous dq reference frame is without a superscript.

The dynamics of the converter current i_c^s are given by

$$L_f \frac{di_c^s}{dt} = u_c^s - R_f i_c^s - u_g^s \quad (1)$$

where u_c^s is the converter voltage, u_g^s is the grid voltage at the PCC, and L_f and R_f are the inductance and resistance of a filter between the converter and PCC, respectively. The converter voltage is assumed to equal its delayed reference

$$u_c^s(t) = u_{c,\text{ref}}^s(t - T_d) \quad (2)$$

where T_d models the system delay including one sampling period T_s delay and additional $0.5T_s$ delay originating from the computation and zero-order-hold nature of the pulse-width modulation (PWM), respectively.

It is considered that the grid voltage u_g^s is an unmeasured input to the system, but has the angular frequency of ω_g . Therefore, its dynamics are modeled as¹

$$\frac{du_g^s}{dt} = j\omega_g u_g^s. \quad (3)$$

III. CONTROL SYSTEM

The control functions of the proposed grid-voltage sensorless control system, shown in Fig. 1, are described below. The controllers operate in the synchronous reference frame tied to the estimated grid voltage.

A. Grid-Voltage Estimator

A reduced-order estimator for the unmeasured grid voltage is based on the model (3) as

$$\frac{d\hat{u}_g^s}{dt} = j\omega_g \hat{u}_g^s + k_o \tilde{u}_g^s \quad (4)$$

where $\hat{u}_g^s$ is the estimated grid voltage, k_o is the estimator gain, and $\tilde{u}_g^s$ is the estimation error obtained from (1) as

$$\tilde{u}_g^s = u_c^s - \hat{L}_f \frac{di_c^s}{dt} - \hat{R}_f i_c^s - \hat{u}_g^s \quad (5)$$

where $\hat{L}_f$ and $\hat{R}_f$ are the system parameter estimates. Hence, the proposed voltage estimator, comprising (2), (4) and (5), requires only inputs from the measured converter current i_c^s and the voltage reference $u_{c,\text{ref}}^s$ as depicted in Figs. 1 and 2. Furthermore, a block diagram of the estimator in Fig. 2 shows

¹Negative sequence component u_{g-}^s can be modeled similarly $du_{g-}^s/dt = -j\omega_g u_{g-}^s$ if it is of interest.

Fig. 3. Current controller.

that the derivative of the current present in (5) can be avoided due the integrator in the estimator.

From a control theory perspective, the estimator structure is a reduced-order model-based observer [23] for the system model comprising (1) and (3). With an additional state for the negative sequence, the reduced-order estimator shown in Fig. 2 can also be extended for positive- and negative-sequence estimation as in [10].

B. PLL

A conventional synchronous-reference-frame phase-locked loop (PLL) calculates the reference-frame angle $\hat{\vartheta}_g$ as

$$\hat{\vartheta}_g = \frac{1}{s} \left[\underbrace{\omega_{gN} + F_p(s) \text{Im} \left\{ \frac{\hat{u}_g}{|\hat{u}_g|} \right\}}_{=\hat{\omega}_g} \right], \quad \hat{u}_g = e^{-j\hat{\vartheta}_g} \hat{u}_g^s \quad (6)$$

where ω_{gN} is the nominal grid angular frequency, $F_p(s)$ is a PLL controller, such as proportional [$F_p(s) = k_p$] or proportional-integral [$F_p(s) = k_p + k_i/s$] controller, and $s = d/dt$. In addition, filtering for voltage harmonics can be implemented in the PLL if needed.

C. Current Controller

The control law of the current controller shown in Fig. 3 is

$$u_{c,\text{ref}} = \hat{R}_f i_{\text{ref}} + R_a (i_{\text{ref}} - i_c) + j\omega_g \hat{L}_f i_c + \hat{u}_g \quad (7)$$

where R_a is the proportional gain, and $\hat{R}_f i_{\text{ref}}$ and $j\omega_g \hat{L}_f i_c$ compensate for the resistive and reactive steady-state voltage drops across the inductive filter, respectively. The estimated voltage in the control law is essentially grid-voltage feedforward with low-pass characteristics. The proportional gain is selected similarly as in [22]

$$R_a = \alpha_c \hat{L}_f - \hat{R}_f \quad (8)$$

where α_c is the desired closed-loop bandwidth of the current-reference tracking. In practice, $\hat{R}_f$ can be set to zero in (8), as it is negligible compared to $\alpha_c \hat{L}_f$.

D. Reference Calculation and Current Limitation

The reference-calculation block shown in Fig. 1 provides the current reference i_{ref} for the current controller, and it is calculated as

$$i_{\text{ref}} = \text{sat}\{i'_{\text{ref}}\}, \quad i'_{\text{ref}} = \underbrace{\frac{2 p_{g,\text{ref}}}{3 u_{g,\text{ref}}}}_{=i_{p,\text{ref}}} + i_{v,\text{ref}} \quad (9)$$

Fig. 4. AC-bus voltage controller.

where $p_{g,\text{ref}}$ is the active power reference, $u_{g,\text{ref}}$ is the PCC voltage reference, and $i_{v,\text{ref}}$ is the output of the ac-bus voltage controller. The function $\text{sat}\{\cdot\}$ denotes saturation and limits the magnitude $|i_{\text{ref}}|$ to its maximum permissible value i_{max} that enables the converter current i_c limitation. The saturation function can be implemented in various ways depending on the priorities of d and q components when limiting the magnitude of i_{ref} . In this paper, the q component $i_{q,\text{ref}} = \text{Im}\{i_{\text{ref}}\}$, related to the reactive power flow, is prioritized. It is first saturated to its maximum value $i_{q,\text{max}} < i_{\text{max}}$ as

$$i_{q,\text{ref}} = \begin{cases} \text{sgn}\{i'_{q,\text{ref}}\} \cdot i_{q,\text{max}} & \text{if } |i'_{q,\text{ref}}| > i_{q,\text{max}} \\ i'_{q,\text{ref}} & \text{otherwise} \end{cases} \quad (10)$$

and then, the d component is limited as

$$i_{d,\text{ref}} = \begin{cases} \text{sgn}\{i'_{d,\text{ref}}\} \sqrt{i_{\text{max}}^2 - i_{q,\text{ref}}^2} & \text{if } i_{d,\text{ref}}^2 + i_{q,\text{ref}}^2 > i_{\text{max}}^2 \\ i'_{d,\text{ref}} & \text{otherwise.} \end{cases} \quad (11)$$

E. Voltage Controller

The ac-bus voltage control law is

$$i_{v,\text{ref}} = G_a(u_{g,\text{ref}} - \hat{u}_g) - jF_v(s)(u_{g,\text{ref}} - \text{Re}\{\hat{u}_g\}) \quad (12)$$

where $F_v(s) = k_v/s$ is an integral controller and the gains are

$$G_a = 1/(R_a + \hat{R}_f) \quad k_v = \omega_g/R_a. \quad (13)$$

In order to avoid integrator windup in $F_v(s)$, the magnitude of the integral state is limited to $i_{q,\text{max}}$ similar to (10). The voltage control law resembles the one proposed in [22]. Differences between (12) and the corresponding controller in [22] are explained in the Appendix.

IV. CLOSED-LOOP SYSTEM

The closed-loop grid-voltage sensorless control system is further analyzed below.

A. Estimator

Let us first analyze the voltage estimator. Substituting u_c^s , solved from (1), into (5) and the resulting estimation error $\tilde{u}_g^s$ into (4), the estimated voltage can be expressed as

$$\hat{u}_g^s = \frac{k_o}{s + k_o - j\omega_g} u_g^s + \frac{k_o(s\tilde{L}_f + \tilde{R}_f)}{s + k_o - j\omega_g} i_c^s \quad (14)$$

where $\tilde{L}_f = L_f - \hat{L}_f$ and $\tilde{R}_f = R_f - \hat{R}_f$ are the inductance and resistance modeling errors, respectively. Moreover, the estimated voltage in synchronous coordinates is

$$\hat{u}_g = \frac{k_o}{s + k_o} u_g + \frac{k_o[(s + j\omega_g)\tilde{L}_f + \tilde{R}_f]}{s + k_o} i_c \quad (15)$$

that in the steady state ($s = 0$) is

$$\hat{u}_g = u_g + (\tilde{R}_f + j\omega_g\tilde{L}_f)i_c. \quad (16)$$

Therefore, the estimated voltage has an error depending on the current operating point if the model-parameter estimates in the estimator are inaccurate. In any case, with the accurate parameter estimates ($\hat{L}_f = L_f$, $\hat{R}_f = R_f$) and with the voltage estimator gain

$$k_o = \alpha_f \quad (17)$$

the voltage estimator becomes a low-pass filter

$$\hat{u}_g = H(s)u_g \quad H(s) = \frac{\alpha_f}{s + \alpha_f} \quad (18)$$

whose bandwidth is α_f . Then, $\hat{u}_g$ does not depend on i_c .

B. Current-Controlled System

In order to keep expressions reasonable, it is assumed that the current-controlled system operates in the linear region and the delays in the control loop can be neglected ($u_c = u_{c,\text{ref}}$). Then, the closed-loop current-controlled system is obtained from (1), (7), and (15) as

$$i_c = G_c(s)i_{\text{ref}} - Y_c(s)u_g \quad (19)$$

where $G_c(s)$ is the reference-tracking transfer function and $Y_c(s)$ is the closed-loop admittance. When the gain k_o is selected according to (17), the transfer functions become

$$G_c(s) = \frac{(\hat{R}_f + R_a)(s + \alpha_f)}{s(L_f + R_a + R_f + j\omega_g\tilde{L}_f) + \alpha_f(s\hat{L}_f + R_a + \hat{R}_f)} \quad (20)$$

$$Y_c(s) = \frac{s}{s(L_f + R_a + R_f + j\omega_g\tilde{L}_f) + \alpha_f(s\hat{L}_f + R_a + \hat{R}_f)}. \quad (21)$$

It is remarkable that even under inaccurate parameter estimates ($\hat{L}_f \neq L_f$ and $\hat{R}_f \neq R_f$), the steady-state gains of the transfer functions (20) and (21) are $G_c(0) = 1$ and $Y_c(0) = 0$, respectively. Therefore, in steady state $i_c = i_{\text{ref}}$. This property originates from the selected current-control structure where the voltage estimator is an input-equivalent disturbance estimator. The disturbance estimator can be seen as an alternative way to implement an integrator in the current-control system.

Furthermore, when the proportional gain R_a is selected according to (8), and $\hat{L}_f = L_f$ and $\hat{R}_f = R_f$, the transfer functions become

$$G_c(s) = \frac{\alpha_c}{s + \alpha_c}, \quad Y_c(s) = \frac{s}{L_f(s + \alpha_f)(s + \alpha_c)}. \quad (22)$$

C. Voltage-Controlled System

When the current reference is not saturated, $i_{\text{ref}} = i'_{\text{ref}}$, a closed-loop system with the voltage controller is obtained from (9), (12), and (19) as

$$i_c = G_c(s)[i_{p,\text{ref}} + G_a(u_{g,\text{ref}} - \hat{u}_g) - jF_v(s)(u_{g,\text{ref}} - \text{Re}\{\hat{u}_g\})] - Y_c(s)u_g \quad (23)$$

TABLE I
NOMINAL PARAMETERS OF A 12.5-kVA CONVERTER SYSTEM

Variable/parameter	Value	Value (p.u.)
Rated power	12.5 kVA	1
Rated voltage (phase-to-neutral, peak)	$\sqrt{2/3} \cdot 400$ V	1
Rated current (per phase, peak)	$\sqrt{2} \cdot 18$ A	1
Maximum current $i_{\max}$	$1.3 \cdot \sqrt{2} \cdot 18$ A	1.3
Maximum reactive current $i_{q,\max}$	$1.0 \cdot \sqrt{2} \cdot 18$ A	1.0
Angular grid frequency ω_g	$2\pi \cdot 50$ rad/s	1
Base impedance Z_b	12.8Ω	1
Filter inductance L_f	3.3 mH	0.081
Filter resistance R_f	0.51Ω	0.04
Filter capacitance C_f	$8.8 \mu\text{F}$	0.036
Switching frequency	5 kHz	100
Sampling frequency	10 kHz	200

where $\hat{\mathbf{u}}_g$ is given by (15). When $\hat{L}_f = L_f$ and $\hat{R}_f = R_f$, the voltage estimate is $\hat{\mathbf{u}}_g = H(s)\mathbf{u}_g$, cf. (18). Then, the closed-loop system is

$$\mathbf{i}_c = G_c(s)i_{p,\text{ref}} + G_a G_c(s)u_{g,\text{ref}} - [G_a G_c(s)H(s) + Y_c(s)]\mathbf{u}_g - jG_c(s)F_v(s)[u_{g,\text{ref}} - H(s)\text{Re}\{\mathbf{u}_g\}] \quad (24)$$

where the transfer functions $G_a G_c(s)$ and $G_a G_c(s)H(s) + Y_c(s)$ are

$$G_a G_c(s) = G_a G_c(s)H(s) + Y_c(s) = \frac{1}{sL_f + R_a + R_f} \quad (25)$$

when G_a is chosen according to (13). Therefore,

$$\mathbf{i}_c = G_c(s)i_{p,\text{ref}} + G_a G_c(s)(u_{g,\text{ref}} - \mathbf{u}_g) - jG_c(s)F_v(s)[u_{g,\text{ref}} - H(s)\text{Re}\{\mathbf{u}_g\}]. \quad (26)$$

In the steady state (assuming accurate model parameters), the ac-bus voltage controller (12) regulates $\text{Re}\{\mathbf{u}_g\}$ to $u_{g,\text{ref}}$ and the PLL (6) forces $\text{Im}\{\mathbf{u}_g\}$ to 0. Therefore, it follows $\mathbf{u}_g = u_{g,\text{ref}}$ and $\text{Re}\{\mathbf{i}_c\} = i_{p,\text{ref}}$ in the steady state. Furthermore, the active power $p_g = 3/2 \cdot \text{Re}\{\mathbf{u}_g \mathbf{i}_c^*\} = p_{\text{ref}}$ in the steady state (* denotes complex conjugate). The steady-state reactive current component is determined by the difference of the actual grid voltage e_g and the voltage reference together with the grid impedance and the injected active power. In an inductive grid, with an inductance L_g , the reactive current component can be calculated as

$$\text{Im}\{\mathbf{i}_c\} = \frac{\sqrt{|e_g|^2 - (\omega_g L_g i_{p,\text{ref}})^2} - u_{g,\text{ref}}}{\omega_g L_g}. \quad (27)$$

This points out that, in strong grids where L_g is small ($L_g \approx 0$ p.u.), the ac-bus voltage control is not recommended due to resulting large reactive currents when the voltage reference is not equal to the actual grid voltage ($u_{g,\text{ref}} \neq |e_g|$).

V. EXPERIMENTAL RESULTS

The proposed control system is experimentally tested with a 400-V 12.5-kVA grid converter system. The system parameters are given in Table I. The dc-bus voltage is controlled to 650 V by another back-to-back connected converter. Furthermore, to reduce switching harmonics in the grid current, a filter capacitor (C_f) is connected in parallel with the grid at the PCC. Thus,

the filter is effectively an LC filter and forms an LCL circuit with the grid inductance L_g . The experiments are carried out with different grid inductance values that are defined by the short-circuit ratio (SCR). Here, the SCR is defined as seen from the converter terminals as $\text{SCR} = L_b/(L_f + L_g)$, where $L_b = Z_b/\omega_g$ is the base inductance.

The control system is discretized using the forward Euler method $s \rightarrow (z - 1)/T_s$, for simplicity. In addition, the phase lag of the fundamental-frequency component $e^{-j\omega_g T_d}$, originating from the delay in voltage production (2), is compensated for in the current controller (7) as $\mathbf{u}_{c,\text{ref}} \rightarrow e^{j\omega_g T_d} \mathbf{u}_{c,\text{ref}}$. The control system parameters are:

- Filter parameter estimates $\hat{R}_f$ and $\hat{L}_f$ equal R_f and L_f given in Table I, and $\hat{R}_f$ models also losses of the converter.
- The bandwidth of the voltage estimation is selected as $\alpha_f = 8$ p.u. resulting in the estimator gain (17).
- The bandwidth of the current-control loop is $\alpha_c = 8$ p.u., and the corresponding gain is calculated according to (8).
- In the PLL controller, a proportional gain is used, $F_p(s) = \alpha_p$, where $\alpha_p = 0.1$ p.u. is the closed-loop bandwidth of the PLL.
- Voltage controller gains are calculated as in (13).
- The voltage reference is set to $u_{g,\text{ref}} = 0.99$ p.u. corresponding to the actual voltage level in the laboratory.

A. Current Control Mode

Fig. 5 shows the measured power step responses for $\text{SCR} = 5$ ($L_g = 0.12$ p.u.) and $\text{SCR} = 1$ ($L_g = 0.92$ p.u.) when the ac-bus voltage controller is first disabled. Then, the system is current controlled emulating fully grid-following mode of operation. Despite the current follows its reference, the PCC voltage drops in the weak grid ($\text{SCR} = 1$) when the current is increased, limiting the power that can be transmitted. In addition, in the first two power transients, the converter voltage (and its reference) hits the boundaries of the realizable voltage hexagon, limiting the rise time of the current and leading to oscillation at the sixth harmonic in u_{gd} and $\hat{u}_{gd}$.

The fault-ride-through (FRT) performance of the current-controlled converter was also tested, and the response is shown in Fig. 6. The ac-bus voltage e_g behind the grid inductance L_g dips down to 0.5 p.u. emulating a symmetric fault. It can be seen that the current is controlled without an error. However, the active power drops significantly. This originates from the voltage dip at the PCC that is further amplified by the active current injection when $\text{SCR} = 1$.

B. Grid-Voltage Supporting Mode

Fig. 7 shows measured power step responses for $\text{SCR} = 5$ and $\text{SCR} = 1$ when the ac-bus voltage controller is enabled. As can be seen, control of the voltage leads to reactive current injection (seen in i_{cq}) in the weak grid ($\text{SCR} = 1$), enabling full power transmit. In addition, in the case of $\text{SCR} = 1$, the power transients are faster compared to Fig. 5 due to a current-reference-shaping effect of the voltage controller when $u_{g,\text{ref}} \neq \mathbf{u}_g$. Similar power-step tests have also been

Fig. 5. Power reference steps when the ac-bus voltage controller is disabled for SCR = 5 and SCR = 1.

Fig. 7. Power reference steps when the ac-bus voltage controller is enabled for SCR = 5 and SCR = 1.

Fig. 6. FRT when the ac-bus voltage controller is disabled for SCR = 5 and SCR = 1.

Fig. 8. FRT when the ac-bus voltage controller is enabled for SCR = 5 and SCR = 1.

performed for a corresponding controller having the grid-voltage sensors in [22]. Compared to the results in [22], the proposed controller provides similar dynamic performance with less sensors.

For further comparison, the FRT performance of the voltage-controlled converter is shown in Fig. 8 for the symmetric fault down to 0.5 p.u. Due to the grid-supporting characteristics of the voltage control, the fault leads to the reactive current injection up to its maximum limit ($i_{q,\max} = 1$ p.u.) for SCR = 5, which can bring the PCC voltage slightly up from 0.5 p.u. On the other hand, when the SCR = 1, the voltage-restoring magnitude of reactive current remains below the maximum limit. Then, the PCC voltage is restored to its

reference value $u_{g,\text{ref}}$ during the fault. Owing to the voltage control, the active power drop can be avoided during the dip unlike in the current-controlled case (cf. Fig. 6).

Finally, Fig. 9 depicts the power control error $p_{\text{ref}} - p_g$ and voltage estimation errors $u_{gd} - \hat{u}_{gd}$ and $u_{gq} - \hat{u}_{gq}$ during the first power step of Fig. 7. The settling time of the power is 3 ms in the case of SCR = 5, whereas it is 33 ms for SCR = 1. As can be seen, the estimation and power-control errors are approximately zero in steady state.

VI. CONCLUSIONS

This paper presented a grid-voltage sensorless control system capable to operate in current control and grid-voltage supporting modes. The experimental tests showed that the

Fig. 9. Power control and estimation errors when the ac-bus voltage controller is enabled for SCR = 5 and SCR = 1. The gray dash-dotted lines marks the 5-% settling limits.

dynamic performance of the sensorless control system is comparable to its counterpart [22] that has the grid-voltage sensors. In weak grids, the voltage support enables higher active power to be transmitted through the converter.

APPENDIX

The voltage controller in [22] originates from a robust PSC design, and it is $\mathbf{i}_{v,\text{ref}} = Y'_v(s)(\mathbf{u}_{g,\text{ref}} - \mathbf{u}_g) - \mathbf{j}F'_v(s)(\mathbf{u}_{g,\text{ref}} - \text{Re}\{\mathbf{u}_g\})$ where $Y'_v(s) = G_a(1 + \alpha_a/s)H(s)$ and $F'_v(s) = (k_v/s) \cdot H(s)$. In addition, $H(s)$ is a low-pass filter, and G_a , α_a and k_v are the controller tuning parameters. In this work, the PCC voltage is estimated unlike in [22]. Therefore, a low-pass characteristics [cf. (18)] naturally exist in the estimator, and the low-pass filter is not needed in (12). Furthermore, the integrator in the original Y'_v is not included ($\alpha_a = 0$) in (12), since the power controller of PSC yielding frequency droop is not included in (6).

Nevertheless, the proposed control system can be easily changed to implement a voltage sensorless PSC variant with the following two modifications. Firstly, instead of the PLL, the angle $\hat{\vartheta}_g$ is obtained from the power controller [22]

$$\hat{\vartheta}_g = 1/s \cdot [\omega_{gN} + K_p(p_{g,\text{ref}} - \hat{p}_g)] \quad (28)$$

where K_p is the controller gain and the PCC power can be calculated as $\hat{p}_g = 3/2 \cdot \text{Re}\{\hat{\mathbf{u}}_g^s \mathbf{i}_c^{s*}\}$ or $\hat{p}_g = 3/2 \cdot (\text{Re}\{\mathbf{u}_c^s \mathbf{i}_c^{s*}\} - \hat{R}_f |\mathbf{i}_c^s|^2)$. Secondly, the voltage controller (12) is changed to a symmetric proportional-integral controller

$$\mathbf{i}_{v,\text{ref}} = G_a(1 + \alpha_a/s)(\mathbf{u}_{g,\text{ref}} - \hat{\mathbf{u}}_g). \quad (29)$$

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