Design and Operation of a Phase-Locked Loop with Kalman Estimator-Based Filter for Single-Phase Applications

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Abstract—This paper describes the design procedure of a Phase-Locked Loop (PLL) preceded by a Kalman estimator-based filter. It provides a highly accurate and fast estimate of the grid frequency and phase angle in grid-connected power electronic applications. Moreover, it enables the transformation to the rotational dq frame for various control purposes. Closed-form equations allow for a performance driven design for both PLL and estimator. By using a variable sample frequency controlled by the PLL, the Kalman filter is always operated around its center frequency, which is the rated grid frequency. The new topology is compared to other published single-phase PLL designs and its operation is verified by both simulations and experiments. The PLL exhibits excellent performance even under severely distorted utility grid voltage conditions. This robustness makes it very suitable for use in systems connected to grids having an important share of non-linear loads.

I. INTRODUCTION

The phase angle of the utility grid voltage is a piece of information used in a wide range of applications with grid-connected inverters, like active harmonic filters, active front-ends, uninterruptible power supplies and distributed generation units. It is used for the transformation of measured quantities like voltage and current to a rotating reference frame, which has shown to have some interesting features regarding control purposes. The phase angle and grid frequency are obtained by a Phase-Locked Loop (PLL).

Phenomena like harmonics, frequency variations and voltage unbalance often occur in the utility grid. Any PLL operating based on grid voltage or current measurements must be designed to be able to phase lock quickly and produce a low distortion output under all imaginable grid conditions.

This paper presents the design procedure of such a robust PLL. A Kalman filter is placed in front of the PLL in order to ensure that the PLL input at all times matches an ideal sinusoidal waveform as closely as possible, even when the voltage is highly distorted by the presence of harmonics. This ensures fast and low distortion operation of the PLL. The design and operation of both the PLL and the Kalman estimator-based filter are described into detail and verified by simulation and in an experimental setup under various voltage conditions. An interesting application in which the robust PLL has been successfully applied is a voltage and frequency droop control method for parallel inverters, as described in [1]. Here the PLL determines the phase angle used for the Park transformation and it provides a very accurate estimate of the grid frequency, used in the frequency droop control.

II. KALMAN FILTER

It is well known that in a balanced, sinusoidal and stationary three-phase system, currents and voltages can be transformed to αβ components using the Clark transformation. By applying the Park transformation AC quantities are transformed to DC, enabling the use of the fast and precise vector or field-oriented control. Active and reactive power are obtained by simple calculations on instantaneous currents and voltages. On the other hand, in single-phase systems, Clark and Park transformation cannot directly be applied and amplitude, phase angle, active and reactive power cannot be obtained from instantaneous values.

In literature, some techniques for estimating αβ components from a single-phase input signal are described, some of which are compared in [2] and [3]. The most straightforward technique incorporates a transport delay shifting the input signal over 90° to obtain the β component. Some other methods obtaining the β component are described, like the Hilbert transform, the inverse Park transformation or the use of a so-called generalized integrator [4], [5]. However, all of these methods show some important shortcomings. They all need an extra filter to ensure that the input signal has minimal non-fundamental frequency components. Some methods are strongly non-linear and require complex controllers.

In this paper, it is shown that the αβ components of a single-phase system are estimated fast and accurately using a Kalman-based filter. This technique is based on [6], where a Kalman filter is used for voltage harmonics tracking. In this section, the Kalman filter is used in an unconventional way, as knowledge of measurement and process noise is not required. Instead, formulas are described that allow to directly design the Kalman-based filter given a desired time constant and taking into account the most prominent harmonic components in the signal, ensuring good operation even under distorted utility grid conditions.

In this paper, the output components α and β of the Kalman-based filter are renamed to e and f, as to underline the difference with the Clark transformation in balanced three-phase systems. In contrast to the αβ components of the Clark transformation, the ef components are not obtained instantaneously, but through a filter. These ef components are useful not only in single-phase systems, but also in unbalanced three-phase systems, where ef components are obtained for each of the direct, inverse and zero-sequence components, and systems containing harmonics, where ef components are obtained for each harmonic.

A. Single phase systems without harmonics or DC

1) e and f estimation using the ’generalized integrator’

In single phase systems, amplitude, active and reactive power are traditionally obtained through the use of low pass filters and peak value detectors. These algorithms exhibit a trade-off between speed of convergence and ripple content. This trade-off is circumvented by using a second-order infinite gain band-pass filter with a structure described by:

\[ G(s) = k \frac{s}{s^2 + \omega^2} \]  \hspace{1cm} (1)

This structure is also called a “generalized integrator”. The system has a resonant angular frequency \( \omega \approx 2\pi f \). By adding a feedback loop (Fig. 1), two quadrature components e and f of a sinusoidal input signal with angular frequency \( \omega \) are estimated. The transfer functions to the ef components are:

\[ G_e(s) = k \frac{s}{s^2 + ks + \omega^2}; \quad G_f(s) = k \frac{s}{s^2 + ks + \omega^2} \]  \hspace{1cm} (2)
Fig. 2 shows the step response to a 50 Hz signal ($\omega = 100\pi$, $k = 100$). The input signal is represented by the dashed line, the ef output components by the solid. Comparison with the function $y = 1/\left(t\tau + 1\right)$ with $\tau = 20$ ms (dotted line) shows that the estimation of the ef components occurs with a time constant of approximately 20 ms. Further analysis shows that the time constant is approximated by $\tau = 2/k$.

For small $\tau/\omega$ ratios, the estimation performs not as good, as an offset is present in the initial estimation of the f component (Fig. 3) for $\tau = 2$ ms and $\omega = 100\pi$. The reason for this offset is that only the e state is corrected by the error input, whereas the f component is only indirectly adjusted. As a result, the convergence occurs much slower than could be expected from the time constant $\tau = 2$ ms (dotted line). As $\tau$ further decreases, the performance becomes even worse as the initial offset in the f component is larger and disappears more slowly.

2) e and f estimation using the Kalman state estimator

By providing an extra link with gain $k_f$ from the error to the f-state (Fig. 4), this initial offset is prevented. The transfer functions to the ef components are:

$$ G_e(s) = \frac{k_e s - ak_f}{s^2 + ks + \omega^2 - ak_f}; \quad G_f(s) = \frac{k_f s + ak_e}{s^2 + ks + \omega^2 - ak_f} \quad (3) $$

The optimal gains $k_e$ and $k_f$ are calculated by designing a Kalman state estimator given a state-space model of the oscillator and process and measurement noise covariance data. The oscillator model is shown graphically inside the dotted box of Fig. 4. It has one measured output $y$ (e component), no measured inputs $u$, but two white noise inputs $w_e$, $w_f$. The continuous-time state-space model of the oscillator is:

$$ x = Ax + Bu + Gw \quad \text{(state equation)} $$
$$ y_e = Cx + Dw + Hv + v \quad \text{(measurement equation)} \quad (4) $$

with known inputs $u$ and process and measurement white Gaussian noise $w$ and $v$ satisfying:

$$ E(w) = 0; \quad E(v) = 0 $$
$$ E(wv^T) = Q; \quad E(vv^T) = R; \quad E(wv^T) = 0 $$

The $A$, $[B \ G] (= [G])$, $C$ and $[D \ H] (= [H])$ matrices are:

$$ A = \begin{bmatrix} 0 & -\omega \\ \omega & 0 \end{bmatrix}; \quad B = \begin{bmatrix} \frac{1}{\omega} \\ 0 \end{bmatrix}; \quad \frac{C}{10}; \quad D = \frac{H}{10}; \quad H = 0 \quad (6) $$

As there are no measured inputs, both $B$ and $D$ are empty matrices. As there is only one measured output $e$, the measurement noise covariance data $R$ is represented by a scalar $r$. As there are two noise inputs $w_e$ and $w_f$, the process noise covariance data is represented by a 2-by-2 diagonal matrix $Q$. Further, measurement and process noise are assumed to be mutually uncorrelated:

$$ R = r; \quad Q = \begin{bmatrix} q & 0 \\ 0 & q \end{bmatrix}; \quad N = 0 \quad (7) $$

The Kalman estimator is obtained by the MATLAB function $\text{kalman.m}$ given the oscillator model and covariance data (Fig. 4). The gains $k_e$ and $k_f$ can be calculated by analytically solving the Riccati equation obtaining the Kalman gain $L$:

$$ L = \frac{k_e}{k_f}; \quad k_f = \omega \left(1 - \sqrt{1 + \frac{q}{\omega^2 - r}}\right); \quad k_e = \frac{2q}{r \sqrt{k_f^2 + k_e^2}} \quad (8) $$

The input is the measured oscillating signal and the outputs are the two states of the oscillator, the first being an estimate of the measured signal, the second lagging with 90°. These states are thus equivalent to the e and f states described above.

When $q = 1$ and $r = 2 \cdot 10^{-4}$, the step response of the Kalman estimator to a 50 Hz input signal is very similar to that of the generalized integrator with feedback loop (Fig. 2). The estimation of the ef components occurs with a time constant of 20 ms. In general, the convergence time constant is:

$$ \tau = \frac{2}{k_f} \approx \frac{2r}{q} \quad (9) $$

Whereas the difference between both methods is marginal for a time constant of 20 ms, the response of the Kalman estimator for $\tau = 2$ ms clearly shows an improved performance (Fig. 5 vs. Fig. 3). In contrast to the result obtained by the generalized integrator with feedback loop, there is no significant initial offset present and the convergence rate is now more or less in accordance with what could be expected from the time constant $\tau = 2$ ms. Further decreasing $\tau$ does not lead to a deteriorated performance.

3) Discrete-time equivalent of the Kalman state estimator

As the estimator is to be implemented in real-time on a DSP, a discrete equivalent of the Kalman estimator is developed, based on the discrete-time model of the oscillator. The block diagram is given by Fig. 6.

The discrete-time state-space model of the oscillator is:

$$ A = \begin{bmatrix} \cos(\alpha_	au) & -\sin(\alpha_	au) \\ \sin(\alpha_	au) & \cos(\alpha_	au) \end{bmatrix}; \quad B = \begin{bmatrix} 1/\omega \\ 0 \end{bmatrix}; \quad C = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}; \quad (10) $$

$$ D = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix}; \quad H = 0 $$

with $t$, the sampling time and (7) the covariance data.

The Kalman estimator is again obtained using the MATLAB function $\text{kalman.m}$ given the discrete-time oscillator model and the covariance data (Fig. 6). Analysis of this estimator shows that the convergence time constant $\tau$ is (if $\tau \gg t$):

$$ \tau \approx t \sqrt{\frac{2r}{q}} \approx \sqrt{k_f^2 + k_e^2} \quad (11) $$

In Fig. 7, the results of the discrete-time model with $t = 1$ ms are compared to those of the continuous-time model.
B. Single phase systems with harmonics and DC

A main advantage in using the Kalman estimator structure lies in the response under distorted conditions, as it is easily extended with additional states to estimate DC as well as harmonic components of the input signal. Even if only the estimate of the fundamental component is needed, it is advantageous to estimate the most important harmonic components as well, as this improves the fundamental component tracking considerably. The computational cost is relatively modest as the Kalman estimator is fully linear. The estimation of harmonic components results in two extra states per harmonic and one state for the DC component.

As an example, a rather highly distorted 50 Hz harmonic signal is considered consisting of a DC, a fundamental and a third harmonic component, all having amplitude 1. The discrete-time state space model is:

$$\mathbf{A} = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 \\ 0 & \cos(\omega_0 t) & -\sin(\omega_0 t) & 0 & 0 \\ 0 & \sin(\omega_0 t) & \cos(\omega_0 t) & 0 & 0 \\ 0 & 0 & 0 & \cos(3\omega_0 t) & -\sin(3\omega_0 t) \\ 0 & 0 & 0 & \sin(3\omega_0 t) & \cos(3\omega_0 t) \end{bmatrix};$$

$$\mathbf{B} \mathbf{G} = \begin{bmatrix} t_f \end{bmatrix}; \quad \mathbf{C} = \begin{bmatrix} 1 & 0 & 1 & 0 \end{bmatrix}; \quad \mathbf{D} \mathbf{H} = 0$$

The covariance data are:

$$\mathbf{R} = \begin{bmatrix} q_0 & 0 & 0 & 0 & 0 \\ 0 & q_1 & 0 & 0 & 0 \\ 0 & 0 & q_1 & 0 & 0 \\ 0 & 0 & 0 & q_1 & 0 \\ 0 & 0 & 0 & 0 & q_1 \end{bmatrix}; \quad \mathbf{Q} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix}; \quad \mathbf{N} = 0$$

where $q_0$, $q_1$, and $q_3$ are the process covariances associated with DC, fundamental and third harmonic components, respectively. The structure of the estimator is shown in Fig. 8. Although this structure is of computational order $O(x)$ with $x$ the number of harmonics, the resulting Kalman state estimator matrix A-LC has a number of non-zero elements of computational order $O(x^2)$. Therefore, it is desirable to implement the structure of Fig. 8, resulting in a significantly reduced computation time.

Again, the Kalman estimator is obtained by using the Matlab function kalman.m. Detailed analysis shows that the convergence time constants $\tau_0$, $\tau_1$ and $\tau_2$ of the DC, the fundamental and the third harmonic components, respectively, are approximately (if $\tau_{0,1,3} \gg \tau_z$):

$$\tau_0 \equiv t_s \sqrt{\frac{r}{q_0}}; \quad \tau_1 \equiv t_s \sqrt{\frac{2r}{q_1}} \equiv \frac{2t_s}{k_{1s} + k_{1f}}; \quad \tau_3 \equiv t_s \sqrt{\frac{2r}{q_3}} \equiv \frac{2t_s}{k_{3s} + k_{3f}}$$

In Fig. 9 the response of the dc, fundamental and third harmonic components are shown together with their sum for $\tau_0 = \tau_1 = \tau_3 = 20$ ms and $t_s = 100$ $\mu$s. The convergence of all components occurs with a time constant of about 20 ms.

Although the description of the Kalman estimator above focuses on single-phase systems, it is applicable to three-phase systems as well. Here, one can use a Kalman estimator for each phase, or, in case of a three-phase system without neutral, a Kalman estimator which takes into account the interdependence between the voltages and currents of the three phases, resulting in a lower number of states (see [7]).

III. PHASE-LOCKED LOOP

A. Structure and design

The transformation of single- or three-phase voltages or currents to a rotating reference frame (referred to as the dq reference frame) has proven to be very useful for control purposes. The Clark transformation is applied to obtain the $d\beta$ components. In single-phase systems, one could use a Kalman estimator to obtain the $d\beta$ components, which are similar to the $q\phi$ components in three-phase systems. The $d\beta$ components are calculated using the Park transformation matrix:

$$\begin{bmatrix} u_d \\ u_{\beta} \end{bmatrix} = \begin{bmatrix} \cos(\theta) & \sin(\theta) \\ -\sin(\theta) & \cos(\theta) \end{bmatrix} \begin{bmatrix} u_d \\ u_{\beta} \end{bmatrix}$$

In order to synchronize this transformation with the fundamental frequency of the input signal, the phase angle $\theta$ needs to be known. This task is usually accomplished by a phase-locked loop (PLL). The design of a PLL is described in literature, e.g. [8] and [9]. The procedure below is, though straightforward, slightly different from this approach.

The $q$-component is calculated using (17) and a PI controller is used to drive the $q$-component to zero. As a result of this, the $d$-component becomes equal to a DC value with amplitude $U_{\text{peak}}$. The structure is represented in Fig. 10(a) and
is equivalent to the model shown in Fig. 10(b) as:

$$u_x = U_{\text{peak}} \cdot \cos(\theta') \quad u_y = U_{\text{peak}} \cdot \sin(\theta')$$

(17)

with $\theta'$ the angle between the e-axis and the rotating voltage vector $U$ (Fig. 11). $\theta$ represents the angle between the $dq$ reference frame and the e-axis. The PLL tries to equate the angles $\theta$ and $\theta'$. Since $\sin(\theta' - \theta) \approx \theta' - \theta$, the closed-loop system from $\theta'$ to $\theta$ approximately is:

$$\frac{\theta}{\theta'} = \frac{k_p \cdot s + k_i}{s^2 + 2 \cdot \zeta \cdot \omega_n \cdot k_p \cdot s + \omega_n^2}$$

(18)

with $a$ the location of the zero, $\zeta$ the damping ratio and $\omega_n$ the natural frequency. The settling time $t_{\text{set}}$ for a second order system is approximately [10]:

$$t_{\text{set}} = \frac{4}{\zeta \cdot \omega_n}$$

(19)

Thus, given the settling time, damping ratio and voltage amplitude, proportional gain $k_p$ and integral gain $k_i$ are:

$$k_p = \frac{8}{t_{\text{set}} \cdot U_{\text{peak}}} \quad k_i = \frac{16}{\zeta^2 \cdot t_{\text{set}}^2 \cdot U_{\text{peak}}}$$

(20)

Note that, because of the zero, the system exhibits overshoot for all values of $\zeta$, even when $\zeta$ is greater than 1. However, for $\zeta \geq 1$, the system roots are real and there is no oscillating behavior, resulting in a single overshoot. For the transfer function of (19), it can be shown that:

$$a = \frac{1}{\zeta^2 \cdot \omega_n^2}$$

(21)

Thus, the percent overshoot of the system is solely determined by $\zeta$. For $\zeta = 1$, the overshoot is about 13%, which is acceptable. The settling time is chosen equal to several grid frequency periods (e.g. $t_{\text{set}} = 60$ ms).

The analogue scheme of Fig. 10(a) is discretized using the backward Euler transformation. Anti-windup is implemented in order to limit the integrator output to e.g. 1 Hz if the grid frequency is never to exceed the rated frequency by this value. The sole purpose of this anti-windup is to dampen the response in case sudden large frequency or phase angle variations occur, causing less overshoot.

In order to prevent $\theta$ from increasing to infinity, causing overflow, the integration of $\omega$ to obtain $\theta$ is somewhat altered. Inside the integration loop, the nearest integer less than or equal to $\theta/(2\pi)$ is subtracted from $\theta/(2\pi)$. As a result, the value of $\theta$ always remains between 0 and $2\pi$. Fig. 12 shows the detailed digital PLL implementation.

### B. PLL operation under distorted voltage conditions

The operation of three-phase PLL algorithms under distorted voltage conditions is examined in [8] and [9]. The three-phase voltages are transformed to the synchronous reference frame by application of the Clark and Park transformation, where the PLL calculates the transformation angle for the Park transformation.

Fig. 13 shows the operation of the PLL without any filter in front of it. The input signal is a three-phase voltage with 5th harmonic amplitude equal to 5% of the rated frequency component appearing at time $t = 0$ ms (top graph). The second graph shows the voltages in the $d\beta$ frame. The last two graphs show the angle error and the output frequency of the PLL, with parameters as calculated above. Clearly, the frequency estimation is severely affected by the occurrence of the 5th harmonic, compromising its further use.

Fig. 14 shows the behavior in response to the same input

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**Fig. 8:** Block diagram equivalence of the ef estimation using the Kalman estimator with harmonics and DC.

**Fig. 9:** Response of the dc (top), the fundamental (second), the third harmonic components (third) of the Kalman estimator and the sum of these components (bottom) with time constants $\tau_0$, $\tau_1$, $\tau_2 = 20$ ms ($\tau = 100$ µs).

**Fig. 10:** (a) implemented, (b) equivalent control model of the PLL system.

**Fig. 11:** Meaning of the angles $\theta$ and $\theta'$.

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signal of the PLL with Kalman estimator in the front, taking into account the presence of the 5th harmonic component. As the fundamental voltage component estimation is much more accurate, the angle error is driven to zero and the frequency estimation converges quickly to its 50 Hz rated value.

IV. SYNCHRONIZATION OF THE ALGORITHM EXECUTION RATE WITH THE GRID FREQUENCY

The use of a Kalman filter ensures the correct operation of the PLL under distorted voltage conditions. However, as the Kalman filter is tuned to one specific value of the grid frequency (e.g. 50 Hz), there will be a deviation between the real fundamental frequency component and the estimated one in case the grid frequency is not exactly 50 Hz. This deviation results in a nonzero phase angle estimation error and a ripple on the estimated frequency.

Fig. 15 illustrates this problem. At time = 0 ms, a 1 Hz frequency step change occurs (top). The estimated waveform (solid grey) clearly shows a phase delay in comparison to the 51 Hz input signal. The estimated frequency (bottom) shows an oscillating estimation error, which is a result from the nonzero angle error (middle).

These errors can be alleviated by increasing the time constant of the Kalman filter and the settling time of the PLL, but only at the cost of slower convergence. An alternative solution is to dynamically adjust the algorithm sample time, synchronizing the sample frequency with the grid frequency. As a result, the Kalman estimator algorithm sees the voltage as a 50 Hz signal, though the real frequency may be different. Of course, this method requires that is technically possible to dynamically adjust the sample frequency. Fig. 16 shows the extended PLL scheme, including the synchronization of the sample frequency. The integral controller gain $k$ determines the synchronization time constant: $\tau_{\text{sync}} \approx 1/k$.

The multiplication factor $a$ in the proportional term (Fig. 16) could be chosen either equal to zero (thus omitting the proportional controller), or equal to the time constant $\tau_1$ of the estimator (as expressed in (15)), thus canceling a pole. In the latter case, the frequency signal contains some minor high-frequency components, which are absent if no proportional control is added. However, as is shown in Section V, use of a proportional term is necessary in order to avoid overshoot and oscillatory behavior when the synchronization time constant nearly equals the Kalman estimator time constant.

V. EXPERIMENTAL RESULTS

The PLL and Kalman estimator schemes described above are also tested in an experimental setup. A voltage source inverter platform, as described in [11], is used to generate a predefined voltage waveform. A separate platform is used to measure this voltage. The Kalman filter estimates the fundamental component of the measured signal and the PLL

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**Fig. 12:** Detailed digital implementation of the PLL.

**Fig. 13:** PLL response to the appearance of a 5% fifth harmonic at time = 0 ms in a three phase system.

**Fig. 14:** Response of PLL and Kalman filter to the appearance of a 5% fifth harmonic at time = 0 ms in a single phase system.

**Fig. 15:** Response of the Kalman estimator and PLL to a 1 Hz frequency step change at time = 0 ms.

**Fig. 16:** PLL scheme with Kalman estimator and synchronization of the sample frequency.
performs a phase-lock, thereby giving a high resolution estimate for the fundamental frequency as an internal signal. In fact, the accuracy of the frequency estimate is only limited by the accuracy of the DSP clock crystal, which in this case is 100 ppm (i.e. 5 mHz for a 50 Hz grid frequency).

First a comparison is made between the PLL scheme using a transport delay in order to obtain αβ components, the PLL scheme with a Kalman filter which estimates 2nd, 3rd, 5th and 7th harmonic voltages and a PLL scheme with a Kalman filter not estimating any harmonics. Fig. 17 shows the response of these three schemes to a magnitude step change of the 5th harmonic voltage (from 0 to 5 % of the magnitude of the 50 Hz voltage component). Clearly, the scheme with the transport delay needs an additional filter for accurate operation under distorted grid conditions. In case the Kalman filter estimates no harmonics, the steady state frequency value contains a ripple, which almost disappears when the Kalman filter does estimate harmonic voltages. Also note that the scales are different: the scale for the upper frequency measurement (transport delay) is five times larger than the one for the other two measurements.

Another comparison was made between a PLL scheme without and with synchronization. For the PLL scheme using synchronization, a further distinction was made by varying the synchronization time constant and turning on/off the proportional controller gain. The response of these five schemes to a step in the frequency from 50 Hz to 51 Hz is shown in Fig. 18. It is clear that the use of synchronization is necessary for an accurate frequency estimate. In case the synchronization time constant is significantly larger than the time constant of the Kalman filter, it is acceptable to omit the proportional term. However, in case both time constants are nearly equal, the absence of the proportional term leads to overshoot and oscillatory behavior, as is shown in Fig. 18.

**VI. CONCLUSION**

In this paper, the design of a PLL with a Kalman-based estimator for single-phase applications is discussed. Equations are given for the calculation of the parameters of the Kalman estimator and the PLL. The operation of the proposed scheme is verified by simulation and by testing in an experimental setup. The combination of the PLL with a Kalman-based estimator shows superior behavior when compared to other single-phase PLL techniques published in literature. The experiments also show the influence of certain parameters on the response of the system. By the presented technique, a very accurate and fast estimation of the frequency is obtained, even under highly distorted grid conditions. Other applications include the use of the phase angle for control purposes, as it enables the transformation to the rotational dq reference frame for single-phase grid-connected systems.

**REFERENCES**


