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A pre-filter based PLL for three-phase grid connected applications

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ABSTRACT

This paper proposes a phase estimator based on a pre-filter based phase locked loop (PF-PLL), which overcomes serious limitations of existing PLL routines under frequency variation, distorted power grid conditions and measurement errors. This is achieved by a chain of pre-filters, which systematically attenuates all distortion components and extracts the fundamental, positive-sequence signal. The reported PF-PLL is capable of fast and precise phase estimation, even with unbalanced and highly distorted threephase electrical variables, which may be measured with low-cost transducers of reduced accuracy. Consequently, it is appropriate for control of all grid connected equipment, including the dynamic voltage restorer (DVR), where it is often required to synchronize with line currents, which experience much higher levels of harmonic distortion than voltages. This paper presents a detailed analysis of the PF-PLL, its evaluation through simulation and experiments, using a fixed-point DSP based laboratory prototype.

1. Introduction

Phase angle estimation of the AC utility voltage or current is critical for reliable operation of grid synchronized electrical equipment such as UPS, custom power devices and electronic energy meters [1]. This information is extracted from the measured periodic voltage/current signal using a phase locked loop (PLL). The classical construct of a PLL [2] with a controlled oscillator – and output synchronized with the input through feedback – is still used in off-grid systems [3]. For power system applications, the oscillator was directly derived from Schmidt triggers driven by the power line voltages [4,5]. However, these methods are not reliable, especially in the presence of distortions and noise in the measured signals. Another software based method uses a rotating phasor of impedance to calculate the deviation of frequency [6] but the computation time depends upon the estimation error.

A different class of approaches [7–10], more suitable for grid-connected applications, uses a synchronous rotating frame (SRF) based PLL. This SRF-PLL requires transformation of the three-phase voltages with an SRF, which is phase locked with the utility voltage. Deficiencies of the SRF-PLL were highlighted in [11,12], under unbalance and harmonic distortion in the measured signals. These were claimed to be overcome by the double decoupled SRF (DDSRF)-PLL [11] by using two synchronously rotating frames having exactly opposing angular velocities, with a controller designed to handle small-signal disturbances. In a practical situation, the input quantities are measured using voltage/current transducers, which usually introduce dc offsets in the measured output. Although dc offsets severely degrade the PLL performance [12], this problem has not been solved previously.

In this paper, a new pre-filter based PLL (PF-PLL) for grid-synchronized applications is reported. The PF-PLL is capable of maintaining excellent phase lock under unbalance, harmonic distortions and dc offsets in the measured signal. It has fast dynamic response and its locking time is independent of the type and magnitude of disturbance inputs. The underlying theoretical analysis is presented and simulation results provided. It is shown that this method ensures reliable phase-lock with even discontinuous input signals (high distortion) and large frequency deviation. Extensive experimental results are provided for validation of analytical and simulation results.

2. The SRF PLL

As the instantaneous phase angle (θ) of a periodic input signal is not directly measurable, it has to be estimated from measured instantaneous values of the input. To compute an accurate estimate (θ^*) of the actual input phase (θ), the estimation error ($\Delta\theta$), defined in (1), must be arbitrarily minimized.

$$\Delta\theta = (\theta^* - \theta). \tag{1}$$

The estimation error is expressed in terms of measured variables using the SRF approach [7]. Here, a set of balanced, three-phase

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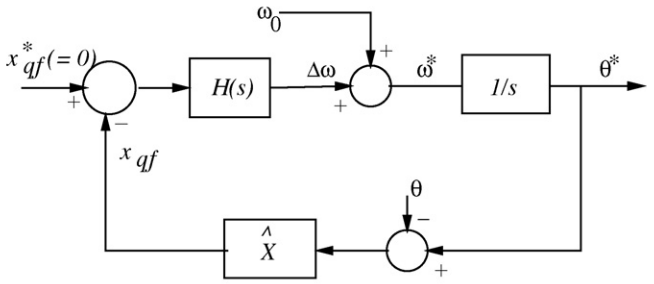

Fig. 1. Control model of PLL system: basic loop.

signals with amplitude X,

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = X \begin{bmatrix} \cos\ \theta \\ \cos(\theta - 2\pi/3) \\ \cos(\theta + 2\pi/3) \end{bmatrix} = X \begin{bmatrix} \cos(\omega t) \\ \cos(\omega t - 2\pi/3) \\ \cos(\omega t + 2\pi/3) \end{bmatrix} \tag{2}$$

are referred to a rotating reference frame having angular phase θ^*, with constant time derivative (angular frequency), ω^*, using the following (Park's) transformation

$$\sqrt{\frac{3}{2}} \begin{bmatrix} x_{df} \\ x_{qf} \end{bmatrix} = \begin{bmatrix} \cos(\theta^*) & \sin(\theta^*) \\ \cos(\theta^* - 2\pi/3) & \sin(\theta^* - 2\pi/3) \\ \cos(\theta^* + 2\pi/3) & \sin(\theta^* + 2\pi/3) \end{bmatrix}^T \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix}. \tag{3}$$

The corresponding transformed variables are as follows [11].

$$\begin{bmatrix} x_{df} \\ x_{qf} \end{bmatrix} = \sqrt{\frac{3}{2}} X \begin{bmatrix} \cos(\Delta\theta) \\ \sin(\Delta\theta) \end{bmatrix} = \hat{X} \begin{bmatrix} \cos(\Delta\theta) \\ \sin(\Delta\theta) \end{bmatrix}. \tag{4}$$

The second row of (4) defines an observable variable (x_{qf}) for the phase estimator. For small values of $\Delta\theta$, x_{qf} is a linear function of the observed state $\Delta\theta$. Phase lock, equivalent to zero estimation error, is achieved by regulating x_{qf} to zero through a suitable dynamic gain, $H(s)$. Fig. 1 shows the schematic of this estimation loop, with loop gain H_{oll}. Magnitude of the measured signal, $\hat{X}$, appears as a scalar gain in the feedback path. Considering discrete-time realization with a P-I controller and modeling the sampling delay, T_s, as a first-order lag [7], the loop gain is

$$H_{oll}(s) = \hat{X} H_{ol}(s) = \hat{X} K_{pll} \left(\frac{1 + sT_{pll}}{sT_{pll}} \right) \left(\frac{1}{1 + sT_s} \right) \left(\frac{1}{s} \right) \tag{5}$$

where K_{pll} and T_{pll} are the controller parameters. In [7], the symmetrical optimum approach is used to obtain a unique closed-loop bandwidth, ω_b. For the set of parameters given in Table 1, bode plot of H_{oll} is shown in Fig. 2. A phase margin of 84° indicates stable operation, with settling time of 30 ms. This method gives excellent phase lock under ideal conditions of balanced and undistorted utility but is adversely affected with the slightest deviation from ideal conditions.

3. SRF-PLL under utility distortion

Principal disturbances in the measured signals (voltages or currents) include (a) load and voltage unbalance, (b) harmonic distortions, (c) variation in supply frequency and (d) dc offsets in measured signals. Being generally not identical for all the measured phase variables, these offsets also adversely affect PLL performance.

Table 1
Controller parameters for SRF-PLL.

T_{pll}	K_{pll}	T_s	$\hat{X}$
81 ms	1111	100 µs	1.0

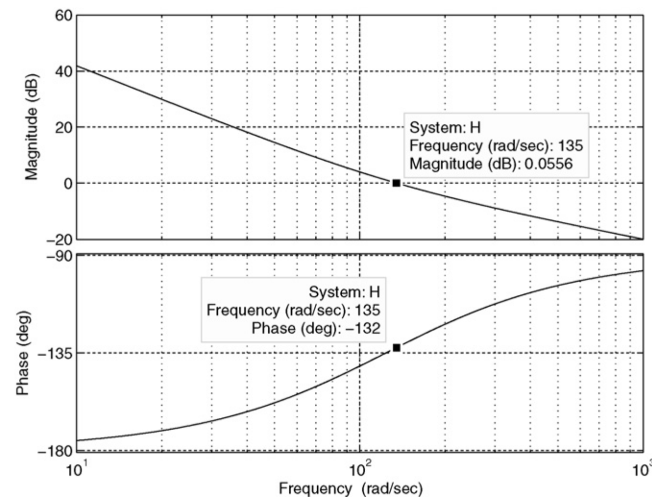

Fig. 2. Bode plot of the loop transfer function of SRF-PLL.

The following analysis of the transformed signals explains these facts.

3.1. Effect of utility distortions

Let $x_{(a,b,c)}$ be a set of three-phase quantities having fundamental positive-sequence components, $x_{+(a,b,c)}$, fundamental negative sequence components, $x_{-(a,b,c)}$, fundamental zero sequence components, $x_{0(a,b,c)}$, dc offsets, $\bar{x}_{(a,b,c)}$ and harmonic frequency components, $x_{h(a,b,c)}$. Denoting the fundamental frequency as f, the composite signal is

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \underbrace{X_+ \begin{bmatrix} \cos\ \omega t \\ \cos(\omega t - 2\pi/3) \\ \cos(\omega t + 2\pi/3) \end{bmatrix}}_{x_{+(a,b,c)}} + \underbrace{X_- \begin{bmatrix} \cos\ \omega t \\ \cos(\omega t + 2\pi/3) \\ \cos(\omega t - 2\pi/3) \end{bmatrix}}_{x_{-(a,b,c)}}$$

$$+ \underbrace{X_0 \begin{bmatrix} \cos\ \omega t \\ \cos\ \omega t \\ \cos\ \omega t \end{bmatrix}}_{x_{0(a,b,c)}} + \underbrace{\begin{bmatrix} \bar{x}_a \\ \bar{x}_b \\ \bar{x}_c \end{bmatrix}}_{\bar{x}_{(a,b,c)}} + \underbrace{X_h \begin{bmatrix} \cosh(\omega t) \\ \cosh(\omega t - 2\pi/3) \\ \cosh(\omega t + 2\pi/3) \end{bmatrix}}_{x_{h(a,b,c)}} \tag{6}$$

where

$$\omega = 2\pi f \quad \text{and} \quad h = (6k \pm 1), \quad k = 1, 2, 3, \ldots \tag{7}$$

Using (3), transformation relative to a SRF rotating at angular frequency ω^* (=$2\pi f^*$) causes frequency shift in each component as follows.

$$\sqrt{\frac{2}{3}} \begin{bmatrix} x_{df} \\ x_{qf} \end{bmatrix} = X_+ \begin{bmatrix} \cos(\omega \sim \omega^*)t \\ \sin(\omega \sim \omega^*)t \end{bmatrix} + X_- \begin{bmatrix} \cos(\omega + \omega^*)t \\ \sin(\omega + \omega^*)t \end{bmatrix}$$

$$+ \bar{X} \begin{bmatrix} \cos(\omega^* t) \\ \sin(\omega^* t) \end{bmatrix} + X_h \begin{bmatrix} \cos\{(\omega^* \pm h\omega)t\} \\ \sin\{(\omega^* \pm h\omega)t\} \end{bmatrix} \tag{8}$$

where $\bar{X}$ is a linear function of $(\bar{x}_a, \bar{x}_b, \bar{x}_c)$ and '$\sim$' indicates magnitude of difference between two quantities. Apart from random noise which is filtered by the PI stage, (8) gives the complete description of the transform outputs. Frequency of the positive sequence transforms is the difference frequency $(\omega \sim \omega^*)$ which is very small compared to the negative sequence transform frequency, for small frequency deviations.