Frequency Tracking in Power Networks in the Presence of Harmonics

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Abstract

Three new techniques for frequency measurement are proposed in the paper. The first is a modified zero crossing method using curve fitting of voltage samples. The second method is based on polynomial fitting of the DFT quasi-stationary phasor data for calculation of the rate of change of the positive sequence phase angle. The third method operates on a complex signal obtained by the standard technique of quadrature demodulation. All three methods are characterized by immunity to reasonable amounts of noise and harmonics in power systems. The performance of the proposed techniques is illustrated on several scenarios by computer simulation.

Keywords: Phase measurements, frequency measurements, power system stability, power system harmonics.

1 Introduction

The problem of determination of accurate frequency in power networks has grown more complex in the recent past. There are multiple reasons for this. The dynamic balance between load and generation, which is a prerequisite for stable power system operation, has become more difficult to maintain because the expansion of the transmission network does not follow the growth of the system. The direct consequence is that the security margins are generally smaller, and quite often power systems operate at the brink of instability, possibly resulting in a blackout. Such operating practices imposed by the practical reasons are further aggravated due to the effects of deregulation. Non-utility generation and wheeling may reduce the stability margins of a normally secure system. It is therefore very important for utilities to develop means to monitor and control the dynamics of the power system.

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The hardware which allows tracking of very fast system dynamics is emerging from new technologies such as GPS satellite receivers (which provide the required $1\mu s$ accuracy for system-wide measurement synchronization), faster computers for processing of system data (parallel architectures), and efficient communication networks (dedicated fiber optic, or digital telehone network, enhanced with data compression for increased information throughput). Direct measurement of the system state and frequency is a very important component of such a system. The hardware platforms are already developed and are undergoing extensive field testing. Some of the underlying principles and most obvious applications are described in [1] [2].

Some of the solutions for the newly created situation are giving rise to new problems. The increased application of power electronics in power systems have allowed leaps forward in transmission capacity (such as HVDC) and flexible control of the system dynamics (such as FACTS elements). But those devices are generators of harmonics, which are corrupting the purity of the 60 Hz sine waves that should, in theory, be the only frequency component in the power networks. In addition, many industrial customers are creating harmonics by using power electronics equipment, arc furnaces, etc. Harmonics, which used to create problems in distribution networks only, are becoming a big nuisance in transmission networks, so that some utilities are already building harmonic measurement systems for transmission networks [3].

Among the first techniques for frequency measurement were those based on zero crossing. They were gradually abandoned due to their sensitivity to noise, presence of DC components in the signal, and harmonics. However, their inherent simplicity cannot be matched by any other technique. As will be shown in this paper, when combined with a data smoothing technique, zero crossing may produce surprisingly good performance. A variation of the same method involves frequency multiplication of the measured signal using PLL, which reduces the measurement time, but does not have very good resolution or dynamic properties. Phadke et al. [2] propose the Discrete Fourier Transform of the voltage samples to be used recursively for calculation of a stationary phasor, and positive sequence phasor rotation to be used for measurement of the frequency. The algorithm is inherently insensitive to harmonics because of the application of the DFT, but as proposed in [2], it is vulnerable to noise, and requires long measurement windows when frequency deviation from nominal is small. Girgis et al. [7]-[8] have proposed to treat the frequency as a stochastic signal and have applied a two-stage algorithm based on a combination of an adaptive extended Kalman filter and an adaptive linear Kalman filter. The objective is to use the measurement as an underfrequency load shedding relay. Although the noise performance of the method is good, it is not meant to be insensitive to harmonics in the measured signal. Sachdev and Giray [5]-[6] propose the the use of a least squares technique after the approximation of the measured signal with the truncated Taylor expansion. Due to the combined effects of approximations, the method may be sensitive to the presence of harmonics and noise in the signal. Kezunovic et al. [9]-[10] propose two new techniques based on digital signal processing and quadratic forms of sample data. The authors claim good noise performance of the algorithm, but not immunity to harmonics. Kamwa and Grondin [12] have proposed recursive least squares and recursive least mean squares for dynamic estimation with an objective to track both voltage phasor and frequency. They utilize band-pass filters to tune out DC and harmonics from the signal. Eckart et al. [4] propose a definition of the instantaneous frequency as angular velocity of the rotating voltage space phasor, similar to the scheme proposed earlier for a positive sequence voltage phasor by Phadke et al. [2]. They utilize four connected FIR-filters to suppress the effects of noise and harmonics and linear observer to extract the electromechanical component of frequency deviation. The obtained results are good, but the proposed scheme is very complex.

Three new techniques will be investigated in this paper. They differ in complexity and accuracy, but all can be considered for implementation, depending on the measurement requirements. The first is a modified zero crossing technique based on curve fitting of voltage samples to enhance noise immunity. The second technique is based on the positive sequence voltage phasors obtained by DFT [2], smoothed via minimum least square polynomial fitting of the quasi-stationnary phasor data. The third technique emerged from a recently proposed method for measurement of the phase angle by demodulation and filtering [14], [15].

2 Definitions of Frequency

Let us assume that the measured signal consists of a fundamental and harmonics

$$v(t) = \sum_{n=1}^{\infty} V_n \cos(n\omega_0 t + \delta_n(t)) \tag{1}$$

where ω_0 is a synchronous frequency, V_n are magnitudes of harmonics, t is time, n harmonic order, and $\delta_n(t)$ are the time varying phase angles of the harmonics. Considering that $\delta_n : R \mapsto R$ is a function of time, instantaneous frequency deviation may be defined as

$$f_1 = \frac{1}{2\pi}\frac{d\delta_1}{dt} \tag{2}$$

Various disturbances, such as transient oscillations and subsynchronous resonance, may modulate $\delta_1(t)$, which is both the way to represent them in simulations and the reason we are seeking to estimate them from voltage measurements. In the more general case, additive noise $\epsilon(t)$ may be added to (1) to model the corruption of the signal. Its statistical properties are $E\{\epsilon(t)\} = 0$ and $E\{\epsilon^2(t)\} = a + b \mid v(t) \mid$, according to [2], where a and b are constants, and $v(t)$ is the actual magnitude of the measured voltage. This noise model will be used in the simulations. In the explanation of the proposed algorithms, we will omit the noise, since we do not deal with it in an explicit fashion. In the steady state, the frequency is defined as the inverse of the *shortest* time interval $T = f^{-1}$ between two instants when the function takes on the same value for all times

$$(\forall t)(v(t) = v(t-T) = v(t-f^{-1})) \tag{3}$$

Since $\delta(t) \neq const.$, (3) cannot be used in general, but whenever the change of $\delta(t)$ is small enough with respect to the synchronous frequency, (3) can be utilized ($\dot{\delta} \ll \omega_0$). In fact, we will use (3) as a definition of frequency for modified zero-crossing technique.

3 Modified Zero Crossing Technique

Let us assume that the voltage waveform has been sampled and that a sample $v[k]$ is defined as

$$v[k] = v(k\Delta t) = \sum_{n=1}^{\infty} V_n \cos(n\omega_0 k\Delta t + \delta_n(k\Delta t)) \tag{4}$$

We can then define a measurement window $V[k]$ as a set of M consecutive samples such that

$$V[k] = \begin{bmatrix} v[k+1] & v[k+2] & \cdots & v[k+M] \end{bmatrix}^T \tag{5}$$

The triggering of the measurement will be initiated every time the counter determines that exactly one half of the samples are with positive sign (assuming M is an even integer). Let us fit the l-th degree polynomial $p_l : R \mapsto R$

$$p_l(t) = a_0 + a_1 t + a_2 t^2 + \cdots a_l t^l = \sum_{j=0}^{l} a_j t^j \tag{6}$$

using the least squares techique. The solution is obtained from the overdetermined system of linear equations

$$\begin{bmatrix} 1 & (k+1)\Delta t & \cdots & (k+1)^l \Delta t^l \\ 1 & (k+2)\Delta t & \cdots & (k+2)^l \Delta t^l \\ & & \cdots & \\ 1 & (k+M)\Delta t & \cdots & (k+M)^l \Delta t^l \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ \vdots \\ a_l \end{bmatrix} = V[k] \tag{7}$$

or

$$K \cdot \mathbf{a} = V[k] \tag{8}$$

which can be solved using the least squares technique

$$\mathbf{a} = \left(K^T K\right)^{-1} K^T V[k] \tag{9}$$

The solution (8) does not require the inversion of $K^T K$, but only one forward and back substitutions. It is therefore quite possible to apply it in real-time, especially when the degree of the polynomial is reasonably small (to avoid influence on the accuracy of the results due to stability reasons, the polynomial should be of second, or third degree). In the next step, we find the roots of p_l

$$p_l(\hat{t}_1) = a_0 + a_1\hat{t}_1 + \cdots + a_l\hat{t}_l = 0 \tag{10}$$

When the window is moved across the waveform, the times $\hat{t}_j$ will correspond to the approximate zero crossings of the waveform. The difference between every two odd, or even subscripted solutions will represent an integer number of periods of a quasi-steady state waveform

$$\hat{f}_j = \frac{1}{\hat{t}_j - \hat{t}_{j-2}}, \qquad k = 1, 2, ... \tag{11}$$

The *continualization* of the discrete set of estimated frequency (11) may be accomplished by piecewise linearization with surprisingly good results. The smoothing (polynomial fit) very efficiently suppresses noise, and harmonics could have an impact on the results only when their phase angles are fluctuating. It is possible to obtain the frequency information using this technique every half cycle, which is reasonably fast. The method performs very well under steady state conditions, and tracks frequency surprisingly well under transient conditions, as will be shown in the simulations. One potential problem is sensitivity of the method to switching transients in the signal, which may deteriorate its performance for up to 30 cycles following the transient. For steady state frequency measurments and non-switching transients, the method offers remarkable performance.

4 DFT Method with Polynomial Fitting

Given the measurment window (4,5), we calculate the phasor by a recursive Discrete Fourier Transform [1] (assuming non-varying phase angles in (4))

$$\Re\left\{\tilde{V}_1[k]\right\} = V_1 \cos\delta_1 = \frac{2}{N}\sum_{j=1}^{N} v[k+j]\cos\left(\frac{2(k+j)\pi}{N}\right) \tag{12}$$

$$\Im\left\{\tilde{V}_1[k]\right\} = V_1 \sin\delta_1 = \frac{2}{N}\sum_{j=1}^{N} v[k+j]\sin\left(\frac{2(k+j)\pi}{N}\right) \tag{13}$$

or, in matrix notation

$$\Re\left\{\tilde{V}_1\right\} = V[k]^T \cdot C[k] \tag{14}$$

and

$$\Im\left\{\tilde{V}_1\right\} = V[k]^T \cdot S[k] \tag{15}$$

where the vectors $C[k]$ and $S[k]$ are defined as

$$C[k] = \left[\begin{array}{ccc} \cos\left(\frac{2(k+1)\pi}{N}\right) & \cdots & \cos\left(\frac{2(k+N)\pi}{N}\right) \end{array}\right]^T \tag{16}$$

and

$$S[k] = \left[\begin{array}{ccc} \sin\left(\frac{2(k+1)\pi}{N}\right) & \cdots & \sin\left(\frac{2(k+N)\pi}{N}\right) \end{array}\right]^T \tag{17}$$

producing the stationary phasor

$$\tilde{V}_1 = M[k]^T \cdot C[k] + jM[k]^T \cdot S[k] = V_1 e^{j\delta_1} \tag{18}$$

The procedure is an effective filter for harmonics present in the signal. When all three phasors $\tilde{V}_{1a}[k]$, $\tilde{V}_{1b}[k]$, and $\tilde{V}_{1c}[k]$ (obtained from single phase measurements) are available, the positive sequence phasor is calculated as

$$\tilde{V}_1[k] = \tilde{V}_{1a}[k] + \alpha\tilde{V}_{1b}[k] + \alpha^2\tilde{V}_{1c}[k] \tag{19}$$

where

$$1 + \alpha + \alpha^2 = 0 \tag{20}$$

We calculate the phase angle $\delta[k]$ as the argument of $\tilde{V}_1[k]$

$$\delta[k] = arg\left\{\tilde{V}_1[k]\right\} \tag{21}$$

and construct the sliding data window

$$D[k] = \left[\begin{array}{cccc} \delta[k+1] & \delta[k+2] & \cdots & \delta[k+M] \end{array}\right]^T \tag{22}$$

We now fit the polynomial (6) using the least squares technique (7,8,9) and obtain the following *continuous* approximation of $\delta(t)$

$$\delta(t) = a_0 + a_1 t + \cdots + a_l t^l = \sum_{j=0}^{l} a_j t^j \tag{23}$$

The instantaneous frequency $f(t)$ is then calculated using (2)

$$f(t) = \frac{1}{2\pi}\frac{d\delta(t)}{dt} = \frac{1}{2\pi}(a_1 + 2a_2 t + \cdots + l a_l t^{l-1}) \tag{24}$$

and discretized again to represent the time tagged result at the instant $t = k\Delta t$

$$f[k] = f(k\Delta t) = \frac{1}{2\pi}\sum_{j=0}^{l-1} j a_j (k\Delta t)^{j-1} \tag{25}$$

The discrete frequency is evaluated at the end of the data window to provide the most recent estimate of the instantaneous frequency

$$f[k+M] = \frac{1}{2\pi}\sum_{j=0}^{l-1} j a_j \left[(k+M)\Delta t\right]^{j-1} \tag{26}$$

The attractive simplicity of this method makes it very easy for implementation in real-time. The noise and harmonics suppression is very good, as will be shown in simulations. Among the interesting properties of the method are:

- The continuous approximations of the instantaneous frequency are possible and logical.
- The instantaneous frequency prediction may be possible within reasonable range, by extrapolating the polynomial outside the data window.
- DFT efficiently suppresses harmonics.
- The least squares method smooths the effect of noise.

5 Demodulation Technique

We now briefly describe the frequency estimation by demodulation. This technique was recently examined in [14]. Note however that once we obtain the complex signal, we pursue a different approach.

Let one phase of the voltage waveform be

$$x[k] = A\cos(\omega_0 k + \phi[k]) + \epsilon[k],$$

where $\epsilon[k]$ represents higher harmonics and noise. The quantity that we estimate (the deviation from 60 Hz of the fundamental instantaneous frequency) is defined by

$$\Delta f[k] = \frac{f_s(\phi[k+1] - \phi[k-1])}{720}$$

where the phase $\phi[k]$ is expressed in degrees, and f_s is the sampling frequency in Hz. From $x[k]$ we form two new signals

$$y_1[k] = x[k]\cos(\omega_0 k),$$

and

$$y_2[k] = -x[k]\sin(\omega_0 k).$$

$y_1[k]$ and $y_2[k]$ carry the information about the instantaneous frequency in a high (around 120 Hz) and a low frequency signal components (around 0 Hz). We remove this redundancy and filter out the high frequency signal component by an appropriate lowpass filter. The filter yields

$$\tilde{y}_1[k] = \frac{A}{2}\cos(\phi[k]) + \tilde{\epsilon}_1[k]$$

and

$$\tilde{y}_2[k] = \frac{A}{2}\sin(\phi[k]) + \tilde{\epsilon}_2[k],$$

where $\tilde{\epsilon}_1[k]$ and $\tilde{\epsilon}_2[k]$ represent the filtered noise from $y_1[k]$ and $y_2[k]$. In the next step consider the complex signal

$$\tilde{y}[k] = \tilde{y}_1[k] + j\tilde{y}_2[k].$$

It can be shown that for high signal-to-noise ratios $\tilde{y}[k]$ can be approximated by [1]

$$\tilde{y}[k] \simeq \frac{A}{2}e^{j(\phi[k]+\phi_\epsilon[k])},$$

where $\phi_\epsilon[k]$ is phase noise. Now let

$$\begin{aligned} u[k] &= \tilde{y}[k+1]\tilde{y}^*[k-1] \\ &= u_1[k] + ju_2[k] \\ &\simeq \tfrac{A^2}{4}e^{j(\phi[k+1]-\phi[k-1]+\phi_\epsilon[k+1]-\phi_\epsilon[k-1])} \end{aligned}$$

From the above expression, we deduce that we can estimate the difference $\phi[k+1]-\phi[k-1]$ by

$$\phi[k+1]-\phi[k-1] \simeq \arctan\frac{u_2[k]}{u_1[k]}.$$

Therefore,

$$\Delta\hat{f}[k] = \frac{f_s}{720}\arctan\frac{u_2[k]}{u_1[k]} \qquad (27)$$

We can find $\Delta\hat{f}[k]$ from each phase of the signal. If the noise around 60 Hz in the three phases are independent, a reasonable estimate of the instantaneous deviation is the mean of the three estimates.

The procedure can easily handle any number of harmonics present in the measured data. In addition, low frequency components are also insignificant because they are filtered out. The only component that distorts the estimates is the noise at and around the fundamental frequency. This procedure is not computationally demanding. The filtering can be implemented by finite (FIR) or infinite impulse response (IIR) filters [15]. For accurate estimation these filters must have constant group delays at low frequences. FIR filters are easily designed to have this property for all frequencies. It turns out that there are IIR filters with this property too. Bessel filters have maximally flat group delays for low frequencies [13]. Their advantage is that we can achieve significant attenuations with very low order filters.

6 Simulation Results and Discussion

The three methods discussed in sections 3-5 have been tested by computer simulation. Several scenarios have been tried, and their performance documented. Whenever the term *true frequency* is used, it refers to the quantity (2) obtained from the signal, which is represented in the form (1). *Simulated transients* are disturbances of the waveform represented by the equation (1), with the following modulations and disturbances added:

- The tracking abilities need to be tested under transient conditions. A 1 Hz swing was modulated on the nominal frequency, with the maximum value of 1 rad/sec.
- Various amounts of measurement noise, modeled according to [2], were added to the signal. Typical values of standard deviation are of the order of 1 percent.
- Quantization noise was added to the samples. It corresponds to the 12-bit A/D converters used in the phasor measurement system. The sampling rate was 1440 Hz (24 samples per cycle), which allows for 12 harmonics to be present in the signal without aliasing effects.
- A subsynchronous oscillation of 6 Hz was modulated on top of the transient swingof 1 Hz. Even though this type of oscillation is not very common, it would test the tracking performance of the proposed methods in that important frequency region.
- Various amounts of harmonics were used in the signal. Three scenarios are presented in the paper: 5 percent 3rd harmonic, 5 percent distortion from the harmonics (3,5,7,9,11), and 25 percent distortion from the same group of harmonics.

The term *estimated frequency* relates to the results of application of the three proposed techniques for assessment of the true frequency (2,3). The estimated frequency is represented by formulae (11,26,27).

The results are shown in Figures 1–9. All the methods track the modulated swing very well, and only minor deviations have been observed in all scenarios with harmonic levels up to 25 percent. More pronounced was the noise sensitivity, especially when the modified zero crossing method (Figures 1–4) was used. This method is very good for frequency measurements of stationary waveforms, but cannot cope with large amounts of noise in the signal. Both DFT with least squares fitting and demodulation techniques successfully deal with large amounts of noise and harmonics, and could be used for dynamic frequency tracking under very difficult conditions. As expected, longer data sets used for fitting in DFT method produce better immunity to noise and harmonics. The best performance was obtained with 1 cycle DFT window for phasor calculation and 80 sample long window for quadratic polynomial fitting (Figures 5–8). It should be noted that $\frac{1}{2}$ cycle phase shift observed on all estimates based on DFT is due to one cycle measurement window – the estimates correspond to the middle of the measurement window, but they can be calculated only when the whole window is available.

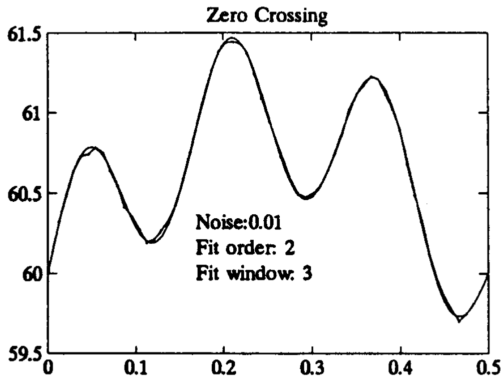

Figure 1: True and estimated frequency during the simulated transient (see text): modified zero crossing method: signal with 5 percent of the 3rd harmonic.

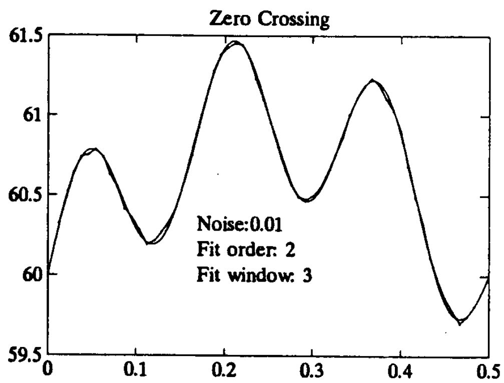

Figure 2: True and estimated frequency during the simulated transient (see text): modified zero crossing method: signal with 5 percent spread over odd harmonics (3rd-11th).

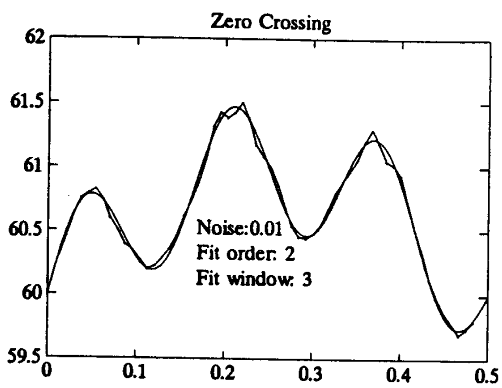

Figure 3: True and estimated frequency during the simulated transient (see text): modified zero crossing method: signal with 25 percent spread over odd harmonics (3rd-11th).

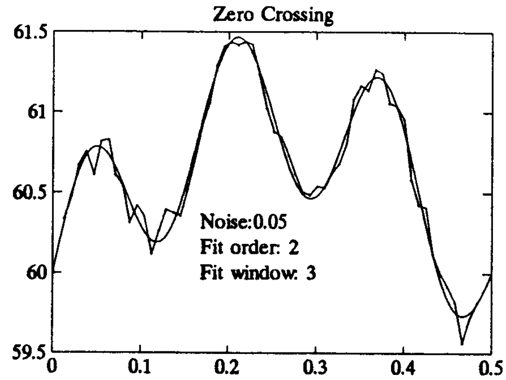

Figure 4: True and estimated frequency during the simulated transient (see text): modified zero crossing method: signal with 5 percent noise, no harmonics.

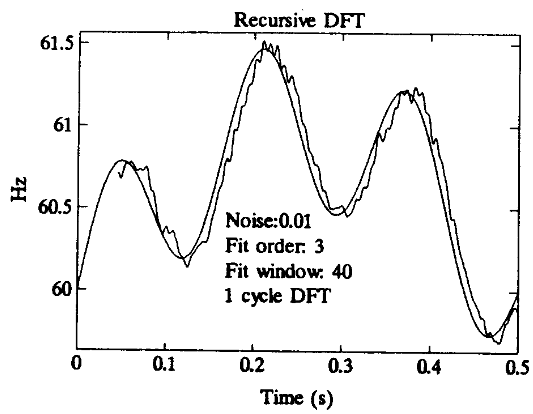

Figure 5: True and estimated frequency during the simulated transient (see text): DFT method with polynomial fit: signal with 5 percent 3rd harmonic.

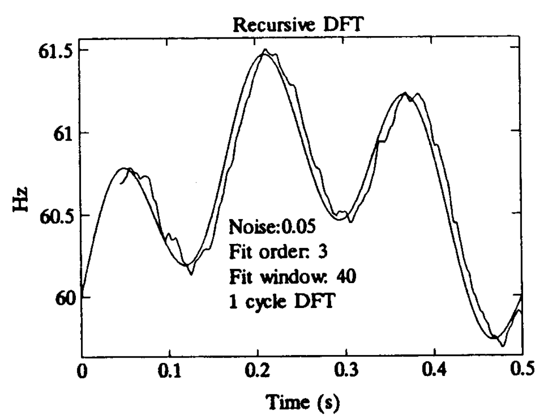

Figure 6: True and estimated frequency during the simulated transient (see text): DFT method with polynomial fit: signal with 5 percent noise, no harmonics.

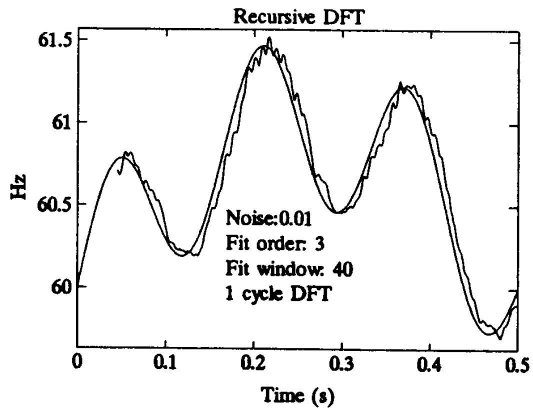

Figure 7: True and estimated frequency during the simulated transient (see text): DFT method with polynomial fit: signal with 5 percent harmonics (3rd-11th).

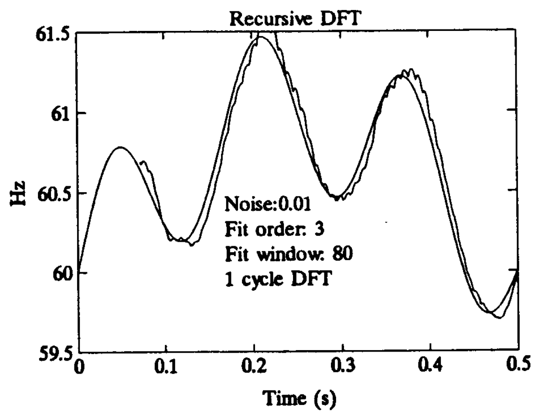

Figure 8: True and estimated frequency during the simulated transient (see text): DFT method with polynomial fit: signal with 25 percent harmonics (3rd-11th).

Some improvement is possible by using advanced extrapolation techniques, which is subject of the ongoing research. The demodulation technique (Figure 9) does not require long measurement windows and reaches excellent performance with a careful design of filters. The accuracy of the estimates is practically independent of the number of harmonics and their magnitudes.

7 Conclusions

Three algorithms for frequency measurement and tracking were proposed and tested – one is based on traditional zero crossing, modified by addition of curve fitting to suppress noise; the second is the Discrete Fourier Transform based method, also reinforced by polynomial fitting; the third is based on phase demodulation and subsequent filtering of the frequency deviation.

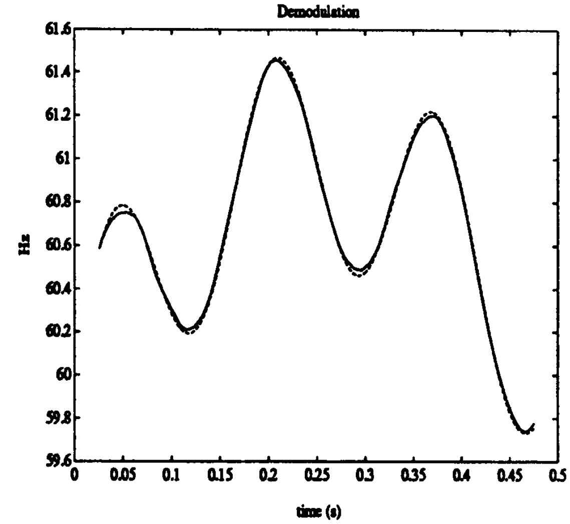

Figure 9: True and estimated frequency during the simulated transient (see text): demodulation method: signal with 25 percent spread over odd harmonics (3rd-11th).

The first algorithm, traditionally considered inaccurate and unreliable, produced surprisingly good tracking performance, and was particularly insensitive to the large amounts of harmonics in the measured signal, but has somewhat poor noise performance. Both DFT- and demodulation-based frequency tracking are capable of transient performance expected for monitoring of the real-time power system dynamics. Further improvements are possible with thorough analysis of the filtering options of the demodulation method.

The proposed techniques can be implemented on a phasor measurement system [1][2], which is centered around a 32-bit microprocessor, and utilizes GPS satellite receivers to provide accurate synchronization signals to the measurement computers. A configuration consisting of multiple phasor measurement units represents a distributed multiprocessor system with a potential for parallel processing of real-time data. Although PMUs are dedicated data acquisition units, careful utilization of the time windows between samplings, and optimization of the communication between measurement units would allow allocation of 20-30 percent of their time to custom data processing, such as the proposed frequency measurement techniques, thus creating an opportunity for real-time monitoring and control of system dynamics.

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