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USING DSP METHODS FOR ACCURATE DYNAMIC MEASUREMENTS IN POWER SYSTEMS

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Abstract - This paper presents the results of comparison of a new method of magnitude spectrum estimation of periodical signals in Power Systems, which uses Discrete Fourier Transform (DFT) and special FIR filters, with other methods of spectrum analysis, based on a preliminary estimation of the dynamically changed signal frequency. The main purpose of the suggested method is to reduce the leakage errors under conditions of desynchronization between the signal and the sample frequencies. Digital modeling of this method shows that accuracy of the estimation of magnitudes of the signal harmonics increases essentially by reference to other methods based on the interpolation of samples. This method can be used especially in Power System dynamics investigation (relay protection, UPS tuning, etc).

Keywords - Power Systems, spectrum, measurement.

1. INTRODUCTION

Accurate dynamic measurements of frequency and magnitude parameters are very important for Power System applications, such as relay protection, harmonic distortions estimation, UPS (Uninterruptible Power Supplies) tuning, etc. Authors of this publication investigated different DSP (Digital Signal Processing) methods for accurate estimation of APLF (Actual Power-Line Frequency) and APLH (Actual Power-Line Harmonics), especially in transient regimes typical for above mentioned applications.

The general review of DSP methods used in Power Systems was given in [1]. Some aspects of particular applications of DSP methods in Power Systems were described in [2] (using weight - integrating A/D converters for high rejection of power - line related noise), [3] (using DSP for average estimation of APLF), [4] (IZC - Integrating Zero - Crossing for APLF estimation with DSP and data acquisition applications), [5] (DFT- Discrete Fourier Transform for frequency and harmonics analysis with interpolation in time and frequency domains). Concurrent DSP methods for dynamic estimation of APLF were described in [6, 7]. They are based on using OPD – Orthogonal Phasor Decomposition. Comparative analysis of OPD and IZC methods for APLF estimation was presented in [8, 9] and is continued here. Last our investigations, published here, are dedicated to comparative analysis of prior time and frequency sample interpolation with prior matched filtering for reducing leakage related errors in APLH estimation.

Description of the interpolation technique, used in DFT, was given before in above mentioned publications and last one [10], and description of the matched filtering technique was shortly presented in [11]. Therefore, we accent here comparison between interpolation and filtering techniques as well as dynamic aspects of APLF and APLH estimation.

In the conclusion, the achieved results described in the paper have to be named and the efficacy of the method pointed out. The restrictions of the procedure, possibility and range of application of the results have to be stated.

In this paper we consider the case of the APLF and APLH estimation for the periodic multifrequency signal with main frequency f and sampling frequency f_s under noise and leakage conditions. The digital model of the signal, which consists of K harmonics and noise, is

$$s(n) = \sum_{k=1}^{K} A_k \cos(\theta_k n + \varphi_k) + \eta(n), \theta_k = \frac{2\pi kf}{f_s}, \quad n = 1,2...N \tag{1}$$

The frequencies of the harmonics equal f , $2f$ … Kf place within $(0 \div f_N)$, where f_N is the Nyquist frequency. The noise $\eta(n)$ is a sequence of normally distributed numbers with mean value equal to 0 and dispersion σ_n. In the following we consider amplitude of the first harmonic $A_1 = 1$ and values of the high harmonics are chosen randomly so that $THD = \sqrt{\sum_{k=2}^{K} A_k^2} / A_1 = const$. The digital measurements of frequency and magnitude parameters, considered here, are based on realization of the following methods: 1) method of orthogonal phasor decomposition with local determination of rotation speed through relation of two orthogonal components; 2) method of signal period estimation as the time interval between zero-crossings of integrated signal (IZC); 3) windowed DFT/FFT method, which uses sample interpolation in time or frequency domain for decreasing leakage, noise and frequency estimation errors. For accuracy estimation of these methods one can use dispersions σ_f and σ_{Ak} of f and A_k

measurements. In all following digital experiments 50 models of $s(n)$ with different randomly chosen values of φ_k, A_k and $\eta(n)$ were generated. Thus σ_f and σ_{Ak} are obtained as a result of 50 experiments.

2. COMPARATIVE ANALYSIS OF APLF ESTIMATION BY USING OPD AND IZC

2.1. The OPD method of frequency estimation

The main principles of this method one can see in Fig.1. This method is based on measuring the angle $\Delta\alpha = \alpha_{n+1} - \alpha_n$ of the phasor rotation between two next samples as shown in Fig.1a. FIR filters F1 and F2 (Fig.1b) are matched to the first harmonic of the signal $s(n)$ with windows corresponding to orthogonal cos and sin components

$$Wc(n) = \cos(\theta_0 n), Ws(n) = \sin(\theta_0 n),\ \theta_0 = \frac{2\pi f_0}{f_s} \qquad (2)$$

correspondingly, where $n = 0,1...N$ and f_0 is the fundamental frequency $f_0 = 50$ Hz. Value of N is a multiple of an integer part of the number of samples in the period of the frequency f_0, i.e.

$$N = Int(f_s / f_0)N_w, \qquad (3)$$

where N_w is an integer number of estimated periods. The input discrete signals of the estimator E are $u_1(n) = a_1\cos(\theta n)$ and $u_2(n) = a_2\sin(\theta n)$. It operates according to the following algorithm:

$$\alpha_n = \mathrm{arctg}\left(\frac{u_2(n)}{u_1(n)}\right),\quad \alpha_n = \mathrm{arctg}\left(\frac{u_2(n+1)}{u_1(n+1)}\right),$$
$$f = f_s\,\frac{\alpha_{n+1} - \alpha_n}{2\pi} \qquad (4)$$

As shown in [9], it is useful (for accuracy increasing) to choose n so that $u_1(n)$ and $u_1(n+1)$ have opposite signs (zero-crossing event). The characteristics of this method depends on N_w. Results of the OPD simulation for $f_s = 3200$ Hz, $SNR = 30, 50, 70$ dB, $N_w = 1,2...10$ show that σ_f changes from 0,25 to 0,02 for $SNR = 30$ dB and from 0,03 to 0,005 for $SNR = 50$ dB. Errors less than 0,005 depend not only on the noise level but also on accuracy of $u_1(n)$, $u_2(n)$ estimation (Fig.1b) including their magnitudes a_1, a_2. For $SNR = 70$ dB found that the value of σ_f is less than 0,006 for all N_w. Other simulation was carried out for $f_s = 3200$ Hz, $SNR = 70$ dB and different distortions (*THD* = 5, 10, 20%). One can show that for above *THD* values σ_f as a function of $N_w = 1,2...10$ changes correspondingly from 0,06 to 0,005 , from 0,18 to 0,01 and from 0,33 to 0,02 because of increasing the high harmonic suppression by F1 and F2 when N_w grows.

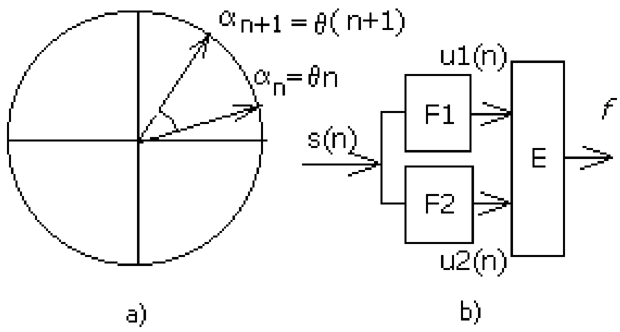

Fig.1 OPD method of frequency estimation.

2.2. The IZC method of frequency estimation.

Zero-crossing method of frequency estimation is based on measuring the distance Δn between two zero-crossing points (Fig.2a) with following frequency determination as

$$f = f_s / \Delta n, \qquad (5)$$

where Δn is obtained by interpolation in zero-crossing points from nearest positive and negative samples.

The main idea of IZC method is using of zero-crossing detection not for original signal, but for priory averaged (integrated) signal, that gives an essential reduction of additive noise and high harmonics influence. The scheme for this method is shown in Fig.2b, where F is the matched FIR filter with window $Ws(n)$ corresponds to (2) and zero-crossing detector (ZCD) which determines Δn and f according to (5). The results of σ_f estimation by IZC in presence of noise $SNR = 30, 50, 70$ dB show that σ_f as a function of $N_w = 1,2...10$ changes correspondingly from 0,001 to 0,0005, from 0,01 to 0,002 and from 0,06 to 0,01. Concerning the influence of high harmonics, IZC simulation shows that σ_f values are negligible up to *THD* = 10% and more less than σ_f values obtained for *THD* = 20% by OPD simulation.

Generally, concerning influence of additive noise and high harmonics APLF estimation by IZC is significantly simpler and more accurate than its estimation by OPD.

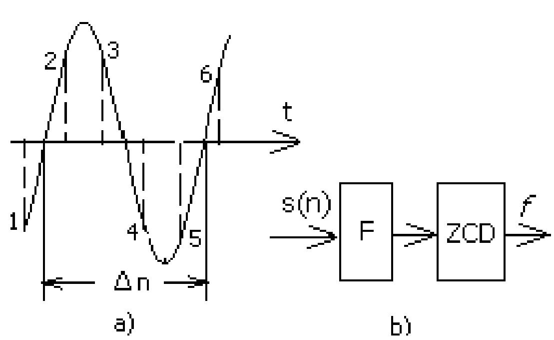

Fig.2. IZC method of frequency estimation.

3. COMPARATIVE ANALYSIS OF APLH ESTIMATION BY DFT/FFT WITH SAMPLE INTERPOLATION IN TIME AND FREQUENCY DOMAINS

3.1. Sample interpolation and resampling in time domain

The main aim of investigation of this method, named quasi-synchronous sample interpolation (QSI), is to study the dependence between accuracy of frequency measurement and accuracy of spectrum analysis. In the simulation program multifrequency periodic signal with random values of harmonic magnitudes was considered. Each realization of it was characterized by following parameters: number of harmonics 6 (including direct current component) ,THD10%, frequency of the signal $f_0 = 50$ Hz. Noise and quantization errors were eliminated. Results of the MATLAB simulation [9] are obtained for cases, where frequency f_0 is determined with relative error δ = 0,005%, 0,0l% ... 0,05% and f_s = 3200Hz, 6400Hz, 12800Hz. Relation to the first harmonic magnitude and 50 realizations of the signal are used in the simulation program for determination of σ_{A_i} values. One can see that frequency errors less than 0,005% don't cause increasing in spectrum analysis errors, so that accurate methods of frequency estimation, especially IZC, can be used in DFT/FFT with QSI. One can show that under above mentioned conditions and frequency deviation Δf =1 Hz relative errors of the first harmonic magnitude will be about 0,005-0,025%. For $\delta \le 0{,}05\%$ values of σ_{A_i} can be 0,005-0,01%.

On the other hand, APLH estimation by QSI can be suffered from dynamic errors caused by dynamically changed APLF. One can decrease these errors by frequency prediction. As shown in [2], the above mentioned errors can be approximately expressed as $e_d = \frac{1}{2f^2}\frac{df}{dt}$, where f is APLF value. It gives e_d about 0,02% for $\frac{df}{dt}$ =1 Hz/sec; for short term frequency prediction e_d can be decreased at least up to 0,002%.

3.2. Sample interpolation in frequency domain

Magnitude spectrum of harmonics estimation was simulated for f_s = 3200, 12800 Hz, estimation was limited to first 4 odd harmonics (1-st, 3-rd, 5-th, 7-th) and noise level *SNR* = 70, 50, 30 dB. Simulations were repeated 50 times and were performed for frequencies from 49 to 51 Hz. Obtained results of σ_{A_i} values of harmonics estimation are represented in the Table 1. As for comparison with DFT/FFT interpolated in time domain by QSI, obtained here results of APLH estimation for *SNR* = 70dB are approximately of the same order value that above mentioned results of APLH estimation with QSI, obtained for multifrequency signal without additive random noise.

One can note that obtained errors of APLH estimation by interpolation in frequency domain do not depend on the accuracy of APLF estimation, but achieving acceptable interpolation errors requires more longer observation time (at least 4 periods). On the other hand, as shown in [10], APLH estimation by QSI requires (for the same aim) use of oversampling (up to 5 times) for observation time of (1-1,5) periods. It means that search of other solution, alternative to above mentioned methods of APLH estimation by interpolation in time or frequency domain, is much desired.

Table 1. Errors of APLH estimation.

Sample frequency	N_{harm}	SNR = 70 dB	SNR = 50 dB	SNR = 30 dB
f_s = 3200 Hz	1st	0,005-0,025%	0,020-0,035%	0,12-0,2%
	3rd	0,15-1,3%	0,5-1,5%	2,5-3,6%
	5th	0,05-0,25%	0,2-0,35%	1,6-2,7%
	7th	0,05-0,25%	0,25-0,4%	2,1-2,9%
f_s = 12800 Hz	1st	0,005-0,025%	0,015-0,035%	0,07-0,12%
	3rd	0,1-1,4%	0,15-1,5%	1,25-2,4%
	5th	0,05-0,35%	0,15-0,35%	0,85-1,4%
	7th	0,05-0,25%	0,12-0,3%	1,1-1,4%

4. SHORT DESCRIPTION OF THE SUGGESTED ALGORITHM FOR APLH ESTIMATION

The basic version of this algorithm, which corresponds to synchronous sampling, uses following DFT operations:

$$u_{Sk}(n) = u(n)\sin(k\varphi n),\ u_{Ck}(n) = u(n)\cos(k\varphi n),$$
$$\varphi = 2\pi f_0 / f_s,\quad n = 0,1...N \tag{6}$$

$$s_k = \frac{a}{N}\sum_{n=0}^{N} u_{Sk}(n),\ c_k = \frac{a}{N}\sum_{n=0}^{N} u_{Ck}(n),$$
$$a = \begin{cases} 1, & k = 0 \\ 2, & k > 0 \end{cases} \tag{7}$$

$$A_k = \sqrt{s_k^2 + c_k^2}, \tag{8}$$

where $u(n)$ - samples of the signal, k - harmonics number, A_k - magnitude of k-th harmonic, N - number of samples in the period $T_0 = 1/f_0$, i.e. integer part of ratio $N_s = f_s / f_0$, where f_s - sample frequency, f_0 - initial value of APLF.

In case of synchronization, $N = N_s$, spectrum of signals (1) consists of frequency components $0, f_0, 2f ... f_N$, and averaging FIR filters with rectangular windows (7) have frequency response with zeros $K(f) = 0$ for all

f_0 harmonics. In case of desynchronization, when f_0 deviates from its initial value, the f_0 harmonics are not completely suppressed by filters (7) (leakage effect) and we must replace them by weighting averaging:

$$s_k = \frac{a}{N}\sum_{n=0}^{N} u_{Sk}(n)w(n),\ c_k = \frac{a}{N}\sum_{n=0}^{N} u_{Ck}(n)w(n), \qquad (9)$$

where $w(n)$ - weighting coefficients of a window of the special FIR filter. Its Z-transform has zeros at all harmonics kf_0:

$$P(z) = \sum_{i=0}^{N} c_i z^{-i} = k_0 \prod_{k=1}^{N'} [z^{-2} + d_k z^{-1} + 1], \qquad (10)$$

where $d_k = 2\cos(k\varphi)$, N' - integer value of $0{,}5N$ and $k_0 = \left(\sum_{i=0}^{N'} c_i\right)^{-1}$, $w(i) = k_0 c(i)$.

One can note, that this DSP modification gives suppression of all signal harmonics in case of f_0, f_s desynchronization for the same number of samples, as for above mentioned basic algorithm, located in the window $w(i)$, which is slightly different from regular prototype.

A single shortage of this algorithm is its dependence on accuracy of APLF estimation, which needs prediction and correction of above mentioned dynamic errors e_d for high rate of frequency change. In order to eliminate this dependence and associated errors, we need an improved modification of the suggested algorithm named FDI (frequency deviation invariant) filtering. It can be realized as a composition of 2 averaging filters, which suppress $if_0 - \Delta f$ and $if_0 + \Delta f$ frequency components, $i = 1,2...N'$, Δf is max value of APLF deviation. Schematically, FDI filtering characteristic is shown in Fig.3.

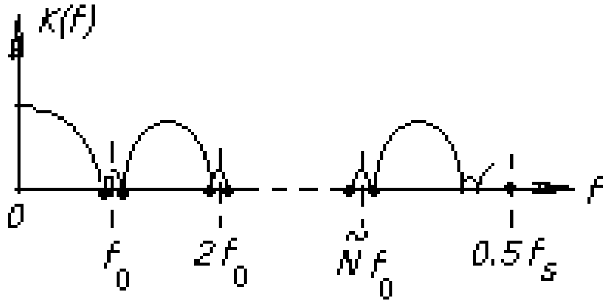

Fig. 3. Frequency response of the matched filter.

5. SIMULATION AND REAL TIME APPLICATION IN POWER SYSTEM INSTRUMENTATION

A simulation program in MATLAB has been used to estimate properties of the suggested algorithm. The program generates multifrequency periodical signal with nominal frequency f_0 = 50 Hz and deviation $\Delta f = \pm 0{,}3$ Hz, the number of harmonics was limited to 8; magnitude of the first harmonic was A_1 = 1 and magnitudes of the higher harmonics were randomly distributed with fixed *THD* = 20% for each realization of the signal (50 realizations were used). An accuracy of APLH estimation was determined as the max value of relative error $\delta_k = | A_k - \tilde{A}_k | / A_1$ %, where $\tilde{A}_k$ is an estimated value of A_k. Two methods with equal length of the signal were compared: the suggested algorithm (the upper line in Table 2) and the interpolation algorithm QSI (the lower line). For preceding f_0 estimation in both cases IZC method was used. As seen from Table 2 the accuracy of the suggested algorithm is essentially better (at least by 10 times) than the accuracy of the compared one, because of absence of interpolation errors. In spite of evident advantages of this algorithm with special filtering for decreasing leakage related errors it needs also the above mentioned improvement (FDI filtering) for dynamic change of APLF, especially in real-time applications.

Table 2. An accuracy comparison of the algorithms for spectrum measurement.

f_s (Hz)	**2000**	**3000**	**4000**	**5000**
Δf = 0,1 Hz	0,0131 0,4900	0,0013 0,2280	0,0036 0,1651	0,0018 0,0912
Δf = 0,2 Hz	0,0138 0,1963	0,0103 0,3350	0,0019 0,2707	0,0018 0,4040
Δf = 0,3 Hz	0,0124 1,5231	0,0160 0,7015	0,0044 0,0207	0,0013 0,2440

6. CONCLUSION

As shown above, the optimal choice of APLF, APLH for power system instrumentation is using IZC with matched FIR filtering for APLF estimation and using DFT/FFT with special FIR filtering (modification FDI) together with simplified orthogonal decomposition, for APLH estimation. Real time applications of above mentioned algorithms need further investigation.

7. ACKNOWLEDGEMENT

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