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ON–LINE INSTANTANEOUS FREQUENCY ESTIMATION AND VOLTAGE/CURRENT COHERENT RESAMPLING METHOD

An innovatory on–line method of power system frequency estimation and voltage/current signals resampling is presented. The method is designed to increase the accuracy of power system equivalent harmonic impedance estimation using simple DAQ systems. Estimation of instantaneous frequency uses 5–th degree polynomial interpolation of filtered signal zero crossings time positions. Final resampling uses cubic splines interpolation for the calculation of output signal values. The details of resampling algorithm are described.

The results of simulation tests demonstrating the typical behavior of a power system were shown as well as results of real voltage signal resampling. The method shows good capabilities of tracking varying harmonic phase angles in reference to the fundamental one while phase/frequency of fundamental harmonic varies too. The computational complexity allows to implement the method as a realtime version on DSP. Comparison of the amount of math operations used in this and another method [1] is shown too.

Keywords: frequency estimation, resampling, interpolation, coherent sampling, power system

1. INTRODUCTION

The requirement of model parameters determining for working power system often appears these days. One of these parameters is the equivalent impedance seen from the point of measurement. The knowledge of this impedance is necessary to calculate harmonic sources on the supplier and customer side, in diagnostics of the power network and for deciding on connecting or not a new load to the line.

Estimates of harmonic impedances base on increments of harmonic voltages and currents phasors, therefore they are especially sensitive to harmonic estimation errors. The increments during normal work of the system are very small, usually below 1%. I connection with low resolution of ADC, this causes significant estimation errors. Another important source of estimation errors is non-coherent sampling, typical for simple DAQ systems. Non-coherent sampling results in harmonic leakage and phase shift of voltage and current harmonics.

The first source of errors cannot be suppressed entirely by a non-invasive method, which uses only natural voltage and current waveforms without intentional disturbances of the system state. However these errors could be reduced using ADC with high resolution.

The second source of errors can be suppressed by coherent sampling (PLL synchronized ADC) or coherent resampling of interpolated non-coherent sampled signals.

Two variants (off-line and on-line) of the resampling algorithms have been developed. The methods and the results are presented in the following sections.

Most of the papers concerning signal resampling touch on interpolating function (reconstruction filter) selection [2, 3]. It is of great importance when aliasing may appear as a result of resampling [4]. In some of the described cases the sample rate conversion ratio is a constant, rational number and is not close to 1 [5]. Such values of sample rate conversion ratio are used for image scaling or for reducing the amount of digital information (e.g. for transmission or storage). Points at which signals have to be resampled are determined in such cases. The problem of finding such points occurs when input and output sample rates differ only slightly and the ratio is variable, which is the case in software radio and in coherent resampling of a power system voltage signal.

The original idea of this paper consists not in selecting the interpolating function but in a completely new way of determining time points at which the resampling of the power network voltage signal have to be performed. It is not a trivial task, as in the case of power network voltage signals the rate conversion ratio varies and is very close to unity. The finding of resampling time points uses interpolation of input signal zero crossing points. An alternative way of doing this is presented in [1]. As it can be assumed that the power network voltage is a band limited (low pass) signal and the output sample rate is very close to the input one, cubic splines are used for final resampling. It is yet possible to use other interpolating functions instead of cubic splines.

Another achievement is the proposal of a new, smooth estimator of instantaneous power system frequency based on above mentioned zero crossing detection. The estimates are smooth (free from noise) in comparison with other algorithms, e.g. [6, 7] and some algorithms presented in [8].

2. PROBLEMS WITH NON - COHERENCY IN HARMONIC IMPEDANCE ESTIMATION

A typical single - phase model of a power system is shown in Fig. 1.

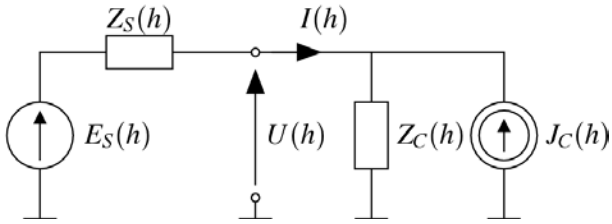

Fig. 1. Single - phase model of a power system for *h* harmonic.

Calculation of the equivalent system impedance $Z_S(h)$ requires $Z_S(h)$ and $E_S(h)$ to be invariant during measurement as well as knowledge of $U(h)$ and $I(h)$ in two different states of the power system [9]. The system states differentiation is the result of customer side ($Z_C(h)$, $J_C(h)$) parameters variability. This situation is described by the following divided differences

$$Z_S(h) = \frac{U(h) - U'(h)}{I'(h) - I(h)} = \frac{\Delta U(h)}{\Delta I(h)}, \tag{1}$$

where h denotes harmonic order, $Z_S(h)$ is estimated power system impedance, $U(h)$, $I(h)$ are voltage and current phasors adequately in the first state of the power system and $U'(h)$, $I'(h)$ are voltage and current phasors in the second state (Z, U, I are complex numbers).

One of the most important problems in estimating of $Z_S(h)$ is non - coherent sampling of voltage $U(t)$ and current $I(t)$ signals which are used then to determine the harmonics $U(h)$, $I(h)$. Coherent sampling of periodic input signal means that an integer number of the signal periods fits in the sampling window.

As it is commonly known, the frequency of the fundamental harmonic is not stable. This is related to a changing ratio of demand/delivery of power. Simple DAQ systems do not make use of PLL for sampling frequency synchronization. They work with the assumption that the fundamental frequency equals a nominal value. In majority of such DAQ systems in use signals are sampled non-coherently.

Non-coherent sampling causes two effects: harmonic leakage and phase shift of harmonics.

The harmonic leakage effect is manifested as magnitude and phase errors of voltage and current harmonics. Although this effect could be limited by proper selection of the shape and length of the window, the problem becomes important when the estimated parameter (here $Z_S(h)$) depends on differences of phasors as in (1), as the differences of phasors in real power system are often small.

The problem of harmonics phase shift arises when consecutive time windows begin with different time offsets from the zero crossing of the fundamental harmonic. When the fundamental frequency is constant and differs from the nominal (e.g. 50 Hz), the phase shift between harmonics and the beginning of the sampling window is linearly increasing/decreasing. This is not a problem when a two-terminal network model is considered, because the numerator and denominator have an identical phase error ξ which compensates as follows:

$$\frac{U}{I} = \frac{|U|e^{j\xi}}{|I|e^{j(\xi-\varphi)}} = \frac{|U|}{|I|}e^{j\varphi} = |Z|e^{j\varphi} = Z . \tag{2}$$

In reality the two-terminal network cannot be used because of unknown source voltage E_S. Then one has to use equation (1) to calculate Z_S. And again, the problem becomes important when using differences ΔU and ΔI of phasors. Unfortunately in general:

$$\frac{\Delta U}{\Delta I} = \frac{|U|-|U'|e^{j(\alpha+\xi)}}{|I'|-|I|e^{j(\beta+\xi)}}e^{j\varphi} \neq \frac{|U|-|U'|e^{j\alpha}}{|I'|-|I|e^{j\beta}}e^{j\varphi} = |Z|e^{j\varphi} = Z . \tag{3}$$

In (2) and (3) the harmonic index h has been omitted for simplicity, φ denotes the phase angle between current and voltage phasors, α, β denote additional phase changes in U and I between two states of the power system.

The influence of both effects on impedance estimation accuracy is considerable, especially for high order harmonics. The necessity of decreasing equivalent harmonic impedance estimation errors requires coherent signal sampling (using e.g. PLL) or signal resampling [10].

3. RESAMPLING METHOD

The input signal $x(t)$ is sampled at constant rate $\frac{1}{T_S}$, with the assumption that one period of the fundamental harmonic of $x(t)$ fits exactly M samples $x(n)$. This assumption may be not fulfilled, as the period of the fundamental harmonic in a power system may differ from nominal value. Thus resampling of samples $x(n)$ has to be done to get M samples $y(k)$ per period. The key issue is the way of calculating of time positions at which the sampled input signal has to be resampled. These positions will be called *resampling moments*.

All the proposed algorithms use detection of zero crossings of the input signal to calculate resampling moments.

3.1. Zero crossing detection

All variants of the method base on period length measured as the time between two consecutive downward zero crossings. The time moment c_i of i-th zero crossing is described by

$$c_i = n_i T_S - d_i , \tag{4}$$

where n_i is the number of sample following the zero crossing moment, T_S is sampling interval of input signal $x(t)$, d_i is time interval between the zero crossing and $n_i T_S$ as shown in Fig. 2. The value of d_i is found using linear interpolation as below:

$$d_i = T_s\left(1 - \frac{x(n_i - 1)}{x(n_i - 1) - x(n_i)}\right). \tag{5}$$

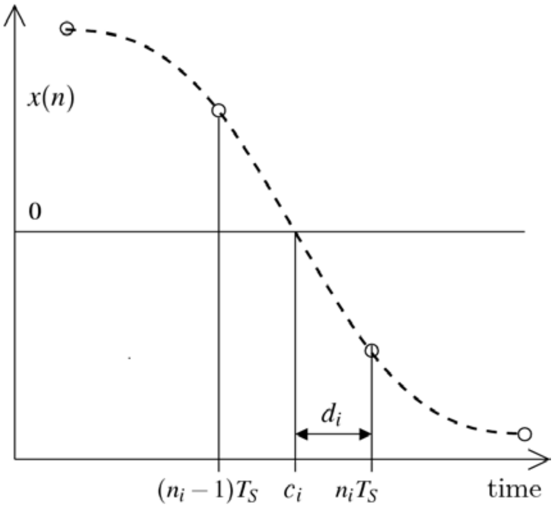

Fig. 2. Detection of *i*-th downward zero crossing instant c_i of the original signal $x(t)$ (dashed line) basing on its samples $x(n)$ (circles).

The length p_i of i - th period of the signal is given by

$$p_i = c_i - c_{i-1}. \tag{6}$$

Signal distortion may cause additional undesirable zero crossings. Therefore, in most cases, signal filtering is required. The filter should be of linear phase and should pass only the fundamental harmonic of the signal. A fragment of the frequency response of the 400-th order FIR filter we used is shown in Fig. 3.

However, when the signal distortion is small, the zero crossing detection can be applied to a raw signal.

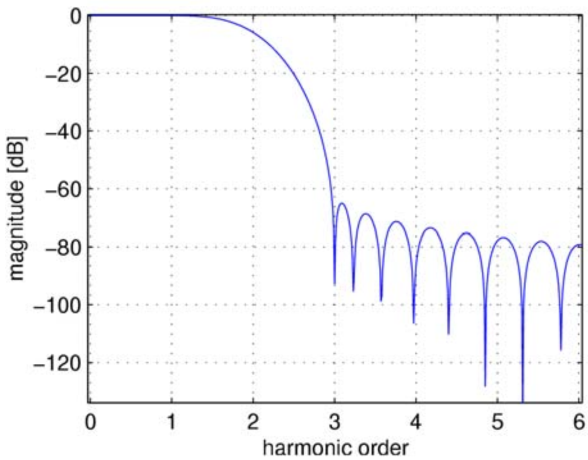

Fig. 3. Frequency response of the FIR filter.

3.2. Variant 1: constant frequency between two following upward zero crossings, off-line

The simplest approach is assuming that the frequency is constant in p_i period. Off-line version initially calculates consecutive zero crossings in the loop. After finding the i-th zero crossing, the p_i period is divided into M equal intervals. Ends of intervals describe equally distributed time moments t_m at which the interpolation of the original signal $x(n)$ is performed. Relative positions of t_m moments (from the beginning of i-th period i.e. from c_{i-1}) are described by formula:

$$t_m = \frac{m}{M} p_i , \tag{7}$$

where $m = 0 \ldots M - 1$.

Absolute positions t_k (from first detected zero crossing c_0) of interpolation time moments are characterized as:

$$t_k = c_i + \frac{mod(k,M)}{M} p_{i+1} \quad i = \frac{k - mod(k,M)}{M}, \tag{8}$$

where $mod(k,M)$ is remnant after division k/M.

Output signal samples $y(k)$ are values of cubic spline passing through input signal samples $x(n)$ calculated at resampling moments t_k (Fig. 4).

Unfortunately as the frequency value is assumed to be constant during one period it may have discontinuities at zero crossing. There can be also discontinuities in the output signal $y(k)$.

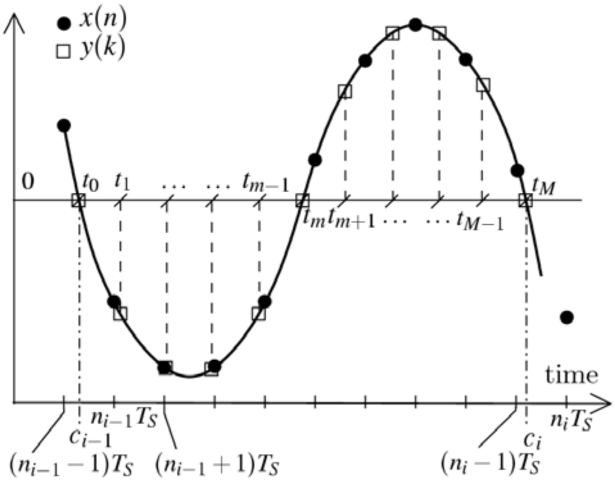

Fig. 4. Resampling of $x(n)$ by interpolation.

3.3. Variant 2: variable frequency between two following downward zero crossings, off-line

As the frequency of the fundamental harmonic in the power system varies in a continuous way, one may consider non-uniform placing of t_m resampling moments during one voltage/current period.

Equation (8) is no longer valid as the resampling moments t_k are now calculated using cubic spline interpolation. Zero crossings c_i are values of equally spaced interpolation nodes. The nodes interval equals M, it is one period of the input signal fundamental harmonic, as we want to get exactly M samples in each period. Then resampling moments t_k are calculated as interpolative values at M equally distributed points in each of all periods. The relationship between t_m and t_k consists in that t_m concern only one period of fundamental harmonic, while t_k are concatenated and unwrapped resampling moments t_m from all consecutive periods. The idea of this interpolation has been shown in Fig. 5.

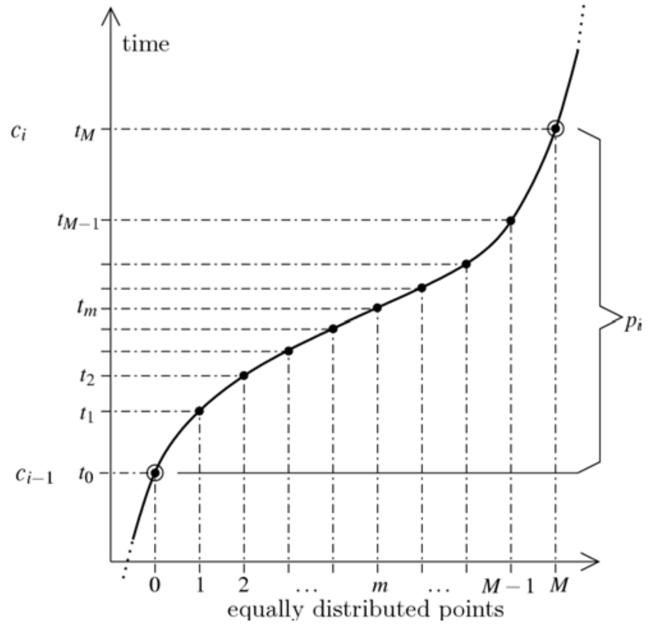

Fig. 5. Interpolation of c_i (zero crossing positions in time) in order to calculate non-uniform placed t_m points.

As in the previous variant, after calculating all t_k resampling moments, the final cubic splines interpolation (resampling) is performed on the whole signal $x(n)$.

This approach is a generalized form of the variant 1. Performing of linear interpolation of zero crossings positions c_i is equivalent to assuming constant frequency within each period.

Having calculated t_k resampling points we can construct a smooth estimator of instantaneous frequency of the fundamental harmonic

$$f(t_k) = \frac{1}{M(t_k - t_{k-1})}. \tag{9}$$

3.4 Variant 3: the on-line version

A drawback of two presented above off-line variants is that they operate on the whole signal. Thus the amount of data used may exceed available memory capacity. The weakness of the previously presented method can be overcome by working on subsets of input signal samples $x(n)$.

The on-line variant works in a loop. It processes the input signal sample by sample and after each downward zero crossing it returns one period (M samples) of the resampled signal. Each loop run consists of the following steps:

1. In this method a sample of the input signal $x(n)$ is read and stored in a buffer of the length of $3M$ samples. The oldest sample is forgotten. If there was not zero crossing between points

$x(n-1)$ and $x(n)$ step 1 is done again. If the downward zero crossing c_i was detected step 2 has to be performed.

2. Period length p_i is calculated (as described in sect. 3.1). The zero crossing c_{i-2} becomes new time origin $t=0$. Lengths of last three periods p_{i-2}, p_{i-1}, p_i stored in memory are utilized to make vector *C* containing times of last four zero crossings relative to $t=0$. The form of vector *C* is: $[-p_{i-2} \quad 0 \quad p_{i-1} \quad p_{i-1}+p_i]$. Auxiliary vector $N=[-M \;\; 0 \;\; M \;\; 2M]$ is also formed.
3. Pairs of vectors elements (N_j, C_j) become interpolation nodes. The nodes are equally spaced by distance *M* so $N=[-M \;\; 0 \;\; M \;\; 2M]$. Fifth degree polynomial *I* on nodes (N_j, C_j) $j=0ots3$ is calculated. Then *M* non-uniformly spaced resampling moments t_m are found as the interpolant *I* values at points $[0 \;\; 1 \; ... \; m \; ... \; M-1]$ (Fig. 5). Resampling moments t_m are the points at which the input signal $x(n)$ has to be resampled.
4. The final resampling (interpolation) of the p_{i-1} period of the input signal samples $x(n)$ is performed. To reduce the computing time, the interpolation uses only input signal samples within and nearby p_{i-1} period, that is values $x(n)$ where $n=[\mathrm{floor}(t_0/T_S)-3; \mathrm{ceil}(t_M/T_S)+3)]$. Moreover, as the time origin $t=0$ may be placed between $x(n)$ samples, the correction of samples time position is necessary before final interpolation is done. The new time positions of interpolation nodes are $nT_S - d_{i-2}$, where d_i is the offset from time origin $t=0$ (c_{i-2}) to the nearest following sample of $x(n)$ (Fig. 5). The final interpolation uses cubic splines to evaluate the pack of *M* samples of the output signal $y(k)$. The *M* samples are returned to the output. After that, a next run of the loop is performed.

Some details have to be explained here. As this is an on-line variant, the interpolation (see point 3.4 above) used to calculate resampling moments t_m differs from that described in sect. 3.3. During each loop run the interpolant *I* is calculated using four nodes saved in vector *C* at positions saved in vector *N*. For the sake of simplicity let us denote C_j to be an element of vector *C*, N_j to be an element of the vector *N* and, where *j* changes from 0 to 3. The interpolation takes place between two zero crossings c_{i-2} and c_{i-1} is in the range N_1 to N_2. In the next loop run the interpolation range is moved one downward zero crossing forward. To ensure continuity of *I* as well as of the first and second derivatives of *I*, the values $I(N_2)$, $I'(N_2)$, $I''(N_2)$ in i-th loop run must be equal to $I(N_1)$, $I'(N_1)$, $I''(N_1)$ adequate in $i+1$ loop run. Central difference quotient of C_1, C_3 is used as approximation of first derivative at N_2 position. Approximation of second derivative at N_2 uses values C_1, C_2, C_3. In this way six conditions describing the fifth order polynomial *I* were obtained. They form a linear system:

$$\begin{cases} I(N_1) & = & C_1 \\ I(N_2) & = & C_2 \\ I'(N_1) & = & (C_2 - C_0)/2h \\ I'(N_2) & = & (C_3 - C_1)/2h \\ I''(N_1) & = & (C_2 - 2C_1 + C_0)/h^2 \\ I''(N_2) & = & (C_3 - 2C_2 + C_1)/h^2 \end{cases}, \qquad (10)$$

where h is an interval between nodes positions *N*. As *h* is constant (equals *M*) the matrix of above-mentioned system (10) is constant, so its inversion have to be calculated only once. Therefore calculation of interpolant *I* coefficients in each one loop run requires only one matrix-vector multiplication (matrix is of the 6×6 size).

The need to have continuous second derivatives is related to instantaneous frequency estimator form (9). Frequency estimation results are shown in Fig. 6. A higher degree interpolating polynomial has been also tested but it results in larger oscillations of frequency estimates.

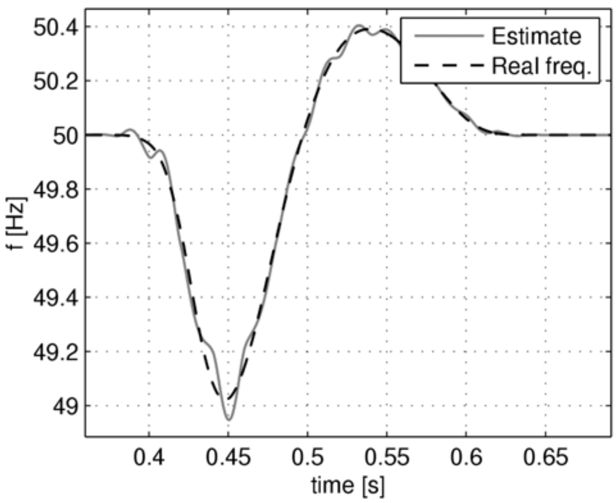

Fig. 6. Real and estimated frequency values.

As the $x(n)$ used for zero crossings detection is the filtered voltage signal $u(n)$, the $u(n)$ has to be equally delayed as the $x(n)$ is. Delaying may be done by an odd length $R + 1$ FIR filter having coefficients $h_j = 1$ when $j = R/2 + 1$ and $h_j = 0$ otherwise. The constant delay introduced by the filter has the length RT_S.

Moreover, input signal frequency deviation should be small enough to keep the fundamental harmonic in the pass band of the filter in use. Fortunately the fundamental voltage harmonic frequency in a normal state of the electric power system changes in small limits [11]. Norm [12] demands the fundamental frequency to be kept in the range 49.5 – 50.5 Hz during 95 % of a week.

The start of the algorithm needs assuming that the length of period p_i, before the first zero crossing detected, is equal to the nominal period of fundamental voltage harmonic in the power system. As it was pointed above, the method uses three last period lengths to calculate resampling positions t_m. This as well as initial assumptions concerning period length implies initialization needs four downward zero crossings of $x(n)$ to be detected before getting correct resampling positions.

It is impossible to determine the total delay of the output signal $y(k)$ in relation to the input signal $x(n)$, because input and output sampling rates are different. Moreover, the output samples are returned in packs of the length of M with variable rate equal to the rate of consecutive downward zero crossing of the input signal. It is important that the first $4M$ samples of the $y(k)$ should be discarded due to initialization.

4. TESTS

To check the correctness of the method (on-line variant) some tests was have been performed on various simulated signals. The test input signals consist of three harmonic of the order 1, 7 and 37 with constant amplitudes 1, 0.4, 0.25 and initial phase angles 0º, 90º, 60º respectively. Then the test signal was modified in the following way:

1. The real fundamental harmonic frequency f_1 of the test signal was constant, different from nominal f_{1N} (assumed during input signal sampling). f_1 = 50.3 Hz, while f_{1N} = 50 Hz. The 7-th and 37-th harmonics phases are linearly changed with reference to the 1-st harmonic phase. The 7-th one changes from 90º at 0 s to 45º at 2 s. 37-th changes from 60º at 0 s to 90º at 2 s.

2. Sinusoidal frequency modulation of input signal: f_1 was modulated with an amplitude of 0.5 Hz around mean 50 Hz with the modulation frequency of 1 Hz. the maximum rate of change of f_1 was relatively small – 0.15 Hz/s.
3. A sudden and relatively large frequency deviation was applied to the input signal. Such a phenomenon could be observed during starting of a big load (e.q. electric motor) connected directly to the transformer. When the temporary power demand during start-up exceeds the transformer capacity, a temporary „sag" of frequency can be observed at load terminals. After „sag" f_1 returns to the nominal value, as shown (dashed line) in Fig. 6. The maximum rate of change of f_1 was large [11] and reaches even 20 Hz/s.
4. The last test used a real 50 Hz signal, from a medium-voltage system, sampled at a 10 kHz rate. As the signal was sampled without any antyaliasing filter it isn't limited in frequency. The mean fundamental frequency of the signal was about 49.97 Hz. The test aim was to show how phase angles and amplitudes of harmonics are tracked before and after resampling.

4.1 Error calculation

The amplitudes and phases of given harmonics were measured on signals before (input signal *x*) and after (output signal *y*) resampling. The aggregated errors from tests 1 - 3 of measured values were calculated as described:

$$ae_X = \max_h \left| abs(X_h) - A_h \right|, \tag{11}$$

$$pe_X = \max_h \left| \arg(X_h) - \varphi_h \right|, \tag{12}$$

$$ae_Y = \max_h \left| abs(Y_h) - A_h \right|, \tag{13}$$

$$pe_Y = \max_h \left| \arg(Y_h) - \varphi_h \right|, \tag{14}$$

where X_h, and Y_h are complex Fourier coefficients of *h* harmonic of *x* and *y* calculated in one realization of FFT. The length of the time window equals *M* samples and it is one period of the nominal frequency f_{1N} . A_h and φ_h are *h* harmonic amplitude and phase. *ae* means amplitude error and *pe* means phase error. The errors values are shown in Fig. 7.

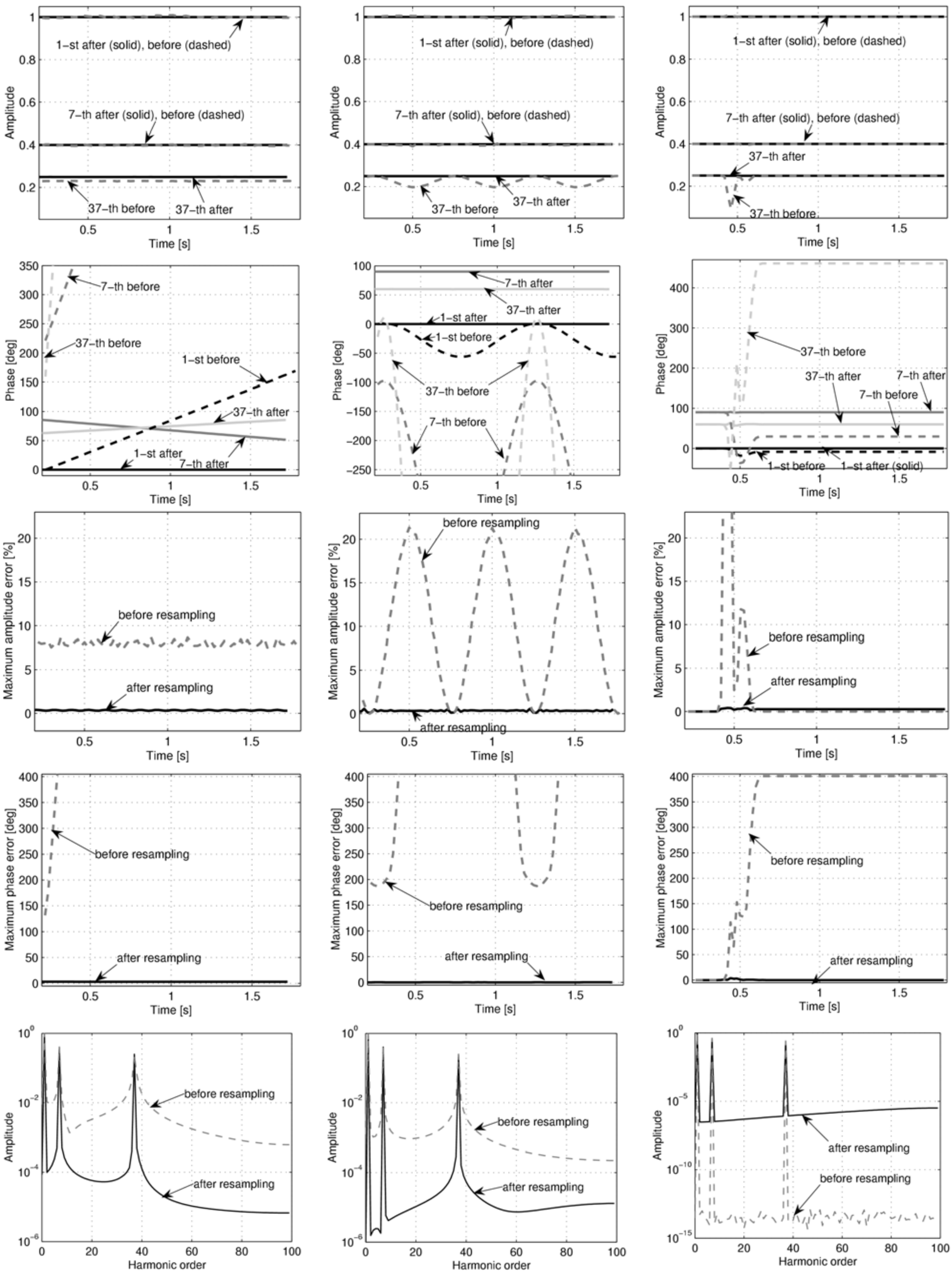

Fig. 7. Simulated measurement results of 1-st, 7-th and 37-th harmonic (before resampling - continuous, after resampling - dashed). Columns from left to right show tests 1 to 3 adequately. First row: amplitudes of harmonic; second: phases; third: max. amplitude meas. errors; fourth: max. phase meas. errors; spectrum of signals from selected realization.

4.2. Results

Test 1 (left column in Fig. 7) shows good tracking capabilities of variable phases of higher harmonics with the 1-st used as reference, even in the presence of permanent phase shift of the 1-st harmonic due to f_1 frequency different from nominal f_{1N}. The maximum phase error after resampling amounts to about 3. The maximum amplitude error after resampling does not exceed 0.4%.

Test 2 (middle column in Fig. 7) shows that mutual relations between phase angles of harmonics are kept after resampling, despite frequency modulation of the fundamental harmonic.

Test 3 (right column in Fig. 7) shows rather good performance of the algorithm in the presence of a sudden and relatively fast fundamental frequency change. It could be observed (second row, right column in Fig. 7) that phase angles of higher harmonics oscillate during such strong disturbance. After the disturbance frequency f_1 returns to the nominal value and then a constant, small error, about 0.01% of 37-th harmonic amplitude, occurs.

Test 4 results cannot be characterized through measurement errors, because real values of harmonics were unknown. But it can be seen in Fig. 8 that after resampling phase angles between 1-st and other harmonic are roughly constant, which is not the case for the signal before resampling. The amplitudes of harmonics before and after resampling have very close values, thus they have not been shown.

All tests use 10 kHz input signal sampling frequency and M = 200 samples per period of output signal fundamental harmonic.

Table 1 shows maximum harmonic measurement errors values calculated during 1.5 seconds long tests 1 - 3.

Tab. 1. Maximum errors values of harmonic measurement from tests 1 - 3 before and after resampling (phase values were unwrapped and may be greater than 360º).

	Test 1		Test 2		Test 3		Harmonic order
	before	after	before	after	before	after	
Max Amplitude Error [%]	0,9	0,001	0,58	0,0001	0,75	0,005	1-st
	1,2	0,013	2,45	0,01	1,35	0,01	7-th
	8,8	0,408	21,3	0,56	63,2	0,43	37-th
Max Phase Error [deg]	168	0,003	56,3	0,003	18	0,09	1-st
	1311	3,13	588	0,021	126	0,63	7-th
	6332	0,16	2166	0,16	401	3,3	37-th

4.3 Computing complexity comparison

There are two main approaches to signal resampling (interpolation). The first one uses polynomial interpolation and the second uses an interpolation filter having sinc impulse response. The Author used the first approach with cubic spline interpolant. Other kinds of interpolants (B-splines and modifications), presented in [5], were not considered here. Other methods of reducing harmonic leakage, using no interpolation, were proposed in [13].

An advantage of the sinc filter approach is that a banded signal keeps its band after resampling. That cannot be told in case of cubic splines interpolation. But the problem with FIR approach is that each output sample (each interpolation) requires calculation of all filter coefficients shifted slightly in time. There are some ways of reducing computing complexity. Calculation time can be reduced by decreasing filter length. Another way consists in precalculated FIR coefficients for some quantized time shifts. This causes some timing errors [1]. Both approaches worsen the spectral properties of ideal FIR filtration. The impact of filter coefficients quantization was studied in [14, 15].

A comparison of the number of multiplications and additions necessary for resampling one period of a signal using sinc FIR filter and cubic splines is shown in Tab. 2. The number of required operations was estimated with the assumptions that calculation of a polynomial value uses Horner's scheme, values of $sin(x)$ function are approximated using a power series of length 5, solving of linear system with tridiagonal matrix uses the optimal method.

Tab. 2. Amount of operations needed for calculating of one period of the output signal (*M* samples) using sinc filter (length *N*) and cubic splines

Type of operation	multiplications	additions
calculation coeffs. of *N* length FIR for *M* output samples	$6NM$	$4NM$
M times convolution of filter coefs. With input signal	NM	NM
overall math operations required for FIR interpolation	$7NM$	$5NM$
solving linear system of 4(*M*-1) equation with tridiagonal matrix	$20M$-24	$12M$-15
M times calculation of value of 3-rd degree spline (polynomial)	$3M$	$3M$
overall math operation required for cubic spline interpolation	$23M$-24	$15M$-15

It can be seen that comparable computing effort for both methods can be reached for FIR of the length about 3 . But such a short FIR filter cannot give good results of resampling.

The presented method (third variant) is an on-line one that does not necessarily mean realtime. In its present form, it can be used for resampling very long signals of voltage and current stored in files. The profit of the on-line variant is that it works on portions of signals instead of loading whole data to memory. Relatively small computing requirements, with reference to the capabilities of present-day DSPs and CPUs, allow to implement the presented method on DSP for realtime operation.

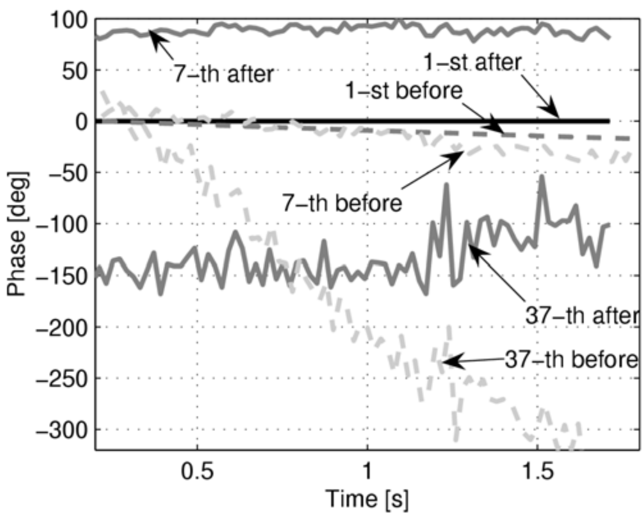

Fig. 8. Harmonics phase angles of the real signal.

5. CONCLUSIONS

The presented method joins well known techniques in an original way thus giving a tool for the estimation of harmonics phase angles and amplitudes of voltage and current as well as the instantaneous frequency of a power system. The tests proved good capabilities of tracking phase angles changes for fundamental harmonic and between harmonics, even in case of large phase deviation.

The novelty is the way of estimation of power system instantaneous frequency using polynomial zero crossings time moments interpolation. The estimate of frequency is smooth in comparison to others [6,7] and requires less computations than e.g. the nonlinear least squares method [16]. In that way it helps to calculate the proper time shift for each output sample.

The correction of non-simultaneous sampling of current and voltage is also possible. The last is needed for correct estimation of power system impedance and harmonic currents and voltages phase angles relations.

The on-line variant of the method can be adapted to realtime form used in DSP based measurement systems.

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METODA ESTYMACJI CHWILOWEJ CZĘSTOTLIWOŚCI ORAZ KOHERENTNEGO REPRÓBKOWANIA SYGNAŁÓW NAPIĘCIA/PRĄDU.

S t r e s z c z e n i e

W artykule przedstawiono innowacyjną metodę estymacji częstotliwości napięcia sieci energetycznej oraz sposób koherentnego repróbkowania sygnału napięcia sieci energetycznej.

Potrzeba przepróbkowania sygnału wynika z faktu, iż częstotliwość napięcia sieci energetycznej nie jest stała lecz zmienia się w sposób ciągły. Proste systemy akwizycji danych najczęściej umożliwiają wybór stałej częstotliwości próbkowania i nie pozwalają synchronizować tejże z aktualną częstotliwością mierzonego sygnału. Fakt ten powoduje rozmycie prążków widma częstotliwościowego sygnału i utrudnia ocenę faz harmonicznych sygnału. Błędy estymacji widma sygnałów sieci energetycznej powodują powstanie znacznych błędów pomiaru zastępczej impedancji sieci energetycznej metodą bierną prezentowaną w [10].

Proponowana metoda wykorzystuje nowatorski sposób wyznaczania punktów czasowych, w których powinien zostać zrepróbkowany sygnał mierzony. Do wyznaczenia czasowych punktów repróbkowania wykorzystano interpolacje kolejnych wartości czasów przejść przez zero mierzonego sygnału. Wariant, pozwalający na repróbkowanie sygnału na bieżąco, wykorzystuje wielomian piątego stopnia do wyznaczenia stałej, całkowitej liczby punktów repróbkowania przypadających na jeden okres badanego sygnału. Warunki nałożone na wielomian interpolujący zapewniają ciągłość jego oraz jego pierwszej i drugiej pochodnej, nawet w warunkach pracy na bieżąco, gdy nie są znane chwile czasu przyszłych przejść sygnału przez zero. Dzięki ciągłości pochodnych wielomianu możliwa jest estymacja chwilowej częstotliwości podstawowej harmonicznej mierzonego sygnału zgodnie z zależnością (9). Otrzymany estymator jest gładki w porównaniu do innych estymatorów bazujących na metodach statystycznych [6, 7]. Repróbkowanie sygnału badanego w uprzednio wyznaczonych punktach czasowych wykorzystuje interpolację funkcjami sklejanymi trzeciego stopnia.

Zaprezentowano wyniki testów symulacyjnych pokazujących typowe zachowanie systemu energetycznego oraz wyniki repróbkowania rzeczywistego sygnału napięcia sieci energetycznej dla kilku wybranych harmonicznych. Metoda ma dobre własności śledzenia zmieniających się kątów fazowych harmonicznych w odniesieniu do fazy podstawowej harmonicznej, nawet w sytuacji gdy faza lub częstotliwość podstawowej harmonicznej się zmienia. Złożoność obliczeniowa pozwala na implementację algorytmu na procesorze sygnałowym w celu zbudowania przyrządu pracującego w czasie rzeczywistym. Przedstawiono porównanie ilości operacji matematycznych potrzebnych do repróbkowania sygnału prezentowaną metodą oraz metodą wykorzystującą interpolację funkcją sinc.