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ELECTRIC POWER COMPONENTS ESTIMATION ACCORDING TO IEEE STANDARD 1459-2010
MERENJE KOMPONENTI ELEKTRIČNE SNAGE PO STANDARDU IEEE 1459-2010

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ABSTRACT

In this paper, the design and implementation of a novel recursive method for the power measurement ac-cording to the IEEE Standard 1459–2010 have been described. The most important parts are adaptive band- and low-pass FIR filters that extract fundamental and dc components, respectively. In addition, by using oversampling techniques and decimation filters, coefficient sensitivity problems of the large-order FIR comb cascade structure are overridden and the parameter estimation accuracy is improved. The sym-metrical components are estimated through a transformation matrix of adaptive phase shifters. The effec-tiveness of the proposed techniques is demonstrated by simulation results.

Keywords: *IEEE Standard 1459-2010, power estimation, cascaded integrator-comb (CIC) filter, oversampling, finite-impulse-response (FIR) comb filter, adaptive phase shifter, symmetrical components.*

REZIME

Merenje u nesinusoidalnim uslovima je u centru istraživanja i mnogo se napora ulaže da se pojam reaktiv-ne snage star više od sedamdeset godina definiše na nov način. Postoji niz pristupa rešavanju problema de-finisanja snaga i/ili pokušaja koncipiranja merne instrumentacije za merenje snaga u sistemima naizmenič-ne struje pod nesinusoidalnim uslovima. Jedini standard koji se odnosi na specifične zahteve za tačnost i odgovarajuće uslove testiranja u prisustvu harmonijskih izobličenja je IEEE Std. 1459–2010, koji ne daje definiciju reaktivne snage u nesinusoidalnim uslovima. Koncept ovog IEEE standarda je baziran na raz-dvajanju snage na fundamentalni i nefundamentalni deo. U literaturi su prisutne različite tehnike za imple-mentaciju standarda IEEE Std. 1459–2010. Ovaj standard je implementiran pomoću dva osnovna prilaza: (1) dvostepeni algoritam sa estimacijom harmonijskih spektara naponskog i strujnog signala u prvom ko-raku i računanjem nepoznatih komponenti snage u drugom koraku i (2) filterska implementacija kombino-vana sa Clarke–Park transformacijom u slučaju trofaznog sistema.U radu je prikazana nova metoda za merenje električnih veličina definisanih standardom IEEE 1459-2010 koristeći drugi pristup. Ključni ele-menti su adaptivni pojasni i niskopropusni FIR filteri koji izdvajaju fundamentalnu i jednosmernu kompo-nentu. U radu su korišćene tehnike oversemplinga i decimacionih filtera, čime se izbegavaju problemi ve-zani za osetljivost na zaokruživanje koeficijenata FIR kaskadnih filtera velikog reda, smanjuje obim nume-ričkih računanja i povećava tačnost merenja. Estimacija simetričnih komponenti vrši se pomoću matrice adaptivnih faznih korektora. U cilju procene performansi algoritma izvršene su računar¬ske simulacije i dati njihovi rezultati.

Ključne reči: *Standard IEEE 1459-2010, merenje snage, cascaded integrator-comb (CIC) filter, oversampling, filter sa konačnim impul¬snim odzivom (FIR) filter, adaptivni fazni korektor, simetrične komponente.*

INTRODUCTION

The concept of power components for sinusoidal single-phase voltages and currents and also for balanced three-phase sinusoidal voltages and currents is well defined. However, the liberalization of the energy market has emphasized a problem of the correct measure of electric energy under non-sinusoidal conditions. This matter is still being actively discussed, and there is not as yet any generalized power theory that can be assumed as a common base for billing purposes, power quality evaluation, harmonic sources detection, and compensation in power systems. Similarly, the standards do not give any specific accuracy requirements and related test conditions in the presence of harmonic distortion. However, metering instrumentation based on classical power theories and designed for sinusoidal waveforms can introduce measurement errors under nonsinusoidal conditions. Thus, the traditional billing quantities are emphasized, being defined as the fundamental active, reactive, and apparent powers, and the related power factor *(Cataliotti et al., 2008)*.

The IEEE Standard 1459-2010 gives definitions for power and energy measurement and their decomposition for designing and using metering instrumentation under sinusoidal, nonsinusoidal, balanced, or unbalanced conditions *(IEEE Std 1459-2010, 2010)*. In the IEEE Standard, the concept is based on splitting power into its fundamental and remaining terms. This approach of separating into fundamental and harmonic parts can be observed for the most important quantities, being normalized as indicators of power quality.

Different approaches were proposed in the literature to implement this standard. This standard has been implemented using two basic approaches: the two-stage algorithm with harmonic spectrum estimation in the first stage *(Chen, 2013; Gherasim et al., 2004; IJ & S Loureiro, 2015; Terzija et al., 2007; Tomic et al., 2010)*, and a time-domain filtering implementation combined with Clarke–Park transformation in the case of the three-phase system *(Emanuel & Milanez, 2006; Cataliotti & Cosentino, 2008; Pigazo & Moreno, 2007; Poljak et al., 2012)*.

Knowledge of the harmonic content of the signal data frame makes the calculation of the power definitions proposed by the IEEE Standard possible, but the standard does not suggest any measurement technique. The spectrum estimation of discrete-time signals (voltage and current) is usually based on procedures employing the fast Fourier transformation (FFT) *(Gherasim et al., 2004; IJ & S Loureiro, 2015)*. However, although the FFT is quite efficient under fixed-frequency conditions, it is well known that FFT loses its accuracy under non-stationary conditions, whereas the fundamental/harmonic frequency may vary over time. This is usually the practice in real measurement systems. To ensure the accuracy of FFT, the sampling interval of the analysis should be an exact integer multiple of the waveform fundamental period. For these reasons, a crucial point in the realization of an accurate measurement is the synchronization process. This is not an easy task in the presence of harmonic and interharmonic distortion.

In addition to disadvantages related to the synchronization of the sampling frequency with the frequency of the signal, the FFT has disadvantages caused by frame implementation. Thus, the FFT processes entire frames of data and cannot provide in-between data. If the calculation is done in a sliding mode, i.e., the FFT is repeatedly applied to a frame of N elements consisting of the last N–1 shifted elements of the previous frame and a single new element, FFT requires intensive computational effort, which complicates its integration on low-cost microcontrollers.

An instrument that is able to perform power measurements according to IEEE 1459 definitions by means of an adaptive resonator-based algorithm for harmonic analysis, together with an external decoupled module for frequency estimation is presented in *(Tomic et al., 2010)*. The resonator bank structure is an attractive tool for implementing transformations. On the basis of these structures, a simple algorithm for harmonic estimation with benefits in reduced complexity and computational efforts is obtained. This structure results in a significant reduction of the computational cost of the harmonic and overall estimation process. The computational effort is significantly less than that of FFT processing in sliding mode. This makes it easy to implement on digital signal processors or programmable logic arrays, using low-level embedded software. The algorithm on-line adapts to input signal frequency drift and overcomes all the synchronization-related problems.

A numerical algorithm for spectrum estimation that considers the system frequency as an unknown parameter of the model to be estimated and, in this way, solves the problem of sensitivity to wide-range frequency variations has been presented in *(Terzija et al., 2007)*. With the introduction of the power frequency in the vector of the unknown model parameters, the model itself becomes nonlinear, and the strategies of nonlinear estimation must be used.

On the other hand, the measurement of IEEE Standard 1459-2010 power quantities can easily be performed by means of time domain filtering techniques for fundamental and harmonics detection *(Emanuel & Milanez, 2006; Cataliotti & Cosentino, 2008; Pigazo & Moreno, 2007; Poljak et al., 2012)*. In the case of the three-phase systems, the Clarke–Park transformation usually should previously be applied. This technique is particularly appropriate for IEEE powers measurement, as it allows one to measure the fundamental and positive-sequence components of voltages and currents, without using any time-to-frequency transformation for voltages and currents spectral analysis. The computational burden is reduced, but the obtained measurements under non-sinusoidal conditions based on usage of low-pass filters for averaging can be erroneous due to low-pass filters (LPFs), as has been demonstrated in *(Pigazo & Moreno, 2007)*. To improve the measurement precision, high-order filters with specific cutoff frequencies, depending on the grid conditions, must be employed. The design process and the performance of the measurement method according to this structure are improved by replacing LPFs with a recursive averaging algorithm (RAA), which evaluates the average values. The algorithm is very efficient, and the computational burden is reduced. In this case, improved estimation fidelity has been achieved, but for frequency excursions, it suffers from the synchronization restrictions as FFT does.

In *(Poljak et al., 2012)*, an alternative filter-based measurement approach with the simple algorithm for simultaneous estimation of the frequency, the magnitudes of voltage and current, both the fundamental and the total (combined) active power, and fundamental reactive power is prescribed. Adaptive bandpass digital finite-impulse-response (FIR) filters are used to minimize the noise effect and to eliminate the effect of the presence of harmonics. To provide the fundamental reactive power estimation, an adaptive phase shifter has been applied to the voltage signal. The estimation of the total active power and true RMS voltage and current was carried out by applying adaptive low-pass FIR filters for averaging of the instantaneous values. The band- and low-pass FIR filter coefficients are adopted depending on the frequency. The good properties have been achieved by cascade connection of the antialiasing cascaded integrator-comb (CIC) filters with the cutoff frequency that is high enough to ensure fast response (up to 8-16 harmonics), and the reduced-order FIR comb filter that is applied on the decimated signal. This way, both coefficients sensitivity problems have been avoided and the parameter estimation accuracy has been improved.

In *(Poljak et al., 2012)*, the use of Clarke-Park transformation is avoided by symmetrical components estimation through transformation matrix of adaptive phase shifters, which makes it possible to get instantaneous symmetrical components independently on the frequency variation with modest computation requirements. The algorithm is not sensitive to power-system frequency changes and to the harmonic distortion of the input signals.

To demonstrate the performance of the proposed algorithm, computer-simulated data records have been processed. It has been found that the proposed algorithm is suitable for real-time applications.

MATERIAL

Basic Power Component Definitions

Single-phase Systems

Consider the following voltage and current functions with respect to time:

$$v(t) = V_0 + \sqrt{2}\sum_{h\neq 0}^{\infty} V_h \sin(h\omega_1 t + \varphi_{vh}) \quad (1)$$

$$i(t) = I_0 + \sqrt{2}\sum_{h\neq 0}^{\infty} I_h \sin(h\omega_1 t + \varphi_{ih}) \quad (2)$$

where $v(t)$ and $i(t)$ are the instantaneous voltage and current, V_h and I_h are the RMS values of the voltage and current harmonic h, and φ_{vh} and φ_{ih} are the phase angles of the voltage and current harmonic h, respectively. ($\omega_1 = 2\pi f_1$, f_1 is the fundamental frequency).

In the IEEE Standard, the concept is based on splitting power into its fundamental and remaining terms. This approach of separating into fundamental and harmonic can be observed for the most important quantities, which are being normalized as indicators of power quality. For example, the corresponding RMS values squared of current and voltage are as follows:

$$V^2 = V_1^2 + V_H^2;\ \ I^2 = I_1^2 + I_H^2 \quad (3)$$

where

$V_H^2 = \sum_{h\neq 1} V_h^2$, $I_H^2 = \sum_{h\neq 1} I_h^2$.

In addition, as a result of the RMS current and voltage separation into the fundamental 50-Hz part, and the non-50 Hz part, the apparent power can be expressed in the following manner:

$$S^2 = S_1^2 + S_N^2 \quad (4)$$

where S_1 and S_N are defined as fundamental and non-fundamental apparent power.

$$S_1^2 = (V_1 I_1)^2 = P_1^2 + Q_1^2 \quad (5)$$

where

$P_1 = V_1 I_1 \cos\varphi_1$, $Q_1 = V_1 I_1 \sin\varphi_1$.

φ_1 is the phase shift between the voltage and current signals of the fundamental component.

The nonfundamental apparent power S_N is an evaluator of harmonic pollution delivered or absorbed by a load and consists of three components, i.e.,

$$S_N^2 = D_I^2 + D_V^2 + S_H^2 = (V_1 I_H)^2 + (V_H I_1)^2 + (V_H I_H)^2 \quad (6)$$

The first term, i.e., current distortion power, D_I, reflects the power increase when the voltage source is perfectly sinusoidal and is usually the dominant component. The second term, which is called voltage distortion power, D_V, represents the effect of voltage distortion on the load bus. In addition, the third term, i.e., harmonic apparent power S_H, quantifies the energy due to cross-products of harmonic voltages and currents.

The harmonic apparent power can be further divided as follows:

$$S_H^2 = (V_H I_H)^2 = P_H^2 + D_H^2 \quad (7)$$

where $P_H = P - P_1 = \sum_{h\neq1} V_h I_h \cos\varphi_h$ is the total harmonic active power, and φ_h the phase shift between the voltage and current harmonic signals. The remaining component D_H is total harmonic nonactive power.

The total nonactive power N may also be defined in a conventional way, i.e.,

$$N^2 = S^2 - P^2. \quad (8)$$

The power factor is defined as the ratio of active power to apparent power, i.e.,

$$PF = P/S = (P_1 + P_H)/S \quad (9)$$

Three-phase Systems

In the general case of a three-phase system, including unbalanced and nonsinusoidal conditions, the effective current is defined as a function of the phase and neutral currents and the effective voltage as a function of the voltages from line to neutral and voltages from line to line. They are related to an equivalent virtual balanced system, which has the same losses as the real unbalanced system *(IEEE Std 1459-2010, 2010)*. The effective voltage and current can be resolved into fundamental and non-fundamental components like in a single-phase system. For three-phase systems, the power quantities are defined, starting from the effective apparent power resolution that is obtained by separating voltage and current fundamental positive sequence components from all other ones that are detrimental *(IEEE Std 1459-2010, 2010)*. Due to lack of space, it is not possible to give a complete overview of all definitions. If interested, the reader must directly refer to the IEEE Standard *(IEEE Std 1459-2010, 2010)*.

By introducing the RMS effective voltage and current in a three-wire system given as *(IEEE Std 1459-2010, 2010)*

$$V_{e,RMS} = \frac{1}{3}\sqrt{V_{ab,RMS}^2 + V_{bc,RMS}^2 + V_{ca,RMS}^2} \quad (10)$$

$$I_{e,RMS} = \frac{1}{\sqrt{3}}\sqrt{I_{a,RMS}^2 + I_{b,RMS}^2 + I_{c,RMS}^2} \quad (11)$$

the effective three-phase apparent power is defined as

$$S_e = 3V_{e,RMS} I_{e,RMS}. \quad (12)$$

The fundamental RMS effective voltage and current are defined in a similar manner as

$$V_{e1,RMS} = \frac{1}{3}\sqrt{V_{ab1,RMS}^2 + V_{bc1,RMS}^2 + V_{ca1,RMS}^2} \quad (13)$$

$$I_{e1,RMS} = \frac{1}{\sqrt{3}}\sqrt{I_{a1,RMS}^2 + I_{b1,RMS}^2 + I_{c1,RMS}^2} \quad (14)$$

The fundamental apparent power becomes

$$S_{e1} = 3V_{e1,RMS} I_{e1,RMS} \quad (15)$$

Having defined S_e and S_{e1}, the nonfundamental apparent power can be written as

$$S_{eN} = \sqrt{S_e^2 - S_{e1}^2} \quad (16)$$

In the three-phase system, the total active power is the algebraic sum of the active powers of all three phases (*a*, *b*, *c*), i.e.,

$$P = P_a + P_b + P_c \quad (17)$$

In the three-phase systems, the fundamental positive-sequence powers are the most important quantities. Power plants generate electric energy whose rate of flow consists of nearly pure positive-sequence power. The fundamental positive-sequence reactive power controls the voltage magnitude along the lines, stability, and power losses. Induction and synchronous motors convert in useful mechanical power only the positive-sequence active power; most of the negative-sequence and the harmonic active powers are converted in harmful forms of power *(Terzija et al., 2007)*. The IEEE Std. 1459 emphasizes the need to separate the positive-sequence fundamental apparent, active, and reactive powers

$$S_1^+ = 3V_1^+ I_1^+,\ P_1^+ = S_1^+ \cos\varphi_1^+,\ Q_1^+ = S_1^+ \sin\varphi_1^+ \quad (18)$$

and to monitor the fundamental positive-sequence power factor

$$PF_1^+ = \cos(\varphi_1^+) = P_1^+/S_1^+ \quad (19)$$

where $\varphi_1^+ = \varphi_{v1}^+ - \varphi_{i1}^+$.

METHOD

Proposed Algorithm

The general block diagram of the proposed algorithm is shown in Fig. 1. As can be seen, the structure consists of decoupled modules: the decimation filters, the adaptive band- and low-pass FIR comb filters, the fundamental frequency estimation module, the adaptive phase shifters and the modules that estimate the fundamental components of the voltage, current and active and reactive power. The obtained decoupled model is linear, which allows the use of linear estimation algorithms. Therefore, the derived algorithm is simple and requires reduced resources for implementation. Fig. 1 shows the general case of the three-phase system. For one-phase systems, the block diagram can simply be obtained by omitting an unnecessary module.

Decimation Filtering

The reduction of a sampling rate is called decimation and consists of two stages: filtering and downsampling. If a signal is not properly bandlimited the overlapping of the repeated replicas of the original spectrum occurs. This effect is called aliasing and may destroy the useful information of the decimated signal. That is why we need filtering to avoid this unwanted effect *(Jovanovic & Diaz-Carmo, 2011)*.

CIC (Cascaded-Integrator-Comb) filte *(Hogenauer, 1981)* r is widely used as the decimation filter due to its simplicity; it requires no multiplication or coefficient storage but rather only additions/subtractions. A z-domain $H_{CIC}(z)$transfer function of CIC filter for D-point moving-average process is

$$H_{CIC}(z) = \left(\frac{1}{D}\frac{1-z^{-D}}{1-z^{-1}}\right)^S \quad (20)$$

The most common method to improve CIC filter antialiasing and image-reject attenuation is by increasing the order *S* of the CIC filter using multiple stages. For the *S* CIC stages in cascade, the overall frequency magnitude response is the product of their individual responses, or *(Lyons, 2014)*

$$\left|H_{\text{CIC,Sth-order}}(e^{j2\pi f})\right| = \left|\frac{\sin(\pi f D)}{\sin(\pi f)}\right|^S \quad (21)$$

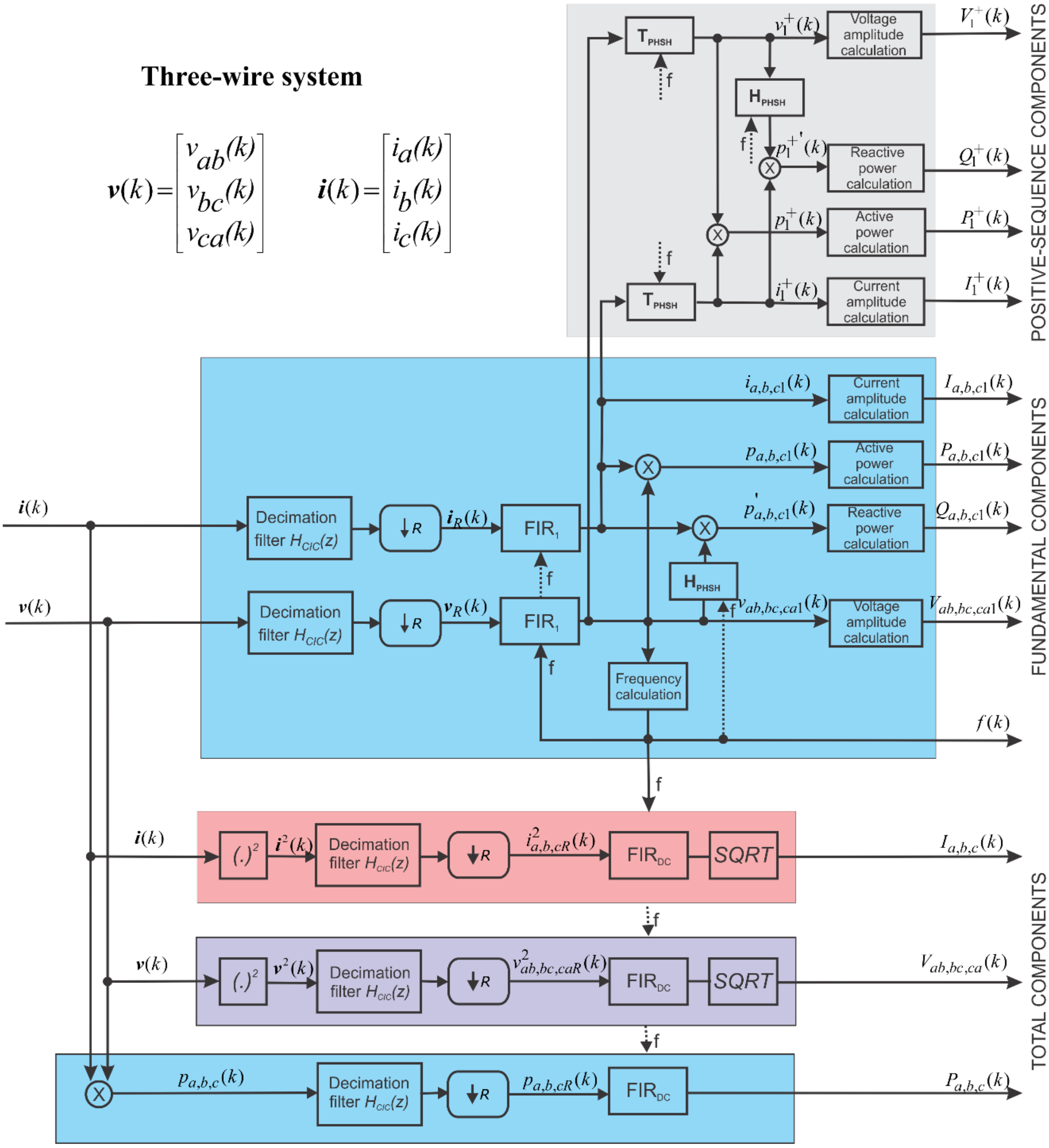

Fig. 1. Block diagram of the three phase system.

Fig. 2 shows the frequency magnitude responses of the first-, second- and third-order (S=1,2,3) CIC decimating filter. Notice increased attenuation of the third-order CIC filter compared with the first-order CIC filter in Fig. 2.

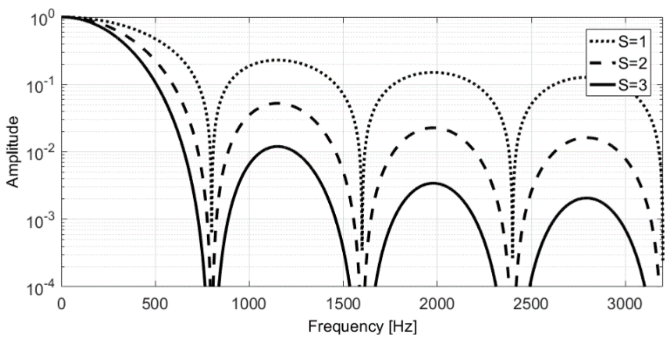

Fig. 2. Frequency responses of the CIC filters for a different number of stages in cascade and $fs = 6400\ Hz$ and D=8.

The CIC filters create errors in magnitude and phase spectrum responses, and they have drooping pass-band gains and wide transition regions. In order to eliminate the side effect of the proposed CIC filter, correction factors were provided.

In the case of signals with a low signal-to-noise ratio (SNR), the accuracy of the algorithm can further be improved by oversampling the input signal combined with high order antialiasing filters and decimation. Each doubling of the sampling frequency will lower the in-band noise by 3 dB, and increase the resolution of the measurement by 1/2 bit.

Adaptive FIR Filtering of Power System Signals

Both the troubling synchronization to the signal fundamental frequency and the use of restricting window functions could be avoided using adaptive digital FIR filters for sinusoidal signals *(Kusljevic, 2010b)*. The FIR digital filters are mainly used to process input voltage and current signals in order to minimize the effect of harmonics presence. It means that the frequency response of the filters must have nulls at the harmonics frequencies that are expected to be present in the signal. Thus, the estimation of the frequency requires the design of new filters at each iteration. The method proposed in *(Kusljevic, 2010b)* uses closed-form solutions for calculating filter coefficients. The complete filter can be realized as a cascade of the second-order subsections that eliminate the dc component and all harmonic frequencies, except the measured one, for which has to have unity gain.

The second-order subsection that eliminates the dc component and angular frequency $\omega_s/2 = \pi/T$ (ω_S is a sampling angular frequency) and has unity gain at the fundamental angular frequency ω_1 is given by the following Z-domain transfer function:

$$H_{10}(z) = \frac{1-z^{-2}}{|1-z_1^{-2}|}, \quad (22)$$

where $|1 - z_1^{-2}| = 2\sin(\omega_1 T)$.

The subsection that rejects the harmonic ω_i and has unity gain at the fundamental angular frequency ω_1 is shown as follows:

$$H_{1i}(z) = \frac{1-2\cos(\omega_i T)z^{-1}+z^{-2}}{|1-2\cos(\omega_i T)z_1{}^{-1}+z_1{}^{-2}|}, \; i = 2,3,\ldots,M \quad (23)$$

where $|1 - 2\cos(\omega_i T)\, z_1{}^{-1} + z_1{}^{-2}| = 2|\cos(\omega_1 T) - \cos(i\omega_1 T)|$ where M denotes the maximum integer part of $\omega_S/(2\omega_1)$. It is equal to the number of the subsections in the cascade.

The transfer function of the complete filter is given as follows:

$$H_1(z) = H_{10}(z) \prod_{i=2}^{M} H_{1i}(z) \quad (24)$$

It can be seen that a calculation of the filter coefficients can easily be implemented if the fundamental angular frequency ω_1 is known.

A lot of literature is available on frequency measurement techniques for power system applications. Some novel recursive methods for measurement of a local system frequency that allow measurement of the instantaneous frequency of real signals for single-phase or three-phase systems have been described in *(López et al., 2008)* and *(Kušljević et al., 2010)*. These methods are based on three consecutive samples in each phase and employed the least squares (LS) and weighted-least-squares (WLS) algorithm, respectively. In this paper, the WLS algorithm *(Kušljević et al., 2010)* is used.

The frequency response of the cascade of the CIC filter and filter (24) for the fundamental harmonic of f_1 = 49.8 Hz and sampling frequency of $f_s = 6.4$ kHz (128 samples per period T_0=1/50=0.02s) are given in Fig. 3.

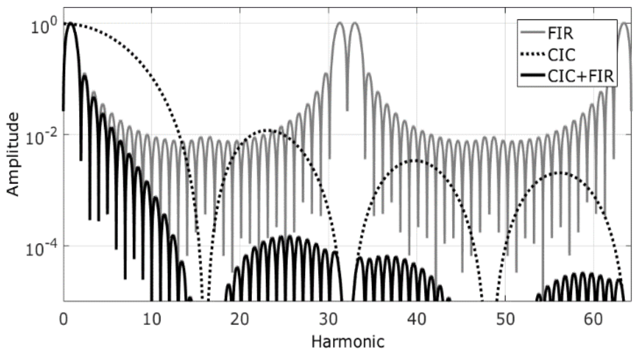

Fig. 3. Frequency responses of the fundamental component filter for the fundamental frequency $f_1 = 49.8$ Hz *and the sampling frequency* $f_s = 6.4$ kHz *and* $D = 8, S = 3, R = 4$.

After the sinusoidal signals of fundamental components are estimated, standard techniques for magnitude estimation could be used.

The instantaneous power *p* is given by the product of voltage *v* and current *i*. The dc component of *p* is the active power *P*, which contains, in addition to the fundamental active power component, harmonic active power components as a result of distorted current and voltage waveforms. The total active power can be measured by means of a low-pass filter. In the same way, the squared RMS values of voltage and current can be calculated by averaging the squared instantaneous voltage and current signals, respectively.

In *(Poljak et al., 2012)*, the performance of the measurement algorithm is improved by the use of the cascade of CIC and an adaptive low pass FIR filter that evaluates the average values of instantaneous power under a wide range of frequency deviations

$$H_{DC}(z) = H_{0DC}(z) \prod_{i=1}^{M} H_{iDC}(z) \quad (25)$$

where $H_{0DC}(z) = (1 + z^{-1})/2$, eliminates angular frequency $\omega_S/2$, and

$$H_{iDC}(z) = \frac{1-2\cos(\omega_i T)z^{-1}+z^{-2}}{2|1-\cos(\omega_i T)|}, \; i = 1,2,\ldots,M. \quad (26)$$

The frequency responses of the filters for dc component for fundamental frequency $f_1 = 49.8$ Hz and a sampling frequency of $f_s = 6.4$ kHz (128 samples per period T_0=1/50=0.02s) are given in Fig. 4.

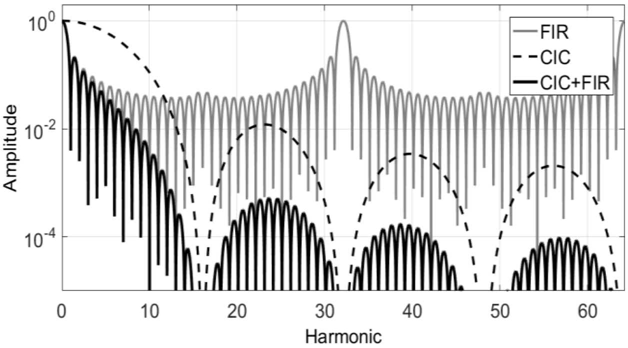

Fig. 4. Frequency responses of dc filter for the fundamental frequency $f_1 = 49.8$ Hz *and the sampling frequency* $f_s = 6.4$ kHz *and* $D = 8, S = 3, R = 4$.

Adaptive Phase Shifters

The measurement of reactive power under sinusoidal conditions, according to the IEEE/IEC definition, requires the introduction of an exact phase shift of $\pi/2$ between the voltage and current signals. A phase shift corrector with the following transfer function can perform a phase shift of $\pi/2$ on the fundamental frequency *(Rebizant et al., 2011)*:

$$H_{\pi/2}(z) = \frac{\cos(l\omega_1 T) - z^{-l}}{\sin(l\omega_1 T)} \quad (27)$$

where the delay l is between one and a quarter of the number of samples per period.

The frequency responses for a sampling frequency equal to $f_s = 800$ Hz (16 samples per period T_0=1/50=0.02s) and *l* varying from 1 to 4 are shown in Fig.5.

It is easy to notice that the filter gain equals unity for ω equal to ω_1 while the argument is equal to $\pi/2$ independent of parameter l. However, when ω is different from ω_1 the argument is not equal to $\pi/2$. It means that during frequency deviations from the nominal value the product of the method is not orthogonal to the signal.

The frequency response of the method depends on an assumed value of parameter l. The minimum value of this parameter equals one and its maximum value results from equation $l\omega_1 = \pi/2$, which depends on sampling frequency. They are a very good illustration of the well-known dilemma: speed–accuracy. As one can see, lower delay values induce stronger suppression of the dc component, but at the same time evoke stronger amplification of the second harmonic (effect of differentiation).

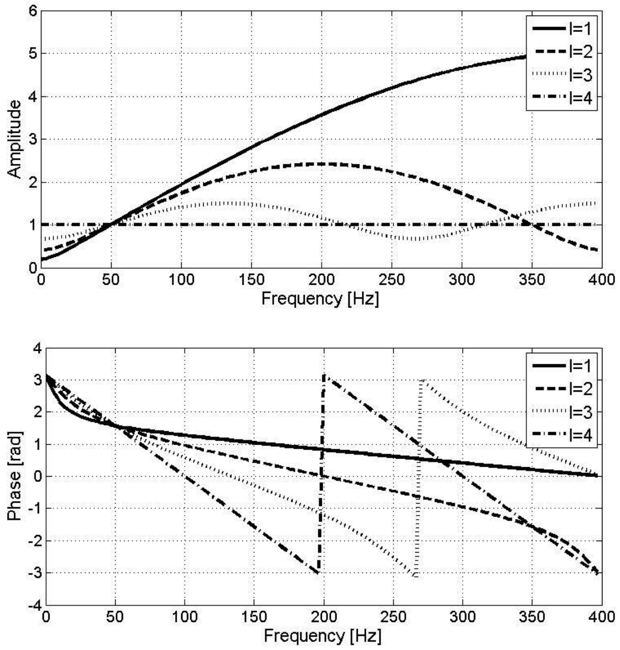

Fig. 5. Frequency responses of $H_{\pi/2}(z)$ for a sampling frequency equal to $f_s = 800$ Hz and l varying from 1 to 4.

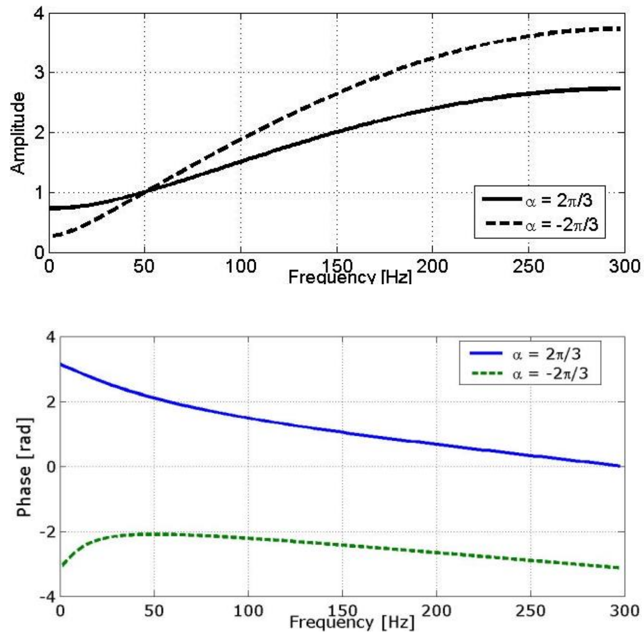

Fig.6. Frequency characteristics of the phase shifters for angle $2\pi/3$ and $-2\pi/3$ for for $l = 1$, f_1=50 Hz and f_S=600 Hz.

Fundamental Positive-Sequence Component Signal Estimation

Transformation of three-phase sinusoidal variables to symmetrical components is realized according to the known matrix equation of complex variables *(Rebizant et al., 2011)*:

$$\underline{\boldsymbol{V}}_{\boldsymbol{S}} = \underline{\boldsymbol{S}}\underline{\boldsymbol{V}}_{\boldsymbol{abc}} \tag{28}$$

where $\underline{\boldsymbol{V}}_{\boldsymbol{S}}$ is a vector of symmetrical components, i.e., zero, positive and negative sequence components, and $\underline{\boldsymbol{V}}_{\boldsymbol{abc}}$ is a vector of three-phase sinusoidal variables.

In turn, the complex matrix of transformation is the following:

$$\underline{\boldsymbol{S}} = \frac{1}{3}\begin{bmatrix} 1 & 1 & 1 \\ 1 & \underline{a} & \underline{a}^2 \\ 1 & \underline{a}^2 & \underline{a} \end{bmatrix} \tag{29}$$

with complex elements: $\underline{a} = e^{j2\pi/3}$; $\underline{a}^2 = e^{-j2\pi/3}$.

Analogous to the transformation of the phasors (28), the transformation of the three-phase signals can be defined as follows (Kušljević, 2007)

$$\begin{bmatrix} V_1^0(z) \\ V_1^+(z) \\ V_1^-(z) \end{bmatrix} = \frac{1}{3}\begin{bmatrix} 1 & 1 & 1 \\ 1 & H_{2\pi/3}(z) & H_{-2\pi/3}(z) \\ 1 & H_{-2\pi/3}(z) & H_{2\pi/3}(z) \end{bmatrix}\begin{bmatrix} V_a(z) \\ V_b(z) \\ V_c(z) \end{bmatrix} \tag{30}$$

where

$$H_{2\pi/3}(z) = \frac{2\sin\left(l\omega_1 T + \frac{2\pi}{3}\right) - \sqrt{3}z^{-l}}{2\sin(l\omega_1 T)} \tag{31}$$

$$H_{-2\pi/3}(z) = \frac{2\sin\left(l\omega_1 T \frac{-2\pi}{3}\right) + \sqrt{3}z^{-l}}{2\sin(l\omega_1 T)} \tag{32}$$

The phase shifters $H_{2\pi/3}(z)$ and $H_{-2\pi/3}(z)$ shift the phase of the fundamental signal component for $2\pi/3$ and $-2\pi/3$, respectively, and they are adaptive in subject to the radian frequency ω_1.

Both the magnitude and phase frequency responses for both $H_{2\pi/3}(z^{-1})$ and $H_{-2\pi/3}(z^{-1})$, for $l = 1$, are given in Fig. 6.

Fundamental Positive-Sequence Powers and RMS Voltage and Current Values Estimation

In *(Kusljevic, 2010b)*, a simple method, which is based on three consecutive samples of fundamental component signals, was used to estimate fundamental power components. This method can be also used for fundamental positive-components estimation. Let us suppose that the fundamental positive-sequence voltage and current signals, which are obtained by (30), have the following forms:

$$v_1^+(k) = \sqrt{2}V_1^+ \sin(\omega_1 kT + \varphi_{v1}^+) \tag{33}$$

$$i_1^+(k) = \sqrt{2}I_1^+ \sin(\omega_1 kT + \varphi_{i1}^+) \tag{34}$$

By performing a square of $v_1^+(k)$, it follows that

$$v_1^{+2}(k) = V_1^{+2} - V_1^{+2}\cos(2\omega_1 kT + 2\varphi_{v1}^+) \tag{35}$$

In the same way, by multiplying $v_1^+(k)$ and $i_1^+(k)$, it leads to the instantaneous fundamental positive-sequence power:

$$p_1^+(k) = 3v_1^+(k)i_1^+(k) = P_1^+ - S_1^+\cos(2\omega_1 kT + \lambda_1^+) \tag{36}$$

where

$S_1^+ = 3V_1^+ I_1^+$, $P_1^+ = S_1^+ \cos\varphi_1^+$,

$\varphi_1^+ = \varphi_{v1}^+ - \varphi_{i1}^+$, $\lambda_1^+ = \varphi_{v1}^+ + \varphi_{i1}^+$,

The fundamental positive-sequence reactive power Q_1^+ can be calculated by the multiplication of the voltage and current signals $v_1^{+\prime}(k)$ and $i_1^{+\prime}(k)$ defined as follows:

$$v_1^{+\prime}(k) = \sqrt{2}V_1^+ \sin(\omega_1 kT + \varphi_{v1}^{+\prime}) \tag{37}$$

$$i_1^{+\prime}(k) = \sqrt{2}I_1^+ \sin(\omega_1 kT + \varphi_{i1}^{+\prime}) \tag{38}$$

where the phases of the voltage and the current are changed such that an additional phase shift of –π/2 is introduced, that is, $\varphi_{v1}^{+\prime} - \varphi_{i1}^{+\prime} = \varphi_{v1}^+ - \varphi_{i1}^+ - \pi/2$.

By multiplying the voltage and current signals defined in (37) and (38), respectively, it follows that

$$p_1^{+\prime}(k) = 3v_1^{+\prime}(k)i_1^{+\prime}(k) = Q_1^+ - S_1^+\cos(2\omega_1 kT + \lambda_1^{+\prime}) \tag{39}$$

where

$S_1^+ = 3V_1^+ I_1^+$, $Q_1^+ = S_1^+\cos\varphi_1^{+\prime} = S_1^+\sin\varphi_1^+$,

$\varphi_1^{+\prime} = \varphi_{v1}^{+\prime} - \varphi_{i1}^{+\prime} = \varphi_1^+ - \pi/2$, $\lambda_1^{+\prime} = \varphi_{v1}^{+\prime} + \varphi_{i1}^{+\prime}$.

Three consecutive samples of the signal $v_1^{+2}(k)$ (the same is valid for $i_1^{+2}(k)$, $p_1^+(k)$ and $p_1^{+\prime}(k)$), which are sampled at instants $k-2$, $k-1$, and k, are connected with the following equation:

$$y(k) = z(k)x(k) \tag{40}$$

For the magnitude of the fundamental positive-sequence voltage component, we have

$$y(k) = v_1^{+2}(k) - 2v_1^{+2}(k-1)\cos(2\omega_1 T) + v_1^{+2}(k-2)$$

$$z(k) = 2(1 - \cos(2\omega_1 T)),\ x(k) = V_1^{+2}(k),$$

where $V_1^+(k)$ represents an estimation of V_1^+ at instant k. The estimation is based on measurements carried out until instant k.

For the fundamental positive-sequence active power, we have

$$y(k) = p_1^+(k) - 2p_1^+(k-1)\cos(2\omega_1 T) + p_1^+(k-2)$$
$$z(k) = 2(1-\cos(2\omega_1 T)),\ x(k) = P_1^+(k). \quad (41)$$

Finally, for fundamental positive-sequence reactive power, we have

$$y(k) = p_1^{+\prime}(k) - 2p_1^{+\prime}(k-1)\cos(2\omega_1 T) + p_1^{+\prime}(k-2)$$
$$z(k) = 2(1-\cos(2\omega_1 T)),\ x(k) = Q_1^+(k). \quad (42)$$

Equation (40) can be solved in the same way using the recursive WLS algorithm *(Kusljevic, 2010a)*. The results are weighted least-mean-square estimations of the quantities based on measurements until the instant *k*.

Performance Evaluation Through Simulation

The performance of the proposed technique has been evaluated using simulated waveforms. The inputs to the algorithm are three-phase voltage and current signals which were sampled at 4 kHz. The proposed algorithm was evaluated using simulated signals with dc components, high harmonic content, and superimposed noise.

An input sinusoidal test signal that is frequency modulated (with frequency step change from 50 to 49.8 Hz at *t*=0*s*) is processed. White noise with SNR of 60dB was added to sinusoidal signals. The obtained results confirm the good dynamics response of the algorithm for the frequency step change, as well as the accuracy of the quantity estimation. We have obtained a technique that provides accurate estimates for SNR=60 dB with frequency estimation errors in the range of 0.002 Hz and power component estimation errors in the range of 0.03% in about 25 ms, Fig. 7.

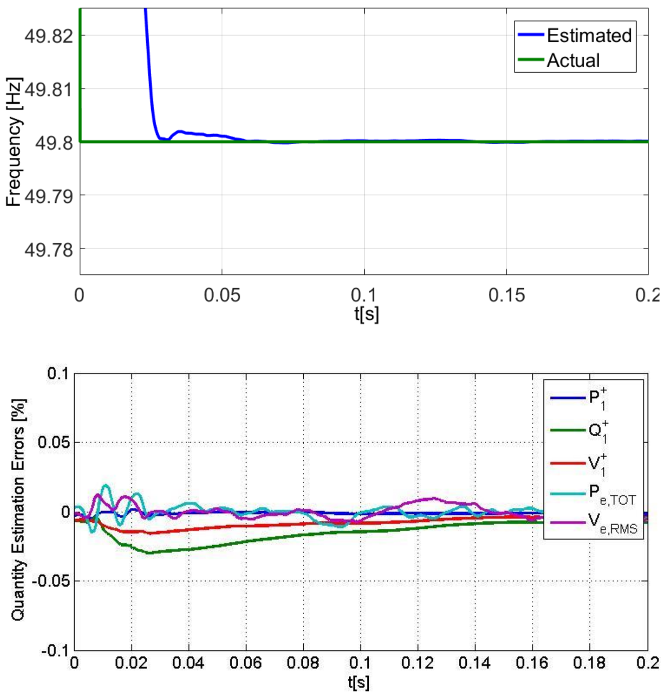

Fig. 7. Estimations for f=50Hz for t<0s and f=49.8Hz for t>0s, with SNR=60 dB.

One of the most interesting network disturbances is voltage dip on one or more phases, in most cases caused by a short-circuit fault, so that one or two phases can reach values equal or close to zero. The short phase voltage dip in phase *a* (zero voltage) initiated at *t*=0*s*. Fig. 8 shows the time response of dynamic simulation results for the proposed algorithm. The good dynamic responses and high accuracy of measurement can be noticed.

The effect of the presence of noise in the signals was studied by estimating the power components of signals that contain noise. A sinusoidal 50-Hz input test signal with the superimposed additive white zero-mean Gaussian noise was used as input for the test. The random noise was selected to obtain a prescribed value of the SNR, which is defined as SNR=20 log (A/(√2 σ)) where A is the magnitude of the signal fundamental component, and σ is the noise standard deviation. The maximum quantity estimation errors in terms of an SNR in the range 40-80 dB, with input signals frequency of 30, 50, and 70 Hz, are shown in Fig. 9. It should be noted that, in practice, the SNR of voltage signal obtained from a power system ranges between 50 dB and 70 dB. At this level of noise, very little error is expected with the proposed technique, as shown in Fig. 9.

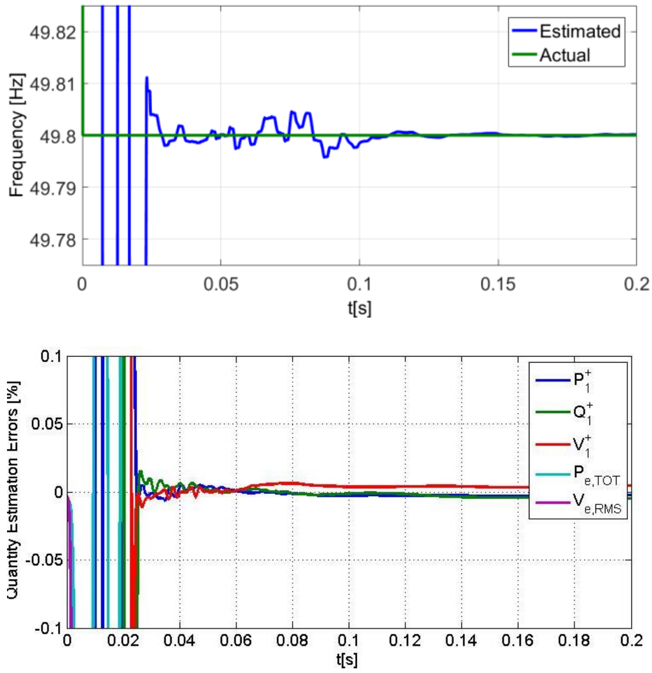

Fig. 8. Estimations for short phase voltage dip with SNR=60 dB and with harmonics presence.

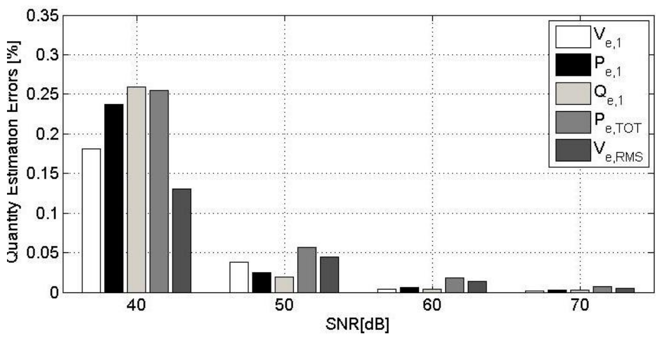

Fig. 9. Maximum estimation errors in terms of the SNR.

To demonstrate the effectiveness of the proposed technique in the presence of harmonics, an input sig-nal having the fundamental frequency of 50 Hz, a third harmonic component in the range of 0%-20% and a fifth harmonic component equal to half of the third component have been used. Fig. 10 shows the max-imum errors for the proposed technique when input signals of 30, 50, and 70 Hz having 0%, 5%, 10%, 15%, and 20% third harmonic were used. The errors of the estimated values are nearly zero, despite the presence of harmonic components. It can be noted that the proposed algorithm has very small power component estimation errors (30 ppm).

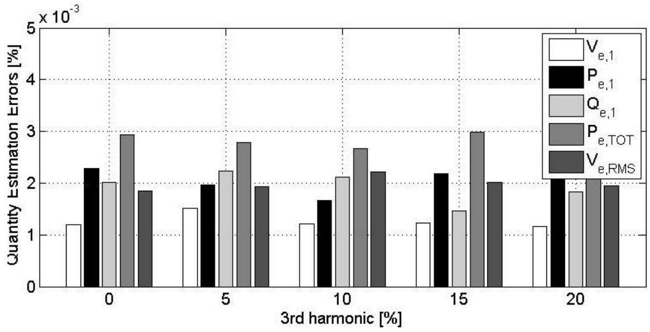

Fig.10. Maximum estimation errors for harmonics presence.

CONCLUSION

The proposed method has been investigated under different conditions and found to be a valuable and efficient tool for the detection of power quantities. In addition, the proposed technique is suitable for measurement for a wide range of frequency changes. The simulation results have shown that the proposed technique provides accurate estimates and offers the possibility to track the quantity changes of nonstationary power signals.

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