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ACTIVE POWER MEASUREMENTS – AN OVERVIEW AND A COMPARISON OF DSP ALGORITHMS BY NONCOHERENT SAMPLING

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Abstract

This paper presents an overview of algorithms for one-phase active power estimation using digital signal processing in the time domain and in the frequency domain, and compares the properties of these algorithms for a sinusoidal test signal. The comparison involves not only algorithms that have already been published, but also a new algorithm. Additional information concerning some known algorithms is also included. We present the results of computer simulations in MATLAB and measurement results gained by means of computer plug-in boards, both multiplexed and using simultaneous signal sampling. The use of new cosine windows with a recently published iterative algorithm is also included, and the influence of additive noise in the test signal is evaluated.

Keywords: power estimation, digital signal processing, non-coherent sampling

1. Introduction

Measurement of power is an important part of electrical measurements and power quality assessment, and many papers have been devoted to various aspects of power measurement (e.g. [1, 2]). This paper presents an overview and a comparison of algorithms for one-phase active power estimation using digital signal processing. The comparison involves not only algorithms that have already been published, but also a new algorithm. Additional information concerning some known algorithms is also included. We present the results of computer simulations and measurement results acquired by means of computer plug-in boards. Both are performed in the MATLAB environment. The paper also includes some examples of the use of cosine windows of new classes designed by a new iterative algorithm recently published by the authors of this paper [3]. The algorithm is implemented in a graphical user interface that is accessible on the web.

Power is estimated by processing two input signals, one corresponding to the voltage on the load, while the other corresponds to the current passing through the load. This paper analyzes the bias caused by the power estimation DSP algorithms that are used, not including the influence of the load voltage and load current converters. The algorithms are analyzed for sinusoidal signals (which are the basic components of any multifrequency signal) and for the general case of non-coherent sampling, i.e. also for a non-integer number of sampled signal periods. The transfer from the time domain to the frequency domain is provided by the Discrete-Time Fourier Transform (DTFT) and the Discrete Fourier Transform (DFT, and its fast version – the Fast Fourier Transform – FFT).

The active power estimation error is influenced by the load phase, the voltage and current DC components, the signal-to-noise ratio (SNR) of the voltage and of the current signals, the number of samples per signal period, the window that is used (window class and window order), and by the non-coherency of the sampling. The influence of the load phase, reduction

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of the influence of multiplexing using linear interpolation, and the influence of non-coherent signal sampling for several classical frequently used windows and processing signals in the time domain are analyzed in [4]. The paper includes additional information, e.g. concerning the efficiency of some recently defined cosine windows [3], the efficiency of better than linear interpolation for error reduction when using a multiplexing DAQ board, and information about a newly proposed method in frequency domain signal processing.

2. Investigated active power estimation methods

The active power is estimated as the average of the instantaneous power in the time domain as

$$P' = \frac{1}{T_M}\int_0^{T_M} v(t)i(t)\mathrm{d}t \cong \frac{1}{N}\sum_{n=0}^{N-1} v(n)i(n), \tag{1}$$

where P' is active power estimate, T_M is time of measurement (in the power definition, one signal period), $v(n)$ and $i(n)$ are sampled versions of continuous-time voltage $v(t)$ and current $i(t)$, and N is the number of acquired samples. The average value (1) can also be found in the frequency domain as the DC component of the Discrete Fourier Transform (DFT) of the instantaneous power. The measurement (sampling) time $T_\mathrm{M} = N\,T_\mathrm{S}$ (T_S being the sampling interval) can also be expressed as $T_\mathrm{M} = (M + \lambda)\,T_\mathrm{SIG}$, where integer M is an integer number of sampled periods (representing the number of signal periods T_SIG sampled completely) and λ is the decimal part of the last sampled signal period ($0 \le \lambda < 1$). There is $P' = P$ for $\lambda = 0$ (coherent sampling), and $P' \neq P$ for $\lambda \neq 0$ [4, 5]. Since exactly coherent sampling is usually not possible (e.g. because of time changes of the power system frequency), the difference between P' and P (i.e. the bias of the measured active power P caused by non-coherent sampling) has to be taken into account. It can be markedly reduced by instantaneous power windowing and additional signal processing. The relative bias of P estimation can be expressed as

$$\delta'_P = \frac{P' - P}{P}. \tag{2}$$

It should be noted that the period of instantaneous power is one half of the voltage and current period.

Additional bias can be generated by non-simultaneous voltage and current input signal sampling when using low-cost multiplexed PC DAQ boards. This bias can be suppressed by various signal interpolations. The effect of linear interpolation has been studied in [4]. A comparison of the efficiency of linear interpolation and spline interpolation (offered e.g. by MATLAB) with processing signals without any interpolation is shown for computer simulation and a physical signal experiment in Fig. 1 and Fig. 2. The Hann window is used in both cases. Fig. 1 shows the results of computer simulation, while Fig. 2 is based on measurements using the low-cost National Instruments PCI 6023E multiplexing DAQ board. The curves for simultaneous sampling and for spline interpolation overlap in the part of Fig. 1 referring to 100 samples per signal period (denoted in the figure as $N_{perT} = 100$). The curves for linear and spline interpolation overlap in the part of Fig. 2 referring to 2000 samples per signal period.

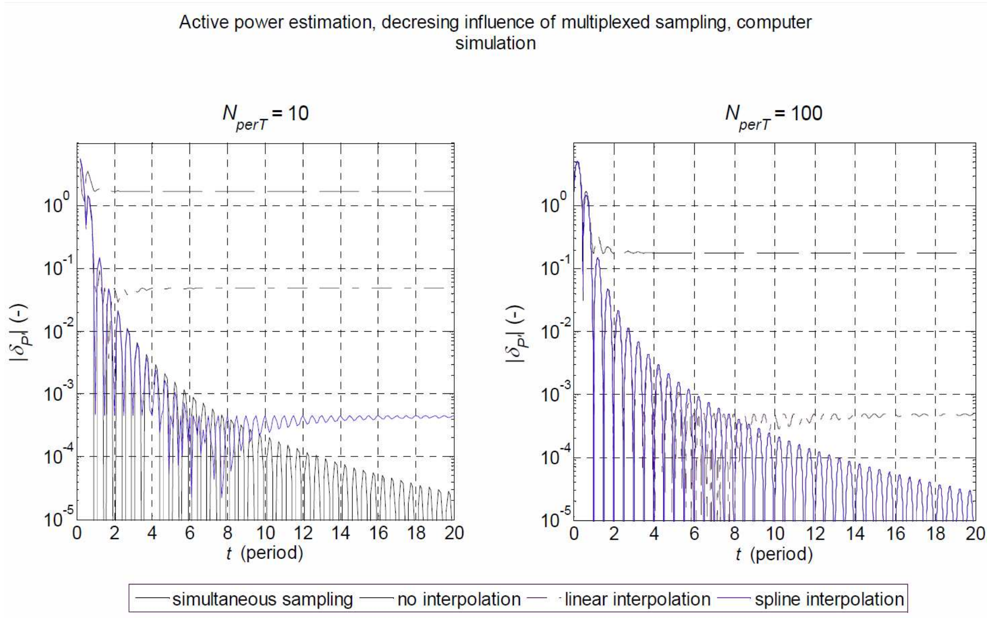

Fig. 1. Relative bias of active power caused by non-simultaneous sampling, Hann window, power estimation in the time domain using windowing (WTD), 10 and 100 samples/period, computer simulation.

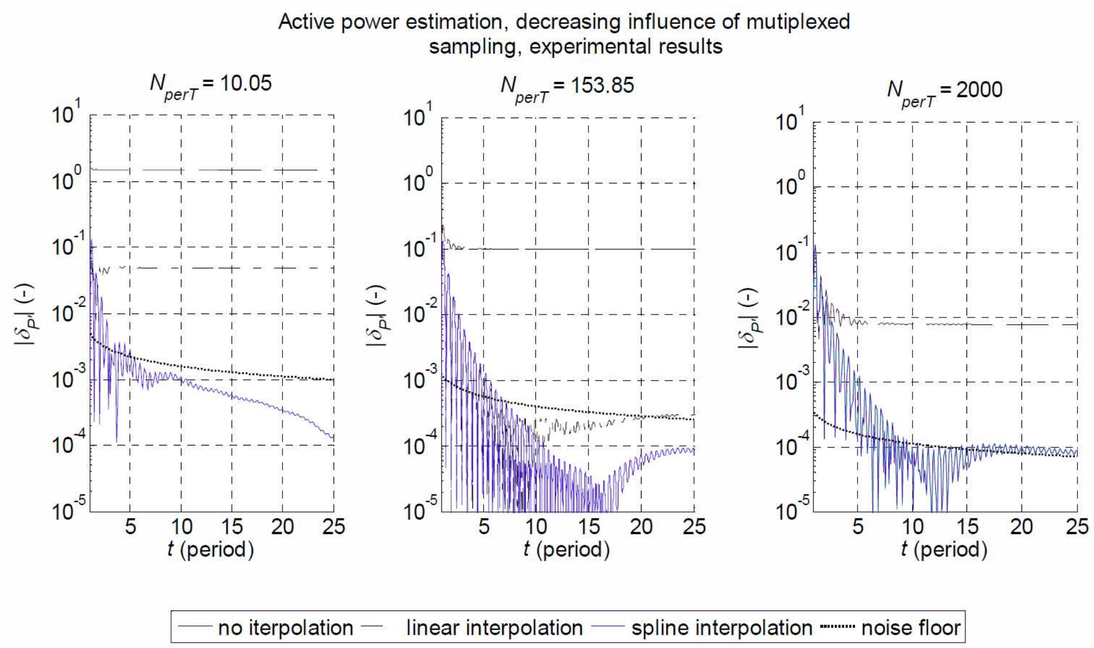

Fig. 2. Relative bias of active power caused by non-simultaneous sampling, Hann window, power estimation method WTD. Experimental results for PCI 6023A DAQ plug-in board.

The effect of non-simultaneous channel sampling can be eliminated by using DAQ plug-in boards with simultaneous channel sampling with a separate ADC (nowadays usually of sigma-delta type) in each channel (the authors of this paper used a PCI 4472 DAQ board), or by using synchronized high-accuracy digital voltmeters for each signal channel sampling, as in [5]. The fundamental cause of the active power uncertainty component corresponding to the noise in the data acquisition channels is the quantization noise in each channel. This can also be expressed by means of the signal-to-noise ratio (SNR) [5].

2.1. Power estimation in the time domain using windowing (WTD)

These algorithms use windowing of the instantaneous power sequence to reduce the power estimation bias. All signal processing takes place in the time domain, i.e. DFT is not used. Detailed information about these algorithms is presented in [4]. Windowing markedly decreases the active power estimation bias with increasing signal sampling time, depending on the window that is used. The choice of a suitable window allows the power to be estimated with low bias in a short time. Windowing is applied here to the instantaneous power. This changes the active power value, so this power estimate has to be corrected by a multiplication factor depending on the window that is used.

The advantages of this method are low processing time and easy applicability for digital signal processors.

2.2. Power estimation using interpolation and re-sampling in the time domain (I/RTD)

This method is not mentioned either in [4] or in [5]. The effect of non-coherent signal sampling can also be reduced by interpolating the sequence of signal samples (i.e. using interpolation in the time domain), truncating an integer number of signal periods and re-sampling the interpolated signal so that an integer number of signal samples are obtained. Any interpolation method can be used; the most widely used are staircase, linear and spline interpolation.

Instead of interpolated signal re-sampling, it is also possible just to truncate the original sample sequence to a length corresponding as closely as possible to an integer number of the original analog signal periods (in this case however FFT algorithms not requiring signal length 2^N samples have to be used). This shortens the required processing time, but residual non-coherency of the sampling leads to a non-zero power estimation bias. A hardware realization of this method is used in some present-day energy analyzers. The application of this method for signal amplitude and phase estimation is described in [6].

A basic condition for successful application of this method is that the signal period time must be known with high accuracy. Frequency estimation has been studied by many authors (recently e.g. in [7, 8, 9]). An error in signal period time (or frequency) estimation is reflected in residual non-coherency of signal sampling and a corresponding power estimation bias. This bias can be reduced by signal windowing and by increasing the signal sampling time.

The I/RTD method provides high-accuracy active power estimation, but it is very sensitive to the presence of sub-harmonic or inter-harmonic components in a multi-frequency signal.

2.3. Power estimation in the frequency domain using windowing (WFD)

Processing a signal in the frequency domain, i.e. after transforming the signal to the frequency spectrum by a suitable type of Fourier transform, offers an illustrative reflection of time-domain operations into the frequency domain. Since the active power is the DC component of the instantaneous power, it can be found either as a mean value of the instantaneous power by integration in the time domain (1), or as the DC component of the normalized DFT spectrum of the instantaneous power. DFT spectrum normalization involves dividing the spectrum by N (using a two-sided spectrum), where N is the DFT length. If a non-integer number of signal periods are sampled, leakage of energy into the surrounding frequency bins causes power estimation bias. The effect of sampling non-coherency can be substantially reduced by instantaneous power windowing, as in the WTD method. The normalized DC DFT component of the windowed instantaneous power has to be corrected by the same multiplication factor as the instantaneous power DC component in the WTD method.

Since the WTD method and the WFD method differ only in the way of finding the instantaneous power DC component, active power estimations by these two methods are practically identical.

2.4. Power estimation in the frequency domain using windowing and interpolation in the frequency domain (WIFD)

In this method, information about all components of windowed instantaneous power spectrum is available to the user for increasing the accuracy of active power estimation. In practice, more than one component of the instantaneous power DFT spectrum in the neighborhood of the DC component is used for active power estimation. If we know the window spectrum and the DTFT and DFT spectrum of the windowed signal spectrum, we can find the individual signal spectrum components in the surroundings of the local DFT spectrum maxima by various interpolated DFT methods (e.g. [10, 11, 12]). In active power estimation, the DC component of the instantaneous power spectrum has to be found. Its fractional frequency bin is zero, and simplified interpolation expressions can therefore be used [5]. If a cosine window w(n) of order L and of length N defined by (3) is used,

$$w(n) = \sum_{r=0}^{L} D_r \cos\left(2\pi r \frac{n}{N}\right), \quad n = 0, 1, \cdots, N-1 \tag{3}$$

the general interpolation formula for an active power estimate can be written as

$$P' = \frac{|X(0)| \cdot |D_0| + |X(1)| \cdot |D_1| + |X(2)| \cdot |D_2| + \cdots}{D_0^2 + D_1^2/2 + D_2^2/2 + \cdots}, \tag{4}$$

where $|X|$ is a row vector composed of the values of the right side of the two-sided DFT magnitude spectrum of the windowed instantaneous power, starting with the DC component $X(0)$ and D_r are window coefficients from (3). Relation (4) can be rewritten in a more suitable form (not presented in [5])

$$P' = |\mathbf{X}| \cdot \mathbf{K}^T. \tag{5}$$

Here **K** is the row vector of the interpolation coefficients. The length of the two vectors $|\boldsymbol{X}|$ and **K** is usually lower than or equal to the number of instantaneous power spectrum lines for nonnegative frequencies within the used window spectrum main-lobe centered on zero frequency. Vector **K** can be expressed as

$$\mathbf{K} = \frac{\left[|D_0||D_1||D_2|...|D_{m-1}|\right]}{D_0{}^2 + \sum_{r=1}^{m-1} \frac{D_r{}^2}{2}}, \tag{6}$$

The interpolation coefficient values found from (6) correspond to the values presented in [5] for window order up to 3. The interpolation coefficients for class 1 Rife-Vincent windows ([13, 14]) and for window orders 0 to 4 are shown in Table 1.

The WFD method corresponds to the 1-point special case of the WIFD method. The WIFD method increases the accuracy of power estimation by windowing and interpolation in the frequency domain. Its processing complexity is higher than the complexities of time-domain methods, but thanks to the fast FFT algorithms and available hardware, the computation time demands may not be markedly increased.

Table 1. Interpolation coefficients for active power estimation using the WIFD method and Rife-Vincent windows of the first class (RV1 windows) [13, 14].

window order L	Interpolation coefficients $\boldsymbol{K}$
0 (rectangle)	$[1]$
1 (Hann)	$[4/3 \quad 4/3]$
2	$[48/35 \quad 64/35 \quad 16/35]$
3	$[320/231 \quad 480/231 \quad 192/231 \quad 32/231]$
4	$[841/604 \quad 929/417 \quad 929/834 \quad 183/575 \quad 22/553]$

2.5. Power estimation in the frequency domain using an estimation of the individual power components (WCFD)

Active power can alternatively be found in the frequency domain by summing the active powers of the individual power spectrum components. These active power components represent the powers of the voltage and the current harmonic components of the same order, including the DC components

$$P^{'} = I_0 V_0 + \sum_{k=1}^{M} \frac{V_k^{'} I_k^{'}}{2} \cos\left(\varphi_{u,k}^{'} - \varphi_{i,k}^{'}\right) . \tag{7}$$

Here M is the number of higher-order signal harmonic components and V', I' are estimates of the voltage and current component *RMS* values, and φ' are estimates of the voltage and the current k-th order harmonic component phase shifts related to the beginning of sampling. Windowing is applied here to the voltage and current signals, and interpolated DFT is used for estimating the frequency and, consequently, the *RMS* values and phases of the individual harmonic components of the voltage and current signals.

It is rather complicated and time-consuming to find the active power by this procedure. Power estimation uncertainty is influenced here basically by the uncertainty of the largest active power value of all processed harmonic components. However, this active power estimation method provides information about the frequency structure of the estimated power, and the possibility to select the frequency band in which the active power should be estimated.

2.6. Power estimation in the frequency domain, using windowing and repeated instantaneous power spectrum component estimation (WSCFD)

This is a newly proposed method that can in some cases increase the accuracy of active power estimation by processing signals in the frequency domain. The principle is as follows. The normalized discrete-time Fourier Transform (DTFT) complex spectrum of a cosine window (3) is

$$W(\theta,\varphi) = \frac{e^{-j\left(\pi\theta\frac{N-1}{N}\right)}}{N} \sum_{r=0}^{L} \frac{D_r}{2} \left(e^{j\pi r\frac{N-1}{N}} \frac{\sin(\pi(\theta-r))}{\sin\left(\frac{\pi(\theta-r)}{N}\right)} + e^{-j\pi r\frac{N-1}{N}} \frac{\sin(\pi(\theta+r))}{\sin\left(\frac{\pi(\theta+r)}{N}\right)} \right) . \tag{8}$$

Here θ and r are the frequencies in the frequency bin (θ is a non-integer real number), and N is the window length in the samples.

After applying a correction multiplicative factor, the estimation bias of the DC component and the remaining components of the instantaneous power spectrum are influenced only by long-range leakage. This leakage can be reduced numerically by repeated spectrum component estimation, in each step after subtracting the largest spectrum component from the instantaneous power spectrum. The largest component frequency, amplitude and phase can be found using the WIFD method, and it can be subtracted from the instantaneous power spectrum (the two mirrored components for positive and negative frequency with all their contributions to the neighboring spectral components are subtracted). The accuracy of the estimation of the remaining components is in some cases increased by this operation.

Starting from the highest component and again subtracting the currently largest component (if this happens to be the DC component, it will be stored and its possible multiple values are accumulated and the remaining number of estimates are realized) may lead step-by-step to increased accuracy of the estimation of the remaining instantaneous power spectrum components. The last estimate should always be the estimate of the DC component.

Three such estimates are sufficient for a one-tone signal with a non-zero DC component. For signals with an unknown frequency structure, a higher number of repeated estimations should be selected and these repetitions can be stopped if the dispersion of the level of the spectral lines is not decreased by more than the selected value after the last estimate.

Even if the principle here described may appear complicated, after a single windowed instantaneous power spectrum computation all subsequent operations are performed on this spectrum. The most time-consuming part of the remaining signal processing is repeated window spectrum computation, but if a low-order window (e.g. a Hann window) is used, this computation is simple.

This method enables us to achieve low active power estimation bias even in the presence of sub-harmonic and inter-harmonic components. However, our experiments showed that this principle is not efficient if there are strong additive noise and strong harmonic components in the instantaneous power spectrum. The advantage of this method in the case of a sub-harmonic component in the signal is shown in Fig. 4.

3. Examples of active power estimation of simulated and measured signals, and a comparison of the investigated active power estimation methods

Fig. 3 and Fig. 4 compare the relative power estimation bias for different input signals for the investigated power estimation methods and for various sampling times.

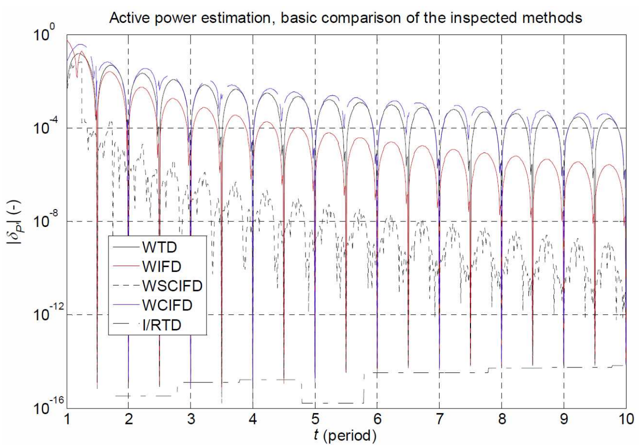

Fig. 3. Dependence of the relative bias of active power estimation for the investigated methods on sampling time. The signal is the fundamental harmonic component only. The sampling time is expressed in input signal periods.

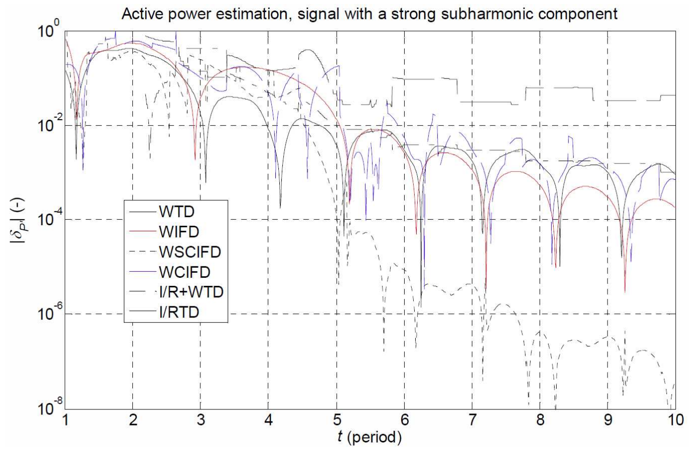

Fig. 4. Dependence of the relative bias of active power estimation for the investigated methods on sampling time. The signal is composed of a fundamental harmonic component and a sub-harmonic component.

Fig. 3 compares the investigated active power estimation methods, using computer simulations for the simplest input signal. Sinusoidal voltage and current waveforms, without DC components and without noise, were used for the simulation. The load impedance phase is 80°. A Hann window and 3-point interpolation is used. As shown in Fig. 3, under these conditions the I/RTD method is the most accurate (using a 3-point Hann window interpolation in signal period estimation and re-sampling for leakage reduction). The second best is the WSCIFD method, using repeated instantaneous power spectrum correction. WIFD is the third best method, again using a 3-point Hann window interpolation in the frequency domain.

Fig. 4 depicts a signal containing a fundamental harmonic component and a sub-harmonic component (with frequency ratio f_{sub}/f_1 = 0.63 and magnitude ratio A_{sub}/A_1=0.5), a zero DC component and a phase of load of 80°. The 3-point Hann window interpolation is used (with the exception of the I/RTD method), and the I/RTD method with windowing (I/R+WTD) is also included.

Fig. 3 and Fig. 4 show that a strong sub-harmonic component influences the bias for all power estimation methods, but the I/RTD method is influenced most, moving from the lowest bias for the conditions of Fig. 3 to the highest bias in the presence of a strong sub-harmonic component.

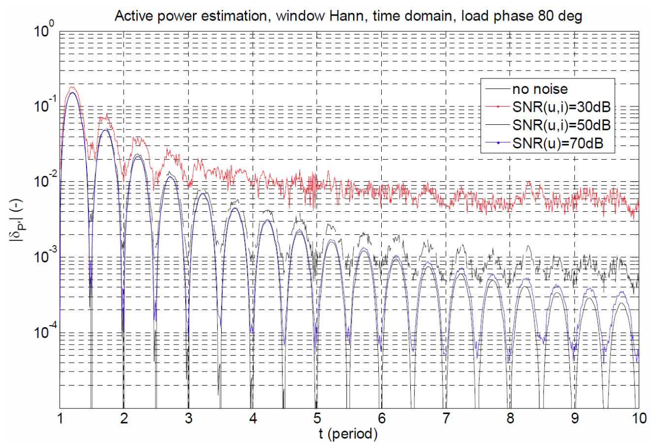

Fig. 5. Example of the influence of the additive noise of the input signals on the active power relative bias for a signal without noise and for SNR 70 dB, 50 dB and 30 dB, computer simulations.

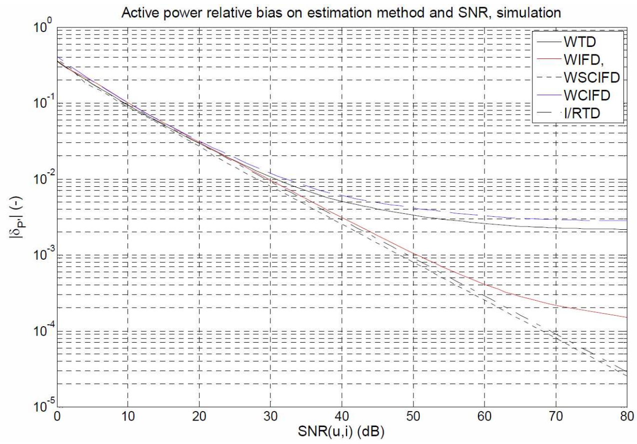

Fig. 6. Comparison of the sensitivity of active power estimation methods to additive noise, expressed as an SNR value.

The influence of additive noise on the active power estimation is shown in Fig. 5 and Fig. 6. The influence of the noise floors corresponding to SNR values is shown for longer times in Fig. 6, where the WCIFD method and the WTD method have the highest power estimation bias for SNR above 30 dB. Fig. 6 compares the sensitivity of the investigated methods with the additive noise expressed in relation to the signal (as SNR). These two figures show the results of a computer simulation for the 80° load phase. Fig.5 is plotted for the WTD method and a Hann window. Fig. 6 is plotted for 4.75 signal periods and for a Hann window (with the exception of the I/RTD method). The SNR values for the voltage and current signals are the same, normally distributed noise is used, and the figure is plotted for 100 simulation repetitions.

Fig. 7 and Fig. 8 compare the results of processing simulated signals and signals generated by an HP3245A two-channel signal generator and measured by two different National Instruments DAQ plug-in boards. A low-cost PCI 6023A multiplexed board (see also Fig. 2) and a PCI 4472 simultaneously sampling board were used. The input ranges of the two boards were 10 V, the sampling frequency was 100 kSa/s for each channel (the 6023A multiplexing DAQ plug-in board was set to 200 kSa/s, and the 4472 board was set to 100 kSa/s). The input signals were sinusoidal 50 Hz with amplitude 5 V, load phase 80°. A Rife-Vincent window of the first class (RV1) [13, 14] and of the second order is used in Fig. 7, and a RV1 window of the third order is used in Fig. 8. The figures are plotted for the WTD power estimation method.

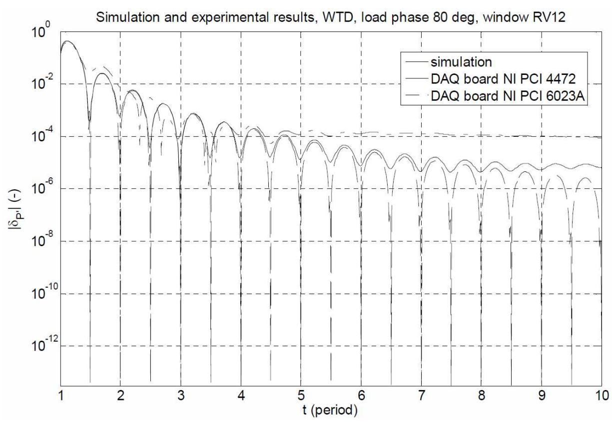

Fig. 7. A comparison of active power estimation by simulation and by measurement, RV1 window [13, 14] of the second order.

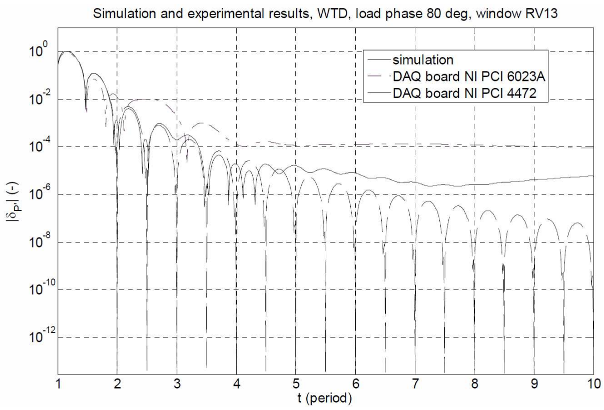

Fig.8. A comparison of active power estimation by simulation and by measurement, RV1window [13, 14] of the third order.

The remaining two figures offer a comparison of various windows for active power estimation. Fig. 9 shows the dependence of the bias on signal length using four 3rd order windows (two of them are RV1 windows [13, 14] and two are from the newly defined cosine window classes C7 and C8 [3]). The results in Fig. 9 were obtained by simulation for a load phase of 60° and by signal processing in the frequency domain (WFD method).

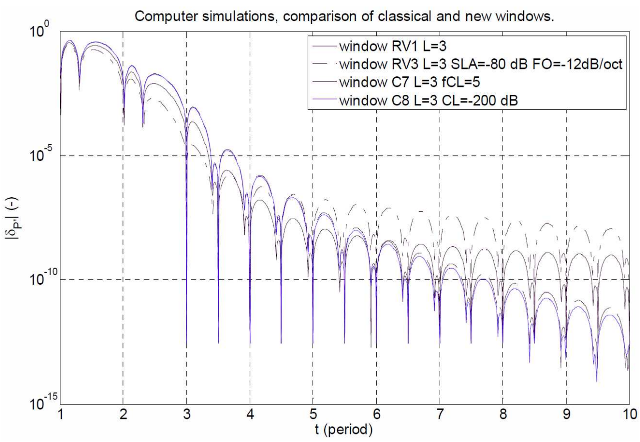

Fig. 9. Influence of the window on the power estimation bias for various windows.

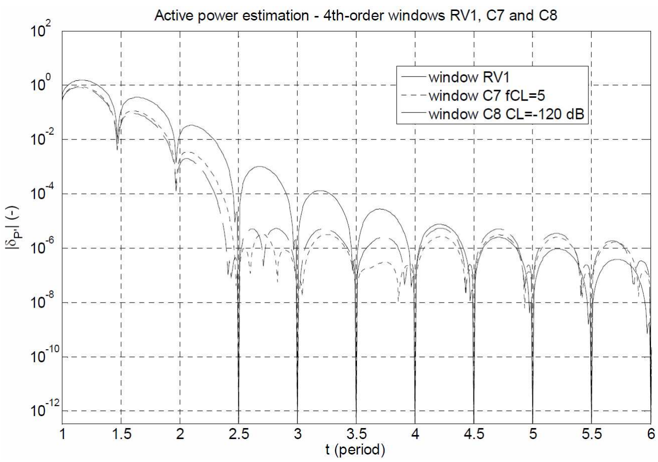

Fig. 10. Showing the advantage of some new windows (C7 and C8 [3]) for active power estimation within a short time.

Fig. 10 shows another comparison of two cosine windows of the new classes C7 and C8 [3] with a classical Rife-Vincent window [13, 14]. All the compared windows are of order 4, and the advantage of windows C7 and C8 for power estimation within a short time (up to four signal periods) can be seen.

Table 2. A comparison of the accuracy and computing time of the investigated active power estimation methods.

Method	**Basic properties, advantages and disadvantages of each method**	**Relative computation time**
WTD	The fastest method, simple power estimation. Mostly a rather low-accuracy method.	1
I/RTD	The most accurate method for signals without a DC component and without inter- or sub-harmonic components (when they are present, the method fails). Its accuracy depends on the accuracy of the signal frequency estimation. This method is used in some energy analyzers.	3,7 – lin. int. (8,7 for spline interpolation)
I/R+WTD	Supplements the I/RTD method with windowing. Depending on the selected window, it suppresses inter- and sub-harmonic components and inaccuracy of signal frequency estimation.	4,3 – lin. int. (9,7 for spline interpolation)
WFD	A fast method that produces (within the numerical accuracy framework of FFT computation) exactly the same estimate as the WTD method.	1,3
WIFD	Uses DFT interpolation based on a selectable number of DFT spectrum lines. A fast method. More accurate than the WTD method. Window order and number of interpolation points selectable according to the additive noise level.	1,3
WCIFD	The slowest and lowest accuracy method, but the only method that allows only the selected power spectrum frequency band to be processed.	47 (for 50 harmonic components)
WSCIFD	A rather slow but high-accuracy method (accuracy comparable to the I/RTD method). Not sensitive even to strong sub-harmonic components, but fails for strong additive noise (i.e. at low SNR. values).	8,7

4. Conclusions

Six active power estimation DSP algorithms (described in section 2) have been compared (see section 3, Fig. 3, Fig. 4, Fig. 6 and Table 2). A sinusoidal test signal (a basic signal of non-harmonic periodic signals) was used. The results of a computer simulation and measurements using two different-type DAQ plug-in boards have been compared (Fig. 7 and Fig. 8). The sensitivity of the methods to a sub-harmonic component (Fig. 4), to the type of window and window order (Fig.9 and Fig.10), and to additive noise (Fig. 5 and Fig. 6) has also been investigated.

Windowing of instantaneous power substantially decreases the active power estimation bias in the case of non-coherent sampling, and substantially reduces the measurement time for the allowed estimation bias. The efficiency of increasing the window order, increasing the number of interpolation points and using more complicated algorithms depends on the noise level in the voltage and current channels. Increasing the window order leads not only to decreasing estimation bias but also (due to the simultaneous increase in the window equivalent noise bandwidth) to a small increase in estimation uncertainty. Computer simulation can be used for finding, for a given DAQ plug-in board and a selected window, the minimum number of signal periods that have to be sampled in order to achieve a power estimation bias level below the noise floor. Classical class 1 Rife-Vincent windows (maximum side-lobe fall-off windows) of a low order (e.g. Hann window) are the best windows for power estimation for sufficiently long signals. The use of a multiplexed DAQ plug-in board may lead to rather high bias for the less than 50 samples per signal period for linear interpolation of signals (spline interpolation is efficient in this case).

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