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High-precision time domain reactive power measurement in the presence of interharmonics

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Abstract: When interharmonics exist in power system signals, large errors emerge in traditional time domain reactive power measurement. In this paper, we present a novel time domain integral method with good effect of restraining interharmonics, synchronization error, and white noise, as well as the principle of the selection of the sampling periods when employing this approach. The current signal and phase-shifted voltage signal are reconstructed after the harmonic components of signals are extracted, so that the interharmonics are filtered. The influence of the synchronization error on the measurement is reduced through removing the weight coefficients of the reactive components. In the simulation, we apply several cosine windows to the proposed method and analyze signals containing both harmonics and interharmonics. The results show that, in the presence of interharmonics, synchronization error, and white noise (with a fundamental signal-to-noise ratio of 40 dB) all together, the relative errors are within the magnitude of 10^{-4}, which perfectly satisfies the practical requirement.

Key words: Cosine window, Interharmonics, Reactive power, Synchronization error, Windowed discrete Hilbert transform
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1 Introduction

The measurement of reactive power is of great significance in power systems. Based on the definition given by Budeanu, current reactive power measurement approaches include mainly methods based on fast Fourier transform (FFT), Walsh functions (Abiyev *et al*., 2007; Abiyev and Dimililer, 2008), or wavelet transform (Yoon and Devaney, 2000; Driesen and Belmans, 2003), and time domain integral methods (Srinivasan, 1987; Makram and Haines, 1991; Saranovac, 2000; Pang *et al*., 2007; Wei *et al*., 2010). Amongst them, the time domain integral method is the most widely used, and it is usually realized in two steps: (1) perform the 90° phase shift on the voltage signal, always by using windowed discrete Hilbert transform (WDHT) (Wei *et al*., 2010; 2011); (2) filter the instantaneous reactive power sequence to obtain the reactive power value.

Within current time domain integral methods, some consider the influence of harmonic distortion on reactive power measurement (Makram and Haines, 1991; Pang *et al*., 2007; Wei *et al*., 2010). The others take into account the influence of the sampling synchronization error, caused by the drift of the fundamental frequency (Srinivasan, 1987; Saranovac, 2000; Wei *et al*., 2010). How to ensure the reactive power measurement accuracy in the presence of interharmonics, however, is still an unknown area. Also, with the increasing use of nonlinear components and periodical time-varying loads, the pollution of interharmonics is becoming more and more severe in power systems (Qi and Wang, 2003; Qian *et al*., 2007). When interharmonics exist in voltage and current signals, they will also appear in the obtained instantaneous reactive power sequence, which severely affects the effect of filtering the sequence, and thus largely decreases the measurement accuracy.

In this paper, we propose a novel time domain integral method with a good effect of restraining interharmonics, synchronization error, and white noise. The main ideas are: (1) truncate the sampled signals with long-duration cosine windows to restrain the interference among interharmonics and harmonics; (2) after extracting harmonic components, reconstruct the current signal and phase-shifted voltage signal to filter the interharmonics; (3) further correct the instantaneous reactive power sequence through calculating the frequency correction value, so that the influence of asynchronous sampling on the measurement is reduced.

2 Discussion of reactive power measurement in the presence of interharmonics

Suppose the period of power system signals $u(t)$ and $i(t)$ is T_1, and their fundamental simulative angular frequency is Ω_1. Sampling them with an interval of T_s, M points per period for m periods, we obtain the N-point discrete signals $u(n)$ and $i(n)$:

$$u(n)=\sum_{k=1}^{H}\sqrt{2}U_k\cos(k\omega_1 n+\alpha_k), \tag{1}$$

$$i(n)=\sum_{k=1}^{H}\sqrt{2}I_k\cos(k\omega_1 n+\beta_k), \tag{2}$$

where ω_1 is the fundamental digital frequency, $\omega_1=\Omega_1 T_s$, U_k, I_k and α_k, β_k are the root mean square (RMS) and initial phases of harmonic k respectively, and H is the maximally analyzed harmonic order of $u(n)$ and $i(n)$.

Budeanu's definition of reactive power is given by (Budeanu, 1927)

$$Q=\sum_{k=1}^{H}Q_k=\sum_{k=1}^{H}U_k I_k\sin(\alpha_k-\beta_k), \tag{3}$$

where Q_k represents the reactive component of harmonic k.

Performing the 90° phase shift on $u(n)$, we obtain $\hat{u}(n)=\sum_{k=1}^{H}\sqrt{2}U_k\sin(k\omega_1 n+\alpha_k)$. Then the instantaneous reactive power sequence is $q(n)=\hat{u}(n)i(n)$, which can eventually be deduced as

$$\begin{aligned}q(n)&=\sum_{k=1}^{H}U_k I_k\sin(\alpha_k-\beta_k)+\sum_{k=1}^{2H}A_k\sin(k\omega_1 n+\psi_k)\\&=Q+\sum_{k=1}^{2H}A_k\sin(k\omega_1 n+\psi_k),\end{aligned} \tag{4}$$

where A_k and ψ_k represent the amplitude and phase of harmonic k, respectively.

As can be seen, $q(n)$ is a periodical signal with the same period as $u(n)$ and $i(n)$. It contains DC and harmonic components. The DC component can be seen as the result of the product-to-sum calculation between the same-order harmonics of $u(n)$ and $i(n)$. Its value is equal to Q, which can be achieved by filtering the harmonic components. This is the theory of the time domain integral method.

Note that we have considered only the harmonics whose frequencies are an integer of the fundamental frequency in the above discussion. In power systems, there also exist interharmonics. IEC Standard 61000-2-2:2002 defines the interharmonic component as "a component having an interharmonic frequency". The 'interharmonic frequency' is defined as "any frequency which is not an integral multiple of the fundamental frequency". When interharmonics exist in the voltage and current signals, they will also appear in the instantaneous reactive power, which affects the filtering effect. In particular, when the frequencies of interharmonics are close to zero, the measurement errors can be larger. By analyzing only the harmonics in the voltage and current signals, the influence that interharmonics have on the measurement can be decreased, which is the essential concern and purpose of this study.

3 Background introduction

3.1 Brief introduction of cosine windows

The general expression of cosine windows is

$$w(n)=\sum_{h=0}^{J-1}(-1)^h a_h\cos\left(\frac{2\pi n}{N}h\right),\ n=0,1,\cdots,N-1, \tag{5}$$

where a_h is the coefficient and J is the number of items. Their values determine the various window functions (Harris, 1978; Nuttall, 1981; Andria *et al.*, 1989).

With the linear-phase characteristic, a_h should meet $\sum_{h=0}^{J-1}(-1)^h a_h = 0$. Then, the discrete time Fourier transform (DTFT) of $w(n)$ can be expressed as

$$\begin{aligned} W(\mathrm{e}^{\mathrm{j}\omega}) &= W_0(\omega)\cdot \mathrm{e}^{-\mathrm{j}N\omega/2} \\ &= \mathrm{e}^{-\mathrm{j}N\omega/2}\sum_{h=0}^{J-1}\frac{1}{2}\frac{a_h \sin\left(\frac{N}{2}\omega\right)\sin\omega}{\sin\left(\frac{\omega}{2}-\frac{\pi h}{N}\right)\sin\left(\frac{\omega}{2}+\frac{\pi h}{N}\right)}, \end{aligned} \quad (6)$$

where $W_0(\omega)$ is a real even function, and the phase constant of cosine windows is $N/2$.

The mainlobe bandwidth of cosine windows is related with J. To be specific, it satisfies

$$\Lambda = 2J\Delta\omega, \quad (7)$$

where $\Delta\omega=2\pi/N$ represents the frequency resolution of discrete Fourier transform (DFT).

3.2 Windowed interpolation algorithm

Take harmonic k of $u(n)$ for analysis:

$$u_k(n) = \sqrt{2}U_k\cos(k\omega_1 n + \alpha_k). \quad (8)$$

Truncating $u_k(n)$ with the cosine window, we obtain $u_{kw}(n)=u_k(n)w(n)$. Based on the convolution theorem in the frequency domain, the DTFT of $u_{kw}(n)$ can be expressed as

$$\begin{aligned} U_{kw}(\mathrm{e}^{\mathrm{j}\omega}) &= \frac{1}{2\pi}U_k(\mathrm{e}^{\mathrm{j}\omega}) * W(\mathrm{e}^{\mathrm{j}\omega}) \\ &= \frac{\sqrt{2}U_k}{2}\Big[W_0(\omega-k\omega_1)\mathrm{e}^{\mathrm{j}(\alpha_k-N(\omega-k\omega_1)/2)} \\ &\quad + W_0(\omega+k\omega_1)\mathrm{e}^{-\mathrm{j}(\alpha_k+N(\omega+k\omega_1)/2)}\Big]. \end{aligned} \quad (9)$$

Due to the duration of $u_k(n)$, $T_w=NT_s$ cannot be an integral multiple of T_1. Then,

$$\frac{NT_s}{T_1} = \frac{2\pi f_1}{2\pi f_s/N} = \frac{\omega_1}{\Delta\omega} = m+\lambda, \quad (10)$$

where λ is the remainder of NT_s/T_1, $\lambda\in[-0.5, 0.5]$.

From Eq. (10), it can be further concluded that the frequency of harmonic k satisfies

$$\omega_k = k\omega_1 = (km+k\lambda)\Delta\omega = (m_k+\lambda_k)\Delta\omega, \quad (11)$$

where $m_k=km$, $\lambda_k=k\lambda$. λ_k is called the frequency correction value, reflecting the synchronization error of the sampling. When the synchronization error is not that large, there is always $|\lambda_k|<1$. Then we can consider that the harmonic k ($\omega=\omega_k$) corresponds to the m_k spectrum line ($\omega=m_k\Delta\omega$) (Huang and Jiang, 2005).

The FFT of $u_{kw}(n)$ is obtained by sampling $U_{kw}(\mathrm{e}^{\mathrm{j}\omega})$ equally in the range of [0, 2π] with a sampling interval of $2\pi/N$ ($\Delta\omega$); thus, we can write the spectrum line of harmonic k as

$$\begin{aligned} U_{kw}(m_k) &= U_{kw}(\mathrm{e}^{\mathrm{j}\omega})\big|_{\omega=m_k\Delta\omega} \\ &= \frac{\sqrt{2}U_k}{2}W_0(-\lambda_k\Delta\omega)\mathrm{e}^{\mathrm{j}(\alpha_k+\pi\lambda_k)} \\ &\quad + W_0\big((2m_k+\lambda_k)\Delta\omega\big)\mathrm{e}^{-\mathrm{j}(\alpha_k+\pi(2m_k+\lambda_k))}. \end{aligned} \quad (12)$$

The second part in Eq. (12) corresponds to the long-range spectrum leakage at the frequency of $m_k\Delta\omega$. Considering the fast sidelobe attenuation of cosine windows, its value is very small and can thus be neglected:

$$U_{kw}(m_k) = \frac{\sqrt{2}U_k}{2}W_0(-\lambda_k\Delta\omega)\mathrm{e}^{\mathrm{j}(\alpha_k+\pi\lambda_k)}. \quad (13)$$

Based on Eq. (13), each harmonic amplitude and phase can thus be calculated after obtaining λ.

4 Reactive power measurement in the presence of interharmonics

4.1 Interharmonic filtering and phase shift of the voltage signal

From this section on, interharmonics are taken into account in all discrete signals $u(n)$ and $i(n)$.

By employing the time domain integral method, we need to perform the Hilbert transform on $u(n)$. A fast WDHT method based on FFT and IFFT (inverse fast Fourier transform) has been proposed, and the realization process proceeds as follows (Wei *et al.*, 2011).

1. Truncate $u(n)$ with cosine window $w(n)$, and accumulate the windowed signal $u_w(n)$ into one period to obtain the M-point signal $u_s(n)$:

$$u_s(n) = u_w(n) + u_w(n+M) + \ldots + u_w(n+(m-1)M),\\ n=0,1,\cdots,M-1.$$

2. Taking the M-point FFT of $u_s(n)$, we obtain the calculated DC and harmonic components of $u(n)$, i.e., $U_w(0)$, $U_w(m_1)$, …, $U_w(m_H)$.

3. Calculate $Z(k)=\begin{cases}U_w(m_k), & k=0,\\ 2U_w(m_k), & k=1,2,\cdots,H,\end{cases}$ take the M-point IFFT of [$Z(0)$, $Z(1)$, …, $Z(H)$, 0, …, 0] (M elements), and divide the derived time domain sequence by m to obtain $u''(n)$.

4. Extending $u''(n)$ periodically for m periods, we obtain the analytic signal of $u(n)$, the imaginary part of which is the phase-shifted signal $\hat{u}(n)$.

As we know, Hilbert transform is a linear transformation which imparts a 90° phase shift of the input signal. It has the transfer function

$$H(\omega)=\begin{cases}-\mathrm{j}, & \omega>0,\\ \mathrm{j}, & \omega<0.\end{cases} \tag{14}$$

Based on Eq. (14), harmonics of $\hat{u}(n)$ can be written as

$$\hat{U}(m_k)=\begin{cases}-\mathrm{j}U_w(m_k), & k=1,2,\ldots,H,\\ \mathrm{j}U_w(m_k), & k=-H,\ldots,-2,-1.\end{cases} \tag{15}$$

Referring to the principle of the windowed interpolation algorithm, Eq. (15) can be expressed as

$$\hat{U}(m_k)=\begin{cases}-\mathrm{j}\dfrac{\sqrt{2}U_k}{2}W_0(-\lambda_k\Delta\omega)\mathrm{e}^{\mathrm{j}(\alpha_k+\pi\lambda_k)}, & \\ \qquad k=1,2,\ldots,H, & \\ \mathrm{j}\dfrac{\sqrt{2}U_{-k}}{2}W_0(\lambda_{-k}\Delta\omega)\mathrm{e}^{-\mathrm{j}(\alpha_{-k}+\pi\lambda_{-k})}, & \\ \qquad k=-H,\ldots,-2,-1. & \end{cases} \tag{16}$$

It has been proved that the N-point signal $\hat{u}(n)$ has the expression of (Wei *et al.*, 2010)

$$\hat{u}(n)=\sum_{k=1}^{H}\sqrt{2}U_kW_0(-\lambda_k\Delta\omega)\sin(k\omega_1'n+\pi\lambda_k+\alpha_k), \tag{17}$$

where $\omega_1'=m\Delta\omega$ is the fundamental frequency.

Compared with the interharmonics-contained and asynchronously sampled $u(n)$, two differences can be found in $\hat{u}(n)$. First, $\hat{u}(n)$ contains only harmonics. Second, $u(n)$ changes to synchronously sampled signal $\hat{u}(n)$ after WDHT. The essential reason for these two changes is that, when taking the FFT of $u_w(n)$ and IFFT of $U_w(m_k)$ in the process of WDHT, only the harmonic components ($\omega=m_k\Delta\omega$) are included in the weighted processing whereas the non-integral harmonic components ($\omega\neq m_k\Delta\omega$) are neglected (supposed to be zero) (Wei *et al.*, 2010).

In WDHT, the spectrum leakage among interharmonics and harmonics may cause large errors in the estimation of $U_w(0)$, $U_w(m_1)$, … , $U_w(m_H)$. We can extend the sampling periods of $u(n)$ and accordingly enlarge the duration of the cosine window to reduce the interference. As long as the duration of $w(n)$ is long enough, we can neglect the spectrum leakage and precisely analyze the single-frequency harmonic component. Detailed analysis of this point is presented in Section 5.

4.2 Interharmonic filtering of the current signal

Referring to the deduction in Section 4.1, the process of interharmonics filtering of $i(n)$ can be summarized as follows:

1. Truncate $i(n)$ with $w(n)$, and accumulate the windowed signal into one period to obtain $i_s(n)$.

2. Taking the M-point FFT of $i_s(n)$, we obtain the M-point frequency-domain components $I_w(0)$, $I_w(m_1)$, …, $I_w(m_{M-1})$.

3. Take the M-point IFFT of [$I_w(0)$, $I_w(m_1)$, …, $I_w(m_{M-1})$] and divide the derived time domain sequence by m to obtain $i''(n)$.

4. Extending $i''(n)$ periodically for m periods, we obtain the N-point interharmonic-filtered signal:

$$i'(n)=\sum_{k=1}^{H}\sqrt{2}I_kW_0(-\lambda_k\Delta\omega)\cos(k\omega_1'n+\pi\lambda_k+\beta_k). \tag{18}$$

4.3 Measuring the reactive power

From Eqs. (17) and (18), the instantaneous reactive power sequence is $q'(n)=\hat{u}(n)i'(n)$ and can be deduced to be

$$q'(n)=\sum_{k=1}^{H}Q_kW_0^2(-\lambda_k\Delta\omega)+\sum_{k=1}^{2H}B_k\sin(k\omega_1'n+\phi_k), \tag{19}$$

where B_k and ϕ_k are the amplitude and phase of harmonic k, respectively.

$q'(n)$ is composed of the DC component and the algebraic sum of harmonic components. It is a synchronously sampled periodic signal of period M. The period-mean calculation can thus be used to obtain the DC component:

$$\frac{1}{M}\sum_{n=0}^{M-1} q'(n) = \sum_{k=1}^{H} Q_k W_0^2(-\lambda_k \Delta\omega). \tag{20}$$

As can be seen from Eq. (20), we need only to use the first M points of $\hat{u}(n)$ and $i'(n)$ to calculate the reactive power; the periodic extension procedure presented in Sections 4.1 and 4.2 can thus be omitted. Also, in Eq. (20), each reactive component contains the weight coefficient $W_0^2(-\lambda_k\Delta\omega)$, which creates errors. The following method can be used to remove the weight coefficient (Wei *et al*., 2010).

Calculating

$$Z'(k) = \begin{cases} U_w(m_k), & k=0, \\ \dfrac{2U_w(m_k)}{W_0^2(-\lambda_k\Delta\omega)}, & k=1,2,\cdots,H, \end{cases}$$

and replacing $Z(k)$ with $Z'(k)$ in WDHT of $u(n)$, we can obtain a new N-point phase-shifted signal $\tilde{u}(n)$:

$$\tilde{u}(n) = \sum_{k=1}^{H} \frac{\sqrt{2}U_k}{W_0(-\lambda_k\Delta\omega)} \sin(k\omega_1' n + \pi\lambda_k + \alpha_k). \tag{21}$$

Then, the new instantaneous reactive power sequence is obtained as $q''(n)=\tilde{u}(n)i'(n)$. Filtering the harmonics of $q''(n)$, we finally obtain the accurate reactive power value:

$$Q = \sum_{k=1}^{H} Q_k = \frac{1}{M}\sum_{n=0}^{M-1} q''(n). \tag{22}$$

To calculate $Z'(k)$, we need to precisely calculate λ under the background of interharmonics. λ is always achieved using various kinds of interpolation algorithms (Liu, 1999; Zhang *et al*., 2001; Agrez, 2002; Pang *et al*., 2003; Huang, 2005; Huang and Jiang, 2005), among which the phase-difference correction method (Liu, 1999; Huang, 2005; Wei *et al*., 2010) is widely used because of its high speed and accuracy. Through experimentation, the measurement precision of λ can be ensured by using the phase-difference correction method even with large interharmonic pollution. To apply the phase-difference correction method to the proposed method, we can adopt the following strategy:

Sampling $i(t)$ and $u(t)$ for m and m+1 periods respectively, we obtain N-point current sequence $i(n)$ and (N+M)-point voltage sequence $u(n)$. The first N points of $u(n)$ correspond to m signal periods and form the sequence $u_{1st}(n)$, and the last N points form the sequence $u_{2nd}(n)$. After windowing $u_{1st}(n)$ and $u_{2nd}(n)$ with $w(n)$, single-spectrum DFT is used to obtain the phases of their fundamental waves α_1 and α_1'. The following formula has proven correct (Liu, 1999; Huang, 2005; Wei *et al*., 2010):

$$\lambda = \frac{(\alpha_1' - \alpha_1) - 2\pi Mm/N}{2\pi M/N} = \frac{m(\alpha_1' - \alpha_1)}{2\pi} - m. \tag{23}$$

5 Principle of selection of sampling periods

When using the proposed method, $\lambda_k\Delta\omega$ in weight coefficients $W_0^2(-\lambda_k\Delta\omega)$ should be within the mainlobe bandwidth of the selected cosine window. Otherwise, the measurement of higher harmonics of the voltage and current signals will be affected severely (some false higher harmonics will emerge). Especially when white noise exists in signals, the errors could be very large, sometimes even leading to burrs in the error curves of reactive power measurement. And, $\lambda_k\Delta\omega$ reaches the maximum value when k=H. Suppose the mainlobe bandwidth of the selected window is $2J\Delta\omega$. Based on the above analysis, we have

$$H|\lambda|\Delta\omega < J\Delta\omega. \tag{24}$$

Suppose the real fundamental frequency is f_1 and the default fundamental frequency in sampling is f_1' ($f_1'=f_s/M$). It is easy to conclude that $|\lambda|=m|(f_1-f_1')/f_1'|$, and Eq. (24) becomes

$$m|(f_1 - f_1')/f_1'| < J/H. \tag{25}$$

The allowed fluctuation range of the fundamental frequency is 49.5–50.5 Hz in power systems, and f_1' is always chosen as 50 Hz in sampling. Therefore, $|(f_1-f_1')/f_1'|$ is maximized to 0.01 when f_1=49.5 Hz or 50.5 Hz. Then,

$$m < 100J/H. \tag{26}$$

In this work, the premise of ensuring the reactive power measurement accuracy is the precise acquisition of harmonic components of the voltage and current signals. Since the amplitudes of interharmonics are much smaller than those of harmonics in power systems, their influence on harmonic measurement can be ignored. In practice, we need only to consider the interference among harmonics, especially that between fundamental and 2nd harmonics. To effectively restrain the spectrum leakage, m should also satisfy (Qi and Wang, 2003)

$$m > \frac{\Delta A - A_1}{BD} + \frac{J + 0.5}{B}, \tag{27}$$

where A_1 (dB) is the maximal sidelobe level of cosine windows, D (dB/sidelobe) represents the speed of sidelobe attenuation, ΔA (dB) is the amplitude resolution between two harmonics, and the span between them is B signal periods.

Generalizing Eqs. (26) and (27) yields the general requirement of m:

$$\frac{\Delta A - A_1}{BD} + \frac{J + 0.5}{B} < m < \frac{100J}{H}. \tag{28}$$

Note that Eq. (28) is not tenable in all situations. For example, set H=20, suppose the maximal interference is the leakage between the fundamental harmonic and the 2nd harmonic (B=1) and the required amplitude resolution is ΔA=−100 dB, and select a two-item Hanning window with J=2, A_1=−32 dB, and D=−6 dB/sidelobe in the measurement. Then, Eq. (28) becomes 13.8<m<10, which is impossible. To better satisfy Eq. (28), we suggest using higher-item cosine windows in the measurement, e.g., four-item cosine windows including the Blackman-Harris window, Nuttall window, and Rife-Vincent window.

In general, the selection of m should balance the real-time sampling and measurement accuracy. Under the restriction of Eq. (28), better spectrum leakage restraining effect can be achieved with a larger m, at the cost of real-time performance degradation. By selecting a smaller m, the real time performance can be ensured but at the cost of lower measurement precision.

6 Simulation and discussion

To illustrate the described method, signals in a single-phase power system which is composed of interharmonics and harmonics (Table 1) were adopted for the simulation in MATLAB. The amplitudes of the signals were selected to accord with the actual condition in practical power systems, and the phases were given arbitrarily. Set the sampling frequency to be 6400 Hz and set H=20.

Table 1 Parameters of the voltage and current signals

Components*	Voltage amplitude (V)	Voltage phase (°)	Current amplitude (A)	Current phase (°)
0.5	2.30	10	0.5	15
1.0	380.00	20	10.0	−20
2.0	5.00	17	0.6	128
3.0	15.00	35	2.0	63
3.4	1.60	33	0.8	145
5.0	11.20	120	1.3	44
5.5	1.52	−57	0.7	−33
6.0	8.40	130	0.7	−22
7.5	1.14	78	0.5	−11
9.0	5.60	−45	0.5	31

* Harmonics and interharmonics whose frequencies are multiples of 50 Hz

In Table 1, the maximal interference is the leakage of the fundamental component on the 2nd harmonic. The ratio between their amplitudes is −37.6 dB. To restrain the leakage within 0.1% (−60 dB), the amplitude resolution should be at least −97.6 dB. According to Eq. (28), when selecting Hanning, Blackman, Blackman-Harris, four-item three-order Nuttall, four-item Rife-Vincent(I), and four-item Rife-Vincent(III) windows respectively in the measurement, m should satisfy 13.4<m<10, 9.9<m<15, 7.3<m<20, 6.0<m<20, 10.6<m<20, and 10.4<m<20 respectively. As shown, m cannot be selected properly when using the Hanning window; in this case, we have to choose the maximal m under the condition of m<10 so that the spectrum leakage can be restrained as much as possible.

When using the above-mentioned windows, we selected m=9, 14, 15, 15, 15, and 15 respectively. Based on Budeanu's definition, the true value of reactive power can be easily computed from Table 1. The absolute values of relative errors of the measurement are shown in Fig. 1. Compared with those of three- and four-item windows, the measurement accuracy of the Hanning window was approximately two orders of magnitude lower. However, in this case, the measurement speed was faster because of the fewer sampling points used. Within the four-item windows, the accuracy was relatively high when employing Rife-Vincent(I) and Rife-Vincent(III) windows.

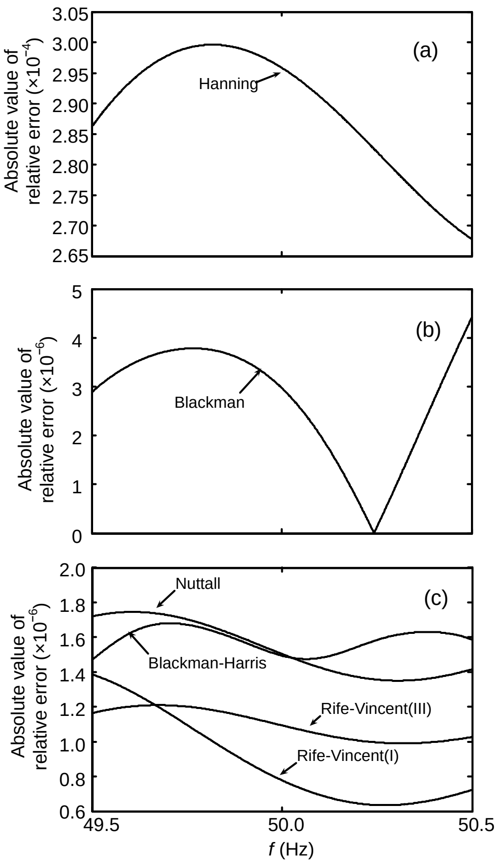

Fig. 1 Relative errors of reactive power measurement in the presence of interharmonics (signal is noise free) for Hanning (a), Blackman (b), and Blackman-Harris, four-item three-order Nuttall, four-item Rife-Vincent(I), and four-item Rife-Vincent(III) (c) with m=9, 14, 15, 15, 15, and 15 respectively

White noise is one of the commonly existing noises in power systems. We added white noises of different degrees to signals (Table 1) to further investigate the measurement precision. The simulation setting was the same as above, and the relative errors can be seen in Fig. 2 (only four-item windows are used) where the fundamental signal-to-noise ratio (SNR) was set to be 60 dB and 40 dB, respectively. Note that the data shown in Fig. 2 are the average of the results of 50 measurements.

In Fig. 2, with weak noise (SNR=60 dB) and relatively strong noise (SNR=40 dB), the relative errors were no more than 7.5×10^{-5} and 8×10^{-4}, respectively. The precision was relatively low compared with that shown in Fig. 1, but still meet the measurement requirement. Also, with the concurrence of interharmonics, synchronization error, and white noise, such ideal measurement results show the reliability and superiority of the proposed method.

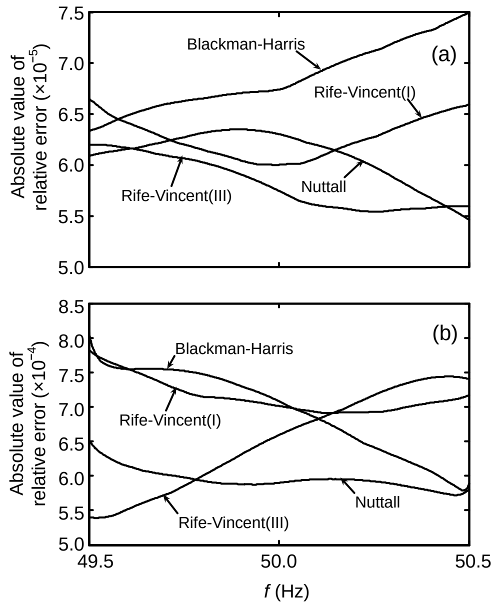

Fig. 2 Relative errors (average of the results of 50 measurements) of reactive power measurement in the presence of interharmonics with noise
(a) SNR=60 dB; (b) SNR=40 dB. Only four-item windows are used

7 Conclusions

The proposed time domain integral method effectively decreases the influence of interharmonics and synchronization errors on reactive power measurement, thus improving the measurement accuracy. Simulation also demonstrates its good effect of restraining the white noise. Therefore, this method has a good prospect of application to real power systems, which contain both interharmonics and white noise. With a proper modification, and by processing the voltage signal in the same way as the current signal, this method can also be applied to active power measurement. The deficiency of this method is that the duration of the sampling window should always reach more than ten signal periods. Thus, much memory space is occupied, and further improvement is needed for real-time measurement.

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