APPLICATION OF DISCRETE AVERAGING TO INCREASE THE ACCURACY OF MULTIPLE MEASUREMENTS

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Abstract. This article discusses the problem of eliminating errors that arise at the processing stage, taking into account the growing need for digital signal processing of measurement results in the oil and gas industry. And investigated the problems with equalizing filtering and discrete averaging in digital signal processing. Therefore, it can be said that the numerical mean operator applied for both types of error correction is intended for any error. This has been proven by modeling in Matlab.

Keywords. Signal spectrum, non-sinusoidal signal, mathematical expressions, digital measurement, noise and interference, digital average, average value.

1. Introduction

Any measured value, such as mains voltage, is a sinusoidal signal. This signal contains only one harmonic. As a result, we get a non-sinusoidal signal on the analog measurement of the value. It is a sine wave composed of extra odd harmonics. These harmonics are caused by noise and interference superimposed on the fundamental signal. Therefore, the measured signals in most cases (80%) are not sinusoidal [2, 3]. Curves The instantaneous values of a continuous waveform are most useful, but not always predictable. Therefore, for the control and analysis of controlled processes and objects, it is of great importance to determine the integral parameters of the signal in real time (ISP). The latter characterize the total amount of substance and energy that must be produced and obtained in production for a certain period of time, operating characteristics and characteristics, are average measured values, etc. Determination of specificity Digital methods and ISP tools must perform discrete averaging values (DA) or discrete integration (DI), constantly changing over time.

2. Problem statement

The operators of preprocessing and display of measurement data have certain filtering (correction) functions. Therefore, for rational hardware and software implementation of correction of filtering problems (CF) and to avoid redundancy of additional data, the introduction of the measurement channel must use the functions of preprocessing and filtering of the representative reference operators appropriately information on measurements [5-12]. Let us consider the case when an automatic control system of an analog-to-digital converter (ADC) is used as an information converter in the Armstrong circuit or for recording some random process events. In this context, the problem of converting the input signal arises. The main method of signal approximation is a step-by-step approach, otherwise called zero-order extrapolation. In this case value of the function *z(t)* at $nT_0<t<(n+1)T_0$ is assumed equal $x^*(nT_0)$, i.e. the value of the function *z(t)* in a concerned time lag is extrapolated on one sample. Approximated *z(t)* signal at the output of a zero order data-hold device (DHD) can be compared with the signal at the ADC input:

$$\varepsilon(t) = x_{inp}(t) - z(t), \tag{1}$$

where $\varepsilon(t)$ is the current value of an aggregate error from an effect of two sequentially switched subchannels: ADC and DHD.

At the DHD output step-by-step restoration of the signal z (t)is provided in that case, if the pulse response of a linear section $k_{ext}(\tau)$ with the transfer functions $K_{ext}(p)$ are a rectangular pulse with duration T_0 and the amplitude equal 1.

Thus and so, in a case of the DHD:

$$k_{ext}(\tau) = 1(\tau) - 1(\tau - T_0), \tag{2}$$

wherever

$$1(\tau) = \begin{cases} 0, \ at & \tau < 0; \\ 1, at & \tau \geq 0; \end{cases} \qquad 1(\tau - T_0) = \begin{cases} 0, \ at & \tau < T_0 \\ 1, at & \tau \geq T_0 \end{cases}. \tag{3}$$

DHD transfer function will be as shown below:

$$K_{ext}(p) = \int_0^\infty \left(1(\tau) - 1(\tau - T_0)\right) e^{-p\tau} d\tau = (1 - e^{-pT_0}) / p. \tag{4}$$

Complex frequency responses (CFR) of the linear section realizing DHD are obtained from (4) substitution p in $j\omega$.

$$K_{ext}(j\omega) = \frac{1 - e^{-j\omega T_0}}{j\omega} = T_0 \frac{\sin(\omega T_0 / 2)}{\omega T_0 / 2} e^{-j\omega T_0 / 2}. \tag{5}$$

It can be seen from (5), that DHD is a low-pass filter with a delay $T_0 / 2$.

3. Solution of the problem

The overall measurement error can be represented as four components, which include both low-frequency everyday noise and high-frequency noise. Therefore, CF should have a band-pass character and consist of at least two consecutive open filters - high-pass (to suppress low-frequency noise) and low-pass (to suppress high-frequency noise) [7, 11]. These are the second sub-channel low-pass filter (DHD), the second sub-channel low-pass filter (DHD), the second sub-channel low-pass filter (DHD); error components. Therefore, we will synthesize in the entire frequency band of the measuring signal, i.e. $-\omega_0 \le \omega \le \omega_0$.

Let high-pass *B(z)* filter is characterized CFR in the form, which is listed below:

$$B(e^{j\omega T_0}) = \sum_{i=0}^{M-1} b(iT_0) e^{-j\omega T_0}, \tag{6}$$

where, $b(iT_0)$ is discrete sample of pulse responses of the filters *B(z)*; M is the filter order.

Let's represent gain frequency characteristic (GFC) of the filter B(z) as:

$$\left|B(e^{j\omega T_0})\right|^2 = \sum_{m=0}^{M-1} b_m^2 + 2\sum_{k=1}^{M-1}\sum_{m=0}^{M-1-k} b_m b_{m+k} \cos(k\omega T_0). \tag{7}$$

Considering the successive inclusion of the high-pass filter B (z) and DHD, variance σ_Σ^2 of an output noise of the signals $\varepsilon_\Sigma(nT_0), n = 1,2,...$ at variance σ_0^2 of input noise of the signals let's define as:

$$\sigma_\Sigma^2 = \sigma_0^2 \frac{T_0}{2\pi} \int_{-\omega_0}^{\omega_0} \left|K_{ext}(e^{j\omega T_0})\right|^2 \left|B(e^{j\omega T_0})\right|^2 d\omega. \tag{8}$$

At the output of two consecutive subchannels of the processing chain we will take the minimum noise variance σ_Σ^2 as the criterion for suppressing the resulting error, i.e. solving the optimization problem for (8), we receive the filter coefficients.

$$\widehat{b}_0 = 1, \quad \widehat{b}_1 = -1, \quad where\ M = 2. \tag{9}$$

$$\widehat{b}_0 = 1, \quad \widehat{b}_1 = -2, \quad \widehat{b}_2 = 1, \quad where\ M = 3. \tag{10}$$

Using the method of mathematical induction and taking into account the special case of the filter B (z), it will be possible to conclude that with super sampling the limiting linear-impulse filter B(z) will approach the finite difference filter (FFD) with the sets of coefficients:

$$b_m = (-1)^m C_{M-1}^m, \quad m = \overline{0, M-1}. \tag{11}$$

As a result, the CFR of the non-recursive filter B (z) with the sets of coefficients (11) will have the following form:

$$B(e^{j\omega T_0}) = \sum_{m=0}^{M-1} (-1)^m C_{M-1}^m e^{-jm\omega T_0}. \tag{12}$$

In formulas (11) and (12), C_{M-1}^m is the set of node coefficients of the quadratic formula (coefficient of each node of the system). In solving the problem of global optimization, the correlation between individual indicators must be taken into account. In this case, the vector of efficiency and quality indicators of the optimized system should be evaluated from the $Q_{opt} = \{q_{1.opt}, q_{2.opt}, \ldots, q_{1.opt}\}$ condition to minimize total losses of $C_\Sigma \to \min$.

Thus, it is possible to estimate the efficiency of the filter B (z) when it is serially connected to DHD by the suppression coefficient of the input noise variance in relation to the variance of the output noise $K = \dfrac{\sigma_0^2}{\sigma_\Sigma^2}$. As a last resort, i.e. at $T_0 \to 0$ is received $\widehat{K} = \lim\limits_{T_0 \to 0} K \to \infty$.

Therefore, with super sampling, the edge pulse filter for the rejection ratio of two series-connected filters approaches infinity and indicates the effectiveness of the series connection of the FFD to the DHD. The above procedure also applies to higher order DFDs.

Among digital methods of measuring ISP - a widely used method of digital processing of the results of direct measurements of instantaneous signal values during averaging (integration). In recent years, interest in this method has grown significantly due to the possibility of inputting computing power into the measurement channels when solving such measurement problems. This problem is further compounded by IRS digital measurements of complex electrical waveforms (non-sinusoidal signals). However, the well-known advantages of electrical control and measurement methods for physical quantities contribute to the more common primary data converters with output in the form of DC and AC currents and voltages. AC signals are more informative and in some case is the only possible forms of obtaining measurement data, especially in objects of conversion and generation of electrical energy [4-11].

Considered the case where a nonsinusoidal signals interacts with a randomly centered signal; to find out what is the spectrum of the product of this signal [9, 10]. Since it accepts a superimposed sine wave on a non-sinusoidal signal source, the odd 13 harmonics are as follows:

$$y(ik) = \sum_{n=1}^{2N+1} sin(\omega knT_0), \quad N = \overline{1,6}$$

ω - carrier frequency of the signals, T_0 - sampling of the frequency (T_0 = 1/1300), к - discrete sample numbers, n - harmonic numbers.

As a random signal that is considered uniformly distributed in the range [0, 2], the signal generated by the built-in rand in Matlab program. The spectrums of the original signals are calculated in the Matlab environment using the fast Fourier transform function fft. Fig. 1 (a, b) and Fig. 1 (v, q) - shows the original signals and the spectrum of the original signals.

To carry out control classes of continuous functions $\tilde{C}^{(m)}[0,T]$, a quadrature formula is used in the form:

$$\bar{I} = \frac{1}{M}\sum_{i=0}^{M-1} y(iT_0), \tag{13}$$

where, $T_0 = \frac{T}{M}$ – is the sampling step in time.

In addition, the resulting product signals are not sinusoidal, and the resulting random signals are obtained, the spectrum of which is shown in Fig. 2 (a). Seen in Fig. 2 (a), the resulting spectrum contains a certain number of harmonics in its structure (13).

We used the method of individually averaging the resulting signal at regular intervals to suppress these harmonics. Calculation’s shown in Fig. 2 (b) confirm that additional harmonics in the spectrum significantly decrease their influence. The signal amplitude was about 2, so we can say that the suppression of the spectrum was almost 1000 times. And this proves the probabilities of correcting the investigated discrete averaging operator over non-sinusoidal signals.

Similar calculation was made in relation to the non-sinusoidal signal of the form (13) and systematic error of about 2 described function of the form:

$$\bar{\varepsilon}(t_k) = \sum_{j=0}^{2(N+1)} a_j t_k^j \tag{14}$$

where t_k - discrete samples bias a_j - polynomial coefficients describing this slowly changing error.

Fig. 3 (a, b) and 3 (v, q) show the original signals and their spectra - offset and non-sinusoidal signal range.

Fig. 4 (a) shows that the spectrum of the resulting signal is equal to the product of a non-sinusoidal signal and systematic errors. The result of control on the received signal is shown in Fig. 4 (b).

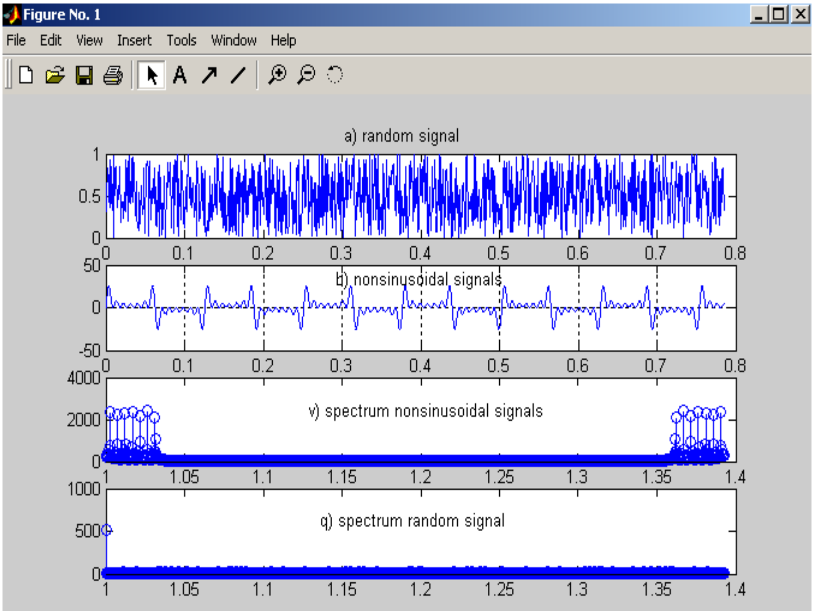

Fig. 1. Spectrum of the input signal
a) the random signal (interference); b) the non-sinusoidal signal;
v) the non-sinusoidal signal spectrum; q) the spectrum of a random signal.

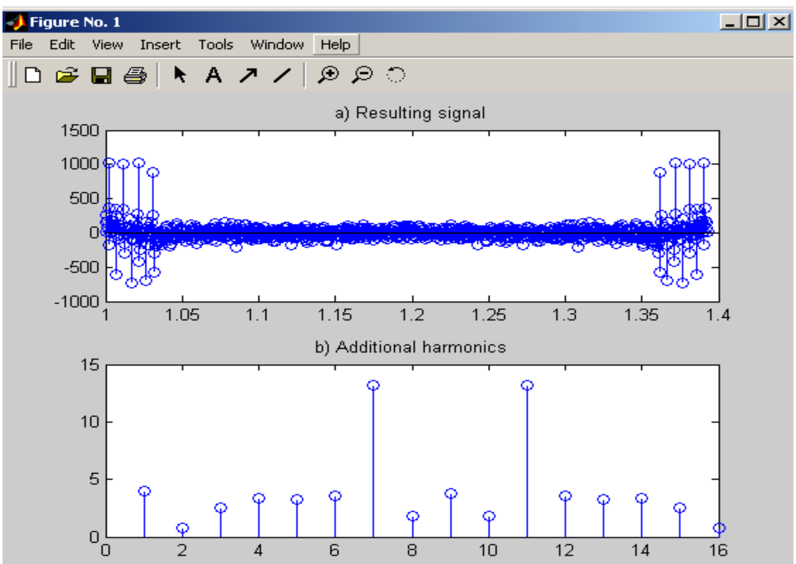

Fig. 2. Significant reduction in the influence of additional harmonics
a) the nonsinusoidal signal spectrum; b) discrete signals spectrum after an averaging

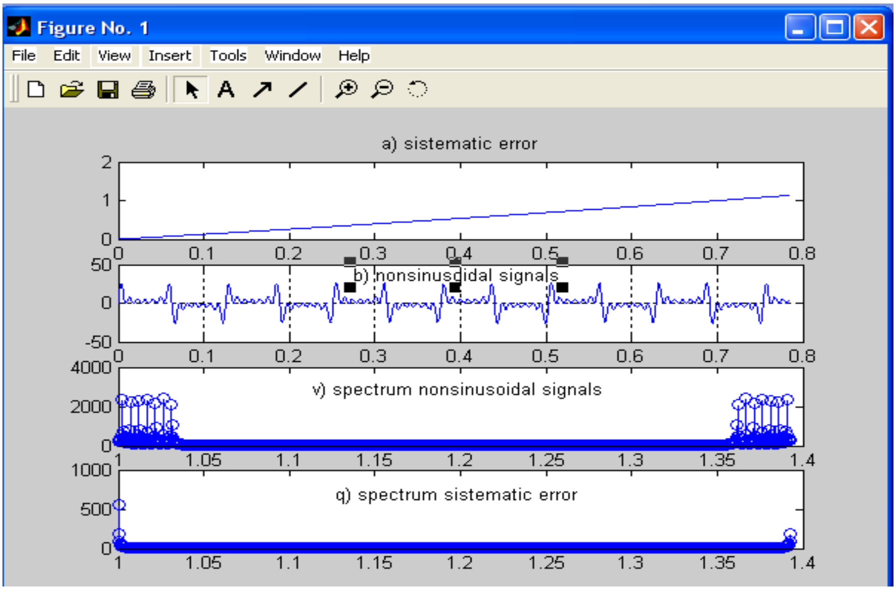

Fig. 3. Original signal spectrum and their spectra - offset and non-sinusoidal range
a) the systematic error; b) the nonsinusoidal signal;
v) the nonsinusoidal signal spectrum; q) the range of systematic errors.

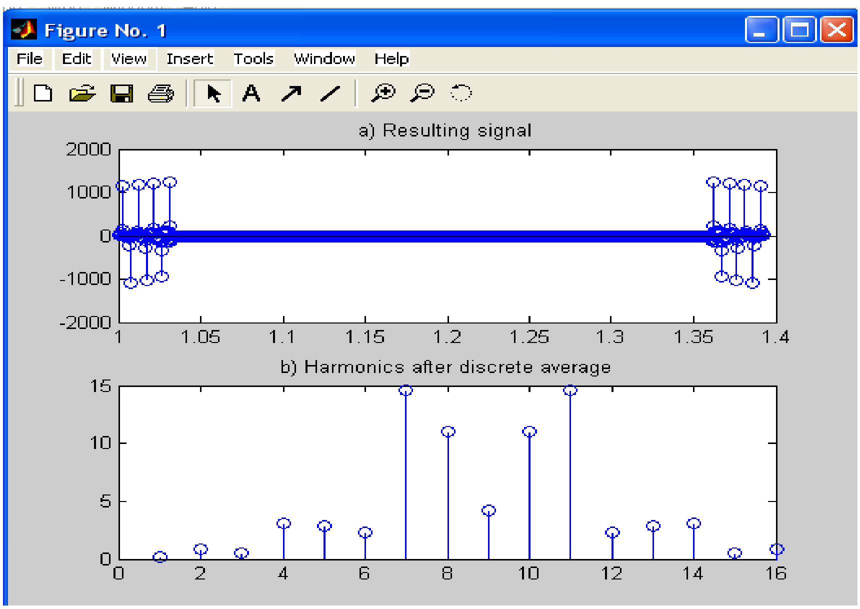

Fig. 4. Spectrum of the resulting signal
a) the spectrum of the input signal; b)the spectrum after digital averaging.

4. Conclusion

Most of the operations that can be performed in the preprocessing and data representation extraction of a modern data measurement system are linear and smoothing, so they have a smoothing function for low-pass filters. The equalization filter should be of a high frequency nature to remove noise in the preprocessing sequences and the input of the presentation data subchannel. The optimal high-pass correcting filters we have synthesized provide sufficiently high suppression coefficients for the corrected variance of the subtracted output noise. Comparative analysis of experimental results with systematic and random errors shows that bias is better suppressed. Therefore, it can be seen that, application of the discrete averaging operator for both types of errors is a correction method for all types' errors.

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