Corrective Filtering of Conversion and Digital Processing Results of Non-Sinusoidal Signals

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Abstract

The systematic approach to increasing the accuracy of information-measuring systems was proposed from the perspective of the concept of corrective filtering. Considering that the information-measuring systems under study are dynamic systems and that the important attributes of such systems are ADC, the modern requirements imposed on these converters were systematized. Based on the analysis of existing works, the possibilities of reducing methodological error were considered, and the importance of cleaning the sub-integral function from noise, that is, reducing instrumental errors, was substantiated. The corrective properties of the discrete averaging operator during digital measurements of the integral parameters of non-sinusoidal signals were studied. The fact that this operator has a corrective character with respect to the systematic and random errors of the instantaneous values of alternating current signals was confirmed by statistical modeling.

Keywords

Corrective filtering, control system, PID controllers, non-sinusoidal signals, conversion processing, digital processing.

AMS Subject Classification: 28A25, 28A20
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Introduction

The development of industrial production in our republic is organically linked to the efficient operation of power plants, which in turn requires that the equipment forming the basis of these facilities meets modern standards. One of the important issues in this direction is the creation and application of measuring instruments that form the basis of efficient information exchange in industrial automation systems. For example, the accuracy of measuring transducers, which are one of the main elements of automatic control and diagnostic tools for electromechanical equipment, is an important factor determining the correctness and efficiency of the decisions made.

In recent years, digital measuring instruments (Mİ) and methods of integral parameters of alternating current signals (ACS), like other areas of measurement technology, have been rationalized both methodologically and algorithmically and technically. However, the achievements in this direction are not significant compared to the successes achieved in the field of creating systems and digital devices for measuring the parameters of DC signals. The main reason for this is several methodological and technical problems that arise during the digital measurement of the integral parameters of ACS.

Due to the rapid increase in the demand for electric energy in current time, one of the important issues in the creation of electric power transmitters and meters with high metrological, technical-economic, constructive-technological and operational indicators is related to the measurement of active power under conditions of non-sinusoidal currents and voltages. Thus, the widespread use of nonlinear elements in electrical equipment is the main reason for the increase in distortions of energy-carrying signals in the power supply lines of the corresponding devices. A lot of theoretical and practical research has been carried out on this issue and certain results have been achieved. However, the problem of measuring active power and energy with the required accuracy within practical technical-economic limitations is still open. To substantiate this conclusion, it is enough to emphasize that the fleet of electromechanical energy meters in both household and industrial facilities is quite extensive today.

ACS in most cases is non-sinusoidal in nature. While the cause of the formation of high harmonics is noise and interference added to the main useful signal during digital measurements of sinusoidal signals, high harmonics in non-sinusoidal signals (NSS) are associated with the processes of generation and consumption of (usually single) electrical energy.

Although the instantaneous value curves of continuous signals are more informative, they are not always easy to understand. Therefore, for the control and analysis of controlled processes and objects, the determination of the integral parameters of signals (IPS) on a real-time scale is of great importance. These parameters characterize the total amount of matter and energy input and output in production over a certain time interval, and the mode indicators consisting of the average values of the measured parameters of the object. The peculiarity of the determination of IPS by digital methods and means is the implementation of discrete averaging (DA) or discrete integration (DI) of quantities that change continuously in the observation interval.

Among the digital measurement methods of IPS, the method of direct measurement results (DO, DI) of instantaneous values of signals in the observation interval is widely used. Recently, interest in this method has increased significantly, which is explained by the possibility of creating virtual devices by introducing computational means into the measurement channels for digital processing of signals. This approach is very important during digital measurements of IPS of non-sinusoidal electrical signals of complex shape. Thus, the well-known advantages of electrical methods of control of physical quantities and their measurement provide the basis for the widespread use of primary information converters in modern automation and control systems, the output signals of which are in the form of alternating current and voltage. ACS is more informative, in some cases, especially at facilities for the production and conversion of electric energy, it is the only possible form of obtaining measurement information. In modern information-measuring practice, mathematical expressions (algorithms) for determining integral parameters, which are widely used in digital processing of ACS, are diverse. When implementing these algorithms using numerical methods and tools, it is necessary to switch from continuous integration to numerical integration. For this purpose, various quadrature formulas are used.

The main errors that arise during digital measurement of integral parameters under NSS conditions include:

1) methodical errors;
2) instrumental errors - deviations;

- deviation in formation of the observation (integration) interval proportional to the period;
- deviation in formation of the sampling step, in the separation of instantaneous signal values by amplitude and time (phase);
- spectral leaks due to a discrepancy between the sampling frequency and the frequencies of higher harmonics;
- noise caused by a discrepancy between the signal source and the parameters of the load (consumer);

- errors caused due to formation of systematic components (trends) in signals when the mode (technical condition) of the load is violated.

Analytical and experimental research and correction of measurement errors arising for the above reasons are of great practical importance.

Currently, digital measurements of IPS are mainly directed at creating virtual devices and information-measuring systems (IMS). However, unfortunately, in the existing works carried out in this direction, a systematic approach to solving the problem has not been effectively used. Therefore, both the analysis of the above-listed instrumental errors and their correction issues have been studied within the framework of the creation of this or that specific device or system.

In terms of the wide application of modern computer and information technologies, the structural and algorithmic unification of digital measuring devices and systems of IPS is on the agenda as an urgent issue to move from concreteness to systematicity. In the presented article the solving of these urgent issues is considered.

To improve the efficiency of control and management of electric power generation and consumption facilities and to develop methods and tools for increasing the accuracy of measurement information formed by NSS conversion and digital processing, the following research was conducted in the main directions:

1. Analysis and systematization of methods and tools for assessing and improving the quality and efficiency indicators of the IMS.
2. Composition of a generalized metrological model of NSS conversion and digital processing tracts and development of a methodology for assessing instrumental errors on this basis.
3. Development of a methodology for determination and assessment of corrective filtering (CF) resources in the process of information exchange despite the unified algorithm and structure of NSS conversion and digital measurement tracts.
4. Development of methods and tools for assessing and improving the efficiency of indicators of CF in the process of digital processing of NSS conversion results based on specific algorithms

Statement of the problem

It is noted that the input signals from the devices used in the field for the automatic control of technological processes are processed based on the program written to the controller and given to the execution mechanisms, and for this, the remote-control building and the central control building are used. Some of the existing facilities in an oil refinery do not have this type of central control room or remote-control building for process control and automatic control. Control of technological devices is carried out by means of "synoptic panels" mounted on the wall or modern DCS control system and PC-based workstations installed in some units.

The devices used in the field are pneumatic-based, old devices that sometimes have inaccurate instructions, malfunctions, and often break down. Since the degree of automation is very low, they try to adjust and manage the process manually. An emergency stop system to prevent accidents and a fire and gas control system to detect toxic gases in the field are not installed. We face similar problems in the crude oil primary processing plant.

The primary crude oil refinery unit, otherwise known as the Electric Atmospheric Vacuum Desalting unit, has been in operation since 1976 and supplies the refinery with feedstock for secondary processing. Instead of a central control room, wall-mounted weather panels and the Honeywell Minitrend Qx/Yokogawa Cx100 system are used to control the plant process. Although pneumatic controllers were used in their early days, electrical type units supplied by Emerson or Foxboro are more commonly used today. In addition, domestic regulators and recorders are devices from Russian companies, as well as Honeywell [1-4]. The unit does not have an ESD emergency stop system and an FGS fire and gas control system. Field processes are carried out manually by operators according to accepted procedures. Honeywell devices are most often used as sensors, and Masoneilan valves are used as actuators. The device does not have analyzers for product quality control. Instead, there are measurement points at which samples are taken and analyzed in the laboratory [5, 6].

Since there is no automatic control system, it is impossible to monitor the process values through the central control building or any HMI, automatically control the system parameters and continuously monitor the process on-line, each time the field operator must manually control the process from the site to monitor the parameters need to be adjusted. This leads to loss of time and workers. At critical moments, the process cannot be controlled by management and the process efficiency statistics are low [7-10].

Another important importance of controlling process parameters is that the raw materials of newly constructed facilities use the products produced at that facility and make the necessary adjustments depending on the incoming product. For this reason, information about the pressure, temperature and composition of the product being manufactured must be known in advance. For example, the product for a diesel fuel hydrotreater will come from a refinery. For diesel to produce products in accordance with the intended Euro V standards, necessary adjustments must be made to account for incoming raw materials. Since this cannot be done with the current control system, the old control system must be replaced with a new one. The new control system being created should ensure the operation of many existing and newly built oil refinery units for a period of 4-5 years, that is, until a new oil refining unit is built and put into operation [11, 12].

Thus, to eliminate the above problems and improve the quality of the system, work was planned based on the “Modernization and Reconstruction Project” and a project was drawn up for the consistent implementation of the planned work to achieve a certain goal.

Based on the above, the project consists of 3 main stages, which include the design stage, the procurement stage and the construction and commissioning stage.

The design phase consists of four main parts:

Concept – the concept provided by the customer is refined and the cost and duration of the project are determined accordingly. A list of required documents will be displayed.

Input engineering submission - preparation of documents for use by departments, sending to the customer, correction of discrepancies received by the customer, and then preparation of procurement forms associated with the purchase.

Detail Design – Documents prepared in the preliminary design are rechecked and transferred to the customer.

The next stage is to correct changes in documents during the project and provide the necessary support to the customer during construction and commissioning [13-17].

The field of mechanical engineering has various branches such as NEC, electrical, mechanical, process, construction and safety. Each of the branches implements the project in accordance with the client's requirements in accordance with its responsibilities [18].

Solution to the problem

Any functional node or processing operator used in measurement information exchange tracts in the IMS has this and other filtering properties. Therefore, during the study and development of these tracts, these properties should be revealed and analyzed and used effectively as a metrological resource. If these resources do not allow meeting the requirements for the efficiency and quality indicators of the IMS, then additional methods and means of corrective filtering should be used.

Naturally, the main purpose of all functional nodes included in the “measurement-processing” tract is to transmit the measurement information carried in the output signal x(t) of the primary transducer (PT) to the output with as little loss (distortion) as possible [19].
Corrective filtering is understood as a method of processing the useful signal and the error signal (noise) together to estimate the informative parameter of the measurement signal (current value, function or functional) with the highest possible accuracy [18-20]. Therefore, the efficiency of the CF, determined by the criterion of the proximity of the “measurement-processing” result to the true value of the informative parameter of the signal, is an indicator of the level of accuracy that can be provided by the IMS. Accuracy is the most objective indicator of the efficiency of the work of any MI.
The aspects that should be considered in the formulation and solution of the CF problem are systematized in [22]:
- The input signal of each sub-tract of the IMS can be considered as an “additive mixture” of the measurement signal and noise (error);
- The most adequate model of the error is a non-stationary process in the continuous time domain, and a non-stationary sequence in the discrete time domain;
- Due to the variability (dynamics) of the operating conditions of the “measurement-processing” tract, the stochasticity of the processes taking place in it (due to noise), the a priori information about the useful measurement signal and noise is always incomplete;
- Due to the diversity and specificity of the problems of converting and processing measurement information, the metrological, dynamic and operational efficiency of most components used in the “measurement-processing” tracts is limited;
- The indicated limitation and “non-ideality” lead to the formation, transformation and accumulation (accumulation) of various types of errors along the tract;
– Most mean square error operators are linear, which is why they are sensitive to both components of the additive "signal + noise" mixture being processed.
Due to the above facts and factors, quasi-optimality rather than optimality is achieved in the formulation and solution of the CF problem in the IMS.
In order to reveal and analyze the CF effect in the "measurement-processing" tracts of the studied İMS, it is important to analyze the transformation (or processing) operators $H_x(p)$ and $H_\varepsilon(p)$ in the time or frequency domains, which characterize the sensitivity of any functional node or processing operator to the useful measurement signal (x) and noise (error signal) (ε).
Naturally, taking into account the relevant restrictions, the degree of usefulness for these operators should be sought analytically or trivially in the form of amplitude$|H_x(p)| \to 1$ and phase $\phi_x(p) \to 0$, for the useful measurement signal, and $|H_\varepsilon(p)| \to 0$ for the noise (where p is the operator corresponding to the time or frequency domain).
The issue of correcting measurement results in the Mİ has already become a classic issue (many authors call this issue “error correction”) and is always in the center of attention of specialists working in this field. This is natural, because the input and output “product” of any information exchange system is information. Therefore, the effectiveness of the decisions made directly depends on the accuracy of the information.
Methods and means of CF of measurement information were analyzed comparatively in [23]. CF methods and means are divided into two main classes, depending on the nature of the measurement signals processed in the sub-tracts of the “measurement-processing” tract and the information exchange processes performed (analog, discrete and digital):
- structural;
- algorithmic.

Structural methods and means serve to increase the accuracy and stability of the Mİ by introducing additional elements and blocks (for example, a feedback device) into the structure of the created functional node.
Algorithmic methods are implemented using software tools, and their implementation is oriented towards classical and modern filtering algorithms (mainly digital filters) [14, 23, 24].

Corrective filtering resources in the real "measurement-processing" tract

Let us examine the modules included in the real "measurement-processing" tract in the metrological model shown in Figure 1 from the perspective of CF resources.

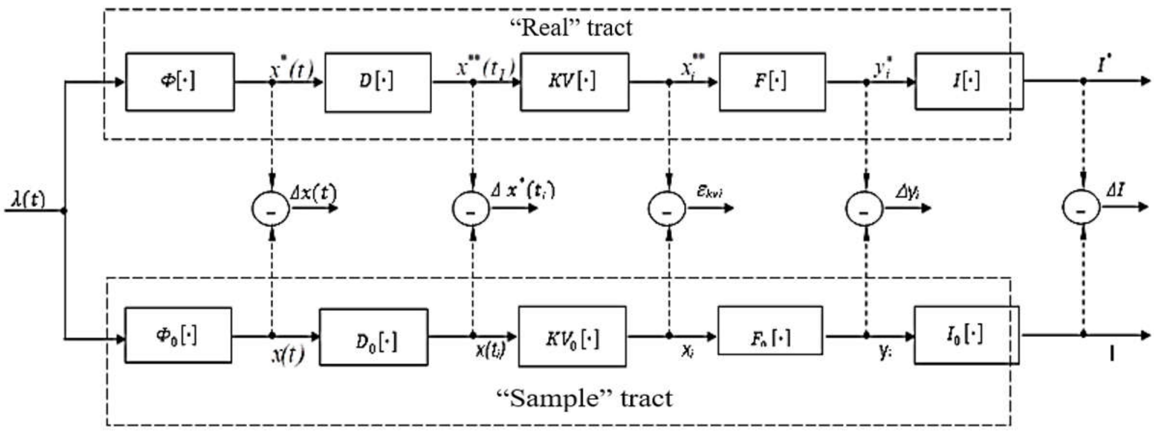

Figure 1: Metrological model of a single-channel measurement-processing tract

The primary transducers of measurement information are the core of the Mİ, and the improvement of their metrological characteristics is mainly oriented towards the use of structural measurement methods and means. This orientation is explained by the fact that the conversion processes taking place in the PT are fundamentally analog in nature. Most of the PT (alternating current and voltage transducers) applied in the subject area (power and electrotechnical facilities) are also analog in nature. Their basis is formed by measuring transformers, shunts, dividers and analog filters, for which constructive-technological CF methods and means are more appropriate.
The main problem in creating an PT is to ensure the stability of the parameters of this function (a_0, a_1, ..., a_{p-1}) by achieving some degree of linearity of their conversion functions (i.e., limiting the order of the polynomial $\Phi_{nom}[\lambda(t)]$ to (p-1).
A general formulation of this problem in the frequency domain is given in [25-28]. In this case, the authors consider the measurement signal received at the output of the PT as the result of the passage (transmission) of any exciting effects (processes) through the dynamic system.
A dynamic system is described by one or another (generally multi-order, nonlinear, variable or with random coefficients) differential equation. In specific cases, a nonlinear system can be linearized.
The most complex and at the same time the most widespread of the measurement signals processed in the IMS is the continuous function of time x(t). This function is generally the realization of a non-stationary random process x(t).
x(t) process at the output of a dynamic system described by a linear differential equation can be considered a stationary process if the operator of the system is time invariant and its input process λ(t) is stationary. The energy spectrum of the output process of a dynamic system that meets these conditions is expressed as follows:

$$S_x(\omega) = |W(j\omega)|^2 \cdot S_\lambda(\omega),$$

where: $W(j\omega)$ - complex transfer function (frequency response) of a dynamic system, and $S_\lambda(\omega)$ - is the energy spectrum of the exciting effect on the system.
In general, the following notation can be used:

$$W(j\omega) = W_{AO}(j\omega) \cdot W_{PT}(j\omega).$$

where $W_{AO}(j\omega)$ and $W_{PT}(j\omega)$ - are the complex frequency response of AO and PT, respectively. From the last expression it can be made decision that the following criteria should be used to select the PT within the frequency band $\omega \in \omega_{AO}$ of the AO:

$$|W_{PT}(j\omega)| = 1,$$

$$\phi(j\omega) = 0, \quad 0 \le \omega \le \omega_{AO}.$$

It is precisely if these conditions meet that the output process of the dynamic system X(t) and its realization x(t) carry information about the exciting effect λ(t) on the AO. Thus, following equal is valid.

$$S_x(\omega) = |W_{AO}(j\omega)|^2 \cdot S_\lambda(\omega)$$

It is clear, that the correct solution of the above problem is possible based on complete a priori information about the excitation effects λ(t) on the AO and its complex transfer function $W_{AO}(j\omega)$. Otherwise, triviality and deviation from optimality will occur in the selection of the PT. In this case, as a way out of the situation, it is necessary to solve the inverse measurement problem $\lambda(t) \to \Phi^{-1}[x^x(t)]$ on the basis of a priori information about the nominal transfer function of the PT $x(t) = \Phi_{nom}[\lambda(t)]$ (usually this information is provided in the technical data of the PT).
It is necessary to determine the CF resources of the DI in terms of the possibilities of reducing its error. In this case, two resources of the CF attract attention:
1) sampling step T_0;
2) finite differences $\Delta^1 x(t_i)$ and $\Delta^1[\Delta x(t_i)]$.
T_0 also affects the methodical error of the DA. But it is noteworthy that reducing step T_0 is advisable from the point of view of suppressing both the methodical error $\Delta I_m(M, y)$ and the instrumental sampling error $\Delta x^*(iT_0)$. Thus, the following effects are available.

$$\lim_{T_0 \to 0} \Delta I_m(M, y) \to 0, \tag{1a}$$

$$\lim_{T_0 \to 0} \Delta x^*(iT_0) \to 0 \tag{1b}$$

The existence of conditions (1) indicates the adequacy of the corresponding error estimation expressions (the condition T_0=0 is characteristic of an analog system, i.e., a transition from discrete integration (Σ) to continuous integration is made, and discretization is inevitable in this case). On the other hand, a transition to the prototypes of DA and Dİ are performed to the “limit impulse” (excessive discretization $T_0 \to 0$).

In real discrete and digital systems, the discretization step T_0 is limited from below (due to the inertia of functional nodes) and from above (spectral distortions). Therefore, depending on the algorithms for further processing of discretization results (i.e., the problem being solved), experts always pay attention to the evaluation of the discretization step based on one or another criterion (optimal, quasi-optimal, rational, etc.).
From the point of view of systematicity, the generalized criterion of the greatest inclination obtained in [29] attracts attention. Based on this criterion, a discretization step T_m is searched for that can be omitted in a given error band (Δ) during approximation, such that the time interval depends on the maximum rate of change (V_{max}) and the acceleration (a) of the approximated signal:

$$T_m = \frac{2\Delta}{V_{max}\frac{V_{max}}{a}}. \tag{2}$$

In [30], the author insists on originality in solving the approximation problem and justifies the use of the derivatives of the measurement signal f(t) as the most acceptable characteristic for an objective assessment of the discretization frequency f_D under conditions of a priori uncertainty:

$$f_D = \frac{1}{\pi}\sqrt{|f"(t)_{max}||_D|}. \tag{3}$$

Here: $f"(t)$ - is the second derivative of the signal, Δ_D is the discretization error.
[31] used the Taylor series decomposition of the correlation function to estimate the variance δ_Δ^2 of the signal reconstruction error based on various basis functions. Based on this approximation, the following normalized quantity is used as a criterion for the proximity of the process obtained because of the reconstruction to the initial process:

$$\delta_\Delta^2 = \frac{\sigma_\Delta^2}{\sigma_x^2}. \tag{4}$$

Where $\delta_x^2 = \int_0^2 G(\omega)\,d\omega$ – is dispersion of initial process (signal).
In the linear approximation of a process that is infinitely differentiable in the mean square sense and has a correlation function $K(\tau) = \frac{\sin\omega_y\tau}{\omega_y\tau}$, the quantity (4) is evaluated as follows

$$\delta_\Delta^2 = \frac{1}{75}\cdot\left(\frac{\omega_Y}{F_0}\right)^4, \tag{5}$$

where ω_Y- is upper frequency of spectrum; $F_0 = 1/T_0$- is discretization frequency.
The final problem to be solved is not the recovery of the discretized signal, but the estimation of its integral parameters by means of DA. Therefore, the discretization step must be evaluated, selected, and justified corresponding to this problem.
By study of the probability-correlation criteria of discretization [32] (searching for the Nemirovsky functional), it was shown that if the correlation interval $\tau = \int_0^\infty r(t)dt$ of a random stationary process with mean square differentiability and variance δ_x^2 is known, then the mean square value of the DA error Δ^2 can be estimated by the following formulas:

$$\Delta^2 = M[(\bar{x}-\tilde{x})^2] = \begin{cases}\sigma_x^2(\Delta T_0)^2/2T^2, & T_0 < 2\tau,\\ \sigma_x^2\,(T_0-2\tau)/T, & T_0 > 2\tau.\end{cases} \tag{6}$$

Here

$$\bar{x} = \frac{1}{T}\int_{t_1}^{t_2} x(t)\text{dt}, T = t_2 - t_1; \tilde{x} = \frac{1}{N}\sum_{i=n_1+1}^{n_2} x(iT_0); N = \frac{T}{T_0} = n_2 - n_1; \; n_2 = \frac{t_2}{T_0}, \; n_1 = \frac{t_1}{T_0}.$$

In expression (6), the relative errors of the formulas correspond to the upper and lower conditions are evaluated in the following limits, respectively

$$\delta_1 = \frac{r'^{'(0)}}{3\sigma_0^2}$$

and

$$\delta_2 = 2\tau^2/T(T_0 - 2\tau)$$

(here $r'^{'(t)}$- is the second derivative of the normalized r(t) correlation function).
Taking expression (6) into account, the following formula was obtained for the required discretization step:

$$T_0 = 2\tau + T\Delta_{bb}^2/\sigma_x^2. \tag{7}$$

here Δ_{bb}- is a valid DA error.
The first term on the right side of expression (7) is the value of the discretization step $T_{01} = 2\tau$ such that the dispersion of DA is equal to $\sigma_x^2/2n^2$. When the discretization step is smaller than the interval 2τ, the equivalence condition of analog and discrete averaging is obtained.
For example, if $R_{xx}(t) = \sigma_x^2 \, exp\{-\alpha t^2\}, \alpha = 0{,}5$, at the T=251 interval of this process, under the condition $\Delta_{bb} = 0{,}005\sigma_x^2$, we get τ≈1.25 and for the desired discretization step is obtained

$$T_0 = 2\tau + T\Delta_{bb}^2/\sigma_x^2 \approx 3{,}7$$

However, if we accept $T_0 = T_{01} = 2\tau = 2{,}5$, then Δ^2 sharply is reduced following value:

$$\Delta^2 = \frac{\sigma^2 T_0^2}{2T^2} = 0{,}00005\sigma_x^2$$

It is more appropriate to use the following formula to estimate the discretization step during DA of signals belonging to the class $\tilde{C}^{(m)}[O,T]$

$$T_0 = \sqrt[k]{\Delta_{mh}(M,y)/\varepsilon_q(M,y)}, \;\; q = \overline{1{,}5}. \tag{8}$$

Where $\Delta_{mh}(M,y)$ - is the upper (limit) value of the methodological error of DA; $\varepsilon_q(M,y)$ - is specifically chosen functional.
The value determined by expression (8) is the upper limit of the discretization step. When real functional nodes of the IMS allow for balancing the “accuracy-speed” relationship, choosing the parameter T_0 below the limit estimated by expression (8) leads to an additional beneficial effect in terms of increasing the metrological efficiency of the system. To analyze this effect, let us examine the CF property of the finite difference operator $\Delta^1[\cdot]$, which determines the instrumental error. For this purpose, let us consider the following expression:

$$\Delta x^*(iT_0) = T_0^2\{\Delta_m^1 x(iT_0) + \Delta_m^1[\Delta x(iT_0)]\}. \tag{9}$$

where: $\Delta_m^1 x(iT_0) = \frac{1}{2}[x(iT_0) - x((i-1)T_0)]$ and $\Delta_m^1[\Delta x(iT_0)] = \frac{1}{2}[\Delta x(iT_0) - \Delta x((i-1)T_0)]$ – are the input signal $x(t)$ of the DI and the inherent error $\Delta x(t)$ at its input are normalized first-order finite differences.

Arbitrary-order normalized finite difference filters (NFDF) were studied in [33].If the energy spectrum of the input signal $x(t)$ of the DI is $G_x(\omega)$, and the energy spectrum of the hereditary error Δx(t) is $G_{\Delta x}(\omega)$, and assuming the stationarity of these processes, we can write for the energy spectra of the corresponding sequences obtained as a result of filtering them through a first-order NFDF:

$$D_x(e^{j\omega T_0}) = G_x(\omega)\left|H_\Delta(e^{j\omega T_0})\right|^2, \tag{10 a}$$

$$D_{\Delta x}(e^{j\omega T_0}) = G_{\Delta x}(\omega)\left|H_\Delta(e^{j\omega T_0})\right|^2. \tag{10b}$$

Where $\left|H_\Delta(e^{j\omega T_0})\right|^2 = sin^2\left(\frac{\omega T_0}{2}\right)$- is the square of the modulus of the complex frequency response of the first order NFDF.

As can be seen, as a high-frequency filter, the NFDF suppresses the frequency complexes $\omega T_0 < \pi$ of the spectra of both the $\{x(iT_0)\}$ and $\{\Delta x(iT_0)\}$ sequences included in the right-hand side of expression (9) (Fig. 2), and the frequency complex $\omega T_0 = \pi$ is transmitted to the output with a sensitivity of $K(\pi) = 1$.

Thus, the use of the DI in the considered "measurement-processing" tract is appropriate in terms of its specific error $T_0^2 \Delta_m^1 x(iT_0)$ and the corrective filtering of the transformed inherit error $T_0^2 \Delta_m^1[\Delta x(iT_0)]$ at its output (both $T_0 \to min$ and the filtering capability of the NFDF).

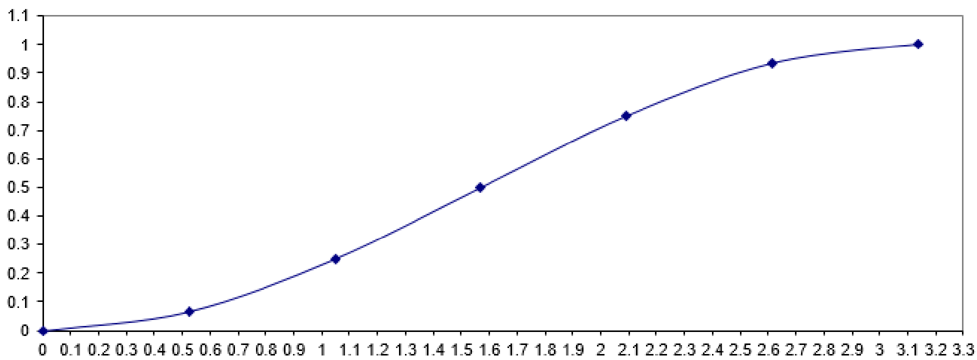

Figure 2: The process of suppressing low-frequency components

It is known that the model of the quantizing module (Fig. 3) is actually nonlinear.

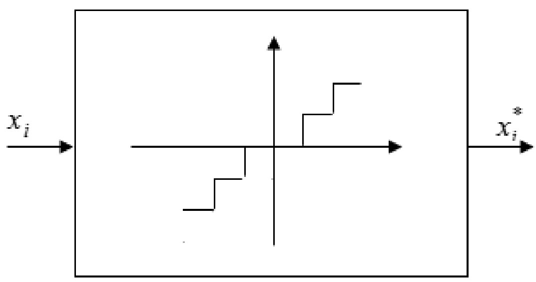

Figure 3: Nonlinear transform function of the quantizer module

The transition from a nonlinear model to a linear model is possible if at least 4 assumptions will made about the input sequence $\{x_i\}$ and the quantization noise $\{\varepsilon_{qui}\}$ and at the same time, the condition $\rho = \sigma_x/h < 0{,}5$ is satisfied [31].

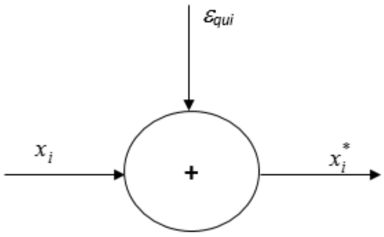

Figure 4: Linear model of the quantizer

On the other hand, the models of the quantizer and quantization noise must be justified depending on the procedure for further processing (using) the output sequence $\{x_i^*\}$ of the ADC. In this sense, it is necessary to distinguish two cases:
- digital measurement of the instantaneous value xi of the continuous signal x(t) (single-point measurements);
- digital measurement of the sequence of instantaneous values $\{x_i\}$ of the continuous signal x(t) (multi-point measurements) and processing $F[\{x_i\}]$.
Naturally, in the first case, the requirements imposed on the ADC are more stringent, and accordingly, the selection and justification of the quantizer and quantization noise models require greater attention.
In the second case, the shortcomings in the selection of models can be eliminated to one degree or another, depending on the CF effect of the operator $F[\cdot]$.
In most existing studies, quantization noise has been evaluated in relation to the problem of continuous signal reconstruction. In [21], the operator $F[\cdot]$ is used for interpolation, and in [11], the effect of the total result of quantization noise on the measurement accuracy was studied when this operator is used for the procedure of estimating the first and second order moments of random signals.
Methods and means of correcting the errors of the ADC (including quantization noise) during single-point digital measurements of continuous signals have been thoroughly studied and developed in [12] and [31]. Thus, quantization noise in single-point measurements can lead to a significant distortion of the measurement result. For example, during single-point digital measurements of random signals, the dispersion of the quantization noise of a linear ADC with the number of orders (levels) m and the dynamic range (scale) L is estimated as follows [24]:

$$\sigma_{qu}^2 \approx \frac{L^2}{3 \cdot 2^{2(m+1)}}.$$

Considering the CF property of the DA operator used in the multi-point "measurement-processing" process under study, we can choose a linear model of the quantizer.
At this point, one more issue should be paid attention to. Estimate the final error of the ADC, it should be noted that the discretization $\Delta x^*(iT_0)$ and the corrective filtering of quantization errors ε_{kvi} require a generalized, rather than an individual, approach.

$$\Delta_{\sum i} = \Delta x^*(iT_0) + \varepsilon_{kvi}, i = 1,2,3, \dots \quad (11)$$

This approach, in turn, requires the search for common effects in overcoming the complexities of the final error.

In this sense, it is appropriate to choose the discretization step T0, and the methods and means of over-discretization widely used in modern information exchange systems [33] attract attention. From the point of view of increasing the metrological efficiency of analog-to-digital and digital-to-analog conversions, the concept of over-discretization should be transferred to the accuracy resource of the acceleration (speed) resource. That is, through a fast rough (with large quantization noise) quantizer (Δ or $\Delta\sum$ modulators), an analog-to-digital conversion is performed at a frequency of excess (above the Kotelnikov frequency f_k) discretization frequency f_0, and the resulting sequence is passed through a digital low-frequency filter and decimated again to the frequency f_k (Fig. 4).

In the block diagram shown in Fig. 4, HF - is a high frequency filter that limits the frequency of the input signal; D - is a discretizer; $\Delta\sum$ - is a delta (Δ)-sigma ($\sum$) modulator (quantizer); LDF - is a low-pass (decimating) digital filter.

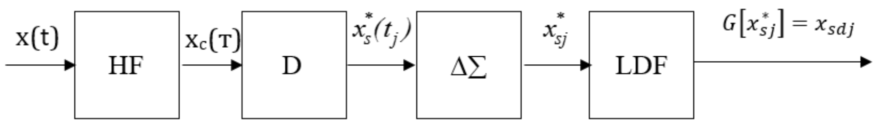

Figure 5: Block diagram of ADC based on over-discretization

The spectrum of the input signal x(t) is limited from above by HF to $0 \le f \le f_0/2$. The frequency-normalized signal xs(t) is discretized (with frequency $f_0=1/T_0$) and quantized. The sequence $\{x_{sj}^*\}$ is digitally filtered by LDF and converted into a sequence $\{x_{sdj}^*\}$ with frequency $f_k=f_0/2V$.

The metrological efficiency of the ADC depends on the excess discretization factor $V=f_0/2f_k$.

Assuming a uniform distribution of the quantization noise in the frequency range $-f_0/2 \le f \le f_0/2$ (Fig. 5), we determine its spectral density $S_{kv}(f) = S_0 = const$ from the following dependence

$$\int_{-f_0/2}^{f_0/2} S_{qu}^2(f)df = \int_{-f_0/2}^{f_0/2} S_0^2 df \,. \tag{12}$$

The left side of expression (12) determines the strength of the quantization noise. On the other hand, despite the noise model corresponding to a centered stationary process, this strength is equal to the variance $h^2/12$. Accepting this conclusion, we write from expression (12)

$$\int_{-f_0/2}^{f_0/2} S_0^2 df = S_0^2 \cdot f_0 = \frac{h^2}{12}. \tag{13}$$

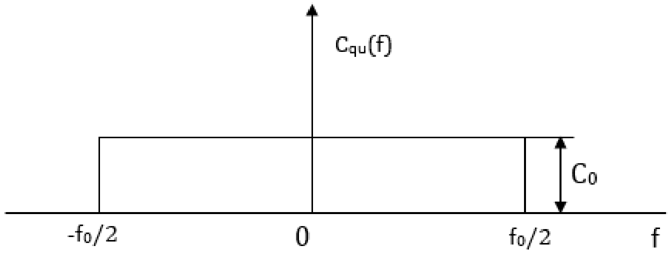

Figure 6: Spectral density of quantization noise

From this expression we get the following formula:

$$S_0 = \left(\frac{h}{\sqrt{12}}\right)\sqrt{\frac{1}{f_0}}. \tag{14}$$

Accept the complex frequency response of the LDF as $G\left(e^{j2\pi fT_0}\right)$, let us express the energy spectrum of the noise of its output sequence (assuming stationarity) as follows:

$$S_{dqu}(2\pi f t_0) = S_0^2\left|G\left(e^{j2\pi fT_0}\right)\right|^2 \tag{15}$$

Considering the LDF as an "ideal" digital low-pass filter (this idealization is possible in error estimation) we will get

$$G\left(e^{j2\pi fT_0}\right) = \begin{cases} 1, & \text{-f}_k \le f \le f_k, \\ 0, & \text{f} < \text{-f}_k \text{ and f} > f_k \end{cases}$$

For the dispersion of the quantization noise after decimation at the output of the ADC, we will get

$$\sigma_{dqu}^2 = \int_{-f_k}^{f_k} \text{S}_0^2 df = \frac{h^2}{12V}. \tag{16}$$

As can be seen, oversampling leads to the effect of correcting the quantization error of the ADC. For example, when V = 2, the power (dispersion) of the quantization noise decreases by 3 dB, which is equivalent to increasing the number of digits of the base quantizer by 0,5 bits.

It should also be noted that the change to over-discretization leads to a reduction in the discretization step, i.e. $T_0 = 1/2Vf_k$. This, in turn, allows us to more quickly reduce the output error estimated by the expression (9) of the DI (see: in Fig. 4, the discretizer module operates at a frequency of f_0).

Thus, using the ADC built based on over-discretization (Fig. 4), we have given the principle of solving the problem of a complex approach to the CF of both discretization and quantization errors.

The errors of a digital functional converter can be divided into two groups:

1) methodological (approximation), instrumental and computational (quantization) errors arising due to the inaccurate implementation of the operator $F[\cdot]$;

2) instrumental error due to the transformation of the final noise $\{\Delta x^*(t_i) + \varepsilon_{qui}\}$ included in the output sequence of the ADC.

At the modern level of development of computer technology, the reduction of errors included in the first group does not cause difficulties both from a methodological and instrumental point of view. However, errors included in the second group attract attention from the point of view of increasing the metrological efficiency of the "measurement-processing" tracts we have studied.

These errors exist even in the case of an exact implementation of the operator $F[\cdot]$ and are associated with the transmitting of noise of input sequence to the output.
Let us expand the analysis by considering expression (9):

$$\Delta y_{1i} = T_0 \Delta^1 F(x_i) \cdot \left[\Delta x^*(iT_0) + \varepsilon_{qui}\right]. \tag{17}$$

This expression shows that two resources are possible for the CF of the considered instrumental error of the digital functional converter:
- reducing the discretization step T_0 (switching to over-discretization);
- taking advantage of the corrective filtering property of the first-order finite difference operator $\Delta^1 F(x_i) = F(x_i) - F(x_{i-1})$.
Both resources are important, interconnected, and a complex approach to them is necessary.
To use the first resource, it is enough to include the digital functional converter in the subtract shown in Figure 4 (between the $\Delta \sum^{\Sigma} mod$ block and the LDF). In this case, the CF effect of over-discretization also applies to the digital functional converter.
If the microprocessor computing device that implements the digital functional conversion cannot operate at the over-discretization frequency f_0 in terms of speedup, then the CF effect should be sought in the first-order finite differences. This effect depends both on the degree of smoothness (differentiation) of the function F(x) and on the corrective property (suppression of low-frequency noise) of the finite difference operator.
It is known from discrete mathematics that finite differences reduce the order of the polynomial F(x). Taking this fact into account, let us apply the first-order finite difference operator to the polynomial:

$$\Delta^1 F(x_i) = \sum_{k=1}^{N} g_k x_i^k - \sum_{k=1}^{N} g_k x_{i-1}^k \qquad i = 0, 1, 2, \ldots \tag{19}$$

Let us express the right side of this expression as follows

$$\Delta^1 F(x_i) = g_1(x_i - x_{i-1}) + \sum_{k=2}^{N} g_k \left(x_i^k - x_{i-1}^k\right) \qquad i = 0, 1, 2, \ldots \tag{20}$$

From the last expression it can be seen that the CF of the instrumental error of a digital functional converter depends on the smoothness of its input signal. As a special case, if x(t) is a constant signal (described by a zero-order polynomial P_0(t), then $\Delta^1 P_0(t) \equiv 0$ is obtained.
Taking account, the metrological model shown in Fig. 2.1, let us consider the real input sequence $\{x_i^{**}\}$ of the operator $F[\cdot]$ on the right-hand side of the expression (20). First, let us express the sequence $\{x_i^{**}\}$ as follows:

$$x_i^{**} = x_i + \frac{1}{2}\Delta^1(x_i + \Delta x_i) + \varepsilon_{qui} \tag{21}$$

Let us write this expression considering on the right side of expression (20)

$$\Delta^1 F(x_i^{**}) = g_1 \Delta^1 x_i^* + \sum_{k=2}^{N} g_k [(x_i^*)^k - (x_{i-1}^*)^k]. \tag{22}$$

where $\Delta^1 x_i^* = \Delta^1 x_i + \frac{1}{2}\Delta^2(x_i + \Delta x_i) + \Delta^1 \varepsilon_{qui}$; $\Delta^2(x_i + \Delta x_i) = x_i + \Delta x_i - 2(x_{i-1} + \Delta x_{i-1}) + x_{i-2} + \Delta x_{i-2}$ – is the second order finite differences.
Let us write the last expression in the following form:

$$\Delta^1 F(x_i^{**}) = g_1 \left[\Delta^1 x_i + \frac{1}{2}\Delta^2 x_i + \frac{1}{2}\Delta^2(\Delta x_i) + \Delta^1 \varepsilon_{qui}\right] + \sum_{k=2}^{N} g_k [(x_i^*)^k - (x_{i-1}^*)^k]. \tag{23}$$

If we take into account that the coefficients of the polynomial (20) decrease as the order k increases (for example, when using the Taylor series, $g_k = F^{(k)}(0)/k!$, then expression (23) can be reduced to the following form

$$\Delta^1 F(x_i^{**}) \approx g_1 \left(\Delta^1 x_i + \frac{1}{2}\Delta^2 x_i + \frac{1}{2}\Delta^2(\Delta x_i) + \Delta^1 \varepsilon_{qui}\right) \quad (24)$$

As can be seen, the digital functional converter has the CF property within the framework of the finite differences (first, second order) of the useful signal (x) and noise (Δx, εqui) involved in the exchange of measurement information in the previous sub-tract of the "measurement-processing" tract.

The above-mentioned CF resources are of great importance in terms of increasing the metrological efficiency of researched IMS.

Conclusion

The systematic approach to increasing the accuracy of information-measuring systems was proposed from the perspective of the concept of corrective filtering. Considering that the information-measuring systems under study are dynamic systems and that the important attributes of such systems are ADC, the modern requirements imposed on these converters were systematized. Based on the analysis of existing works, the possibilities of reducing methodological error were considered, and the importance of cleaning the sub-integral function from noise, that is, reducing instrumental errors, was substantiated.

To solve the problem of converting and digital processing of non-sinusoidal signals based on a systematic approach, the errors arising in the conversion and digital processing tracts of non-sinusoidal signals were analyzed, and a methodology for their evaluation was developed.

Based on the unified structure of the information exchange tracts of information-measuring systems, a generalized metrological model of the "measurement-processing" tract was developed, and the possibilities of corrective filtering in measurement-processing processes were determined. In the "Measurement-processing" tract, the errors introduced into the output of the corresponding devices, that is, the errors created by the initial information transmitter, quantizer, discretizer, digital converter, integrator during the evaluation of integral parameters, were determined. The corrective properties of the discrete averaging operator during digital measurements of the integral parameters of non-sinusoidal signals were studied. The fact that this operator has a corrective character with respect to the systematic and random errors of the instantaneous values of alternating current signals was confirmed by statistical modeling.

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