Algorithmic methods of increasing the accuracy of analog blocks of measuring systems

O A Jumaev, M T Ismoilov, G B Mahmudov and M F Shermurodova

Department of Automation and Control, Faculty of Automation, Navoi State Mining Institute, Navoi, Uzbekistan

E-mail: jumaev5216@mail.ru

Abstract. The main requirements for an intelligent measuring channel are considered in the article when designing automated control and control systems. The methods for estimating errors are classified and the constituent groups are classified, and some methods for calculating the errors of measuring means and measuring modules of automatic control systems are given.

1. Introduction

Metrological characteristics of information-measuring systems (IMS) are determined from the combination of characteristics of the elements of the measuring channel. IMS tools with various functionalities and hardware implementations are combined with an identical structure, which differs only in the specific performance of individual devices. The main functional units included in the IMS are: measuring systems (amplifier, MS, ADC), computing part (microprocessor, RAM, ROM), auxiliary part (power supply, input-output devices).

2. Method

Development trends of devices for an analog input of information determine the following structure of the measuring channel (figure 1). A physical quantity (PQ) is converted by a primary measuring transducer (PMT) into current or voltage, then this signal is amplified by a unifying transducer (UT), to transmit this signal to the ADC inputs. Analog filters (AF) are usually installed before the ADC. After the ADC, information signals in digital form are processed by a microprocessor (MP). The results of the measured parameter are transmitted via the interface channel to the control and control systems (CS) [1].

![Figure 1. Block diagram of one measuring channel](image)

The generalized structural diagram of the measuring channel software and hardware corresponding to the structural diagram according to the mathematical model of analog-to-digital conversion operations is shown in figure 2, where X (t) is the input action, Y (t) is the true value of the output signal, ωₐ(t) is the weight characteristic of the analog components of the channel (ω(t) is the equivalent weight characteristic of the digital elements of the channel (digital pre-amplifiers)), ωᵣ(t) is the weight...
characteristic of the block of calculations and processing (block of intermediate results), $E_q^*(n)$ is the quantization error by level in the analog-to-digital converter[2,3].

![Diagram](image.png)

**Figure 2.** The structure of the firmware of the measuring channel

Characteristics of the reaction of the analog elements $Y_a(t)$ to the input signal $X(t)$:

$$Y_a(t) = \int_{-\infty}^{\infty} \omega_a(t)X(t-\mu)d\mu,$$

where the equivalent weight characteristic of the analog components of the channel and the signal conversion at the ADC output will be:

$$Y_c^*(n) = X_c(nT_s) + E_c^*(n),$$

here: $X_c(t)$ is the conversion signal at the input of the ADC; $E_c^*(n)$ - level quantization error.

The signal at the output of the digital part of the system is defined as

$$Y_d^*(n) = \sum_{l=0}^{\infty} \omega_d^*(l)X_d(n-l).$$

The converted signal $Y_r(t)$ at the IR output is

$$Y_r(t) = \sum_{n=\infty}^{E[t/T_s]} X_r^*(n)\omega_r(t-nT_s).$$

where: $X_r^*(n)$ is the converted discrete parameter at the input of the signal processing device; $\omega_r(t)$ is the weight characteristic of the signal processing device; $E[\cdot]$ is the integer part of the number. Applying the replacement of variables of the form

$t-nT_s = kT_s + Ts$, $\text{где } k = E[t/T_s]-n$, $a \xi = t/Ts - E[t/Ts],\text{ expression (2.16) can be transformed to the form:}$

$$Y_r(t) = \sum_{k=0}^{\infty} \omega_r[(k+\xi)T_s]X_r^*(E\left[\frac{t}{T_s}\right]-k).$$

The variable $\xi$ in expression (4) can have any numerical values in the range from 0 to 1 (0≤ $\xi$ < 1). We obtain the expression for the adjustable parameter $X(t)$ of the firmware MC $X(t)$ $Y_r(t) = (t)$, $X_r^*(n) = Y_d^*(n)$.

$$X[m,\xi] = \sum_{k=0}^{\infty} \omega_r[k,\xi]\sum_{l=0}^{\infty} \omega_d^*(l)X_d^*(m-k-l).$$

Bearing in mind that $X_d^*(n) = Y_c^*$

$$X[m,\xi] = \sum_{n=0}^{\infty} \omega_w[n,\xi]\left(X_c[(m-n)T_c] + E_q^*(m-n)\right).$$
In turn, \( X_c[nT_s] = Y_d[nT_s] \), so we obtain the final expression of the output parameter of the MC software and hardware

\[
X[m, \xi] = \sum_{n=0}^{\infty} \omega_w [n, \xi] \times \left( E_d^*(m - n) + \int_{0}^{\infty} \omega_a(\mu)X[(m - n)T_s - \mu]d\mu \right).
\]

Given that the mathematical expectation of quantization error by level is 0, we get:

\[
m_x(t) = \sum_{n=0}^{\infty} \omega_w [n, \xi] \int_{0}^{\infty} \omega_a(\mu)m_x[(m - n)T_s - \mu]d\mu,
\]

(6)

Here: \( m_x(t) \) is the mathematical expectation of the input action \( X(t) \) of the hardware and software of the measuring path, which, according to the accepted mathematical model of measuring information, generally depends on time.

We obtain an expression for the correlation function of the output parameter \( X(t) \), which by definition is equal to:

\[
K_x(t, t + \tau) = M\{X^*(t)X(t + \tau)\} = M\{X^*[k, \xi]X^*[k + m, \Psi]\},
\]

where \( X^*(t) \) is the corrected random signal.

3. Result and discussion

Thus, one of the most significant factors limiting the accuracy of a measuring device (MD) at a low level of the output signal of the conversion circuit is the effect of the intrinsic noise of the primary converter (PC), which is related to thermal noise and is due to the presence of a signal equivalent to internal resistance \( R_3 \) as a source of signal (resistance of the mass of liquid between the electrodes). This noise arises in any circuit having an equivalent active resistance, and its value depends only on the resistance value and does not depend on the physical nature of the noise source \([4]\). The spectral density of thermal noise is frequency independent. The noise of the operational amplifier is characterized by noise voltages and currents reduced to the input and having a power spectral density of \( S_u(\omega) = S_i(\omega)R_a \), respectively. The total power spectral density of the equivalent total noise over the voltage reduced to the input of the measuring device will be \([5]\):

\[
s(\omega) = S_u(\omega) + S_i(\omega)R_a + 4kTRE.
\]

where \( k \) is the Boltzmann constant; \( T \) is the absolute temperature; \( \omega \) is the angular frequency.

Given the indicated ratio, the first two members of its right-hand side are the noise of the input amplifier (MU), the third is the noise of the PC.

In view of the dependence of the spectral density on the noise frequency, the PP is white noise, the noise of the input amplifier has two components - white noise and flicker noise, while the dependence of both \( S_u \) and \( S_f \) on the frequency, \( \omega \) has the form

\[
S_u = S_{u^\delta}(1 + \frac{\omega_0}{\omega});
\]

\[
S_f = S_{f^\delta}(1 + \frac{\omega_0}{\omega}),
\]

where \( S_{u^\delta}, S_{f^\delta} \) characterize the amount of white noise; \( \omega_0 \) is the frequency below which flicker noise prevails \([6]\).

The frequency \( \omega_0 \) for current noise is usually higher than for voltage noise. Passing through the MD conversion circuits, the noise spectrum will change in accordance with the frequency response of the conversion circuit, which is determined by the conversion method and parameters of the timing diagram.

Thus, for the resulting noise value \( U_{N_{out}}(t) \) at time \( t_6 \) we have
where $U_{N_{\text{out}}}(t)$ is the instantaneous voltage of noise brought to the input of the MD; $\tau$ – integration time constant; $k_0$ – gain of the conversion path from the input of the MD to the input of the integrator.

In expression (7), the time is counted from the beginning of the period. In the form corresponding to any $n$-th period, it can be written, given that in this case, the time $t_6$ will correspond to the time instant $nT$ of the end of the $n$-th period:

$$U_{N_{\text{out}}}(nt) = \frac{k_0}{\tau} \left\{ \int_{t_2}^{T/2} U_{N_{\text{in}}}(t)dt - \int_{t_2-T/2}^{T-\tau-t_i} U_{N_{\text{in}}}(t)dt \right\}$$

where $T$ is the period of the power supply current; $t_i$ is the integration time, and $t_i = t_3 - t_2 = t_6 - t_5$

expression (8) corresponds to samples at time instants $n$-th of a continuous function of the form

$$U_{N_{\text{out}}}(t) = \frac{k_0}{\tau} \left[ \int_{t}^{T/2} U_{N_{\text{in}}}(t)dt - \int_{T/2-t_i}^{T-t_i} U_{N_{\text{in}}}(t)dt \right]$$

Expressions (7) and (8) give the same voltage values of the noise passing through the conversion path at the end of each period. Having found the Laplace image of function (8) in accordance with the rules of operational calculus, and also using the theorem of operational calculus of delay in the domain of originals and the theorem of splitting a certain integral, we can obtain the frequency response of the transformation path for noise $G(j\omega)$

$$G(j\omega) = \frac{k_0}{\tau} \frac{1}{j\omega} \left[ e^{-j\omega T/2} - e^{-j\omega (T/2+t_i)} - 1 + e^{-j\omega t_i} \right] = \frac{k_0}{\pi} \frac{1}{j\omega} \left[ 1 - e^{-j\omega t_i} \right]$$

Frequency response module $|G(j\omega)|$ determined to take into account (13)

$$|G(j\omega)| = 2k_0/\omega \sqrt{(1 - cos\omega t_i) \left( 1 - cos\omega \frac{T}{2} \right)}$$

To obtain the frequency response in a form independent of specific times $t_i$ and $T$, we introduce the reduced value of the frequency $\Omega$

$$\Omega = \omega t_i$$

 wherein $\omega = \frac{T}{2} = m\Omega$,

where $m = \frac{T}{2t_i}$ – is a coefficient depending on the ratio of the integration time and the power supply period of the MD.

The expression for the frequency response module related to the static transmission coefficient equal to $\frac{2k_0T_i}{T}$ will take the form

$$|G(\Omega)| = \sqrt{(1 - cos\Omega)(1 - cosm\Omega)/\Omega}$$
In figure 3 shows the curves of the module of the frequency response of the conversion path for noise at three values of T. According to the data of the curves, one can judge the frequency region in which the noise with the greatest weight passes to the output of the conversion path.

As can be seen in figure 3, the noise component with zero frequency does not pass through the conversion path, in contrast to the constant component of the useful signal. This should significantly reduce the passage of flicker noise, whose density at the zero frequency increases unlimitedly, which is confirmed experimentally with a decrease in the frequency \( \omega \) up to \( 10^6 \) Hz. [5]. At the same time, the presence of a maximum frequency response in the low-frequency region \( (\Omega = \frac{\omega}{2}) \) is unfavorable from the point of view of flicker noise transmission.

Dispersion \( D_{out} \) (the average square of the effective value of the noise voltage at the output of the conversion path) [7] is determined by the expression

\[
D_{out} = \int_0^\infty |G(j\omega)|^2 S(\omega) d\omega. \tag{13}
\]

The square root of the variance is the rms current value of the noise voltage. Dividing this value by the static transmission coefficient of the useful signal, we obtain the noise voltage reduced to the input and representing the mean square absolute error of measuring the input signal of the MD.

Due to the fact that the components of the noise voltage (white noise and flicker noise of the input amplifier, PC noise, which also applies to white noise, voltage noise and current noise of the input amplifier) have different weights and can be reduced by various means, it is advisable to find the equation for each individually.

For white noise with density \( S_w \), taking into account the frequency response (11), (12), relation (13) takes the form

\[
D_w = 4k_0^2 \tau S_w / \tau^2 \int_0^\infty \frac{(1-\cos\Omega)(1-\cos\Omega)(d\Omega)}{\Omega^2}; \tag{14}
\]

for flicker noise density \( S_f \omega_0 / \omega \)

\[
D_f = 4k_0^2 \tau S_f / \tau^2 \int_0^{\infty} (1 - \cos\Omega)(1 - \cos m\Omega)(d\Omega)/\Omega^3 \tag{15}
\]
4. Conclusion
The integral on the right-hand side of expression (14) calculated analytically is $\pi/2$, that is, independent of $T$. The integral on the right-hand side of expression (15) is a function $F(T)$ of the number $T$ and is analytically not solvable.

In conclusion, it can be noted that the results obtained allow us to resolve the contradiction between accuracy and power consumption by optimally choosing the method and parameters of analog-to-digital conversion and the type of operational amplifier.

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