REMOTE ERRORS CORRECTION OF MULTI-CHANNEL CYBER-PHYSICAL MEASURING SYSTEMS

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Abstract: **The method of remote auto calibration measurement channels cyber-physical systems in real time using code-controlled measures is described. Additive error components as measurement as well the reference channels is adjusted by switching invert method. It is shown that the resulting measurement error virtually determined by code-controlled measures error that could it serve as a basis for implementing remote auto calibration and operative check of metrological characteristics of measuring channels cyber-physical systems. The results of the experimental study model code-controlled measures showed good coincidence with the theoretical model.**

Key words: **Code-controlled measures, cyber-physical systems, measurement channels, operative check of metrological characteristics, remote auto calibration, switching invert method**

I. INTRODUCTION

The ultimate purpose of measurement processes operative checking in cyber-physical systems (CPS) is to improve the metrological reliability. This procedure implementation results from two of its scientific and technical aspects. The first is quantitative characteristics – reliability monitoring, and the second – holding frequency. The checking reliability reflects the degree of objectivity of test results and is expressed in probability terms of making the right decisions $D=1-(\alpha-\beta)$, where α is the probability of wrong decision "out of tolerance"; β is the probability of wrong decision "within tolerance". Measuring instruments (MU) errors Δ_x are continuous random variables with density distribution $p(\Delta_x)$. The same procedure of control is that the controlled setting of the MU Δ_x can be in one of two states: H1 is the error Δ_x MU is the "admission" $\Delta_{xmin} \leq \Delta_x \leq \Delta_{xmax}$ and H2 is the error Δ_x MU is "out of tolerance" $\Delta_x \geq \Delta_{xmin}$, $\Delta_x \geq \Delta_{xmax}$, where Δ_{xmin}, Δ_{xmax} is the minimum and maximum error values respectively [1]. Error Δ_{ck} checking MU is also a continuous random variable with density distribution $p(\Delta_{ck})$. Since the general density distribution errors MU and specific checking devices are generally unknown, it is advisable to adopt equally laws for their distribution. It is known that under symmetry tolerance field error $\pm d$ and by coincidence expectation $M(\Delta_x)$ of the tolerance field center, only if the ratio $d \geq \Delta_{xal}$, $\Delta_{ck} \leq d-\Delta_{xal}$ are the

errors MU checking is the perfect reliability and $D=1$, where Δ_{xal} – the allowed error value for a while MU checking [1]. In most cases the allowed error value MU is symmetrical, $M(\Delta_x)=0$ and for a traditional metrology ratio $\Delta_{ck} \leq \Delta_{xal}/5$, $d=\Delta_{xal}$, we received $\Delta_{ck} \leq 0$, $8\Delta_{xal}$ error value of specific checking device. To control the error value over the entire range of measurement MU, it is advisable to choose the same checked point as in the metrological verification MU.

Therefore, to ensure high reliability of monitoring $D = 1$ we should in practice ensure symmetrical boundaries of allowed error value MU in actual use of mathematical expectation $M(\Delta_x) = 0$. Equality is zero expectation errors MU in actual use can be achieved only if the correction of systematic error component.

II. ANALYTICAL REVIEW OF AVAILABLE TECHNIQUES TO IMPROVE THE METROLOGICAL RELIABILITY OF MEASURING CHANNELS CPS

The most effective way to improve the CPS quality operation is to increase the calibration frequency. It is often realized by verification the checking MU through more accurate one or calibrator applying to identify and exclusion of any bias (errors). The analysis showed that there are two ways to increase metrological reliability [2]. The first one is embed a reference element connected in each measurement cycle. The second method is performed periodically to calibrate by the external standard.

One of the ways to improve the measurements quality now is to establish the optimal interchecking or intercalibration intervals [3]. Moreover, the conventional approach based on the principle of interverification is interval appointment for all MUs of the same type by the duration optimizing of these intervals. It is to minimize the total costs of MU maintenance (verification, inspection and downtime MU and unreliable measurements results) [4, 5]. However, it is difficult to find such an economic relationship for most MU types. Because of the diverse requirements the same MU type can be used for different purposes, so the consequences of inaccurate measurement results can differ dramatically. For those parameters settlement is practically not required information about temporal

instability components MU, on the conditions and the intensity of MU using, on the economic consequences of inaccurate measurement results etc. That leads to inappropriate practical application of these complex and labor-intensive methods because of the derived results of low probability [3]. Often the intercalibration MU intervals declared by manufacturers in the competitive environment look incredibly long.

To predict the MU metrological operability in working phase it is appropriate to use the metrological checks protocols. This received information varies for different ways of checking. Metrological stability MU can be indirectly checked only if to check the stability standards and followed suitable calibration MU. However, this method is practically used only to verify the working standards. So, in the CPS the best way by industrial measurements quality improving is to develop methods and algorithms for metrological characteristics operational control of measuring channels. The operational MU CPS control implementation will significantly reduce the share of metrological faulty MU, operating in industrial processes and significantly increase the reliability of measurements in the CPS.

III. ANALYTICAL REVIEW OF EXISTING METHODS FOR THE METROLOGICAL RELIABILITY IMPROVING OF MEASURING CHANNELS CPS

Analysis reveals that for on-situ operational checking of industrial MU errors preference should be given to following methods: reconciliation homogeneous MU using common source; reconciliation circular in homogeneous groups MU; mutual verification by stable MU; reconciliation MU using industrial calibrator [6]. Obviously, the calibrator implement method can be used for operational control of measuring channels in CPS because of their dispersion.

Choose method for the additive component error correction.

Industrial measurements are accompanied by the noise and usually significant influence of normal and common interference. Shielding, galvanic isolation of the measuring signal sources and secondary MU parts (the measurement information devices for display, storing, transmission, storage, processing, etc.) and measuring circuits' equipotential protection are generally used to reduce the common mode interference effect. The averaging conversion method is the most accepted for noise mitigation. The digital integration of partial transformation results is presented in an attractive way by using a given conversion cycle ADC or delta-sigma ADC (DSADC) due to the modern microelectronic hardware simplicity and small cost [7]. The ADC conversion function nonlinearity is several tenth of ppm and it is negligible versus allowed error value. To correct the multiplicative and additive components the auto-calibration method is used which based on the periodically disconnect measured signals from ADC

input during calibration. This causes reduction of the ADC conversion time and measurement data losing at auto calibration duration. To reduce the ADC errors in working conditions the auto-calibration should be performed as often as possible that will significantly decrease measurement period. Since the maximum error value of transfer coefficient ADC after auto-calibration does not exceed ±0, 5 units of least significant digit, and their maximum value of the additive error component (AEC) is no more than ±10 μV [7], it becomes an obvious need to reduce the latter, especially when converting low signal levels. For example, this AEC value can be transform into near ±1 K for the thermoelectric thermometer applying.

The crucial advantage of digital averaging in DSADC is a possibility of simple changes in the frequency pole values and implementation of large values (upper 80 dB) factor weakening in the vicinity dominant frequency noise (typically the mains power frequency) and even the relatively wide frequency band of influence [7]. The inverting input method is used to correct the AEC for the negligible level in DSADC. But it doubles the conversion time and ADC output code is updated near a few hertz [7]. Thus the unadjusted value of AEC is several microvolts for whole work operating temperature changes. In our opinion, this is due to the infra low noise influence of ADC elements. AEC unadjusted value will be determined by noise parameters of used components equivalent and the frequency value of the establishing zero procedure.

Dispersion D noise of output voltage electronic devices is in the ratio [8]:

$$D = \int_0^\infty S(w) |G_t(w)|^2 \, dw, \qquad (1)$$

where $S(\omega)$ is the noise input spectral density; $G_t(\omega)$ is the device current frequency response.

The noise spectral density $S(\omega)$ at the electronic devices input is approximated by amount densities of white noise and hyperbolic noise (flicker noise) [8] $S(w) = S_0 \left(1 + \dfrac{w_0}{w}\right)$, where S_0 is the white noise spectral density (constant versus frequency); ω_0 is the frequency coupling of white and flicker noise; ω is the current noise frequency value. The output voltage $y_\omega(t)$ of current value of the electric device is based on the Duhamel integral (convolution) provided that at the input a complex harmonic oscillation $e^{j\omega t}$ is given. We suppose that the time origin is selected in the end of auto-calibration. Then the current value of a square module current frequency characteristic delta-sigma ADC (DSADC) is recognized as:

$$|G_t(w)|^2 = y_{win}(t) \, y_{-win}(t) = 2k^2 \left(\frac{w_i}{w}\right)^2 (1 - \cos wt), \quad (2)$$

where $g_{in}(t\text{-}\tau)=k/\tau_i=k\omega_i$ is the device pulse weight function for i time interval; k is the transfer coefficient of the non-inertial scale input device; $\tau_i=R_iC_i$ is the

integrator time constant during integration input signal; $\omega_i = 1/\tau_i$.

Taking into account ratio (2), variance D_{nn} and D_{fn} of output noise voltage DSADC, caused by white and flicker noise of both input scale device and integrator amplifier at the time t_0 after the next errors auto-correction are find on inappropriate integrals tables:

$$D_{nn} = 2\left(k^2 S_{0in} + S_{0it}\right)w_i^2 \frac{1-\cos wt}{w^2} dw =$$
$$= 2p\left(k^2 S_{0in} + S_{0it}\right)w_i^2 t \qquad (3)$$

$$D_{fn} = 2\int_0^\infty \left(k^2 S_{0in}w_{0in} + S_{0it}w_{0it}\right)w_i^2 \frac{1-\cos wt}{w^3} dw =$$
$$= 4(k^2 S_{0in}w_{\backslash 0in} + S_{0it}w_{0it})w_i^2 t^2 \int_0^\infty \frac{\sin^2 \Omega}{w^3} dw \qquad (4)$$

where S_{0in}, S_{0it} is the white noise spectral density of input scale device and integrator amplifier; $\Omega = 0{,}5wt$.

Integral (4) is not confined to the tabular one and for it value analytical defined is necessary the replacement at the integration lower limits α and upper on β, carry out integration by parts and made on to the limit conditions $a \to 0$, $b \to \infty$, then given

$$D_{fn} = 4\left(k^2 S_{0in}w_{0in} + S_{0it}w_{0it}\right) \int_{a\to 0}^{b\to\infty} \frac{\sin^2 \Omega}{w^3} dw =$$
$$= 4\left(k^2 S_{0in}w_{0in} + S_{0it}w_{0it}\right)_i^2 [3 + Ci(2\Omega)],$$

where W_{0in}, W_{0it} is the frequency coupling of white and flicker noises of input scale device and integrator amplifier respectively; t is the time interval after auto calibration DSADC; $Ci(...)$ is the cosine integral. The integral cosine asymptotic behavior is fed formulas [8] for $2\Omega \rangle\rangle 1$, $(2\Omega \to \infty)$: $\lim\limits_{\Omega\to\infty} Ci(2\Omega) = \lim\limits_{\Omega\to\infty}(\sin 2\Omega/2\Omega) = 0$ and for $2\Omega \to 0$ $\lim\limits_{2\Omega\to 0} Ci(2\Omega) = = \lim\limits_{2\Omega\to 0} \ln 2\Omega g = \infty$. This uncertainty to resolve into account that the threshold argument cosine integral mean auto calibration frequency DSADC $w_k = 2p/t_k$, that principally can not be zero. Cosine integral in $w = w_k = 2p/t_k$, $t = 0$ ratio may be submitted [8] $Ci(w_k t) = \ln\left|1 - \frac{t}{t_k}\right|$, where t_k is the inter calibration interval.

In addition to these sources, noise ADC zero-comparator (ZC) lead to instability threshold triggering it. ZC pulse weight function for noise signal $g_{ZC}(t)=1$. By analogy on the top made analysis it defined the dispersions D_{nzc}, D_{fzc} output voltage DSADC, caused by white and flicker noise served as

$$D_{nzc} = 4S_{ozc} \int_0^\infty \frac{\sin^2 \Omega}{w^2} dw = \frac{p}{2} S_{ozc} t,$$

$$D_{fzc} = S_{ozc}w_{ozc}t^2\left(3 - \ln\left|1 - \frac{t}{t_k}\right|\right),$$

where S_{ozc}, ω_{ozc} is the spectral density of white noise and frequency coupling of white noise and noise with hyperbolic spectral density of ZC.

Finally, the output signal DSADC variance D_k caused the main elements noise is determined by the expression:

$$D_k = \left[W_i + 0{,}5pS_{ozc}t + S_{ozc}w_{ozc}t^2\left[3 - \ln\left|1 - \frac{t}{t_k}\right|\right]\right] \qquad (5)$$

where

$$W_i = 2w_i^2\left[p\left(k^2 S_{0in} + S_{0it}\right) + \left(k^2 S_{0in}w_{0in} + S_{0it}w_{0it}\right)\right].$$

The inter calibration interval maximum duration t_k can be estimated for the set and reduced into input dispersion value D_{al} (or standard deviation σ_{al}) conversion results DSADC and measured voltage $U_x=0$ for the following conditions: $D_k = At_k^2 + Bt_k \leq D_{al}$, where

$$A = 2\left[S_{0it}w_{0in} + S_{0it}w_{0it}/k^2\right]w_i^2[3 - \ln 2pg]\rangle 0;$$
$$B = \left(S_{0in} + S_{0it}/k^2\right)w_i^2\rangle 0.$$

The physical realization condition (the discriminant inherent value $d = B^2 + 4AD_{al}\rangle 0$) is performed if the positive coefficient A and true this inequality $t_k \leq \frac{B + \sqrt{d}}{2A}$. That will determine the maximum possible time between auto-calibration DSADC for a given maximum input voltage offset.

IV. METROLOGICAL RELIABILITY INCREASING FOR VOLTAGE MEASURING DEVICES AT WORKING CONDITIONS CPS

In general, the error MU is described by a polynomial model [8]. For the most of industrial MU it can be limited to only three components separately and considers temporal drift and changes its constituent's values when changing the working conditions:

$$\Delta_x = \Delta_{x0} + \Delta_{xt} + \Delta_{xp} = \Delta_0 + d_S X + eX^2 +$$
$$+ \Delta_{0t} + d_{St}X + e_t X^2 + \Delta_{0p} + d_{Sp}X + e_p X^2 \qquad (6)$$

where Δ_x is the absolute MU error just t moment in working conditions, vector environmental parameters which change as a time function $\overline{p_t} = F(p,t)$; $\Delta_{x0} = \Delta_0 + d_S X + eX^2$ is the MU absolute error at the initial time moment t_0 (after metrological verification or calibration) in normal operating conditions; $\Delta_{xt} = \Delta_{0t} + d_{St}X + e_t X^2$ is the absolute error change for time interval $t-t_0$, which passed after the MU metrological verification or calibration provided in normal operating conditions; $\Delta_{xp} = \Delta_{0p} + d_{Sp}X + e_p X^2$ is the absolute error change for time interval $t-t_0$, which passed after the MU metrological verification or calibration provided in working conditions, which the

environment parameter vector varied as a time function $\overline{p_t} = F(p,t)$; $\Delta_0, \Delta_{0t}, \Delta_{0p}$ is the MU AEC of the initial time moment t_0 (after metrological verification or calibration) in normal operation and its possible changes at time t and working conditions, the environment parameter vector which was $\overline{p_t} = F(p,t)$ accordingly; d_S, e is the multiplier and quadratic coefficients of MU error polynomial model in the initial time t_0 (after metrological verification or calibration) in normal operating conditions respectively; d_{St}, e_t, d_{Sp}, e_p is the multiplicative and quadratic coefficients of MU error polynomial model at time t in working conditions, which the environment parameter vector was $\overline{p_t} = F(p,t)$ accordingly; X is the measuring quantity mean.

The ratio (6) analysis shows that it may correct errors temporal changes of this type MU based on the accumulation and processing (approximation) the experimental data arrays, similar to existing methods [2-5]. We should take into account the impact factors variety and their relatively large ranges in working conditions and the no predictive information on the possible nature of their changes. Thus, we can conclude practical impossibility of designing common methods for correcting temporal changes errors MU in the working conditions at the modern measurement technology. However, the up-to-day microelectronics and information technology fundamentally changes the approach to the MU measuring loop structure where, in general, any physical quantities are converted into electrical signals. First of all, it is possible due to the MU intellectualization and automatic adjustments the measuring scheme errors using known methods: exemplary signals, iterative, additive and multiplicative correction and auxiliary measurements [8]. Any improving accuracy of MU methods implementation is limited by the electronic switches residual parameters which reconfigurate input circuit MU. These electronic keys residual parameters determine the unadjusted error values of measurement results. AEC MU typically has the greatest value and the most intensive changes both in time and when environment conditions are changed. It is theoretically shown that by switching inverting method the unadjusted voltmeter AEC value can be reduced to negligible level and will not exceed (for produced MOS polarity switch chip) $\Delta_{0ki} = 2R_x \Delta I_{bk} + (4I_{bk} + I_{bx})\Delta R_k$ [8], where Δ_{0ki} is the unadjusted AEC; Rx, ΔR_k is the signal source impedance and maximum resistances match of switch polarity keys accordingly; $I_{bk}, \Delta I_{bk}$ is the maximum leakage current and their match of switch polarity keys (at the most unfavorable impact factors values in the working conditions) respectively; I_{bx} is the maximum MU input current.

Long industrial experience in operation (since 1983 up today) of A565 and CR7701 digital device types (instrument-making factory "Mukachivprylad") showed that the AEC unadjusted value in working conditions will not exceed a few tenths of µV. It allowed to refuse from of the zero manual installation for these devices [8]. Indeed, the digital MU measuring time is much smaller compared to the maximum temporal changes of their error impact factors and as fast as possible error values drift time. This shows the effectiveness of AEC adjustment by inverting switching method in real-time and MU working conditions. It is obvious that in the ratio (6) we must understand that AEC value is equal zero. In addition, even the degree of nonlinearity of MU general transformation function is adjusted and non-linear error component is significantly reduced (6). Thus, the industrial MU time and temperature stability are mainly determined by only the scale elements stability if inverting and switching methods are implemented. For the consumer's stable scale elements the CR7701 type devices are guaranteed without calibrations time up to 5000 hours or nearly two years of their one shift work, while the metrological checks frequency is one year.

The digital MU metrological reliability can be increased in such traditional ways as by using: the most stable components; auxiliary measurement method or constructive-technological methods of the elimination of impact environmental factors (screening, thermo isolation, sealing, etc.). All these ways of metrological reliability of MU improving is evolutionary and their implementation is limited by modern technical, technological and economic factors. New opportunities are revealed in case of usage of only one precise and stable scale block MU, especially made in the integrated technology. As you know, the metrological characteristics MU which are built on the compensation and differential measurement method basis, are mainly determined by the accuracy and stability of the code-control voltage measures that have long massively made as DAC. To build highly stable digital voltmeters preference should be given to differential measurement method. For errors correction both the measuring and reference channels we proposed switching and inverting voltage method (see Fig. 1). The essence of the highly differential measurement method is to find the code measurement result as algebraic sum codes precise and stable code-control measure and weighted measurement code of the measured difference and reference voltages [1, 8]. This signal summer error of both measuring channels significantly affects the resulting error. Unlike existing schemes we proposed to carry out the summarized these signals in a sequential circle. Code measurement result N_{xi-1} is in each conversion cycle as the algebraic sum:

$$N_{xi} = N_{xi-1} + N_{xi-1} k_{1H}/k_{2H}, \qquad (7)$$

where N_i is the measurement result code at i conversion cycle; N_{i-1} is the code of measuring and reference voltages at i-1 conversion cycle; k_{1H}, k_{2H} is the nominal values of transformation coefficient PA at first and the second conversion cycles accordingly.
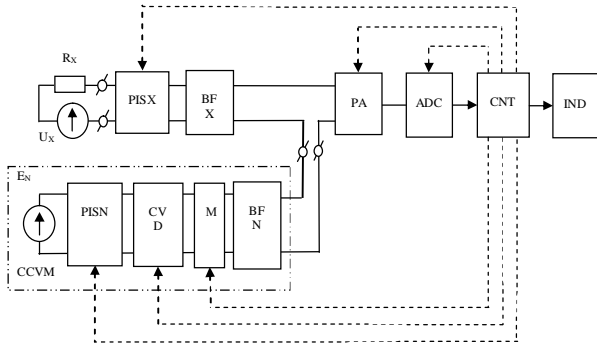
Fig. 1. Digital voltmeter scheme with remote auto calibration. PISX, PISN – polarity inverse switch of measuring and reference voltages accordingly; M – scale transducers; CVD – code control divider; BFN, BFX – buffers of reference and measuring voltage accordingly; PA – programmable amplifier; ADC – analog to digital converter; CNT – voltmeter controller; IND – indication block

In the first cycle by the inverting switching method it give code N_1 approximate value of the measured voltage U_X, while to CVD given code value N_1 previous measurement result:

$$N_1 = N_{11} - N_{12} = 2k_{ADC}k_1 U_X =$$
$$= 2k_{1H}k_{ADCH}U_X(1 + d_1 + d_{ADC}) \quad (8)$$

where $N_{11} = k_1 k_{ADC}[(U_X + \Delta_X) + \Delta_{M2}]$,

$N_{12} = k_1 k_{ADC}[(-U_X + \Delta_X)k_1 + \Delta_{M2}]$ are the conversion result codes at direct and opposite connect polarity PISX and PISN; $k_1 = k_{1H}(1 + d_1)$, $k_{ADC} = k_{ADCH}(1 + d_{ADC})$ is the transformation coefficient PA and ADC respectively (subscript "H" designation of their nominal value); d_1, d_{ADC} are the relative errors of transformation coefficient accordingly k_{1PA}, k_{ADC}; Δ_X, Δ_{M2} are the equivalent AECs of channels respectively measuring U_x and reference E_N voltages at zero control code.

In the second conversion cycle code N_1 is fed to CVD, which is converted into a voltage $U_N = (m_H E_{NH} + \Delta_{M1})(1 + d_C)\dfrac{N_1}{N_{mk}}$, where m is the scale transformation coefficient of reference voltage CVD; Δ_{M1} is the AEC CVD, reduced to it input; $d_C = d_E + d_{CV} + d_m$ is the relative error CVD; d_E, d_{CV}, d_m are the relative errors of reference voltage source, CVD and scale m coefficient; $N_{mk} = 2k_{1H}k_{ADCH}U_{Xm}$ is the maximum code nominal value CVD, which respond to the maximum nominal value of measuring voltage U_{Xm}.

Further it changes mentioned k_{2PA} conversion coefficient PA and the difference of the measured and reference voltages is converted into the code N_2 by inverting switching method:

$$N_2 = N_{21} - N_{22} =$$
$$= 2k_2 k_{ADC}\left[U_X - m_H E_{NH}\frac{N_1}{N_{mk}}(1 + d_C)\right], \quad (9)$$

where

$$N_{21} = k_2 k_{ADC}\left\{U_X + \Delta_X - (mE_N + \Delta_{M1})(1 + d_C)\frac{N_1}{N_{mk}} + \Delta_{M2}\right\}$$

$$N_{22} = k_2 k_{ADC}\left\{-U_X + \Delta_X + (mE_N - \Delta_{M1})(1 + d_C)\frac{N_1}{N_{mk}} + \Delta_{M2}\right\}$$

are the codes for positive and negative polarity of measuring voltage accordingly; k_2 is the conversion coefficient PA at second measuring cycle.

Considering apparent scale value $\dfrac{N_{mk}}{m_H E_{NH}} = 2k_1 k_{ADC}$, equality (9) takes as:

$$N_2 = 2k_{2H}k_{ADCH}\left[U_X - m_H E_{NH}\frac{N_1}{N_{mk}}(1 + d_C)\right] \times$$
$$\times (1 + d_2 + d_{ADC}) \quad ,(10)$$

where d_2 is the relative error of transformation coefficient PA at second conversion cycle (subscript "H" designation of their nominal value).

Code N_x measurement result find the ratio (7):

$$N_X = N_1 + N_2\frac{k_{1H}}{k_{2H}} \cong 2k_{1H}k_{ADC}(1 - d_C)U_X \cdot \quad (11)$$

Code value measurement result will be determined by the code-controlled reference voltage measure accuracy as it can be seen from the ratio (11) for to the second order value of the smallest accuracy. Obviously, this measure CVD structure is very similar to the circuit serially fabricated voltage DAC and can be produced as a single integrated circuit. This voltage measure can be protected from exposure to electromagnetic fields and radiations by screening and from the moisture effects and harmful effects of aggressive chemicals by sealing and etc.

To increase the developed voltmeter measuring we proposed to accept the current code value of the previous conversion result for the code control CVD formation. Thus, at each transformation cycles it will change the connection polarity both the measured and reference voltages while the conversion coefficient value k_2 PA remain unchanged and we have the code series $N_{xi-1}=$ $=N_{xi-2}+(N_{i-2}-N_{i-1})k_{1H}/k_{2H}$; $N_{xi}=N_{xi-1}+(N_{i-1}-N_i)k_{1H}/k_{2H}$; $N_{xi+1}=N_{xi}+(N_i-N_{i+1})k_{1H}/k_{2H}$ and etc, where N_{i-2}, N_{i-1}, N_i, N_{i+1} are the conversion code of adjacent i-2, i-1, i and i+1 cycles. In addition, for each transformation adjacent cycles pairs will be checked held constant measured voltage: $N_{i-2}-N_{i-1} \leq \Delta_{max}$, $N_{i-1}-N_i \leq \Delta_{max}$, $N_i-N_{i+1} \leq \Delta_{max}$ and etc., For example, if at three related transformation cycles the abovementioned conditions are not met, it may mean changing the measured parameter value. After the new value measured signal is settled the equations (8)–(11) algorithm a measured voltage new value is determined to improve voltmeter performance. If

multivalue bit DSADC implement both transmission coefficients can be equal to each other $k_{1H}=k_{2H}$, that simplifies hardware and measurement result code procedure definition.

V. THE EXPERIMENTAL VERIFICATION RESULTS OF THE PROPOSED METHOD OF REMOTE ERRORS CORRECTION

The voltage calibrator experimental model with AEC automatic adjustment was made to test the basic principles of the proposed method. The classical calibrator scheme consists of such traditional elements: a reference voltage source, a voltage buffer and code-controlled voltage divider. For adjusting AEC in the developed calibrator structure we used the double inversion method of converted reference voltage and then averaging the output calibrator voltage. Thus, at the input of first one and output of the second voltage buffers, electronic polarity switches based chips MOS keys were applied. As voltage reference the chip was used with a nominal value 2, 5 V of output voltage. Code-controlled divider was implemented as series of the same nominal values resistors with conversion coefficients: 0; 0.25; 0.50; 0.75; 1.00. To check the depth error correction the imitator AEC was used which output voltage value was approximately 20 mV. It was included in the random link circuit voltage calibrator but between both polarity switches. Polarity switches were controlled by frequency signal near several tens of hertz. Output voltage calibrator was measured by precision voltmeter which had the least significant bit (LSB) ±1 μV. Voltage calibrator model was investigated by measuring its output voltage for all conversion coefficients divider and the AEC simulator source was connected and disconnected from the scheme. The experimental results of calibrator voltage model is showed that voltage values differed no more than one unit voltmeter LSB for all conversion coefficients and both connected and disconnected AEC simulator source. This indicates a good coincidence of theoretical and experimental results.

VII. CONCLUSIONS

The analysis showed that traditional methods of maintenance and operational control of metrological characteristics of measuring channels are practically difficult to apply because of CPS spatial dispersion. The differential measurement method can be applied for CPS measuring channels realization using code-control measure and summation of both measuring and reference signals directly at the unit input. It is shown that proposed technical solution makes it possible to carry out operational control of CPS measuring channels metrological characteristics if the code-control measure is made as a compact and removable unit.

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