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PHASE CONVERTER OF COMPOSING DISPLACEMENT

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Minimax strategy of mechatronic converters efficiency improving relative to error decrease with velocity increase at the same time provides common dataware level rise. The analysis of usage possibilities of different type position transducers (PT) gives the advantages of resolvers. The subsequent processing of their output signals is performed by "Resolver-to-Digit" Converter (RDC) which provides displacement digital equivalent and digital or analog signals specifying its velocity and acceleration. With appearance new electronic component, including microcontroller, for increasing of accuracy of the measurements of the corner of the tumbling resolver possible to use such methods, which earlier realize was not on. Given article contains the offers about the most further development of the technical decisions for RDC of the corner. Replaceable elementary substracting section, traditionally used in all, without exception, schemes of the automatic regulation, on more "intellectual" statistical section, possible avoid the influence signals in sidebar to feedback on accuracy of the transformation.

Keywords: stepping and brushless motors, resolver-to digital converters, substracting and statistical section, E-operation method, transfer function.

Introduction

Modern stage of robotics development is defined by microprocessor (MP) usage. To interface them with any end effector mechatronic converters (MC) are used for converting μP output signals into manipulator unit motion. This technology which has evolved over the last 40 years and brought to its highest level in Japan is crucial to increasing the competitiveness of any nation in world markets [1, 2]. Minimax strategy of MC efficiency improving relative to error decrease with speed of response increase at the same time provides dataware level rise. At dataware absence uncontrolled dispacements take place. Dataware next stage provides the presence of internal feedback that is realized in stepping motor (SM). Its invariance to load change must be provided in programming. Reduction of sensitivity to load variation and speed of response rise is achieved in local closed MC by dataware level increase at the cost of local feedback introduction. Gained quantitative growth of factors leads to qualitative changes in energy converter which transforms into brushless motor (BM) [3].

Local feedback usage does not eliminate the possibility of unacceptable loss of Information in conversion. Counteracting of external and internal disturbances is reached by using positional feedback allowing to judge reliably on absolute displacement value. Displacement optimized control requires the following dataware level providing additional forming of displacement velocity. PID algorithm control provides acceleration data production [4]. The analysis of usage details and functional possibilities of different type PT gives the advantages of brush or brushless resolvers [5–7].

1. Common Dataware with RDC

From the point of view of common dataware of BM or SM control in μP MC multicomponent converter design is of interest providing versatile usage of PT output signals resolvers *R* operating in analog phase conversion mode. RDC is specified by expanded functional possibilities in comparison with known digital tachometer types. RDC functional diagram is shown in Fig. 1.

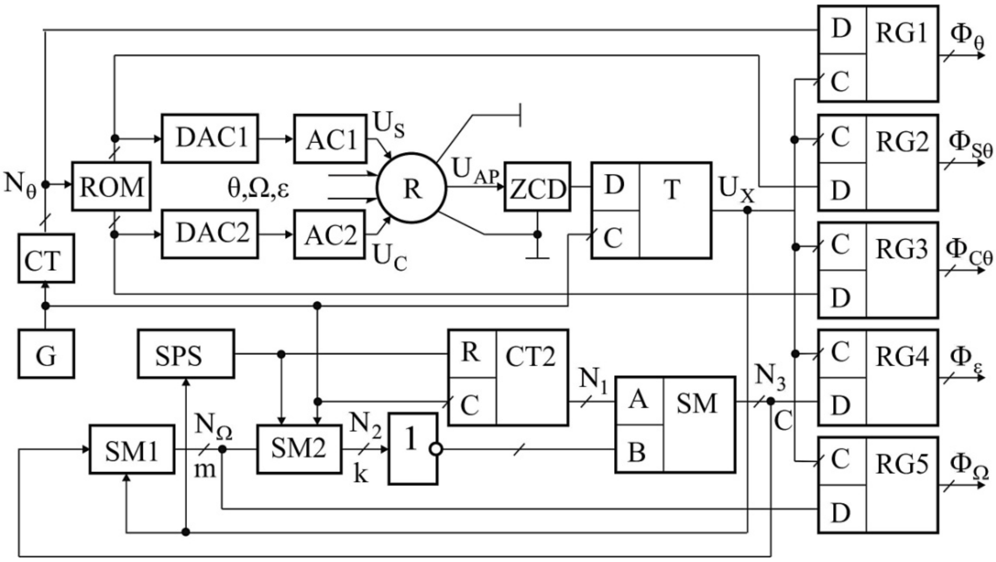


Fig. 1. Functional diagram of RDC

Slew angle Θ digital equivalent is formed from phase shift between *R* reference voltage and its output signal $U_{AP} = U_{\max}\sin(\omega_0 t - \Theta)$, where ω_0 is reference voltage frequency. The procedure of N_Θ forming does not differ from the known one [8, 9]. Angle conversion loop has zero crossing detector (*ZCD*) converting U_{AP} into rectangular pulses, *D*-trigger timed by pulse edge of pulse generator *G* and *ZCD* gating output signal so code recording moment In output registers *RG*1 and *RG*2 doesn't enter transfer stages in counter *CT*2, accumulators *SM*1 and *SM*2.

The signal at the trigger output *T* is the following: $U_x = \operatorname{sign}\sin(\omega_0 t - \Theta)$.

The value of linearly increasing code is registered in $CT2X = f{\cdot}t$, which complies with signal edge *U*. Thus we can write that $N_\Theta = f{\cdot}\Theta{\cdot}\omega_0^{-1}$, i.e. *RG*3 output code is a digital equivalent of angle Θ. The singularity of RDC is conversion loop construction for velocity Ω and acceleration *ε* and their interfacing with angular loop is effected via short pulse shaper *SPS* and binary summary counter *CT*2 with digit number N_1 and *R* reset input.

Velocity digital equivalent is calculated without any intrinsic error. This is achieved by the fact that deviation code of *P* output signal period is not received as slew rate equivalent N_Ω but subjected to conversion by closed digital system. The system includes in the loop series accumulators *SM*1, *SM*2, inverter unit and full adder *SM*3. Accumulators *SM*1, *SM*2 in their turn are designed on closed in ring binary full adders and registers in which data are registered on pulse edge coming to its timing inputs.

Accumulator *SM*1 with digit number *m* fulfills the role of a digital integrator assigning to RDC the first order astatism. Its operation is described by the difference equation

$$N_\Omega[n+1] = N_\Omega[n] + N_3[n],$$

where $N_\Omega[n+1]$ is *SM*1 output code after $(n+1)$ impulse arriving; $N_\Omega[n]$ and $N_3[n]$ are *SM*1 output and input codes respectfully after *n*-pulse arriving.

Accumulator *SM*2 is designed on the base of adder and register but differs from *SM*1 by the fact that the register resets by *SPS* output pulse. In statics with *R* rotor fixed its output signal period is $T_x = T_0 = 2\pi{\cdot}\omega_0^{-1}$.

With *R* rotor rotating with rate Ω the period is

$$T_x = 2\pi\left(\omega_0 - \Omega\right)^{-1} = 2^r f^{-1}\left(1 - \bar{\Omega}\right), \qquad (1)$$

where $\bar{\Omega} = \Omega \cdot \omega_0^{-1}$ is a relative slewing rate.

All further positions are valid for small Ω, i.e. for a case when $[\Omega] \ll \omega_0$.

For velocity and acceleration definition the period increment $T = T_x - T_0$ of RDC output signal T_x versus reference T_0 is measured first. For this *CT*2 resets by U_x approaching at *CT*2 output additional code is generated in proportion with period difference $N_1[n] = f(T_x - T_0) = 2\pi\overline{\Omega}f(\omega_0 - \Omega)^{-1}$.

At the same time code adding in *SM*2 takes place for every impulse *G* and at the same moment $N_2[n] = N_\Omega f T_x = 2\pi N_\Omega f(\omega_0 - \Omega)^{-1}$. Thus *SM*2 multiplies *SM*1 output code N_Ω by *R* output signal period T_x.

The digit capacity *SM*2 is $k > m$. The output code *SM*1 with digit capacity m goes to least significant digit *SM*2 input, to the rest most significant digit *SM*2 $(k - m)$ sign digit of *SM*1 is given. Transfer factor $k_t = 2^{m-k}$ is realized at the expense of digit matrix shift between *SM*1, *SM*2 and *SM*3. Accumulator *SM*2 operates only in the mode of additional code N_Ω summing and forming code N_2 which is subtracted from period increment code N_1 with the help of invertors unit and adder *SM*3. Code N_3 is generated by subtraction of these codes and at the moment of signal edge U_x arrival corrects *SM*1 content (code N_Ω) so that mismatch $N_3[n]$ is tending to zero. At a steady state $N_1[n] = N_2[n]$, i.e. $N_\Omega = \Omega$. Mismatch $N_\varepsilon = N_3[n]$ developed at the moment of signal edge U_x arrival at *SM*3 output is registered in *RG*2. As N_ε is an input code of *SM*1 digital integrator the output code N_Ω of which is proportional to velocity Ω, then $\mathrm{N}_3 = N_\varepsilon \approx \overline{\varepsilon}$ where $\overline{\varepsilon} = 2\pi\varepsilon\omega_0^{-2}$, i.e. N_ε at the *RG*4 output in a first approximation is proportional to acceleration ε. The higher ω_0 the more rigorous the rule is. As N_Ω and N_ε shaping is realized in digital closed loop stability analysis was performed which showed the opportunity of its provision without any correction with minimax approach to intrinsic error and angular frequency selection [10].

2. The dynamics of phase RDC

Since the forming N_Ω and N_ε occurs in the closed-loop digital system, it's required to investigate its dynamic properties. For this purpose let's use *z*-transform method [11] and perform our study in two steps [7].

At the first step let's consider stabilized conditions in terms of speed.

In this case polyfunctional phase displacement transducer can be considered as linear pulse system with constant coefficients for which a necessary and sufficient condition of stability will be as follows:

$$|z_i| < 1,$$

where z_i – the roots of the characteristic equation

$$1 + W(z) = 0. \tag{2}$$

In order to determine transfer function of the open-loop system $W(z)$ let's use linearized equivalent structural schematic (Fig. 2, a) not taking into account level quantization:

$$W(z) = W_1(z) \cdot W_2(z), \tag{3}$$

where $W_1(z)$ and $W_2(z)$ respectively transfer functions (TF) *SM*1 and *SM*2 adders.

Counter adder *SM*1, which functions as digital generator, can be described by TF $W_1(z) = (z-1)^{-1}$. Counter adder *SM*2, which is returned to zero in each transformation cycle, in terms of dynamics occurs to be an amplifying instantaneous element with transfer function which takes into account that output code *SM*1 is sent to the inputs of least significant bits of *SM*2. With taking into account equation (3) TF of the open-loop system will be:

$$W(z) = 2^{m-k} f T \cdot (z-1)^{-1} = 2^l \cdot \left[(1 - \overline{\Omega})(z-1)\right]^{-1}, \tag{4}$$

where $l = r + m - k$ – parameter is an integer.

The characteristic equation (2) for (4) takes the form $(z-1)(1-\Omega) + 2^l = 0$. It has the same root:

$$z_1 = 2^l \cdot (1-\Omega)^{-1} + 1. \tag{5}$$

In order to find optimum parameter value from stability domain we carry out estimation of the transition process under stepwise speed variation from 0 to $\Omega = \text{const}$. This case is interesting when we use secondary converter in multiplexed RDC operating in multi-channel system from several primary phase shift modules.

The image of input signal of the system [11]:

$$X(z) = z\Delta Tf \cdot (z-1)^{-1} = z2^r\bar{\Omega} \cdot \left[(z-1)(1-\bar{\Omega})\right]^{-1}, \tag{6}$$

and the image of its output signal:

$$Y(z) = W_3(z)X(z), \tag{7}$$

where $W_3(z) = W_1(z)\left(1+W_1(z)\right)^{-1}$.

Taking into account (3), (4) and (6), (7) can be transformed as follows:

$$Y(z) = 2^r\bar{\Omega}\, z \cdot \left[(z-1)(1-\bar{\Omega})\right]^{-1} \cdot \left(z-(1-\bar{\Omega}-2^l)(1-\bar{\Omega})^{-1}\right). \tag{8}$$

Applying inverse z-transform to equation (8) we can find a lattice function which describes variation of the output code $N_\Omega[n]$ in transition process:

$$Y[n] = N_\Omega[n] = \bar{\Omega} \cdot 2^{k-m}\left[1-\left(1-2^l-\bar{\Omega}\right)/\left(1-\bar{\Omega}\right)\right]. \tag{9}$$

Analysis of (9) for the parameter values shows that the transition process: at $l = 1$ is oscillatory, and its duration is long; at $l < 0$ is monotonous, and its duration decreases with increase of l; at $l = 0$ have minimum duration; if $\Omega > 0$, then the process is oscillatory, if $\Omega < 0$, then – monotonous, if $\Omega = 0$, then it ends after one clock period.

Thus, the optimum value of the parameter is $l = 0$. In this case relative dynamic error of velocity measurements will be: $\delta_\Omega[n] = (-\bar{\Omega})^n \cdot (1-\Omega)^{-n}$.

As we have studied stability using linearized model, it's required to check for the condition of absence of periodic behavior which can occur in digital systems due to level quantization. In order to exclude periodic behavior, phase characteristic $\psi(\lambda)$ of the open-loop system at fixed frequencies:

$$\lambda_N = 2T_x^{-1}\,\mathrm{tg}(0,5\pi), \tag{10}$$

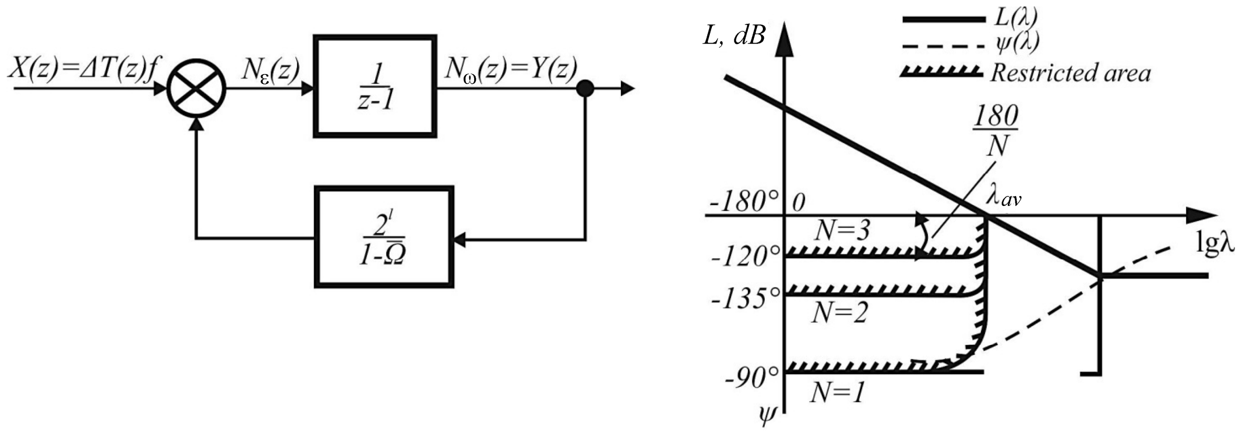
shall not enter forbidden regions, drawn up for $N=$const, provided that $N = 1, 2, 3\ldots$ – half of relative oscillation period, and λ – quasi-frequency [11].

For determining frequency transfer function $W(j\lambda)$ using TF (4) we use a well-known substitution [11]: $z = (1+0,5j\lambda T_x)\cdot(1-0,5j\lambda T_x)^{-1}$.

Frequency transfer function of the system:

$$W(j\lambda) = 2lj\lambda T_x \cdot (1-\bar{\Omega})^{-1} \cdot (1-jT_x \cdot (2\lambda)^{-1}). \tag{11}$$

Logarithmic frequency characteristics for $l \le 0$ drawn up using equation (11) shown in Fig. 2, b. The case $l = 1$ ($\Omega < 0$) is not considered as it has no practical importance due to long duration of the transition process and overcorrection $\sigma \approx 100$ %.



a) b)

Fig. 2. The linearized equivalent circuit structure (a) and logarithmic frequency response (b) polyfunctional phase RDC

From $N = 1$ and $N = 2$ it follows that the frequency is higher than cutoff frequency, therefore the condition of absence of periodic behavior is fulfilled. For $N \geq 3$ phase characteristic doesn't enter forbidden region, so how $\varphi(\lambda) > -135°$ at $\lambda \leq \lambda_{av}$. Therefore, the absence of symmetric periodic behavior is proved.

At the second step of our dynamics research, we will consider such operating conditions of phase RDC when acceleration value is constant. Current value of Ω in n-th period:

$$\Omega[n] = n\varepsilon T_x[n] = nC ,$$

where $C = \varepsilon T_x[n] = \text{const}$.

Lattice function of acceleration $N_\varepsilon\left[n\right]$ can be found from structural schematic shown in Fig. 2, a:

$$N_\varepsilon[n] = \Delta T[n] f - 2^l N_\Omega[n] \cdot \left(1 - \overline{\Omega})[n]\right)^{-1} .$$

Applying passage to the limit [11]: $\varphi\left[n\right] = \lim\limits_{z \to 1}\left(z-1\right)\varphi\left(z\right)$, we obtain stabilized value of acceleration code:

$$N_\varepsilon = 2^r T_x \varepsilon \cdot \left(2^l \omega_0 - T_x \varepsilon\right)^{-1} = \left(1-\overline{\Omega}\right)^{-1} \overline{\varepsilon} \cdot 2^{k-m} \cdot \left(1 - \left(1-\overline{\Omega}\right)^{-1} \overline{\varepsilon} \cdot 2^{-l}\right)^{-1} ,$$

where $\overline{\varepsilon} = 2\pi\varepsilon\omega_0^{-2}$.

From which it follows that during acceleration measuring a systematic error can occur which decreases with increasing of parameter l. Minimum error occurs with the optimum, in terms of response time, value of parameter $l = 0$. Analysis of drawing up peculiarities and dynamic properties of cyclic-type phase RDC shows that without the need for correction of dynamic characteristics it provides stable operation when minimax approach is utilized for selection of such its parameters as systematic error and response time.

3. Synthesis of the digital tracking algorithm with regressive statistical unit in the feedback loop for phase angle transducer of high accuracy resolver

The results of studying the effects of stationary random fluctuations on the accuracy of the tracking system, including the RDC are presented, for example, in papers [13–15]. They also propose methods for countering random discrete processes with known mathematical expectation and the correlation function. Feasible synthesis of digital witness regression algorithm static link in the feedback circuit. To do this, it is proposed to substitute the conventional differential unit, traditionally used in all without any exception structures of the tracking systems in the automatic control theory [15] for the unit with the program mathematical regression treatment (Fig. 3) of the total number of previous error values, that occur during the control procedure in real time.

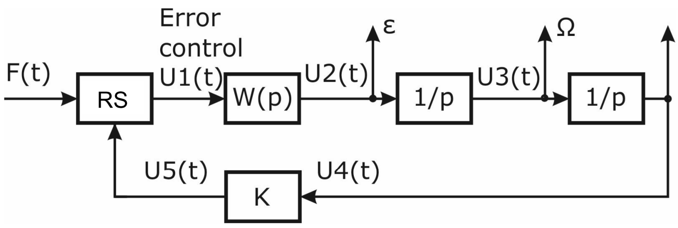


Fig. 3. The structure of the proposed multifunctional digital tracking algorithm with the RS unit in the feedback loop

We write the system of equations describing this structure:

$$W\left(p\right) = k_1 \cdot \left(1 + T_1 \cdot p\right)^{-1} . \tag{12}$$

$$\begin{cases} U1\left(p\right) = R\left(F\left(p\right), U5\left(p\right)\right); \\ U2\left(p\right) = W\left(p\right) \cdot U1\left(p\right); \\ U3\left(p\right) = U2\left(p\right) \cdot p^{-1}; \\ U4\left(p\right) = U3\left(p\right) \cdot p^{-1}; \\ U5\left(p\right) = U4\left(p\right) \cdot K_2. \end{cases} \tag{13}$$

Consider the first equation of the system (13). It corresponds to a mathematical model of a regression statistical unit in the feedback loop. We are using as a transfer function of R error objective function (*EOF*), reminiscent of the function of the Kalman filter, but with mutually independent coefficients α and β:

$$EOF = R\left(F(t), U5(t)\right) = \sum_{i=1}^{N}\left(F(t) - \alpha \cdot U5(t) + \beta\right)^2, \quad (14)$$

where $F(t)$ [rad] – targeting signal; $U5(t)$ – a feedback signal; α [rad/B], β [rad] – dimensional regression coefficients generally depend on N; N – a number of measuring points that (almost always) is known in advance.

Otherwise, on accumulating the points value of the regression error will change over time, but still being within the meaning of the $EOF = U1(t) = 0$.

The problem of determining the coefficients α and β in (14) is restricted to finding the minimum of the EOF, so you should write down and solve the linear system of equations consisting of the partial differential coefficientsof EOF in accordance with α and β. Equating to zero the equation to find the extremum, after simple transformations we obtain:

$$\begin{cases} \alpha \cdot \sum_{i=1}^{N} U5^2(t_i) - \beta \cdot \sum_{i=1}^{N} U5(t_i) = \sum_{i=1}^{N} U5(t_i) \cdot F(t_i); \\ \alpha \cdot \sum_{i=1}^{N} U5(t_i) - \beta \cdot N = \sum_{i=1}^{N} F(t_i). \end{cases} \quad (15)$$

Solving the system (15), we obtain:

$$\alpha = \frac{\begin{vmatrix} \sum_{i=1}^{N} U5(t_i) \cdot F(t_i) & -\sum_{i=1}^{N} U5(t_i) \\ \sum_{i=1}^{N} F(t_i) & -N \end{vmatrix}}{\Delta}; \quad \beta = \frac{\begin{vmatrix} \sum_{i=1}^{N} U5^2(t_i) & \sum_{i=1}^{N} U5(t_i) \cdot F(t_i) \\ \sum_{i=1}^{N} U5(t_i) & \sum_{i=1}^{N} F(t_i) \end{vmatrix}}{\Delta}, \quad (16)$$

where

$$\Delta = \begin{vmatrix} \sum_{i=1}^{N} U5^2(t_i) & \sum_{i=1}^{N} U5(t_i) \\ \sum_{i=1}^{N} U5(t_i) & N \end{vmatrix}. \quad (17)$$

Since the EOF (14) is represented by an even function, the current error must be calculated as follows:

$$U1(t) = F(t) - \alpha_0 \cdot U5(t) + \beta_0, \quad (18)$$

substituting values $\alpha_0 = \alpha,\ \beta_0 = \beta$ calculated from (16).

To calculate the values of α and β in real-time demands specific computing resources from the processor on which the system will be implemented.

We now turn to the rest of the equations of the system (13) by applying the method of algebraic E operator transformation of differential equations. This efficient method of converting the differential equations to the form of algebraic difference equations is described in [16]. Recall that E- operator, similar to the Laplace operator allows you to “replace” differentiation by the set of algebraic operations. Moreover, the operator p^k must be replaced by the formula $(1-E)^k \cdot \Delta t^{-k}$ and then output variable explicitly written in the form of differential algebraic equation with a parameter of time.

Consider the second equation of the system (13). Physically, it expresses the value of the angular acceleration ε of the resolver barrel rotation.

In view of (17) corresponds to:

$$U2(p) = U1 \cdot k_1 (1 + T_1 p)^{-1}. \quad (19)$$

Using E method of transformation of an operator equation in the differential algebraic equation, we get:

$$U2(t) = \left(k_1 \cdot U1(t) + U2(t) \cdot E \cdot T_1 / \Delta t\right)\left(1 + T_1 \Delta t^{-1}\right). \quad (20)$$

Consider the third equation of the system (13). Physically, it expresses the value of the angular speed Ω of the resolver barrel. Using E method, we get the difference algebraic equation:

$$U3(t) = U2(t)\Delta t + E \cdot U3(t). \quad (21)$$

Consider the fourth equation of the system (13). Physically, it expresses the value of the barrel deflection angle in the resolver [17]. Using E method, we obtain finite-difference equation:

$$U4(t) = U3(t)\Delta t + E \cdot U4(t). \tag{22}$$

The fifth equation of the system (13). It corresponds to a very simple algebraic equation:

$$U5(t) = U4(t) \cdot K_2. \tag{23}$$

Now with the obtained equations (18, 20–23), you can easily construct a flowchart of the algorithm of the digital phase RDC (Fig. 4) with regressive statistical unit in the feedback loop.

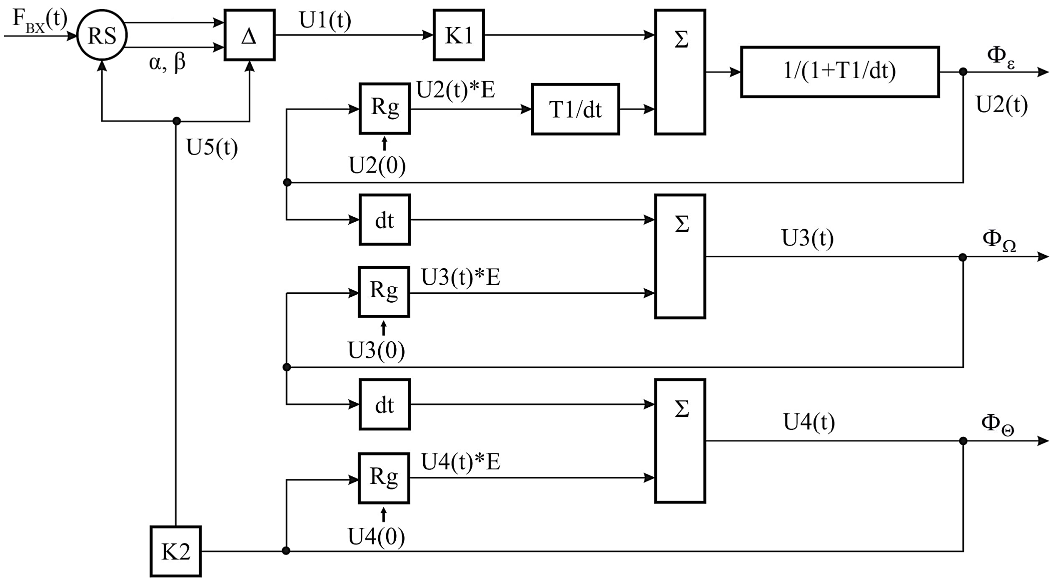


Fig. 4. The flowchart of the digital phase RDC with RS unit in the feedback loop

We recall that all coefficients are numeric constants calculated in the programming process prior to the firmware in the microprocessor flash- memory.

4. Microcontroller version of the RDC

The first step of electronization allows to simplify electromechanical part of electromechatronic converter due to exclusion of reduction gear, speed-voltage generator and accelerometer. However this leads to a certain complication of microelectronic part. Feasibility of this approach is explained by the use of microcontroller in indicating part of polyfunctional phase displacement transducer, which functional algorithm harmoniously fits the modern trend to replace electromechatronic converter hardware with its programmable analogues.

Microcontroller is selected according to the following requirements:

- availability of two modules of digital-to-analog converters;
- availability of analogue comparator module;
- availability of two 16-bit timers;
- availability of required communication interfaces.

Functional diagram of the RDC using microcontroller (MC) presented in Fig. 5.

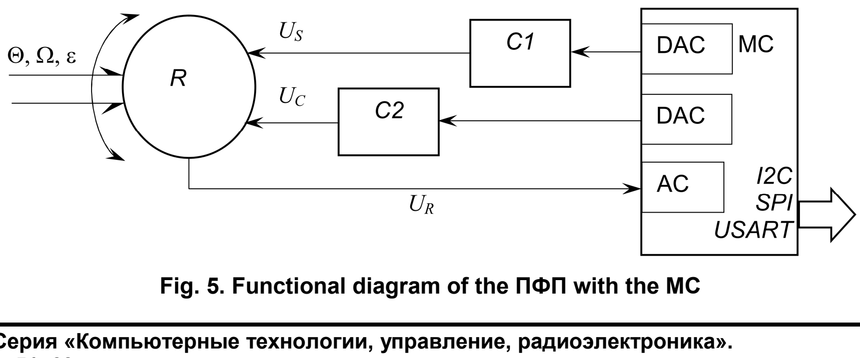


Fig. 5. Functional diagram of the ПФП with the MC

Bit synchronizing of microcontroller operation is performed by the internal clock generator which pulses are clocking for both timers-counters. The internal memory of microcontroller programs contains firmware which generates reference signals U_S and U_C send them from *DAC*-modules output via converters *C*1 and *C*2 to phase shift module. The total count of reference of timer-counter *TMR*1 from 0 to overflow corresponds to reference signal cycle. Output voltage of phase RDC U_R is applied to the input of analogue comparator which detects its zero transients. Reference voltage applied to the inverting input of analogue comparator is the voltage which corresponds to zero voltage level U_R. Comparator output U_0 forms a signal equal to 1 on positive half-wave U_R and a signal equal to 0 on negative half-wave U_R. Timing diagrams of polyfunctional phase RDC operation is shown on the Fig. 6. Timer *TMR*2 is applied for tracking the signal U_R from the output of phase RDC. If reference timer *TMR*1 is a complete 16-bit timer-counter, then when the speed of phase shift module is higher than the speed of reference signal, the overflow of timer *TMR*2 will occur. Therefore, to determine phase shift module parameters it's practicable to use timer *TMR*2 with preliminary dividing coefficient equal to 2.

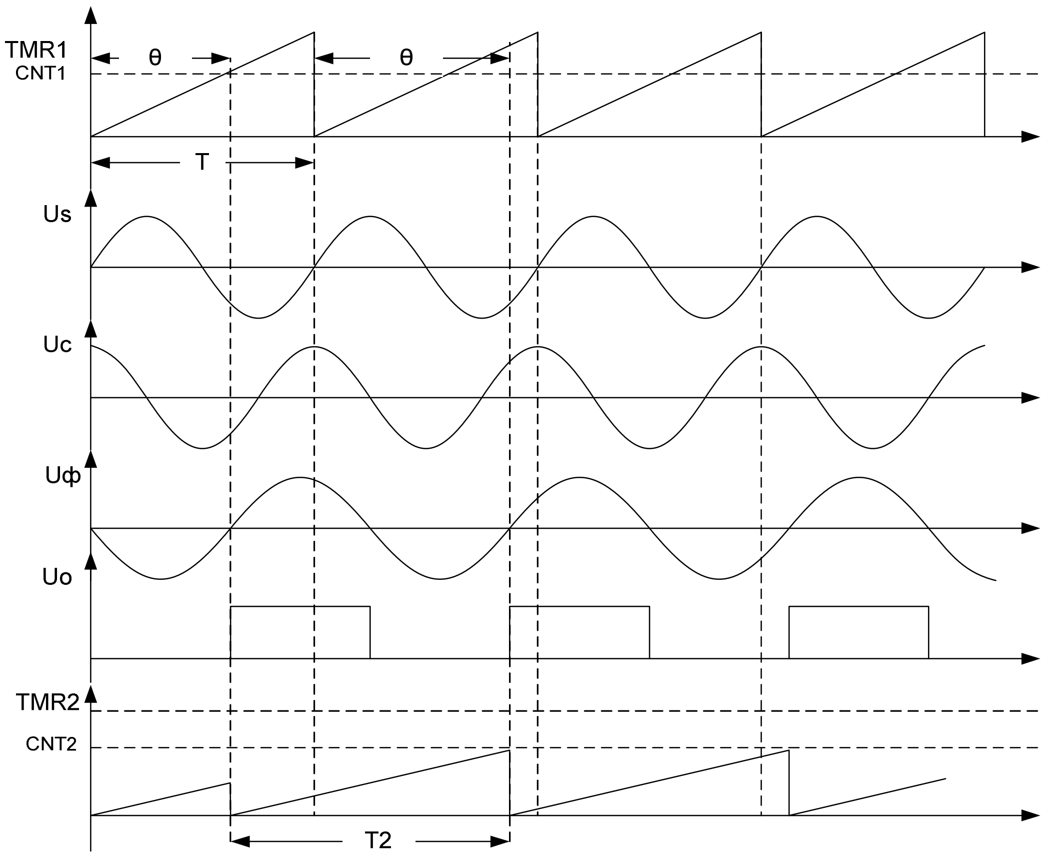


Fig. 6. Timing diagrams of polyfunctional phase RDC

Assuming clock frequency to be f_{OSC}, one cycle of counting of timer *TMR1*: $t_{TMR1} = f_{OSC}^{-1}$, reference time of timer *TMR*2 with preliminary dividing coefficient equal to 2: $t_{TMR2} = 2f_{OSC}^{-1}$.

Signal cycle U_R is determined by its rising edge U_0 at the output of analogue comparator *AC*. Processing of the described signal is practicable to perform through detection of its interruptions when it changes from 0 to 1 at the comparator output. In this moment of time value *CNT*2 of timer *TMR*2 is recorded, which determines time interval:

$$T2 = CNT2 \cdot t_{TMR2}. \quad (24)$$

Phase shift is determined by value *CNT*1 of timer *TMR*1 at the point of rising edge of the signal from comparator output: $\Theta = CNT1 \cdot t_{TMR1}$.

Module of conversion of 16-bit complete timer-counter $N = 2^{16}$, which determines the cycle of reference frequency:

$$T = N \cdot t_{TMR1}. \quad (25)$$

Reference frequency is determined as follows: $\omega_0 = 2\pi \cdot T^{-1} = 2\pi \cdot f_{OSC} \cdot N^{-1}$.

Rotating frequency of phase shift module *R*:

$$\omega = 2\pi\left(T2 - T\right)^{-1}. \quad (26)$$

Substituting (24) and (25) into (26) we obtain the following:

$$\Omega = \frac{2\pi \cdot f_{OSC}}{CNT2 \cdot 2 - N}. \quad (27)$$

Speed value in (27) is a signed value which determines the rotation direction of phase shift module.

Angular acceleration is determined by the difference equation: $\varepsilon = \dfrac{\omega_k - \omega_{k-1}}{T2_k - T2_{k-1}}$.

In order to reduce noise component a digital filter can be used which does not contribute to phase shift in the resulting signal, e.g. αβ-filter. At each transformation step the following parameters are determined: $\Delta\Omega_k = \Omega_k - \widehat{\Omega}_{k-1}$, $\Delta T2_k = T2_k - T2_{k-1}$. Angular speed at the next transformation step: $\widehat{\Omega}_k = \widehat{\Omega}_{k-1} + \alpha \cdot \Delta\Omega_k$. Angular acceleration at the next transformation step: $\widehat{\varepsilon}_k = \widehat{\varepsilon}_{k-1} + \beta\Delta\Omega_k\left(\Delta T2_k\right)^{-1}$.

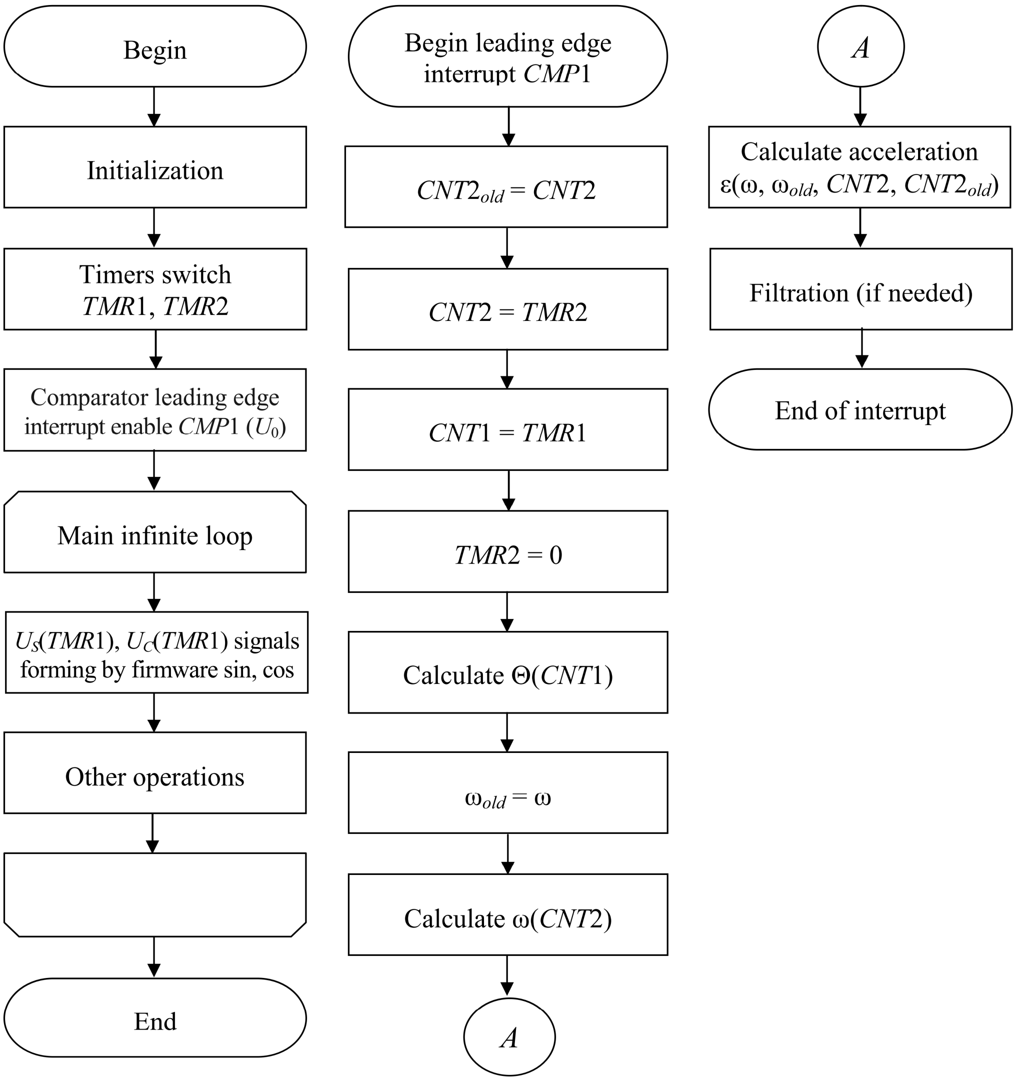


Fig. 7. Algorithm of the microcode program

Coefficients α and β are determined by model approach from the range 0…1. Utilizing of such a variant of digital filtering significantly increases noise resistance of polyfunctional phase displacement transducer, which forms digital equivalents of displacement in closed-loop digital system. For implementation of the device was used microcontroller STM8L151C8 manufactured by ST Electronics. Execution algorithm of the microcode program of the microcontroller is shown at Fig. 7.

Conclusion

The second step of electronization provides for improvement of primary sensor through moving from kinesthetic electromechanical angle-sensing transducers to generator microelectronic magnetic field sensors caused by the following reasons:

- the cost of electromechanical sine-cosine angle-sensing transducer is three orders higher than the cost of magnetic field sensor with comparable error;
- precision electromechanical proximity angle-sensing transducer has complicated design and complex technology of manufacturing due to a special copper wire required for its production and special precision bearings with limited service life;
- the use of direct transformation when forming displacement components of the structure limits utilization of polyfunctional phase displacement transducers in electromechatronic converters with high noise level;
- maximum speed of water action for such polyfunctional phase displacement transducer does not exceed $90' \cdot \mathrm{sec}^{-1}$ due to fundamental limitations inherent to cyclic type converters “angle – phase – code” [1, 8];
- the absence of quaternary winding in a number of electromechanical angle-sensing transducers that excludes their use as phase shift modules.

Elimination of these limitations when moving from kinesthetic electromechanical primary sensors to generator magnetic field sensors involves the use of cyclic or tracking structures of amplitude-type analog-to-digital converters which have enhanced noise resistance as compared to the phase RDC. The peculiarities of drawing up and operation of amplitude self-organizing multi-component converter which form Φ_{Θ}, Φ_{Ω}, Φ_{ε}, $\Phi_{S\Theta}$ and $\Phi_{C\Theta}$ displacement by the method of analog-digital integration in closed-loop electronic tracking system are shown in. To the second step of electronization of electromechatronic converters contributes circular and linear variants of primary sensors for magnetic field sensors [19].

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ФАЗОВЫЙ КОНВЕРТОР СОСТАВЛЯЮЩИХ ПЕРЕМЕЩЕНИЯ

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Минимаксная стратегия проектирования мехатронных систем предусматривает снижение погрешности измерения скорости при условии применения единого информационного обеспечения. Анализ функциональных и метрологических возможностей различных типов датчиков положения отдает предпочтение резольверам. Преобразование их выходных сигналов «Resolver-to-Digit Convertor» (RDC) формирует цифровой эквивалент перемещения и аналоговые эквиваленты составляющих его скорости и ускорения. С появлением новых электронных компонентов, в том числе микроконтроллеров, для повышения точности измерений угла поворота резольвера можно применить методы, которые раньше реализовать было невозможно. Представленная статья содержит предложения о дальнейшем развитии технических решений для многофункциональных фазовых преобразователей угла. Одним из ключевых моментов статьи является замена элементарного разностного звена, традиционно используемого во всех без исключения структурных схемах автоматического регулирования, на более «интеллектуальное» статистическое звено. Это позволяет формировать текущее значение ошибки управления с помощью математического регрессионного анализа всех ее предшествующих значений и, тем самым, устранить влияние случайных флуктуаций сигналов в контуре обратной связи на точность преобразования.

Ключевые слова: шаговый и вентильный электродвигатель, резольвер, разностный и статистический алгоритм, E-операторный метод, передаточная функция.

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