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FUNCTIONAL DIGITAL-TO-ANALOG CONVERTERS - NEW OPPORTUNITIES OF INSTRUMENT MAKING

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Abstract. The digital-to-analog converters (DAC) are widely applied for transformation of signals in an informational - measurement technology, radio engineering and instrument making. They are a potent resort of increase readout accuracy. However existing a DAC can execute only linear transformation of a kind $U_{out}=kU_{ref}$. At the same time in different areas of engineering it is necessary with a split-hair accuracy to replicate non-linear functional connections. For example, such necessity arises at construction digital-controlled of phase changes and calibrators of a phase, as the relation of a phase shift to change of a controlled variable is always non-linear. The most preferential way for implementation of this operation - usage a DAC.

Keywords: functional digital-to-analog converter, calibrators of a phase,

1. FUNCTIONAL DAC WITH USAGE PIECEWISE LINEAR APPROXIMATION

For reproduction of non-linear functional connections and simulation them with the help a linear DAC we have used two kinds of approximating: piecewise linear approximating and approximating by power serieses. Each of the mentioned kinds of approximating guesses the way of a hardware representation. However both ways for increase of a discretization of a reproduced function will use a standard multidigit DAC.

Let's consider the first way, in which one the linear approximating of a reproduced functional connection $y = f(x)$ in an interval will be used $[a+i\,(b-a)/n,\ a+(i+1)\,(b-a)/n]$. Here $[a, b]$ - section, which one is staggered, $h = (b-a)/n$ - value of an interval of approximating of a functional connection $f(x)$, $i=0, 1, 2..., n-1$ (i - number of an interval of approximating, n - number of sections of approximating of relation $f(x)$). We suppose, that on a section of approximating the function is nonnegative, though the given way can be amplated on negative values of a function.

As the multidigit linear DAC is necessary to use any multiplying DAC with a constant input resistance R_0 [1, 2]. For example, the DAC operating a resistor matrix *R-2R* and controlled by a binary code, or voltage divider with shunting decades controlled by a decimal code .

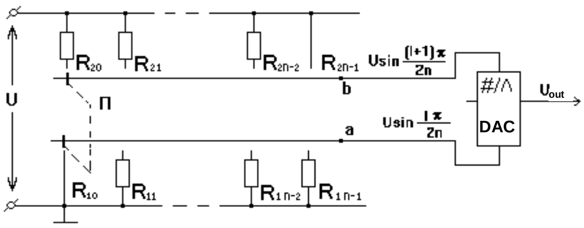

Fig. 1. The functional digital-to-analog converter

In a fig. 1 the scheme illustrating usage of linear approximating for reproduction of a functional connection $f(x)$ is adduced.

If to suspect, that $U = \max\limits_{[a,b]} f(x)$ then the values of resistance R_{1i}, R_{2i}, R_0 will be connected among themselves by ratio:

$$R_{1i} = \frac{R_0 f(a+ih)}{f(a+(i+1)h) - f(a+ih)}, \qquad (1)$$

$$R_{2i} = R_0 \frac{\max\limits_{[a,b]} f(x) - f(a+(i+1)h)}{f(a+(i+1)h) - f(a+ih)}. \qquad (2)$$

Voltage in the scheme (fig. 1) thus to be arranged as follows:

$$U_a = U\frac{f(a+ih)}{\max\limits_{[a,b]} f(x)},\ U_b = U\frac{f(a+(i+1)h)}{\max\limits_{[a,b]} f(x)}, \qquad (3)$$

where U - voltage made to a non-linear DAC.

With the help a linear DAC the output voltage U_{out} changes from value U_a up to, U_b, approximately replicating in the *i* interval relation *f(x)* with a given quantum. For change of a control interval of a functional connection $f(x)$ (the most significant bit in the nonlinear DAC) will be used a key S (fig. 1).

Naturally, that at such construction a functional DAC the last has the methodical error called by linear approximating of a functional connection $f(x)$ in the interval $[a+i\frac{(b-a)}{n},\ a+(i+1)\frac{(b-a)}{n}]$. This error is determined by expression

$$\Delta = f(x) - y_{1i}, \qquad (4)$$

where y_{1i} is the equation of the straight line, approximating relation $f(x)$ in the interval $[a+i\frac{(b-a)}{n},\ a+(i+1)\frac{(b-a)}{n}]$. Let's enter notation

$$\Delta f = f\left(a+(i+1)\frac{(b-a)}{n}\right) - f\left(a+i\frac{(b-a)}{n}\right), \qquad (5)$$

then the equation of the straight line will be written to a kind:

$$y_{1i} = \frac{\Delta f}{h}(x-(a+ih)) + f(a+ih). \qquad (6)$$

Here $a+i\frac{(b-a)}{n} \le x \le a+(i+1)\frac{(b-a)}{n}$. The methodical error of reproduction of a function $f(x)$ will depend first of all on its kind and selected section of approximating, and also from numbers and quantities of sites of approximating.

Nonlinear DAC, the approximating functions $sin\varphi$ and $cos\varphi$ are widely applied at construction of calibrators of a phase and phase changes (sine-cosine potentiometric phase changes), in which for formation of the output voltage will be realised the ratio:

$$\dot{U} = U\cos\varphi + jU\sin\varphi = Ue^{j\varphi}. \qquad (7)$$

2. DISCRETE SINE-COSINE POTENTIOMETER

As an example we shall consider the construction of the functional digital-analog converter - discrete clone of the sine-cosine potentiometer.

The known sine-cosine potentiometers have found a use in radio engineering, to an informational - measurement technology [3]. They use contoured winding of a wire and have a sliding contact, that is non-technological, does not allow to receive the good metrology characteristics and does not enable to apply them in microelectronics.

The fulfilment of the sine-cosine potentiometer from a set of resistors and keys would result in necessity of application a plenty of resistors of miscellaneous ratings and keys. So, for the discrete potentiometer replicating sine relation of resistance from change of an input code in range from 0 up to 90° and discretization 0.1°, would need 900 resistors and 901 keys.

Quite apparently, that the similar discrete sine potentiometers are very bulky and, besides at such implementation of the sine potentiometer impossible are to made by digit-by-digit regulation of a reproduced functional connection.

From these lacks the scheme , reduced in a fig. 2, constructed on the basis non-linear DAC — of the digital-to-analog converter replicating non-linear dependence $sin\varphi$ is free.

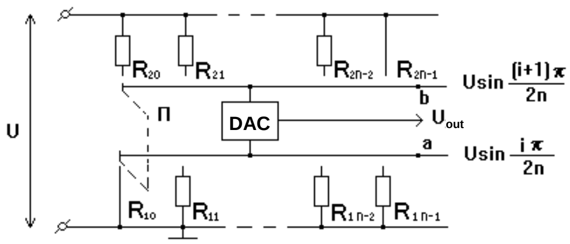

Fig. 2. A functional DAC is discrete analog of sine potentiometer.

With the help of a twin key the linear multidigit DAC, is hooked up to resistors R_{1i} and R_{2i} - and will derivate the converter modelling relation $sin\varphi$ in the interval from $i\pi/2n$ up to $(i+1)\pi/2n$. Here $\pi/2n$ is the interval of approximation for the functional dependences $sin\varphi$, $i=0,1,2..., n-1$ (i is the number of of approximating intervals, n is the number of sections of approximating of relation $\sin\varphi$ in the interval from 0 up to $\pi/2$). Any DAC with a constant input resistance R_0 can be utilized as a multidigit liner DAC.

The resistances R_{1i}, R_{2i}, R_0 are interconnected by the relations

$$R_{1i} = \frac{R_0 \cdot sin\frac{i\pi}{2n}}{2\,sin\frac{\pi}{4n} \cdot cos\frac{(2i+1)\pi}{4n}}; \qquad R_{2i} = \frac{R_0\left[1 - sin\frac{(i+1)\pi}{2n}\right]}{2\,sin\frac{\pi}{4n} \cdot cos\frac{(2i+1)\pi}{4n}}. \qquad (8)$$

The voltage in the scheme of a fig. 2 will be arranged as follows:

$$U_a = U\,sin\frac{i\pi}{2n},\ U_b = U\,sin\frac{(i+1)\pi}{2n}, \qquad (9)$$

where U is the voltage made to the discrete analog of the sine potentiometer - the functional DAC.

With the help the linear DAC the output voltage U_{out} changes from value $U\,sin\frac{i\pi}{2n}$ up to $U\,sin\frac{(i+1)\pi}{2n}$, approximately reproducing $sin\varphi$ by the necessary number of bits within the i-th range. Switch S is used to change the interval within which $sin\varphi$ (the most significant bit in the nonlinear DAC) is regulated.

Naturally, that at such construction of discrete analog of the sine potentiometer the last has methodical error called by linear approximating of relation $sin\varphi$ in an interval $[i\pi/2n,\ (i+1)\pi/2n]$. This error is determined by expression

$$\Delta = \sin\varphi - y_{1i}\,, \qquad (10)$$

where y_{1i} - equation of a straight line, approximating relation $sin\varphi$ in an interval$[i\pi/2n, (i+1)\pi/2n]$. :

$$y_{1i} = (i+1)\sin\frac{i\pi}{2n} - i\sin\frac{(i+1)\pi}{2n} + \qquad (11)$$
$$+ 4\frac{n\varphi}{\pi}\cdot\cos\frac{(2i+1)\pi}{4n}\cdot\sin\frac{\pi}{4n}$$

Here $i\pi/2n \le \varphi \le (i+1)\pi/2n$.

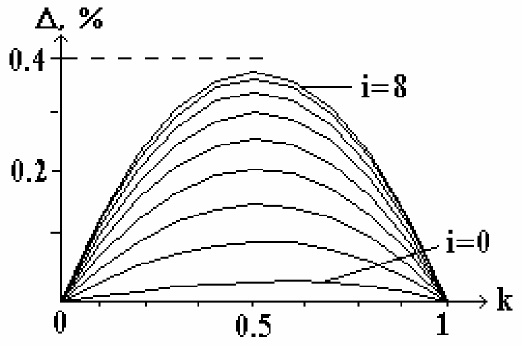

Fig. 3. Distribution of the methodical error of the discrete analog of the sine potentiometer inside the ranges of approximating ($n=9$).

Figure 3 shows the dependance of the methodical error of $sin\varphi$ within the approximation interval (Δ) for $n=9$. In this case, the most significant bit for control of the phase shift corresponded to $10°$. For convenience, we introduce a new variable k from the relation

$\varphi = (i+k)\cdot\pi/2n$. here $0 \le k \le 1$.

The greatest concern is introduced by the maximum error received at approximating of the last interval of relation $sin\varphi$. In tab. 1 the maximum magnitudes of error are adduced depending on number of intervals of approximating.

TABLE 1. Methodical error for different number of intervals

n	3	9	18	36	72
Δ, %	3,29	0,38	0,09	0,02	0,006

So, in a case $n=9$, she makes value less than 0,4 %, that can be reasonable to many operational uses.

In case of simultaneous usage of sine and cosine converters, as it is made in activity [4] for construction of the phase change, the methodical error of reproduction of a phase shift receives small, as the errors of converters have one sign. So, at decade regulation of a phase shift ($n=9$) the methodical error called by approximating of relations $sin\varphi$ and $cos\varphi$, makes all of half-minute.

The discrete analog of the sine potentiometer (fig. 2) enables to change relation $sin\varphi$ within the limits of one quadrant. All same members will be used and for regulation of relation $sin\varphi$ in all four quadrants, but in the third and fourth quadrants for obtaining negative values the inverter will be used or in an input circuit (the input voltage) is inverted, or in an output circuit (the voltage output) is inverted. Besides all circuit components of a fig. 2 are applied and to construction of discrete clone of the cosine potentiometer, as $cos\varphi=sin(\pi/2+\varphi)$. Thus, the scheme, reduced in a fig. 2, can almost completely supply simultaneous simulation both sine, and cosine converters.

The reviewed way is applicable for construction digital-controlled of potentiometric and bridge phase changes.

3. FUNCTIONAL DAC WITH USAGE POLINOMIAL APPROXIMATION

The second way operating powermode approximating, is encompassed by following. On a considered section $[a, b]$ the function $f(x)$ is changed with a polynomial

$$P_n(x) = a_0 + a_1x + a_2x^2 + \ldots + a_nx^n\,. \qquad (12)$$

The simulation of a polynomial implements a coherent DAC cascadely (fig. 4) [4, 5].

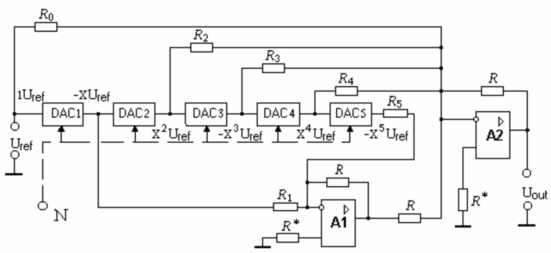

Fig. 4. The scheme of the functional digital-to-analog converter

In quality a DAC figured on the scheme, apply a multiplying DAC, which one can work with two-polar reference voltage. The output voltage such a DAC is determined under the formula:

$$U_{вых} = -U_{on}\cdot\frac{R_N}{R}\frac{N}{N_{max}}\,, \qquad (13)$$

where N - current digital code, which one varies within the limits from 0 up to $N_{max}-1$; $N_{max}=2^m$, m - digit capacity a DAC; R - resistance of a matrix of resistors; R_N - resistance of the resistor in a feedback loop OPA a DAC; U_{ref} - reference voltage. Attitude $A = \frac{R_N}{R}$ call as a scale factor or scaling ratio. It can be changed over a wide range, changing R_N.

In the scheme, reduced in a fig. 4, at submission on digital inputs a DAC of a code N, on an output 1-st DAC is reshaped pressure voltage, equal:

$$U_1 = -U_{ref}A\frac{N}{N_{max}}\,. \qquad (14)$$

This voltage is input for 2-nd DAC, and the voltage, on its output will be determined by a ratio:

$$U_2 = -U_1 A\frac{N}{N_{max}} = U_{ref}\left(-A\frac{N}{N_{max}}\right)^2. \quad (15)$$

Prolonging this series, for the k DAC is possible to record:

$$U_k = U_{ref}\left(-A\frac{N}{N_{max}}\right)^k. \quad (16)$$

The voltage from outputs a DAC through resistors R_1, R_2, ..., R_5 move on an input of the summator $A2$. For maintenance of the indispensable sign, the voltage from outputs the maiden and fifth DAC pass through the inverter $A1$. Follow-up on the summator through the resistor R_0 the reference voltage moves. On an output of the summator the voltage U_{out} is reshaped:

$$U_{out} = -\left(\frac{R}{R_0}U_{\cdot} + \frac{R}{R_1}U_1 + \frac{R}{R_2}U_2 + \right.$$
$$\left. + \frac{R}{R_3}U_3 + \frac{R}{R_4}U_4 + \frac{R}{R_5}U_5\right) \quad (17)$$

or, with allowance for of previous equation:

$$U_{out} = -U_{ref}\left(\frac{R}{R_0} - \frac{R}{R_1}\left(-A\frac{N}{N_{max}}\right) + \frac{R}{R_2}\left(-A\frac{N}{N_{max}}\right)^2 + \right.$$
$$\left. + \frac{R}{R_3}\left(-A\frac{N}{N_{max}}\right)^3 + \frac{R}{R_4}\left(-A\frac{N}{N_{max}}\right)^4 - \frac{R}{R_5}\left(-A\frac{N}{N_{max}}\right)^5\right) \quad (18)$$

If to designate

$$x = A\frac{N}{N_{max}}, \ |a_i| = \frac{R}{R_i}, \quad (19)$$

last equation will accept a kind:

$$U_{out} = -U_{ref}(a_0 + a_1x + a_2x^2 + \ldots + a_5x^5) \approx -U_{ref}f(x). \quad (20)$$

The factors of the polynomial sold by the given scheme, have following signs: $a_0>0$, $a_1>0$, $a_2>0$, $a_3<0$, $a_4>0$, $a_5>0$. If the factors have other signs, the scheme undergoes only minor alteration. Thus, we receive voltage output of a proportional approximable function $f(x)$.

It is necessary to indicate limitation, which one is superimposed by a cascading a DAC on range of change of a scale factor A. At submission on a chain a DAC of a code, close to N_{max}, voltage output k the DAC is proportional $U_{ref}(-A)^k$. For $A>1$ this value increases on geometrical progression and can result in to a saturation of operational amplifiers. Therefore is expedient to establish A equal 1 and to approximate a function recognizing that argument x changes from 0 up to 1.

From above mentioned follows, that for reproduction of a polynomial $P_n(x)$ of a degree n the cascadely live DAC is necessary n. The factors of a polynomial will be realised by selection of resistors R, R_0, R_1... R_k, and the signs of addends are established through the inverter.

That the error called by approximating (methodical error), was minimum, it is necessary in appropriate way to pick up factors of a polynomial.

We review three methods of calculus of these factors.

The most widespread method of approximating of a function $f(x)$ –its Taylor series expansion. In a general view this decomposing of a function $f(x)$ in neighborhood of a point x_0 implements under the formula:

$$f(x) = f(x_0) + \frac{f^{(1)}(x_0)}{1!}(x-x_0) +$$
$$+ \frac{f^{(2)}(x_0)}{2!}(x-x_0)^2 + \ldots + \frac{f^{(n)}(x_0)}{n!}(x-x_0)^n + \ldots$$

The decomposing of a function in a Taylor series is not alone. There is a capability to decompose a function in a series on generalized polynomials, for example, on polynomials of Legendre, Chebyshev, Jacobi, Hermite or Laguerre. Here we shall stay on polynomials of Chebyshev, as they give a best approximation.

For finding of factors C_k of decomposing on polynomials of Chebyshev $f(x) = \sum_{k=0}^{n} C_k T_k(x)$ is used the following formulas:

$$C_0 = \frac{1}{\pi}\int_{-1}^{1} f(x)\frac{dx}{\sqrt{1-x^2}} = \frac{1}{\pi}\int_0^{\pi} f(\cos\Theta)d\Theta \quad (21)$$

$$C_k = \frac{2}{\pi}\int_{-1}^{1}\frac{f(x)T_k(x)dx}{\sqrt{1-x^2}} = \frac{2}{\pi}\int_0^{\pi} f(\cos\Theta)\cos k\Theta d\Theta, \ k>0. \quad (22)$$

Substituting instead of degrees x of their expression through polynomials of Chebyshev, and then, resulting like terms at polynomials of one degree, we shall receive a required polynomial

$$P_n(x) = a_0 + a_1x + a_2x^2 + \ldots + a_nx^n. \quad (23)$$

To evaluate error produced by decomposing a functions $f(x)$ in a series on polynomials of Chebyshev in a general view, it is very difficult. But according to the general theorems of the theory of approximating this decomposing gives a best approximation from all possible. It is necessary also to mark, that the decomposing of a function $f(x)$ on polynomials of Chebyshev is possible only for functions having a continuous maiden derivative on a section [-1, 1]. This condition provides convergence of a series to a function $f(x)$. For all elementary functions this condition is executed.

The following method is based on the theory of interpolation. In this case build a polynomial, which one in $n+1$ given points x_0, x_1, . . . x_n, receives values $f(x_0)$, $f(x_1)$, . . . $f(x_n)$,and in remaining points of a section [a, b], inhering of a range of definition $f(x)$, approximately introduces a function $f(x)$ to some accuracy. For finding of factors of a polynomial a_k the set of equations is made:

$$a_0 + a_1 x_i + a_2 x_i^2 + \cdots + a_n x_i^n = f(x_i), \quad (i = 0, 1, 2, \ldots, n), \tag{24}$$

which one is easily decided, for example, method of Cramer.

4. THE CALIBRATOR OF A PHASE WITH A LINEAR TRANSFORMATION OF THE ESCAPE CODE IN A PHASE SHIFT

The widespread occurrence was received by calibrators of a phase, the principle of operation which one is encompassed by summation of two sinusoidal voltage shifted one concerning other on 90° . The control band of a phase shift thus makes *0-90°*, and its dilating up to *360°* implements the introducing of the commutator of reference voltages.

At regulation of a phase shift within the limits of 0-90° the voltage output of the calibrator of a phase is reshaped pursuant to relation

$$\dot{U}_{out} = k_1 U_{in} + jk_2 U_{in}, \tag{25}$$

where U_{in} - amplitude of reference voltages;
k_1 and k_2 - weighting coefficients.

Amplitude and phase of voltage output are connected to weighting coefficients k_1 and k_2 by ratio

$$U_{out} = U_{in}\sqrt{k_1^2 + k_2^2}; \tag{26}$$

$$\varphi = arctg\frac{k_2}{k_1}. \tag{27}$$

The calibrators of a phases controlled by a digital code, should provide a linear transformation of the escape code in a phase shift of voltage output. Besides in the majority of practical cases it is necessary to supply a constance of amplitude of voltage output in all control band of a phase shift. For fulfilment of these requirements the weighting coefficients k_1 and k_2 should be connected to the escape code by non-linear dependences.

In the calibrator of a phase, the skeleton diagram which one is submitted in a fig.. 5, the output voltage is the sum of two sinusoidal voltage $\dot{U}_S$ and $\dot{U}_C$, shifted one concerning other on 90°. They are reshaped of input sinusoidal voltage $\dot{U}_{in}$ by a circuit of cascadely live digital/analog converters DAC1 - DAC7 and inverting summators A1 - A4.

The transfer function an i- DAP $H_i(\theta)$ expresses by linear dependence

$$H_i(\theta) = b_i \alpha\theta, \tag{28}$$

where b_i - scale of transformation an *i*-DAC;

$$\theta = N/N_{max}; \alpha = \pi/2; \tag{29}$$

N and N_{max} - current and maximum values of the escape code.

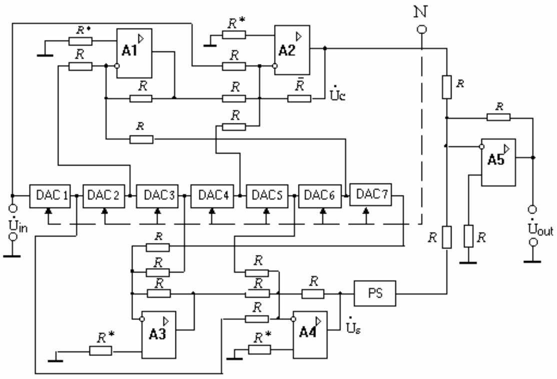

Fig. 5. The calibrator of a phase with a linear transformation of the control code in a phase shift

The phases DAC, used in the calibrator, have small output resistance, that enables their cascading without violation of normal operational mode each separately taken a DAC. Therefore transfer function *n* the cascadely live DAC with a sufficient accuracy rate is possible to write to a kind

$$H_n(\theta) = \prod_{i=1}^{n} H_i(\theta) = (\alpha\theta)^n \prod_{i=1}^{n} b_i. \tag{30}$$

On an input the first DAC being an input of the calibrator of a phase, from the external generator acts sinusoidal voltage U_{in}. Of it the circuit a cascadely live DAC reshapes voltage, the amplitudes U_n are connected which one to the escape code by a ratio

$$U_n = U_{in} H_n(\theta) = U_{in} x^n a_n, \tag{31}$$

where $a_n = \prod_{i=1}^{n} b_i; x = \alpha\theta.$

The voltage $\dot{U}_n$ from outputs a DAC and input voltage will be used for formation of two inphase voltage $\dot{U}_s$ and $\dot{U}_c$, and the voltage $\dot{U}_c$ is reshaped of input voltage and voltage from outputs a DAC with even numbers, and the voltage output an odd DAC will be used for formation of voltage $\dot{U}_s$. Supposing weighting coefficients of summarized voltage $\dot{U}_n$ and U_{in} by equal unit and allowing padding inverting of the conforming voltage by summators *A1* and *A3,* the relations of a range of stress $\dot{U}_s$ and $\dot{U}_c$ from the escape code pursuant to (4.5) can be presented by the way

$$U_s = -U_{in}(a_1 x - a_3 x^3 + a_5 x^5 - a_7 x^7) = U_{in} k_2; \tag{32}$$

$$U_c = -U_{in}(1 - a_2 x^2 + a_4 x^4 - a_6 x^6) = U_{in} k_1. \tag{33}$$

Thus, the ranges of stress $\dot{U}_s$ and $\dot{U}_c$ are connected to amplitude of sinusoidal voltage U_{in}, calibrator, acting on an input, of a phase, factors k_1 and k_2, which one in turn have non-linear dependence from the escape code.

$$k_1 = 1 - a_2x^2 + a_4x^4 - a_6x^6; \qquad (34)$$

$$k_2 = a_1x - a_3x^3 + a_5x^5 - a_7x^7. \qquad (35)$$

To receive a phase shift 90° between voltage $\dot{U}_s$ and $\dot{U}_c$, one of them, $\dot{U}_s$, moves on an input of the phase change PC, as a result of which on inputs of the summator A5 the voltage U_c and jU_s act. The voltage output of the summator A5, being voltage output of the calibrator of a phase, is described by a ratio (25).

The phase $\varphi_0 = x = \alpha\theta$ is determined by the escape code, and the amplitude of voltage output depends only on amplitude of input voltage U_{in} (in the reviewed case $U_{out}=U_{in}$).

The difference between computational both set values of a phase and amplitude of voltage output will be the less, than the functional connections *cosx* and *sinx* are more precisely modelled.

The functions *cosx* and *sinx* can be submitted by the way sums of a power series, and the more precisely, than more members of a power series thus less range of change of argument *x* will be used and than. When preset number of the members of a series and range of change of a variable *x*, the problem of minimization of error of simulation is reduced to precise calculus of factors at the summarized members of a series.

In tab. 1 the values of factors a_n, counted for three cases are adduced, when a maximum degree *x* and number a DAC in the scheme of the calibrator of a phase are accordingly peer seven, six and.

Methodical phase error and instability of amplitude of voltage output are determined by ratio

$$\Delta\varphi = \alpha\theta - arctg\frac{k_2}{k_1}; \qquad (36)$$

$$\delta U_{out} = \frac{U_{in} - U_{out}}{U_{in}} = 1 - \sqrt{k_1^2 + k_2^2} \qquad (37).$$

The diagrams of these relations are adduced in a fig. 6. (a, b) (where *m* - number a DAC in the scheme), and their maximum ratings are indicated in tab. 2.

TABLE 2. Values of coefficients and the maximal values of errors for m=5,6,7

Coefficients a_n		$\delta\ U_{out}$, %	$\Delta\ \varphi$, °
$a_1 = 1{,}000$ $a_2 = 0{,}4999$ $a_3 = 0{,}1666$ $a_4 = 0{,}0416$	$a_5 = 0{,}0083$ $a_6 = 0{,}0013$ $a_7 = 0{,}0002$	0,005	0,003
$a_1 = 1{,}000$ $a_2 = 0{,}4999$ $a_3 = 0{,}1661$	$a_4 = 0{,}0416$ $a_5 = 0{,}0076$ $a_6 = 0{,}0013$	0,016	0,005
$a_1 = 1{,}000$ $a_2 = 0{,}4967$ $a_3 = 0{,}16605$	$a_4 = 0{,}03705$ $a_5 = 0{,}00761$	0,06	0,06

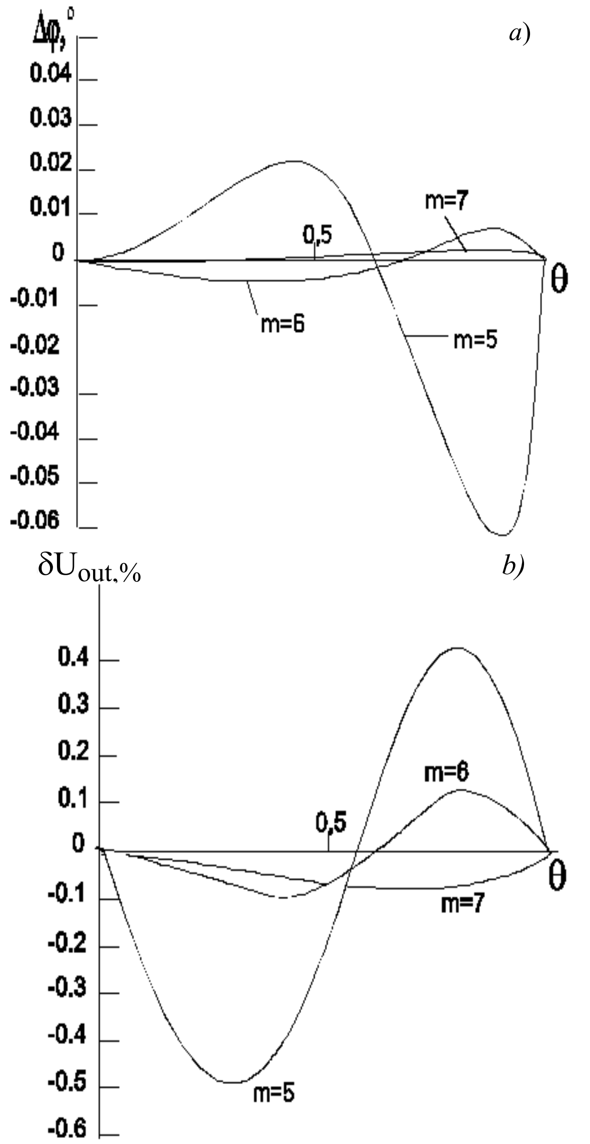

Fig. 6. Relations of error of a phase shift $\Delta\varphi$ (*a*) and module δU_{out} of voltage of an output (б)

The analysis of the obtained outcomes demonstrates, that already at usage six DAC is possible is to constructed by the calibrator of a phase with the high metrology characteristics. The error of the calibrator of a phase basically will be determined by error of set-up a DAC on a given scale of transformation, spurious phase shifts in a DAC and summators on high frequencies, accuracy of maintenance of a phase shift 90° .

The experimental check of a breadboard of the calibrator of a phase has shown, that in range of 30 Hz - 20 kHz the intrinsic error does not exceed 0,1° and reaches maximum rating on a higher frequency of the indicated range (test of a breadboard conducted with usage of the calibrator of a phase, elapsed metrology certification in Russian science-researching institute of metrology).

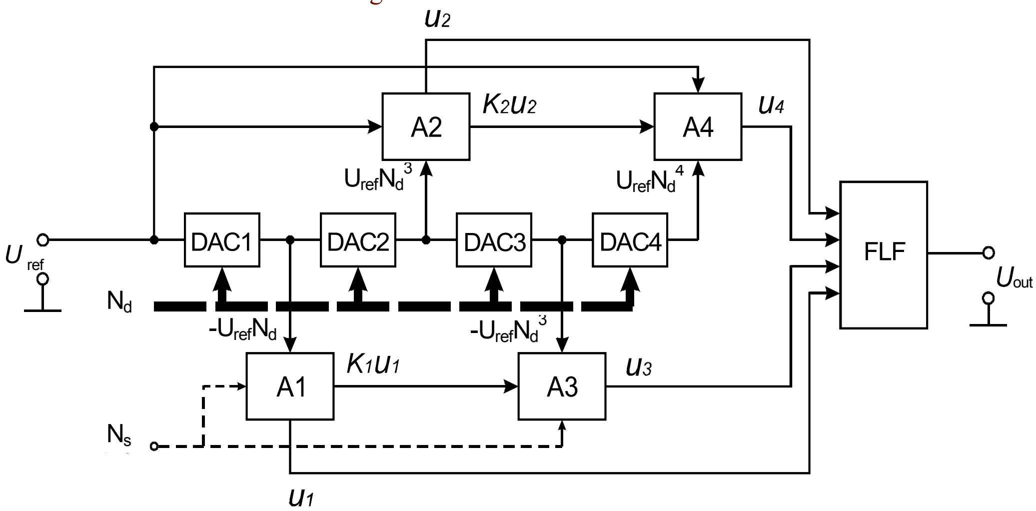

Fig. 7. The functional circuit of a frequency and phase multiplier

5. THE BROADBAND DIGITAL-CONTROLLED MULTIPLIER OF FREQUENCY AND PHASE

Nonlinear resonant frequency multipliers which are characterized by a narrow working strip of frequencies, have are widely used. It is possible to expand a range of frequencies using the transformation of a harmonious signal determined by Chebyshev polynomial $T_k(x)=cos(k \times arccos(x))$ which provide on an output only the given harmonic . This principle take as a basis of a considered frequency multiplier. Reproduction of a polynomial of Chebyshev is made by the indicated above way.

The functional scheme of the broadband frequency and phase multiplier are represented in a fig. 7. The circuit consists of digital-to-analog converters (DAC), scale converters A1, A2, A3, A4 and the analog filter of low frequencies (FLF). The basis of the circuit is made with a chain in cascade connected digital-to-analog converters DAC1-DAC4 modelling power dependencies of a output voltage from controlling code . Scale converters produce summation of output voltage of DAC with Chebyshev coefficients The filter of low frequencies is executed broadband and serves for allocation of a necessary harmonic.

Application of an offered method in comparison with known ways allows essentially to increase speed and accuracy Chebyshev transformations of an input signal. It, in turn, promotes expansion of a working range of frequencies.

6. CONCLUSION

In this work two universal methods of construction functional DAC are considered. The first one based on partial-linear approximation nonlinear function. The second one uses polinominal approximation and consecutive inclusion of multiplying DAC. Using as the functions of transformation of functional DAC sin(x) and cos(x) it is possible to construct the digital-controlled calibrators of a phase. In the work are considered scheme realizations for each of the specified methods. Methodical errors of functional DAC and calibrators of a phase are considered, depending on a degree of approaching to approximated functions. On the basis of functional DAC the broadband multiplier of frequency and phase is developed.

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