

Development of a digital signal processing model using a frequency synthesizer and synthesis of quadrature conversion circuits

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Abstract. An important issue of improving the quality of the functioning of measuring systems is the improvement of the conversion methods and the algorithmization of the processes for eliminating various noises and interferences that arise in the tasks of analog-to-digital information conversion. The article analyzes the algorithms for assessing the accuracy of digital signal processing (DSP) in a complex of issues due to the development of digital and digital-analog element base, as well as the development of methods and algorithms for processing and generating signals.

1 Introduction

Modern systems use various methods of digital information processing and circuit solutions to replace the classical adaptive structures of the organization of interface communications. At the same time, new requirements arise for designing noise-resistant interface systems based on microprocessor controllers for creating information processing networks.

For the dynamic expansion of the ranges of analog modules, in particular in their composition of the compensation of the unbalance of quadrature mixers (CS), we consider a compensation algorithm based on the calculation of the parameters of the output signals of the CS by complex digital spectral analysis.

2 Method

Noting that the parameters a and b at the initial zero frequencies are equal to the values of the transformations of the corresponding real signals at the output of the quadrature mixer. To determine the parameters of the discrepancy with the specified algorithm, it is necessary to provide a reference signal in the form of a harmonic oscillation[1].

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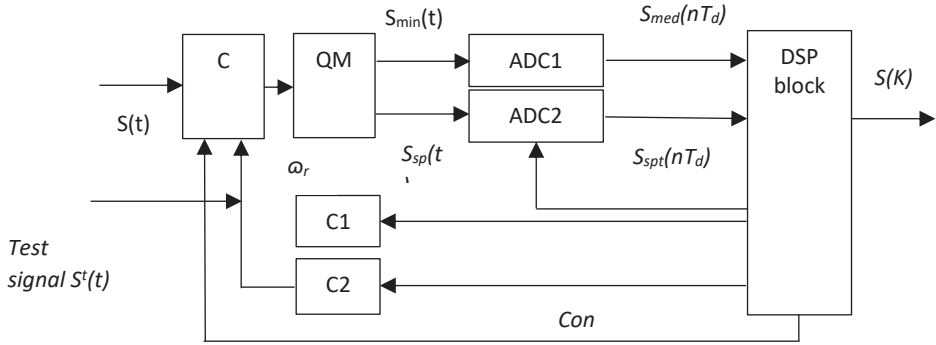


Fig.1. Functional diagram of a quadrature transformation module with perturbation compensation. C - signal switching devices; QM - quadrature mixer; C1 - frequency synthesis devices for the receiving device; C2 - frequency synthesis device for test signal.

The control signal generated by the DSP unit reformulates the $S(t)$ input signal received by the signal switch in accordance with the parameters of the reference test signal generated by the C2 synthesizer. In-phase and quadrature components of the displacement results are calculated in the DSP unit. The obtained results of processing the displacements with the reference and the parameters of the discrepancy of the amplitude-frequency domain are calculated. These parametric discrepancies are taken into account when processing the real signal $S(t)$. The procedure carried out, called CS calibration, is performed cyclically.

The functional diagram of the device of the quadrature conversion module with compensation of disturbances is shown in Fig. 1 [2,4,7,9].

In-phase and quadrature components of discrete signals corresponding to the test signal at the output of a real CS are described by the following expressions;

$$S_{cpu}^T(nT_d) = S_{cpd}^T(nT_d) + \eta_{cd} [S_{cpd}^T(nT_d)] \quad (1)$$

$$S_{spu}^T(nT_d) = S_{spd}^T(nT_d) + \eta_{cd} [S_{spd}^T(nT_d)]$$

$$S_{cpd}^T(nT_d) = a + U_0 \cos[\Delta\omega nT_d + \varphi_0] \quad (2)$$

$$S_{spd}^T(nT_d) = b + (1 + \Delta k)U_0 \sin[\Delta\omega nT_d + \varphi_0 + \Delta\varphi]$$

At the same time, it is necessary to take into account the resulting quantization noise in the common-mode part $\eta_{cd} [S_{cpd}^T(nT_d)]$ and also, it is necessary to take into account the noise in the quadrature mixer $\eta_{cd} [S_{spd}^T(nT_d)]$.

Having normalized in the time domain of the complex discrete Fourier transform of the reduced (1 and 2) functions, we obtain [3,10]:

$$\bar{S}_{c(s)pu}(k) = \frac{1}{N} \sum_{n=0}^N S_{c(s)pu}^T(n) e^{j\frac{2\pi}{N}nk} \quad (3)$$

Assuming if in a quadrature transformation, there are no unbalances in the results of the transformation. then the real values of the common - mode and quadrature components of the signals at the outputs of the CS can be determined[11,12]:

$$S_{cud}^T(nT_d) = U_0 \cos[\Delta\omega nT_d + \varphi_0] = S_{c\varnothing d}(n)$$

$$S_{sud}^T(nT_d) = U_0 \sin[\Delta\omega nT_d + \varphi_0] = S_{s\varnothing d}(n) \tag{4}$$

If we accept the signal in (4) as a reference, and then the discrete Fourier transforms of this signal will also turn out to be a reference:

$$\bar{S}_{c(s)\varnothing d}(k) = \frac{1}{N} \sum_{n=0}^{N-1} S_{c(s)\varnothing d}(n) e^{j\frac{2\pi}{N}nk} \tag{5}$$

The above expressions do not take into account quantization noise, and the selected frequency band must be wide enough to analyze real signals. Further, comparing the natural decomposition coefficients previously obtained by formulas (3) and (5), we obtain the actual real deviation parameters[13,16,17],

a^* , b^* , $\Delta\varphi^*(i)$, $\Delta k^*(i)$ at the i -th frequency point:

$$\dot{a} = |\bar{S}_{cpu}(0)| - |\bar{S}_{c\varnothing d}(0)| \tag{6}$$

$$\dot{b} = |\bar{S}_{spu}(0)| - |\bar{S}_{s\varnothing d}(0)| \tag{7}$$

$$\Delta\dot{\varphi}(i) = \arg \bar{S}_{spu}(i) - \arg \bar{S}_{s\varnothing d}(i) \tag{8}$$

$$\Delta\dot{k}(i) = \frac{\bar{S}_{spu}(i)}{\bar{S}_{s\varnothing d}(i)} - 1 \tag{9}$$

Then, accordingly, the amplitude spectra of the common-mode and quadrature components in the common-mode channel of the complex area of the test and reference signal can be taken away, as shown in Fig.2 a, b and similar spectra for the test, as well as the reference signals, respectively, are shown in Fig.3 a, b[14,15,21]

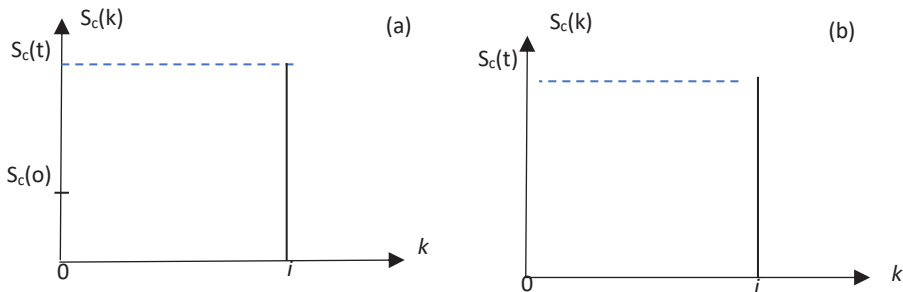


Fig.2 (a) Spectra for the test signal in the common-mode region; b) the spectra of the reference signal in the common-mode part of the harmonic.

It is recommended to always take into account the influence of quantization noise. In particular, the presence of quantization noise of the specified test signals at the input of the

considered quadrature mixer may somewhat limit the accuracy of the evaluation of expressions (6-9). Since the algorithm for setting the reference signal is modeled taking into account the expected disturbances, it can be assumed that it does not contain distortions.

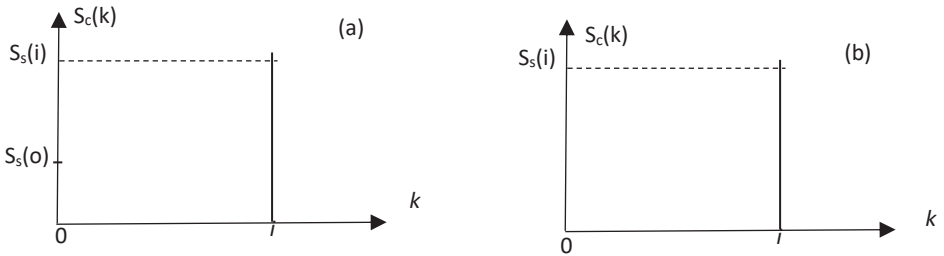


Fig. 3. (a) the spectrum corresponding to the test signal; b) the spectra of the reference signal in the quadrature space of oscillations.

We assume that if the output of the CS signal spectra is in an ideal form[18,22,23,25]:

$$\bar{S}_{c(s)pu}(k) = \frac{1}{N} \sum_{n=0}^{N-1} S_{c(s)pu}^T(n) e^{j\frac{2\pi}{N}nk} \quad (10)$$

It is possible to imagine a deviation compensation algorithm for a test signal:

$$\bar{S}_{ck}(k) = \bar{S}_{cpu}(k) e^{j\varphi_{cpu}(k)} - a(0) \quad (11)$$

$$\bar{S}_{sk}(k) = S_{spp}(k) [1 - \Delta k^*] e^{j[\varphi_{spp}(k) - \Delta\varphi^*(k)]} - b(0) \quad (12)$$

In the future, according to the specified algorithm, the switch K is switched to the real specified signal under study, i.e., according to the results of the obtained deviation parameters (10. and 11), the results of the study are corrected. To develop a system implementing digital filtering algorithms, various circuit solutions were used: digital control devices, operational amplifiers with a different set of frequency characteristics and signal conversion coefficients. In particular, the Nyquist digital quadrature transformation method was widely used in practical research on the dynamic expansion of the ranges of analog modules[19,20,26].

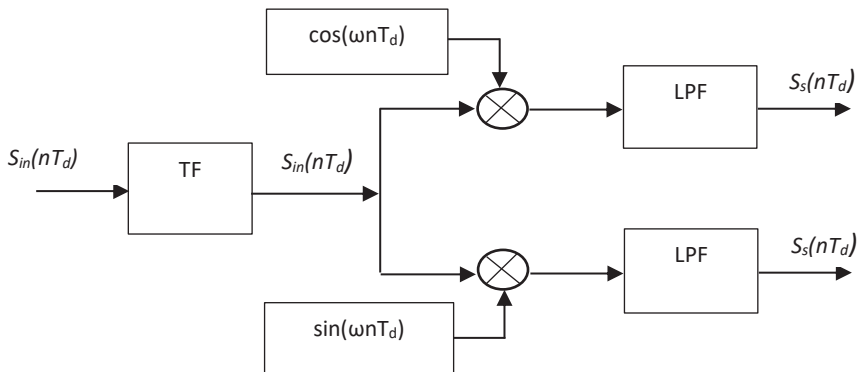


Fig.4. Block representation of quadrature signal conversion.

The main essence of the study is the preliminary restriction with a bandpass filter (PF) in

the Nyquist frequency bandpass filter (PF) and the further distribution of the signal spectrum on common-mode oscillations $S_c(T_d)$, as well as quadrature components $S_s(nT_d)$. To determine these two components: in-phase $S_c(nT_d)$ and quadrature $S_s(nT_d)$ the input signal is multiplied by the reference signals $\cos(\omega nT_d)$ and $\sin(\omega nT_d)$

The implementing scheme of the quadrature converter is shown in Fig. 4[24,27].

Let's consider analytical solutions for the schematic representation of quadrature signal conversion. Let the input signal of the quadrature transformation be represented by the following expression $S_c(nT_d) = A(nT_d)\cos(\omega_0 nT_d + \varphi_0)$ and accordingly, the reference signals are given:

$$S_{op_cos}(nT_d) = \cos(\omega_0 nT_d + \varphi_0);$$

$$S_{op_sin}(nT_d) = \sin(\omega_0 nT_d + \varphi_0)$$

then the result of multiplying the input signal with the reference signals can be represented as

$$\begin{aligned} S_c(nT_d) \cdot S_{op_cos}(nT_d) &= A(nT_d)\cos(\omega_0 nT_d + \varphi_0)\cos(\omega_0 nT_d) = \\ &= \frac{A(nT_d)}{2}\cos(2\omega_0 nT_d + \varphi_0) + \frac{A(nT_d)}{2}\cos(\varphi_0) \\ S_c(nT_d) \cdot S_{op_sin}(nT_d) &= A(nT_d)\cos(\omega_0 nT_d + \varphi_0)\sin(\omega_0 nT_d) = \\ &= \frac{A(nT_d)}{2}\sin(2\omega_0 nT_d + \varphi_0) + \frac{A(nT_d)}{2}\sin(\varphi_0) \end{aligned} \tag{13}$$

The quadrature components of the signal in a certain frequency spectrum will be carried out, respectively, at the difference and total frequency limits. To remove the total frequency components, low-pass filters (LPF) are used. In the final result we have:

$$S_c(nT_d) = \frac{A(nT_d)}{2}\cos(\varphi_0) \text{ and } S_s(nT_d) = \frac{A(nT_d)}{2}\sin(\varphi_0)$$

The structural foundations of the LPF are filters with a finite impulse response (FIR). Then the amplitude characteristics of the input signal is found by the formula:

$$2 \cdot \sqrt{S_s(nT_d)^2 + S_c(nT_d)^2} = 2 \sqrt{\frac{A^2(nT_d)}{4}(\cos^2(\varphi_0) + \sin^2(\varphi_0))} = 2 \sqrt{\frac{A^2(nT_d)}{4}} = A(nT_d) \tag{14}$$

Thus, when the common-mode and quadrature components are known, it will be possible to determine the initial phase of a given input signal:

$$\arctan\left(\frac{S_c(nT_d)}{S_s(nT_d)}\right) = \arctan\left(\frac{\frac{A(nT_d)}{2}\sin(\varphi_0)}{\frac{A(nT_d)}{2}\cos(\varphi_0)}\right) = \arctan(\tan(\varphi_0)) = \varphi_0 \tag{15}$$

3 Conclusion

Thus, it can be concluded that the conversion of an analog signal into a discrete form with quantization is one of the most common types of tasks in automated information processing. At the same time, the choice of sampling frequency in them is determined by the range of the

signal spectrum.

If, for example, the spectrum of the signal suitable for conversion is limited by the frequency f_{max} , to restore the signal without loss of information content, the sampling rate should be 2 times higher. In other words, if the highest frequency harmonic of the input analog signal has a period T , then for this harmonic period it should correspond to two sampling counts.

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