Estimations of the Signal Information Parameters in Radio Engineering Systems

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Abstract. Radio engineering systems are used to monitor the environment and operate in difficult conditions with a large number of interferences of natural and artificial origin. Contamination of the frequency space by radiation sources places high demands on methods and means of signal processing of such systems, in particular methods of estimation of information parameters of signals, their amplitudes, phases, frequencies, modulation depths. A new approach is proposed to estimate the phase difference in VOR systems and the difference in the modulation depth of the radar signals of ground beacons of the ILS instrumental landing system using the maximum likelihood method, and based on the Fourier transform. One of the applications of the considered method in onboard computers for estimating the difference in modulation depth is shown. The stability conditions for the optimal estimation of the harmonic signal frequency to action and the parameters of impulse noise are analytically determined, which was confirmed by statistical modeling.

Keywords: Estimation Of A Signal Parameters, Phase, Frequency, Impulse Noise, Robust Method, Navigation Systems.

1 Introduction

An increase in air traffic intensity, a high saturation of aviation with radio engineering means of communication, control, navigation and landing, far and near radar, an increase in the information saturation of the radio frequency spectrum as a whole, and an increase in industrial interference leads to a constant complication of the interference situation in airport areas. Under these conditions, the noise immunity of the onboard equipment of short-range navigation and landing systems is deteriorating.

The most of radio engineering systems signal processing methods are based on the Gaussian model, for which a number of optimal solutions are obtained. However, in most cases signal and interference distributions are different from the Gaussian model. This leads to a significant decrease in the real performance of systems compared to theoretical.

Of particular importance is the efficiency and noise immunity of signal processing channels in flight radio systems [8-10], in particular in short-range navigation systems operating in radio frequency contaminated conditions [4].

The article solves the problem of synthesizing new post-detector algorithms for evaluating the information parameters of radio signals in short-range navigation systems (navigation beacons VOR, ILS) in digital channels of radio receivers. The algorithms are obtained by statistical synthesis using the maximum likelihood method and provide high noise immunity and accuracy.

2 Digital signal processing in on-board equipment of navigation system VOR

2.1 Features of the beacon VOR

Digital signal processing (DSP) in receivers VOR is considered as their non-linear transformation in order to measure the phase difference of the receiver detected signals. The phase estimation algorithm is further synthesized using the statistical maximum likelihood method, which provides good noise immunity and accuracy at low signal-to-noise power ratios (SNR).

The information signal of the VOR receiver at the output of the detector of the high-frequency amplifier is the sum of the signals of the channel of the reference phase (RPH) and the channel of the variable phase (VPH) [4]. The signal of the RPH channel is a frequency-modulated oscillation of the form:

$$U_{oph}(t) = U^0_{oph} \sin(2\pi f_0 t + \frac{f_{dev}}{f_0} \cos\left(f_{30} t + \varphi^0_{oph}\right) + \varphi) \,, \qquad (1)$$

where U^0_{oph} - amplitude of FM oscillations; f_0- subcarrier frequency; φ_0- its initial phase; f_{dev}- subcarrier deviation frequency; f_{30}- frequency modulating oscillations; φ - its initial phase.

The signal of the variable phase (VPH) channel is harmonic oscillation with amplitude U^0_{vph}, initial phase φ^0_{vph} and frequency f_{30} equal to:

$$U_{vph}(t) = U^0_{vph} \sin(2\pi f_{30} t + \varphi^0_{vph}) \,, \qquad (2)$$

Useful information (azimuth to the beacon), which must be detected by processing these signals, is enclosed in the difference $\Delta\varphi$ between the current phase of the modulating oscillation of the RPH channel $\varphi_{oph}(t) = f_{30} t + \varphi^0_{oph}$ and the phase of the VPH channel $\varphi_{vph}(t) = f_{30} t + \varphi^0_{vph}$, then:

$$\Delta\varphi = \varphi_{oph}(t) - \varphi_{vph}(t) \,, \qquad (3)$$

The nominal frequency values are as follows: f_{30} = 30Hz, f_0 = 9960Hz, f_{dev} = 480Hz.

According to the recommendations of the international documents "ARINC-711" and the requirements of the "Airworthiness Norms of airplanes NLGS-3", the error of azimuth measurement by VOR equipment at nominal values of the input signal with a probability of 0.95% should not exceed 0.3 ° [4].

In the presence of interfering factors, this error may be 0.5 °. The following factors are related to interfering factors: drift of a frequency 30 Hz, which is not more than 1.5% of the nominal value (f_{30}= (30.00 ± 0.45) Hz); departure of the subcarrier frequency from the nominal value, with f_0 = (9960 ± 100) Hz; drift of frequency deviation from the nominal value, with f_{dev} = (480 ± 30) Hz; the interference, consisting of spectral components in the range 0 - 20 kHz with a total amplitude, does not exceed the amplitude of the useful signals.

The block diagram of the digital signal processing path implemented in the on-board navigation equipment VOR-85-01 is shown in Fig. 1. The demodulation of FM oscillation occurs on the basis of fixing the moments of its intersection of the zero level. The final processing is carried out by the calculator

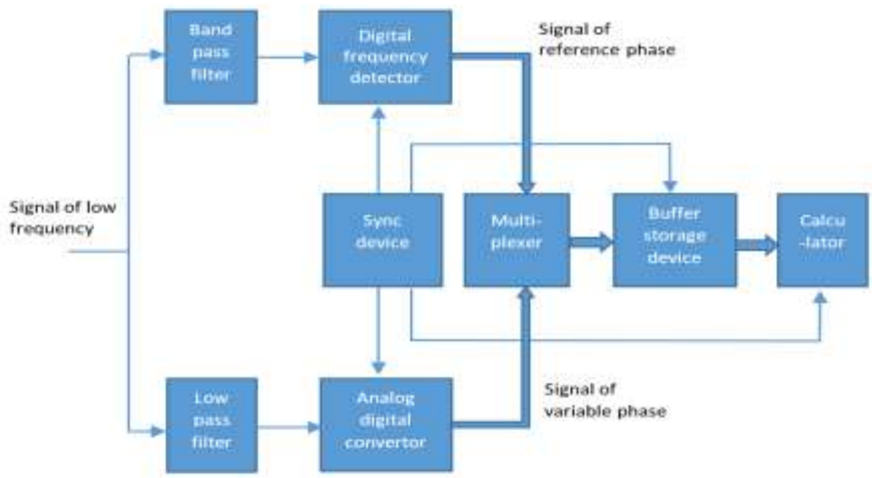


Fig. 1. The block diagram of the digital signal processing path implemented in the on-board navigation equipment VOR-85-01.

It is proposed to use the maximum likelihood method for calculating the phase difference with the solution of the likelihood equation according to the Newton-Raphson algorithm, which has high noise immunity and estimation accuracy.

2.2 Synthesis of the algorithm for estimating the phase difference by the maximum likelihood method

The statistical model of the input signal (sample of values) (x_1, ..., x_2) is taken as an additive mixture of Gaussian noise with dispersion (σ^2) and a useful harmonic signal with known amplitude (U) and frequency ($\omega_0 = 2*pi*f_0$). It is necessary to evaluate the phase of the input signal from a sample of values.

The likelihood function is generally written as

$$f(x_1,..x_n,t_1,..t_n) = \frac{1}{(\sqrt{2\pi\sigma^2})^n} \exp\left\{-\sum_{i=1}^{n}\left[x_i - U\cos(\omega_0 t_i + \varphi)\right]/2\sigma^2\right\} \quad (4)$$

It is required to determine the coordinate of the function maximum (4) by the parameter φ.

After differentiation, we obtain the likelihood equation

$$\frac{\partial}{\partial\varphi}\ln f(x_1,..x_n,t_1,..t_n/\varphi,U,\omega_0) = \\ = \frac{1}{\sigma^2}\sum_{i=1}^{n}\left[x_i - U\cos(\omega_0 t_i + \varphi)\right]\sin(\omega_0 t_i + \varphi)U = 0 \quad (4)$$

Solving this equation using the Newton – Raphson method, we obtain

$$\varphi_{k+1} = \varphi_k - \frac{\sum_{i=1}^{n}\left[x_i - U\cos(\omega_0 t_i + \varphi)\right]\sin(\omega_0 t_i + \varphi_k)}{\frac{\partial}{\partial\varphi}\left[\sum_{i=1}^{n}\left[x_i - U\cos(\omega_0 t_i + \varphi)\right]\sin(\omega_0 t_i + \varphi_k)\right]\Big|_{\varphi=\varphi_k}} \quad (5)$$

Transforming this expression, we obtain an iterative rule of the phase estimation algorithm.

$$\varphi_{k+1} = \varphi_k - \frac{\sum_{i=1}^{n}\left[x_i - U\cos(\omega_0 t_i + \varphi)\right]\sin(\omega_0 t_i + \varphi_k)}{\sum_{i=1}^{n}\left[x_i - \cos(\omega_0 t_i + \varphi)\right] - U\cos(2(\omega_0 t_i + \varphi_k))} \quad (6)$$

To determine the phase difference, it's necessary to calculate the formula (6) for each signal (for signal of the reference phase and for signal of the variable phase) and calculate the difference between them.

2.2 The research of noise immunity of the Newton-Raphson method

To evaluate the noise immunity of the Newton – Raphson method, a software model was developed where the input signal is an additive mixture of the harmonic signal *S(t)* and white noise η

$$x(t) = S(t) + \eta = U\cdot\cos(\omega_0 t + \varphi) + \eta \quad (7)$$

where U - amplitude; $\omega_0 = 2\cdot\pi\cdot f_0$- frequency; φ- its initial phase; .

The mathematical model of white Gaussian noise was modeled as

$$\eta = \sqrt{-2\cdot D\cdot\ln(RND(1))}\cdot\cos(2\pi\cdot RND(1)) + M, \quad (8)$$

where M – mathematical expectation, D – dispersion, RND (1) - values of the random number generator with uniform distribution on the interval (0 ... 1).

Results of modelling for various levels of the Signal-to-Noise Ratio (SNR) are shown in Fig.2.

Also, the calculations were performed for the input signal with a different initial phase shift. For example, with φ_0 = 60.0 ° and a noise variance of D = 0.3, a result of 60.90783° was obtained at the first iteration. Note that none of the analog phase capture methods can provide such accuracy with such a high noise level.

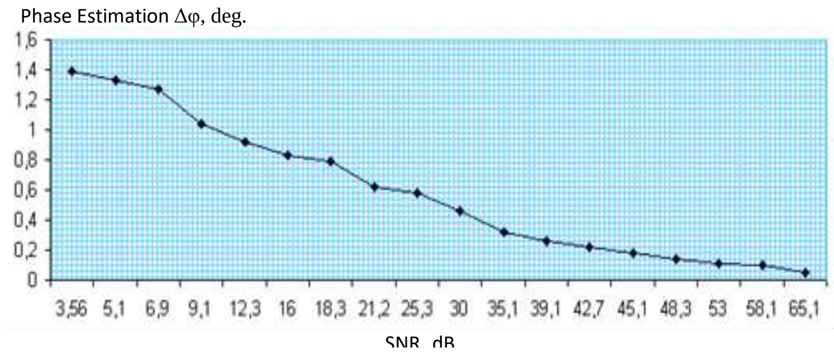


Fig. 2. Results of modelling.

3 Synthesis Of A Frequency Estimation Algorithm In A Frequency Detector

3.1 The Need To Take Into Account The Impact Of Pulsed Interference On Frequency Estimation

The traditional solution for a digital frequency detector includes an algorithm for estimating the signal frequency by counting zero crossings for a certain period of time. For its simplicity, this method has insufficient noise immunity and is inoperative at low signal-to-noise ratios.

Improving the accuracy and noise immunity of a digital frequency detector can be obtained by statistical synthesis of the optimal algorithm for estimating the frequency of a harmonic signal

In the actual operating conditions of the radio technic systems, one of the important problems is to ensure the stability of algorithms for estimating the information parameters of useful signals to the effect of chaotic impulse interference (HII) [1], since their appearance can lead to complete distortion of the calculation results.

This applies in particular to the quasi-optimal frequency estimation (QOF) algorithm of the harmonic signal (hereinafter referred to as the "signal"), which is proposed in [6]. This paper uses an equidistant model of the signal state in recurrent form

$s_i = \alpha s_{i-1} - s_{i-2}, i = \overline{1,n}$ (n is the sample size) in which the parameter $\alpha = 2\cos(2\pi f\tau)$, $\tau = t_i - t_{i-1}$ associated with the instantaneous signal frequency f, and the observational model based on additive Gaussian noise (hereinafter referred to as "noise") η_i is presented as

$$x_i = s_i + \eta_i,\ i = \overline{1,n} \tag{9}$$

Synthesized by the maximum likelihood method, the algorithm of frequency estimation is based on the solution of a quadratic equation $\alpha^2 - B\alpha - 2 = 0$, in which only the coefficient B depends on the input counts. This coefficient is a statistic of the form

$$B\left(x_1, ..., x_N\right) = \sum_{i=3}^{N}\left[\left(x_i + x_{i\text{-}2}\right)^2 - 2x_{i\text{-}1}^2\right] / \sum_{i=3}^{N}\left(x_i x_{i\text{-}1} + x_{i\text{-}2} x_{i-1}\right) \tag{10}$$

By one of the roots of the equation (10)

$$\alpha^* = \frac{B}{2} + \sqrt{\frac{B^2}{4} + 2} \tag{11}$$

The estimated frequency of the signal is determined: $\gamma^* = \arccos(\alpha^* / 2)$. The absolute radial frequency estimate of a signal $\omega^* = 2\pi f$ is uniquely related to its normalized value as

$$\omega^* = \frac{\gamma^*}{\tau} \tag{12}$$

where is the sampling interval, so in the following we will consider only the normalized frequency, for which in this section we use the abbreviated term "frequency".

The frequency detector consists of an analog-to-discrete converter (ADC), a sliding window SW of size n, in which a bundle of signal is formed, an instantaneous frequency value ω^* calculator and an instantaneous and carrier frequency ω_0 difference calculator - $\Delta\omega^*$.

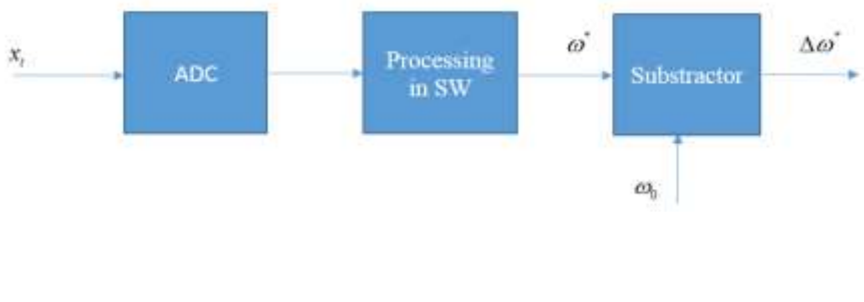


Fig. 3. Block diagram of frequency detector based on frequency estimator in sliding window

In Fig. results of FM signal processing by frequency detector based on frequency estimator (10-12) and frequency estimator based on counting of zero-level crossings in a sliding window of size n = 200 are presented. Carrier frequency 1MHz, frequency deviation 486 Hz. The upper graph shows the modulating signal, the average signal of the frequency estimator (*) and the lower signal of the estimation-cha based on the counter 0-sections. Noiseless situation was simulated. The estimation error for the algorithm (10-12) was 1.8%, for the 0-crossing algorithm - 10%.

In Fig. results of operation of frequency detectors in the situation with noise SNR = 11dB are given. The estimation error in this case for algorithm (10-12) was 7%, for algorithm 0-sections - 22%.

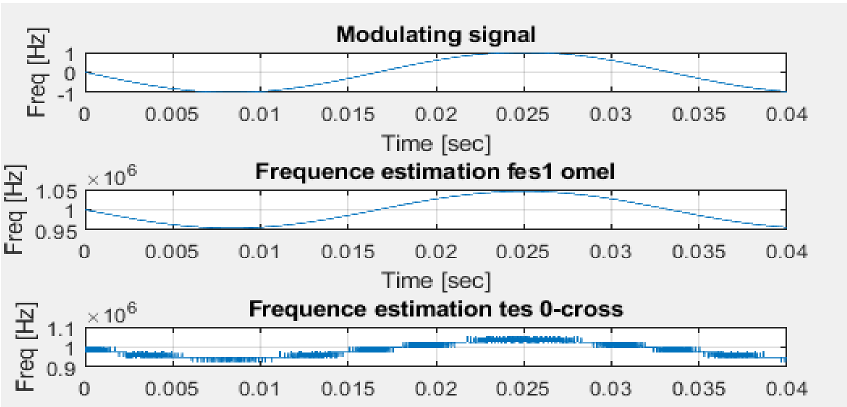


Fig. 4. Simulation of frequency detectors based on the frequency estimator and 0-section counter in a sliding window (n = 200) without noise

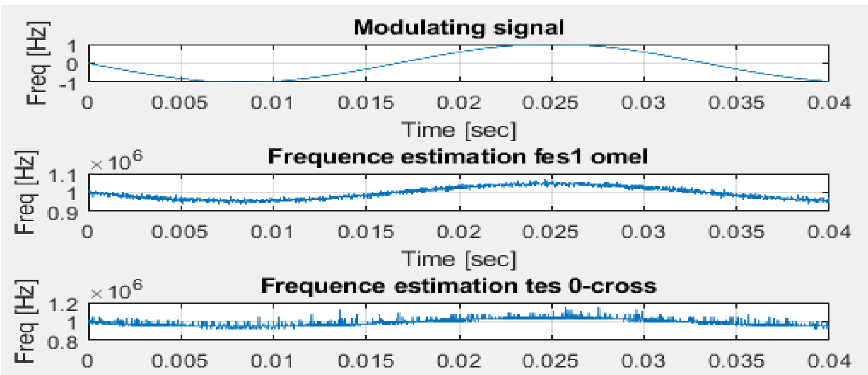


Fig. 5. Simulation of frequency detectors based on frequency estimator and 0-section counter in sliding window (n = 200) with noise. SNR = 11dB

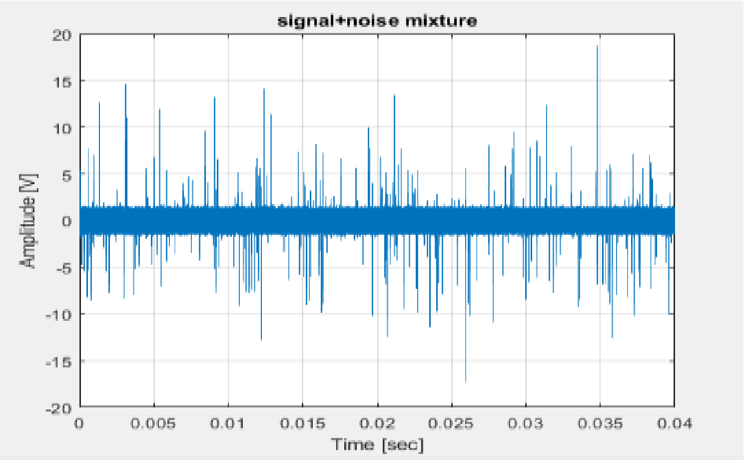


Fig. 6. Implementation of a mixture of FM signal, Gaussian noise and chaotic impulse interference. SNR = 11dB, RMS HIP = 5V, probability of HIP action p = 0.001.

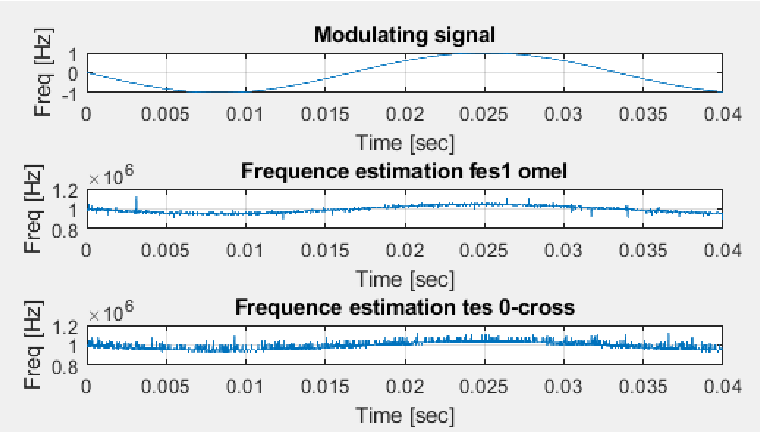


Fig. 7. Modeling of frequency detectors based on frequency estimator and 0-section counter in sliding window (n = 200) with noise. SNR = 11dB and impulse interference with parameters: RMS HIP = 5V, probability of HIP action p = 0.001.

The influence of chaotic impulse interference on the efficiency of signal frequency estimation is investigated. Under these conditions, the observation model looks like:

$$x_i = s_i + \eta_i + \vartheta_i,\ i = \overline{1,n} \tag{13}$$

Impulse interference ϑ_i is distributed by law

$$f_p(x) = (1-p)\delta(x) + \frac{p}{\sqrt{2\pi}\sigma_\vartheta}\exp\left(-\frac{x^2}{2\sigma_\vartheta^2}\right),$$

where σ_ϑ^2 is the conditional variance of the impulse noise, p is the probability of occurrence of the impulse noise.

The analysis of the results of simulation of frequency detectors shows the significant advantages of the proposed FD over the FD based on the 0-section counter.

4 Algorithms For Estimating The Modulation Coefficient Of Signals In Navigation-Landing Systems

4.1 Signals Of Instrumental Landing Systems

The position of the aircraft relative to a given descent path is determined by the angular deviations $\Delta\varphi$, $\Delta\theta$ in the horizontal and vertical planes. Angular deviations are measured relative to the course and glide path planes, the intersection of which gives a predetermined path, using the dependence of the spatial modulation depth coefficient. The spatial dependence of the modulation depth on the angles φ, θ is set by course and glide path beacons with the corresponding radiation pattern forms.

The task of the onboard radio is to isolate the signal, filter it and determine the modulation coefficients.

A signal is generated in space using radio beacons of Instrumental Landing Systems (ILS):

$$S(t) = U_0 \cdot (1+M_1\cos(\Omega_1 t+ \varphi_1) + M_2\cos(\Omega_2 t + \varphi_2)) \cdot \cos(\omega_0 t + \varphi_0), \qquad (14)$$

where: U_0 is the signal amplitude, which depends on the radiation pattern at the receiving point;

M_1, M_2 - spatial modulation depth coefficients;

Ω_1, Ω_2 are the frequencies of the modulating signals 150 Hz and 90 Hz;

φ_1, φ_2 are the phases of the corresponding signals;

ω_0, φ_0 - frequency and phase of the carrier wave 330 MHz;

The informative parameter that needs to be determined is the difference in the depth of spatial modulation (DDM)

$$DDM = M_1 - M_2.$$

4.2 Algorithm for estimation of the informative parameters of ILS

The received signal at the input of the radio receiver is an additive mixture of the useful signal and Gaussian interference η

$$X(t) = S(t) + \eta, \tag{15}$$

The task is to isolate the information parameter from this mixture.

To solve it, we use a statistical approach based on a method of the maximum likelihood. We have an input sample of values $x_i = X(t_i)$. To simplify the calculations, we assume that the amplitude U_0, the frequencies Ω_1, Ω_2, ω_0 and the corresponding phases φ_1, φ_2, φ_0 of the signal are known.

We compose the likelihood function with according to the vector of parameters $\overline{M} = (M_1, M_2)$.

$$f(x_1, x_2, ..., x_n / \overline{M}) = \frac{1}{(\sqrt{2\pi\sigma^2})^n} \cdot \exp\left\{ -\frac{\sum_{i=1}^{n}\left[x_i - S(t_i \mid \overline{M})\right]^2}{2\sigma^2} \right\}. \tag{15}$$

We optimize these equations by finding the maximum of the likelihood function by differentiating the logarithm of the function (15). We obtain the following system of likelihood equations:

$$\begin{cases} \frac{\partial}{\partial M_1} \ln f_1(x_1, .. x_n, / M_1) = \\ -\frac{1}{2\sigma^2} \sum_{i=1}^{n} \{x_i - U_0[1 + M_1 \cos(\Omega_1 t_i) + M_2 \cos(\Omega_2 t_i) \cos(\Omega_0 t_i)]\} \times [\cos(\Omega_1 t_i) \cdot \cos(\Omega_0 t_i)] = 0, \\ \frac{\partial}{\partial M_2} \ln f_2(x_1, .. x_n, / M_2) = \\ -\frac{1}{2\sigma^2} \sum_{i=1}^{n} \{x_i - U_0[1 + M_1 \cos(\Omega_1 t_i) + M_2 \cos(\Omega_2 t_i) \cos(\Omega_0 t_i)]\} \times [\cos(\Omega_2 t_i) \cdot \cos(\Omega_0 t_i)] = 0. \end{cases} \tag{16}$$

or

$$\begin{cases} \sum_{i=1}^{n} \{x_i - U_0[1 + M_1 \cos(\Omega_1 t_i) + M_2 \cos(\Omega_2 t_i) \cos(\Omega_0 t_i)]\} \cdot [\cos(\Omega_1 t_i) \cdot \cos(\Omega_0 t_i)] = 0, \\ \sum_{i=1}^{n} \{x_i - U_0[1 + M_1 \cos(\Omega_1 t_i) + M_2 \cos(\Omega_2 t_i) \cos(\Omega_0 t_i)]\} \cdot [\cos(\Omega_2 t_i) \cdot \cos(\Omega_0 t_i)] = 0. \end{cases} \tag{17}$$

The use of traditional methods for solving such a system of equations (17) is very laborious, since they can be used under conditions of parametric definiteness of the problem and with a small sample; otherwise, calculations are difficult to perform, and their speed decreases. Therefore, we further consider the features of using numerical methods for solving nonlinear equations.

4.3 Simple Iteration Method

This method consists of two stages: determining the initial approximation and the iterative process itself. Considering that the modulation coefficient can take a value from 0 to 1, we take $M = 0$ as the initial approximation. Next, we substitute this value into the equation and calculate the correction. For the next approximation, we take the obtained value of the modulation coefficient.

$$\begin{cases} M_1(j+1) = M_1(j) - \\ -\frac{1}{i}\sum_{i=1}^{n}\left\{x_i - \mathrm{U}_0\left[1+\mathrm{M}_1\cos(\Omega_1 \mathrm{t_i}) + \mathrm{M}_2\cos(\Omega_2 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\right\}\cdot\left[\cos(\Omega_1 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right] \\ M_2(j+1) = M_2(j) - \\ -\frac{1}{i}\sum_{i=1}^{n}\left\{x_i - \mathrm{U}_0\left[1+\mathrm{M}_1\cos(\Omega_1 \mathrm{t_i}) + \mathrm{M}_2\cos(\Omega_2 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\right\}\cdot\left[\cos(\Omega_2 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right] \end{cases} \quad (18)$$

The criterion for the end of the iterative process (18) is to select a condition until the approximate value $M_i^{(k)}$ differs from the previous one $M_i^{(k\text{-}1)}$ by an acceptable value of accuracy E.

$$\max_{1\le i\le n}\frac{\left|\mathrm{M}_i^{(k)} - \mathrm{M}_i^{(k\text{-}i)}\right|}{\mathrm{M}_i^{(k)}} < E$$

The simple iteration method has fast convergence and computation speed.

4.4 Newton-Raphson Method

We use yet another numerical Newton-Raphson method to solve the system of equations (17). We can write the iterative equation in the form

$$M_1(j+1) = M_1(j) - \frac{-\sum_{i=1}^{n}\left\{x_i - \mathrm{U}_0\left[1+\mathrm{M}_1(\mathrm{j})\cos(\Omega_1 \mathrm{t_i}) + \mathrm{M}_2(\mathrm{j})\cos(\Omega_2 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\right\}\left[\cos(\Omega_1 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]}{\frac{\partial}{\partial M_1}\left[\sum_{i=1}^{n}\left\{x_i - \mathrm{U}_0\left[1+\mathrm{M}_1(\mathrm{j})\cos(\Omega_1 \mathrm{t_i}) + \mathrm{M}_2(\mathrm{j})\cos(\Omega_2 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\right\}\left[\cos(\Omega_1 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\right]\Bigg|_{M_1 = M_{\blacklozenge}(j)}} \quad (19)$$

Transforming it, we obtain an algorithm for estimating the modulation depth coefficient:

$$M_1(j+1) = M_1(j) - \frac{\sum_{i=1}^{n}\left\{x_i - \mathrm{U}_0\left[1+\mathrm{M}_1(\mathrm{j})\cos(\Omega_1 \mathrm{t_i}) + \mathrm{M}_2(\mathrm{j})\cos(\Omega_2 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\right\}\left[\cos(\Omega_1 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]}{\sum_{i=1}^{n}(-\mathrm{U}_0)\cdot\left[\cos(\Omega_1 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]\left[\cos(\Omega_1 \mathrm{t_i})\cos(\Omega_0 \mathrm{t_i})\right]} \quad (20)$$

The value of M2 is similar. Accordingly, to determine the difference in the modulation depth coefficient, it will be necessary to calculate the formula (19) for each coefficient and calculate the difference between them DDM = M1 - M2.

The calculations were carried out for $M_1 = 0.2$ and $M_2 = 0.2$.

4.5 Fourier Transform

The use of the Newton-Raphson method for solving systems of equations in radio engineering problems allows one to obtain a sufficiently high accuracy in estimating information parameters. However, it is time-consuming and difficult to build a computing process.

Considering that when assessing the modulation depth coefficient in the ILS-85 instrumental landing system, an allowable error in the estimation of signal parameters of at least 2% was established, we can use simpler and more easily implemented calculators methods that provide a given accuracy. One such approach to determining the parameters of radio engineering signals is to use the Fourier transform.

In the ILS-85 instrument landing system, the task of determining the modulation depth coefficient is simplified by the fact that the carrier frequency and modulation frequencies are known in advance and do not need additional determination. First, we take the phases of the corresponding signals $\varphi_1, \varphi_2, \varphi_0$ known. Using the Fourier transform, we find the signal spectrum and harmonic amplitudes corresponding to the carrier frequency Ω_0 and frequencies $\Omega_0 \pm \Omega_1, \Omega_0 \pm \Omega_2$. The modulation depth coefficient is found from the ratio of the amplitudes of these harmonics.

The research results are presented in the form of curves of the root-mean-square deviation (RMS) of the DDM estimate from the sample size, the signal-to-noise ratio $U^2/2\sigma^2$ and from the value of the estimated DDM parameter itself.

When changing the signal sample, it can be seen that even with a small number of signal samples (400-500 samples), the DDM estimate is 0.001, i.e. It is within acceptable limits with a sufficient margin of accuracy. As can be seen from Fig. 8, when the SNR decreases, the DDM estimate changes slightly and at the level of 20 dB it equals 0.0005, which is an order of magnitude higher than the permissible value.

Thus, the RMS of the DDM assessment is no more than 0.0005 and, accordingly, does not exceed the permissible value. This gives us a sufficient margin of accuracy and reason to use the considered methods for assessing DDM in the design of radar landing systems.

It should be noted that the RMS of DDM estimates obtained using the maximum likelihood method is less than when using the Fourier transform of the signal, but they require a lot of time.

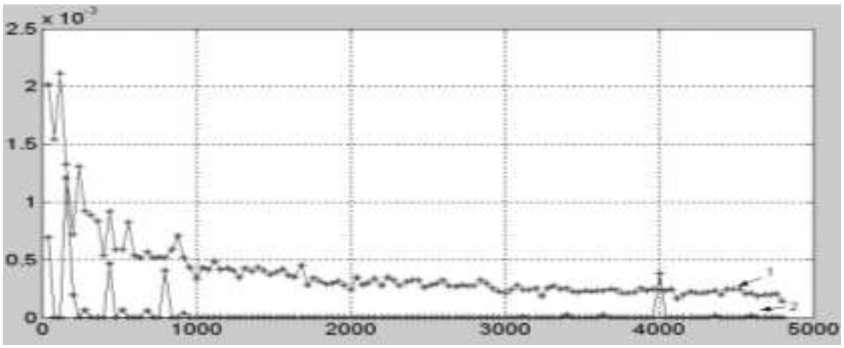
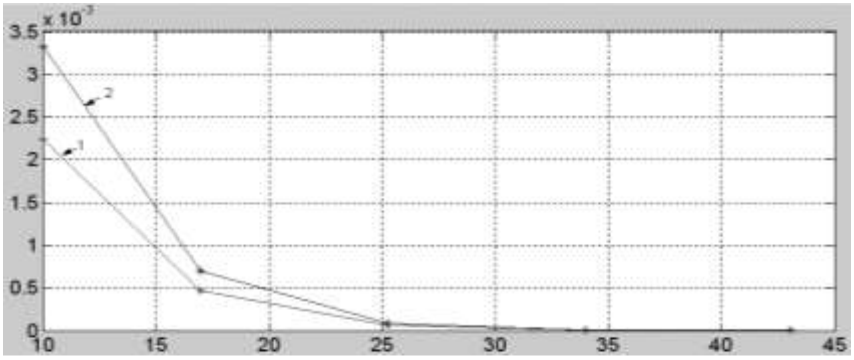


Fig. 8. RMS for DDM assessment when changing SNR

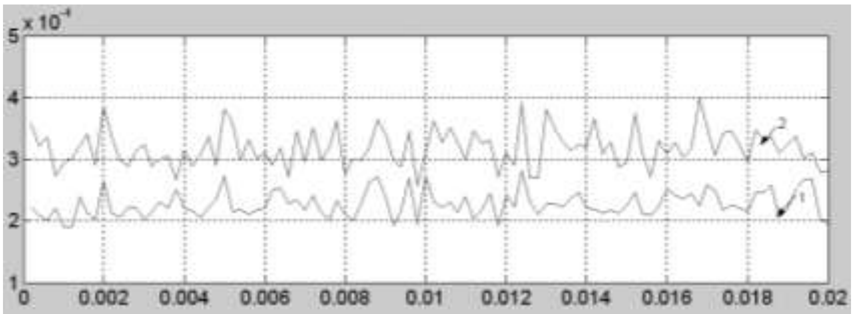
RMS



N

Fig. 9. RMS of the DDM assessment when changing the signal sample

RMS



DDM

Fig. 10. RMS of the DDM assessment when changing the value of the DDM

Relative error

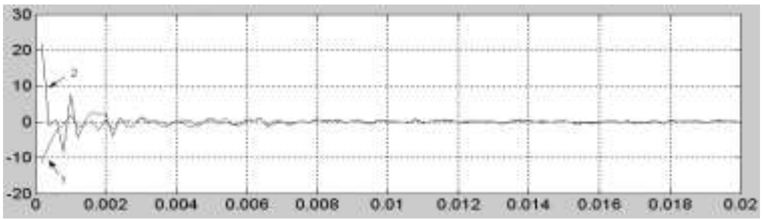


Fig. 12. The relative error of the estimate of the DDM from the value of the DDM

5 Conclusions

Using presented methods of radio signals estimation (phase, frequency, depth modulation coefficient) allows to increase the technical characteristics of radio engineering systems with respect to signal processing. So, these processing algorithms allows to evaluate information parameters in more difficult interference conditions.

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