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METHOD OF THE INCREASING THE DETECTION SYSTEM AND RECOGNITION OF DIGITAL RADIOSIGNALS

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МЕТОД ПОВЫШЕНИЯ ПОМЕХОУСТОЙЧИВОСТИ СИСТЕМЫ ОБНАРУЖЕНИЯ, РАСПОЗНАВАНИЯ И ЛОКАЛИЗАЦИИ ЦИФРОВЫХ СИГНАЛОВ В ИНФОРМАЦИОННЫХ СИСТЕМАХ

Summary. In the process of detecting and recognizing a digital radio signal, a topical issue is increasing noise immunity. The features of the use of low frequency filters with quadratic and linear response dependence on the input signal are investigated in the article. It is shown that the principle of operation of filters is that the summation process is performed. In this case, the useful signal is summed up coherently and the interference signal is incoherent, ie, the useful signal increases and the interference signal decreases.

When acting on the input of linear and quadratic filters of rectangular pulse, which simulates the signals of modern digital means, the necessary parameters for further use of the input and output signals are determined: mathematical expectation, correlation coefficient, variance, root mean square deviation, ratio of signals to noise in temporal and spectral form. The payoff coefficient, which shows the efficiency of using low-pass filters, is calculated.

The graphs of the envelope voltage at the output of the ideal bandpass filter when exposed to the input of a rectangular pulse with different duration - the signal of the means of silent receiving of information.

The filtration process is simulated at different correlation coefficients. This confirmed the possibility of isolating the signal of the means of silent receiving of information by the method of determining the two-dimensional density of the likelihood of interference signal against the background of the common signal.

The process of increasing the noise immunity of the system as a whole is investigated. It is proved that the use in the process of signal processing of narrow-banded filters of low frequency allows to increase the noise immunity of the system of detection and recognition of digital radio air signals by 23%.

Аннотация. В процессе выявления, распознавания и локализации сигнала средств негласного получения информации в информационных системах актуальным вопросом является повышение помехоустойчивости. В статье исследованы особенности использования фильтров низкой частоты с квадратичной и линейной зависимостью отклика на входной сигнал. При воздействии на вход линейного и квадратичного фильтров прямоугольного импульса, который имитирует сигнал современных цифровых средств негласного получения информации, определены необходимые для дальнейшего использования параметры входных и выходных сигналов: математическое ожидание, коэффициент корреляции, дисперсия, среднеквадратическое отклонение, отношение сигнала к помехе во временном и спектральном виде. Вычислено коэффициент выигрыша, которой показывает эффективность использования фильтров низкой частоты.Приведены графики огибающей напряжения на выходе идеального полосового фильтра при воздействии на вход прямоугольного импульса с разной продолжительностью – сигнала средств негласного получения информации.

Проведено моделирование процесса фильтрации при разных коэффициентах корреляции. Это подтвердило возможность выделения сигнала средств негласного получения информации методом определения двухмерной плотности вероятности сигнала помехи на фоне общего сигнала.

Исследуется процесс повышения помехоустойчивости системы в целом. Доказано, что использование в процессе обработки сигналов узкополосных фильтров низкой частоты позволяет добиться повышения помехоустойчивости системы определения, распознавания и локализации средств негласного получения информации на 23%.

Keywords: noise immunity, filter, mathematical expectation, variance, modeling.

Ключевые слова: помехоустойчивость, фильтр, математическое ожидание, дисперсия, моделирование.

Introduction

Radio interference means any kind of electrical oscillation which, when penetrated into or from inside the radio, complicates the determination of the radio signal. The signal and interference, acting simultaneously on the input of the receiver, are reproduced at the output of the latter as a random oscillatory process. As a result, the signal parameters cannot be precisely determined. Normal signal detection is only possible with a certain ratio of signal power and interference at the receiver output. The minimum signal strength at which a satisfactory signal determination is provided depends on the interference level. This amount of power characterizes the sensitivity of the receiver. The ability of a radio receiver to receive a signal of a given quality in the presence of interference is called noise immunity. Improving the noise immunity of radios is one of the major and most complex problems of radio engineering. In order to successfully solve it, it is necessary to study the properties and nature of the influence of the interference on the signal, and then to determine the ways of attenuating their influence on the quality of signal determination.

Issues of overcoming interference also have their peculiarities in the process of detecting and recognizing digital radio signal. To this end, consider the issue of noise immunity in the study of the above processes.

Analysis of recent publications and problem statement

A considerable number of publications are devoted to the issue of noise immunity. Thus, in [1], the technical methods of improving radio efficiency related to noise immunity are considered. The methods of increase of noise protection and noise immunity are considered and the factors that shape them. The most dangerous interferences affecting the radio station are the relaying factors, when the correlation function of the useful signal and the interference take on large values compared with the values for the interference of the pseudo-probable sequence and the harmonic interference. It is shown that variants of coding of the source of information do not fundamentally affect the stability of radio stations during the action of these interferences. However, noise immunity issues are not addressed when probable digital signals are detected.In [2], the process of noise immunity of a typical detection path composed of sequentially included modules is considered: an ideal bandpass filter, a quadratic detector, and an ideal integrator. The described technique for determining the probabilistic characteristics of detection can be applied to the study of typical path composed of other elements of considerable practical interest. However, the issue of the effect of interference on a rectangular signal that is similar to a digital signal is not addressed.In article [3], using the methods of statistical radio engineering, the noise immunity of receiving signals with quadrature amplitude modulation in the presence of noise and harmonic interference is analyzed. The dependences of the bit error probability on the signal-to-noise ratio, the noise intensity, and its decomposition relative to the center frequency of the useful signal spectrum are obtained. It is shown that the reception of signals with quadrature amplitude modulation is greatly impaired in the presence of harmonic interference, and with the increase of signal positionality this influence is enhanced. However, the determination of digital radio signals is not considered. In article [4], based on distributed models, a method of bringing voice signals to a single amplitude and time window is proposed. Distributed clustered schemes of voice signals training are also proposed for forming reference models of speech voice sounds. These methods make it possible to quickly convert quasiperiodic sections of different lengths into a single window of amplitude and time for further comparison, and to determine the optimal number of clusters, which increases the likelihood of clustering. The proposed methods can be used in signal recognition systems. In [5], an optimization model for the measurement of power in circuits was developed on the basis of studies conducted in MATLAB. The proposed algorithms can be used to develop the characteristics of various information signals, including digital signals from modern devices.In [6] investigated the effect of multiray propagation of radio waves on the transmission of audio content through channels with normal and lognormal interference distribution using GSM and WiMAX wireless technologies. For researching in the MATLAB Simulink software environment, appropriate models of transceivers have been built using the elements of the Communication System Toolbox library. At the same time, we do not use low frequency filters with quadratic and linear dependence of the response on the input signal.In [7], a technique for the interaction of mobile technical objects in the process of data flow transfer under conditions of powerful electromagnetic field is proposed.The work [8] is devoted to increasing the noise immunity of information messages under the conditions of powerful electromagnetic interference by the use of complex signal-code structures. This increases the volume and speed of information transfer. As a result of encoding information with super short pulses in wireless information transmission systems, a quantitative and qualitative evaluation of the effectiveness of the proposed method was carried out. However, signal filtering methods are not addressed in this paper.In [9] the results of studies on increasing the signal-to-noise ratio in mobile communication systems are highlighted. This direction is realized through the use of methods of dynamic change of transmitter power, organization of multiple access and dynamic distribution of communication channels. However, the issue of digital signal recognition is not resolved.From the analysis of modern literature, we can conclude that the problems of noise immunity, which have their own peculiarities in the process of detecting and recognizing digital signal of digital radio broadcasting, are practically not considered. Therefore, it seems appropriate to investigate the issue of noise immunity in the automated detection system and the recognition of digital radio broadcasts.

Presentation of main material

Almost all methods of noise immunity receive signals based on the principle of signal averaging and interference. This principle is that the summation process is performed. Moreover, the useful signal is summed up coherently, and the noise signal is incoherent. For the purpose of averaging the useful signal and interference, linear systems of two types are used: narrow band filters and low frequency filters. It is possible to optimize low pass filters or narrow band filters.

To consider the issue of interference filtering, let us assume that the narrowband filter itself does not distort the signal that has passed through it. An ideal bandpass filter is a filter with an amplitude-frequency response of the type:

$$K(\omega)=\begin{cases}1 \quad якщо \quad \omega_0-\dfrac{\Delta\omega}{2}\le|\omega|\le\omega_0+\dfrac{\Delta\omega}{2}\\ 0 \quad якщо \quad \left]-\infty,\omega_0-\dfrac{\Delta\omega}{2}\right[\ U\ \left]\omega_0+\dfrac{\Delta\omega}{2},\infty\right[\end{cases}, \tag{1}$$

where $\Delta\omega$ – filter bandwidth.

For an ideal filter, effective band $\Delta\ \omega_e$ and band $0,707-\Delta\ \omega_{\sqrt{2}}$, is equal to the filter transparency band $\Delta\omega$.

For filters, the assumption is that $\Delta\omega << \Delta\omega_0$.

The frequency response of the expression for (1) is the impulse transition characteristic, which will be determined by the expression:

$$h_\delta(t) = \frac{\Delta\omega}{\pi} \cdot \frac{\sin\frac{\Delta\omega t}{2}}{\frac{\Delta\omega t}{2}} \cos\omega_0 t \quad (2)$$

Given that the digital signal is not a clear pulse [10], it is possible to calculate the envelope voltage at the output of an ideal filter when exposed to a rectangular pulse of duration:

$$x(t) = \begin{cases} X_m \cos\omega_0 t & якщо \quad 0 \le t \le T \\ 0 \quad якщо & \left]-\infty, 0\right[U \left]T, \infty\right[\end{cases}, \quad (3)$$

where X_m – the envelope signal $x(t)$ at the inlet of the filter.

Using the envelope voltage theorem of the narrowband filter, we write the expression for the envelope voltage at the output of the filter:

$$Y_m(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} K_{fn}(j\omega) S_{X_m}(j\omega) e^{j\omega t} dt, \quad (4)$$

where $S_{X_m}(j\omega) = \int_{-\infty}^{\infty} X_m e^{-j\omega t} dt$ – the amplitude spectrum of the envelope signal $x(t)$,

K_{fn} – complex factor of low frequency filter transmission:

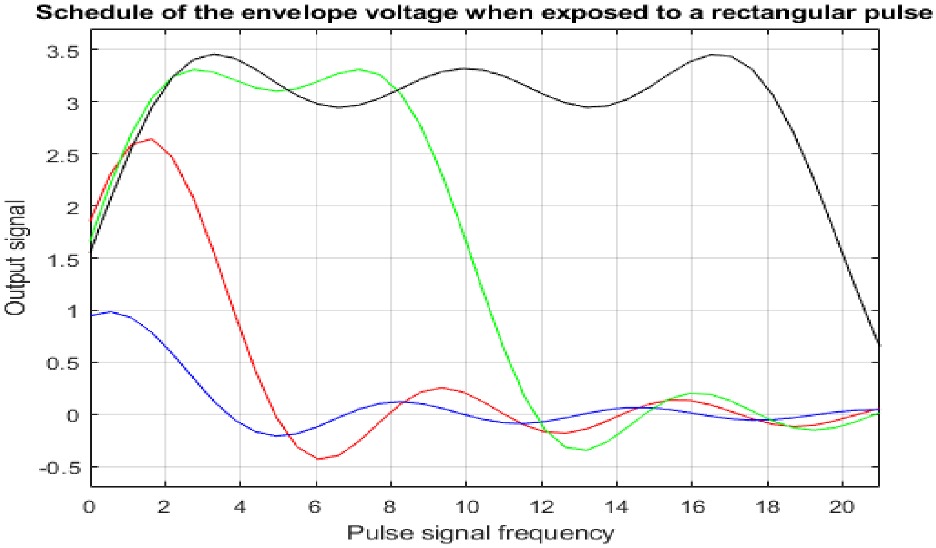
$$K_{fn}(j\omega) = \begin{cases} 1 \quad якщо & -\frac{\Delta\omega}{2} \le |\omega| \le \frac{\Delta\omega}{2} \\ 0 \quad якщо & \left]-\infty, \frac{\Delta\omega}{2}\right[U \left]\frac{\Delta\omega}{2}, \infty\right[\end{cases} \quad (5)$$

Substituting expression (5) into expression (4), we get the expression:

$$Y_m(t) = \frac{X_m}{2\pi} \left(Si(\Delta\omega t) - Si(\Delta\omega(t-T)) \right), \quad (6)$$

where $Si(z) = \int_0^z \frac{\sin t}{t} dt$ – integral sinus [11].

In Pic. 1 dependency graphs of the duration of the influencing rectangular pulse (blue color - pulse duration T = 1, red color - T = 10, green color - T = 15 and black color - T = 20) on the frequency range (filter bandwidth).



Pic. 1. Graph of the envelope voltage when exposed to a rectangular pulse signal

The graphs show significant differences between the input rectangular pulse and the output signal. The distortion of the input impulse increases as its duration increases. This distortion of the pulse shape can be characterized by the duration of the envelope of the impulse of the filter output to the duration of the envelope of the output impulse.

This indicates that short-term rectangular signals can be isolated using a bandpass filter [12].

To further calculate the interference signal, we determine the autocorrelation coefficient of the white noise passed through the bandpass filter:

$$R_w(\tau) = \frac{\int_0^\infty K^2(\omega)\cos\omega t dt}{\int_0^\infty K^2(\omega) d\omega} . \quad (7)$$

After substituting expression (1) into expression (7) we obtain:

$$R_w(\tau) = \frac{1}{\Delta\omega}\int_{\omega_0-\frac{\Delta\omega}{2}}^{\omega_0+\frac{\Delta\omega}{2}} \cos\omega\tau d\omega = \frac{\sin\left(\omega_0+\frac{\Delta\omega}{2}\right)\cdot\tau - \sin\left(\omega_0-\frac{\Delta\omega}{2}\right)\cdot\tau}{\Delta\omega\tau} \quad (8)$$

or

$$R_w(\tau) = r_w(\tau)\cos\omega_0\tau , \quad (9)$$

where $r_w(\tau) = \frac{\sin(\Delta\omega\frac{\tau}{2})}{\Delta\omega\frac{\tau}{2}}$ – the envelope of the autocorrelation coefficient of the process at the output of the bandpass filter.

Due to the fact that the signal of digital means of silent retrieval of information is a signal of a rectangular pulse, with an envelope of duration *T*, the expression has the form [13]:

$$y_s = \begin{cases} A\cos(\omega_0 + \varphi_0) \quad ,0 \le t \le T \\ 0 \qquad]-\infty, t[U]t, \infty[\end{cases} \quad . \tag{10}$$

Then the numerical characteristics of the quadratic filter filtering process will look like:

$$m_1[z_{\Sigma 0}(t)] = \begin{cases} A_1\sigma^2_{yN}(1+q^2), \quad 0 \le t \le T \\ A_1\sigma^2_{yN} = m_1[z_{N0}(t)],]-\infty, t[U]t, \infty[\end{cases} , \tag{11}$$

where $m_1[z_{0N}(t)]$ – mathematical expectation of low-frequency noise fluctuation.

$$R_{z\Sigma 0}(t, t+\tau) = \begin{cases} \dfrac{r^2_{yN}(\tau) + 2q^2 r_{yN}(\tau)}{1 + 2q^2} = R_{z\Sigma 0}(\tau) \ , \quad 0 \le t \le (T-\tau) \\ r^2_{yN}(\tau) = R_{zN0}(\tau), \quad]-\infty, t[U]t, \infty[\end{cases} \tag{12}$$

$R_{zN0}(\tau)$ – autocorrelation coefficient of low-frequency noise interference.

$$D_{z\Sigma N} = \sigma^2_{z\Sigma N}(t) = \begin{cases} A_1^2\sigma^4_{yN}(1 + 2q^2) = \sigma^2_{z\Sigma 0}, \quad 0 \le t \le T \\ A_1^2\sigma^4_{yN} = \sigma^2_{zN0}, \quad]-\infty, t[U]t, \infty[\end{cases} \tag{13}$$

A process in which the mathematical expectation and the correlation function are independent of time at a fixed fixed time interval is called quasi-stationary. Then the process at the output of the filter will not affect its additive amount of signal and interference and will be quasi-stationary [14].

For a linear filter, the numerical parameters of the filtering process take the form:

$$m_1[z_{\Sigma 0}(t)] = \frac{A_1\sigma^2_{y\Sigma}}{\sqrt{2\pi}} \ ; \tag{14}$$

$$R_{z\Sigma 0}(\tau) \approx r^2_{yN}(\tau); \tag{15}$$

$$\sigma^2_{z\Sigma 0} = \frac{A_1\sigma^2_{y\Sigma}}{8\pi} \ , \tag{16}$$

where $\sigma^2_{y\Sigma} = D_{y\Sigma}$ – the variance of the total process at the inlet of the filter. It is determined by:

$$\sigma^2_{y\Sigma} = \sigma^2_{yS} + \sigma^2_{yN} \ , \tag{17}$$

where $D_{ys} = \sigma^2_{ys}, D{=}_{ys}\sigma^2_{yN}$ – signal dispersion and interference at the filter input.

We calculate the mutual correlation functions $z_{N0}(t), \quad z_{\Sigma 0}(t)$ of the output signals.

In the case of an interference signal, the mathematical expectation for a second order mixed signal $z_N(t), \quad z_\Sigma(t)$ will be determined by the expression:

$$m_1[z_N(t_1), z_\Sigma(t_2)] = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} z_N(t_1) z_\Sigma(t_2) w_2[y_N(t_1), y_\Sigma(t_2)] \times$$
$$\times d\,y_N(t_1) d\,y_\Sigma(t_2) = A_2^2 \int_0^{\infty}\int_0^{\infty} y_N(t_1) y_\Sigma(t_2) w_2[y_N(t_1), y_\Sigma(t_2)] \times$$
$$\times d\,y_N(t_1) d\,y_\Sigma(t_2) \tag{18}$$

where $w_2[z_N(t_1), z_\Sigma(t_2)]$ – two-dimensional probability density of stationary normal processes $y_N(t), \quad y_\Sigma(t)$. Given that the autocorrelation coefficient of both signals is the same and equal to $R_{yN}(\tau)$, it is possible to write an expression for $w_2[z_N(t_1), z_\Sigma(t_2)]$ – two-dimensional probability density in the form:

$$w_2[z_N(t_1), z_\Sigma(t_2)] = \frac{1}{2\pi\ \sigma_{yN}\sigma_{y\Sigma}\sqrt{1-R_{yN}^2(\tau)}} \cdot \exp\times$$
$$\times\left(-\, {}^{1}\!\big/_{2(1-R_{yN}^2(\tau)} \left[\frac{y_N^2(t_1)}{\sigma_{yN}^2} - 2R_{yN}(\tau)\frac{y_N(t_1)y_\Sigma(t_2)}{\sigma_{yN}\sigma_{y\Sigma}} + \frac{y_\Sigma^2(t_1)}{\sigma_{y\Sigma}^2} \right.\right) \tag{19}$$

Performing the substitution of the form: $\dfrac{y_N(t_1)}{\sigma_{yN}} = x, \quad \dfrac{y_N(t_2)}{\sigma_{y\Sigma}} = y$ we get the expression:

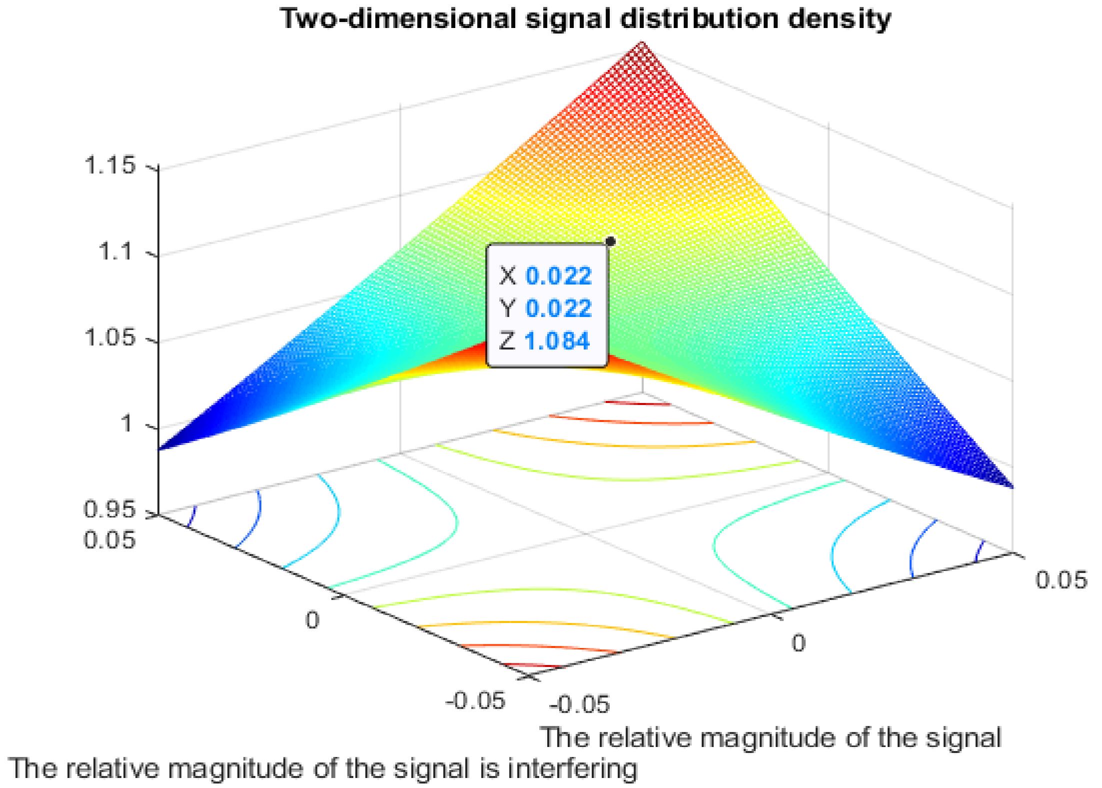
$$m_1[z_N(t_1), z_\Sigma(t_2)] = \frac{A_2\sigma_{yN}\sigma_{y\Sigma}}{2\pi\ \sqrt{1-R_{yN}^2(\tau)}} \cdot \int_0^{\infty}\int_0^{\infty} \exp\left(-\frac{x^2 - 2R_{yN}(\tau)xy + y^2}{2(1-R_{yN}^2(\tau))} \right) dxdy \tag{20}$$

In order to determine the effect of the correlation coefficient (signal-to-noise ratio) on the mathematical expectation (ie, the effect of noise-to-signal), we will simulate the process.

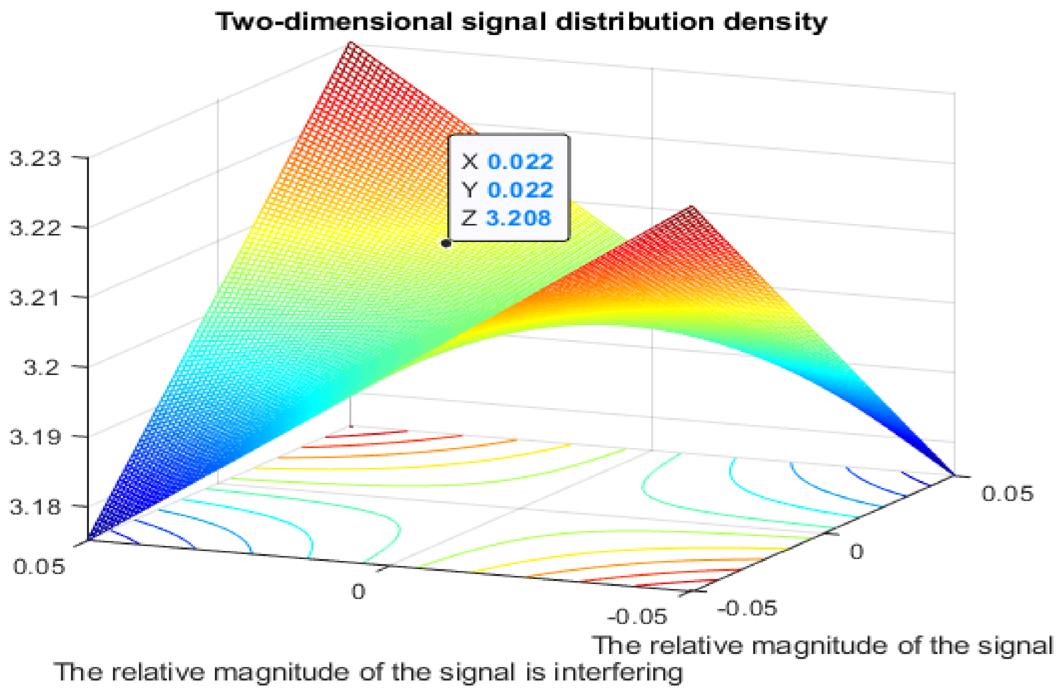
To evaluate the strength of communication in the theory of correlation, the scale of English mathematics Cheddock is used: weak - from 0.1 to 0.3; moderate - from 0.3 to 0.5; noticeable - from 0.5 to 0.7; high - from 0.7 to 0.9; very high (strong) - from 0.9 to 1.0.

Therefore, we consistently choose the correlation coefficient for weak, moderate and high bond strength of communication, respectively [15].

The simulation results are shown in Pic. 2-3:



Pic. 2. Two-dimensional density of signal distribution at R_{yN} *= 0.3 (moderate dependence)*



Pic. 3. Two-dimensional density of signal distribution at R_{yN} *= 0.9 (moderate dependence)*

To analyze the results obtained, on each of the graphs took the point with the same coordinates of the relative values of signal and interference.

As we can see from the above graphs Pic. 2-3, with increasing correlation dependence from weak to high, the magnitude of the signal distribution density increases. This indicates the ability to distinguish between signal and interference, reducing interference by filtering.

To determine the signal-to-noise ratio at the output of a typical path when exposed to its input noise additive $N(t)$ and signal $S(t)$ we have:

$$x(t) = S(t) + N(t). \quad (21)$$

Suppose that the signal and the interference are stationary white noise, with zero mathematical expectation $m_1(S(t)) = m_1(N(t)) = 0$. Signal and interference are uncorrelated: $m_1(S(t)(N(t)) = 0$ and defined over a long period of time. Then it is possible to write the expressions:

$$D_\Sigma = \sigma_\Sigma^2 = \Delta f_e S_\Sigma; \quad D_s = \sigma_s^2 = \Delta f_e S_s; \quad D_N = \sigma_N^2 = \Delta f_e S_N , \qquad (22)$$

where Δf_e – effective filter transparency band;

$D_\Sigma = \sigma_\Sigma^2$ – variance and root mean square deviation of the signal mixture;

$D_s = \sigma_s^2$ – dispersion and root mean square deviation of the signal;

$D_N = \sigma_N^2$ – dispersion and standard deviation of interference;

$S_\Sigma; S_s; S_N$ – spectral densities, respectively, of a mixture of signal and noise, signal and noise.

The assumptions we have made are as follows:

$$\sigma_\Sigma^2 = \sigma_s^2 + \sigma_N^2 \text{ or } D_\Sigma = D_S + D_N . \qquad (23)$$

The low-frequency component of the voltage at the output of the tract, detected at the time of reference $t = T$, denote $u_{\Sigma 0}$, the interference voltage u_{N0}, the voltage sum of the signal $u_{\Sigma 0}$. It should be noted that u_{N0} and $u_{\Sigma 0}$ are random variables.

The appearance of a signal at the input of the tract, detected at time $t = T$, can change the mathematical expectation of the low-frequency component voltage at the output of the tract, from magnitude $m_1[u_{N0}(t)]$ to $m_1[u_{\Sigma 0}(t)]$. This signal increase is called a useful signal. Write for him the expression:

$$C = m_1[u_{\Sigma 0}(T)] - m_1[u_{N0}(T)] = \Delta\ m_1[u_0(T)] . \qquad (24)$$

In such a case, the interference at the same instant of time T will be determined by the mean square value of fluctuation of the random probable magnitude:

$$N = \sigma_{u\Sigma}(T) = \left(m_1\left[u_{\Sigma 0}{}^2\ (T)\right] - m_1{}^2\ \left[u_{\Sigma 0}(T)\right]\right)^{\frac{1}{2}} . \qquad (25)$$

The signal / interference ratio for $t = T$ will be:

$$\frac{C}{N} = \frac{m_1[u_{\Sigma 0}(T)] - m_1[u_{N0}(T)]}{\left(m_1\left[u_{\Sigma 0}{}^2\ (T)\right] - m_1{}^2\ \left[u_{\Sigma 0}(T)\right]\right)^{\frac{1}{2}}} . \qquad (26)$$

Expressions (17 - 19) are the definition of signal, interference and signal / interference ratio at the output of the receiving path. In the future, our task will be to determine the signal, interference and their correlation through the corresponding parameters at the input of the receiving path.

There are two methods for determining this relationship: spectral and temporal.

In the time method, the voltage at the output of the receiving path at time $t = T$ will be determined by the expression:

$$u(T) = \int_0^T h_\delta(T - t) z(t) dt , \qquad (27)$$

where h_δ – impulse transient response of the filter, $z(t)$ – input voltage.

The mathematical expectation of this voltage when exposed to the input of a mixture of signal and interference will be:

$$m_1\left[u_\Sigma(T)\right] = \int_0^T h_\delta(T-t) m_1\left[z_\Sigma(t)\right] dt \quad . \tag{28}$$

Due to the fact that z_Σ – is a stationary process, its mathematical expectation is independent of time, then we have:

$$m_1\left[u_\Sigma(T)\right] = m_1\left[z_\Sigma(t)\right]\int_0^T h_\delta(T-t)dt = m_1\left[z_\Sigma(t)\right]\int_0^T h_\delta(t)dt \quad . \tag{29}$$

Similarly, it is possible to determine mathematical expectation when exposed only to interference:

$$m_1\left[u_N(T)\right] = m_1\left[z_N(t)\right]\int_0^T h_\delta(t)dt \quad . \tag{30}$$

Substituting expressions (18) and (19) into expression (13) we obtain:

$$C = \Delta\ m_1\left[z_0(T)\right]\int_0^T h_\delta(t)dt \quad , \tag{31}$$

where $\Delta\ m_1\left[z_0(t)\right] = m_1\left[u_{\Sigma 0}(t)\right] - m_1\left[u_{N0}(t)\right]$ increasing the mathematical expectation of the low-frequency component voltage at the output of the filter.

The dispersion of the fluctuations at the output of the low-pass filter is determined by:

$$D_{u\Sigma} = \sigma^2_{u\Sigma} = \left(m_1\left[u_\Sigma{}^2\ (T)\right] - m_1{}^2\left[u_\Sigma(T)\right]\right) \quad . \tag{32}$$

From the expression:

$$u^2_\Sigma(T) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_\delta(T-t_1,T)h_\delta(T-t_2,T)z_\Sigma(t_1)z_\Sigma(t_2)dt_1dt_2 \tag{33}$$

we have:

$$m_1[u^2_\Sigma(T)] = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_\delta(T-t_1,T)h_\delta(T-t_2,T)m_1[z_\Sigma(t_1)z_\Sigma(t_2)]dt_1dt_2 \quad ; \tag{34}$$

$$\begin{aligned} D_{u\Sigma} = \sigma^2_{u\Sigma}(T) = &\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_\delta(T-t_1,T)h_\delta(T-t_2,T)\times \\ &\times m_1[z_\Sigma(t_1)z_\Sigma(t_2)]dt_1dt_2 - m_1^2[z_\Sigma(t)]\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_\delta(T-t_1,T)h_\delta(T-t_2,T)\times \\ &\times h_\delta(T-t_2,T) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_\delta(T-t_1,T)h_\delta(T-t_2,T)\times \\ &\times m_1[z_\Sigma(t_1)z_\Sigma(t_2)]dt_1dt_2 - m_1^2[z_\Sigma(t)]dt_1dt_2 \end{aligned} \tag{35}$$

We will replace: $\tau = t_2 - t;\ \ d\tau = dt_2;\ \ t = t_2;\ \ dt = dt_1$ Then we will have:

$$D_{u\Sigma} = \sigma_{u\Sigma}^2 = \sigma_{z\Sigma}^2 [\int_{-\infty}^{\infty} Q_h(\tau,T) R_{z\Sigma}(\tau) d\tau], \quad (36)$$

Where

$$Q_h(\tau,T) = \int_{-\infty}^{\infty} h_\delta(T-t\ ,T) h_\delta(T-t-\tau,T) dt; \quad (37)$$

$$R_z(\tau) = \frac{m_1[z_\Sigma(t) z_\Sigma(t+\tau)] - m_1{}^2\ [z_\Sigma(t)]}{\sigma_{z\Sigma}^2}, \quad (38)$$

where $R_{z\Sigma}(\tau)$ – autocorrelation coefficient, $D_{z\Sigma} = \sigma_{z\Sigma}^2$ – the variance of the process when exposed to the input of the sum of signal and interference.

Assume that according to expression (37) the interference: $N = \sigma_{u\Sigma}(T)$ then we will have:

$$N = \sigma_{u\Sigma}(T) = \sigma_{\Sigma 0}[\int_{-\infty}^{\infty} Q_h(\tau,T) R_{z\Sigma}(\tau) d\tau]^{\frac{1}{2}}. \quad (39)$$

Due to the fact that energy spectrum is the main factor in determining the signal of the means of silent receiving of information , we will find the expression for interference in spectral form.

To do this, we use the Wiener-Hinchin theorem, which establishes the relationship between the correlation function and the power spectral density $g(\omega)$:

$$K(\tau) = \int_{-\infty}^{\infty} g(\omega) e^{j\omega t} d\omega. \quad (40)$$

Then we get:

$$N = \left(\int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h_\delta(T-t,T) h_\delta(T-t-\tau,T) dt \times [\int_{-\infty}^{\infty} g_{z\Sigma 0}(\omega) e^{j\omega\tau} d\omega] d\tau \right)^{\frac{1}{2}} =$$

$$= \left(\int_{-\infty}^{\infty} g_{z\Sigma 0}(\omega) d\omega \int_{-\infty}^{\infty} h_\delta(T-t,T) dt \int_{-\infty}^{\infty} h_\delta(T-t-\tau,T) e^{j\omega\tau} d\tau \right)^{\frac{1}{2}} =$$

$$= \left(\int_{-\infty}^{\infty} g_{z\Sigma 0}(\omega) K_T(j\omega) d\omega \int_{-\infty}^{\infty} h_\delta(T-t,T) e^{j\omega(T-\tau)} dt \right)^{\frac{1}{2}} =$$

$$= \left(\int_{-\infty}^{\infty} g_{z\Sigma 0}(\omega) \left| K_T(j\omega) \right|^2 d\omega \right)^{\frac{1}{2}} = \left(\int_{-\infty}^{\infty} G_{z\Sigma 0}(\omega) \left| K_T(j\omega) \right|^2 d\omega \right)^{\frac{1}{2}} \quad (41)$$

where $K_T(j\omega)$ – low frequency filter complex frequency response;

$g_{z\Sigma 0}(\omega)$ – spectral power density of low-frequency fluctuations at the output of the filter over the entire frequency axis.

$$G_{z\Sigma 0}(\omega)=\begin{cases}2g_{z\Sigma 0}(\omega)=\dfrac{2\sigma^2_{z\Sigma 0}}{\pi}\int\limits_0^{\infty}R_{z\Sigma 0}(\tau)\cos\omega\tau d\tau, & \omega>0\\ 0, & \omega<0\end{cases}, \quad (42)$$

where $G_{z\Sigma 0}(\omega)$ – the spectral power density of the signal $z_{\Sigma 0}(t)$, determined only in the region of positive frequencies.

Thus, the expression for the signal-to-noise ratio at the output of a typical radio path in the mode of detecting a signal against the background of the signal takes the form:

Temporarily:

$$\frac{C}{N}=\frac{\Delta\ m_1[z_0(t)]\int\limits_0^T h_\delta(t)dt}{\sigma_{\Sigma 0}[\int\limits_{-\infty}^{\infty}Q_h(\tau,T)R_{z\Sigma 0}(\tau)d\tau]^{\frac{1}{2}}}. \quad (43)$$

Spectral recording form:

$$\frac{C}{N}=\frac{\Delta\ m_1[z_0(t)]K_T(0)}{[\int\limits_{-\infty}^{\infty}G_{z\Sigma 0}(\omega)\left|K_T(j\omega)\right|^2 d\omega]^{\frac{1}{2}}}. \quad (44)$$

In addition to the signal-to-noise ratio, the filter's characteristic is the winning ratio, which is determined by the expression:

$$K_B=\frac{C\ /\ N вих}{C\ /\ N вх}. \quad (45)$$

Substituting expression (43) into expression (45), we obtain:

$$K_B=[\frac{\Delta\ m_1[u_0(T)]}{\sigma_{u\Sigma 0}(T)}]/[\frac{\Delta\ m_1[z_0(T)]}{\sigma_{z\Sigma 0}(T)}]=[\frac{\Delta\ m_1[u_0(T)]}{\Delta\ m_1[z_0(T)]}]/[\frac{\sigma_{u\Sigma 0}(T)}{\sigma_{z\Sigma 0}(T)}], \quad (46)$$

where $\frac{\Delta\ m_1[u_0(T)]}{\Delta\ m_1[z_0(T)]}]$, $[\frac{\sigma_{u\Sigma 0}(T)}{\sigma_{z\Sigma 0}(T)}]$ – determine the increase in mathematical expectation and the average square deviation of low-frequency fluctuations as a result of processing the input signal by a low-pass filter.

Thus, in order to increase the noise immunity of the detection and recognition system, it is necessary to use a low pass filter. This significantly lowers or completely eliminates the analysis of low-frequency interference.

Analysis of trends in the development of modern means of silent retrieval of information show trends in the transition of their work in the high frequency range. That is, the information transmission signal is shifted to the high frequency range, in which the process of detecting and recognizing digital signals is quite complicated.

By eliminating the low-frequency interference analysis, we will already significantly increase the system's overall noise immunity.

Conclusions

The peculiarities of the use of low-pass filters to increase the noise immunity of an automated system for detecting and recognizing digital airwaves are investigated. It is shown that the principle of operation of filters is that the summation process is performed. In this case, the useful signal is summed up coherently, and the noise signal is incoherent. That is, when summing up, the useful signal increases and the interference signal decreases.

Taking into account the peculiarities of the digital signal, the signal parameters are defined (mathematical expectation, correlation coefficient, variance, root mean square deviation) and the outputs of linear and quadratic filters at the influence on the input of a rectangular pulse that simulates the signal of modern digital means of silent receiving of information are determined.

The graphs of the envelope voltage at the output of the perfect bandpass filter with the influence on the input of a rectangular pulse (digital signal) of different duration are obtained.

The results of the simulation of the filtering process, with different correlation coefficients, confirmed the possibility of selection of the digital signal by the method of determining the two-dimensional probability density of the signal of interference on the background of the common signal.

It is proved that the use in the process of signal processing of low bandwidth filters of low frequency allows to increase the noise immunity of the system of detection and recognition of digital radio airwaves signals by 23%.

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