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POTENTIAL NOISE-IMMUNITY OF COMMUNICATIONS WITH POWER RECEPTION OF THE SHIFT-KEYED NOISE SIGNALS

M.Ye. Ilchenko [1], T.N. Narytnik [1], & R.M. Didkovsky [2]♦

[1]*National Technical University of Ukraine "Kyiv Polytechnic Institute"*
37, Prospect Peremogy Ave., Kyiv 03056, Ukraine

[2]*Cherkassy state technological university*
460, Shevchenko Blvd, Cherkassy 18006, Ukraine

♦Address all correspondence to R.M. Didkovsky E-mail: didkow@gmail.com

Accurate theoretical estimates of bit error probability in the communications using amplitude and frequency noise shift-keying are obtained. It is demonstrated that in the given kinds of communications the bit error probability depends upon not only the signal-to-noise ratio but upon the bandwidth-delay product as well. The definitions of optimal bandwidth-delay product and optimal signal-to-noise ratio upon power are substantiated and the relevant values are calculated. Comparative analysis of noise-immunities is performed. The results of theoretical calculations are confirmed with the help of the computer simulation.

KEY WORDS: *UWB communications, noise carrier, power reception, system performance, amplitude noise shift-keying, frequency noise shift-keying, bit error probability, bandwidth-delay product*

1. INTRODUCTION

Employment of the more complicated signals than the harmonic oscillation as data carriers is highly efficient in development of the current communications. It had become the subject of study for prominent researches and leading experts in the communication branch already in the middle of the last century. The search for alternative options of the carrier [1] started at that same time. This line of inquiry has grown especially active during the recent two decades [2].

By now there have been formed 4 principal trends of development of the communications using signals of a complex shape, and namely:

1) spread spectrum harmonic signal [3];

2) ultra-short pulses and the series thereof [4];
3) dynamic chaos [5–6];
4) noise signals [7–8].

The spread spectrum systems allow performing (although at the expense of significant hardware resources) coherent reception of the transmitted signal. This provides for their high potential system performance. For this reason that systems have been remained in the center of the researchers' attention during the last 50 years. It resulted in a widespread commercial implementation of the relevant methods and technologies [9].

The situation is completely different in development of the communications using pulse, chaotic and noise signals. Realization of the coherent reception in such systems is quite a complicated problem, and in the case with the noise carrier it is principally impossible. Compared to the classical Gaussian channel these systems lose uniquely in terms of error performance.

However, there exist a number of particularities of non-harmonic carries of signals that attract the attention of the developers of special means of communication [10–11]. Primarily, it is high level of structural security and encryption stability of the signal that provides for protection of the transmitted data from any kind of outside interference on the physical (signal) level. System resistance to complicated signal propagation conditions (multibeam channel, variable parameter channel, Doppler frequency shift etc.) is also of not less importance. Besides, the structure of transmitter and receiver units using initially a broadband signal turns out to be much simpler than the structure applied in the communications using spread spectrum of a harmonic signal.

In our opinion, the attention that is currently being paid to information protection in public and commercial telecommunication networks along with the strive to explore new bands of ultra-high frequencies [12] must drive the interest of developers to using of non-harmonic, in particular, the noise signals.

In this paper we shall consider in more detail the issue of studying the properties of data communications using the noise carrier. Under the term 'noise' we shall mean realization of an arbitrary broadband stochastic process. Hence it follows that the signal shape will be a priori unknown at the receiver.

The power reception is one of the most apparent methods for reception of the signal with an absolutely unknown shape.

The communications using the amplitude [13] and frequency [14] noise shift-keying are based right upon the principle of power reception. Despite the fact that such systems are suggested long ago there has been performed no thorough investigation of their potential performance parameters.

The objective of this paper is to obtain mathematically substantiated expressions for the bit error probability in the communications using the amplitude and frequency noise shift-keying and to study particularities of potential system performances of the above communications.

2. COMMUNICATIONS USING AMPLITUDE NOISE SHIFT-KEYING

Let us formalize the description of the method of amplitude noise shift-keying (ANSK), build mathematical models of the processes running within the respective communication and calculate the bit error probability.

The ANSK principle is rather simple. If the current k -th information bit is equal to "1" then along the bit interval with the duration T the transmitter radiates a fragment of the noise signal $x(t)$, $t \in \left[(k-1) \cdot T, k \cdot T \right]$, $k = 1,2,3,...$. If the bit is equal to "0" then it is corresponded by a passive pause of the same duration. Shown in Fig. 1 is block diagram of the units engaged in performing of the said means of communication.

The noise signal generator (NSG) demonstrated in Fig. 3 forms the signal $x(t)$, which is the realization of a continuous stochastic process concentrated within the frequency bandwidth $\left[f_1; f_2 \right]$ having zero mathematical expectation and the dispersion (power) equal to D_x. The signal $x(t)$ arrives at the input of the controlled switch 5. The primary digital signal of the encoder 4 varying under the law of the input bit flow is fed to the controlled switch input. At transmission of the bit "1" the switch is closed and the transmitter output signal is $y(t) = x(t)$. At transmission of "0" the switch is in the released position and $y(t) = 0$.

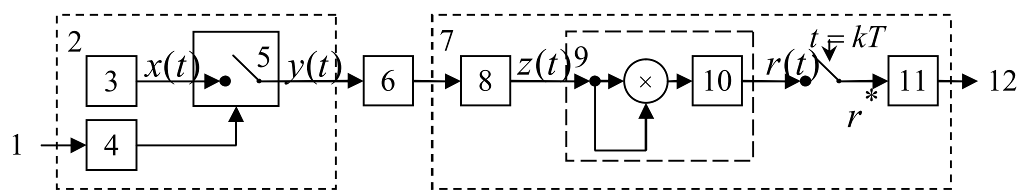

FIG. 1: Block diagram of the communication using ANSK: 1 – transmitted bit; 2 – transmitter; 3 – NSG; 4 – encoder; 5 – switch; 6 – channel; 7 – receiver; 8 – bandpass filter; 9 – demodulator; 10 – integrating unit; 11 – detector; 12 – received bit

Via the communication channel 6 the signal $y(t)$ is fed to the input of the receiver 7. We shall apply the simplest classical model of the loss-free channel where there are operating additive interferences in the form of Gaussian white noise with the one-side spectral density N_0. The additive mix of the payload signal $y(t)$ and the interference (noise) propagate through the bandpass filter 8 with the throughput bandwidth $\left[f_1; f_2 \right]$. At the output of the filter we have the signal

$$z(t) = y(t) + n(t), \tag{1}$$

where $n(t)$ is the noise propagated through the input filter. Then the signal $n(t)$ is the realization of Gaussian random process with the restricted frequency bandwidth

$[f_1; f_2]$, zero mathematical expectation and the dispersion $D_n = N_0 \cdot F$, where $F = f_2 - f_1$ is the width of the signal frequency band.

The signal $z(t)$ is fed to the input of the demodulator 9 containing the squaring unit and the integrating unit 10, at which the integration period length corresponds to the length of the bit interval T. At the demodulator output the random function is observed

$$r(t) = \int_{t-T}^{t} z^2(\tau)\, d\tau .$$

At the end of each bit interval (at $t = kT$) the switch controlled by the synchronization system, performs strobbing of the function $r(t)$. The obtained in the above manner value of $r^* = r(kT)$ enters the input of the threshold detector 11. If $r^* > \gamma$ then it is accepted the hypothesis H_1 about receiving of the bit “1”, otherwise, if $r^* < \gamma$, – the hypothesis H_0 about receiving of the bit “0”.

In its unfolded representation the value r^* under the condition of transmitting “0” and “1” can be put down as follows

$$r^*(0) = \int_{(k-1)T}^{kT} n^2(t)\, dt , \qquad r^*(1) = \int_{(k-1)T}^{kT} (x(t) + n(t))^2\, dt . \tag{2}$$

Under the condition of equiprobable occurrence of “0” and “1” in the input bit flow and the simple loss function, the information bit error probability during transmission/reception is calculated upon the formula

$$P_b = \frac{1}{2} \cdot \left(P\left(r^*(0) > \gamma\right) + P\left(r^*(1) < \gamma\right) \right), \tag{3}$$

where the optimal threshold value γ is determined from the condition $P_b \to \min$.

Let us calculate the bit error probability P_b. For that purpose we proceed to the discrete time scale. Considering that the signal (1) possesses a limited spectrum then under the Kotelnikov theorem that signal can be represented by its sampling counts following one another with the period of $\Delta t = 1/(2F)$. $N = 2FT = 2B$ counts will be positioned along one bit interval. The value of $B = FT$ is called the bandwidth-delay product.

Under the discrete time scale the values (2) have the following analogs

$$r^{*\Delta}(0) = \sum_{i=(k-1)N+1}^{kN} n_i^2 , \qquad r^{*\Delta}(1) = \sum_{i=(k-1)N+1}^{kN} \left(x_i + n_i\right)^2 . \tag{4}$$

To simplify the computing we omitted the multiplier Δt in the formulas (4), which is common for the values $r^{*\Delta}(0)$ and $r^{*\Delta}(1)$. This exerts no influence upon calculation of the bit error probability P_b.

Based on the assumption about the properties of the signal $n(t)$ the counts $n_i = n\left(\left(i-1\right)\Delta t\right)$, $i = 1,2,...$ are independent Gaussian random values with zero mathematical expectation and the dispersion D_n. In this case the probability distribution density function for the value $r^{*\Delta}(0)$ can be obtained by scaling of the distribution density χ^2:

$$p_{r(0)}(v) = \frac{1}{D_n} \cdot \frac{1}{2^{\frac{N}{2}} \cdot \Gamma\left(\frac{N}{2}\right)} \cdot \exp\left(-\frac{v}{2D_n}\right) \cdot \left(\frac{v}{D_n}\right)^{\frac{N}{2}-1} =$$

$$= \frac{1}{D_n} \cdot \frac{1}{2^B \cdot \Gamma\left(B\right)} \cdot \exp\left(-\frac{v}{2D_n}\right) \cdot \left(\frac{v}{D_n}\right)^{B-1}, \tag{5}$$

where $\Gamma(\cdot)$ is the Gamma function.

If it is assumed that distribution of the stochastic process generating the signal $x(t)$ is also of the Gaussian mode then the counts $z_i = x_i + n_i$ of the signal arriving at the input of the demodulator will be independent realizations of the Gaussian random value with zero mathematical expectation and the dispersion equal to $D_x + D_n$. Then similar to (5) we obtain

$$p_{r(1)}(v) = \frac{1}{D_x + D_n} \cdot \frac{1}{2^B \cdot \Gamma\left(B\right)} \cdot e^{-\frac{v}{2(D_x+D_n)}} \cdot \left(\frac{v}{D_x + D_n}\right)^{B-1}. \tag{6}$$

The optimal threshold γ will be determined in this case as the solution to the equation

$$p_{r(1)}(v) = p_{r(0)}(v). \tag{7}$$

Substituting into (7) the right-hand side parts of the equations (5) and (6) and solving the obtained equation we find

$$\gamma = \frac{2BD_n\left(D_x + D_n\right)}{D_x} \cdot \ln\frac{D_x + D_n}{D_n}. \tag{8}$$

Then the equation (3) could be rewritten in the suitable representation for calculations

$$P_b = \frac{1}{2}\left(\int_0^{\gamma} p_{r(1)}(v)\, dv + \int_{\gamma}^{+\infty} p_{r(0)}(v)\, dv \right). \tag{9}$$

Analyzing the expression (9) as well as the expressions (5), (6) and (8) included therein we can draw the conclusion that for the communications using ANSK the bit error probability depends in a rather complicated manner upon the signal power D_x and the noise power D_n, and upon the bandwidth-delay product B too. More detailed studies show that the bit error probability P_b is the function of two independent parameters – the signal-to-noise ratio upon power $\rho^2 = D_x / D_n$ and the bandwidth-delay product B. The results of calculation of P_b as the function of ρ^2 and B are illustrated in Fig. 2(a).

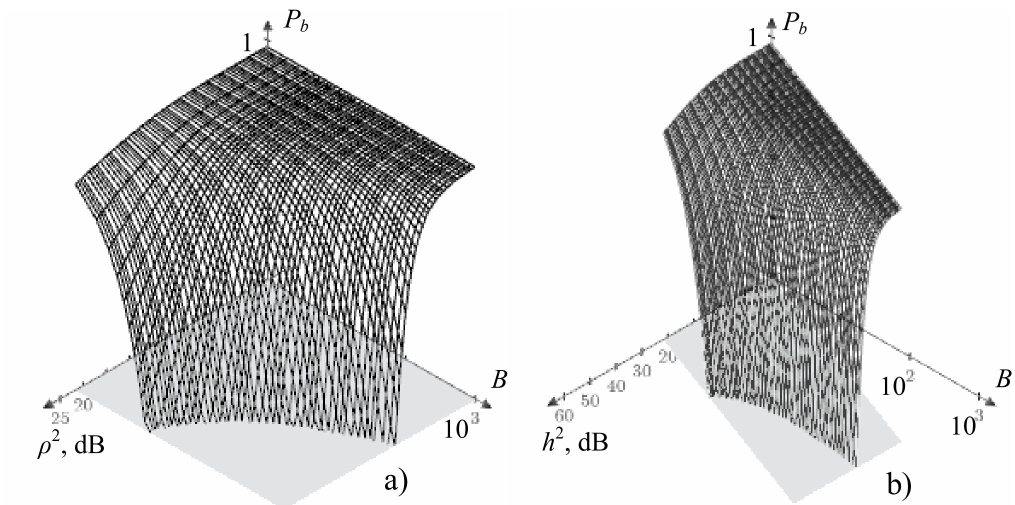

FIG. 2: Bit error probability in the communications using ANSK: $P_b = P_b\left(\rho^2, B\right)$ dependence (a); $P_b = P_b\left(h^2, B\right)$ dependence (b)

However, for the classical digital communication systems (with binary correlation reception of the signals) the normalized value of the signal-to-noise ratio [15] is the traditional indicator of system performance

$$h^2 = \frac{E_b}{N_0} = \frac{D_x \cdot T}{D_n / F} = \frac{D_x}{D_n} \cdot FT = \rho^2 \cdot B .$$

In such systems the bit error probability can be put down as the single argument function h^2. To provide for facilitation of comparison of the obtained results with the classic ones we proceed to the system of coordinates $\left(h^2, B\right)$. The fragment of the surface $P_b = P_b\left(h^2, B\right)$ is shown in Fig. 2(b).

We investigate the cross-sections of the surface $P_b = P_b\left(h^2, B\right)$ at the fixed B (Fig. 3(a)) and the fixed h^2 (Fig. 3(b)).

From Fig. 3(a) it is evident that with the increase of the bandwidth-delay product B the curves $P_b = P_b\left(h^2\right)$ vary their shapes: they decrease slower at small h^2 and faster – at large h^2. As the result, there exists a single value of h^2, at which the curve $P_b = P_b(h^2)$ correspondent to the set value of B provides for the minimal value of the bit error probability P_b among the entire family of curves.

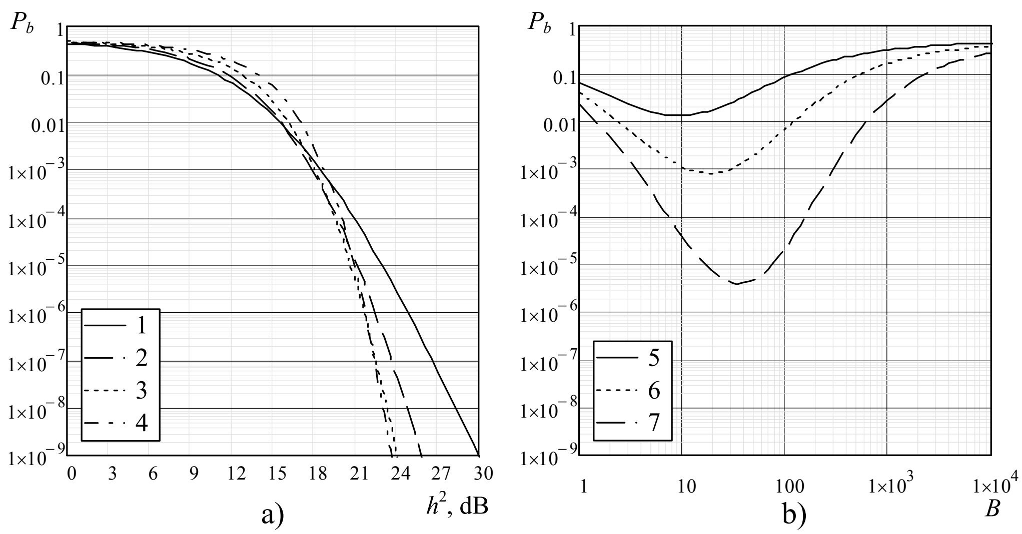

FIG. 3: Dependence of $P_b = P_b\left(h^2\right)$ at the fixed bandwidth-delay product B (1 - $B = 8$; 2 - $B = 16$; 3 - $B = 32$; 4 - $B = 64$) (a) and dependence of $P_b = P_b(B)$ at the fixed h^2 (5 - $h^2 = 15.051$ dB, 6 - $h^2 = 18.062$ dB, 7 - $h^2 = 21.072$ dB) (b) for the communication using ANSK

The same result can be interpreted as the existence of the sole minimum of the bit error probability P_b upon the curves $P_b = P_b(B)$ at any fixed h^2 that is clearly illustrated in Fig. 3(b). The value of the bandwidth-delay product, at which the minimum of the bit error probability is attained for the set h^2, is called the optimal bandwidth-delay product and designated as B_{opt}.

The calculations demonstrate that there exists a simple linear correlation between the value of h^2 and the optimal bandwidth-delay product B_{opt}. For the communications using ANSK we have

$$B_{opt} \approx 0.281 \cdot h^2 . \tag{10}$$

Taking into account the equation (10) and the fact that $h^2 = \rho^2 \cdot B$ it seems reasonable to introduce into the consideration the value of $\rho_{opt}^2 = 1/0.281 = 3.556$ as the optimal signal-to-noise ratio upon power. The value of ρ_{opt}^2 depends solely upon the shift-keying technique of the noise signal.

Now we can formulate the algorithm for setting of optimal parameters in the communication system. Let there be set the signal frequency bandwidth F and the maximally allowable value of the bit error probability $P_{b\ \max}$. Thus in order to attain the most power efficient operation mode of the system it is necessary to perform the following stages:

1) measure the power of the noise D_n;

2) adjust the signal power D_x in a way that the signal-to-noise ratio upon power at the receiver input was equal to ρ_{opt}^2;

3) select the bit interval duration T sufficient for obtaining of the signal-to-noise ratio $h^2 = \rho_{opt}^2 \cdot F \cdot T$ providing for the bit error probability $P_b \le P_{b\ \max}$.

Therefore, optimal adjustment of the value of the signal-to-noise ratio h^2 has to be performed by means of variation of the bandwidth-delay product at the fixed value of the signal power D_x providing for fulfillment of the equation $\rho^2 = \rho_{opt}^2$. In that case, any selected value of the bandwidth-delay product B will be optimal for the resulting value of h^2.

It should be noted that the system performance of the communications using ADNSK is substantially dependent upon the probability distribution of the stochastic process, the realization of which is represented by the signal $x(t)$. This fact is easily clarified if we calculate mathematical expectation and dispersion of the value $r^{*\Delta}(1)$:

$$M_{r(1)} = N\left(D_x + D_n\right),\ D_{r(1)} = N\left(D_x^2\left(\gamma_4^{(x)} + 2\right) + 4D_x D_n + 2D_n^2\right),$$

where $\gamma_4^{(x)}$ is the fourth cumulative coefficient or the coefficient of excess [16] of the signal $x(t)$. Hence it is clear that decreasing of $\gamma_4^{(x)}$ results in decreasing of dispersion of the value $r^{*\Delta}(1)$ and, thus, in decreasing of the bit error probability P_b.

At the Gaussian distribution of $x(t)$ the coefficient $\gamma_4^{(x)} = 0$. The coefficient of excess attains its minimal value $\gamma_4^{(x)} = -2$ for distribution of the binary alternative, the probability density of which has the following form

$$p_x(\xi) = \frac{1}{2}\left(\delta\left(\xi + \sqrt{D_x}\right) + \delta\left(\xi - \sqrt{D_x}\right)\right),$$

where $\delta(\cdot)$ is the Dirac delta function.

In the general case (for an arbitrary distribution of $x(t)$) it is impossible to obtain a compact expression for the density $p_{r(1)}(\nu)$ similar to (6).

We designate (same as before) $p_x(\xi)$ – the probability distribution density function of independent counts of the signal $x(t)$. Assuming that the noise still has the Gaussian distribution, for one count we obtain the density of

$$p_{r(1)}(\nu)\Big|_{N=1} = \frac{1}{\sqrt{2\pi D_n \nu}} \int_{-\infty}^{+\infty} p_x(\xi) \cdot \exp\left(-\frac{\left(\sqrt{\nu} - \xi\right)^2}{2D_n}\right) d\xi. \qquad (11)$$

The density $p_{r(1)}(\nu)$ for an arbitrary N can be found with the help of the set of characteristic functions [17].

The characteristic function of the random value $r^{*\Delta}(1)$ at $N = 1$ is determined as the inverse Fourier transform for the function $p_{r(1)}(\nu)\Big|_{N=1}$

$$\varphi(t)\Big|_{N=1} = \int_{-\infty}^{\infty} e^{it\nu} \cdot p_{r(1)}(\nu)\Big|_{N=1} d\nu. \qquad (12)$$

It is important to mention that the characteristic function of the sum of independent random values is equal to the product of the respective characteristic functions, thus

$$\varphi(t)\Big|_{N=k} = \varphi^k(t)\Big|_{N=1}. \qquad (13)$$

The unicity theorem and the inverse transform theorem [17] are true for the characteristic function, therefore, the density function is determined unambiguously

$$p_{r(1)}(\nu)\Big|_{N=k} = \frac{1}{2\pi} \int_{-\infty}^{\infty} e^{-it\nu} \varphi(t)\Big|_{N=k} dt. \qquad (14)$$

Sequential realization of calculations upon the formulas (11)-(14) solves the set problem of finding the function $p_{r(1)}(v)$.

Having in our disposal the expressions (5) and (14) for the functions $p_{r(0)}(v)$ and $p_{r(1)}(v)$ correspondingly, subsequently from (7) we find the optimal threshold value γ and calculate the bit error probability according to (9). Therefore, we can obtain estimation of the system performance of the communication using ADNSK for an arbitrary probability distribution of the noise signal generator.

3. COMMUNICATION USING FREQUENCY NOISE SHIFT-KEYING

Now let us consider the communication using the frequency noise shift-keying (FNSK). The relevant block diagram is provided in Fig. 4.

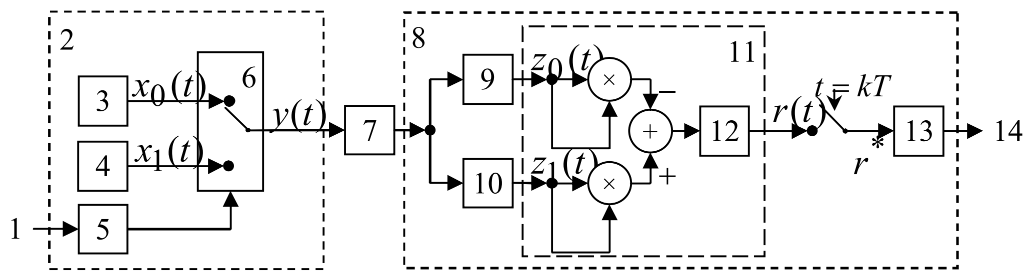

FIG. 4: Block diagram of the communication using FNSK: 1 – transmitted bit; 2 – transmitter; 3 – NSG-0; 4 – NSG-1; 5 – encoder; 6 – switch; 7 – channel; 8 – receiver; 9, 10 – bandpass filters; 11 – demodulator; 12 – integrating unit; 13 – detector; 14 – received bit

The transmitter of the communication using FNSK is equipped with two noise signal generators NSG-0 and NSG-1, which are designated in Figs. 3 and 4. The output signals of the generators – $x_0(t)$ and $x_1(t)$ – occupy different frequency bandwidths $[f_{01}, f_{02}]$ and $[f_{11}, f_{12}]$ correspondingly. To simplify our computing we shall assume that the powers of the signals are equal and amount to D_x, the bandwidths are not overlapping and have equal widths $F = f_{02} - f_{01} = f_{12} - f_{11}$.

The switch 6 controlled by the signal from the encoder supplies to the transmitter output the fragment of the signal $x_0(t)$ with the duration T at transmission of the bit "0" and the signal $x_1(t)$ at transmission of "1". The channel model remains unchanged.

The input signal of the transmitter arrives upon two bandpass filters 9 and 10. The throughput bandwidths of the filters are $[f_{01}, f_{02}]$ and $[f_{11}, f_{12}]$ correspondingly. Thus, at transmission of, for example, the bit "1" the following signal is observed at the output of the filter 10

$$z_1(t) = x_1(t) + n_1(t),$$

where $n_1(t)$ is the noise that underwent filtering within the throughput bandwidth $[f_{11}, f_{12}]$. At that, at the output of the filter 9 there will be the signal $n_0(t)$ representing the noise within the frequency bandwidth $[f_{01}, f_{02}]$.

Therefore, at the end of the k -th bit interval the following value will be fed to the input of the detector 13 at transmission of “0”

$$r^*(0) = \int_{(k-1)T}^{kT} \left(n_1^2(t) - \left(x_0(t) + n_0(t) \right)^2 \right) dt,$$

and at transmission of “1”

$$r^*(1) = \int_{(k-1)T}^{kT} \left(\left(x_1(t) + n_1(t) \right)^2 - n_0^2(t) \right) dt. \tag{15}$$

If the probabilities distribution of stochastic signals $x_0(t)$, $x_1(t)$, $n_0(t)$ and $n_1(t)$ are symmetrical with respect to zero (and we shall assume them to be of that kind), then distributions of the values $r^*(0)$ and $r^*(1)$ are symmetrical towards each other with respect to zero. Hence it follows that the optimal threshold level of the comparator at reception of the signals with FNSK is equal to zero and does not depend upon the noise level in the communication channel. Besides, to calculate the bit error probability of the receiver it would be sufficient to investigate one of the values only, for example, $r^*(1)$.

Under the discrete time scale the expression (15) acquires the following form

$$r^{*\Delta}(1) = \sum_{i=(k-1)N+1}^{kN} \left(\left(x_{1i} + n_{1i} \right)^2 - n_{0i}^2 \right). \tag{16}$$

Let as assume that the signals $x_0(t)$ and $x_1(t)$ have the Gaussian distribution. Then the function of probability distribution density of the value $r^{*\Delta}(1)$ determined by the equation (16) will have the following representation

$$p_{r(1)}(v) = \frac{1}{\left(D_x + D_n\right) D_n} \cdot \frac{1}{2^{2B}\, \Gamma^2(B)} \times$$
$$\times \int_0^{+\infty} \exp\left(-\frac{\theta}{2\left(D_x + D_n\right)} - \frac{\theta - v}{2D_n} \right) \cdot \left(\frac{\theta \cdot (\theta - v)}{\left(D_x + D_n\right) D_n} \right)^{B-1} \cdot \eta(\theta - v)\, d\theta,$$

where $\eta(t) = \begin{cases} 0, & t < 0, \\ 1, & t \ge 0 \end{cases}$ is the Heaviside function (the inclusion function).

Considering the symmetry of the densities $p_{r(1)}(\nu)$ and $p_{r(0)}(\nu)$, the bit error probability P_b is calculated upon the formula

$$P_b = \frac{1}{2}\left(\int_{0}^{+\infty} p_{r(0)}(\nu)\, d\nu + \int_{-\infty}^{0} p_{r(1)}(\nu)\, d\nu \right) = \int_{-\infty}^{0} p_{r(1)}(\nu)\, d\nu \,. \qquad (17)$$

Like in the case with ADNSK the bit error probability P_b of the function of the signal-to-noise ratio h^2 and the bandwidth-delay product B. Figure 5 provides cross-sections of the surface $P_b = P_b\left(h^2, B\right)$ at the fixed B and at the fixed h^2.

From Fig. 5 it is evident that the dependence of P_b upon h^2 and B for the communication using FNSK possesses the same particularities than for the communication using ADNSK (cf. Fig. 3). In particular, the definitions of the optimal bandwidth-delay product B_{opt} and the optimal signal-to-noise ratio upon power ρ^2_{opt} remain actual. The differences occur in numerical correlations, thus, for the communication using FNSK we have

$$B_{opt} \approx 0.313 \cdot h^2 \,,$$

whence $\rho^2_{opt} = 3.2$ that is by 11% less than for the communication using ANSK.

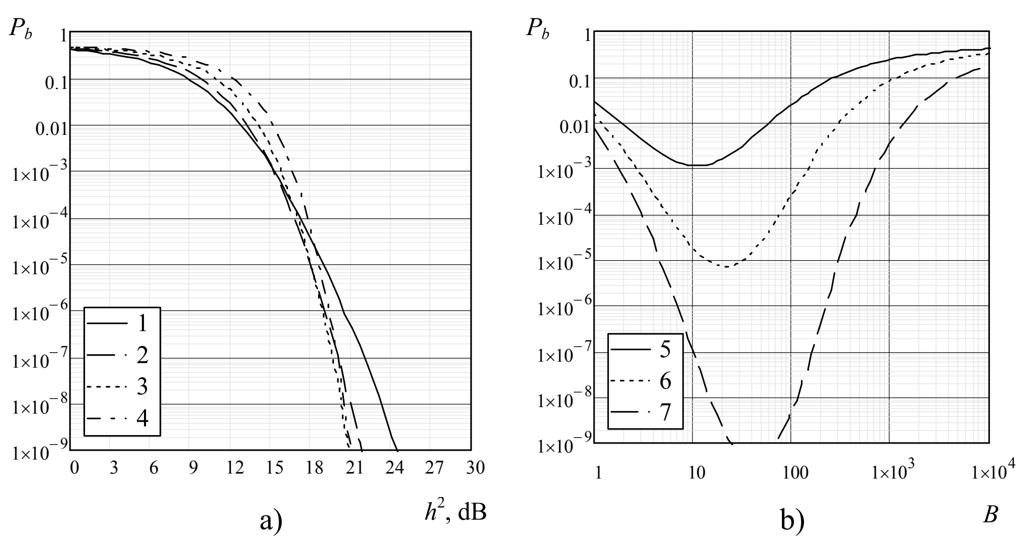

FIG. 5: Dependence of $P_b = P_b\left(h^2\right)$ at the fixed bandwidth-delay product B (1 - $B = 8$; 2 - $B = 16$; 3 - $B = 32$; 4 - $B = 64$) (a) and dependence of $P_b = P_b(B)$ at the fixed h^2 (5 - $h^2 = 15.051$ dB, 6 - $h^2 = 18.062$ dB, 7 - $h^2 = 21.072$ dB) (b) for the communication using FNSK

The above calculations are true for the Gaussian noise carrier. But if the NSG-0 and NSG-1 possess other probability distributions then it would be necessary to begin calculation of the bit error probability with finding of the distribution density function $p_{r(1)}(\nu)\big|_{N=1}$ for one count similar to (11). Then, using the method of characteristic functions (formulas (12)-(14)) we find the function $p_{r(1)}(\nu)$ for an arbitrary N, and finally we calculate the bit error probability P_b upon the formula (17).

4. COMPARATIVE ANALYSIS OF THE ERROR PERFORMANCES

Comparison of system performances of the considered communications at optimal parameters, i.e., under the most power-efficient operation mode, represents a special interest. Considering that in this case the value of the signal-to-noise ratio upon power is fixed $\rho^2 = \rho_{opt}^2$, then the error performance is characterized by one curve $P_b = P_b\left(h^2\right)$ only; in so doing the increase of the parameter h^2 occurs here due to the increase of the value of the bandwidth-delay product B (Fig. 6).

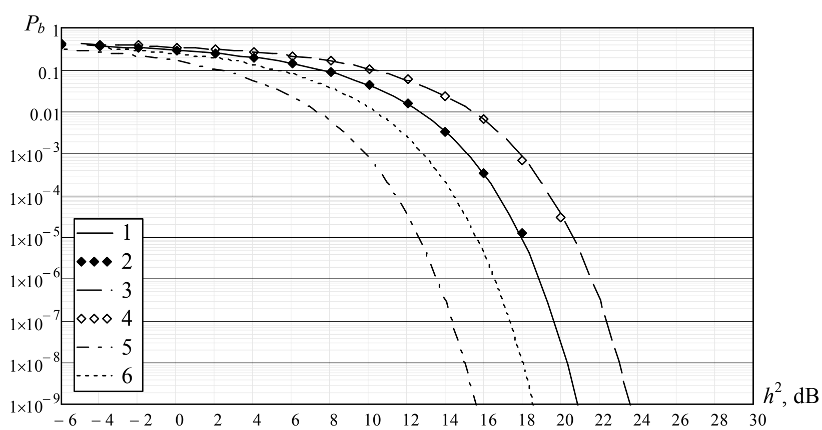

FIG. 6: Dependences of $P_b = P_b\left(h^2\right)$ for FNSK (1 - theory, 2 - experiment) and ANSK (3 - theory, 4 - experiment) at optimal parameters. For the purpose of comparison the curves for the frequency (5) and the amplitude (6) noise shift-keying of the deterministic signal are provided

It should be noted that the curves $P_b = P_b\left(h^2\right)$ obtained at optimal parameters have the classical shape that provides for an explicit demonstration of the power loss of one system with respect to the other.

Analysis of the results shows that the communication using ANSK is losing 2.773 dB to the communication using FNSK. Whereas, FNSK is losing 5.289 dB to the classical communication using the frequency noise shift-keying (curve 5 in Fig. 6), and ANSK is losing 5.051 dB to the communication using the amplitude noise shift-keying (curve 6 in Fig. 6) of the deterministic signal.

Figure 6 provides also the estimates of the bit error probability in the communications using ADNSK and FNSK obtained with the help of a series of the computer simulation. It is evident from the Figure that the experimental results form a rather accurate match with the results of theoretical calculations that witnesses for adequacy of the suggested models and calculation techniques.

5. CONCLUSIONS

The mathematical models suggested in this paper allow performing an adequate estimation and a detailed study of error immunity values of the communications using power reception of noise signals.

Communications using ADNSK and FNSK discussed above possess a number of particular features that make them essentially different from the classical communication systems with binary correlation reception of the deterministic signals. The information bit error probability during transmission/reception for the above kind of communications depends not only upon the signal-to-noise ratio but upon the bandwidth-delay product as well. There occur specific definitions of the optimal bandwidth-delay product and the optimal signal-to-noise ratio upon power.

The structure of transmitter/receiver units applied in the communication using ADNSK is simpler compared to that applied in the communication using FNSK. However, ANSK loses to FNSK in terms of system performance. Besides, the optimal threshold level of the comparator in the receiver of the communication using ADNSK is different from zero and dependent upon the noise power. Therefore, setting of the optimal threshold requires additional efforts at reception of the signal.

Comparatively simple means of formation of the broadband noise signal in combination with the incoherent power reception make the communications using ADNSK and FNSK quite attractive for development of non-expensive and resistant to unfavorable for signal propagation operational environment data telecommunication systems, despite the fact that they demonstrate a quite substantial loss in terms of the system performance.

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