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COMPRESSION OF COMMUNICATION CHANNELS BASED ON SUPER-RAYLEIGH RESOLUTION OF SIGNALS IN TERMS OF TIME OF ARRIVAL

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The paper considers several methods for time compression of narrow-band communication lines based on the super-Rayleigh resolution of pulse signals.

The main way of meeting the growing requirements for capacity of communication channels consists in expanding their bandwidth. On the other hand, the use of procedures of super-Rayleigh resolution of signals opens the possibility for an alternative approach oriented to comparatively narrow-band data-handling buses.

The purpose of this work is to outline the conceptual basis of the time compression of information flows by using the methods of super-Rayleigh resolution of pulse signals in terms of their time of arrival.

The circuitry selected for realization of the procedures suggested below is the Software Radio technology [1] as applied to a cable communication channel (Fig. 1). Here we consider the cases of strict synchronization of communication channels and asynchronous mode of reception, which are characterized by quite different principles of signal processing.

In the first case the synchronizing packet of pulses makes it possible to perform, with an accuracy of a quantization period, the fixing of time position of the information bits. It means that the time of arrival of the message signals is known, provided that the transmitting side forms the signals in the same time scale as taken for ADC clock generator in the message receiver. If these assumptions are consistent then the time separation of the pulse signals of a transmitter's coded message can be carried out with regard to their subsequent super-Rayleigh resolution in the receiver, when the envelope of every pulse signal is formed in conformity with a prescribed rule.

If, in the simplest case, we are oriented to the amplitude coding of data messages, when some discrete values of signal amplitude are put into one-to-one correspondence with a particular coded combination of information symbols, then the decoding procedure in the receiver can be reduced to a trivial estimation of the amplitude of every video pulse in a multi-signal packet. In the deterministic approach the required amplitudes can be determined from an expression [2]

$$\tilde{a}_m = \frac{\det_m}{\det}, \quad (1)$$

where $\det = \begin{vmatrix} K(s_1 - z_1) & K(s_1 - z_2) & \dots & K(s_1 - z_M) \\ K(s_2 - z_1) & K(s_2 - z_2) & \dots & K(s_2 - z_M) \\ \dots & \dots & \dots & \dots \\ K(s_M - z_1) & K(s_M - z_2) & \dots & K(s_M - z_M) \end{vmatrix}$, $m = 1, 2, \dots, M$; $\det_m$ is the partial determinant obtained from det by replacing the respective column by the vector of free terms $[B] = [U_1\ U_2\ U_3 \dots\ U_M]^T$, where M is the number of pulses in the data message and U_n is the nth voltage sample from the whole number taken for processing at the output of the

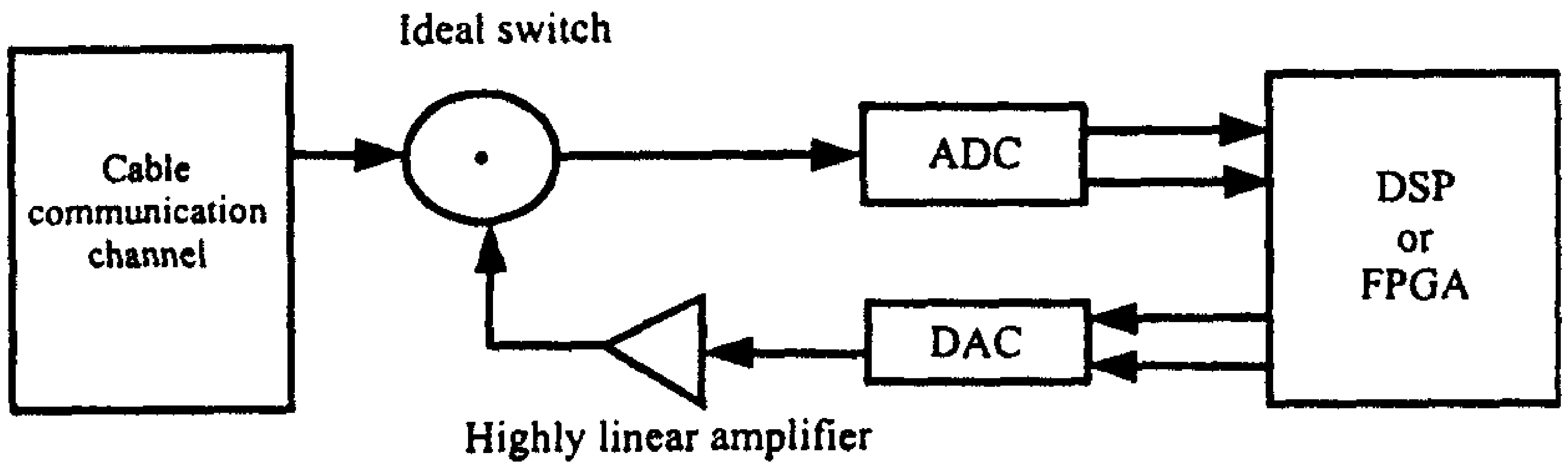

Fig. 1

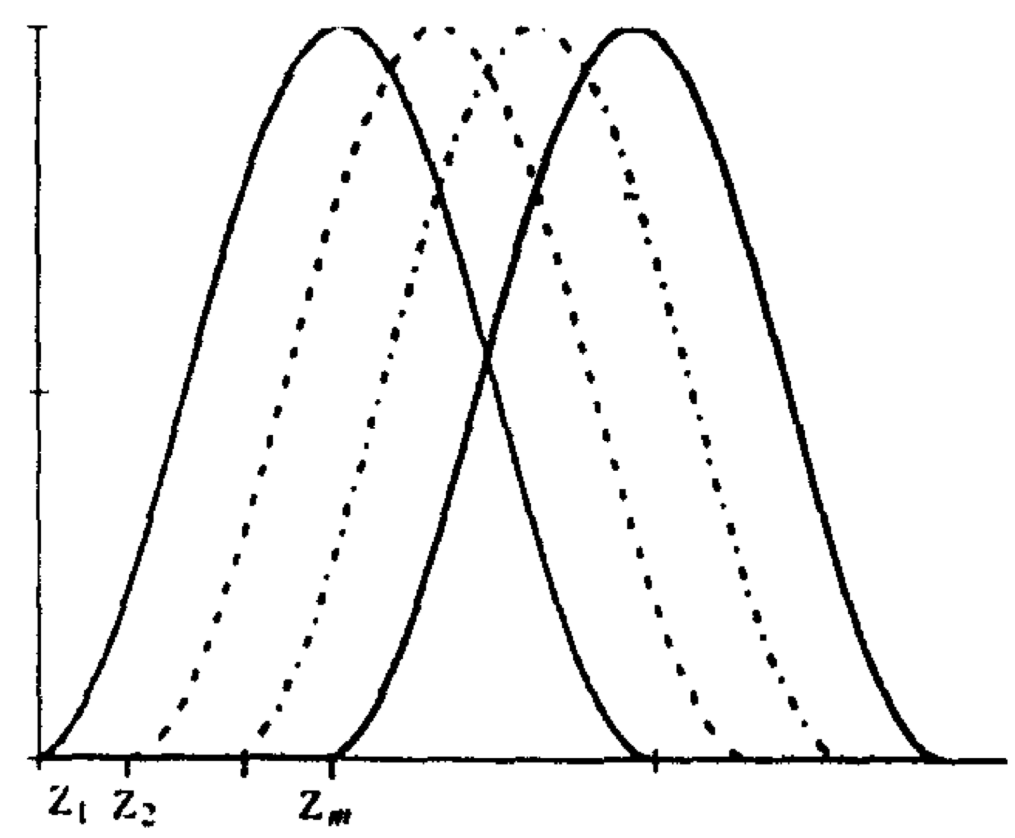

Fig. 2

analog-digital converter (ADC); $K(s_n - z_m)$ is the normalized discrete envelope of the pulse signal for the ADC mth sample; s_n is the ordinal number of the ADC reading in a signal sample expressed in the sampling periods; and z_m is the time position of the mth pulse expressed in the ADC sampling periods (Fig. 2).

The potential accuracy of determination of pulse signal amplitudes in the lock-in mode of operation for a prescribed envelope function depends on the noise variance, separation of the pulses in time, and on their duration in ADC readings. It should be noted that, as follows from the Cramer-Rao lower bound, pulse amplitudes have no effect on the absolute error of amplitude measurement.

In asynchronous mode of operation the peculiar feature of signal reception consists in unknown time shift of the first signal of the pulse packet with respect to the starting point of the measurement sample. In this case the shift of all the rest of the pulses with respect to the first one is assumed known. For such reception conditions the task of signal amplitude estimation has to be preceded by measurement of time of arrival of the first pulse in the coded message. The respective measurement procedure can be easily synthesized by analogy with [3] based on the maximum likelihood method. In the general case, for M video pulses the estimate of time position of the first signal in the packet can be obtained by some iterative procedure, based on minimization of the function

$$F_M = -\begin{vmatrix} 0 & W_1 & \cdots & W_M \\ W_1 & Q_{11} & \cdots & Q_{1M} \\ \vdots & \vdots & \cdots & \vdots \\ W_M & Q_{M1} & \cdots & Q_{MM} \end{vmatrix} = \max \qquad (2)$$

where

$$W_n = \sum_{s=z_n}^{z_n+N-1} U_s \cdot K_{sn}, Q_{mn} = Q_{nm} = \sum_{s=z_m}^{z_n+N-1} K_{sm} K_{sn},$$

$$Q_{11} = Q_{mm} = \ldots = Q_{MM} = \sum_{s=z_1}^{z_1+N-1} K_{s1}^2 = \sum_{s=z_m}^{z_m+N-1} K_{sm}^2,$$

$K_{sm} = \begin{cases} K(s - z_1 - \Delta_m) & \text{at} \quad z_1 + \Delta_m \le s \le z_1 + \Delta_m + N \\ 0 & \text{at} \quad s < z_1 + \Delta_m, s > z_1 + \Delta_m + N \end{cases}$ is the normalized discrete function of the envelope, s is the ordinal number of an ADC reading in a signal sample expressed in the sampling periods; z_1 is the first of ADC readings within the limits of existence of the first signal in the packet, N is the length of pulses in the ADC samples; and Δ_m is the known time shift of the mth pulse with respect to the first one expressed in the periods of ADC samplings.

The values of time of arrival of each signal produced in the course of maximization (2) should be substituted in (1) to calculate the amplitude estimates.

The process of message decoding reduces to comparison of signals' measured amplitudes with the preset threshold levels. When the physical carriers of information are radio pulses, the channel traffic capacity can be improved by the amplitude-phase coding of messages, when we establish a correspondence between quadrature components of signal amplitudes and specially coded packets. The relevant operation of forming the quadrature components of signal voltages in the receiver may be implemented either in analog or in digital form. In the former case we merely perform the analog multiplication of the received signals by the reference oscillation of the given frequency followed by analog-digital conversion of each quadrature component. Since, depending on the proportion between the carrier frequencies of the received signals and of the reference oscillation, the quadratures produced will take the form of radio or video pulses, the relationships for the amplitude component estimates will somewhat differ from those considered above.

At the video-signal representation of the quadratures, expression (1) reduces to the form [4]

$$\tilde{a}_m^{c(s)} = \frac{\det_m^{c(s)}}{\det} \tag{3}$$

where the determinant det remains identical to that used in (1); $\det_m^{c(s)}$ is the partial determinant produced from det by replacing the respective column by the vector of free terms $[B^c] = [U_1^c U_2^c U_3^c \ldots U_M^c]^T$ or $[B^s] = [U_1^s U_2^s U_3^s \ldots U_M^s]^T$; U_n^c, U_n^s is the nth, among M samples taken for processing, cosine or sine quadrature component of voltage at the output of the analog-digital converter (ADC).

Optimization of amplitude estimates based on the maximum likelihood method, with the video-pulse format of the quadratures, leads to the following relation [4]:

$$\tilde{a}_{m_{\text{opt}}}^{c(s)} = \frac{\det_m^{c(s)}}{\det} \tag{4}$$

where

$$\det = \begin{vmatrix} \sum\limits_{s=z_1}^{z_1+N-1} K_{s1}^2 & \sum\limits_{s=z_2}^{z_1+N-1} K_{s1} K_{s2} & \cdots & \sum\limits_{s=z_M}^{z_1+N-1} K_{s1} K_{sM} \\ \sum\limits_{s=z_2}^{z_1+N-1} K_{s1} K_{s2} & \sum\limits_{s=z_2}^{z_2+N-1} K_{s2}^2 & \cdots & \sum\limits_{s=z_M}^{z_2+N-1} K_{s2} K_{sM} \\ \cdots & \cdots & \cdots & \cdots \\ \sum\limits_{s=z_M}^{z_1+N-1} K_{s1} K_{sM} & \sum\limits_{s=z_M}^{z_2+N-1} K_{s2} K_{sM} & \cdots & \sum\limits_{s=z_M}^{z_M+N-1} K_{sM}^2 \end{vmatrix},$$

and $\det_m^{c(s)}$ is a partial determinant obtained from det by replacing the respective column by the vector of free terms $[B^c] = [W_1^c W_2^c W_3^c \ldots W_M^c]^T$ or $[B^s] = [W_1^s W_2^s W_3^s \ldots W_M^s]^T$; $\dot{W}_n = \sum\limits_{s=z_n}^{z_n+N-1} (U_s^c + \mathrm{j} \cdot U_s^s) K_{sn}$.

In the case of radio-pulse variant of quadrature components' generation, in relation (4) we introduce the determinant

$$\det = \begin{vmatrix} Q_{11} & Q_{12} & Q_{13} & \cdots & Q_{1M} \\ Q_{21} & Q_{22} & Q_{23} & \cdots & Q_{2M} \\ \vdots & \vdots & \vdots & \cdots & \vdots \\ Q_{M1} & Q_{M2} & Q_{M3} & \cdots & Q_{MM} \end{vmatrix} \tag{5}$$

and $\det_m^{c(s)}$ obtained from (5) by replacing the respective column $\{Q_{1m}, Q_{2m}, \ldots, Q_{Mm}\}^T$ by the vector of free terms $[B^c] = [W_1^c W_2^c W_3^c \ldots W_M^c]^T$ or $[B^s] = [W_1^s W_2^s W_3^s \ldots W_M^s]^T$, where

$$\dot{W}_n = \sum_{s=z_n}^{z_n+N-1} (U_s^c + \mathrm{j} U_s^s) \overset{*}{K}_{sn}, Q_{mn} = Q_{nm} = \sum_{s=z_n}^{z_n+N-1} \dot{K}_{sm} \overset{*}{K}_{sn},$$

$$\dot{K}_{sm} = K_{sm} \cdot \cos(\omega \cdot \Delta t(s - z_m)) + \mathrm{j} \cdot K_{sm} \sin(\omega \cdot \Delta t\,(s - z_m)),$$

$K^{*}_{sm} = K_{sm} \cdot \cos\left(\omega \cdot \Delta t(s - z_m)\right) - \mathrm{j} \cdot K_{sm} \sin\left(\omega \cdot \Delta t(s - z_m)\right)$ is the value which is complex-conjugate to K_{sm}; Δt is the sampling period; z_m is the known location of the mth pulse in time expressed in the sample periods of ADC; $z_m = z_1 + \Delta_m$, and ω is the radio signal carrier frequency. The rest of the notations are the same as in (1)–(4).

The operations of time synchronization at asynchronous reception are modified in the same manner. For the video-pulse representation of quadrature components, the estimate of the time of arrival of the first pulse in the packet can be obtained after maximizing the function [4]

$$F_M = -\begin{vmatrix} 0 & W^{*}_1 & W^{*}_2 & \cdots & W^{*}_M \\ \dot{W}_1 & Q_{11} & Q_{12} & \cdots & Q_{1M} \\ \dot{W}_2 & Q_{21} & Q_{22} & \cdots & Q_{2M} \\ \vdots & \vdots & \vdots & \cdots & \vdots \\ \dot{W}_M & Q_{M1} & Q_{M2} & \cdots & Q_{MM} \end{vmatrix} = \max, \qquad (6)$$

$$\dot{W}_n = \sum_{s=z_n}^{z_n+N-1} \left(U^c_s + \mathrm{j}U^s_s\right) K_{sn}, Q_{mn} = Q_{nm} = \sum_{s=z_m}^{z_n+N-1} K_{sm} K_{sn},$$

$$Q_{11} = Q_{mm} = \ldots = Q_{MM} = \sum_{s=z_1}^{z_1+N-1} K^2_{s1} = \sum_{s=z_m}^{z_m+N-1} K^2_{sm}.$$

In the case of radio pulses, expression (6) has to be modified by the following values:

$$\dot{W}_n = \sum_{s=z_n}^{z_n+N-1} \left(U^c_s + \mathrm{j}U^s_s\right) K^{*}_{sn},\ Q_{mn} = Q_{nm} = \sum_{s=z_m}^{z_n+N-1} \dot{K}_{sm} K^{*}_{sn}. \qquad (7)$$

An alternative method of formation of the quadrature components in the receiver consists in using the discrete Hilbert filtration under the sliding window conditions based on the signal mixture voltages obtained in the course of analog-digital conversion. If we neglect the transient processes in the Hilbert filter [5], the above operations of signal processing remain valid. However, if we have to allow for distortion at the edges of the pulse envelopes arising in one of Hilbert's components [5], then in relationships (5) and (7), instead of the complex envelopes, we must use the expression

$$\tilde{\dot{K}}_{sm} = \left[K^c_{sm} + \mathrm{j}K^s_{sm}\right] \cdot \left[\cos p_{sm} + \mathrm{j}\sin p_{sm}\right] =$$
$$= K^c_{sm} \cos p_{sm} - K^s_{sm} \sin p_{sm} + \mathrm{j}\left[K^s_{sm} \cos p_{sm} + K^c_{sm} \sin p_{sm}\right]$$

while for those complex-conjugate to them,

$$\tilde{K}^{*}_{sm} = \left[K^c_{sm} - \mathrm{j}K^s_{sm}\right] \cdot \left[\cos p_{sm} - \mathrm{j}\sin p_{sm}\right] =$$
$$= K^c_{sm} \cos p_{sm} - K^s_{sm} \sin p_{sm} + \mathrm{j}\left[K^s_{sm} \cos p_{sm} + K^c_{sm} \sin p_{sm}\right]$$

where

$$\dot{K}_{sm} = K^c_{sm} + \mathrm{j} \cdot K^s_{sm} = \begin{cases} K(s - z_m), z_{kf_m} < s < z_{nr_m}, \\ K(s - z_m) + \mathrm{j} \cdot \left(K(s - z_m) + \Delta K(s - z_m)\right), \\ z_m \le s \le z_{kf_m}, z_{nr_m} \le s \le z_{k_m}, \\ 0, s < z_m s > z_{k_m}, \end{cases}$$

s is the ordinal number of the complex sample; $z_m = z_1 + \Delta_m$, where z_m is the number of the first of ADC samples within the limits of existence of the mth pulse; z_{kf_m} is the number of the last sample from ADC obtained at its leading edge; z_{nr_m} is the number of the first sample from ADC during the trailing edge of the mth signal; z_{k_m} is the moment of termination of the mth pulse; $K(s - z_m)$ is the discrete envelope, normed to its maximum, of the mth initial real-valued signal; $\Delta K(s - z_m)$ is the distortion of the normalized discrete function of envelope because of DHT (discrete Hilbert transform) transient process in the sine quadrature on the leading and trailing edges of the mth pulse; and $p_{sm} = \omega \cdot \Delta t\,(s - z_m)$, where Δt is the signal sampling period.

In this case K^c_{sm} and K^s_{sm} describe the variation laws for the envelope of the mth pulse in quadratures. Other notations are as follows:

$$\dot{W}_n = \sum_{s=s_n}^{s_n+N-1} (U^c_s + jU^s_s)\tilde{K}^*_{sn} = \sum_{s=s_n}^{s_n+N-1} \{U^c_s[K^c_{sn}\cos p_{sn} - K^s_{sn}\sin p_{sn}] +$$

$$+U^s_s[K^s_{sn}\cos p_{sn} + K^c_{sn}\cdot\sin p_{sn}] + j(U^s_s[K^c_{sn}\cos p_{sn} - K^s_{sn}\sin p_{sn}] -$$

$$-U^c_s\cdot[K^s_{sn}\cos p_{sn} + K^c_{sn}\sin p_{sn}])\},$$

$$\dot{Q}_{mn} = \sum_{s=s_m}^{s_n+N-1} \tilde{K}_{sm}\,\tilde{K}^*_{sn} = \sum_{s=s_m}^{s_n+N-1} \{[K^c_{sm}K^c_{sn} + K^s_{sm}K^s_{sn}]\cos(p_{sm} - p_{sn}) +$$

$$+[K^c_{sm}\cdot K^s_{sn} - K^c_{sm}\cdot K^c_{sn}]\sin(p_{sm} - p_{sn})\} + j\sum_{s=s_m}^{s_n+N-1} \{[K^c_{sm}\cdot K^c_{sn} + K^s_{sm}\cdot K^s_{sn}]\times$$

$$\times\sin(p_{sm} - p_{sn}) + [K^s_{sm}K^c_{sn} - K^c_{sm}K^s_{sn}]\cdot\cos(p_{sm} - p_{sn})\}.$$

The process of linking, within the conception suggested, is almost identical to traditional approaches. To improve the efficiency under interference conditions, the application of modulated signals must be preceded by estimating the interference situation in the communication line. Thus, in the absence of the transmitter signals, we may perform a measurement of the interference amplitude components and use the measurement results for adaptation of the modulated signal level.

Estimation of a parasitic distortion introduced by the line and arising, for instance, because of rereflections, is best performed by transmitter test signals, whose quadrature components have some fixed values known at the receiving end. For the test we may use a pulse packet with at least two signals of nonzero amplitude.

In order to estimate the limiting potentialities of the suggested approach to improving the capacity of the narrow-band communication channels, one may employ the methods considered in [6], [7] as applied to radar situations. The distinguishing feature is that when we form the Cramer-Rao lower bound for the variance of nonbiased estimates of pulse signal amplitudes depending on synchronous or asynchronous reception conditions, the derivatives of the likelihood functional with respect to time of signal arrival are not used at all, or may be substituted only for the first of the pulses.

As an example of application of this method for analysis of limiting potentialities of the time compression of communication channels under strict lock-in conditions, we can present the results of investigation of the relation between the mean square errors (MSE) in estimating the amplitudes of video pulses whose envelopes $\sin^2 x$ are uniformly distributed in the packet on the interval of existence of the first signal.

Assuming that for reliable decoding of a data message the interval between two neighboring levels of signal amplitudes must exceed six MSE of their estimates (σ_{a_m}), consider a case of a 4-digit amplitude coding, when the whole range of possible values of the normalized amplitudes is broken into 16 steps Δ_{a_m}. At a preset signal-to-noise ratio the maximum attainable compression of the pulses in terms of their time of arrival Δz_m will be achieved by exhaustion of probable time shifts between them until at least in a single channel the condition $\Delta_{a_m} \geq 6\cdot\sigma_{a_m}$ is violated.

For the case of uniform location of three and more pulses in the packet, the major part will be played by MSE of the estimates of amplitudes in the inner group of signals: their measurement conditions, at a large number of superpositions and for a long enough pulse, will be the worst. Moreover, this situation may be taken into account when we select the quantization step of amplitude components in the process of message coding: for the pulses in the middle of the packet the amplitude increments must be set smaller than for the bordering signals. For an 8-pulse packet, for example, we may adaptively pass from 4PSK-modulation in the central group of signals to 16PSK-coding at the periphery.

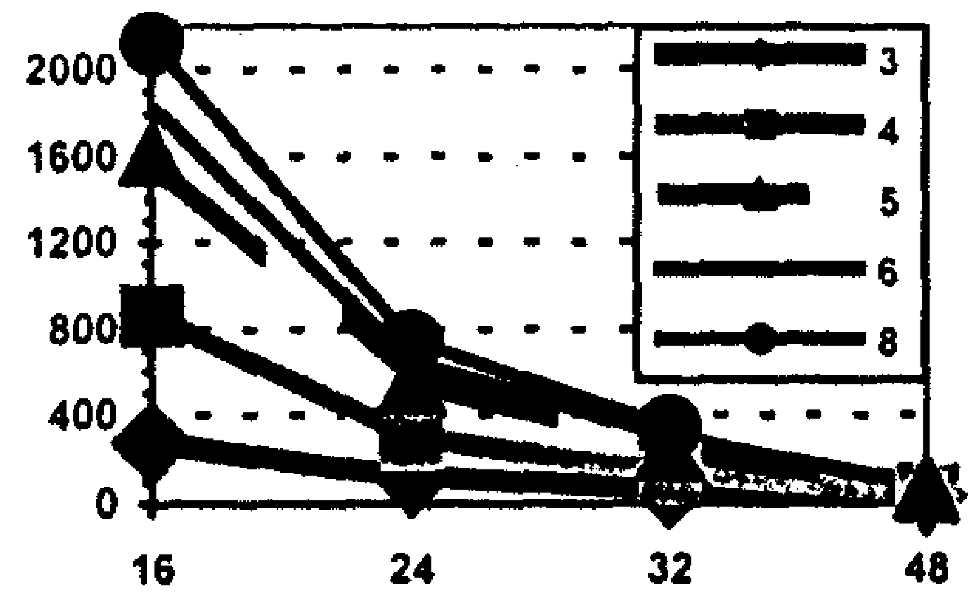

Fig. 3

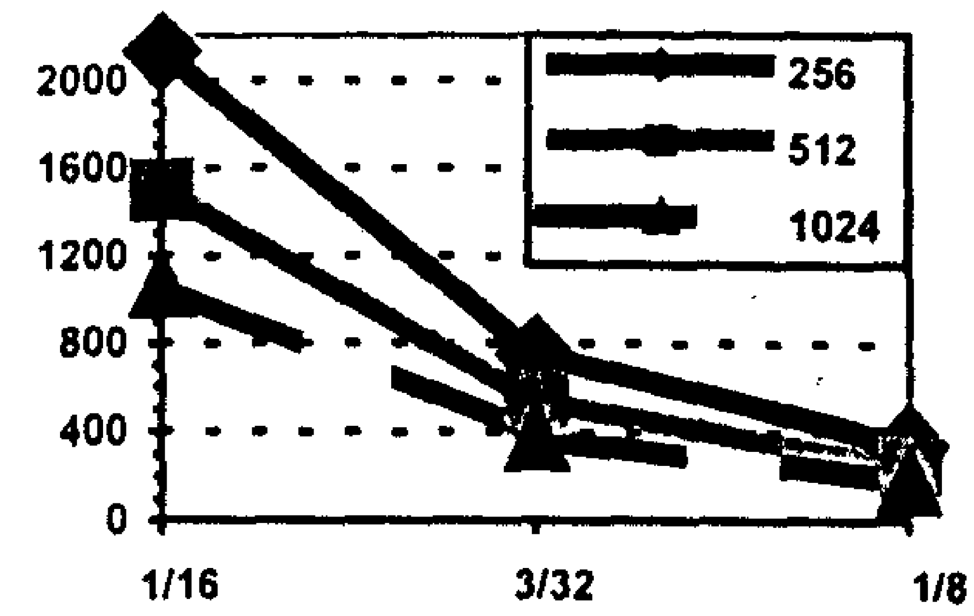

Fig. 4

Figure 3 illustrates application of the proposed method for analysis of maximum attainable compression of communication channels for three to eight pulses of 256-sample duration. Here we can see the dependence of the required amplitude of pulses (the vertical axis) on the time shift between the signals in ADC samples (horizontal axis) in the case of noise variance equal to unity. The calculations were performed using Mathcad 2001, with regard to the MSE of amplitude measurement of the central pulse in the signal packet. In the case of eight signals arranged with intervals equal to 24 samples of ADC, the required signal-to-noise ratio in terms of voltage is 750. Assuming that the mean square deviation of noise equals a single quantum of ADC, realization of the 8-pulse compression will demand only a 12-digit analog-digital converter.

In our efforts to analyze the effect of signal duration on the marginal traffic capacity of the processing methods, it would be convenient to express the time interval between pulses in a packet in relative duration units. Figure 4 shows the calculated dependencies of the normalized marginal time shift between eight signals on the signal-to-noise ratio (in terms of voltage) for several values of signal duration N ($N = 256, 512, 1024$). It can be easily seen that as the signal duration increases the requirements on SNR will soften, which makes it possible to implement a more dense arrangement of pulses along the time axis, or to use ADC with a smaller number of digits.

The results reported above point to the possibility of a multiple increase in traffic capacity of narrow-band communication channels based on the methods of signal processing considered in this paper.

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