CONTENTS

VOLUME 41 NUMBER 9 1998

RADIOELECTRONICS AND COMMUNICATIONS SYSTEMS

Radioelectronics and Communications Systems
Vol. 41, No. 9, pp. 43–47, 1998

Izvestiya VUZ. Radioelektronika
Vol. 41, No. 9, pp. 61–68, 1998
UDC 621.396.967

AUTOCORRELATION METHODS FOR THE CREATION OF MOVING WINDOWS IN PULSE RANGING TASKS

V. I. Slyusar

Kiev, Ukraine

The paper describes some new procedures for range measurement based on creation of moving windows by means of decomposition of the signal into sections with a constant or variable correlation interval and subsequent union of partial correlation sums. A nonlinear transformation of the squared module of coherency is considered.

One of the methods of receiving periodic radio signals backgrounded by noise is generation of the autocorrelation function of the signal mixture [1] when we use differences between the correlation function of the chaotic noise decreasing over time and the periodic function of the continuous signal correlation. Below we consider the respective digital versions of autocorrelation procedures for radio pulse processing and their nonlinear modifications.

Consider a real-valued signal with unmodulated carrier. In the course of the analog-to-digital conversion of the signal we create N discrete readings on its duration interval. By convention, let us divide the radio pulse into halves and create their cross-correlation function with the correlation interval equal to half of the signal duration, i.e., $N/2$. In doing so, the fragmentation of the pulse will be done in such a manner that the product of readings of both of its parts is always of the same sign. As a result, we obtain the simplest balanced correlation procedure in the form:

$$F(s_1) = \sum_{s=s_1}^{s_1+\frac{N}{2}-1} U_s U_{s+\frac{N}{2}}, \tag{1}$$

where s is the ordinal number of the ADC reading, and s_1 is the number of the first reading within the signal's experience.

Maximization of sum (1) by means of enumeration of the s_1 values in the course of moving the processing window along the array of readings permits us to determine the time position of the radio pulse. This method of range measurement is sensitive to the radio signal carrier frequency but invariant to its envelope shape. Despite the insignificance of the differences in the correlation functions of the noise and signal within interval $N/2$, this technique allows us to enhance the range measurement accuracy compared to the moving window of the type described in [2]:

$$F(s_1) = \sum_{s=s_1}^{s_1+N-1} U_s^2 \tag{2}$$

The advantage achievable may be estimated from Fig. 1 displaying the results of the mathematical simulation of procedures (1) and (2) with different signal-to-noise ratios A and signal durations N in the case of a rectangular envelope. Along the vertical axis the mean square errors of range measurement in the ADC readings are plotted, the lines corresponding to processing (2), and rectangles — to procedure (1).

Other approaches to the creation of this type of moving windows may be reduced to various methods of signal decomposition into equal duration segments, with the subsequent joining of their respective weighted correlation sums having either fixed or changing correlation interval. For the dichotomous decomposition of the rectangular radio pulse an

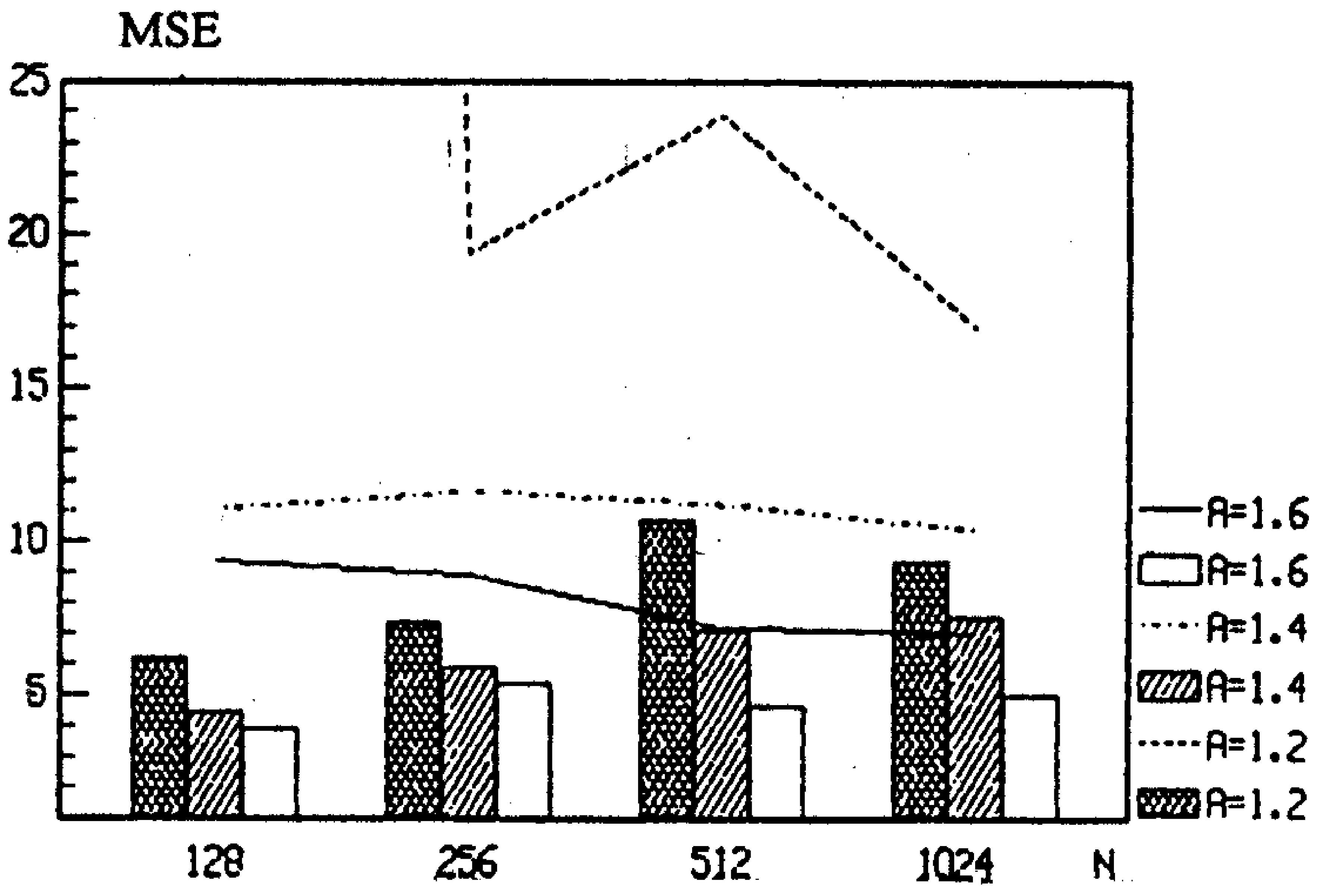

Fig. 1

alternative to procedure (1) may become the accumulation with the weights which are inverse to the values of the noise correlation function. In this case the correlation step varies from $N-1$ to 1 while the processing procedure itself, at even number or readings within the signal duration, takes the form:

$$F(s_1)=\sum_{s=1}^{\frac{N}{2}} U_{s_1+s-1}\, U_{s_1+N-s}\, K^{-1}\,(N-1-2\,[s-1]),$$

where $K(N-1-2\,[s-1])$ is the noise correlation function.

A further generalization of the approach considered in the event of multicomponent representation of radio pulses implies the elimination of the restrictions inherent in the length of the correlated fragments. In this case the interfragment correlation sums of readings for the segments coinciding in their duration may be balanced, and their joining may be performed with various weighting functions. Particularly, for the bell-shaped radio pulse whose envelope is broken into fragments symmetrical about a maximum, the correlation sum

$$F(s_1)=\sum_{r=1}^{R} \gamma_r \sum_{s_r=1}^{N_r} U_{s_r 1}\, U_{s_r 2},$$

where R is the number of pairs of the correlated fragments over the pulse duration interval, r is the ordinal number of the correlated pair of fragments $U_{s_r 1}$ and $U_{s_r 2}$ are the voltages of the first and second correlated fragments of the rth pair in the sth time reading, and N_r is the length of a fragment of the rth pair in the ADC readings.

The autocorrelation methods considered have much in common with well-known interference procedures of direction finding [3]. For this reason many conclusions concerning the accuracy properties of these procedures and their energy losses may also be applicable to the above methods of the pulse-type range measurement.

Another generalization of the above methods may be in the complex-valued representation of the ADC readings. Irrespective of the method of shaping the digital analytical signal [4], in all correlation sums one of the cofactors will be complex-valued while the other is complex conjugate with it. For example, relationship (1) in this case will be transformed into:

$$F(s_1)=\sum_{s=s_1}^{s_1+\frac{N}{2}-1} U_s\, U^{*}_{s+\frac{N}{2}},$$

where * denotes the complex conjugation operator.

Apart from radio signals with unmodulated carrier, the autocorrelation moving windows may be applied to sophisticated signals. For instance, for linear frequency modulated (LFM) pulse it will be enough to perform correlation using irregular step, and try to obtain, as before, an invariant sign of the reading product in the correlation sum. The maximum of the latter will occur at full coincidence of the processing window with the time position of the LFM signal. It is essential that this procedure be accompanied by shaping of the response which is compressed in time, similar to the so-called matched processing [5], while the use of weighting of partially correlated products makes it possible to control the level of side lobes of the compressed pulse.

In addition to purely autocorrelation convolution of the above type, in the field of the moving window we may combine the fragments with matched processing [5], non-coherent summing (2) and the procedures under consideration. Finally, of importance are the procedures for creating the moving correlation sum based on exhaustive search for all nonrecurring combinations of readings. The analytical expression for the respective algorithm as applied to the real-valued representation of the ADC readings will be as follows:

$$F(s_1) = \sum_{m=1}^{N-1} \gamma_m \left(U_m \sum_{p=m+1}^{N} \eta_p U_p\right),$$

where γ_m and η_p are weighting coefficients, which are not necessarily of the same sign.

This approach is accompanied by less power loss compared to procedures of type (1) but requires considerable computational time.

The methods of radio pulse processing considered are invariant to the initial phase of the signal and may be used in single-channel receiving devices of radar, supersonic, and hydroacoustic measuring devices. With analog-to-digital conversion speedy enough, maximum possible accuracy of estimation of the radio pulse time position may be in proportion to signal carrier frequency instead of its spectral width. In this context the approaches considered are preferable to post-detection procedures of the matched filtration of narrow-band video pulses.

As for the resolving power, in the autocorrelation procedures of range measurement considered, like method (2), it may be insufficient for a number of applications. At the same time, according to [6], in the spectral estimation problems there exists a simple technique for enhancing the visually perceived resolution based on replacing the initial spectral function $P_A(f)$ by a new one:

$$P_B(f) = [I - P_A(f)]^{-1}. \tag{3}$$

In this case we may state that in its appearance the relationship $P_B(f)$ will look like a spectrum with "higher resolution" [6].

Thus, of importance is the extension of the method of nonlinear transformation of decision function (3) to the case of range measurement with a check of the possibility to improve, based on the above transformation, the visually perceived resolution of the range measurement procedures of the correlation type.

A peculiar feature of the above approach is the non-normalized nature of the functions to be maximized which does not permit their direct substitution into (3) as $P_A(f)$.

At the same time, a new method of processing is suggested in [6] making it possible to overcome this obstacle by creating the so-called squared module of coherency (SMC):

$$\text{SMC} = \frac{P_{12}(f)\, P_{12}^{*}(f)}{P_{11}(f)\, P_{22}(f)}, \tag{4}$$

where P_{11}, P_{22} are the averaged autospectra of signals of the 1st and 2nd channels, respectively, and P_{12} is the averaged mutual (interchannel) spectrum.

Since the SMC values are confined to the interval from 0 to 1, it is expedient to use, instead of (3), a processing which is equivalent from the information standpoint:

$$F = [I - \text{SMC}]^{-1}. \tag{5}$$

It should be noted that, as applied to the range measurement problem, creation of SMC may be carried out by several methods. In what follows the main attention will be paid to the class of the autocorrelation type measurement procedures based on calculation of the mutual correlation of the pulse signal fragments.

In the simplest case we may confine ourselves to the decomposition of the pulse into halves, assigning R equidistant segments to each half. In this case it becomes possible to carry out in each spectrum the averaging of the mutual and autospectra which, similar to that in [6], in the event of the single-frequency filling of the radio pulse, may be expressed in the following form:

$$P_{nm} = \frac{1}{2R} \sum_{r=1}^{R} \sum_{\substack{n,m=1 \\ m \neq n}}^{2} X_{rn}(f) X^*_{rm}(f),$$

$$P_{nn} = \frac{1}{2R} \sum_{r=1}^{R} \sum_{n=1}^{2} X_{rn}(f) X^*_{rn}(f), \tag{6}$$

where X_{rn} is the M-vector of the discrete Fourier transform (DFT) of the M-dimensional array of complex-valued readings in the rth segment of the nth half of the signal.

The segmentwise averaging, in conformity to [6], is the necessary condition to make SMC sensitive to existence of non-coherent components. Otherwise, if the averaging is not done, SMC = 1 for any type of signals, and its application in the measurement tasks becomes unfeasible.

As was shown by the simulation results, in order to maintain the validity of procedure (5) in the case when two or more pulses are superimposed, the mutual spectra P_{nm} have to be calculated segment-by-segment with enumeration from the center of the signal to its edges. When the envelope is non-rectangular, it is expedient to vary the above segments in terms of their duration so that their length is minimum in the signal center, and maximum at its peripheral parts. This method permits us to accumulate the signal-to-noise ratio on the segments where it decreases due to envelope changes.

Another method of creating SMC reduces to the replacement of the DFT by the procedure of the matched filtration of the radio pulse segments, in the form

$$\widetilde{X}_{rn} = \sum_{s=s_{1rn}}^{s_{1rn}+M_{rn}-1} K_{rn}(s - s_{1rn}) \left[U^c_{srn} \cos P_{srn} + U^s_{srn} \sin P_{srn}\right] +$$

$$+ j \sum_{s=s_{1rn}}^{s_{1rn}+M_{rn}-1} K_{rn}(s - s_{1rn}) \left[U^s_{srn} \cos P_{srn} - U^c_{srn} \sin P_{srn}\right], \tag{7}$$

where $K_{rn}(s - s_{1rn})$ is the normalized discrete function of the radio pulse envelope in the rth segment, s_{1rn} is the number of the ADC reading corresponding to the beginning of the rth segment, $P_{srn} = 2\pi f \Delta t (s - s_{1rn})$, M is duration of the rth segment in the ADC readings, $U^{c(s)}_{srn}$ are the squared components of the sth reading of complex-valued voltages of the signal in the rth segment, Δt is the digitization period, f is the carrier frequency, and n is the number of the signal half.

When using in (6) only a single value of frequency f, the available DFT vector X_{rn} differs from $\widetilde{X}_{rn}$ in (7) only in the presence in the latter of weighting taking into account the discrete function of the envelope.

When creating SMC (4), the further extension of the above approach consists in that the calculation of the coherency factor is performed in a pair of channels, or with the use of two echo-signals, including those having different carrier frequencies. In this case it is possible not to decompose the pulses for processing into fragments but to use them as a whole, by setting the duration of the moving window equal to their duration. To obtain the matched filtration in each window we have to set its own frequency f which is assumed to be known. In addition, when calculating SMC, the role of one signal may be played by some standard mathematical model.

Still another feature is interesting. In the case of narrow-band envelope, the effective duration of the pulse depends on the signal-to-noise ratio: the greater the ratio, the larger part of the signal exceeds the noise level. So the processing window adjusted to a signal with a larger amplitude will not be matched to the pulse with a weak level. In this connection, it is hardly possible that the use of the coherency factor for the echogram visualization will lead to the reduction of the half-tone spectrum since the signal level decrease in the case of the non-rectangular envelope shape will be accompanied by pronounced deviations of the coherency factor from 1. On the other hand, in the "lens" mode of operation after the detection of peaks of function (5), we may process their neighborhood, gradually decreasing the dimension of the vector X_{rn} ($\widetilde{X}_{rn}$),

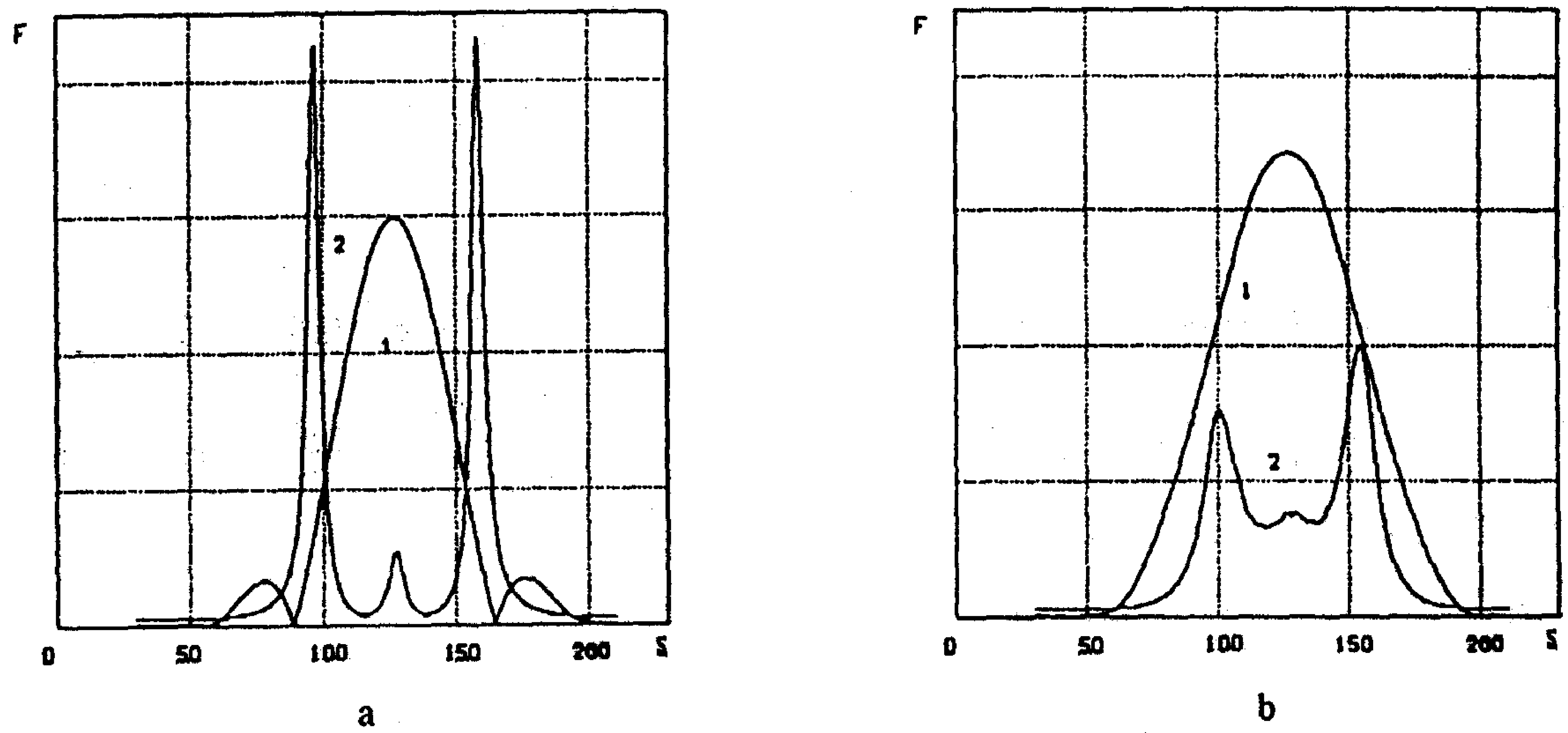

Fig. 2

in order to reach maximum proximity of the coherency factor to the unit level. It opens the possibility to improve the selective properties of the measurement procedure at the weak levels of the signals. The same principle may be applied to the measurement of the echo-pulse duration: it is sufficient to fix the length of the processing window (the data vector length) at the moment of minimum deviation of the coherency factor from unity.

Since the literature does not contain the comparison of procedures with and without nonlinear transformation of the decision function, such a comparison would be of interest. Figure 2 shows the results of mathematical modeling of the procedure consisting in maximization of function (1) (curve *1*) and algorithm (5) (curve *2*) when expression (7) is used for creating SMC. In this case the duration of the radio pulse with envelope $\sin^2 x$ was set equal (as applied to the pulse base) to 100 ADC readings. The discrepancy in the signal arrival times was 53 digitization periods. For both pulses the same signal-to-noise ratio was taken: 28 dB in terms of voltage. The position (a) corresponds to the zero difference of the initial phases of the signals while the position (b) displays the same situation for the case when the phase difference of the carriers is 90°.

In both cases curve *2* demonstrates a pronounced enhancement of resolving capacity taking place at favorable phase relationships and different times of the signal arrivals when the difference exceeds half of the signal duration. However, a radical solution of the problem of signal separation may be obtained only when using special super-resolution procedures. The range estimates thus obtained are not so sensitive to the difference of the carrier initial phases and describe adequately the time distribution of signals with smaller distances between their sources.

REFERENCES

1. A. A. Kharkevitch, Selected Works, Vol. 3 [in Russian], Nauka, Moscow, 1973.
2. V. I. Slyusar, V. I. Pokrovskii, V. F. Sakhno, et al., "The digital range finder," Patent of Russian Federation No. 2069003. IPC G01S 7/28, published in Bulletin of Inventions, no. 31, 10 November 1996.
3. V. V. Karavayev and V. V. Sazonov, The Statistical Theory of Passive Location [in Russian], Radio i Svyaz', Moscow, 1987.
4. M. I. Zhodzishskii, R. B. Mazepa, Ye. P. Ovsyannikov et al., Digital Radio-Receiving Systems - A Handbook [in Russian], (ed. by M. I. Zhodzishskii), Radio i Svyaz', Moscow, 1990.
5. V. I. Slyusar, V. I. Pokrovskii, V. F. Sakhno et al., "A digital method for optimum reception of linear frequency-modulated pulses," Patent of Russian Federation No. 2042956. ICI G01S 7/285, 13/10, published in Bulletin of Inventions, no. 24, 1995.
6. S. L. Marple, Jr., Digital Spectral Analysis and Its Applications [Russian translation], Mir, Moscow, 1990.

29 January 1997