Intra-Period Signal Processing in a Synthetic Aperture Radar

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***Abstract*—One introduces a technique for solving the problem of a synthetic aperture radar tracking when the model of the reflected signal in the form of a discrete normal random process is applied. It is shown that, in a general case, the optimal processing of the signals reflected from the resolved surface element should consist in multichannel linear filtering using the weight coefficients that are generated accounting for the correlation properties of the reflected signal and the receiver noise. The processing procedure should also include the calculation of the linear filter output signal power, followed by the weighting summation of the results. These operations are carried out in terms of the problem of an intra-period processing of the chirp signals in a synthetic aperture radar. The maximum achievable values of the radiometric resolution and the relative level of interfering signals from the surrounding surface elements have been determined for the optimal processing of the chirp signal. The results obtained can be used while designing the promising signal processing devices incorporated in a synthetic aperture radar.**

***Index Terms*—Synthetic aperture radar, intra-period processing, compression of chirp signals, correlation function, radiometric resolution, interfering signal level**

I. Introduction

The technique of a signal processing in a synthetic aperture radar (SAR) for obtaining a "mosaic" of radio brightness is one of the main problems of a SAR technology in terms of maintaining both the best performance characteristics of radio observations and the easiest digital processing hardware implementation procedure [1]-[4].

In general, all received signal samples related to a certain resolution element should be used together in a single processing procedure. However, it is permissible (as it is already usually performed) to consider two independent procedures:

– intra-period processing (pulse compression) [5]-[9],

– inter-period processing (aperture synthesis) [10], [11].

This paper deals with the intra-period processing of SAR signals. Such processing consists in a generation of a compression filter response to a SAR probe signal. The width of the main peak of this function and the relative level of the side peaks generally characterize the radio signal capacity and the method of signal processing that would meet the requirements to the quality characteristics of a radio observation. Particularly, it is obvious that a decrease in the level of side peaks makes it possible to increase the dynamic range of a reproducible radio brightness, i.e. to reduce the level of mutual interferences caused by the reflection of signals from various surface elements. The width of the main peak directly influences the delay resolution [12].

To reduce the level of the side peaks of the compression filter response, “weighting” of the signal spectrum by Hemming function, or by some other functions, is applied [9]. Usually, to evaluate the quality of the intra-period processing of SAR signals, the analysis of the compression filter response, taking into account the used weighting processing, is considered to be enough. In this case, the signal delay resolution is defined by the main peak cross-section width at a particular level, and the mutual interferences caused by the signals coming from resolved delay elements are estimated by the sum of the energy in the side lobes divided by the sum of the energy in the main lobe that is the Integrated Side Lobe Ratio (ISLR).

Meanwhile, the modern methods for analyzing radar observation tasks and calculating SAR parameters are based mainly on the model of a reflecting surface in the form of the point scattering targets [1], [11]. On this basis, in particular, numerical values of the surface resolution are determined. However, in the practice of using SAR, a considerable attention is paid to the tasks of observing an extended rough surface (mapping, etc.) [3], [4]. In these cases, the surface model in the form of point targets cannot be considered applicable to the observation task, and the model of the observed object in the form of an extended rough surface should be used instead [13]. Accordingly, the reflected signal is represented as a discrete random normal process and its statistical properties should be taken into account when processing the signal [14].

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II. PROCESSING TECHNIQUE

SAR probing signal reflected from the resolved surface element can be considered as a normal random process, its correlation function depending on the element's location in the irradiation zone specified by range and azimuth. The observation task is to extract (evaluate) these signals and determine their intensity.

It is assumed that the total signal $y(t)$ observed within a finite interval ΔT is represented by a sequence of complex samples y_i ($i=1,2,\ldots,M$), equal to the sum of the samples of the uncorrelated normal noise $\mathbf{n}$ and elementary signals $\mathbf{s}_l$ ($l=1,2,\ldots,L$) with the preset normalized correlation matrices $\mathbf{R}_l=\left(1/\sigma_l^2\right)\overline{\mathbf{s}_l\mathbf{s}_l^+}$, while $\overline{\mathbf{s}_l\mathbf{s}_k^+}=0$ if $l\neq k$. Here σ_l^2 is l signal power, the bar above being the statistical averaging operation, and "+" symbol denotes transporting with complex conjugation.

The posterior probability density of the signal vectors $\mathbf{s}_l$ takes the form

$$w(\mathbf{s}_1,\ldots,\mathbf{s}_L|\mathbf{y})=K\exp\left[-\frac{1}{2\sigma_0^2}\left(\mathbf{y}-\sum_{l=1}^{L}\mathbf{s}_l\right)^+\times\right.$$
$$\left.\times\left(\mathbf{y}-\sum_{l=1}^{L}\mathbf{s}_l\right)-\sum_{l=1}^{L}\frac{1}{2\sigma_l^2}\mathbf{s}_l^+\mathbf{R}_l^{-1}\mathbf{s}_l\right],$$

where K is the normalizing coefficient, σ_0^2 is the dispersion of the noise samples, $\mathbf{R}_l^{-1}$ is the matrix inverse to $\mathbf{R}_l$.

The optimal estimates of $\mathbf{s}_l$ vectors realizations are the mathematical expectations of the posterior distribution [15]. By applying the variational calculus methods [16], for the estimate $\hat{\mathbf{s}}_k$ of the signal reflected by the k-th surface element, one gets [17]:

$$\hat{\mathbf{s}}_k=\sigma_k^2\mathbf{R}_k\left(\sigma_0^2\mathbf{I}+\sum_{l=1}^{L}\sigma_l^2\mathbf{R}_l\right)^{-1}\mathbf{y}=\mathbf{Q}_k\mathbf{y}\,, \tag{1}$$

where $\mathbf{I}$ is the unit matrix, $\mathbf{R}_k$ is the normalized correlation matrix of $\mathbf{s}_k$ signal, and σ_k^2 is the dispersion of the signal coming from the element number k.

In the general case, the calculation of $\hat{\mathbf{s}}_k$ vector estimate requires the application of the matrix procedure for processing the samples $\mathbf{y}$ of the observed signal. Otherwise, the processing matrix $\mathbf{Q}_k$ can be represented as [10], [18]

$$\mathbf{Q}_k=\sigma_k^2\mathbf{R}_k\mathbf{R}_y^{-1}, \tag{2}$$

where

$$\mathbf{R}_y=\sigma_0^2\mathbf{I}+\sum_{l=1}^{L}\sigma_l^2\mathbf{R}_l \tag{3}$$

is the correlation matrix of the total signal $\mathbf{y}$.

Power of the extracted realization $\hat{\mathbf{s}}_k$ is

$$P_k=\hat{\mathbf{s}}_k^+\hat{\mathbf{s}}_k=\mathbf{y}^+\mathbf{Q}_k^+\mathbf{Q}_k\mathbf{y}=\mathbf{y}^+\mathbf{W}_k\mathbf{y}\,. \tag{4}$$

The expression (4) demonstrates the final result of processing of the received wave $\mathbf{y}$ when estimating the intensity of the signal reflected by the k-th surface element. The specific form of the processing matrix

$$\mathbf{W}_k=\mathbf{Q}_k^+\mathbf{Q}_k \tag{5}$$

depends on the accepted model of a priori intensity distribution σ_l^2. In the simplest case, one can accept that

$$\sigma_l^2=\text{const}\,. \tag{6}$$

The operation (4) can be represented in the form

$$P_k=\sum_{m=1}^{M}\lambda_m\left|\sum_{i=1}^{M}v_{mi}^*y_i\right|^2, \tag{7}$$

where λ_m and $\mathbf{v}_m$ are the eigenvalues and eigenvectors of the matrix $\mathbf{W}_k$ and M is the number of the samples within the observation interval (the size of the correlation matrix being $M\times M$), while "*" symbol denotes complex conjugation.

From (7), it follows that signal $\mathbf{y}$ processing should include the multichannel linear filtering [17] with the weighting functions $\mathbf{v}_m$ (M channels), the calculation of the linear filters output signal power and the following summation of the results with the weighting coefficients λ_m. The practical number of the partial sums (i.e., the eigenvalues λ_m differing significantly from zero), which should be taken into account in (7), depends on the ratio of the number of samples M and the $\mathbf{s}_k$ signal correlation interval.

III. INTRA-PERIOD PROCESSING OF THE CHIRP SIGNALS

Now one can apply the described processing procedure to solve the problem of the chirp pulse compressing. The complex modulating function of the chirped pulse with the duration τ_p and the frequency deviation ΔF_s can be represented as follows [19]

$$S_0(t)=\exp\left(j\pi\Delta F_s t^2/\tau_p\right),\qquad t\in\left[-\tau_p/2,\tau_p/2\right]. \tag{8}$$

The complex modulating function of the signal reflected from the surface element corresponding to the delay interval $\Delta\tau$ takes the form (in the absence of the Doppler frequency shift):

$$S(t)=\int_{\tau_n-\Delta\tau/2}^{\tau_n+\Delta\tau/2}\rho(\tau)S_0(t-\tau)\mathrm{d}\tau\,.$$

Here τ_n is the average delay of the signal reflected by the n-th elementary band; $\rho(\tau)$ is the complex random function describing the reflection coefficient within a delay resolution element. It can be accepted that $\overline{\rho(\tau_1)\rho^*(\tau_2)}=\delta(\tau_2-\tau_1)$ [14].

If the interval of the time quantization of the complex signal samples is chosen equal to

$$\Delta\tau_d=1/\Delta F_s\,, \tag{9}$$

then the correlation matrix of the reflected pulse samples includes the following elements:

$$R_{pq}=\iint_{\tau_n-\Delta\tau/2}^{\tau_n+\Delta\tau/2}\overline{\rho(\tau_1)\rho^*(\tau_2)}S_0(t_p-\tau_1)S_0^*(t_q-\tau_2)\mathrm{d}\tau_1\mathrm{d}\tau_2=$$
$$=\int_{\tau_n-\Delta\tau/2}^{\tau_n+\Delta\tau/2}S_0(t_p-\tau)S_0^*(t_q-\tau)\mathrm{d}\tau\,,$$

where $t_p=(p-M/2)\Delta\tau_d$, $t_q=(q-M/2)\Delta\tau_d$, $p,q=\overline{1,M}$; M is the number of samples per probing period.

By applying (8), (9), for the correlation matrix of the chirp signal samples one can get

$$R_{lpq}=S_{pq}\exp[-j2\pi(p-q)\tau_l/B\Delta\tau_d]\times$$
$$\times\exp\left[j\pi\left((p-M/2)^2-(q-M/2)^2\right)/B\right]. \tag{10}$$

In (10), the notations are the following: $S_{pq}=\mathrm{sinc}[\pi(p-q)\Delta\tau/B\Delta\tau_d]$; $B=\Delta F_s\tau_p=M$ is the chirp signal base; $\Delta\tau$ is the resolution by the delay; τ_l is the average delay from l-th resolution element; $\mathrm{sinc}(x)=\sin(x)/x$, if $x\neq 0$, and $\mathrm{sinc}(x)=1$, if $x=0$. In practice, the samples are taken with a shorter interval $\Delta\tau_d$, but, to simplify the calculations, one will further use the formula (10).

In accordance with the technique presented in Section 2, the matrix $\mathbf{Q}_k$ (2), where $\mathbf{R}_k$ corresponds to (10) and the correlation matrix $\mathbf{R}_y$ of the received signal $\mathbf{y}$ corresponds to (3), is used for linear filtering of the received signal samples $\mathbf{y}$ while generating the estimate $\hat{\mathbf{s}}_k=\mathbf{Q}_k\mathbf{y}$ of reflections from the k-th element.

The result of intra-period processing of the samples within the interval ΔT should be considered the generation of N_1 complex values

$$z_m=\sqrt{\lambda_m}\mathbf{v}_m^+\mathbf{y} \tag{11}$$

for each resolution element. In (11), λ_n and $\mathbf{v}_n$ are eigenvalues and eigenvectors of the matrix $\mathbf{W}_k$ (5).

When using a model (6), one can take that $\mathbf{R}_y=\mathbf{I}$. In this case, in accordance with (2) one gets $\mathbf{Q}=\mathbf{R}_k$, where the multiplier σ_k^2 is omitted. Then the processing matrix (5) takes the form

$$\mathbf{W}_k=\mathbf{R}_k^+\mathbf{R}_k\,, \tag{12}$$

where $\mathbf{R}_k$ corresponds to (10).

During the subsequent inter-period processing, the obtained N_1 values of z_n (11) should be used independently and be combined only at the final stage of the aperture synthesis. Formally, N_1 coincides with the number of signal samples within the interval ΔT. However, the matrix $\mathbf{W}_k$ includes a small number of eigenvalues that are significantly different from zero, which is determined by the ratio between the correlation interval of the signal $\mathbf{s}_k$ reflected by the k-th resolution element and the total processing time ΔT. This number is close to the value $N_1\approx\Delta F_s\Delta\tau$ and determines the redundancy of the deviation ΔF_s relative to the required resolution $\Delta\tau$, i.e. the possibility of the incoherent accumulation of the linear filtering results.

The elements of the weighting functions $\mathbf{v}_m$ of the linear filtration (11) can be represented as follows

$$v_{mp}=V_{mp}\exp\left(-j\frac{2\pi p}{B}\frac{\tau_l}{\Delta\tau_d}\right)\exp\left[j\frac{\pi}{B}\left(p-\frac{M}{2}\right)^2\right], \tag{13}$$

where V_{mp} are the elements of the eigenvectors belonging to the matrix with the elements S_{pq} (10).

It follows from (13) that for the conditional value $\tau_l=0$, the eigenvectors of the matrix $\mathbf{R}_k$ include the elements $v_{mp}=V_{mp}\exp\left[j\pi(p-M/2)^2/B\right]$. The eigenvalues μ_m of the matrix $\mathbf{R}_k$ are equal to the eigenvalues of the matrix $\mathbf{S}$ with the elements S_{pq} (10). Under $\Delta\tau=\Delta\tau_d$, the first four eigenvalues normalized to the maximum value μ_1, one gets

$$\mu_1^0=1,\quad \mu_2^0=0.26,\quad \mu_3^0=0.0145,$$
$$\mu_4^0=0.000273. \tag{14}$$

Formally, the intra-period signal processing should consist of obtaining M samples

$$s_m=\mathbf{v}_m^+\mathbf{y}\,,\qquad m=\overline{1,M}$$

at the output of the linear filters with the pulse responses $\mathbf{v}_m$. In this case, the power of the useful signal component at the output of the m-th filter is $p_m = \overline{|s_m|^2} = \sigma^2 \mu_m$.

In accordance with (7), this value with the weighting coefficient λ_m equal to the eigenvalue of the matrix (12) is included in the final result: $\lambda_m = \mu_m^2$. Thus, at the output of the *m-th* channel processing, the signal power σ^2 has a "weight" corresponding to μ_m^3. And it follows from (14) that only one filter with an pulse response $\mathbf{v}_1$ can be used that means applying the vector processing only.

In Fig. 1, the results of calculating the scalar of the filtration eigenvector $|\mathbf{v}_1|$ corresponding to the maximum eigenvalue μ_1 are shown for the case when the parameters are the following: $\Delta F_s = 10$ MHz; the pulse duration is $\tau_p = 12.1$ µs; the signal base is $B = 121$; the sampling interval is $\Delta T = \tau_p$; the number of samples within the sampling interval is $M = B$; the time resolution is $\Delta\tau = \Delta\tau_d$. In fact, it represents a time (or frequency) weighting function commonly used in the chirp signal filtering. However, in this case, there is no arbitrariness in its choice (the Hamming function, the Taylor function, etc.). Form of the vector $\mathbf{v}_1$ magnitude specifies the optimal weighting function for the required resolution parameters.

The subsequent eigenvalues of the processing matrix become comparable with the first one, if one uses a signal modulation bandwidth that is redundant in comparison with the required delay resolution ($\Delta\tau > \Delta\tau_d$) allowing for an additional smoothing of the speckle noise.

For example, for $\Delta\tau = 3\Delta\tau_d$, the first six eigenvalues are equal to $\mu_1^0 = 1$, $\mu_2^0 = 0.9697$, $\mu_3^0 = 0.7335$, $\mu_4^0 = 0.2629$, $\mu_5^0 = 0.0349$, $\mu_6^0 = 0.0022$, and only the first three of them are the essential ones. In this case, the multichannel (matrix) processing with the weighting functions (Fig. 2) provides the best signal extraction. To that end, the independent results of the signal processing when these signals come from the compression filter output having the responses $\mathbf{v}_m$ must be summed up with the weighting coefficients μ_m^2.

The above ratios make it possible to obtain general formulas for both the maximum achievable radiometric resolution and the ratio of the output power of the useful signal to the power of interfering signals from neighboring surface elements as well as to estimate the efficiency of intra-period processing.

IV. The Radiometric Resolution

Radiometric resolution is an important quality characteristic for surface search radar systems. The concept of radiometric resolution has no generally accepted definition. By and large, the radiometric resolution

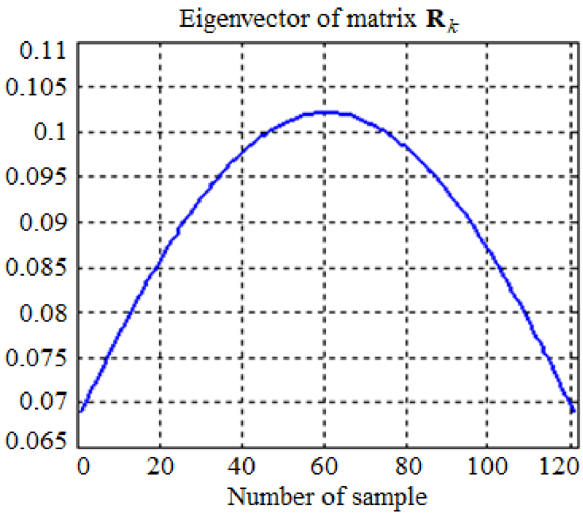


Fig. 1. The scalar of the eigenvector $|\mathbf{v}_1|$ ($\Delta\tau = \Delta\tau_d$).

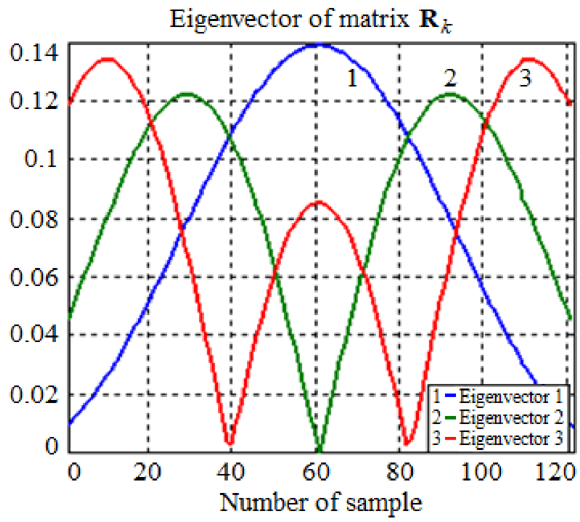


Fig. 2. The scalars of the eigenvectors $|\mathbf{v}_1|$ – $|\mathbf{v}_3|$ ($\Delta\tau = 3\Delta\tau_d$).

("contrast sensitivity") refers to the minimum ratio between the intensities of the two signals at which their difference can be quite reliable [2]. In many cases, the ratio of the effective value of fluctuations σ_y of the sample processing result $\mathbf{y} = \mathbf{s}_k + \mathbf{n}$ to the mathematical expectation $\overline{P}_k$ of the useful signal $\mathbf{s}_k$ is taken as a numerical measure of this parameter [11], [20].

In this case, the radiometric resolution δ is defined as

$$\delta = \left(\overline{P}_k + \sigma_y\right)/\overline{P}_k = 1 + \sigma_y/\overline{P}_k ,$$

The mathematical expectation of the selected signal $\hat{\mathbf{s}}_k$ power (4) and the effective value σ_y of fluctuations in accordance with [15] are determined as follows:

$$\overline{P}_k = \overline{\mathbf{y}^+ \mathbf{W}_k \mathbf{y}} = \sigma_k^2 tr \mathbf{W}_k \mathbf{R}_y , \quad \sigma_y = \sqrt{tr \mathbf{W}_k^2 \mathbf{R}_y^2} , \tag{15}$$

where *tr* is the operation of calculating the matrix trace.

With this in mind, one can get

$$\sigma_y^2 / \overline{P}_k^2 = tr\mathbf{W}_k^2\mathbf{R}_y^2 \Big/ \left(\sigma_k^2 tr\mathbf{W}_k^2\mathbf{R}_k^2\right)^2 ,$$

where $\mathbf{R}_y = \sigma_0^2\mathbf{I} + \sigma_k^2\mathbf{R}_k$.

Under the resolution by the delay and according to (12) one can write

$$\frac{\sigma_y^2}{\overline{P}_k^2} = \frac{tr\mathbf{R}_k^4\left(\sigma_0^2\mathbf{I} + \sigma_k^2\mathbf{R}_k\right)^2}{\left(\sigma_k^2 tr\mathbf{R}_k^3\right)^2} = \left(1 + \frac{\sigma_0^2}{\sigma_k^2}\right)^2 \frac{\sum_{m=1}^{M}\mu_m^6}{\left(\sum_{m=1}^{M}\mu_m^3\right)^2} .$$

As a rule, the value of the radiometric resolution δ is expressed in the logarithmic units:

$$\delta = 10\lg\left[1 + \left(1 + 1/q^2\right)/K\right], \quad (16)$$

where $K = \sum_{m=1}^{M}\mu_m^3 \Big/ \sqrt{\sum_{m=1}^{M}\mu_m^6}$ and $q^2 = \sigma_k^2/\sigma_0^2$ is the signal-to-noise ratio in the samples.

The coefficient $K > 1$ determines the improvement of the radiometric image resolution when using the algorithm (7). It is possible due to the summation of the independent results of processing the signals from the outputs of the compression filters having responses $\mathbf{v}_m$ with the weight coefficients μ_m^2 (see Section 3). With an excessive signal modulation bandwidth ΔF_s ($\Delta\tau > \Delta\tau_d$), a significant improvement in the radiometric resolution of the image can be achieved.

In Fig. 3, one can see the results of calculating the dependence of the radiometric resolution (16) on the signal-to-noise ratio. The time resolutions are $\Delta\tau = \Delta\tau_d$ (curve 1), $\Delta\tau = 2\Delta\tau_d$ (curve 2) and $\Delta\tau = 3\Delta\tau_d$ (curve 3). The rest of the calculation parameters are the same as for Fig. 1.

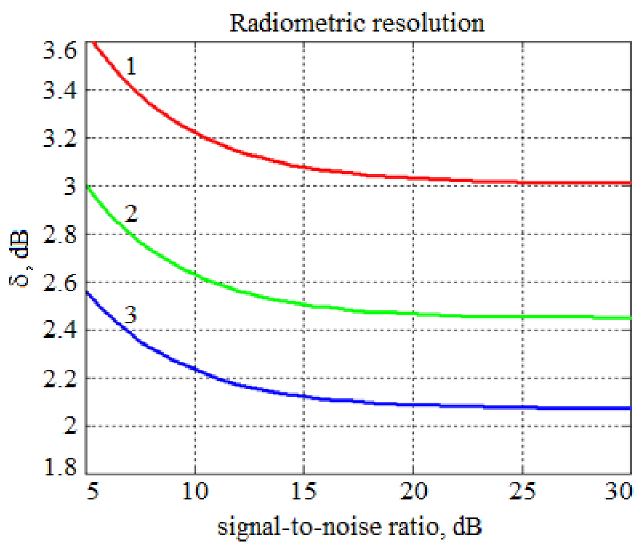


Fig. 3. The dependence of the radiometric resolution on the signal-to-noise ratio.

It follows from Fig. 3 that with an excess modulation bandwidth of the signal (curves 2, 3), the value of δ decreases, i.e. the radiometric resolution is increased. Increasing the signal-to-noise ratio by more than 20 dB does not make a lot of difference to the value of the radiometric resolution.

The obtained relation (16) refines and generalizes the approximate formula $\delta = 10\lg\left[1 + \left(1 + 1/q^2\right)/\sqrt{N}\right]$ that, as a rule, is used to estimate the radiometric resolution when summing N independent samples in a resolution element [2], [11]. The expression (16) does not include N number of incoherent accumulations, since such a procedure is not provided at all in the algorithm (7). Incoherent accumulation in (7) is implemented by means of weight summation of the output results of the filtering channels. In this case, instead of a number N, the number of eigenvalues of the processing matrix $\mathbf{W}_k$ is used, specifically the ones that significantly differ from zero. It should be noted that the specified number depends on both the actual time of element observation in the radiation area and the chosen resolution.

V. The Level of the Interfering Signals Coming From the Surrounding Surface Elements

The efficiency of the intra-period processing can be estimated by the relative power of the signals $\hat{\mathbf{s}}_k$ (1) at the output of the linear filter under unit values of σ_k^2.

If the useful signals $\mathbf{s}_l$ are considered as forming the input signal $\mathbf{y}$, then, taking into account (15), one can write

$$\overline{P}_l = \overline{\mathbf{s}_l^+\mathbf{W}_k\mathbf{s}_l} = tr\left(\mathbf{W}_k\mathbf{R}_l\right).$$

Thus, introducing different values of l, it is possible to determine the power of the signals coming from the various elements at the output of the processing channel tuned to the k-th element. The relative level of interfering signals can be calculated from the relation

$$\overline{P}_l / \overline{P}_k = tr\left(\mathbf{W}_k\mathbf{R}_l\right) / tr\left(\mathbf{W}_k\mathbf{R}_k\right), \quad (17)$$

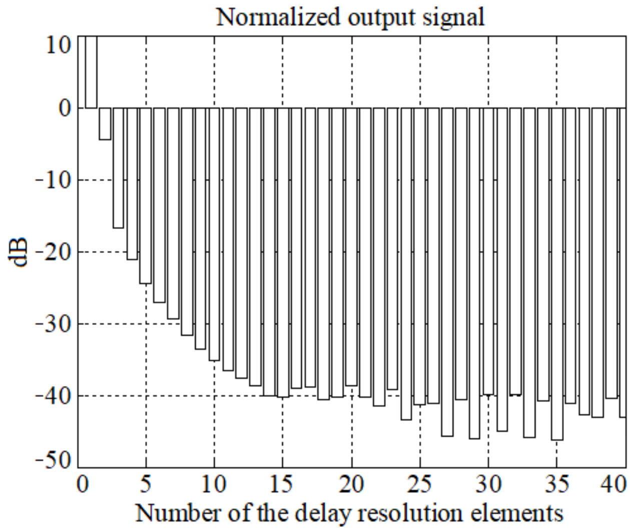
where the elements of the matrix $\mathbf{R}_k$ are determined from (10) under $\tau_l = 0$.

As the quality characteristic of the delay (range) resolution one can consider the ratio between the output power of the k-th signal ($l = k$) and the power of the signal coming from the neighboring element ($l = k + 1$). Taking conditionally that $k = 0$, one gets

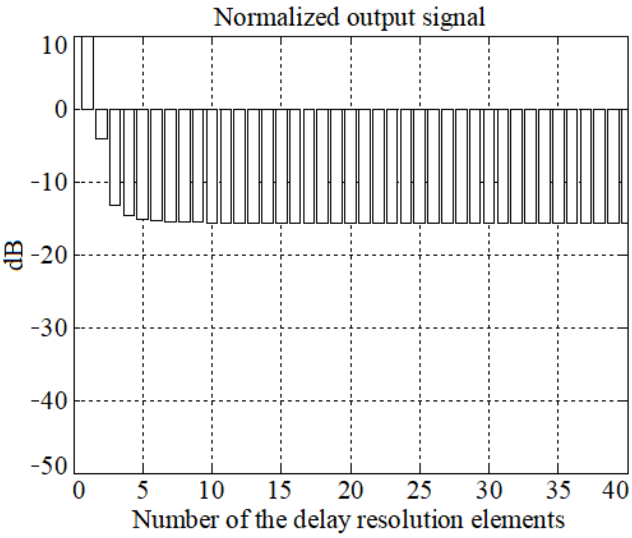
$$\Delta = tr\left(\mathbf{W}_0\mathbf{R}_0\right) / tr\left(\mathbf{W}_0\mathbf{R}_1\right).$$

If the value of Δ is satisfactory, then it should be considered that the specified value of the delay resolution $\Delta\tau$ is provided.

More precisely, the sum of powers of the interfering signals $P_{\Sigma} = \sum_{l \neq 0} tr(\mathbf{W}_0 \mathbf{R}_l)$ can be taken into account and, thus, the value



a)



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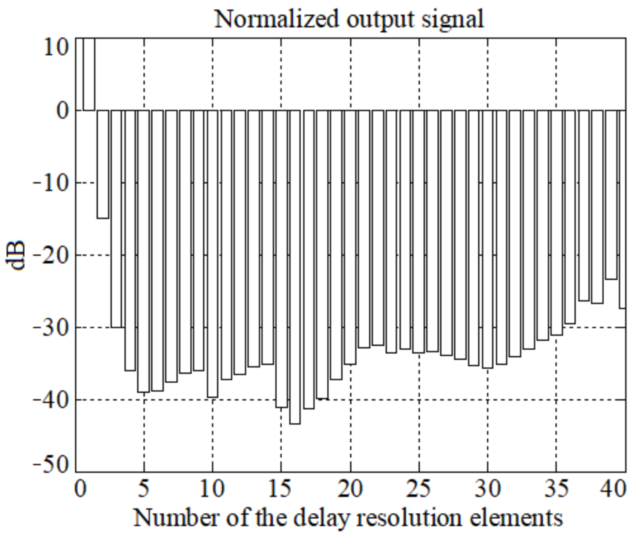
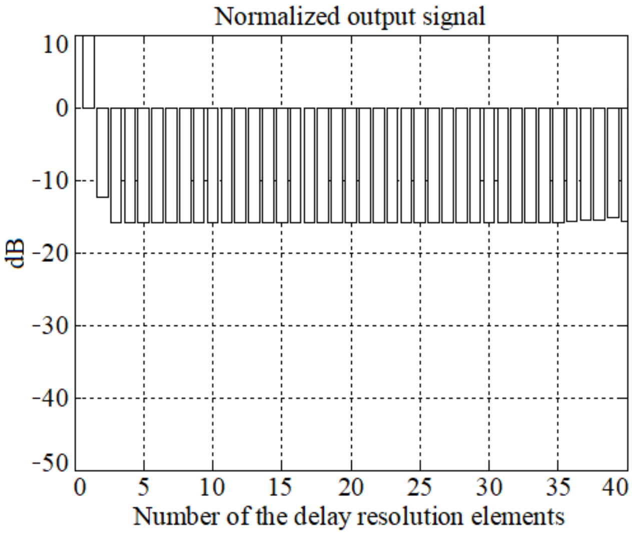


c)



d)

Fig. 4. The relative power of the signals $\hat{\mathbf{s}}_k$ at the output of the linear filter: a) $\Delta\tau = \Delta\tau_d$, q^2 = 50 dB; b) $\Delta\tau = \Delta\tau_d$, q^2 = 25 dB; c) $\Delta\tau = 3\Delta\tau_d$, q^2 = 50 dB; d) $\Delta\tau = 3\Delta\tau_d$, q^2 = 25 dB.

$$\mathrm{ISLR} = tr(\mathbf{W}_0 \mathbf{R}_0) \Big/ \sum_{l \neq 0} tr(\mathbf{W}_0 \mathbf{R}_l)$$

can be determined. However, as a rule, such an improvement is irrelevant.

In Fig. 4, there are presented the calculation results of the relative power (17) of the signals $\hat{\mathbf{s}}_k$ (1) at the output of the linear filter. The time resolutions are $\Delta\tau = \Delta\tau_d$ (Fig. 4a, Fig. 4b) and (Fig. 4c, Fig. 4d). The signal-to-noise ratio in the samples is 50 dB (Fig. 4a, Fig. 4c) or 25 dB (Fig. 4b, Fig. 4d). The rest of the calculation parameters are the same as for Fig. 1.

From Fig. 4a, it follows that the nearest peak (17) is –14 dB. When the signal modulation bandwidth is excessive in comparison with the required resolution by the delay (Fig. 4c), the nearest peak becomes equal to –26 dB and this allows improving the radar image in subsequent aperture synthesis.

In accordance with the graphs in Fig. 4, when increasing the noise power, the ratio between the main signal power (from the selected band) and the power of the nearest interfering signals (from the neighboring bands) deteriorates by 1 dB when $\Delta\tau = \Delta\tau_d$ (Fig. 4b) and by 2 dB when $\Delta\tau = 3\Delta\tau_d$ (Fig. 4d).

VI. Conclusion

The performed analysis allows substantiating the structure of the intra-period SAR signal processing channel based on the observation object model presented as the extended rough surface. The results of the analysis show that in order to achieve the limiting quality characteristics, it is necessary to use multichannel processing devices that take into account the correlation functions of the signals reflected by each of the resolution elements on the surface. The obtained ratios for the main quality characteristics of intra-period processing make it possible to estimate the potentially attainable performance indicators when conducting radio observations using SAR. The examples of numerical calculations confirm the usefulness of the results obtained for the practical design of the devices for SAR signal processing. In the modern conditions of the widespread application of computing facilities in SAR signal processing devices, the implementation of the introduced intra-period processing procedure does not give rise to any special difficulties.

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