Optimization of the digital rejection filter

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Abstract. **The digital rejection filter (RF) is offered in the form of the device for subtraction of weighted samples in the non-delayed channel and results of group accumulation of samples in the delayed channel. The RF optimization task is considered with the group sample accumulation in the delayed channel. The optimal relationships between RF parameters and correlation properties of interference are discussed, which corresponds to minimum of interference remainders. The influence of bit grid finiteness of the analog-to-digital converter (ADC) on effectiveness of interference rejection is studied. The expression is suggested for the minimal number of bits, which can be used for a choice of ADC type with account of given losses in effectiveness of interference rejection and required operation speed. The RF adaptation principles under condition of *a priori* uncertainty of interference correlation parameters are discussed. The analysis of the adaptive RF effectiveness is carried out depending on correlation properties of interference and the volume of the learning sample. From relations obtained, it follows that losses in effectiveness of interference rejection, which are caused by adaptation errors, can be restricted in advance by the given value by means of appropriate choice of the learning sample volume.**

Keywords: adaptation, quantization, feedback, learning sample, optimization, bit grid, rejection filter, correlation coefficient, clutter

I. INTRODUCTION

Extraction of the moving target signals on the background of correlated (passive) interference caused by the spurious reflections (clutter) is the one of relevant and difficult tasks of arrived data processing, which are usually solved in the radar systems (RS) of various destinations [1]. The passive interference seriously disturbs the RS normal operation, leading to overloading of receiver path, as well as to masking and, in the end, to moving target signal missing. Protection methods against passive interference depend on the RS type and the probing signal used. This problem is the most effectively solved in so-called pulse-Doppler radars with low off-duty factor of the probing signal, or in RS with quasi-continuous emission, in which the pulses with high repetition frequency are used (up to some tens and even hundreds kilohertz). At that, the off-duty factor does not usually exceed to 20. Selection of moving targets on the background of spurious reflections in such RSs is based on the Doppler effect usage [1].

Processing of received data is performed in the multi-channel system for range and Doppler frequency. Initially, the range gating (selection) is carried out. For each range interval under selection, the set of Doppler filters covers the whole possible range of Doppler frequencies of signals, which are reflected from moving targets. Owing to Doppler frequency difference between the interference and the useful signal, the tracking of the useful signal is performed in Doppler filters, which are interference-free. This provides the best selection of moving targets on the clutter background. At that, we can achieve the unambiguous measurement of the target radial velocity with high resolution and accuracy. The range measurement, however, relates to ambiguity, which can be eliminated by the special method application, which makes the signal processing more complicated [1].

The unambiguous range measurement for large number of targets by simple approaches and with high resolution is achieved in coherent-pulse RS with probing pulses of high off-duty factors, which causes the wide application of such RSs in practice [1]. The low pulse repetition frequency selecting from the condition of unambiguous range measurement leads to close location of the comb spectral components filter, which complicates the moving target signal selection on the background of interference, which power is large compared to the signal. In this case, the main operation of the received signal processing is the rejection of interference spectral components, and RF is the main unit of the appropriate processing system [1].

Utilization of the digital signal processing technique allows implementation of the under-optimal processor on the base of the digital filter for interference suppression, and leaded to realization of the RF with adaptation to the clutter Doppler phase [2-5]. Development of digital methods and devices for digital signal processing proceeds discussions in modern scientific-technology literature.

The parametric *a priori* uncertainty at signal detection on the clutter background makes essential difficulties at effective detection of moving targets. This leads to necessity of the adaptive Bayesian approach [6] utilization based on estimation of unknown interference parameters according to the learning samples, which are formed by samples on adjacent bins on range or the Doppler frequency. The adaptive detection of moving target signals on the clutter background, which is arose by the spurious reflections from the lengthy objects, was considered in [7].

Manuscript submitted on 12 May, 2017.

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The point (in range) interference corresponding in duration to signals, which reflected from the point target and which cannot be distinguished from signals reflected from moving targets, requires the special attention. Problems of surveillance radar protection against interference of such a type are considered in [8-11]. In [12], by means of the maximal likelihood method, the estimation algorithm is synthesized for the coefficient of inter-period correlation of the point clutter. The formula for estimation accuracy is obtained, which depends on the correlation coefficient and the volume of learning sample. Digital filters providing rejection of the point interference are of interest.

Digital RF, which suppresses the correlated interference, increases the level of non-correlated interference (proper noises), which leads to losses in SNR and interference/noise ratio. To reduce these losses in RF we offer to use the group sample accumulation, which can be realized with the help of the switched delayed feedback [13].

The structural diagram of digital RF in the form of subtraction device of weighted samples in non-delayed channel and results of group sample accumulation in the delayed channel is offered below. The problem of such RF optimization is considered and optimal relationships are described between RF parameters and correlation properties of interference, which correspond to minimum of interference residuals. The influence of finiteness of the bit grid of analog-to-digital converter (ADC) on effectiveness of interference rejection is considered. The relationship for minimal number of bits is obtained, by which the ADC type can be chosen taking into account the given losses in effectiveness of interference rejection and required operation speed. The principles of RF adaptation under conditions of a priori uncertainty of the correlation parameters of interference are described. The effectiveness analysis of the adaptive RF depending on correlation properties of interference and the volume of learning sample is performed.

II. STRUCTURE AND PARAMETER OPTIMIZATION OF RF

The structural diagram of RF with group sample accumulation with the help of the switched delayed feedback is presented in the Figure 1, where Cal is the calculator of the weighting coefficient a, $\times$ is the multiplier unit, Sw is a switcher, CB is the control block, Σ is a summer, SD is the storage device (for repetition period T), and Cm is a commutator.

Samples of the point interference, which arrive from ADC, in the j-th period (with account of quantization errors) have a form $\tilde{u}_j = u_j + \xi_j$, where ξ_j are samples of the quantization noise. After weighting in the multiplier unit $\times$ of the non-delayed RF channel with the weight coefficient a, we obtain $a(u_j + \xi_j)$. In the delayed channel, taking into consideration the implementation with the help of the first summer Σ, the storage device SD and the commutator Cm of the group sample accumulation with N periods, we have

$$\sum_{k=1}^{N}(u_{j-k} + \xi_{j-k}).$$

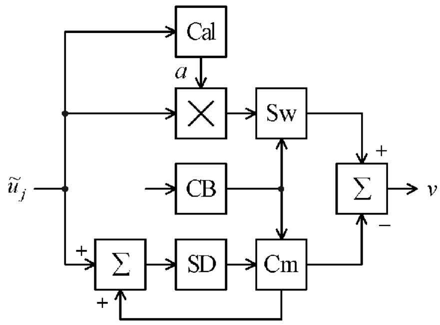

Figure 1. The structural diagram of RF

After closing of the switch Sw (according to the command from CB) and switching of Cm to the input of the second summer Σ of RF, the following quantity is calculated at the output of the second summer

$$v = a(u_j + \xi_j) - \sum_{k=1}^{N}(u_{j-k} + \xi_{j-k}) =$$
$$= au_j - \sum_{k=1}^{N} u_{j-k} + a\xi_j - \sum_{k=1}^{N}\xi_{j-k}.$$

The standard deviation of the RF output is $\sigma_v^2 = \overline{v^2}$. Taking into account the absence of quantization noise correlation with the process under quantization, we obtain

$$\sigma_v^2 = a^2\overline{u_j^2} - 2a\sum_{k=1}^{N}\overline{u_j u_{j-k}} +$$
$$+ \overline{\left(\sum_{k=1}^{N} u_{j-k}\right)^2} + a^2\overline{\xi_j^2} + \overline{\left(\sum_{k=1}^{N}\xi_{j-k}\right)^2}.$$

Assuming that interference characteristics are stationary and tasking into consideration the absence inter-period correlation and the uniform distribution law of quantization noise, we finally obtain

$$\sigma_v^2 = (a^2 + N)\sigma^2 - 2a\sigma^2\sum_{k=1}^{N}\rho(kT) +$$
$$+ 2\sigma^2\sum_{k=1}^{N}\sum_{l=1}^{N-k}\rho[(k-l)T] + (a^2 + N)\frac{\delta^2}{12}, \tag{1}$$

where σ^2 is the interference standard deviation on the RF input, $\rho(kT)$ are coefficients of inter-period interference correlation, δ is the ADC quantization step.

As we see from (1), effectiveness of interference suppression is determined by coefficients of its inter-period correlation and RF parameters a and N. It is interesting to find the optimal relation between these quantities corresponding to minimum of interference remainders. For this, we write the equation

$$\frac{\partial\sigma_v^2}{\partial a} = 2a\sigma^2 - 2\sigma^2\sum_{k=1}^{N}\rho(kT) = 0,$$

from the solution of which we find out the optimal value of a

$$a = a_{\text{opt}} = \sum_{k=1}^{N}\rho(kT). \tag{2}$$

Since $\rho(kT) < 1$, we have $a < N$. For the non-fluctuating interference $\rho(kT)=1$ and $a=N$, which is evident and confirms the trustworthiness of considered transformations.

At exponential function of interference correlation, we have: $\rho(kT)=\rho^k$, where $\rho=\exp(-T/\tau_c)$, τ_c is correlation time. In this case, the optimal value of weight coefficient is

$$a=a_{\text{opt}}=\sum_{k=1}^{N}\rho^k, \tag{3}$$

which simplifies its determination.

At presence of *a priori* information about the ρ value, we can choose in advance the optimal value of a coefficient. Under conditions of *a priori* ambiguity, the a value should be chosen according to the minimax rule orienting on the some average ρ value from the *a priori* expected interval of its variation, or to use the adaptive methods for a choice according to estimated value of the correlation coefficient $\hat{\rho}$, which is determined in the appropriate measuring system according to the operation algorithm [19]. (The adaptive a choice is considered below).

III. CHOICE OF RF DIGIT CAPACITY

Now we consider the influence of the ADC bit grid on effectiveness of interference suppression. Quantization with the δ step leas to increase of non-compensated interference remainders, which have the random character due to inter-period interference fluctuations and the presence of the receiver proper noise. The growth of non-compensated interference remainders may be taken into account by introduction of the discrete quantization noise, which is included in (1) with account of passage through RF.

Let us designate the standard deviation of the interference itself in (1) as

$$\sigma_{v\text{cl}}^2=(a^2+N)\sigma^2-2a\sigma^2\sum_{k=1}^{N}\rho(kT)+$$
$$+2\sigma^2\sum_{k=1}^{N}\sum_{l=1}^{N-k}\rho[(k-l)T].$$

Then, losses in effectiveness of interference compensation can be estimated according to the ratio of standard deviation of interference residuals with and without account of the quantization noise:

$$\frac{\sigma_v^2}{\sigma_{v\text{cl}}^2}=\frac{\sigma_{v\text{cl}}^2+(a^2+N)\delta^2/12}{\sigma_{v\text{cl}}^2}=1+\frac{\delta^2}{12}\cdot\frac{a^2+N}{\sigma_{v\text{cl}}^2}.$$

As we see, losses depend on the quantization step δ, RF parameters a and N, and also on the value of reminders of interference itself at the RF output. Ultimately, losses are determined by the ratio of quantization noise and interference reminders at the RF output. However, from the point of view of further processing, the loss value should be considered with account of the proper noise. The proper noise passes through RF in the similar manner as the quantization noise, i.e., its standard deviation σ_n^2 varies by (a^2+N) times. Then, the standard deviation of interference residuals at the RF output takes a form (taking into account the proper noise and the quantization noise):

$$\sigma_{v\text{n}}^2=\sigma_{v\text{cl}}^2+(a^2+N)(\sigma_\text{n}^2+\delta^2/12).$$

At that, losses caused by quantization errors correspond to expression

$$\frac{\sigma_{v\text{n}}^2}{\sigma_{\tilde{v}}^2}=\frac{\sigma_{v\text{cl}}^2+(a^2+N)(\sigma_\text{n}^2+\delta^2/12)}{\sigma_{v\text{cl}}^2+(a^2+N)\sigma_\text{n}^2}=$$
$$=1+\frac{(a^2+N)\delta^2}{12[\sigma_{v\text{cl}}^2+(a^2+N)\sigma_\text{n}^2]},$$

where $\sigma_{\tilde{v}}^2$ is the standard deviation of interference and the proper noise (at RF output) without account of the quantization noise.

In this case, losses are determined by the ratio of the quantization noise and a sum of interference reminders and the proper noise at RF output. The largest losses will be at ideal compensation of the non-fluctuating interference, i.e., at $\sigma_{v\text{cl}}^2=0$. Then, we have

$$\Delta=\frac{\sigma_{v\text{n}}^2}{\sigma_{\tilde{v}}^2}=1+\frac{\delta^2}{12\sigma_\text{n}^2}.$$

The loss value now depends on the ratio of the quantization noise and the proper noise. The level of the quantization noise is determined by the dynamic range u_{dr} and the number of bits ν of ADC. At that, the quantization step is $\delta=u_{\text{dr}}/2^\nu$. The ADC dynamic range should be chosen for the sum of the signal u_s, interference and noise with account of the interference and noise range, not less than $3(\sigma+\sigma_\text{n})$, i.e., $u_{\text{dr}}=u_\text{s}+3(\sigma+\sigma_\text{n})$. The loss value $\Delta=\sigma_{v\text{n}}^2/\sigma_{\tilde{v}}^2$ corresponds to the minimal bit number:

$$\nu=\frac{1}{2}\log_2\frac{u_{\text{dr}}^2}{12(\Delta-1)\sigma_\text{n}^2},$$

according to which the ADC type can be chosen, taking into consideration the given losses Δ and the required operation speed.

At suppression of the fluctuating interference, we must take into consideration the losses caused by the proper noise. These losses value (without account the quantization noise) is determined by equation:

$$\frac{\sigma_{\tilde{v}}^2}{\sigma_{v\text{cl}}^2}=\frac{\sigma_{v\text{cl}}^2+(a^2+N)\sigma_\text{n}^2}{\sigma_{v\text{cl}}^2}=1+\frac{(a^2+N)\sigma_\text{n}^2}{\sigma_{v\text{cl}}^2}=1+\frac{\lambda}{\mu},$$

where $\lambda=\sigma_\text{n}^2/\sigma^2$ is the input ratio noise/interference,

$$\mu=1-\frac{2a\sum_{k=1}^{N}\rho(kT)}{a^2+N}+\frac{2\sum_{k=1}^{N}\sum_{l=1}^{N-k}\rho[(k-l)T]}{a^2+N}\;-$$

is the normalized coefficient of interference suppression.

Since $\mu<1$, the account of the proper noise influence on the interference compensation effectiveness corresponds to loss growth. If the interference is suppressed below the level of the proper noise, we can neglect by interference reminders and at further processing of the moving object signal to consider the proper noise influence only.

IV. RF ADAPTATION

The above-described solution of the RF optimization task with group delayed sample accumulation depending on correlation properties of the point interference allows minimization of its reminders at the RF output and under conditions of *a priori* uncertainty of interference correlation parameters.

During accumulation in the delayed channel of samples with N adjacent repetition periods T and weighting of samples by the weight coefficient a in –non-delayed RF channel, the optimal value $a = a_{\text{opt}}$, determined in general case by (2), while at the exponential function of interference correlation – by (3), corresponds to minimum of reminders.

At presence of *a priori* information about the ρ value, the optimal value of a may be chosen in advance. Under conditions of *a priori* uncertainty it is necessary to use adaptive methods for the a value choice.

In accordance with the adaptive Bayesian approach, the unknown quantities $\rho(kT)$ are replaced by their consistent estimations $\hat{\rho}(kT)$ [6]. At known form of the correlation function, it is enough to find estimation of the correlation coefficient $\hat{\rho}$. Then, in the case of exponential correlation function, we have for estimated value of the optimal weight coefficient:

$$\hat{a} = \sum_{k=1}^{N} \hat{\rho}^{k} . \tag{4}$$

As the $\hat{\rho}$ estimation, we should use estimates of maximal likelihood (EML) obtaining on the base of both direct and indirect algorithms of sampled values' processing u_j, $j = \overline{1, n}$. Direct algorithms contain the multiplication operation of initial samples [12]:

$$\hat{\rho} = \sum_{j=2}^{n} u_{j-1} u_j \Big/ \sum_{j=2}^{n} u_j^2 .$$

Indirect (summing-subtracting) algorithms are free from this operation:

$$\hat{\rho} = 1 - \frac{\sum_{j=2}^{n} (u_{j-1} - u_j)^2}{\sum_{j=2}^{n} (u_{j-1}^2 + u_j^2)} .$$

Modified algorithms deal with summed (subtracted) values:

$$\hat{\rho} = \frac{\sum_{j=2}^{n} [(u_{j-1} + u_j)^2 - (u_{j-1} - u_j)^2]}{\sum_{j=2}^{n} [(u_{j-1} + u_j)^2 + (u_{j-1} - u_j)^2]} .$$

Calculation of correlation coefficient $\hat{\rho}$ estimation according to one of presented algorithms and the optimal weighting coefficient on algorithm (4) is performed in the calculator Cal (see Figure 1).

From EML consistency condition [14]

$$P\{\lim_{n\to\infty} \hat{\rho} = \rho\} = 1$$

it follows that with growth of learning sample volume n, the $\hat{\rho}$ estimation with unit probability converges to the true value of parameter under estimation. Hence, EML usage (instead unknown parameters) leads to adaptive algorithms, which have the convergence property to the appropriate algorithms at known parameters.

At adaptive interference compensation, the RF output value is

$$v = \hat{a} u_j - \sum_{k=1}^{N} u_{j-k} .$$

The standard deviation of the RF output quantity is

$$\sigma_v^2 = \overline{v^2} = \overline{\hat{a}^2 u_j^2} - 2\overline{\hat{a} \sum_{k=1}^{N} u_j u_{j-k}} + \overline{\left(\sum_{k=1}^{N} u_{j-k} \right)^2} .$$

Because of the fact that determination of the $\hat{\rho}$ estimation is based on the averaging of n samples, the mutual correlation of the $\hat{\rho}$ estimation and the separate sample u_l is practically absent. From these, it follows that the mutual correlation of quantity $\hat{a}$ and u_l is absent. Then, assuming that interference characteristics are stationary, we obtain

$$\sigma_v^2 = (\overline{\hat{a}^2} + N)\sigma^2 - 2\bar{\hat{a}}\sigma^2 \sum_{k=1}^{N} \rho(kT) +$$
$$+ 2\sigma^2 \sum_{k=1}^{N-1} (N-k)\rho(kT) , \tag{5}$$

where σ^2 is the interference standard deviation at RF input.

Using asymptotic properties of EML $\hat{\rho}$, we perform the appropriate averaging in (5). Equation (4) can be represented in the form of functional transform $\hat{a} = f(\hat{\rho})$. Let us consider the linear approximation of the $\hat{a} = f(\hat{\rho})$ function in vicinity of ρ in the form

$$\hat{a} = a + a'(\hat{\rho} - \rho) ,$$

where

$$a = f(\rho) , \; a' = f'(\rho) = \partial f(\rho) / \partial \rho .$$

According to (3), it is easy to find

$$a' = b = \sum_{k=1}^{N} k \rho^{k-1} .$$

Taking into consideration the asymptotic normality of the $\hat{\rho}$ estimation distribution with the mean value ρ and the standard deviation $\sigma_{\hat{\rho}}^2$, we obtain

$$\overline{\hat{a}^2} = \overline{[a + b(\hat{\rho} - \rho)]^2} = a^2 + b^2 \sigma_{\hat{\rho}}^2 . \tag{6}$$

The standard deviation $\sigma_{\hat{\rho}}^2$ is determined in accordance with the expression obtained in [12]:

$$\sigma_{\hat{\rho}}^2 = \frac{(1-\rho^2)^2}{(n-1)(1+\rho^2)} .$$

We would like to note than for direct and indirect algorithm, the measurement accuracy is the same [12], which justifies their equivalence from the point of view of measurement accuracy.

Taking into account equations (5) and (6) and the obvious equality $\overline{\hat{a}} = a$, we finally obtain for interference suppression by the adaptive RF:

$$\frac{\sigma_v^2}{\sigma^2} = a^2 + b^2\sigma_{\hat{\rho}}^2 + N - 2a\sum_{k=1}^{N}\rho(kT) + 2\sum_{k=1}^{N-1}(N-k)\rho(kT) . \qquad (7)$$

Equation (7) allows estimation the effectiveness of the adaptive RF as a function of correlation properties of interference, the volume of learning sample and RF parameters. As wee see, the adaptation errors determining by the $\sigma_{\hat{\rho}}^2$ standard deviation value, increase the interference reminders at RF output, resulting in appropriate losses in effectiveness of its suppression. Calculation show that at $n \geq 4$, losses do not exceed a fraction of decibel in the wide range of interference parameters variation.

V. Conclusion

Optimization of RF parameters in accordance with correlation properties of fluctuating interference allows minimization of their reminders at the RF output, and the proposed choice of ADC digit capacity allows limitation of losses, caused by quantization, by the required value.

Adaptation of the offered rejection filter allows minimization of reminders of the correlated interference at its output under conditions of parametric *a priori* uncertainty, while the losses caused by the adaptation errors in the effectiveness of interference suppression, as it follows from equation obtained, can be limited by the given in advance value by means of appropriate choice of the learning sample volume.

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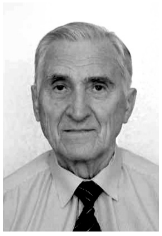

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