The Method of Open Space Selection of Signals for Redcom Systems

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***Abstract*—The article proposes the method of simultaneous reception of radar and telecommunication OFDM (N-OFDM) signals by a receiving digital antenna, that is based on the spatial selection of signal sources. Two-Stage demodulation of signals is proposed. The angular coordinates of radiation sources are obtained at the first. Their amplitudes are estimated at second. The proposed method is generalized to cases of using a linear and flat digital antenna array. The efficiency of the proposed method is verified by estimating its potential accuracy based on the calculations of the Cramer-Rao lower bound and mathematical modeling**

Keywords—solving radar tasks, data transmission, the integrated radar and telecommunication system

I. Introduction

In recent years in the development of radar and telecommunication systems has been formed a direction that foresees the common use of OFDM (N-OFDM) signals, both in solving radar tasks and for data transmission, including using the MIMO principle [1-5]. The idea of creating an integrated radar-telecommunication system (RedCom) was proposed in [2-4, 6-8]. Several papers described the concept [7], structure [6-8], and modes of operation [7], based on the distribution of communication and radar tasks in time. The question of the functioning of the integrated radar and telecommunication system provided that jointly received radar and telecommunication services are not sufficiently investigated by the authors. The complexity of the demodulation of a mixture of different signals is the effect of a rectangular radio pulse signal on the OFDM (N-OFDM) signal response on the outputs of frequency filters synthesized by the Fast Fourier Transform (FFT) operation, which will be perceived as impulse noise and can be induced to a forced reduction QAM modulation order in the OFDM (N-OFDM) packet transmission. Therefore, it is advisable to switch to OFDM (N-OFDM) and pulsed signals by means of a digital antenna array (DAA). The proposed idea requires the development of appropriate algorithmic support and clarification of the conditions for its application.

II. The Privacy Problem

The purpose of the article is to improve the algorithmic provision of radar and OFDM (N-OFDM) signals demodulation, in case of use on the receiving side of the digital antenna array and to investigate the potential precision of their demodulation on base of the proposed method.

III. Proposed Solution

Consider a linear DAA with an equidistant placement of antenna elements in it. It should be noted that in this case, the task of allocating pulsed signals from the mixture with continuous signals, which are conditionally OFDM signals (N-OFDM), should be solved step by step. In the first stage, for example, in connection mode, the angular coordinates of the signal sources must be obtained. For this purpose, known radar guidance methods, including multi-signal over-differentiation algorithms, may be used. The received information about directions of arrival of signals allows proceeding to the next stage of processing - estimation of general signal amplitudes of each type. For this purpose, it is advisable to use the procedure proposed in [9], for which we will show the vector of voltage signals of the spatial receiving channels of the DAA in separate moments of time at the outputs of the digital diagramming module in the form:

$$U = QW + n, \tag{1}$$

where $Q = [Q_S \mid Q_P]$ – block matrix of the values of the directional patterns of the secondary spatial channels of the DAA in the directions to the OFDM (N-OFDM) (block Q_S) and pulse signal reflector (block Q_P) signals sources; $W = [W_S \mid W_P]^{\mathrm{T}}$- block-vector of generalized amplitudes of N-OFDM signals (block W_S) and amplitudes of pulse signals (block W_P); "T" - is the symbol of the matrix transposition operation, n - noise voltage vector.

For the selection of OFDM (N-OFDM) signals in the background of reflected pulse air targets, in the formation of an optimal estimation of the generalized amplitudes $\tilde{W} = (Q^T Q)^{-1} Q^T U$ vector, a segment of the vector $\tilde{W}$ corresponding to the OFDM (N-OFDM), block W_S, is calculated separately. Similarly, for the selection of pulsed signals in the OFDM (N-OFDM) background, the pulse signal amplitude segment is calculated - block W_P. Further on the received in the series of an analog-to-digital converter (ADC) samplings, estimates of the generalized OFDM amplitudes (N-OFDM) and pulse signals perform

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FFT procedures, which allows to synthesize the frequency filters necessary for the spectral selection of subcarriers of OFDM (N-OFDM) signals and measure the radial velocity of air targets [10-15].

In the general case, in the presence of M sources of OFDM (N-OFDM) signals and P pulse reflectors, matrix blocks Q of the directivity pattern for R of the secondary spatial channels of the DAA can be written as:

$$Q_S = \begin{bmatrix} Q_1(x_{1S}) & Q_1(x_{2S}) & \cdots & Q_1(x_{MS}) \\ Q_2(x_{1S}) & Q_2(x_{2S}) & \cdots & Q_2(x_{MS}) \\ \vdots & \vdots & \vdots & \vdots \\ Q_R(x_{1S}) & Q_R(x_{2S}) & \cdots & Q_R(x_{MS}) \end{bmatrix}, \qquad (2)$$

$$Q_P = \begin{bmatrix} Q_1(x_{1P}) & Q_1(x_{2P}) & \cdots & Q_1(x_{PP}) \\ Q_2(x_{1P}) & Q_2(x_{2P}) & \cdots & Q_2(x_{PP}) \\ \vdots & \vdots & \vdots & \vdots \\ Q_R(x_{1P}) & Q_R(x_{2P}) & \cdots & Q_R(x_{PP}) \end{bmatrix}, \qquad (3)$$

where $Q_r(x_m) = \left[sin\left(\frac{R}{2}[r - x_m]\right)\right]\left[sin\frac{1}{2}(r - x_m)\right]^{-1}$ – the diagram of the r-th secondary spatial channel synthesized by the Fast Fourier Transform (FFT) operation; $x_{mS(pP)} = \frac{2\pi}{\lambda} d\left(r - \frac{R-1}{2}\right) sin\,\theta_{mS(pP)}$ - generalized angular coordinate of *m-th* or *p-th* sources of signals relative to the normal to the DAA; λ - the wavelength of the central carrier signals N-OFDM and pulse signals; d - a distance between antenna elements of the DAA; R - the number of elements in the antenna array; $\theta_{mS(pP)}$ - angular coordinates of *m-th* or *p-th* sources of signals relative to the normal to the DAA.

The difference between further methods of processing is the different principles of forming FFT filters depending on the type of signals. In particular, for the selection of subcarriers OFDM (N-OFDM) signals FFTs should be carried out on the basis of reference unit estimates W_S, which are formed sequentially in time, at intervals of their follow-up. For pulse signals, FFT is performed over the arrays of reference blocks W_P, of the estimates taken for each of the gates of the range through the period of repetition of pulsed signals. For OFDM (N-OFDM) signals, the FFT procedure is performed once for a character interval, and for pulse signals - T times, where T - the number of gates in the range within the interval of its unambiguous measurement. Based on the received output voltages of synthesized frequency filters, demodulation OFDM (N-OFDM) signals is further demodulated, and estimates of trajectory parameters of air traffic movement are calculated.

In general, we consider the statements in the case of a plane equidistant DAA, which has the orientation diagrams of the secondary spatial channels, factorized in the form of a product of the directivity diagram in two angular planes. In the case of the structure of the DAA with $R \times K$ elements, the expression (1) can be rewritten using the block matrix product of Khatri-Rao:

$$U = \left[Q \;[\blacksquare]\; V\right] W + n, \qquad (4)$$

where $[\blacksquare]$ – is the symbol of the Khatri-Rao block product; $V = [V_S \mid V_P]$ – block matrix of values of the directional patterns of the secondary spatial channels of the DAA in a second angular plane, where

$$V_S = \begin{bmatrix} V_1(y_{1S}) & V_1(y_{2S}) & \cdots & V_1(y_{MS}) \\ V_2(y_{1S}) & V_2(y_{2S}) & \cdots & V_2(y_{MS}) \\ \vdots & \vdots & \vdots & \vdots \\ V_K(y_{1S}) & V_K(y_{2S}) & \cdots & V_K(y_{MS}) \end{bmatrix}, \qquad (5)$$

$$V_P = \begin{bmatrix} V_1(y_{1P}) & V_1(y_{2P}) & \cdots & V_1(y_{PP}) \\ V_2(y_{1P}) & V_2(y_{2P}) & \cdots & V_2(y_{PP}) \\ \vdots & \vdots & \vdots & \vdots \\ V_K(y_{1P}) & V_K(y_{2P}) & \cdots & V_K(y_{PP}) \end{bmatrix}, \qquad (6)$$

$$V_k(y_m) = \left[\sin\left(\frac{K}{2}[k - y_m]\right)\right]\left[\sin\frac{1}{2}(k - y_m)\right]^{-1}, \qquad (7)$$

$$y_{mS(pP)} = \frac{2\pi}{\lambda} d_y\left(k - \frac{K-1}{2}\right)\sin\theta_{mS(pP)} \sin\varepsilon_{mS(pP)}, \qquad (8)$$

$\varepsilon_{mS(pP)}$ – angular coordinates of *m-th* or *p-th* sources of signals, that relative to the normal to the DAA in the second angular plane; d_y – the distance between the antenna elements of the DAA in the corresponding angular plane.

In addition, as an argument to the directional diagram of the matrix Q should be used a generalized angular coordinate

$$x_{mS(pP)} = \frac{2\pi}{\lambda} d_x\left(r - \frac{R-1}{2}\right) sin\,\theta_{mS(pP)}\, cos\,\varepsilon_{mS(pP)}. \qquad (9)$$

Prospects for further research were the study of the potential of the proposed methods of signal processing based on the lower limit of Cramer-Rao, which would provide separate selection of communication signals and locations in radar and telecommunication MIMO systems while simultaneously solving communication and radar problems. The expression for the Cramer-Rao lower bound with respect to estimates of the dispersion of the vector of generalized amplitudes W can be written as:

$$\sigma_W^2 \geq \sigma_n^2 diag\left[[Q_S \mid Q_P]^T [Q_S \mid Q_P]\right]^{-1}, \qquad (10)$$

where σ_n^2 - the dispersion of noise in a separate time reference signal of the signal mix on the output of the second spatial channel of the linear DAA; $diag[Z]$ - a vector formed from the diagonal elements of the matrix Z.

When transferring the expression (10) to the dispersion of noise at the output of ADC, account should be taken of the effect of increasing the dispersion of noise when FFT is performed according to the expression $\sigma_n^2 = R\sigma_{ADC}^2$, where R - the dimension of spatial fast Fourier transform (FFT), in this case this is the number of elements of the DAA; σ_{ADC}^2 - the dispersion of noise on the output of the ADC. Therefore, the Equation (10) will be rewritten as:

$$\sigma_W^2 \geq \sigma_{ADC}^2 \cdot R \cdot diag\left[\left[Q_S \mid Q_P\right]^T \left[Q_S \mid Q_P\right]\right]^{-1}. \quad (11)$$

In the case of the use of a plane equidistant DAA from $R \times K$ elements having the orientation diagrams of the secondary spatial channels, factorized in the form of the product of the directivity patterns in two angular planes, the expression (10) can be rewritten to the form:

$$\sigma_W^2 \geq \sigma_n^2 diag\left[\left(Q\ [\blacksquare]\ V\right)^T \left(Q\ [\blacksquare]\ V\right)\right]^{-1} \quad (12)$$

Taking into account the identities known from [4]

$$\left[A\ [\blacksquare] B\right]^T = A^T [\square]\, B^T; \quad (13)$$

$$\left(\left[A_{ji}\right][\square]\left[B_{ji}\right]\right)\left(\left[K_{ik}\right][\blacksquare]\left[M_{ik}\right]\right) = \left[P_{jk} = \sum_i \left\{\left(A_{ji} \cdot K_{ik}\right) \circ \left(B_{ji} \cdot M_{ik}\right)\right\}\right] \quad (14)$$

where $[\square]$ - symbol of block face product of block matrices (the transposed block matrix product of Khatri-Rao) [4], the expression (12) can be rewritten in the form:

$$\sigma_W^2 \geq \sigma_n^2 diag\left[\left(Q^{\mathrm{T}} [\square]\ V^T\right)\left(Q\ [\blacksquare]\ V\right)\right]^{-1} = \sigma_n^2 diag\left[\left(\begin{bmatrix} Q_S^T \\ Q_P^T \end{bmatrix} [\square] \begin{bmatrix} V_S^T \\ V_P^T \end{bmatrix}\right)\left(\left[Q_S \mid Q_P\right]\ [\blacksquare]\ \left[V_S \mid V_P\right]\right)\right]^{-1} \quad (15)$$

or

$$\sigma_W^2 \geq \sigma_n^2 diag\left[\begin{bmatrix} Q_S^T Q_S \circ V_S^T V_S & Q_S^T Q_P \circ V_S^T V_P \\ Q_P^T Q_S \circ V_P^T V_S & Q_P^T Q_P \circ V_P^T V_P \end{bmatrix}\right]^{-1}, \quad (16)$$

where $\circ$ - symbol of Hadamard product.

Recalculation to noise at the output of the ADC is performed according to the expression $\sigma_n^2 = R \cdot K \cdot \sigma_{ADC}^2$, where R and K dimensional spatial fast Fourier transform (FFT) for two angular coordinates.

In particular, for expression (16) obtain:

$$\sigma_W^2 \geq RK\sigma_{ADC}^2 diag\left[\begin{bmatrix} Q_S^T Q_S \circ V_S^T V_S & Q_S^T Q_P \circ V_S^T V_P \\ Q_P^T Q_S \circ V_P^T V_S & Q_P^T Q_P \circ V_P^T V_P \end{bmatrix}\right]^{-1} \quad (17)$$

Inequalities (17) characterize the lower bound of the variances for estimates of the quadrature components of the generalized amplitudes of the signals obtained by expression:

$$W^c = Re\left(\left\{\left(Q\ [\blacksquare]\ V\right)^T \left(Q\ [\blacksquare]\ V\right)\right\}^{-1} \left(Q\ [\blacksquare]\ V\right)^T U\right), \quad (18)$$

$$W^S = Im\left(\left\{\left(Q\ [\blacksquare]\ V\right)^T \left(Q\ [\blacksquare]\ V\right)\right\}^{-1} \left(Q\ [\blacksquare]\ V\right)^T U\right), \quad (19)$$

where

$$\mathrm{W}^c = \left[W_{1S}^c\ \ W_{2S}^c\ \cdots\ W_{MS}^c \mid W_{1P}^c\ \ W_{2P}^c\ \cdots\ W_{PP}^c\right]^{\mathrm{T}},$$

$$\mathrm{W}^s = \left[W_{1S}^s\ \ W_{2S}^s\ \cdots\ W_{MS}^s \mid W_{1P}^s\ \ W_{2P}^s\ \cdots\ W_{PP}^s\right]^{\mathrm{T}}.$$

The obtained dispersion estimates can then serve as noise dispersions in calculating the accuracy of impulse and OFDM (N-OFDM) signal estimation at the subsequent stages of their digital processing. In particular, in the case of executing a series of sequentially formed time subvectors of estimates of the generalized amplitudes of N-OFDM signals

$$\mathrm{W_{Sn}} = \mathrm{W_{Sn}^C} + j\mathrm{W_{Sn}^S} = \left[W_{1\mathrm{Sn}}^c\ \ W_{2\mathrm{Sn}}^c\ \cdots\ W_{MSn}^c\right]^{\mathrm{T}} + j\left[W_{1\mathrm{Sn}}^s\ \ W_{2\mathrm{Sn}}^s\ \cdots\ W_{MSn}^s\right]^{\mathrm{T}}. \quad (20)$$

Fast Fourier Transform (FFT) expression for the Cramer-Rao lower bound dispersion of OFDM (N-OFDM) amplitude estimates of signals with the same frequency grid for all revenue streams has the form:

$$\sigma_A^2 \geq \sigma_w^2 \otimes \left(N \cdot diag\left[F^T F\right]^{-1}\right) \quad (21)$$

or

$$\sigma_W^2 \geq RK\sigma_{ADC}^2 diag\left[\begin{bmatrix} Q_S^T Q_S \circ V_S^T V_S & Q_S^T Q_P \circ V_S^T V_P \\ Q_P^T Q_S \circ V_P^T V_S & Q_P^T Q_P \circ V_P^T V_P \end{bmatrix}\right]^{-1} \otimes \left(N \cdot diag\left[F^T F\right]^{-1}\right), \quad (22)$$

where $F = \begin{bmatrix} \dot{F}_1(\omega_1) & \cdots & \dot{F}_1(\omega_T) \\ \vdots & \cdots & \vdots \\ \dot{F}_N(\omega_1) & \cdots & \dot{F}_N(\omega_T) \end{bmatrix}$ – matrix of amplitude-frequency characteristics (AFC) $\dot{F}_n(\omega_t)$ synthesized by the operation of FFT N frequency filters at carrier T frequencies of N-OFDM signals, $\otimes$ is the symbol of the Kroneker product of matrices. Similarly, the dispersion estimates of pulse signal parameters, in particular, Doppler frequency shift, of the time of arrival can be calculated.

IV. Discussion Of Future Extensions

In the course of further research, the properties of the obtained analytical relations by means of mathematical modeling in the Mathcad package were determined and the limits of the OFDM (N-OFDM) demodulation signals are established on the background of the action of the pulse signals of selection of air targets. At the same time, for the simplification of calculations, the assumption was made about the absence of active interference.

In the course of mathematical modeling, the mean square deviation (MSD) of the amplitudes estimates was determined in quants of the ADC. At the same time, the ADC bit size does not matter, because, for any bit, the size of the ADC quantum with the MSD noise must be aligned (for example, they should be equal in magnitude). Angular diversity of signal sources was set in the fractions of the main petal of the direction diagram of the DAA. At the same time, in order to reduce the volume of calculations during research, the step of changing the angular diversity of signal sources was set equal to 0.1 width of the main petal of the direction diagram in the range of angular distances from 1 to 0.1 width of the secondary beam of the direction diagram and 0.01 - in the range from 0.1 to 0.01 beam width.

Figures 1 to 3 illustrate results of mathematical modeling of the process of estimating the quadrature component amplitudes of the OFDM signal in a separate reference frame of the ADC on the background of the impulse radiation source with a different angular distance between them (horizontal axis). The values obtained at this MSD value of the estimates of the amplitude components are deferred vertically. The number of antenna elements in the DAA for the indicated illustrations was 4, 16 and 64. In addition to the values obtained by the MSD, the estimates of the amplitudes in the figures show the calculated value of the Cramer-Rao lower bound (CRLB) and the limits of its confidence interval for the reliability of 0.95.

The obtained results confirm the well-known scientific position that when the number of antenna elements of the DAA increases in R times of the CRLB, the values of the MSD of the amplitudes are reduced in $\sqrt{R}$ times. Most precisely this pattern works for orthogonal angular dispersion, that is, the width of the secondary ray, but its manifestation is noticeable and with smaller differences in the angular coordinates.

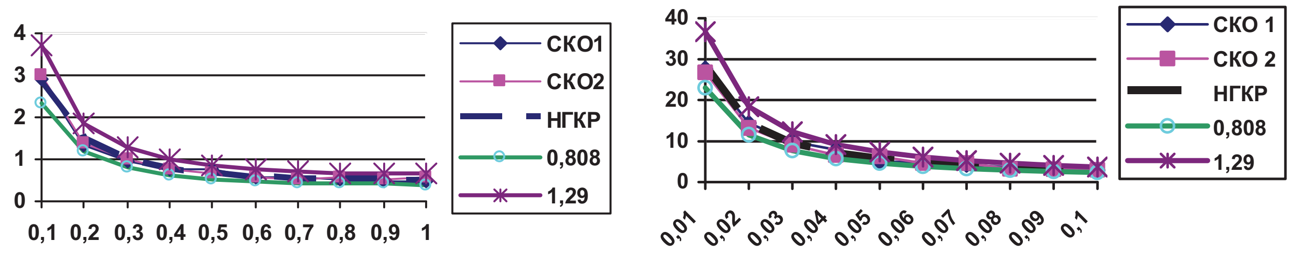

Fig. 1. MSD of error estimation of signal amplitudes after the first stage of 2-step evaluation (4-element DAA).

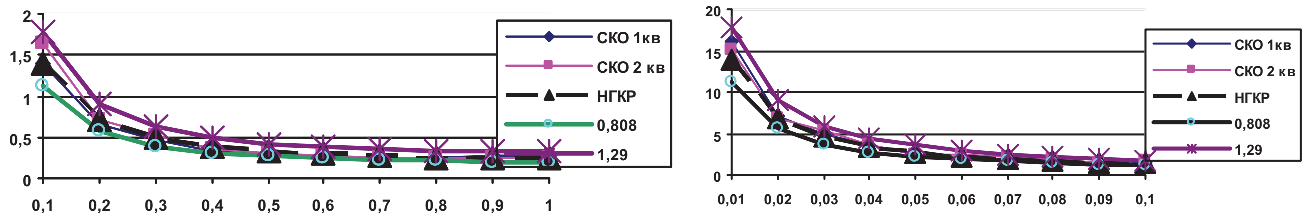

Fig. 2. MSD of error estimation of signal amplitudes after the first stage of 2-step evaluation (16-element DAA).

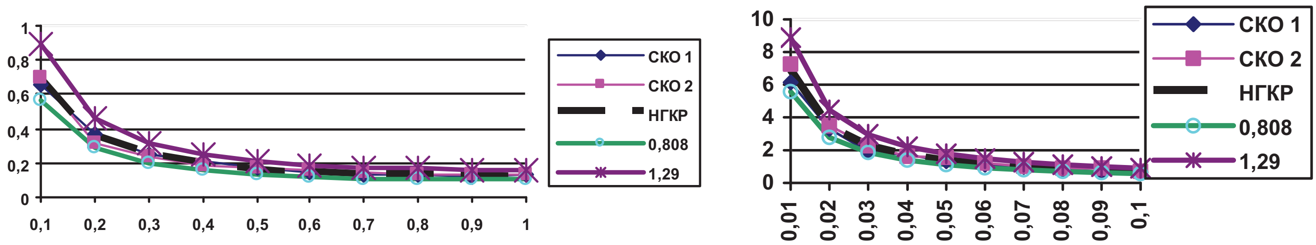

Fig. 3. MSD of error estimation of signal amplitudes after the first stage of 2-step evaluation (64-element DAA).

For example, at corner distributions of signal sources at 0.1 wavelength of the secondary beam of the directivity diagram, the upper limit of the confidence interval is about 1.8 quanta of ADC for a 16-cell DAA and 0.9 for a 64-element DAA, an increase in the number of antenna elements in 4 times leads to a two-fold decrease in the magnitude of the amplitudes MSD.

For the first time for the two-stage evaluation of the signal parameters, it has been established that irrespective of the number of antenna elements of the DAA in the angular resolution of two sources of less than 0.1 of the width of the secondary beam of the directional diagram, the value of the CRLB for the MSD of the quadrature component of the amplitudes of signals decreases in direct proportion to the reduction of angular diversity. For example, in Fig. 3, the upper limit of the confidence interval of the MSD (0.9 quanta of ADC) at a corner spacing of 0.01 is in 10 times smaller than the corresponding value (9 quanta of ADT) for angular dispersion of 0.1. The same can be traced on charts for the 16-element and 4-element DAA.

Based on the obtained results, it is possible to formulate an approximate empirical dependence for calculating the potentially possible value of the MSD estimation of the quadrature component of the signal amplitudes σ_a for the case of angular diversity of two sources by a value less than 0.05 of the width of the secondary ray of the directional diagram of the DAA:

$$\sigma_a \approx 0{,}552\frac{\sigma_{ADC}}{\Delta\sqrt{R}}, \quad (23)$$

where σ_{ADC} - the MSD noise at the output of the ADC in the fractions of the ADC quantum, Δ - the angular separation of radiation sources in the fractions of the secondary beam of the directional diagram of the DAA, R - the number of antenna elements in the DAA.

The proportional size of the angular spacing on the number of antenna elements of the DAA in determining the CRLB of MSD amplitude components is possible approximately to the angular distance of 0.2 bars of the beam of the directional diagram of the DAA. For example, according to the results of the conducted mathematical experiment, it was found out that at an angular spacing of 0.1 in 64-element DAA, the value of the CRLB of MSD amplitudes (0.6937914361) is approximately equal to the value of CRLB MSD (0.70878204683) for a 16-element DAA at angular intervals between sources 0,2. That is, in the specified interval of changes in angular distances with a fixed value of MSD amplitudes achieve a decrease in the maximum permissible angular diversity of signal sources in N times, that requires an increase in the number of antenna elements of the DAA in N^2 times.

In general, the conducted mathematical experiment has allowed to certify that at angular spacings of sources of pulsed and OFDM signals at a value, of the width of the secondary beam of the directional diagram of the DAA that not less than 0,75, the presence of the second source almost does not affect the accuracy of the estimation of amplitudes of signals. The same applies to the effects of active interference. If N-bit ADCs are used in one of the DAAs, one of whose digits are significant, then in the specified spatial sector, provided that there is no output signal of the signal mixture beyond the limits of the ADC`s aperture, it is potentially possible to provide a detachment from the impulse or interference source at a level of 6 × (N-1) dB. For example, for a 16-bit ADC we are talking about a 6 × 15 = 90 dB limit, which is sufficient for the selection of small-scale air targets using pulse signal with the simultaneous reception of OFDM (N-OFDM) communication signals. If the specified corner segment is narrowed, then the efficiency of the rejection from the additional source of signals, which is the indicator of the level of MSD of the amplitudes, will deteriorate.

It should also be noted that due to the accumulation of OFDM signals in time in the synthesis of frequency filters using the FFT, the specified values of the delay from the pulse (noise) signal will increase. However, as stated in [9], in practical implementation of the proposed approach, the estimates of the achievable level of the mutual influence inhibition of the sources of two signals will need to be adjusted with taking into account the negative factors due to the not-identical of the amplitude-frequency characteristics of the receiving channels and the directional patterns of the DAA antenna elements, limited accuracy of information about angular coordinates of signal sources, nonlinearity of analog segment of receiving channels in combination with non-linearity of aperture of ADC, etc.

Conclusion

Thus, the considered mathematical device allows to form mathematical models of the responses of receiving linear and plane DAA of radar and telecommunication systems in the case of simultaneous receipt of reflected signals from air targets of the pulses and OFDM (N-OFDM) signals providing data transmission functions. The proposed method of digital signal processing allows you to divide the pulse and N-OFDM signals into the receiving system for further measurement of trajectory parameters of air traffic movement and demodulation of information messages, which confirms the results of mathematical modeling. The obtained relations can be generalized to the case of multi-position signal processing, which requires to complicate the block structure of the corresponding matrices and to introduce additional indexes in their elements.

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