ISSN 0735-2727, Radioelectronics and Communications Systems, 2007, Vol. 50, No. 3, pp. 161–167. © Allerton Press, Inc., 2007.
Original Russian Text © V. I. Slyusar, A. N. Dubik, S. V. Voloshko, 2007, published in Izvestiya VUZ. Radioelektronika, 2007, Vol. 50, No. 3, pp. 61–70.

The MIMO Method for Transmission of Telecode Information

V. I. Slyusar, A. N. Dubik, and S. V. Voloshko
Kiev, Ukraine

Received in final form January 19, 2007

Abstract—A new method is suggested for signal processing in the pulsed MIMO system using the space of beams in the receiving digital antenna array. The paper also describes an approach to simultaneous treatment of the communication and radiolocation tasks based on combination of radio-pulse transmission of telecode data and multifrequency OFDM-sounding of air space.

DOI: 10.3103/S0735272707030089

A serious problem in multiposition radiolocation is transmission of radar information to the central control station. Particularly, severe requirements, in terms of data exchange speed, are imposed on the system of telecode communication if we deal with technology of sensor networks or cooperative processing of signals, when each of the radar system positions is regarded as an element of a widely spaced digital antenna array.

The purpose of this paper is to disclose a new approach to implementation of the system of telecode communication based on the methods used in MIMO (Multiple Input—Multiple Output) communication systems [1].

This work is an extension of the method described in [2] and is devoted to synthesis of procedures of processing of multisignal mixture at the output of the digital pattern-shaping system of demodulation of multisignal mixture at the output of the receiving digital antenna array (DAA) in the case of pulsed mode of operation of the MIMO system.

It is known that processing of signals in the receiving DAA can be performed after digital pattern-shaping, with transition to the “space of beams”, i.e., at the outputs of secondary spatial channels synthesized with the aid of the fast Fourier transform. Figure 1 illustrates directivity characteristics of the synthesized (with the aid of FFT) secondary spatial channels for a four dipole array. This approach has become the subject of much investigation, particularly, in connection with the design of MIMO systems. An example of such inquiry is a dissertation [3] devoted to preprocessing of signals, where we can also find a reference to another work [4] supposedly considered as pioneering in preprocessing of MIMO signals. It should be said nevertheless that the idea of such preliminary processing of signals in the reception segment of MIMO system was proposed quite independently in [2, 5].

By analogy with radiolocation, the advantage of the “space of beams” decision consists in spatial coherent accumulation of the received signals. As a result, in the presence of Gaussian noncorrelated noise, the signal-to-noise ratio for the output voltage of these secondary spatial channels can be raised by a factor proportional to the square root taken of the number of reception channels used in the antenna array (for example, by a factor of two in the case of four elements).

If applied to the telecode communication system, this technique also allows for effective protection from active interference acting, for instance, over one of major beams of the synthesized secondary reception channels. As for selection of the pulsed mode of operation, this principle offers a whole number of advantages over OFDM signals widely used in Communication Standards 802.16-2004, 802.16e, and draft version of Standard 802.11n. Particularly, as distinct from the OFDM method, in the pulsed mode of operation there is no necessity in observance of orthogonality of signals’ carrier frequencies. As a result, we can narrow the spectral band of the communication radio line—to set the same carrier frequency for all pulses radiated, and perform communication with movable objects, for example, with pilotless flying vehicles. Moreover, the suggested approach to construction of MIMO system permits to improve immunity of communication channels to unsanctioned access, and to raise the rate of data transmission to longer distances (against several hundred meters inherent in the known MIMO-system implementations based on OFDM).

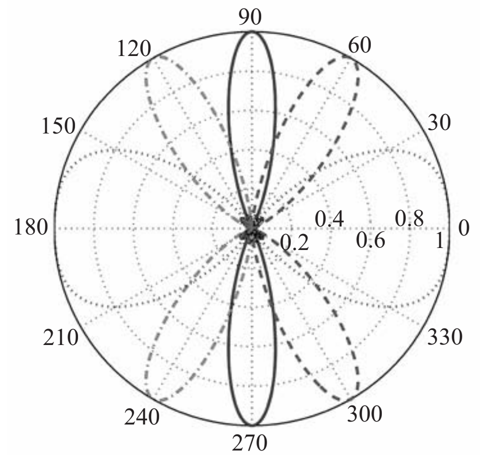

Fig. 1

On the other hand, there is a possibility for simultaneous application, in the telecode systems of data transmission, of the pulsed signals and their OFDM-alternative—to combine the treatment of communication tasks and those of MIMO radiolocation. It is known that application of MIMO systems is convenient not only in communication but also in so-called MIMO radar systems. The latter differ from ordinary systems in that every transmission channel uses autonomous radiation of its own sounding signal. It the last few years this approach has been discussed in foreign publications intensively [6]. In multiposition MIMO location of air targets, the above-mentioned multifrequency signal with orthogonal carriers (OFDM) is used. Being more advantageous from the energy viewpoint, the continuous sounding can be used for detecting multiple targets at far distances to the radar system, and reveal the spatial sectors and frequency ranges in which radiation sources are acting. The subsystem of sounding stations, radiating a very weak continuous signal, can be rather efficiently masked from anti-radar missiles. The pulsed sounding in the suggested method is intended only for data transmission tasks. Except for operation in the pulsed conditions dealing with large ranges, we thus do away with the problem of reticence of pulsed sounding radio aids (due to low radiation power).

It should be noted that reception of pulsed signals in DAA could be performed rather efficiently against the background of continuous reflections, whose frequency is coincident with the carrier. Here we can point to several approaches to suppression of these reflections, and combination of these techniques permits to give up the extension of the nomenclature of reception stations (we mean the use of different volumes of the accumulated signal samples). In pulsed reception conditions, the accumulation time is considerably less (by several orders) and is limited by pulse duration. The energy loss in this case can be covered through bringing nearer the far boundary of telecode data transmission, or by improving the spatial selectivity of radiation pattern of the receiving antenna system. The coherent accumulation of continuous signals, due to its large duration, leads to reducing the pulse signals to the background noise level. Since the pulsed sounding does not introduce any substantial changes in processing of continuous reflections from air target, selection of information-bearing pulsed signals is more important. The use of digital processing makes it possible to overcome this problem: for example, we can subtract the continuous signal from the voltages of pulsed-continuous mixture by any known method of jamming compensation.

As a whole, the above approach is somewhat new and, in the scientific aspect, deserves a more detailed investigation: From the viewpoint of its numerous possible applications, it may be regarded as a new and promising line of scientific research.

Among several realizations of MIMO pulsed method in telecode communications, of particular interest is its modification aimed at raising the communication range. Its essence consists in radiation, by each transmission channel, of solitary pulses shifted in time, so that at the reception side we have a mixture of signals overlapped in time. In this event the transmitting antenna, during processing the many-pulse packet, is regarded in the reception segment as a point-like radiator.

In all cases mentioned above, when we pass to formal description of linear reception array instead of tabular values of discrete functions corresponding to directive properties of secondary spatial channels $Q_r(x_m)$, we have to use the relationships of directivity characteristics of a spatial FFT-filter [2]:

$$Q_r(x_m) = \frac{\sin \frac{R}{2}\left(\frac{2d}{\lambda}\pi \sin x_m - \frac{2r}{R}\pi\right)}{\sin \frac{1}{2}\left(\frac{2d}{\lambda}\pi \sin x_m - \frac{2r}{R}\pi\right)}, \tag{1}$$

where R is the number of antenna elements, d is the distance between antenna elements in the array, and λ is the wavelength.

An important issue, associated with application of the spatial FFT at the stage of preprocessing of a received packet of pulsed signals containing telecode data, is the need in elimination of spurious phase increments. The latter may arise due to operations of weighted phased accumulation of spatial samples. Analysis of analytical description of responses of the synthesized FFT-channels shows that the above phase distortion can be by convention divided in two groups.

The first one depends on ordinal number of a synthesized FFT-beam and can be compensated by phase correction of voltage responses of secondary spatial channels. Provided the phase center of the signal sample and of the linear DAA corresponds to the first (in ordinal numbers) antenna element (channel), for such phase correction we have perform the required phase shift of voltages of secondary spatial channels—to multiply them by the complex value

$$\exp\left(j\pi \frac{r(R-1)}{R}\right), \tag{2}$$

where r is the current number of the secondary spatial channel of the linear equidistant DAA ($r = 1,\ldots, R$), and R is the dimension (number of points) of FFT.

Owing to this phase turn, we can apply, for description of analytical responses of the receiving DAA in the space of beams, the same relationships [2] as in the case of ordinary MIMO system.

The second group of phase distortions, arising during FFT implementation, is invariant to ordinal numbers of secondary spatial channels and, being the same for all responses of FFT-beams, is dictated only by the selected phase center of DAA and by direction of pulse signal arrival. Its compensation should be performed at the stage of assessment of quadrature components of signal amplitudes.

Then, after correcting the responses of secondary spatial channels, the amplitude estimation reduces to treatment of equation systems set up based on samples of the signal mixture voltages. For this purpose we can use the deterministic approach, when the measurement noise is not taken into account, and the algebraic equation system sought for is resolved by the well-known Kramer rule. The main prerequisite for obtaining this solution is equality between the unknown amplitudes of signals and the number of voltage digital samples used in the processing. The latter file (of the relevant dimension) can be obtained either by usage of the required number of reception channels' responses in a single time sample, or with the aid of their sequential sampling in time.

Let us illustrate the respective variant of the sampling procedure by an example of a two-element antenna array used for reception of a pulse packet of four signals, whose arrangement in time permits to pick out at least a single pair of voltage samples (for instance, with ordinal numbers t and $(t + 1)$), where all the pulses are present simultaneously. Assume also that each of the four transmission channels radiates a solitary pulse, and the discrete samples of the functions of pulse signal envelopes $K(s_t - z_m)$ are known exactly (with an accuracy of discretization period). With the aid of the model of reception DAA response, considered above, we can set up, based on this pair of samples of our array, a system of four equations

$$\begin{cases} \dot U_{1,t} = \dot a_1 Q_1(x_1)K(s_t - z_1) + \dot a_2 Q_1(x_2)K(s_t - z_2) + \dot a_3 Q_1(x_3)K(s_t - z_3) + \dot a_4 Q_1(x_4)K(s_t - z_4), \\ \dot U_{1,t+1} = \dot a_1 Q_1(x_1)K(s_{t+1} - z_1) + \dot a_2 Q_1(x_2)K(s_{t+1} - z_2) + \dot a_3 Q_1(x_3)K(s_{t+1} - z_3) + \\ + \dot a_4 Q_1(x_4)K(s_{t+1} - z_4), \\ \dot U_{2,t} = \dot a_1 Q_2(x_1)K(s_t - z_1) + \dot a_2 Q_2(x_2)K(s_t - z_2) + \dot a_3 Q_2(x_3)K(s_t - z_3) + \\ + \dot a_4 Q_2(x_4)K(s_t - z_4), \\ \dot U_{2,t+1} = \dot a_1 Q_2(x_1)K(s_{t+1} - z_1) + \dot a_2 Q_2(x_2)K(s_{t+1} - z_2) + \dot a_3 Q_2(x_3)K(s_{t+1} - z_3) + \\ + \dot a_4 Q_2(x_4)K(s_{t+1} - z_4), \end{cases} \tag{3}$$

where z_m is the first of ADC samples, obtained during existence of the mth pulse, and s_t is the ordinal number of ADC sample.

The solution to equation system (3) in terms of unknown amplitudes of the pulsed information carriers represents the following estimates:

$$\dot{a}_1 = \frac{\begin{vmatrix} \dot{U}_{1,t} & Q_1(x_2)K(s_t - z_2) & Q_1(x_3)K(s_t - z_3) & Q_1(x_4)K(s_t - z_4) \\ \dot{U}_{1,t+1} & Q_1(x_2)K(s_{t+1} - z_2) & Q_1(x_3)K(s_{t+1} - z_3) & Q_1(x_4)K(s_{t+1} - z_4) \\ \dot{U}_{2,t} & Q_2(x_2)K(s_t - z_2) & Q_2(x_3)K(s_t - z_3) & Q_2(x_4)K(s_t - z_4) \\ \dot{U}_{2,t+1} & Q_2(x_2)K(s_{t+1} - z_2) & Q_2(x_3)K(s_{t+1} - z_3) & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}}{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & Q_1(x_2)K(s_t - z_2) & Q_1(x_3)K(s_t - z_3) & Q_1(x_4)K(s_t - z_4) \\ Q_1(x_1)K(s_{t+1} - z_1) & Q_1(x_2)K(s_{t+1} - z_2) & Q_1(x_3)K(s_{t+1} - z_3) & Q_1(x_4)K(s_{t+1} - z_4) \\ Q_2(x_1)K(s_t - z_1) & Q_2(x_2)K(s_t - z_2) & Q_2(x_3)K(s_t - z_3) & Q_2(x_4)K(s_t - z_4) \\ Q_2(x_1)K(s_{t+1} - z_1) & Q_2(x_2)K(s_{t+1} - z_2) & Q_2(x_3)K(s_{t+1} - z_3) & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}},$$

$$\dot{a}_2 = \frac{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & \dot{U}_{1,t} & Q_1(x_3)K(s_t - z_3) & Q_1(x_4)K(s_t - z_4) \\ Q_1(x_1)K(s_{t+1} - z_1) & \dot{U}_{1,t+1} & Q_1(x_3)K(s_{t+1} - z_3) & Q_1(x_4)K(s_{t+1} - z_4) \\ Q_2(x_1)K(s_t - z_1) & \dot{U}_{2,t} & Q_2(x_3)K(s_t - z_3) & Q_2(x_4)K(s_t - z_4) \\ Q_2(x_1)K(s_{t+1} - z_1) & \dot{U}_{2,t+1} & Q_2(x_3)K(s_{t+1} - z_3) & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}}{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & Q_1(x_2)K(s_t - z_2) & Q_1(x_3)K(s_t - z_3) & Q_1(x_4)K(s_t - z_4) \\ Q_1(x_1)K(s_{t+1} - z_1) & Q_1(x_2)K(s_{t+1} - z_2) & Q_1(x_3)K(s_{t+1} - z_3) & Q_1(x_4)K(s_{t+1} - z_4) \\ Q_2(x_1)K(s_t - z_1) & Q_2(x_2)K(s_t - z_2) & Q_2(x_3)K(s_t - z_3) & Q_2(x_4)K(s_t - z_4) \\ Q_2(x_1)K(s_{t+1} - z_1) & Q_2(x_2)K(s_{t+1} - z_2) & Q_2(x_3)K(s_{t+1} - z_3) & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}},$$

$$\dot{a}_3 = \frac{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & Q_1(x_2)K(s_t - z_2) & \dot{U}_{1,t} & Q_1(x_4)K(s_t - z_4) \\ Q_1(x_1)K(s_{t+1} - z_1) & Q_1(x_2)K(s_{t+1} - z_2) & \dot{U}_{1,t+1} & Q_1(x_4)K(s_{t+1} - z_4) \\ Q_2(x_1)K(s_t - z_1) & Q_2(x_2)K(s_t - z_2) & \dot{U}_{2,t} & Q_2(x_4)K(s_t - z_4) \\ Q_2(x_1)K(s_{t+1} - z_1) & Q_2(x_2)K(s_{t+1} - z_2) & \dot{U}_{2,t+1} & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}}{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & Q_1(x_2)K(s_t - z_2) & Q_1(x_3)K(s_t - z_3) & Q_1(x_4)K(s_t - z_4) \\ Q_1(x_1)K(s_{t+1} - z_1) & Q_1(x_2)K(s_{t+1} - z_2) & Q_1(x_3)K(s_{t+1} - z_3) & Q_1(x_4)K(s_{t+1} - z_4) \\ Q_2(x_1)K(s_t - z_1) & Q_2(x_2)K(s_t - z_2) & Q_2(x_3)K(s_t - z_3) & Q_2(x_4)K(s_t - z_4) \\ Q_2(x_1)K(s_{t+1} - z_1) & Q_2(x_2)K(s_{t+1} - z_2) & Q_2(x_3)K(s_{t+1} - z_3) & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}},$$

$$\dot{a}_4 = \frac{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & Q_1(x_2)K(s_t - z_2) & Q_1(x_3)K(s_t - z_3) & \dot{U}_{1,t} \\ Q_1(x_1)K(s_{t+1} - z_1) & Q_1(x_2)K(s_{t+1} - z_2) & Q_1(x_3)K(s_{t+1} - z_3) & \dot{U}_{1,t+1} \\ Q_2(x_1)K(s_t - z_1) & Q_2(x_2)K(s_t - z_2) & Q_2(x_3)K(s_t - z_3) & \dot{U}_{2,t} \\ Q_2(x_1)K(s_{t+1} - z_1) & Q_2(x_2)K(s_{t+1} - z_2) & Q_2(x_3)K(s_{t+1} - z_3) & \dot{U}_{2,t+1} \end{vmatrix}}{\begin{vmatrix} Q_1(x_1)K(s_t - z_1) & Q_1(x_2)K(s_t - z_2) & Q_1(x_3)K(s_t - z_3) & Q_1(x_4)K(s_t - z_4) \\ Q_1(x_1)K(s_{t+1} - z_1) & Q_1(x_2)K(s_{t+1} - z_2) & Q_1(x_3)K(s_{t+1} - z_3) & Q_1(x_4)K(s_{t+1} - z_4) \\ Q_2(x_1)K(s_t - z_1) & Q_2(x_2)K(s_t - z_2) & Q_2(x_3)K(s_t - z_3) & Q_2(x_4)K(s_t - z_4) \\ Q_2(x_1)K(s_{t+1} - z_1) & Q_2(x_2)K(s_{t+1} - z_2) & Q_2(x_3)K(s_{t+1} - z_3) & Q_2(x_4)K(s_{t+1} - z_4) \end{vmatrix}}. \quad (4)$$

To obtain the optimal estimates of signal amplitude components, we suggest the maximum likelihood method. In the presence of Gaussian non-correlated noise the method can be reduced to minimization of the scalar function

$$L = \{\dot{\mathbf{U}} - P\dot{A}\}^* \{\dot{\mathbf{U}} - P\dot{A}\} = \min,$$

where $\dot{\mathbf{U}}$ is the vector of complex-valued samples of signal mixture voltages at ADC output; P is the signal matrix, whose entries represent the product of the directivity characteristics of the secondary spatial channels $Q_r(x_m)$ by discrete samples of functions of pulse signal envelopes with regard for their known (with an accuracy of the discretization period) mutual position in time; and $\dot{A}$ is the vector of quadrature components of signal amplitudes.

The sought estimates of the vectors of quadrature components of amplitudes can be represented in the known form

$$A^c = \mathrm{Re}(\{P^T P\}^{-1} P^T \dot{\mathbf{U}}), \quad A^s = \mathrm{Im}(\{P^T P\}^{-1} P^T \dot{\mathbf{U}}), \tag{5}$$

where $A^c = [a_1^c \cdots a_M^c]^T$, $A^s = [a_1^s \cdots a_M^s]^T$; Re is the real part of the complex vector; Im is its imaginary part; T denotes transposition; and a_m^c, a_m^s are quadrature components of signal amplitudes.

It is essential that estimates (5) can be obtained from a redundant number of voltage samples, which exceeds the dimension of the file necessary for setting up the normal system of equations. Because of this, such variant is preferable to the deterministic alternative considered above.

The estimates of quadrature components of partial pulse signal amplitudes, obtained from (4) and (5), involve phase errors arising from the spatial FFT procedure and not depending on the number of a spatial channel. Now it is pertinent to come back to the issue of elimination of this phase distortion. We could introduce a compensating correction into ultimate response at FFT-beam output, or create phase predistortion of signal sample readings prior to spatial FFT. However, in our situation it is expedient to combine compensation of phase distortions of both types within the framework of a single procedure, so that the respective phase shift should be performed after synthesis of the “space of beams”.

In the case under consideration, when the phase center of the linear DAA corresponds to the first antenna element (channel), every complex-valued sample of voltages at the output of the virtual reception channel must be shifted in phase.

To make the analytical derivation simpler, the compensation will be performed after implementing the spatial FFT. In order to realize correction of M pulsed signals, we have to carry out several preliminary transformations of vector estimates (5) of amplitude components of received signals (these components have been obtained in the course of demodulation). The transformation include the following operations:

a) to pool the column vectors of quadrature components’ estimates (5) into a single block matrix

$$\widehat{A} = [A^c \;\vdots\; A^s] = \begin{bmatrix} a_1^c & \vdots & a_1^s \\ \vdots & & \vdots \\ a_M^c & \vdots & a_M^s \end{bmatrix}; \tag{6}$$

b) to transpose the generated block matrix of the quadrature component estimates:

$$\widehat{A}^T = \begin{bmatrix} A^{c^T} \\ \cdots \\ A^{s^T} \end{bmatrix} = \begin{bmatrix} [a_1^c & \cdots & a_M^c] \\ \cdots & \cdots & \cdots \\ [a_1^s & \cdots & a_M^s] \end{bmatrix}, \tag{7}$$

c) to decompose into vectors [7] the transposed block matrix of the amplitude components

$$\mathrm{vec}[A^T] = \mathrm{vec}\begin{bmatrix} A^{c^T} \\ \hline A^{s^T} \end{bmatrix} = \mathrm{vec}\begin{bmatrix} [a_1^c & \cdots & a_M^c] \\ \hline [a_1^s & \cdots & a_M^s] \end{bmatrix} = \begin{bmatrix} a_1^c \\ a_1^s \\ \hline a_2^c \\ a_2^s \\ \hline \vdots \\ \hline a_M^c \\ a_M^s \end{bmatrix} \tag{8}$$

Now let us put down the sought expression for amplitude correction of M signals based on the procedure of ordinary block matrix product:

$$[\widetilde{A}_m] = \begin{bmatrix} \widetilde{a}_1^c \\ \widetilde{a}_1^s \\ \hline \cdots \\ \hline \widetilde{a}_m^c \\ \widetilde{a}_m^s \\ \hline \cdots \\ \hline \widetilde{a}_M^c \\ \widetilde{a}_M^s \end{bmatrix} = [C_m][\times](\mathrm{vec}[A^T])$$

$$= \begin{bmatrix} \cos\frac{K_1}{2} & \sin\frac{K_1}{2} \\ -\sin\frac{K_1}{2} & \cos\frac{K_1}{2} \\ \hline \cdots & \cdots \\ \hline \cos\frac{K_m}{2} & \sin\frac{K_m}{2} \\ -\sin\frac{K_m}{2} & \cos\frac{K_m}{2} \\ \hline \cdots & \cdots \\ \hline \cos\frac{K_M}{2} & \sin\frac{K_M}{2} \\ -\sin\frac{K_M}{2} & \cos\frac{K_M}{2} \end{bmatrix} [\times] \begin{bmatrix} \widehat{a}_1^c \\ \widehat{a}_1^s \\ \hline \cdots \\ \hline \widehat{a}_m^c \\ \widehat{a}_m^s \\ \hline \cdots \\ \hline \widehat{a}_M^c \\ \widehat{a}_M^s \end{bmatrix} = \begin{bmatrix} \widehat{a}_1^c \cos\frac{K_1}{2} + \widehat{a}_1^s \sin\frac{K_1}{2} \\ \widehat{a}_1^s \cos\frac{K_1}{2} - \widehat{a}_1^c \sin\frac{K_1}{2} \\ \hline \cdots \\ \hline \widehat{a}_m^c \cos\frac{K_m}{2} + \widehat{a}_m^s \sin\frac{K_m}{2} \\ \widehat{a}_m^s \cos\frac{K_m}{2} - \widehat{a}_m^c \sin\frac{K_m}{2} \\ \hline \cdots \\ \hline \widehat{a}_M^c \cos\frac{K_M}{2} + \widehat{a}_M^s \sin\frac{K_M}{2} \\ \widehat{a}_M^s \cos\frac{K_M}{2} - \widehat{a}_M^c \sin\frac{K_M}{2} \end{bmatrix}, \tag{9}$$

where [×] denotes the block ordinary matrix product.

The argument of the correcting multiplier K_m will depend on the type of antenna array (linear or plane DAA). For linear DAA this argument has to involve a phase correction χ_m for compensation of distortion at the outputs of synthesized spatial filters:

$$\chi_n = (R-1)\frac{2\pi d}{\lambda}\sin\theta_n, \tag{10}$$

where θ_n is angular direction of arrival of the nth signal.

Thus the expression permitting to determine K_m in the case of linear DAA has the form $K_m = \chi_m$.

A peculiar feature of design decisions, often considered in literature on MIMO systems, is the use of linear antenna arrays both in the receiving and transmitting segments. In the general case, however, in MIMO channels we may face scattering of signals not only in the horizontal but also in vertical plane. Particularly, the two-coordinate re-reflection of signals must be considered when extending the MIMO principle to simultaneous treatment of communication and radar tasks for broken country. In this event the "multibeamness" of radio wave propagation results in emergence of signals coming to the receiving array at various angles both in the horizontal and vertical plane. Because of this, it would be interesting to extend the considered variants of mathematical description of output signals of the receiving DAA to the two-dimensional case of shaping the secondary spatial channels (the "space of beams").

The suggested approach to construction of communication systems based on MIMO-principle adds to functional abilities of multiposition radar systems, and opens new prospects for their development.

REFERENCES

1. V. I. Slyusar, MIMO Systems: Design Principles and Signal Processing, Elektronika: Nauka, Tekhnologiya, Biznes, No. 8, 52 (2005).
2. V. I. Slyusar and A. N. Dubik, Izv. VUZ. Radioelektronika **49**(3), 75 (2006).
3. Pallav Sudarshan, PhD Thesis in Philosophy Electrical Engineering (North Carolina State University, 2004), http://www.lib.ncsu.edu/theses/available/etd-01042005-200617/unrestricted/etd.pdf.
4. A. F. Molisch, X. Zhang, S. Y. Kung, J. Zhang, "FFT-Based Hybrid Antenna Selection Schemes for Spatially Correlated MIMO Channels" in: *IEEE Int. Symposium on Personal, Indoor and Mobile Radio Commun., September 2003* (Beijing, China, 2003).
5. V. I. Slyusar and I. V. Titov, Izv. VUZ. Radioelektronika **47**(8), 14 (2006).
6. D. J. Rabideau and P. A. Parker, in: Project Report DAR-4 "Ubiquitous MIMO Multifunction Array Radar... and the Role of Time-Energy Management in Radar" (Massachusetts Institute of Technology—Lincoln Laboratory, 10 March 2004), http://handle.dtic.mil/100.2/ADA421233.
7. V. I. Slyusar, Izv. VUZ. Radioelektronika **46**(10), 15 (2003).