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CORRECTION OF CHARACTERISTICS OF TRANSMITTING CHANNELS IN AN ACTIVE DIGITAL ANTENNA ARRAY

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The paper describes a new method for correction of amplitude-frequency responses of DAA transmitting channels in the case of their nonidentity.

In digital shaping of beams in antenna arrays used both for reception and transmission (in communications, for example), a necessity arises for correction of amplitude-frequency characteristics of transmission channels of an active DAA, whose functional diagram is shown in Fig. 1. However, until recently this issue has received little attention. Thus the purpose of this paper is development of methods of correction of characteristics of active DAA transmission channels when the correction is performed based on special test (pilot) signals radiated by the antennas.

The principle of the proposed correction consists in preliminary weighting of quadrature components of signal voltages, applied to digital-to-analog converters (DAC), with the aid of complex correction coefficients (CC) in conformity to an expression similar to that given in [1–3] for a flat array:

$$\tilde{U}^c_{rq_i} = U^c_{rq_i} \cdot \alpha^c_{rq} - U^s_{rq_i} \cdot \alpha^s_{rq}, \quad \tilde{U}^s_{rq_i} = U^s_{rq_i} \cdot \alpha^c_{rq} + U^c_{rq_i} \cdot \alpha^s_{rq}, \tag{1}$$

where α^c_{rq} and α^s_{rq} are cosine and sine components of the correction coefficient of the response of the DAA rqth channel of the transmitting DAA located in the rth row of the qth column; $U^c_{rq_i}$ and $U^s_{rq_i}$ are uncorrected quadrature components of the signal voltages applied to the DAC at the ith time instant, where these components correspond to the required phase-amplitude field distribution on the aperture of the transmitting antenna array; and $\tilde{U}^c_{rq_i}$ and $\tilde{U}^s_{rq_i}$ are corrected values of the orthogonal components of signal voltages applied to the DAC.

To calculate the correction coefficients in (1), we may use estimates of the amplitude components of signals radiated by the transmitting active DAA which is phased in its normal direction. Then the procedure of CC calculation, for instance, for a flat DAA composed of $R \times Q$ elements, can be written, as in [1–3], in the form

$$\alpha^c_{rq} = \frac{A^c_{rq} \cdot V^c + A^s_{rq} \cdot V^s}{A^{c^2}_{rq} + A^{s^2}_{rq}}, \quad \alpha^s_{rq} = \frac{A^c_{rq} \cdot V^s - A^s_{rq} \cdot V^c}{A^{c^2}_{rq} + A^{s^2}_{rq}}, \tag{2}$$

where A^c_{rq} and A^s_{rq} are estimates of the amplitude components of the signal radiated by the rqth channel; and $V^{c(s)}$ are the quadrature components of voltages of the channel selected as the standard (reference) one.

In order to measure the amplitudes A^c_{rq} and A^s_{rq} of the radiated signals, we propose to use the concept embedded in communication systems of MIMO type (Multiple–Input–Multiple–Output). Here every partial channel of the transmitting DAA, in the event of their simultaneous radiation, is regarded as an independent source of message [Bharath Bhoopalam. MIMO Channel Capacity and Space Time Coding. —http://www.ece.drexel.edu/courses/ECE-T612/Talks/B_Bhoopalam.pdf].

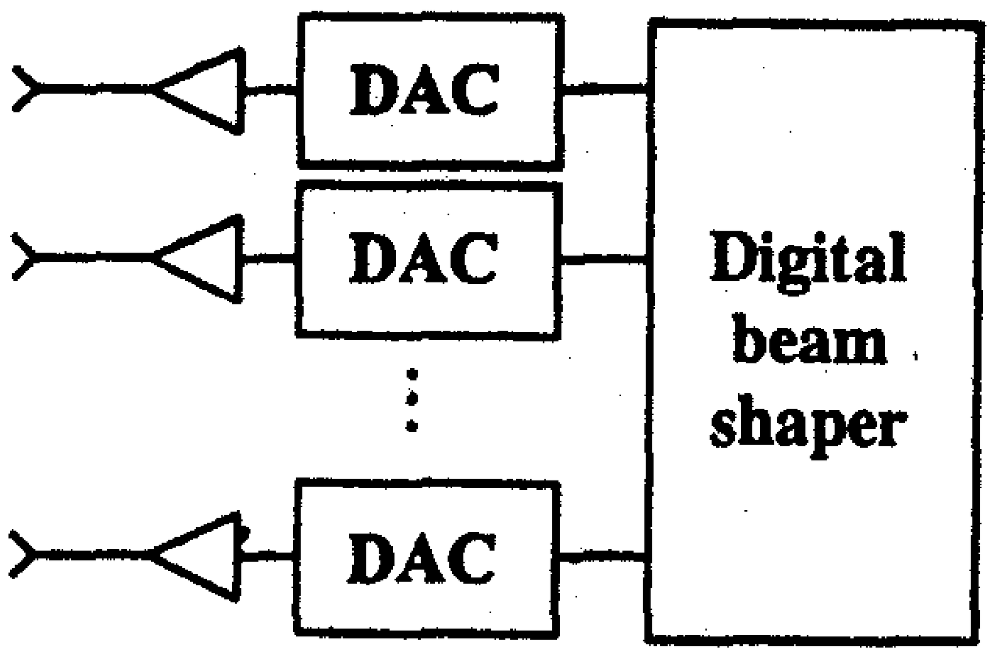


Fig. 1

In this case the role of measurer of transfer characteristics of an active array will be best played by the receiving system, which later will interact with the antenna under test. In particular, it may be some other, reception DAA, whose channel characteristics have been corrected beforehand by the methods suggested in [1–3]. In the general case, N transmitters may be coupled to R receivers ($R \geq N$). We shall consider this very case.

Of course, in this situation we could feed the channels of the transmitting DAA one after another, i.e., pass to the MISO type system (Multiple–Input–Single–Output). However, the time of testing grows substantially, and in principle we cannot guarantee that the channels' parameters remain unchanged during the whole test. In addition, the separate work of the channels during radiation gives no way of checking their mutual influence. Another alternative, called SIMO (Single–Input–Multiple–Output), implies the use of a single-channel receiver, rather than a passive DAA, as a measurer. Although such a solution seems attractive due to its cheapness, here we come up against some errors arising from different lengths of signal paths from radiator outputs of the active DAA to the phase center of the receiver antenna element. When dealing with a passive DAA, these errors are absent due to cophasal summation of voltages in the totality of reception channels, and shaping of a so-called amplitude multi-channel analyzer.

Since the distance between radiators in antenna arrays usually does not exceed several wavelengths of the carrier frequency employed for transmission, in our efforts to resolve the problem of measurement of parameters of signals simultaneously formed by every antenna element of the DAA under test, we cannot use the angular resolution of sources in a rather wide range of distances in the Rayleigh sense. On the other hand, in order to obtain the super-Rayleigh resolution in the angular coordinate in the event of large distance between the sources, too large a signal-to-noise ratio is required. For this reason, the maximum range (distance) of MIMO-systems is low enough and rarely exceeds a hundred meters.

In order to raise the estimation accuracy of quadrature components of amplitudes, we propose to use, in the transmitting channels, the orthogonal-in-frequency signals — as in the OFDM method. In this case for all transmitting modules we set the same duration of OFDM frame, and the permissible number of signals $T = K/N$ in each module is defined by the ratio between the number K of FFT points used for shaping the array of frequency filters in the receiving DAA, and the number N of the transmitting channels under test.

It is essential that for the method of correction proposed in this work the nominal values of signal frequencies in the transmitting DAA channels must not repeat. In other words, for every transmitting channel we have to allocate its own set of orthogonal frequencies, not overlapping with the totality of frequencies used for other transmitters. Fulfillment of these conditions enables us to implement the separate measurement, in the receiving DAA, of quadrature components of amplitudes of signals formed by partial radiators of the active array under test, in shifting the radiators as close as possible to each other along the generalized angular coordinate. This makes it possible to establish a considerable spacing between both arrays — the DAA to be tested and the measuring one, limited only by the required accuracy of the quadrature components' estimation (by the signal-to-noise ratio).

For evaluation of the correction coefficients, we suggest applying continuous signals, equal in amplitude and phase, to digital-to-analog converters of DAA transmitting channels. Calculation of the correction coefficients for subsequent introduction of predistortions in the transmitting channels will be performed for identical input coded sequences radiated into the free space by the partial channels of the active DAA, and under the assumption that the frequencies of the radiated signals and angular directions of their arrival are exactly known.

To calculate the quadrature components of the amplitudes, represent the responses of FFT-filters, formed in one channel of the receiving OFDM-system, as

$$U = PA + n \tag{3}$$

where A is the vector of signals' complex amplitudes,

$$P = \begin{bmatrix} V_1(f_1) & V_1(f_2) & V_1(f_3) & \dots & V_1(f_M) \\ V_2(f_1) & V_2(f_2) & V_2(f_3) & \dots & V_2(f_M) \\ V_3(f_1) & V_3(f_2) & V_3(f_3) & \dots & V_3(f_M) \\ \dots & \dots & \dots & \dots & \dots \\ V_D(f_1) & V_D(f_2) & V_D(f_3) & \dots & V_D(f_M) \end{bmatrix}$$

is the AFR-matrix of D FFT-filters used for reception of pilot signals out of T synthesized ones ($D \le T$),

$$V_i(f_m) = [\sin T[(i\pi/T) - f_m]] \times [\sin[(i\pi/T) - f_m]]^{-1}$$

is the value of AFR of the ith frequency filter, synthesized by FFT, at the frequency f_m, $m = \overline{1, M}$ is the signal number, and f_m is a frequency (frequencies), from the totality of allocated ones, expressed in fractions of the width of FFT-filter response.

Then the estimates of amplitude components, corresponding to model (3), can be easily obtained by the maximum likelihood method if we differentiate, with respect to the vector of unknown amplitudes A, the function of residues

$$L = (U - PA)^*(U - PA) = \min. \tag{4}$$

Upon resolving the respective likelihood equation, the sought vector of the amplitude components' estimates takes the known form

$$A = (P^H P)^{-1} P^H U. \tag{5}$$

If we use a linear DAA for reception of OFDM-signals, and the signals from G transmitting channels of the tested linear active array arrive simultaneously, then the response of such OFDM-system can be represented with the aid of (4) provided that the matrix P is written as

$$P = F[\otimes]V \tag{6}$$

where $[\otimes]$ denotes the block Kronecker product;

$$F = \begin{bmatrix} F_1(x_1) & F_1(x_2) & & F_1(x_G) \\ \vdots & \vdots & \cdots & \vdots \\ F_R(x_1) & F_R(x_2) & & F_R(x_G) \end{bmatrix}$$

is the block matrix of columns composed of the values of directivity characteristics (CD) of secondary spatial channels of the receiving DAA, when these channels are obtained after digital pattern-shaping with the aid of FFT, in the direction of phase centers of antenna elements x_m of the transmitting DAA; and

$$V = \begin{bmatrix} V_1(f_{11}) & \cdots & V_1(f_{1M}) & V_1(f_{21}) & \cdots & V_1(f_{2M}) & & V_1(f_{G1}) & \cdots & V_1(f_{GM}) \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots & \cdots & \vdots & \ddots & \vdots \\ V_D(f_{11}) & \cdots & V_D(f_{1M}) & V_D(f_{21}) & \cdots & V_D(f_{2M}) & & V_D(f_{G1}) & \cdots & V_D(f_{GM}) \end{bmatrix}$$

is the block matrix of AFR values of D FFT-filters for the gth direction at the mth frequency (f_{gm}).

After some simple rearrangements, the matrix P in (6) can be written in a more compact form

$$P = \begin{bmatrix} F_1(x_1)\tilde{V}_1 & & F_1(x_G)\tilde{V}_G \\ \vdots & \cdots & \vdots \\ F_{R_X}(x_1)\tilde{V}_1 & & F_{R_X}(x_G)\tilde{V}_G \end{bmatrix} \tag{7}$$

where $\widetilde{V}_1 = \begin{bmatrix} V_1(f_{11}) & \cdots & V_1(f_{1M}) \\ \vdots & \ddots & \vdots \\ V_D(f_{11}) & \cdots & V_D(f_{1M}) \end{bmatrix}$, $\widetilde{V}_G = \begin{bmatrix} V_1(f_{G1}) & \cdots & V_1(f_{GM}) \\ \cdots & \ddots & \vdots \\ V_D(f_{G1}) & \cdots & V_D(f_{GM}) \end{bmatrix}$.

Since for all G angular directions of signal reception we use different sets of orthogonal frequencies, expression (6) cannot be simplified by replacing the blocked modification of Kronecker's product by its traditional version, i.e., we may not use the notation $P = F \otimes V$, where the matrix V is identical to P present in (3).

Verification of applicability of the procedure for calculation of quadrature components of radiated signal amplitudes (5) with regard to matrix P (6)–(7) was performed with the use of Mathcad package. We checked the amplitudes of a three-element active DAA, where for each tested channel we imitated radiation of 8 signals orthogonal in frequency. Figure 2 displays an example of the results obtained during the simulation. Figure 2*a* illustrates distribution of three groups of signals along the frequency axis over outputs of 64 FFT-filters synthesized in every secondary spatial channel of the receiving DAA. Figure 2*b* depicts the spatial response of the three-element DAA, where the horizontal axis gives angular coordinates (in degrees) with respect to the array normal. Figure 2*c* gives a three-dimensional view of the resulting response of the receiving DAA in the "angle—frequency" coordinates. As the number of channels in the transmitting DAA grows, the pattern along the angular coordinate shows little difference from that shown in Fig. 2*b*, while in the frequency domain the totality of orthogonal signal groups increases.

Similar to (6) and (7), for a planar DAA composed of $R\times K$ antenna elements with factored directivity characteristics, which is well-suited for checking the characteristics of transmission channels of a planar active DAA, we may write

$$P = Z[\otimes]V$$

where the matrix Z can be expressed through the Khatri-Rao product in the form $Z = [F \blacksquare Q]$, "$\blacksquare$" denotes the Khatri-Rao matrix product [4]; while F and Q are, respectively, the CD matrices in terms of azimuth x and angle of site y, in which the number of columns equals the number G of the tested transmitting channels of the active DAA (i.e., signals' arrival directions):

$$F = \begin{bmatrix} F_1(x_1) & F_1(x_2) & & F_1(x_G) \\ \vdots & \vdots & \cdots & \vdots \\ F_R(x_1) & F_R(x_2) & & F_R(x_G) \end{bmatrix}, \quad Q = \begin{bmatrix} Q_1(y_1) & Q_1(y_2) & & Q_1(y_G) \\ \vdots & \vdots & \cdots & \vdots \\ Q_K(y_1) & Q_K(y_2) & & Q_K(y_G) \end{bmatrix}.$$

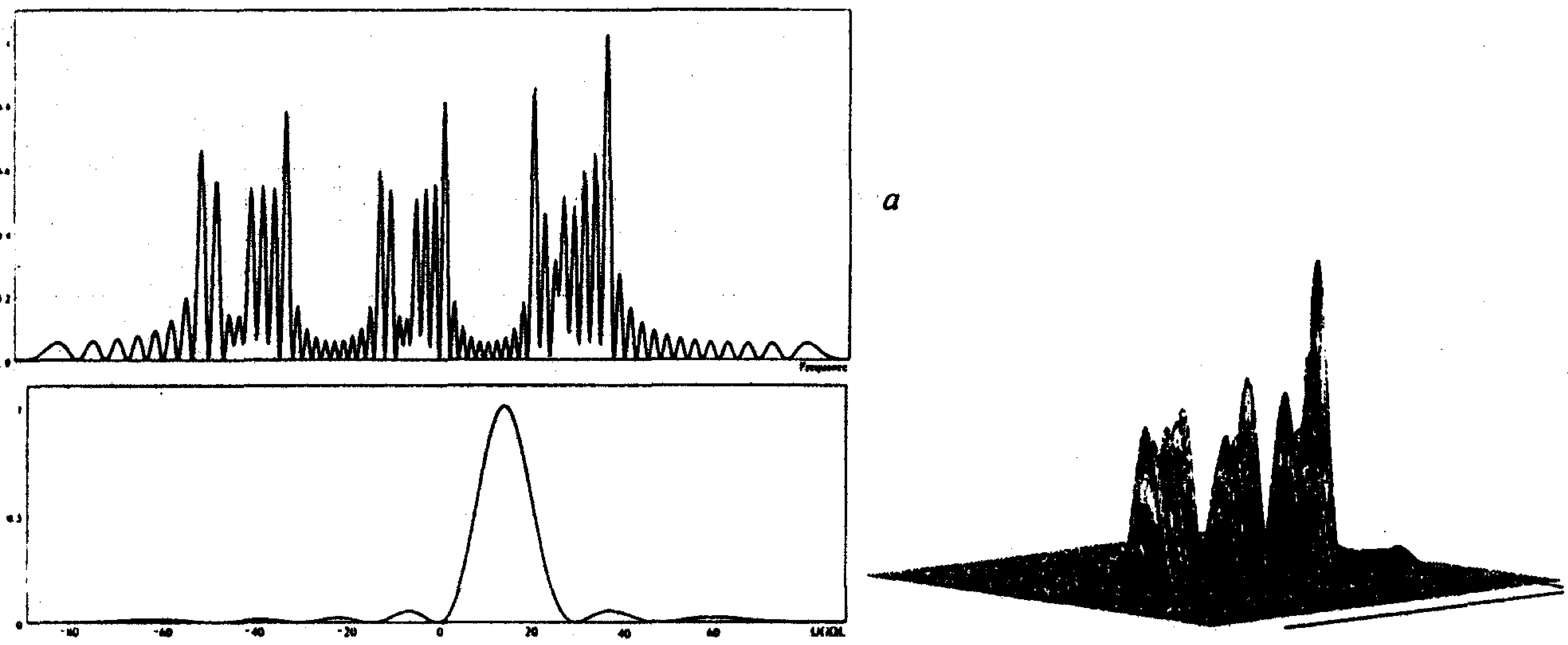
Here

$$Z = [F \blacksquare Q] = \begin{bmatrix} F_1(x_1)\begin{bmatrix} Q_1(y_1) \\ \vdots \\ Q_K(y_1) \end{bmatrix} & F_1(x_2)\begin{bmatrix} Q_1(y_2) \\ \vdots \\ Q_K(y_2) \end{bmatrix} & & F_1(x_G)\begin{bmatrix} Q_1(y_G) \\ \vdots \\ Q_K(y_G) \end{bmatrix} \\ \vdots & \vdots & \cdots & \vdots \\ F_R(x_1)\begin{bmatrix} Q_1(y_1) \\ \vdots \\ Q_K(y_1) \end{bmatrix} & F_R(x_2)\begin{bmatrix} Q_1(y_2) \\ \vdots \\ Q_K(y_2) \end{bmatrix} & & F_R(x_G)\begin{bmatrix} Q_1(y_G) \\ \vdots \\ Q_K(y_G) \end{bmatrix} \end{bmatrix}.$$

For correction of characteristics of the transmitting channels over the whole totality of operating frequencies, we developed a cyclic procedure of calculation of complex CC by successively changing the standard values in frequency groups of the channels to be tested. In future, in case of necessity, one can average quadrature components of CC over the whole totality M of operating frequencies:

$$\alpha_{cor}^{c(s)} = \frac{1}{M}\sum_{m=1}^{M} \alpha_{rq_m}^{c(s)}$$

where $\alpha_{rq_m}^{c(s)}$ are quadrature components of correction coefficients, calculated, in conformity with (2), by the mth packet of frequencies for the rqth transmission channel of the DAA under test.



a *b* *c*

Fig. 2

This approach may be called frequency-independent, since, in the end, the dependence of the correction coefficients on frequency will be brought to some level. The method is applicable to relatively narrow-band communication or radar systems. In the communication tasks based on OFDM-protocols, one may use an alternative variant consisting of generation of the CC matrix based on the $\alpha_{rq_m}^{c(s)}$ values stored for all channels and all working frequencies. Here the correction procedure reduces to element-wise multiplication of the CC matrix by the matrix of complex amplitudes of signals prior to applying the latter to the input of the inverse Fourier transformation and to DAC.

A peculiar feature of the suggested method of correction is the all-through control not only of the transmitting DAA, but of the whole communication line as well, including the reception channels of the testing array. As a result, we can compensate, at the transmitting end, the impact of AFR of the receiving path of the testing DAA by amplitude variation of those radiated signals whose frequencies correspond to AFR slopes of the reception channels.

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