CONTENTS

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	PAGES RUSSIAN	ENGLISH
Frequency-stabilized semiconductor sources of electromagnetic oscillation in the millimeter range of wavelengths. Part 1: diode sources. L. V. Kasatkin and V. P. Rukin	3	1
Design of analog networks by control theory methods. Part 1: theory. A. M. Zemlyak	18	11
Frequency-space apodization of surface acoustic wave transducers. Ye. A. Nelin	29	18
Analysis of linear networks in the Walsh transformation basis. A. I. Rybin	36	23
Measurement of partial directivity characteristics of reception channels in digital antenna arrays. V. I. Slyusar and A. F. Kozlov	41	27
Structural synthesis of multichannel adaptive rejection filters and analysis of their efficiency. P. A. Bakulev, V. A. Fyodorov, and N. D. Shestakov	48	32
Optimal processing of nonequidistant signals. D. I. Popov	54	37
Experimental investigation of random errors in measurement of characteristics of low-noise amplifiers in the millimeter range. I. K. Sunduchkov	61	42
Compensation of frequency fluctuations in a quartz standard arising from nonstationary temperature field inside a thermostat. Yu. I. Yevdokimenko and N. I. Svitenko	69	48
An acoustooptical corrector of time distortion of analog signals. A. R. Gasanov and Kh. I. Abdullayev	73	51
A multi-dimensional adaptive tracking system. V. I. Ishchenko and I. V. Zimchuk	76	53

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Vol. 47, No. 5, pp. 41–47, 2004
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MEASUREMENT OF PARTIAL DIRECTIVITY CHARACTERISTICS OF RECEPTION CHANNELS IN DIGITAL ANTENNA ARRAYS

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A new technique is suggested for measurement of directivity characteristics of reception channels in a digital antenna array when the channels are not identical.

Accurate measurement of echo signal parameters in digital antenna arrays (DAA) presumes the knowledge of partial directivity characteristics (DC) of reception channels. In order to estimate the DC as applied to phased antenna arrays (PAA), a rich variety of methods have been suggested. However, application of the same methods to DAA is hardly possible because of different circuit solutions in the above classes of antenna systems. This can be explained first of all by a peculiar analytical description of the responses of PAA reception channels. According to [1], such a response, being related to the time instant s, represents the total voltage

$$y_s = \sum_{n=0}^{N-1} C_n a_n + \varepsilon_s$$

where C_n are some known coefficients; a_n is the amplitude of excitation of the nth channel; and ε_s is the measurement error.

Summation of the channel voltages in PAA with shaping of only several beams makes impossible the measurement of directivity characteristics of the whole totality of N channels with the aid of a single time sampling. In DAA the response of every primary channel exists autonomously and can be expressed independent of other channel's voltages. Because of this, the technique used in [2] for estimating DC through solution of a system of equations set up by the voltages of the reception channels, can be applied in the case of DAA only after a radical adaptation.

As a rule, the known techniques of DC estimation are oriented to a "well stabilized" source of the test signal [2, 3], otherwise they are not optimal for DAA and may give considerable errors. So, of interest is development of procedures for measurement of directivity characteristics of the primary channels of the receiving DAA, which, despite poor stability of the test signal, could guarantee an estimation accuracy close to that potentially attainable.

Note that the algorithms suggested in this paper can be employed not only in methods dealing with an immobile antenna, but also with its rotation [2]. In our calculations this fact will not be considered in detail. As for the digital antenna array configuration, we assume that it represents a string of equidistantly arranged elements.

In real-valued description of directivity characteristics in such an antenna, the test signal voltage at the output of the kth reception channel at the sth time instant can be written in the form

$$\dot{U}_{k_s} = U^c_{k_s} + jU^s_{k_s} = \dot{a}_s F_k(x)\exp(jx_k) + \dot{n}_{k_s}, \tag{1}$$

where $\dot{a}_s = a^c_s + ja^s_s$ is the complex amplitude of the test signal at the sth time instant ($a^c_s = a \cdot \cos\varphi$ and $a^s_s = a \cdot \sin\varphi$ are the quadrature components, while φ is the signal's initial phase); $F_k(x)$ is the value of the directivity characteristic of the kth primary channel in the direction x; $x_k = x(k-1)$, $x = (2\pi/\lambda)d\sin\theta$ is the generalized angular coordinate of the test source with respect to the normal to DAA; λ is the test signal wavelength; d is the distance between elements of the equidistant array; θ is deviation of the direction toward the test signal source counted from the normal to the array; and $\dot{n}_{k_s}$ is the complex value of noise in the sth sample of ADC.

Using the method of least squares (MLS) for synthesis of the measurement procedure, let us minimize, in terms of unknowns a_s^c, a_s^s , and $F_k(x)$, the following functional:

$$M = \sum_{k=1}^{K} \left\{ U_{k_s}^c - F_k(x)[a_s^c \cos x_k - a_s^s \sin x_k] \right\}^2 + \sum_{k=1}^{K} \left\{ U_{k_s}^s - F_k(x)[a_s^c \sin x_k + a_s^s \cos x_k] \right\}^2 = \min \quad (2)$$

deduced from (1) based on trigonometric identities.

Here several approaches are possible. The essence of the first one, according to [4], consists in replacing (2) by its information equivalent:

$$\widetilde{M} = \hat{a}_s^c \sum_{k=1}^{K} F_k(x)[U_{k_s}^c \cos x_k + U_{k_s}^s \sin x_k] + \hat{a}_s^s \sum_{k=1}^{K} F_k(x)[U_{k_s}^s \cos x_k - U_{k_s}^c \sin x_k] = \max. \quad (3)$$

As a result, with the aid of estimates $\hat{a}_s^c$, $\hat{a}_s^s$, meeting the condition (2), we can easily modify version (3) into

$$\widetilde{M} = \left[\left\{ \sum_{k=1}^{K} F_k(x)[U_{k_s}^c \cos x_k + U_{k_s}^s \sin x_k] \right\}^2 + \left\{ \sum_{k=1}^{K} F_k(x)[U_{k_s}^s \cos x_k - U_{k_s}^c \sin x_k] \right\}^2 \right] \cdot \left[\sum_{k=1}^{K} F_k^2(x) \right]^{-1} = \max. \quad (4)$$

Thus, we may assert that the problem of measurement of the directivity characteristics has been reduced to maximization of expression (4), for example, through enumeration of possible values of $F_k(x)$. Since the range of $F_k(x)$ variations is naturally limited to the interval from 0 to 1, this search, in the event of a small number of primary channels, can be performed on an ordinary commercial computer, especially when the solution of this problem does not require operation in real time conditions.

For a large number of DAA receiving elements, another, i.e., non-iterative approach is of interest. It is based on estimating, instead of absolute values of directivity characteristics, their relative values normalized to the characteristic of the reference (bench mark) channel.

For the reference channel one may take an arbitrary rth receiver on DAA. For any direction x we may assume, without any prejudice, that the reference channel characteristic $F_r(x) = 1$. With regard for the above, the required estimates of $F_k(x)$ of the rest of the channels can be determined from (2) by solving the equation $\partial M / \partial F_k(x) = 0$.

Then we come to

$$F_k(x) = \frac{\hat{a}_s^c [U_{k_s}^c \cos x_k + U_{k_s}^s \sin x_k] + \hat{a}_s^s [U_{k_s}^s \cos x_k - U_{k_s}^c \sin x_k]}{\hat{a}_s^{c^2} + \hat{a}_s^{s^2}}$$

where for the values of the quadrature components of the amplitude $\hat{a}_s^c$ and $\hat{a}_s^s$ we can use the MLS single-sample estimates of the type

$$\begin{cases} \hat{a}_s^c = U_{r_s}^c \cos x_r + U_{r_s}^s \sin x_r, \\ \hat{a}_s^s = U_{r_s}^s \cos x_r - U_{r_s}^c \sin x_r, \end{cases} \quad (5)$$

obtained by the voltage of the DAA reference channel with its ordinal number r.

The problem becomes more complicated when we try to estimate the characteristics in the event of their complex-valued representation. In this case the response of the reception channel to the test signal can be expressed analytically in the form

$$\dot{U}_{k_s} = U^c_{k_s} + jU^s_{k_s} = \dot{a}_s \dot{F}_k(x)\exp(jx_k) + \dot{n}_{k_s} \tag{6}$$

where $\dot{F}_k(x) = F^c_k(x) + j\cdot F^s_k(x)$ while $F^c_k(x)$ and $F^s_k(x)$ are quadrature components of the complex directivity characteristic of the DAA kth element.

Because of inderdeterminacy of the equation system, which can be set up based on primary channel voltages (6), the single-sample procedure of measurement of the characteristics $F_k(x)$ can be carried out only by referring their values to the reference channel parameters. With all this in mind, let us estimate the quadrature components of the characteristics of the primary channels $F^c_k(x)$ and $F^s_k(x)$ with the aid of the method of least squares. Minimization of the relation

$$M = \sum_{k=1}^{K}\left\{U^c_{k_s} - [a^c_s F^c_k(x) - a^s_s F^s_k(x)]\cos x_k + [a^s_s F^c_k(x) - a^c_s F^s_k(x)]\sin x_k\right\}^2 +$$
$$+\sum_{k=1}^{K}\left\{U^s_{k_s} - [a^c_s F^c_k(x) - a^s_s F^s_k(x)]\sin x_k - [a^s_s F^c_k(x) - a^c_s F^s_k(x)]\cos x_k\right\}^2 = \min,$$

after some simple rearrangement, gives us

$$\begin{cases} \underset{k\neq r}{F^c_k(x)} = \dfrac{\hat{a}^c_s[U^c_{k_s}\cos x_k + U^s_{k_s}\sin x_k] + \hat{a}^s_s[U^s_{k_s}\cos x_k - U^c_{k_s}\sin x_k]}{\hat{a}^{c^2}_s + \hat{a}^{s^2}_s}, \\ \underset{k\neq r}{F^s_k(x)} = \dfrac{\hat{a}^c_s[U^s_{k_s}\cos x_k - U^c_{k_s}\sin x_k] - \hat{a}^s_s[U^c_{k_s}\cos x_k + U^s_{k_s}\sin x_k]}{\hat{a}^{c^2}_s + \hat{a}^{s^2}_s}, \end{cases}$$

where, as before, we use expressions (5) for the amplitude estimates.

An obvious disadvantage of all single-sample algorithms is vulnerability of the operation of estimation of the amplitude components of test signal (5) to the impact of noise outbursts in the absence of averaging over several realizations. So, in order to improve the characteristic measurement accuracy, it is desirable to employ a chain of ADC samples.

In practice this approach can be implemented as follows. In every primary channel of DAA, based on T consecutive time samples of complex voltages, we construct a system of frequency filters, for instance, by the FFT-operation associated with the method of OFDM-communication.

In the reference channel, with the aid of filter responses, for each direction x we can estimate by any known technique the signal frequency ω, and then — the amplitude of the test source signal. Frequency estimation may be omitted if for the test signal we use the pilot signal of the OFDM-burst coinciding with the maximum of the FFT filter AFR, and we can assure a high stability of the frequency.

The quadrature components of the signal amplitudes can be obtained, for example, within the MLS framework, provided that $F_r(x) = 1$:

$$\hat{a}^c = \frac{\sum_{t=1}^{T} f_t(\omega)\left\{U^c_{r_t}\cos x_r + U^s_{r_t}\sin x_r\right\}}{\sum_{t=1}^{T} f_t^2(\omega)},$$

$$\hat{a}^s = \frac{\sum_{t=1}^{T} f_t(\omega)\left\{U^s_{r_t}\cos x_r - U^c_{r_t}\sin x_r\right\}}{\sum_{t=1}^{T} f_t^2(\omega)},$$

where $f_r(\omega) = \sin T(\omega - \omega_r)/\sin(\omega - \omega_r)$, ω_r is the "resonant" frequency of the *r*th FFT-filter, and $U^{c(s)}_{r_t}$ are quadrature components of the response voltage of the *r*th frequency filter in the *r*th reception channel of DAA, which has been selected as reference. Otherwise,

$$\hat{a}^c = \frac{U^c_{r_t} \cos x_r + U^s_{r_t} \sin x_r}{T}; \quad \hat{a}^s = \frac{U^s_{r_t} \cos x_r - U^c_{r_t} \sin x_r}{T}, \tag{7}$$

if a signal from the orthogonal frequency array is used.

After that, based on known values of the amplitude components $\hat{a}^c_s$, $\hat{a}^s_s$, and of the frequency ω of the test signal, we calculate the relative magnitudes of directivity characteristics of $(k-1)$ channels, referred to the $F_r(x)$ value. With the use, for example, of MLS, for the real representation of the directivity characteristic, the necessary estimate takes the form

$$\underset{k \neq r}{F_k}(x) = \frac{\sum_{t=1}^{T} f_t(\omega)\left\{\hat{a}^c [U^c_{k_t} \cos x_k + U^s_{k_t} \sin x_k] + \hat{a}^s [U^s_{k_t} \cos x_k - U^c_{k_t} \sin x_k]\right\}}{\left(\hat{a}^{c^2} + \hat{a}^{s^2}\right) T^2}$$

or, as applied to the frequency which is resonant for the *r*th FFT-filter,

$$\underset{k \neq r}{F_k}(x) = \frac{\hat{a}^c [U^c_{k_t} \cos x_k + U^s_{k_t} \sin x_k] + \hat{a}^s [U^s_{k_t} \cos x_k - U^c_{k_t} \sin x_k]}{\left(\hat{a}^{c^2} + \hat{a}^{s^2}\right) T}.$$

It should be noted that the above methods of measurement can be extended to the case of planar DAA, although their implementation is much more complicated. Similar procedures can be synthesized based on responses of the secondary channels. However, from the viewpoint of computational expenditures they are much less attractive, so that the use of primary channels for this purpose is preferable.

Another issue is noteworthy. Multi-sample procedures of type (7), like the single-sample ones, incidentally, are efficient only in the case when, during a switch from one direction to other, the frequency of the test signal does not change or, within the range of its fluctuations, the AFR of all the reception channels do not alter their shape. Otherwise, estimation of partial DC will be accompanied by systematic errors — in proportion with AFR discrepancies.

One way to overcome this difficulty is to use a so-called frequency correction of DAA. It is based on a preliminary calculation of correction coefficients assigned to primary channels for every of possible frequencies or their groups. The correction is performed with a test source arranged on the normal to the DAA. Some of the correction procedures are described in [5] in detail. Their correlation with DC estimation is as follows.

In the frequency range of interest, for every primary channel, versus the reference one, we determine the correction coefficients by the pilot-signal from the DAA normal. The pilot-signal may represent a harmonic oscillation [5], radio pulses, or a δ-pulse without its carrier frequency. The use of the latter is preferable, since it permits estimation of AFR of all the channels in a wide range of frequencies during a single step.

In the DC measurement conditions, we determine the frequency of the sounding generator signal by the responses of the reference channel. After that we perform the weighting of output voltages of the rest of the reception elements, i.e., multiply them by respective correction coefficients. The voltages thus corrected are used in further computational operations necessary in measurements of directivity characteristics for angular directions, which differ from the normal to the array. In fact, measurement of DC of DAA partial channels makes it possible to estimate their mutual non-identity for various angles of signal reception.

In summation, it should be said that the consistency of the above algorithms has been confirmed experimentally with the aid of an operating breadboard model of DAA. Apart from the tasks of direction finding of radar sources, these methods are of particular interest as a tool for checking the quality of DAA reception channels of various purposes — either in the stage of their design, or under maintenance conditions.

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