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THE USE OF ULTRA WIDEBAND TECHNIQUE'S FOR MARINE RADIOLOCATION'S TASKS

ИСПОЛЬЗОВАНИЕ СВЕРХШИРОКОПОЛОСНОЙ ТЕХНИКИ ДЛЯ ЗАДАЧ МОРСКОЙ РАДИОЛОКАЦИИ

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ABSTRACT

The reception antenna diagram side lobe's level suppression algorithm for marine UWB-radar by means of antenna array with only a few tuning elements of antenna array is considered. The others no tuning elements of array are choosing for obtain a given value of average side lobe level suppression and with given value of antenna directivity without using of numerical optimization procedures. The special algorithm of interaction of these no tuning elements for realizing superdirectivity properties is used. The structural diagram of array is presented. The efficiency of suggested design has been investigated.

Keywords: ultra-wide band, antenna array, weight coefficients, side lobes level, partial diagram, superdirectivity.

РЕФЕРАТ

Метою представленого дослідження є формування алгоритму для пригнічення бічних пелюстків діаграми спрямованості антени судового надширокопasmового (Ultra-wide band – UWB) радіолокатора за допомогою антенних решіток з контрольованими елементами.

Методика дослідження полягає в розробці методів пригнічення бічних пелюстків діаграми спрямованості антени за допомогою використання антенних решіток з обмеженою кількістю налаштованих вагових коефіцієнтів просторового фільтру (елементів антенної решітки). Як відомо, через досить високий рівень бічних пелюстків діаграми спрямованості антени неможливе відокремлення відбитих сигналів від великих цілей (які мають велику ефективну площу розсіювання (ЕПР)) та від малих цілей, що мають малу ЕПР, що розташовані на одній дальності та мають близькі значення азимутальних кутів. UWB - радіолокатори здатні розпізнавати тип і форму цілі, оскільки прийнятий ехо-сигнал несе інформацію не тільки про об'єкт в цілому, але й про його елементи [1].

Результатом дослідження є розрахований алгоритм пригнічення бічних пелюстків діаграми спрямованості антени в зазначених точках за допомогою двох налаштованих вагових коефіцієнтів (першого та останнього W_1, W_N), а всі

інші вагові коефіцієнти W_i прийомної антенної решітки фіксовані (обираються за умови забезпечення необхідного середнього рівня бічних пелюстків діаграми спрямованості – $W_2; W_3; \dots; W_{N-1}$) [6]. На основі отриманих результатів розрахунків побудовано діаграми спрямованості, які відображають отримання пригнічення бічних пелюстків в зазначених точках. Також приведено структурну схему запропонованого алгоритму.

Практичне значення. Представлене дослідження можливо застосовувати у модернізації існуючих та побудові нових систем морської радіолокації. Враховуючи широке застосування над широкосмугових (Ultra-wide band – UWB) технологій для задач формування діаграм спрямованості антен, запропонований алгоритм також дає можливість використання отриманих розрахунків стосовно завдань морської радіолокації.

Ключові слова: над-широкосмуговий, антенна решітка, вагові коефіцієнти, пригнічення бічних пелюстків, парціальна діаграма, над-спрямованість.

Formulation of the problem in general terms and its connection with important scientific practical tasks

UWB radars are used not only ultrashort pulses, but also a very low level of radiated power. Higher range resolution and accuracy of measuring the distance to the target is provided by a UWB radar's wide band. High spatial resolution also provides the ability to isolate the target from interference reflections [1]. The modern UWB radars aren't free from a lot of drawbacks, including high enough side lobes level of radar antenna array diagram. It leads to impossibility of separation reflected signals from the big targets (which have big effective reflecting surface (ERS)) and from the small targets, which have small ERS, and which are situated on the same distance and have azimuth angle's with close value. The solution of this problem can be obtained by means radar antenna pattern formation on the base of antenna array with controlled elements.

The last achievements and publications analysis, in which the solution of the problem is begun and selection of the unsolved aspects of the problem

Optimal methods of radiation pattern linear array with controllable elements are known and involve the adjustment of all elements on the rather complicated algorithms and discussed in [2]. Our research makes it possible to simplify these methods and use simpler algorithm, worked out on the base of approaches, suggested in [15], [16].

The proper choice of antenna construction for corresponding broadband characteristics is considered in [3]. Also, obtaining good broadband characteristics using horn elements in antenna array considered in [3]. Corresponded broadband characteristics of antenna may be achieved by means proper choice of antenna construction [3]. Theoretically antennas which are constructed on the base of biconical elements may be provided pattern as well as impedance which are

practically independent of frequency inside of certain interval [3], [4]. It is known UWB antenna array based on planar dipole elements [5].

In [6], [7], [8] are presented the different algorithms for calculation the fixed weights coefficients for providing the required average antenna pattern side lobes level on the base of mean-square criteria, and more simple ones on the base of weighing functions $\sin(x)$ and $(\sin(x))^2$ [8], [11].

The increasing the width of main lobe of antenna diagram's issues are considered in [8,12]. Our algorithm allow significantly reduce the width of main lobe of antenna diagram's on the base of forming two partial diagrams and using the special interaction between them [9,10].

Therefore, the issues of using ultra wideband (UWB) techniques for marine radiolocation's tasks are relevant, and this determines the choice of this publication.

The paper purpose formulation

The purpose of this paper is to use special algorithm of interaction of non-tuning elements for realizing super directivity properties for UWB radars. The considered algorithm together with suppression antenna diagram side lobes in given points will improve the antenna array resolution and give possibility of separation reflected signals from the targets with different effective reflecting surfaces (ERS).

Presentation of basic research material substantiating scientific results

The reaching of antenna pattern side lobes required suppression level may be realized with the simple methods by using the arrays with limited number of tunable weight coefficients of spatial filter (antenna array elements), for example, when there are only two of such tunable weight coefficients. In this case all weights coefficients W_i of the receiving antenna array, except two (first and last: W_1, W_N), are fixed (selected under the condition of providing the required average antenna pattern side lobe's level) ($W_2, W_3, \dots, W_{N-1}$). Value of the two tunable weights coefficients are selected for providing zero values in two points (θ_1, θ_2) of the reception pattern [6,8].

The expressions, which are describing the reception pattern of linear array antenna $G(\theta)$ for the case considered were obtained in [6], and has the next form:

$$G(\theta) = \sum_{i=1}^N W_i \cdot e^{-j2\pi(d/\lambda)(N-1)\sin\theta} = G_{N-2}(\theta) - \gamma_1 \cdot G_{N-2}(\theta_1) - \gamma_2 \cdot G_{N-2}(\theta_2), \quad (1)$$

where $G_{N-2}(\theta)$ — partial diagram,

$$G_{N-2}(\theta) = \sum_{i=2}^{N-1} W_i \cdot e^{-j2\pi(d/\lambda)(N-1)\sin\theta} \quad (2)$$

$$W_1 = \frac{G_{N-2}(\theta_2) \cdot e^{j2\pi(d/\lambda)(N-1)\sin\theta_1} - G_{N-2}(\theta_1) \cdot e^{j2\pi(d/\lambda)(N-1)\sin\theta_2}}{e^{j2\pi(d/\lambda)(N-1)\sin\theta_2} - e^{j2\pi(d/\lambda)(N-1)\sin\theta_1}} \quad (3)$$

$$W_N = \frac{G_{N-2}(\theta_1) - G_{N-2}(\theta_2)}{e^{j2\pi(d/\lambda)(N-1)\sin\theta_2} - e^{j2\pi(d/\lambda)(N-1)\sin\theta_1}} \quad (4)$$

$$\gamma_1(\theta) = \frac{e^{-j2\pi(d/\lambda)(N-1)\sin\theta_2} - e^{-j2\pi(N-1)(d/\lambda)\sin\theta}}{e^{-j2\pi(d/\lambda)(N-1)\sin\theta_2} - e^{-j2\pi(N-1)(d/\lambda)\sin\theta_1}} \quad (5)$$

$$\gamma_2(\theta) = \frac{e^{-j2\pi(d/\lambda)(N-1)\sin\theta} - e^{-j2\pi(N-1)(d/\lambda)\sin\theta_1}}{e^{-j2\pi(d/\lambda)(N-1)\sin\theta_2} - e^{-j2\pi(N-1)(d/\lambda)\sin\theta_1}}, \quad (6)$$

$\phi = 2\pi d \sin \theta / \lambda$ — signal phase; λ — wave's length; d — distance between antenna's array elements; θ — angle between the normal to the axis of the array antenna and direction of coming signal.

($N-2$)-fixed weight coefficients may be selected under condition of additional suppression average level of reception diagram's side lobes (2) with possible widening the main lobe of antenna array diagram (1). So, the full number of the coefficients, which creates reception array diagram, is equal N . In this resrarch we use ten-elelements antenna array. This number of antenna array elements is used in standart marine radars. Calculation algorithm of the fixed weight coefficients optimized by mean-square criteria is considered in [7], [8].

Losses in antenna's directivity to fully common-mode reception diagram:

$$\rho = \frac{|G(0)|^2}{N \sum_{n=1}^N |W_n|^2} \quad (7)$$

Let's consider the form for different weight functions effect on reception diagram, which allow transforming the reception diagram properties. The expression for the weight coefficient in this case has the next form [6,11]:

$$W_n^{(1)} = \sin \left[\pi \left(\frac{y}{N+1-2} + \frac{n-2}{N+z-2} \right) \right] \quad (8)$$

$$W_n^{(2)} = \left[\sin \left[\pi \left(\frac{y}{N+1-2} + \frac{n-2}{N+z-2} \right) \right] \right]^2, \quad (9)$$

where: $n=2:N-1$; $y=1/[(N-2+1)/2]$; $z=[2y(N-2)-(N-2+1)]/[N-2+1-2y]$.

We can regulate the main lobe widening by parameter 'y'. The bigger value of parameter 'y', so the less main lobe widening. For example, the reception diagrams calculated with (8) and (9) are shown in Fig. 1.

Fig. 1. Partial reception diagram with 1) — $y=1, W_n^{(1)}$; 2) — $y=1, W_n^{(2)}$

The similar type of expressions, as (3), (4) were obtained for the signal time processing in the case of frequency selection [15] and the time-frequency task selection of signal [16].

So, we can correct the reception diagram by different weights functions. This approach doesn't require the implementation of numerical optimization procedures as were described in [6]. Choosing of such kind weighing functions we can get additional average side-lobe suppression, but with widening main lobe. The results of calculation diagram with using tuneable weights coefficients chosen according (3), (4) presented in Fig. 2. As we can see zero values in two given points are provided. The side-lobes value between suppressing points is rather small and satisfy to estimates were gotten in [17].

Fig. 2. Reception diagram side lobes suppression:

1) $-\theta_1 = -0.2576, \theta_2 = -0.2293, \rho = -1.4107 \text{ dB};$

2) $-\theta_1 = -0.2576, \theta_2 = -0.2293, \rho = -7.4690 \text{ dB}, y=1, W_n^{(2)}$

The method of increasing angle selectivity without losses in suppression of given points and in average side-lobe level had been suggested in [9] on the base of approach worked out in [18]. In considered case at the output of $N-2$ antenna array (no tunable part of reception antenna (1) (see (4))) we have complex signals $X_1, X_2, \dots, X_N$ [6]. By the first $N-2$ signals are created the sum:

$$Z_1 = \sum_{i=1}^{N-2} W_{i+1} X_i \quad (10)$$

Beside (8) the second sum is created:

$$Z_2 = \sum_{i=3}^N W_{i-1} X_i \quad (11)$$

Where $X_i = S_i + N_{1i}, S_i = |S_1| e^{j(i-1)\varphi}, \varphi = 2\pi d \sin \theta / \lambda, N_{1i}$ — thermal noise. If root-mean-square value of thermal noise is negligible small and $|S_1| = 1$ coincides with (2) and $Z_2 = Z_1 e^{j2\varphi}$.

Then the sum is created [18]:

$$G_R(\varphi^\wedge) = \left[1 \cdot e^{-j\psi^\wedge} \right] \cdot \begin{bmatrix} Z_1 \\ Z_2 \end{bmatrix}, \quad (12)$$

where $\psi^\wedge = -2 \arctg \cdot \mu \sin \varphi^\wedge / (1 - \mu \cos \varphi^\wedge)$ – resulting phase angle; $\varphi^\wedge$ – resulting signal's phase; $0 \leq \mu \leq 1$ – parameter, that controls the main peak's width.

$$\sin 2\varphi^\wedge = \text{Im}[(Z_1^* Z_2) / (|Z_1| |Z_2|)], \quad \cos 2\varphi^\wedge = \text{Re}[(Z_1^* Z_2) / (|Z_1| |Z_2|)].$$

From (11), using (12), after some transformations we can get:

$$G_R(\varphi) = [G_{N-2}(\varphi) \cdot 2 \cdot |\mu - \cos \varphi|] / \sqrt{1 - 2\mu \cos \varphi + \mu^2}. \quad (13)$$

Considering (11) and (12), (13) may be represented in the form:

$$G_R(\theta) = G_{N-2}(\theta) \cdot G_S(\theta) \quad (14)$$

$$G_S(\theta) = [2 \cdot |\mu - \cos 2\pi \sin \theta (d/\lambda)|] / \sqrt{1 - 2\mu \cos 2\pi \sin \theta (d/\lambda) + \mu^2}, \quad (15)$$

where: $G_{N-2}(\theta)$ is determined by (2).

As we can see from (11) angle selectivity of antenna may be essentially increased by means proper choose the value of μ . The width of the main beam of (11) on the level 0 and level $0.5G_S(0)$ are:

$$\Delta\theta_0 = 2 \arcsin[(\arccos \mu) / (2\pi d/\lambda)], \quad (16)$$

$$\Delta\theta_{0.5} = \arccos[\mu/2 + (\sqrt{2 - \mu^2})/2] / (\pi d/\lambda). \quad (17)$$

From (17) it follows, that if $\mu \rightarrow 1$, so $\Delta\theta_{0.5} \rightarrow 0$. So super selectivity [13,14] may be provided by means (15).

Consider some peculiarities of antenna array working with diagram (15). Due to functional transform (12) linearity of processing and Principle of superposition are breaking under affecting a few signals from different direction. If two interfering signals have close angles of arrival, may be provided good enough suppression of both signals. If the difference of arrival angles is big and intensity one is bigger enough than another, we get the suppression of the bigger signal. These considerations are stay in force for the case of more, then two signals. Thus for providing functionality of proposed principle of selection for multitarget situation special condition should be provided. Which suppose that signals with approximately equal intensity would have small difference of arrival angles, and for signals with essential different angles of arrival would be provided corresponding difference in their intensities. It may be realise by means antenna with diagram $G_R(\theta)$ (14). So, approximately equal intensities will be took place only in narrow angles interval, determined by main lobe beam width of diagram $G_{N-2}(\theta)$. For the signals which have essential difference of arrival directions, weighting of their intensities would be provided by the same diagram ($G_{N-2}(\theta)$).

Thermal noise and errors of practical realization are limited the maximal value of μ and thus limited the minimal value of main lobe beam width. If it is introduced η - noise/interference ratio (supposed equivalent noise, which included thermal noise

and technology errors), then we have $\mu \leq 1/(1+\eta)$. Using this value in (14) we can get restriction on main lobe beam width.

The partial reception diagrams, calculated by (11) with weight coefficients, calculated by (8) and (9), with equal and unequal correction are shown in Fig. 3.

The reception diagrams side lobes suppression in given points, calculated by (11), and for comparison reception diagram 10-element phased array with equal correction, calculated by (2) are shown in Fig. 4. One can see that not only side lobe suppression level was reserved, but and main lobe width is decreased in twice comparatively to simple phased array without amplitude correction.

Fig. 3. The reception diagram $W_i(G_{N-2}(\theta))$, $\mu = 0.95$:
1) – with equal correction; 2) – with unequal correction, $y=1$

Fig. 4. The reception diagram $W_i(G_{N-2}(\theta))$ side lobes suppression in points:

1 – $\theta_1 = -0.2576, \theta_2 = -0.2293, \rho = -7.4690 \text{ dB}$; $y=1, W_n^{(1)}, \mu = 0.95$;

2 – $\theta_1 = -1.2001, \theta_2 = -1.1718, \rho = -0.2238 \text{ dB}$; $y=1, W_n^{(1)}, \mu = 0.95$;

The block diagram of the proposed ten elements radar antenna array algorithm is shown on fig. 5.

Fig. 5. Block diagram of the proposed algorithm

Conclusions

In this paper UWB radar's antenna array design capable to obtain the given side-lobe suppression with controlled value of directivity coefficient and with possible realization of superdirectivity property is suggested. The approach is simple enough for calculations and does not require the implementation of numerical optimization procedures. It's very useful for practical implementation, when it's necessary to get the given side lobes suppression with given main lobe properties. This approach can be applied in the modernization existing marine radar and construction of new marine radar systems on the base UWB techniques. The efficiency of proposed approach is investigated.

REFERENCES

1. N. Scherbak, Ultra-wide band radiolocation. What is this? // Electronics, technology, business. Issue N3/2002.
2. Y. Shirman, V. Mandjos, 1981, Theory and technics of radar information processing under interferences. M. Radio I Svyaz, – P. 416.
3. D. Valderas, J.I. Suncho, D. Puente, C. Ling, X. Chen. Ultra-wideband Antennas (Design and Application) Imperial College Press.2011. P.194.
4. K. Sivaaak, D. McKeown, Ultra-Wide-band Radio Technology, John Wilay&Sons,2004, P.246.

5. B. LiJ. Zhang, Y. Deng, Z. Zhou. Design of a Low-profile Ultra-wideband Antenna Array Based on Planar Dipole Elements, IEEE Radar 2018, pp.0125-0128.
6. Koshevyy V., Shershnova A., 2013, The formation of zero levels of Radiation Pattern linear Antennas Array with minimum quantity of controlling elements, *Proc. 9 Int. Conf. on Antenna Theory and Techniques (ICATT-13)*, Odessa, Ukraine: pp. 264-265.
7. V. Koshevyy, V. Lavrinenko, S. Chuprov, 1975, The efficiency of quasi-filter analysis. «RIPORT », VIMI. №2, – p. 7.
8. V. Koshevyy, A. Shevchenko, 2016, The research of non-tuneable part of antenna array amplitude distribution for side lobes suppression efficiency. 2016 International Conference Radio Electronics & Info Communications (UkrMiCo'2016), National Technical University of Ukraine "Kyiv Polytechnic Institute", Kyiv, Ukraine, pp. 156 – 160.
9. V. Koshevyy, A. Shevchenko, 2017, Radiation Pattern of Linear Antenna Array With Control of Directivity and Super-Selectivity Properties//XI International Conference on Antenna Theory and Techniques (ICATT), Kyiv, Ukraine, pp. 165 – 168, 2017.
10. V. Koshevyy, A. Shevchenko, 2018, Antenna array with superdirectivity properties. International Conference on Electrical, Electronics, Computer, Communication, Mechanical and Computing (EECCMC) – 2018, Priyadarshini Engineering College, Vellore District, Tamil Nadu, India.
11. Koshevyy V., Pashenko O., Improved compound multiphase waveforms with additional amplitude modulation (periodic mode) for marine radars. *The second International Conference on Information and Telecommunication Technologies and Radio Electronics (UkrMiCo'2017)*, 11-15 September 2017, Odessa, Ukraine: pp. 1-6.
12. Huang Yi, Boyle K., Antennas, From Theory to Practice. Wiley, 2008: P. 363.
13. Hansen R.C., 1981, Some New Calculations on Antenna Superdirectivity, *Proceeding of the IEEE*, Vol. 69, No. 10: pp. 1365-1366. (Oct. 1981).
14. Newman E. H., Richmond J.H., Walter C. H., Array size reduction using superdirectivity, *AP – S Session 108*, 1550: pp. 331-334 (June 21, 1976).
15. Koshevyy V., Lavrinenko V., 1981, Target selection on the base of discrete structure with minimum number of tuned elements. "Izvestia VUZ Radioelectronika", vol. 24, №4: pp. 105 – 107.
16. V. Koshevyy, 1982, Moving target systems indication synthesis with the inverse matrix size restrictions. - «Izvestia VUZ. Radioelectronika», т.25, № 3, pp. 84-86.

17. V. Koshevyy, 1981, Some Limited Relations for Cross-Ambiguity Function of Finite Signals, Radio Engineering and Electronics, Vol. 26. Pp. 2588-2599. Dec. 1981.
18. V. Koshevyy, 1983, Optimum Properties of One Stage (Single) MTI Canceller, Radio Engineering, N7, pp. 64-66, Jul.1983.