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WAVEFORM DESIGN FOR MASKING EFFECT REDUCTION IN NOISE RADAR USING VITERBI ALGORITHM

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Signal processing in noise radar is based on the calculation of the correlation between transmitted and received signals. Strong echoes of nearby targets present relatively high sidelobes in the correlation function, thus they can conceal weak echoes of far targets (masking effect). Therefore, finding methods to suppress this effect becomes important. These methods can be implemented at the transmitter or receiver side. There are many new methods developed for the receiver side and a few methods applicable at the transmitter side. The main idea of suppressing side lobes at the transmitter is to design a specific noise signal instead of using pure random noise. A new waveform design to reduce the masking effect is introduced in this paper, which is based on Viterbi algorithm and the produced signals are quantized.

Keywords: masking effect, noise radar, sidelobe, Viterbi algorithm.

I. INTRODUCTION

Pulse compression is an effective method for achieving medium- and high-range resolution in long-range radars.

For a long time, the desired range resolution has been obtained by means of linear frequency modulation (chirp) and matched filtering, in both pulsed and continuous wave (FMCW) radars. Recently, high resolutions are frequently achieved in ultra-wide band radars, often using noise or pseudo-noise signals for target illumination [1–3]. The noise radar technology is a radar technology which uses the noise continuous waveform as a probe signal and correlation (or double spectral) processing of the radar returns as its optimal reception (matched filtering) [4]. Having been considered as early as the 1950s, the concept of noise radar is not new [5]. Noise waveforms enable independent range and velocity resolution, which is regarded as a very significant attribute in the design of surveillance radar systems having moving target indicator (MTI) capabilities. In addition, noise radar guarantees high resolution, low cost, robustness to countermeasures and good electromagnetic compatibility.

Moreover, noise radar waveforms are of interest in LPI (low probability of intercept) radar and repeater jamming. These radars have been used for the measurement of range profiles [6], Doppler estimation [7], detection of buried objects [8], interferometry [9] and inverse synthetic aperture radar (ISAR) and synthetic aperture radar (SAR) imaging [10–12]. A more complete list of the literature on noise radars is provided in [13].

One of the major troubles that should be considered in radar design is weak target echo that may be masked by strong ones. This problem can be easily solved by utilizing range gain control in pulse radars, because of the time separation of near and far echoes. Also it is reduced in FMCW radars by using analogue filters benefiting from the frequency separation of near and far echoes. In contrast with pulse and FMCW radars, there is no frequency or time separation in continuous wave noise radars. Consequently, the sidelobes generated by range compression blocks may reduce radar sensitivity and detection range [14], and in the case of two close targets with significantly different RCS, it may also lead to masking the weak one. Moreover, crosstalk signal and clutter can be considered as important origins of masking effect. Several methods have been developed to counter the masking effect [15]. The crosstalk signal and the ground clutter can be adaptively removed from the received signal with an adaptive lattice filter [16, 17]. The non-zero Doppler clutter can also be removed by using a variation of the previous method [18, 19]. Since this method is not applicable to high-speed targets, stretch processing has been proposed by Misiurewicz and Kulpa [14] and Kulpa and Misiurewicz [20] to overcome the masking effect in noise radars. Another method of suppressing range sidelobe level (RSL), applied to random binary phase coded waveforms, has been introduced by Hong et al. [21]. An apodization filtering technique, developed in [22], achieves sidelobe suppression of greater than 20 dB. Nelander [23] has presented a sidelobe suppression algorithm based on inverse filtering. Sidelobe suppression can also be achieved using an iterative algorithm known as CLEAN [24]. Although all mentioned methods are based on signal processing at the receiver part, there are very few noticeable works focusing on the waveform design at the transmitter part of noise radars. In [25], it is shown that transmission of a sine wave, which is phase or frequency, modulated by random noise waveform leads to improved sidelobe suppression in comparison with transmission of a pure noise waveform. In [26], a method of waveform design with the goal of masking effect suppression has been developed. The proposed waveforms use many short codes to produce a code with the length of the product of the shorter codes lengths. The resulted long code can be arbitrarily long by introducing new shorter codes iteratively. A method for designing chaotic waveforms with parameter optimization for the purpose of complex target detection has been suggested by Carroll; however, it does not concern the masking effect [27]. As the review of noise radar literature shows, most of the algorithms that have been developed in order to decrease the masking effect are applicable in the receiver end. Hence, it should first be clarified that whether masking effect reduction in noise radar is possible by concentrating on waveform design [28].

In this paper, a new waveform design based on Viterbi algorithm to reduce masking effect in random phase modulated radarsis introduced. Uniqueness of the developed algorithm comes from the quantized nature of designed signal which is another step to applicability. In previous waveform design methods which is presented in [29], the output signals are analog phases in range $[-\pi, \pi]$; however in practical radars, phases should be quantized with limited number of bits.

In the following sections, first of Viterbi algorithm is reviewed. second, all necessary parameters are defined. Next, waveform design using Viterbi Algorithm is introduced. Next, simulation results are discussed and eventually paper will be concluded.

II. VITERBI ALGORITHM

The Viterbi algorithm was proposed by Andrew Viterbi in 1967 as a decoding algorithm for convolutional codes over noisy digital communication links. The algorithm has found universal application in decoding the convolutional codes used in both CDMA and GSM digital cellular and etc.It is now also commonly used in speech recognition, keyword spotting, computational linguistics, and bioinformatics [30].

Since that time, it has been recognized as an attractive solution to a variety of digital estimation problems [30].

In its most general form, the VA may be viewed as a solution to the problem of maximum a posteriori probability (MAP) estimation of the state sequence of a finite-state discrete-time Markov process observed in memoryless noise [30].

In the presence of intersymbol interference in communication channels, using the whitening filter in system results [31]:

$$v_k = \sum_{n=0}^{L} f_n I_{k-n} + \eta_k \ . \tag{1}$$

Where $\{\eta_k\}$ is a white Gaussian noise sequence, $\{f_k\}$ is a set of tap coefficients of an equivalent discrete-time transversal filter with length of L and $\{I_k\}$ s the information sequence.

MLSE of the information sequence $\{I_k\}$ is most easily described in terms of the received sequence $\{v_n\}$ at the output of the whitening filter. In the presence of intersymbol interference that spans $L+1$ symbols (L interfering components), the MLSE criterion is equivalent to the problem of estimating the state of a discrete-time finite-state machine [31]. The finite-state machine in this case is the equivalent discrete-time channel with coefficients $\{f_k\}$, and its state at any instant in time is given by the L most recent inputs, i.e., the state at time k is

$$S_k = (I_{k-1}, I_{k-2} .... I_{k-2}) \,. \tag{2}$$

Where $I_k = 0$ for $k \le 0$. Hence, if the information symbols are M-ary, the channel filter has M^L states. Consequently, the channel is described by an M^L -state trellis and the Viterbi algorithm may be used to determine the most probable path through the trellis.

The metrics used in the trellis search are akin to the metrics used in soft-decision decoding of convolutional codes. In brief, we begin with the samples $v_1, v_2, ..., v_{L+1}$, from which we compute the M^{L+1} metrics

$$\sum_{k=1}^{L+1} In\rho(v_k \,|\, I_k, I_{k-1}, ... I_{k-L}) \tag{3}$$

The M^{L+1} possible sequences of $I_{L+1}, I_L, ... I_2, I_1$ are subdivided into M^L groups corresponding to the M^L states $(I_{L+1}, I_L, ..., I_2)$ Note that the M sequences in each group (state) differ in I_1 and correspond to the paths through the trellis that merge at a single node. Form the M sequences in each of the M^L states, we select the sequence with the largest probability (with respect to I_1) and assign to the surviving sequence the metric

$$\begin{aligned} PM_1(I_{L+k}) &= PM_1(I_{L+1}, I_L, ..., I_2) \\ &= \max \sum_{k=1}^{L+1} ln\rho(v_k \,|\, I_k, I_{k-1}, ..., I_{k-L}) \end{aligned} \tag{4}$$

The $M-1$ remaining sequences from each of the M^L groups are discarded. Thus, we are left with M^L surviving sequences and their metrics.

Upon reception of v_{L+2}, the M^L surviving sequences are extended by one stage, and the corresponding M^{L+1} probabilities for the extended sequences are computed using the previous metrics and the new increment, which is $In\rho(v_{L+2} \,|\, I_{L+2}, I_{L+1}, ... I_2)$. Again, the M^{L+1} sequences are subdivided into M^L groups corresponding to the M^L possible states $(I_{L+2}, I_{L+1}, ... I_3)$ and the most probable sequence from each group is selected, while the other $M-1$ sequences are discarded.

The procedure continues with the reception of subsequent signal samples. In general, upon reception of v_{L+k}, the metrics

$$\begin{aligned} PM_k(I_{L+k}) = \max[&Inp\left(v_{L+k} \,|\, I_{L+k}, ..., I_k\right) + \\ &+ PM_{k-1}\left(I_{L+k-1}\right)] \end{aligned} \tag{5}$$

That are computed give the probabilities of the M^L surviving sequences. Thus, as each signal sample is received, the Viterbi algorithm involves first the computation of the M^{L+1} probabilities

$$Inp(v_{L+k} \,|\, I_{L+k}, ... I_k) + PM_{k-1}(I_{L+k-1})$$

corresponding of the M^{L+1} sequences that form the continuations of the M^L surviving sequences from the previous stage of the process. Then the M^{L+1} sequences are subdivided into M^L groups, with each group containing sequences that terminate in the same set of symbols $I_{L+k}, ... I_{k+1}$ and differ in the symbol I_k . From each group of M sequences, we select the one having the largest probability as indicated by (5), while the remaining M-1 sequences are discarded. Thus, we are left again with M^L sequences having the metrics $M_k(I_{L+k})$ [31].

III. DEFINITION OF ISLR

The transmitted and received signals are represented by x_k and y_k respectively. Assuming a maximum delay of N samples for farthest target, signals can be divided into N-sample blocks [32].

To extract information of targets at the receiver, a time correlation between these signals is calculated, which is between successive blocks. The time correlation between block m and $m+1$ is represented:

$$P(i,m)=\sum\nolimits_{k=mN}^{(m+1)N-1} x_k y_{k+i}^{*} \ . \qquad (6)$$

Where $i=0.1,...,N, m=0,1,...,M-1$, symbol * represents complex conjugate and M is the total number of blocks. Ignoring Doppler effect, received signal is a delayed version of transmitted one

$$y_k = x_{k-l} \ . \qquad (7)$$

Suppose phase modulating signal:

$$x_k = e^{j\theta_k} \ . \qquad (8)$$

Using (5), (6) and (7) results:

$$P(i,m)=\sum\nolimits_{k=mN}^{(m+1)N-1} \theta^{j(\theta_k-\theta_{k+i-l})} \ . \qquad (9)$$

Range main lobe take place, when i and l are equal and Range side lobe in other case. Sidelobes are divided into two groups based upon sign of $i-l$, show them with C_p and D_p respectively [32]:

$$C_p=\sum\nolimits_{k=mN}^{(m+1)N-1} e^{j\theta_k - j\theta_{k+p}}, p=1.2....,N$$

$$D_p=\sum\nolimits_{k=(m+1)N}^{(m+2)N-1} e^{j\theta_k - j\theta_{k-p}}, p=1.2....,N \qquad (10)$$

In addition main lobe level is made up with C_0 and D_0 which are equal to N Fig. (1) represents a simple example of Masking weak target main lobe by strong target sidelobe [32].

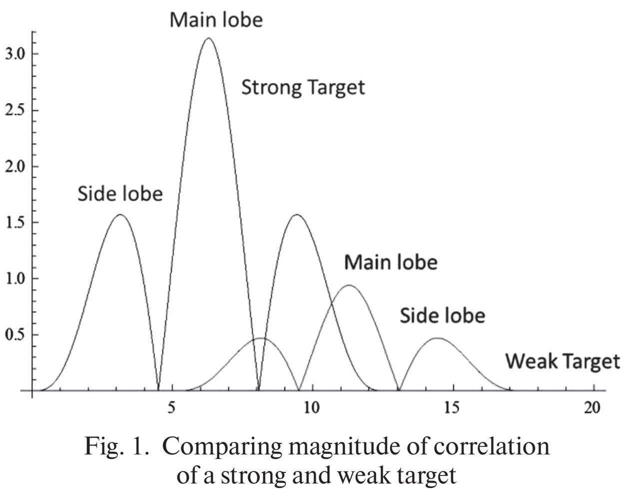

Fig. 1. Comparing magnitude of correlation of a strong and weak target

The Integrated side lobe ratio (ISLR) is defined as the total energy of sidelobes to total energy of main lobe, will equal

$$ISLR=\frac{2NJ}{|D_0|^2+|C_0|^2} \ . \qquad (11)$$

And J is defined as below:

$$J=\frac{1}{2N}\sum\nolimits_{p=1}^{N}(|C_p|^2+|D_p|^2) \ . \qquad (12)$$

ISLR is an appropriate measurement of masking effect, from now on, the purpose is to minimize ISLR. Let's have a more detailed look on ISLR. As mentioned before, $|D_0|$ and $|C_0|$ are constant and equal to N, thus minimization of ISLR and J are equivalent. Expanding eq. (12):

$$\begin{aligned} J=\sum_{k=1}^{N}\sum_{i=1}^{N-1}\sum_{j=i+1}^{N}\{&\cos(\theta_i+\theta_{j+k}-\theta_{i+k}-\theta_j)+ \\ &+\cos(\theta_{j+N}+\theta_{i+N-k}-\theta_{i+N}-\theta_{j+N-k})\} \end{aligned} \qquad (13)$$

Procedure of finding the appropriate phases for waveform design will be block by block, it means set first N-signal block randomly. Find the second N-signal block to minimize the ISLR due to these two blocks, and continue these progress up to end.

To find the transition coefficients among Markov process states, we rewrite J in a new form. That is

$$J=J_{cte}+J_{N+1}+J_{N+2}+...+J_{2N} \ . \qquad (14)$$

Where, J_{cte} is the constant part of J due to the first N signal of first block which have been set and is independent of the second block that is in optimization process $J_m, m=N+1,...,2N$, is part of J which is affected by signals 1 to m.

$$\begin{aligned} J_m=&\sum_{k=m-N}^{N}\sum_{i=1}^{m-k-1}\cos(\theta_m+\theta_i-\theta_{i+k}-\theta_{m-k})+ \\ &\sum_{k=1}^{N}\sum_{i=1}^{m-N-1}\cos(\theta_m+\theta_{i+N-k}-\theta_{i+N}-\theta_{m-k}) \end{aligned} \qquad (15)$$

Implementation of VA to minimize ISLR is discussed in subsequent section.

IV. ALGORITHM

In section II, a general discussion of VA was presented and important parameters were described. Let study this subject in detail.

The proposed waveform design method is based on VA, thus a Markov process, states, transition weights, decision criteria and all other parameters should be defined and a one-to-one correspondence between these parameters and the in hand problem should be made.

Suppose, signals of first block are chosen completely random. Now we are going to find the i^{th} block signals. More precisely, phase of i^{th} block signals. Clearly, due to quantized nature of produced signals, each phase can have 2^{nbits} different values, where $nbits$ is the number of quantization bits. As stated in previous section, ISLR is our criterion in designing the signals. In addition, equivalence of ISLR and parameter J was proved. Thus, the cost functions are chosen based on parameter J. For designing each new block signal, a new Markov process is synthesized, which has N steps (equal to number of signals in a block). It means, each step is matched with equivalent signal. In transition between step k to step $k+1, (k+1)^{tn}$, signal is under investigation.

Ideally, number of states in each step, is number of all different cases of previous signals in the block. However, simple calculations show that is impossible in practical cases due to large consumption of memory and calculations. Thus, parameter L is defined, which is the memory (or buffer) size of algorithm. Parameter L means, in process of calculating different choices of θ_k, different cases of the last L signals are kept, however the other previous signals should have been determined up to that moment. i.e. θ_{k-L-1} should have been determined and costs for θ_k are calculated based on 2^{nbits} different values of θ_k and many different cases for θ_{k-L} to θ_{k-1}. Remember, these different cases are survived path of VA from previous step. Now, θ_{k-L} should be determined based on the θ_{k-L} of the minimum cost survived path of step *k*, and algorithm continues. Remember in step N, all the θ_{N-L} to θ_N should be determined based on the θ_{N-L} to θ_N of the minimum survived path of N^{tn} step.

A General description of proposed algorithm has been described so far. Now it's time to look in more detail to the algorithm. As discussed in section II, survived path in VA should be chosen based on a criterion. In section II it was probability, however as mentioned before, parameter *J* is our cost function and criterion for determining survived path. Due to parameters L and $nbits$, there are $2^{nbits\times L}$ states in each step. Each step is determined with a unique survived path. In addition, each state has a cost, which is sum of corresponding survived path, vectors cost. Let Illustrate the above parameters through an example. Again, suppose we are in transition of k^{tn} step to $(k+1)^{tn}$ step. Thus survived paths and costs of all the states in k^{tn} are determined and we should calculate these parameters for stats of $k+1$ step. There are 2^{nbits} different possibilities for θ_{k+1}. Vectors are the links, connecting states of k^{tn} step to states of $(k+1)^{tn}$ step. There are 2^{nbits} outgoing vectors from each state of k^{tn} step, and 2^{nbits} incoming vectors to each state of step $k+1$. Each state in k^{tn} step is one of the $2^{nbits\times L}$ different cases of θ_{k-L} to θ_{k-1} and each state in $(k+1)^{tn}$ step is one of the $2^{nbits\times L}$ different cases of θ_{k-L+1} to θ_k. Vectors cost of this transition is calculated by Eq. (15). Note that $m=k+N+1$ and θ_m is the corresponding value of θ_{k+1}. Now, each state of step $k+1$ has 2^{nbits} incoming vectors with different costs. If vectors cost are added to states cost of their sources, 2^{nbits} different total cost for each state is found. Clearly, the survived path is the path with minimum cost and the state cost is the minimum of the costs.

Note that the process depends on the parameters: N, L and $nbits$.

V. SIMULATION RESULTS

In this section, simulation results of developed method are presented. Variation of ISLR versus L of VA method for $N=8$ and different values of nbits is plotted in Fig. (2). As expected, increasing nbits leads to better ISLR (reduction in ISLR). In addition, note that ISLR decreases while L increases, which was predictable. This observation comes from the fact that, increasing L, is translated to increasing memory and including more cases that clearly result a better performance and of course imply more computational complexity.

Fig. (3) is similar to Fig. (2), however $N=16$. Those observations are confirmed in this plot again. Another interesting point is a negligible difference between the results of Fig. (2) and Fig. (3), which show a low relationship with N, number of signals in a block.

It should be emphasized that the ISLR improvement shown in following figures are in relative to pure noise sidelobe level. It means, to obtain sidelobe level of designed signal, these numbers should be added to pure noise sidelobe level.

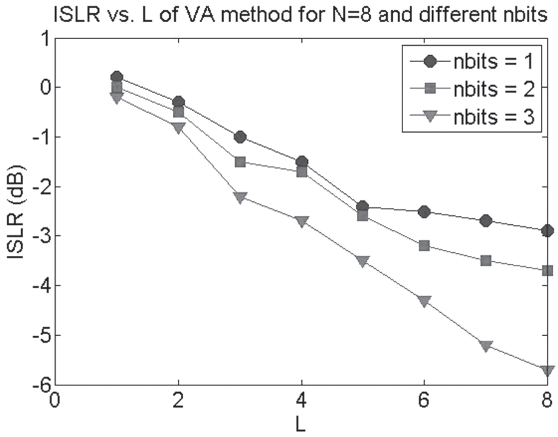

Fig. 2. ISLR vs. L of VA method for $N=8$ and different values of nbits.

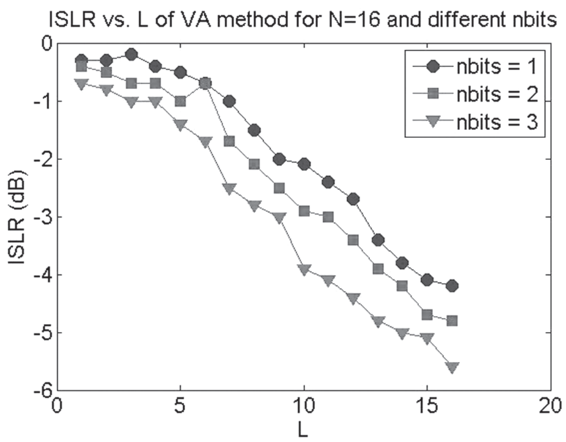

Fig. 3. ISLR vs. L of VA method for N=16 and different values of nbits

Few methods of waveform design for masking effect reduction have proposed till now. As mentioned before, all these methods make analog outputs. To have criterion of developed method performance, one of the best previous methods presented in [29] is compared with VA method, which is named Conventional method in tables in the following.

To have a fair comparison, output phases of Conventional method are quantized with similar number of bits for VA method and the results are represented in Table (1) for case $N=8$ and in Table (2) for $N=16$

Table 1

Performance comparison (ISLR) of VA and Conventional methods for $N = 8$

nbits	Conventional method	VA method
1	-1.1 dB	-2.9 dB
2	-2.7 dB	-3.7 dB
3	-4.5 dB	-5.7 dB

Table 2

Performance comparison (ISLR) of VA and Conventional methods for $N = 16$

nbits	Conventional method	VA method
1	-1.3 dB	-4.2 dB
2	-2.8 dB	-4.8 dB
3	-4.5 dB	-5.6 dB

ISLR of Conventional method is represented in second column of both tables and the third column is ISLR of VA method. First, second and third row of each table is for the cases of number of bits equal to one, two and three respectively. Better performance of developed method is visible for all the conditions and the results are similar for both cases $N = 8,16$.

VI. CONCLUSION

During the correlation process between transmitted and received signals, relatively high sidelobes of strong echoes of nearby targets can conceal weak echoes of far targets (masking effect). There has been a wide study on this subject which led to different methods. These methods can be implemented at the transmitter or receiver side.

There are many new methods developed for the receiver side and a few methods applicable at the transmitter side. The main idea of suppressing side lobes at the transmitter is to design a specific noise signal instead of using purely random noise.

In this paper a new waveform design to reduce the masking effect was introduced which is based on Viterbi Algorithm. The important difference of developed algorithm with other waveform design methods is the quantized nature of the produced signal that makes it more applicable. In addition, simulation results showed the higher performance of developed method in comparison with previous ones.

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Получение сигнала для снижения маскирующего эффекта в шумовом радиолокаторе с использованием алгоритма Витерби / Е. Тохиди, М.Н. Мадж, Х.Х. Джариани, М. Наяби // Прикладная радиоэлектроника: науч.-техн. журнал. – 2013. – Том 12. – № 1. – С. 11–16.

Обработка сигналов в шумовой радиолокации основана на вычислении корреляции между передаваемым и принимаемым сигналами. Сильное эхо от близлежащих целей является причиной относительно высокого уровня боковых лепестков корреляционной функции, таким образом, они могут скрыть слабые отклики от далеко расположенных целей (эффект маскировки). Поэтому становится важным поиск методов для подавления этого эффекта. Эти методы могут быть реализованы на передающей или на приемной стороне. Существует много новых методов, разработанных для канала приемника, и несколько методов, применяемых в канале передатчика. Основная идея подавления боковых лепестков в передатчике заключается в формировании шумового сигнала со специальными свойствами вместо истинно случайного шума. Новый способ формирования сигнала с целью уменьшения эффекта маскировки, который предложен в статье, основан на алгоритме Витерби с квантованием получаемых сигналов.

Ключевые слова: маскирующий эффект, шумовой радиолокатор, боковые лепестки, алгоритм Витерби.

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Отримання сигналу для зниження маскуючого ефекту в шумовому радіолокаторі з використанням алгоритму Вітербі / Е.Тохіді, М.Н. Мадж, Х.Х.Джаріані, М.Наябі // Прикладна радіоелектроніка: наук.-техн. журнал. – 2013. – Том 12. – № 1. – С. 11–16.

Обробка сигналів в шумовій радіолокації заснована на обчисленні кореляції між переданим і прийнятим сигналами. Сильне відлуння від довколишніх цілей є причиною відносно високого рівня бічних пелюсток кореляційної функції, таким чином, вони можуть приховати слабкі відгуки від далеко розташованих цілей (ефект маскування). Тому стає важливим пошук методів для заглушення цього ефекту. Ці методи можуть бути реалізовані на передавальній або на приймальній стороні. Є багато методів, розроблених для каналу приймача, і кілька методів, які застосовуються в каналі передавача. Основна ідея заглушення бічних пелюсток в передавачі полягає у формуванні шумового сигналу зі спеціальними властивостями замість істинно випадкового шуму. Новий спосіб формування сигналу з метою зменшення ефекту маскування, який запропонований в статті, базується на алгоритмі Вітербі з квантуванням одержуваних сигналів.

Ключові слова: ефект маскування, шумовий радіолокатор, бічні пелюстки, алгоритм Вітербі.

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