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Algorythm of out-of-band radiation reduction

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Abstract

The paper represents regular method of synthesis of a bit signal for the wireless communication networks using frequency manipulation for coding of the transferred information. The synthesized signal has two-level envelope and low sidelobe level in the set area of a spectrum. The synthesis method is based on repeated operations of a delay and addition of initial sequence of pulses with rectangular envelope. As a result of synthesis the signal with pulse-width modulation is obtained. Software implementation of the algorithm was created using development environment NI LabVIEW. Temporal and spectral properties of the signal are researched. The signal can also find application in Doppler's ultrasonic systems that use high-powered nonlinear amplifiers.

Keywords - regular synthesis of a bit signal without window processing; low sidelobe level of spectrum; two-level envelope

Introduction

Nowadays most types of wireless networks are based on fixed allocation of frequency bands and hierarchical infrastructure. In developing new methods of communication, it is necessary to take into account the occupation of frequency domain. Transition to less busy domain of higher frequencies means either a reduction of covered area or an increase of the power used. Therefore, the main criteria in developing new systems are the maximum spectral and energy efficiency. This is especially critical for systems focused on operating in the already densely occupied frequency ranges of (0.7–3) and (8–12) GHz.

Wireless data networks of third and fourth generations widely use the technology of orthogonal frequency-division multiplexing (OFDM), which is based on the use of a large number of closely located orthogonal sub-carriers [1 - 6]. With OFDM, each subcarrier is modulated by a conventional modulation scheme (for example, QAM) at a low symbol rate, maintaining total data transfer rate as in conventional single-carrier modulation systems in the same bandwidth.

Such a signal can be viewed as a set of slowly modulated narrowband signals, but only when there is the possibility to distinguish the signals of adjacent subcarriers.

One of the reasons for the deterioration of the frequency resolution is the expansion of signal spectrum due to the nonlinearity of the transmitter’s cascades and the overlaying of subcarrier signals’ spectrum sidelobes. This may result both in adjacent-channel interference (part of the spectrum of a signal emitted by one radio station falls within the band the other) and intersymbol interference (the intersection of the spectrums of subcarriers makes them indistinguishable).

One of the key features of 5G is an opportunistic spectrum access [7 – 9]. Its concept is to give users a temporary access to the unused frequency bands (so-called "white space"). The implementation of such technology will broaden the term “spectrum efficiency”, since we’re talking now just about effective use of a given channel, but about the frequency domain in general. What complicates its application is the necessity to preliminary check the spectrum occupation. At sufficiently high noise power a "false occupation" decision can be made. On the other hand channel fadings may lead to a "false white space" error. More accurate methods require the use of sophisticated algorithms and the increase of integration (estimation) time, which has a disadvantageous effect on performance. This method allows increasing the overall spectral efficiency, but at the same time exacerbates the problem of out-of-band radiation even further.

The easiest way to reduce this kind of interference is to use the guard intervals. However, they degrade the overall spectral efficiency because a part of the frequency band is not used for transmitting useful signals. A more complicated way is the synthesis of signals with specified parameters, namely with determined spectral characteristics. Indeed, if during the formation of, for example, an OFDM-signal we suppress sidelobes of subcarriers it can significantly increase the resolving power and reduce intersymbol interference.

Similar difficulties and much more stringent requirements for their solution are found in Doppler radar and sonar systems. The effectiveness of their solutions is greatly influenced by the choice of the probing signal, its structure and properties. The shape of partial pulse, its duration, the number and placing of pulses in the pulse train – these parameters are determined by the required resolution, measurement accuracy, dynamic range of the radar, etc. The echo signal received by sonar will be delayed in time (proportionally to the distance traveled) and have Doppler frequency shift (proportional to the radial velocity of the target). It is necessary to measure that shift. Sidelobes in this case are interference, resulting in possible false alarm error. Disposing of them while keeping the pulse nature of the signal cannot be physically implemented. Therefore, we should aim to maximally weaken their influence [10 – 13].

On the majority of the above criteria we can justify our choice by analyzing the ambiguity function (AF): width of its main lobe, the distance between the main and side lobes, the position of zeros in the cross section of AF.

The traditional way to reduce the side lobe level (SLL) of AF is weighting, in which predistortions in the form of amplitude modulation are introduced in the received or reference signals. However, the formation of probe signals modulated in amplitude by one of the weighting functions in the non-linear high-power transmission devices involves technical difficulties.

The presence of narrowband interference may complicate it even further. In such cases, sometimes a concentrated suppression of a certain frequency range in the area of AF sidelobes is required. The efficiency of solutions of radar challenges, based on the analysis of Doppler spectrum, is determined by the sidelobe level in the frequency cross section of AF. Therefore it makes sense to synthesize the probing signals with a rectangular envelope,

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ensuring low SLL in cross section of AF along the axis of the Doppler frequencies.

Mathematical description

Suppose that probe signal $x(t)$ is a high frequency harmonic oscillation with f_0 frequency, amplitude-modulated by $w_z(t)$ function:

$$x(t) = \mathrm{Re}\left\{ w_z(t) e^{j2\pi f_0 t} dt \right\}, \quad t \in (0, T_l),$$

where $w_z(t)$ – real function which takes values of 0 and 1; T_l – duration of the probe signal.

Cross section of AF $\chi(F)$ of probe signal $x(t)$ along the frequency axis is described by the equation:

$$\chi(F) = \int_{-\infty}^{\infty} x(t) x_{ref}^{*}(t) e^{-j2\pi F t} dt \,,$$

where $x_{ref}^{*}(t) = exp(-j2\pi f_0 t),\ t \in (0, T_l)$ – reference signal.

After analyzing the above relations, we can see that the $\chi(f)$ function is the Fourier image of the $w_z(t)$ function. This implies that the form of the $w_z(t)$ function defines sidelobe level of both AF and the spectrum of the emitted signal.

Based on this, it can be asserted that the introduction of the weighting function into the emitted and not into received or reference signals has a number of advantages: weighting results not only in reduction of SLL of AF, but also in reduction of the radiated power outside the main lobe (out-of-band radiation level). This allows to us to reduce the influence of side lobes of powerful echo signals invading the operating band.

At the moment this is the most urgent problem for sonars, because the majority of them use high-power thyristor generators [14 – 15]. Thyristors operate on the principle of electronic key, which means that an additional requirement is placed on the emitted signal – a two-level envelope.

Thus, the problem of synthesis of the probe signal is reduced to the choice of a two-level function $w_z(t)$, which will provide the desired sidelobe level in the frequency cross section of AF.

Suppose that the original function $w_0(t)$ describes the signal in the form of pulse train with duration T and number $N+1$ of short pulses of unit amplitude with a Δt repetition period inside a train. It is necessary to reduce the SLL at f_1 frequency in spectral density $W_0(f)$ of that signal. To do this, we multiply $W_0(f)$ by $S_1(f)$, which equals zero at the frequency and is close to zero in the vicinity of f_1. According to one of the properties of the Fourier transform, the resulting spectrum will correspond to a signal, obtained by the convolution of function $w_0(t)$ with $s_1(t)$, spectrum of which is a function $S_1(f)$. If it is necessary to suppress several points of the frequency axis, then new function $w_z(t)$ can be obtained by performing multiple convolutions:

$$w_z(t) = w_0(t) * s_1(t) * s_2(t) * \ldots * s_z(t) = w_0(t) * s(t).$$

The described requirements are met by a function $s_i(t)$, which consists of two short pulses with unit amplitude, separated by an interval τ_i. The amplitude spectrum $S_i(f)$ of this function has the form of

$$S_i(f) = G(f) \left| \cos \pi f \tau_i \right|,$$

where $G(f)$ – module of spectrum of partial (single) pulse.

Function $S_i(f)$ becomes zero at frequencies $f_{i,k}$, equal to

$$f_{ik} = (2k-1)/(2\tau_i) = (2k-1) f_{i1}\,,\ k=1,2,\ldots;\ i=1,2,\ldots,z.$$

Thus, the convolution of the original function $w_0(t)$ with functions $s_i(t)$ leads to the formation of the zeros in the spectrum of the resulting weighting function at frequencies $f_{i,k}$ and to a reduction of sidelobe level near these frequencies.

If we define the frequencies f_{i1} by the equation:

$$f_{i1} = \beta 2^{i-1} / T,\ 1 \le \beta < 2,$$

then the set $\{f_{i,k}\}_{i=l}^{z}$ forms a grid of evenly distributed frequencies with β/T increment.

From the above formulas it follows that the interval τ_i between pulses in the function $s_i(t)$:

$$\tau_i = T / \left(\beta 2^i\right)..$$

Convolution of the original function $w_0(t)$ with functions $\{s_i(t)\}_{i=l}^{z}$ can be obtained as a result of the algorithm:

$$\left.\begin{aligned} w_1(t) &= w_0(t) * s_1(t) = w_0(t) + w_0(t-\tau_1); \\ w_2(t) &= w_1(t) * s_2(t) = w_1(t) + w_1(t-\tau_2); \\ &\ldots \\ w_z(t) &= w_{z-1}(t) * s_z(t) = w_{z-1}(t) + w_{z-1}(t-\tau_z). \end{aligned}\right\} \quad (1)$$

Function $w_z(t)$ will have a rectangular envelope and match the sequence of unit pulses only in case of a mismatch of pulses of summable sequences in time. The condition for such mismatch is interval Δt being aliquant to interval τ_i, which takes place when

$$\tau_i / \Delta t = N / \left(2^i \beta\right) \ne m,\quad m = 1, 2, \ldots$$

Lets examine the spectral properties of the non-periodic pulse sequence $w_z(t)$.

The module of spectrum $W_z(f)$ of function $w_z(t)$ can be described by the following equation [1] :

$$W_z(f) = G(f) \left| \sum_{n=-\infty}^{\infty} \frac{\sin \pi T(f - nN/T)}{\pi T(f - nN/T)} \prod_{i=1}^{z} \cos \pi f T / \left(2^i \beta\right) \right|.$$

In this equation the spectrum in the form of a product of cosines corresponds to the signal $s(t)$, which represents the results of a convolution of z pairs of unit pulses separated by intervals τ_i. While performing operations of convolution a sequence of unit pulses is formed with a repetition period $T / \left(\beta 2^z\right)$ and duration:

$$T' = \sum_{i=1}^{z} \tau_i \left(1 - 2^{-z}\right) / \beta\ .$$

The module of spectrum $S(f)$ of such a limited in duration sequence is described by the equation:

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$$S(f)=G(f)\left|\sum_{n=-\infty}^{\infty}\frac{\sin\pi T\left(1-2^{-z}\right)\left(f-n2^{z}\beta/T\right)\beta}{\pi T\left(1-2^{-z}\right)\left(f-n2^{z}\beta/T\right)\beta}\right|.$$

From these relations it follows that the sidelobe suppression in the ΔF frequency band will be proportional to the value of parameters $\beta 2^{z}$. Assuming $N > \beta 2^{z}$, we find that the frequency band is

$$\Delta F = 2^{z-1}\beta/T.$$

In the ΔF band spectrum of $W_z(f)$ is close to that of the Bartlett weighting function (for β = 1 it practically coincides with the spectrum of this function) due to the fact that both the spectrum of $W_z(f)$ and the spectrum of Bartlett function are obtained by multiplying the functions of the *sin(x)/x* type.

The deeper sidelobe suppression can be achieved by a repeated execution of the algorithm. Number of iterations of this algorithm l is related with the desired sidelobe level D and steepness S of the slope with following relations:

$$D = 13{,}4\left(l+1\right);\ S = 6(l+1),$$

where D – in dB, S – in dB/oct.

Applying the algorithm l times must be executed for such values of β_j, $j=2,3,\ldots,l$, for which the functions $s_{i,j}(t)$ and $w_0(t)$ will not have pulses coincident in time, and the envelope of formed signal remains two-level. This requirement for β_1=1 is met by a set of values of β_j with general term:

$$\beta_i = l/\left[l-1/\left(N2^{(j-1)z}\right)\right].$$

When the algorithm is executed three times (1) a signal in the form of a non-periodic impulse sequence is generated with spectral characteristics in the band ΔF close to the properties of the Parzen weight function.

Let us consider parameters of the generated sequence $w_z(t)$. The duration of the sequence is increased to a value of

$$T_l = T\left(1+\left(1-2^{-z}\right)\sum_{j=1}^{l}\beta_j^{-1}\right).$$

The number of pulses N_l in the sequence $w_z(t)$ is $N_l=(N+1)2^{zl}$.

Above it was assumed that as the initial sequence $w_0(t)$, as well as the sequence $s_{i,j}(t)$ consist of short pulses with a duration $\Delta\tau \ll \Delta t_l$. In practice, to increase the radiated power it is expedient to use rectangular partial pulses with $\Delta\tau = \Delta t_l$. In this case, on most part of the interval T_l pulses of summable sequences $w_{i,j}(t)$ merge into a single pulse, and formed two-level envelope of the signal is essentially a pulse-width modulation signal (PWM).

Analysis of the algorithms software implementation and its results in NI LABVIEW

Synthesis was performed using virtual instrument created in the development environment NI LabVIEW. The algorithm requires user to specify 4 input parameters: the desired peak value, suppression level, frequency resolution and width of suppression band. Suppression level determines the number of algorithms iterations, frequency resolution and suppression band are used to calculate length of pulse train, number of pulses in the train and repetition period, peak value is used in forming initial pulse sequence. Number of iterations and frequency resolution are indirectly used later to calculate initial pulse length.

Figure 1 shows the block diagram of formation of array of β coefficients. They are used in calculation of delays and to adjust them later on after compensating the discreteness errors.

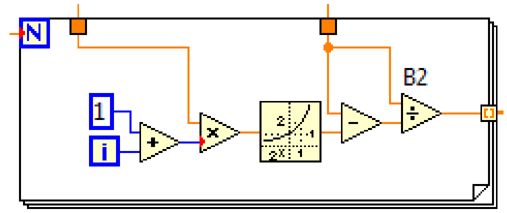

Figure 1 Formation of array of β coefficients

Figure 2 shows the block diagram of formation of delays matrix. Number of algorithm iterations determines the number of rows in the resulting matrix; number of pulses in the train and their length – number of elements in rows.

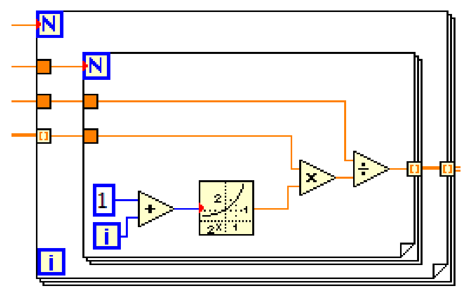

Figure 2 Block-diagram of formation of delays matrix

It is worthy to mention, that initial pulses length is calculated based on the values of the delays in order to minimize pulses intersections during the convolution. Figure 3 shows the block-diagram of this part of the algorithm. Also, due to the specifics of LabVIEW, in case of one algorithm iteration the initial pulses length will be calculated before the delays, in case of 2 or more – vise versa.

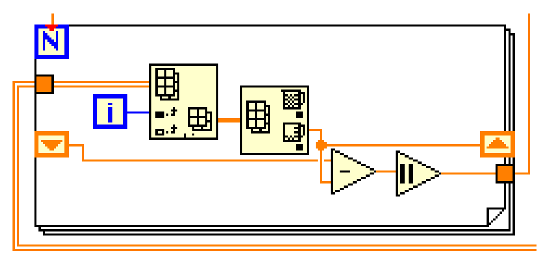

Figure 3 Calculation of the initial pulses length

Main feature of developed software implementation of the algorithm is the core of virtual instrument, shown in figure 4.

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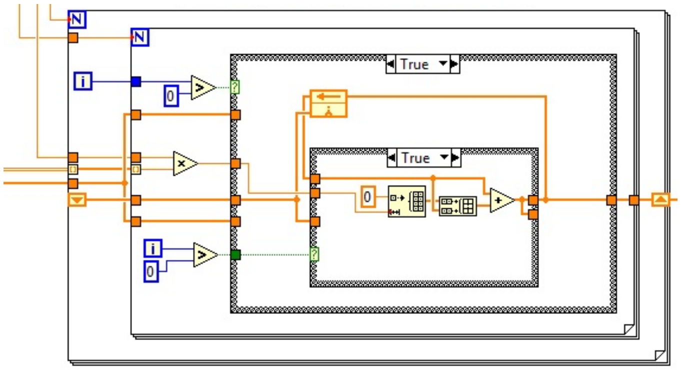

Figure 4 Core of the virtual instrument

In modern CAD the convolution operation (see figure 5) is implemented on the basis of multiplying the Fourier images of signals followed by applying IFFT to the result. The obvious downside of such method is the inefficient usage of RAM and time, wasted on computing intermediate data (Fourier images).

Since we have chosen two Dirac deltas (two short unit pulses) with τ_i delays as the weighting function, the convolution can be implemented by summing original signal with the same signal, delayed by τ_i . It decreases computing time almost 3 times, since instead of 4 operations only 2 are used and "add" executes faster than "multiply".

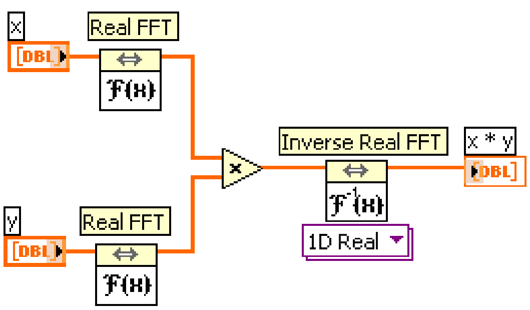

Figure 5 Block-diagram of convolution operation

The core consists of 2 loops with a shift register and a feedback-node and 2 logical case-structures. The initial pulse train is only used during the first convolution of the first iteration of the algorithm. The outer loop is the repeater of the algorithm, used in cases when one iteration is not enough to achieve the required level of suppression. Its iteration terminal (i in blue square) is used to control the outer case-structure and switch the entry data source from the initial train generator to the values stored in shift register.

Inner loop is the software implementation of the convolution operation. Its iteration terminal is used to control the inner case-structure, which copies the entry array of data, delays it for the corresponding value of τ_i and adds it to the non-delayed array. The result matches the operation of convolution with 2 Dirac deltas but doesn't require their implementation and saves a lot of computing time.

This is the stage that shows the error due to discrete nature of initial signal. Figure 6 shows the example of such errors. Since it is impossible to use irrational numbers in software they are being rounded. This causes β coefficients and delays to contain inaccuracies, which in turn in some cases cause fronts of pulses to intersect during the convolutions. This results in appearance of narrow "spikes" with peak value N times greater than specified, where N is a positive integer between 2 and the number of convolutions per iteration.

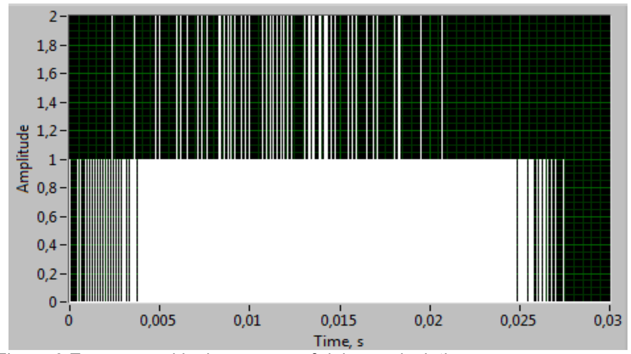

Figure 6 Error caused by inaccuracy of delays calculation

It is noteworthy, that different values of β coefficients lead to a variety of effect: from spikes many times greater in length and/or peak value than specified to no errors at all.

The simplest way to negate errors of this kind is to apply an amplitude limiter after the core. It will ensure the reduction of any spikes down to specified peak value. Figure 7 shows an example of such a limiter made in NI LabVIEW. Figure 8 shows the signal from figure 6 after the application of this limiter.

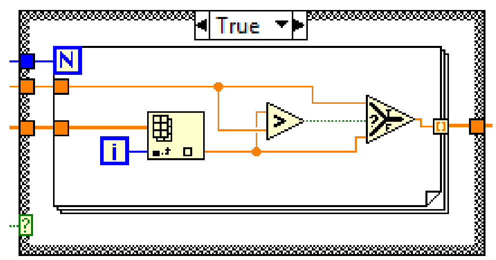

Figure 7 Amplitude limiter

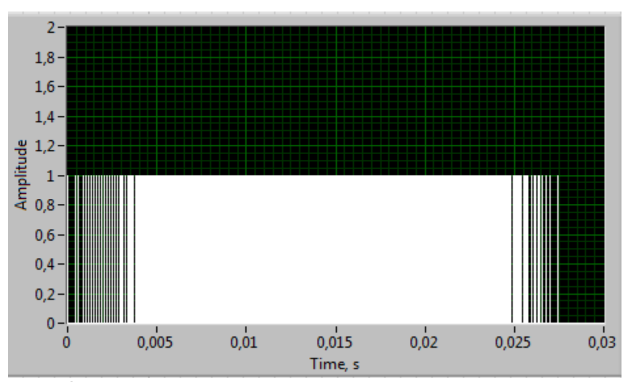

Figure 8 Signal with compensated discreteness errors

But if it is implemented in real systems (not the VI), an amplitude limiter means the loss of power, since we're simply

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cutting a part of the initial signals. In order to minimize the power loss we must choose the β coefficients which correspond to the minimal percentage of energy lost after limitation.

Figure 9 shows a block-diagram of such decision making algorithm. For each set of β coefficients there is a corresponding percentage of energy lost in the limitation. Both arrays are chosen from the multitude signals whose characteristics meet the specified requirements. After the cycle we have an array of β coefficients. In most cases it's better to let the following logic of the algorithm to select the coefficient automatically, since it'll pick the one with the smallest power loss, but it can also be done manually in case, for example, a small increase of power loss will lead to the large increase of compression factor.

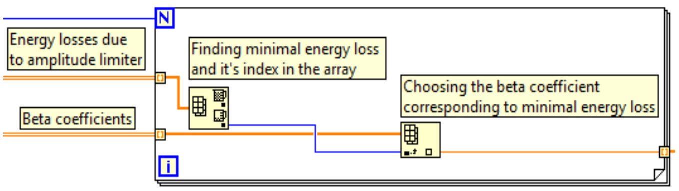

Figure 9 Decision making algorithm

For example, the following figures show the results of the synthesis of the signal envelope with a width Δf of the main lobe of spectrum no more than 100 Hz, suppression band ΔF of 2 kHz and SSL D in that band not more than 40 dB.

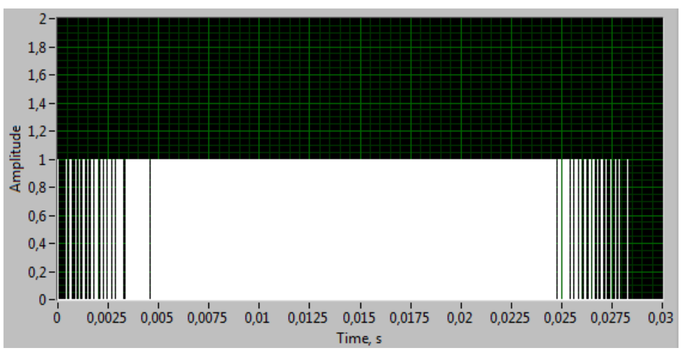

Figure 10 Temporal function of envelope of the synthesized signal

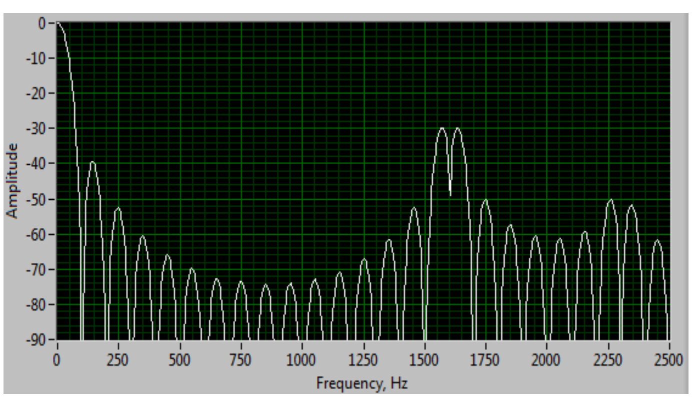

Figure 11 Spectrum of envelope of the synthesized signal

As you can see from the figure 11, sidelobe level in range of 1550-1650 Hz reaches the level of –30 dB. It is possible to additionally suppress them by shifting the zeros of the weighting function. To achieve such effect the algorithms increases the β coefficients, which will decrease τ_i .

Figures 12 and 13 shows the envelope of the synthesized signal with changed delays τ_i .

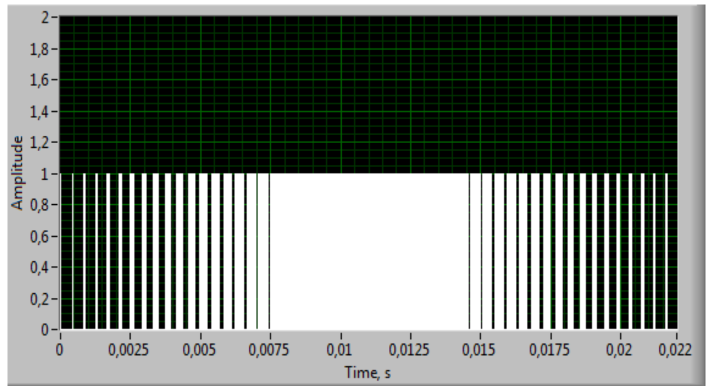

Figure 12 The envelope of the signal after delays correction

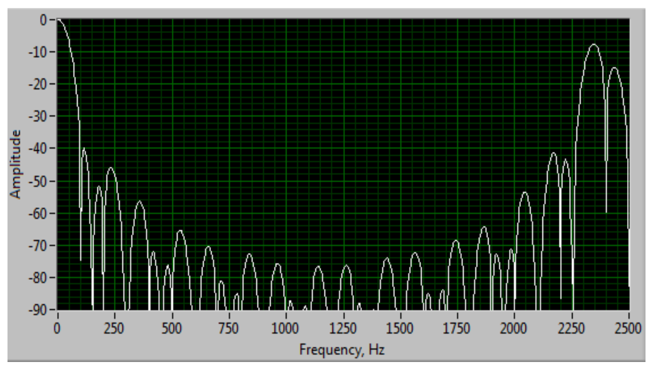

Figure 13 Spectrum of the signal envelope after delays correction

After shifting the zeros of weighting function, the sidelobes in the band of 100-2100 Hz do not exceed –40 dB, satisfying the specified value of D. It is noteworthy that increasing the β leads to a decrease of the duration of resulting envelope, and therefore, to expansion of the main lobe. This is exactly why when generating the original pulse train, its duration is chosen so as to ensure the frequency resolution is no worse than a predetermined value in any case.

Spectrum of envelope of signal after passing the ideal lowpass filter (as a model we use 100th order IIR lowpass filter with a Butterworth frequency response) with a cutoff frequency of 2100 Hz, simulating signal transmission through narrowband antenna during its radiation, and the autocorrelation function of the transformed by filter envelope are shown in figures 14 and 15. The signal delay after the filter is due to its high order.

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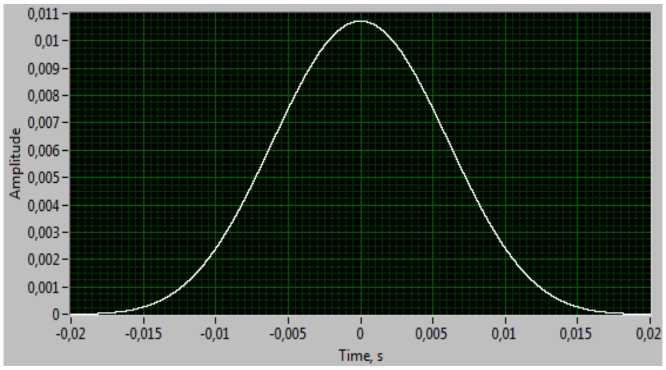

Figure 14 Autocorrelation function of the transformed by filter envelope

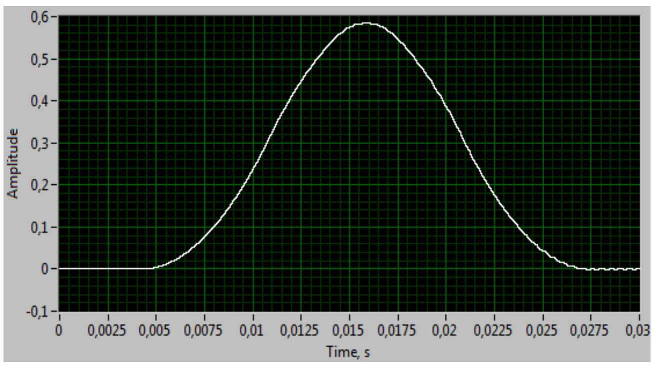

Figure 15 Signal envelope after passing the ideal lowpass filter

Low-pass filter can be reconfigured to simulate the effect of virtually any antenna on the envelope. Autocorrelation function can be used in building matched filters, widely used in radar and sonar systems.

Conclusion

The paper presents a regular algorithm of synthesis of two-level signal envelope with low sidelobe level of its spectrum in predetermined frequency band and its analysis using the virtual instrument built with graphical programming language NI LabVIEW.

Let us study the pros and cons of using this method of signal synthesis.

Pros:
– Simplicity. The whole algorithm is build using the simplest mathematical and logical operations, which makes its future development, improvement and debugging easier.
– Automation. It requires only 4 input parameters to function, will give the most optimal in terms of energy efficiency result and is able to provide alternative variants if special conditions are specified.
– No upper limit. The algorithm is capable of synthesizing a signal with virtually any spectrum parameters. The only limits are placed by available computing resources.

Cons:
– Cyclical nature. In order to achieve the most optimal case in terms of energy efficiency, the algorithm needs to repeat the synthesis.
– Inconstant hardware requirements. The amounts of required memory and processing power grow exponentially as the desired suppression level and/or suppression band/resolution ration grows.
– Compression factor. With the increase of β coefficients, the resulting signals length decreases, which in turn decreases the signals process gain – a crucial parameter in radar and sonar systems.

The examples presented in this paper demonstrate the effectiveness of the algorithm. Its simplicity allows repurposing it for a variety of tasks and areas of application.

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