Ultra-Wideband Radio Systems. Their Peculiarities and Capabilities

Abstract

Theory and applications of Ultrawide Band (UWB) radiosystems which belong to a new intensively developing area in radar and communications are considered in the paper. Main distinguishing features of UWB radiosystems are systematized, while difference between such systems and traditional narrow-band ones is analyzed. Scientific and technical challenges faced by UWB radar designers are shown. Some new technical and circuit solutions used in UWB radiosystems are considered on examples of UWB radars. Possible directions in development of UWB technology and technique for such systems are suggested.

1. Introduction

The conventional radio-technical systems operate at a narrow frequency band and employ harmonic (sinusoidal) signals at their carrier oscillation frequency for data transmission. It came about this way historically, because a simple LC circuit allowed for easy generation of the oscillation at the needed frequency, the sinusoid being the natural oscillation of this circuit. The frequency selection has since remained the main technique for radio channel division, with the majority of radio-technical systems operating at such a band of frequencies, which is substantially smaller than their center frequency. All theory and applied practice of the modern radio engineering is based on this feature for successful operation.

However, the narrowband operation restricts the informational capabilities of the radio-technical systems. As shown by Shannon C. [1], the amount
of data transmitted per unit time $H$ is directly proportional to the data transmission channel bandwidth $\Delta f$:

$$H = \Delta f \cdot \log \left(1 + \frac{\text{power of signal}}{\text{power of noise}}\right).$$

Increasing of the data transmission capabilities of a system calls for widening of its operating band. The alternative to it would be a mere increasing of the data transmission time.

A dramatic growth in data transmission flows in the modern world makes this problem especially challenging. The conventional radio-technical systems, in which the operating bandwidth does not exceed 10% of the center frequency, have actually exhausted their capabilities as regards data transmission. And for this reason alone, one of the ways to develop further information networks would be the using of signals across wide and UWB frequency bands [2–10].

Creation of UWB radio systems, like any other relevant technologies, mandates postulation of certain theoretical fundamentals that should enable to compute their performances accurately and define the requirements for radio system components while designing the hardware. For all that, despite the growing number of published papers that have seen light of late, an orderly UWB system theory is, practically, non-existent. The reason for this is quite objective. The processes of transmission of radio signals differ considerably in the narrowband systems and UWB systems. A study that could be made on those differences shall allow one to understand when the traditional theory could be applicable to design the UWB radio systems or when this theory is unusable and some novel techniques should be in order.

This review paper brings into sharp focus both state-of-the-art informational capabilities that the radio-technical systems acquire when using UWB waveforms and basic features of such UWB radio systems that emerge with employment of those signals.

2. Terminology

In 1990, Defense Advanced Research Projects Agency (DARPA) under US Department of Defense (DoD) introduced the definition [5] of the so-called fractional frequency band$^1$:

$$\eta = \frac{f_{\text{upper}} - f_{\text{lower}}}{f_{\text{upper}} + f_{\text{lower}}}.$$

In accordance with this definition, the UWB systems embrace those systems and waveforms for which

$$0.25 < \eta \leq 1.$$ This definition is widely referred in literature at present.

Nonetheless, this definition as introduced by DARPA has not met with much success. A mere case in point (Fig. 1) and a diagram (Fig. 2) indicate that signals, at the same frequency range $\Delta f = 1/\tau$, that have similar data transmission capacities$^2$ fall, in accordance to DARPA, under different classes$^3$.

At the same time, a feature common to all UWB waveforms is their capability of resolving objects over distance, determining, in this way, their structure. In this connection, it might be expedient to catalog signals according to this feature and refer to ultra-wideband waveforms, only if their spatial duration $c\tau$ ($c$ is the speed of light) becomes smaller than the size $L$ of object illuminated (more often than not, not

Fig. 1. Example of signals of equal duration.

a) $\Delta f = 1/\tau = 1000$ MHz; $\tau = 1$ ns; $f_{\text{carrier}} = 1000$ MHz; $\eta = 0.5$ (UWB signal).

b) $\Delta f = 1/\tau = 1000$ MHz; $\tau = 1$ ns; $f_{\text{carrier}} = 50000$ MHz; $\eta = 0.01$ (Narrowband signal).

Fig. 2. Classification of signals.

$^1$The very same expression multiplied by a factor of two became known as relative frequency band.

$^2$From here on, $\tau$ – a simple signal duration or the autocorrelation function width of a signal with chirp modulation.

$^3$The absence of clarity in the definition of UWB waveforms employed in radars prompted establishment within the framework of the Society IEEE of an International Working Group (UWB Working Group), which carries on R&D on a special reference standard “IEEE Standard P 1672 UWBR Terms and Definitions”.

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smaller than the antenna size)\(^4\): \(c \tau \ll L\). This kind of definition brings about certain inconveniences, since a class of signal (and a class of radio system) becomes dependent on the dimensions of object placed under illumination. However, the inequality \(c \tau \ll L\) plays a major role in more than just formal reference of signals (and systems) to the UWB class. Fulfilling of this inequality makes apparent an important peculiar feature of all UWB systems: \textit{UWB waveform changes in the course of illumination of target, its reflection and reception}. It is this very feature that precludes some methods of the conventional radio engineering from being used in UWB radio systems design.

Below, the consideration is given to such waveforms and systems, in which the inequality \(c \tau \ll L\) is fulfilled.

3. UWB Waveform Variation during Transmission-Reception. General Approach

The narrowband signals have a unique property. With such widely used transformations as addition, subtraction, differentiation and integration the narrowband signal shape, which is determined by the harmonic reference carrier wave law, remains actually unchanged. What really takes place is just a changing of the amplitude of the signal and its shift over time (phase variation). The UWB waveforms undergo changes not only in their parameters, but in their shapes as well, during those (and other) transformations.

Let us now take a look, as an example, at how the UWB waveform changes during target illumination. A schematic of the radar is given in Figure 3. According to this schematic, a multibeam antenna, made as an array of \(N\) radiating elements, is used for transmission, while the receiver is connected to a single-beam (horn or mirror) antenna.

The signal of transmitter in the form of the current pulse \(S_0(t)\) with the duration \(\tau\) comes to each radiating element of the array, the aperture of which being of the size \(L\). During radiation the first change in the signal waveform occurs, namely its differentiation, because with the majority of simple radiating elements the electromagnetic field pulse shape \(S_1(t)\) is a derivative of the current pulse shape in the radiating element. The current pulse in the radiating element has the spatial extent \(c_0 \tau\), where \(c_0\) is a charge motion velocity in the radiating element material. For the UWB waveform \(c_0 \tau \ll L\). This makes the process of differentiation extended over time, running on until the current pulse completes its cycle in the radiating element. As a result, an additional (second) UWB waveform change occurs \(S_2(t)\), since during the time of this cycle, in the space appear more than one \((K)\) electromagnetic field pulses separated by the intervals \(\Delta t_k\). Since the visible radiating element wavelength \(L\) changes relative to the angle \(\theta\) between the plane of the mouth and direction toward the receiving point, then the signal waveform \(S_2(t)\) should differ at different angles of surveillance. As a result, the radiating element spatial field distribution in the far zone (directional pattern over field) changes its position and shape during the process of current pulse motion across the radiating element, i.e. it becomes transient (this issue is considered in a more detail in Chapter 4).

The waveform \(S_2(t)\) is emitted simultaneously by all radiating elements \(N\) of the multi-element antenna. In the process, a delay occurs, in the directions that are different from normal to the antenna mouth, between field pulses, which come in from different radiating elements. For adjacent radiating elements this delay would be \((d/c) \cos \theta\), where \(d\) - a distance between the radiating elements. Resulting from the addition \(N\) of the delayed signals \(S_2(t)\), a new (third) shape is created of the UWB waveform \(S_3(t)\). While performing surveillance at different angles \(\theta\), the resulting waveform \(S_3(t)\) shall have variable, sometimes rather complicated, shape. Fig. 4 from reference [2] shows an instance of the waveform variation of this signal relative to the angle \(\theta\) for an array of four radiating elements. The field for the individual radiating element \(S_2(t)\) is shown in this figure for the sake of simplicity as rectangular video-pulse.

The fourth change in the UWB waveform occurs during its return from local reflective elements \(M\) ("brilliant specks") of target, arranged arbitrarily over the length of the target \(L_1\). The waveform \(S_4(t)\) returns from each of these elements with a different delay \(\Delta t_m\). The sum of the displaced signals vs. time \(S_4(t)\) forms the waveform \(S_3(t)\), the shape of which (the number of maximums, intervals between them \(\Delta t_m\) and their amplitudes) depends on geometry and materials of the target. This waveform is also dependent on impulse response of the target brilliant specks \(h_m\), which may turn out to be frequency filters for this signal. This kind of signal is called the "portrait of the target". The entire portrait is formed in the time \(T = 2L_1/c\), its waveform varying vs. target surveillance angle.

The fifth change in the UWB waveform \(S_5(t)\) occurs during signal reception. Below, Chapter 5 will deal with causes of those changes. They have to do with the dependence of a receiving antenna directional pattern form on that of a field pulse returning from target and on the temporal shift of current pulses, which is induced by the field in near and far ends of the antenna.

\(^4\)From this standpoint, it might be more logical to name the considered waveforms as ultra-short impulse ones, which is done by several authors [11]. Yet, the term “ultra-wideband” has long since gone down as such in literature the world over, and its re-naming can hardly be accounted for at present.
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\[ S(t) = \frac{dS(t)}{dt} \]

\[ S(t) = \sum_{k=1}^{K} S_k(t - \Delta t_k) \]

\[ S(t) = \sum_{n=1}^{N} S_n(t - \frac{nd}{c}\cos \Theta) \]

\[ S(t) = \int_{-\Delta t}^{\Delta t} S(t) \cdot h(t - \Delta t) \cdot \Delta t \]

Fig. 3. Variation of UWB waveform during transmission and reception.

Fig. 4. Field pulse shapes at various angles \( \theta \).
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Signal of the transmitter

Radiated signals

Reflected signal

2.0 4.0 6.0 8.0

2.0 4.0

Radiated signal

First BP

Second BP

Change of the signals reflected from different brilliant points (BP) of the target

Fig. 5. Variation of shapes of real UWB waveforms.

Fig. 6. Model of dipole shoulder.

4. Waveform Variation during Radiation

Let us consider the ultimate radiating element of UWB waveform as linear antenna (a length of wire or a shoulder of dipole) and present the physical interpretation of the processes that cause the changes in the waveform $S_1(t)$ and $S_2(t)$ [12].

A model of this radiating element of the length $L$ is given in Figure 6. Let us separate it into elementary radiators with the linear size $\Delta L$. Designations in this Figure are as follows: $\Delta L_j$ – a $j$-th elementary radiator; $L_j$ – its coordinate; $R$ – a distance as far as the point of surveillance $M$; $\theta$ - an angle between direction to the point of surveillance and plane of the mouth. The current pulse $i(t)$ arises in the point $O$ and propagates along the radiating element. In this way, a series excitation of its aperture takes place.

The electromagnetic field arising in the far zone of the radiating element is associated with the movement of charges through radiating element wire via the known relationship [13]:

$$E(t) = \frac{q}{4\pi\varepsilon} \left[ \frac{1}{c^2} \frac{d^2 \mathbf{r}}{dt^2} \right]$$

where $c$ – the speed of light; $q$ – a charge; $\mathbf{e}_r$ – a unit vector of the charge.

Since $i(t) = dq/dt$, the charge acceleration in the radiating element is induced by the field which is proportional to the first current derivative.

When the current pulse has appeared, the excitation of the first elementary radiator occurs in the point $O$. In the time $R/c$, an electromagnetic field will arise in the far zone, which will look like as follows [14]:

$$E_1(t, \theta) = \frac{Z_0 \sin \theta}{4\pi c R} \times \frac{d}{dt} \left[ i \left( t - \frac{L_1}{c_0} - \frac{R - L_1 \cos \theta}{c} \right) \right] \Delta L \ (1)$$

where $Z_0$ – a characteristic impedance of free space.

The second summand in parentheses determines the delay of the signal in this wire and the third one indicates the delay of the signal in space.

In the time $\Delta L/c$, the current pulse will excite the second elementary radiator. The second electromagnetic field $E_2(t, \theta)$ will emerge that would be similar to (1), replacing $L_1$ with $L_2$.

Travelling along the wire, the current pulse will excite one by one the subsequent elementary radiators. To get a relatively plain physical picture of the processes taking place in the radiating element, we shall introduce certain simplifications. We shall neglect losses in the radiating element wire and losses to radiative radiation; we shall assume that there is no reflection from the end of the radiating element, and upon excitation of the last, $N$-th, elementary radiator the current pulse will be absorbed by the load, and we will act on assumption, as well, that there is no delay in the wire and thus $c_0 = c$. This kind of radiating element will be further known as matched one.

The total field of all elementary radiators in the far

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The zone acquires the following form:

$$E_{\Sigma}(t, \theta) = \frac{Z_0 \sin \theta}{4\pi R} \frac{1}{c} \sum_{j=1}^{N} \int_0^L \frac{dL}{dt} \left[ i \left( t - \frac{L_j}{c} - \frac{R - L_j \cos \theta}{c} \right) \right] \Delta L. \quad (2)$$

We shall now use, instead of the discrete representation of the radiating element, the continuous one. For this purpose, we shall get the length of elementary radiators to strive to zero $\Delta L \to 0$ and their number to strive to infinity $N \to \infty$. Then, in (2) the summation turns into integration:

$$E_{\Sigma}(t, \theta) = \frac{Z_0 \sin \theta}{4\pi R} \frac{1}{c} \int_0^L \frac{dL}{dt} \left[ i \left( t - \frac{L}{c} - \frac{R - L \cos \theta}{c} \right) \right] dL. \quad (3)$$

The expression (3) describes the electric component of the electromagnetic field in the far zone for an extended radiating element excited at one end by the current of arbitrary shape.

The field that is determinable by the derivative of the current over time is formed by various points of the radiating element as far as the pulse travels along the wire. The expression found in parentheses determines its running time, taking into account this delay. By assuming the derivative of this time over $dL$, we shall raise a possibility to make in (3) a replacement of the variables [15]:

$$dt = \frac{\cos \theta - 1}{c} dL.$$

As a result, we obtain the integral of the derived function over the same variable, which is equal to the function itself. Then,

$$E_{\Sigma}(t, \theta) = \frac{Z_0 \sin \theta}{4\pi R} \frac{1}{c} \int \left[ i \left( t - \frac{L}{c} - \frac{R - L \cos \theta}{c} \right) \right]_0^L \frac{dL}{dt} \left[ i \left( t - \frac{L}{c} - \frac{R - L \cos \theta}{c} \right) - i \left( t - \frac{R}{c} \right) \right]. \quad (4)$$

One can see from this expression that, in the general case, the radiating element field must consist of two fields, positive and negative, each of which follows the shape of the exciting current. The shape of the total field hinges on the relationship between the radiating element length $L$ and spatial duration of the exciting pulse $c\tau$. This shape will also depend on the angle of surveillance $\theta$. We shall consider these relationships by citing a specific instance when the exciting current pulse appears as Gaussian curve with unitary amplitude (Figure 7, solid curve):

$$i(t) = \exp \left( -4 \left( \frac{t - \frac{L}{c}}{\tau} \right)^2 \right), \quad (5)$$

where $\tau$ - a pulse duration at level 0.5. The derivative of this pulse is a symmetrical bipolar pulse (Figure 7, dotted curve). By substituting (5) in (3) and performing the differentiation and integration, we shall obtain:

$$E_{\Sigma}(t, \tau, \theta) = \frac{Z_0 \sin \theta}{4\pi R} \left[ \frac{1}{\cos \theta + 1} \right]$$

$$\times \left\{ \exp \left[ -4 \left( \frac{t - \frac{L}{c} - \frac{R - L \cos \theta}{c}}{\tau} \right)^2 \right] \right.$$  

$$\left. - \exp \left[ -4 \left( \frac{t - \frac{R}{c}}{\tau} \right)^2 \right] \right\}. \quad (6)$$

The formula (6) indicates that the radiating field is indeed a sum of two fields, one of which is radiated at the moment when the current pulse enters the radiating element excitation point and the other exactly at the moment when this pulse reaches the end of the radiating element. This process is sometimes explained as radiative radiation from the excitation point and from the radiating element end. However, the physics of phenomena occurring in the radiating element will be somewhat different.

We shall now consider formation of the field $E_{\Sigma}(t, \tau, \theta)$ by using the discrete radiating
element model (Fig. 6). In Figure 8, the solid curve stands for fields excited by elementary radiators in the point $M$ at the angle $\theta = 90^\circ$, the value of $L/c\tau$ being $\gg 1$. One can gather from this Figure that the fields of the elementary radiators are shifted over time due to the current pulse delay and that they have positive and negative half-waves the areas of which are equal. That is why they will be partially compensated during the summation. The degree of this compensation depends on the ratio between radiating element length $L$ and pulse duration $c\tau$. The complete field compensation starts off at the moment of time $t_k = \tau$. Uncompensated will remain only a part of the elementary radiator fields that are located in the vicinity of the excitation point and toward the radiating element end (Figure 6, dotted curve). For this reason, at $L/c\tau \gg 1$ the radiating element field will look like radiation off the power point and off the end. By comparing the current in Figure 7 and field in Figure 8, it is evident that the shape of the field that lingers after the compensation coincides with the shape of the current that excites the radiating element.

As the ratio $L/c\tau$ decreases, the temporal interval during which the field compensation takes place becomes lesser and lesser, as compared to the field pulse duration. The gap between field pulses in Figure 8 becomes ever shorter. In the long run, at $L/c\tau \ll 1$ the compensation actually stops altogether (Figure 9). All of the aperture radiates simultaneously, while the shape of the field resembles more and more the form of the derivative of the current that excites the radiating element. In this way, at $L/c\tau \gg 1$ the shape of the field coincides with that of the current, while at $L/c\tau \ll 1$ it coincides with its derivative.

Figure 10 shows electric components of the radiating element field pulse for various angles of surveillance $\theta$. One can deduce here that, upon changing of the angle of surveillance, the field pulse shape does so as well, since the radiating element projection size varies at a given angle. The angle, at which the field shape can still be viewed as two separate different-polarity pulses with the duration that is equal to the current pulse duration, is determinable by using the expression:

$$\theta_{gr} = \pi - \arccos \left[ \frac{c\tau}{L} - 1 \right].$$

The angle $\theta_{gr}$ depends on the ratio $c\tau/L$. At $c\tau > 2L$, the angle $\theta_{gr}$ takes on negative values. This observation stands to mean that the field in the far zone is not separated into two pulses at any angles whatsoever and, in its shape, it corresponds to the derivative of the Gaussian pulse. As a matter of fact, we have here transition from UWB waveform to narrowband signal.

5. Directional Pattern to Radiative Radiation

The variation of the field pulse shape at different angles of surveillance causes transient temporal behavior of the directional pattern over field. To prove it, we shall use the expression (6). The factor before the square brackets is directional pattern (DP) of the elementary radiator. Inside the square brackets one
shall find the factor of elementary radiator array. It depends on time, angular direction, exciting signal shape and antenna length. The exciting signal shape and antenna length in our case are fixed. To determine the dependence of DP on time, we shall separate several moments of time $t_0, t_1, t_2, t_3, \ldots$ in the interval, during which the current pulse exists in the antenna (Figure 10) and, employing the expression (6), we shall construct the dependence of field on angular coordinates, i.e. instantaneous DPs. Figure 11 shows a family of instantaneous DPs for the shoulder of dipole being excited by the Gaussian current pulse at $L/c \tau \gg 1$.

As one can see from this Figure, the DP maximum changes its direction in the course of existence of the field. Initially, it is directed actually along the axis of the antenna. As the current pulse travels along the wire, this maximum shifts toward the normal to the antenna, with the field amplitude decreasing. The DP scans the space, changing its position from the antenna axis to its normal. During the scanning the DP width and amplitude decrease. The nature of the instantaneous DP family presented in Figure 11 agrees well with the results reported in reference [16] where a rigorous linear antenna model was used for the simulations.

The transient behavior of DP over field makes it unsuitable for computations of the parameters of a radio-technical system, since it does not allow one to determine the directive gain (DG), beam width, etc. Early literature (for example, reference 17) described different options of DP that are used relative to peak amplitude, to peak power, to steepness. However, the most suitable seems to be the energy DP $W_T(\theta)$.

This DP is obtained by way of averaging the power radiated in each angular direction over the time of the current pulse travel through the radiator, describing the density distribution of energy flow being radiated in space $[17,18]$.

$$W_T(\theta) = \frac{1}{Z_0} \int_{-\infty}^{+\infty} E^2(t, \theta) \, dt,$$

where $Z_0 = 120 \pi \approx 377$ Ohm is the characteristic impedance of free space.

The endless limits of integration over time allow one to apply this expression to current pulses of any shape and duration.

To compare the antenna performances, the normalized energy DP comes in very handy:

$$W_{TN} = \frac{W_T(\theta)}{W_{T_{\max}}},$$

where the value of DP in the direction of maximum radiation is calculated according to the formula:

$$W_{T_{\max}} = \frac{1}{Z_0} \int_{-\infty}^{+\infty} E_{\max}^2(t) \, dt.$$
of the radius $R$ should be equal to:

$$W = \frac{R^2}{Z_0} \int_{0}^{2\pi} \int_{0}^{\infty} E^2(\theta, \varphi, t) \sin \theta \, d\theta d\varphi dt.$$ 

By dividing this energy by the surface area of the sphere surrounding the antenna, we shall obtain the energy density flow of equivalent isotropic radiating element:

$$W_{T0} = \frac{W}{4\pi R^2} = \frac{1}{4\pi Z_0} \times \int_{0}^{2\pi} \int_{0}^{\infty} E^2(\theta, \varphi, t) \sin \theta \, d\theta d\varphi dt.$$ 

The obtained expressions now permit to determine the DG of radiating element energy:

$$D_T = 4\pi \int_{0}^{\infty} \frac{E^2_{\text{max}}(t)}{E^2(t, \varphi, t) \sin \theta \, d\theta d\varphi dt}.$$ 

The above linear radiating element for UWB waveforms (Fig. 6) is, as a matter of fact, an antenna with a series excitation. Upon simultaneous excitation of the entire aperture with ultra-short UWB impulse ($c\tau \ll L$), the antenna radiation also acquires unusual properties (a TEM pyramidal horn may be cited as an example of this antenna). The DP for the antenna of this type can be derived from the expression (4) by excluding from the parentheses the summand $L/c\tau$ that determines the current pulse delay due to its travel across the aperture.

Figure 13 shows a family of instantaneous DPs over field for several moments of time $t_0, t_1, t_2, t_3, \ldots$ in the interval, during which the exciting current is running through the antenna. As different from the preceding case, the DP axis does not change its direction, although the DP itself, during the current pulse travel through the radiating element, splits into two diverging rays. This has to do with the fact that, in an antenna with a parallel excitation, like in an antenna with a series excitation, the field in the far zone changes its shape at different angles. Figure 14 presents a family of normalized energy DPs for an antenna with a parallel excitation at different values of the ratio $L/c\tau$.

The above variations of UWB waveforms as radiated by separate radiating element are common in character. Depending on the type and geometry of the antenna, those changes may come up differently. The more complicated the antenna, the more complicated the field shape of radiated UWB waveforms. Even transition from one dipole shoulder to entire dipole and considering of the signal reflections from its ends makes the picture of radiated field noticeably more difficult to grasp, obscuring the understanding of the process of formation of the field pulse and that of the antenna DP. For this reason, the above physical interpretation of the processes occurring in the ultimate antenna during radiation of a “short” UWB waveform is bound to help understand the problems one has to face while designing UWB radio systems.
of the antennas. As a consequence, the receiving antenna directional pattern shall now depend on this arrangement and due to this the DP of this antenna will differ in the modes of transmission and reception of UWB waveform. This circumstance does not allow for the use of the reciprocity theorem in order to determine the directional patterns of UWB receiving antennas relative to their directional patterns in the transmission mode, as it is the norm in the conventional narrowband antenna theory.

![Fig. 15. Setup of antennas.](image)

With a view of determining the dependence of the shape of UWB waveform received on the reciprocal orientation of transmitting and receiving antennas, we shall turn to Figure 15 which shows the setup of those antennas that are arranged as dipoles of the lengths $L_T$ and $L_R$. To simplify this problem, the arrangement has no illuminated target, excluding thereby its influence on performances of the receiving antenna. The receiving antenna is located at a distance $R$ from the transmitting antenna in its far zone (point M in Figure 6). The load and dipole ends have been matched across the signal bandwidth and do not reflect energy. The electric component of the field is radiated by the transmitting antenna in the direction of the receiving antenna at the angle $\theta_T$ between the line of its aperture and direction toward the receiving antenna, being incident on the receiving antenna at the angle $\theta_R$ between the line of its aperture and direction toward the transmitting antenna.

The electric component of the field of one shoulder of radiating dipole is described by the formula (4), which, considering new designations, shall assume the following form (the index with $E$ stands for the number of the shoulder):

$$E_1 (t, \theta_T) = \frac{Z_0 \sin \theta_T}{4\pi R} \left[ \frac{1}{\cos \theta_T - 1} \right] \times \left\{ i \left( t - \frac{L_T}{c} - \frac{R - L_T \cos \theta_T}{c} \right) - i \left( t - \frac{R}{c} \right) \right\}.$$

For the second dipole shoulder, the electromagnetic field in the far zone is determined using the following expression:

$$E_2 (t, \theta_T) = \frac{Z_0 \sin \theta_T}{4\pi R} \left[ \frac{1}{\cos \theta_T + 1} \right] \times \left\{ i \left( t - \frac{R}{c} \right) - i \left( t - \frac{L_T}{c} - \frac{R + L_T \cos \theta_T}{c} \right) \right\}.$$

Then the total field radiated by the dipole would be:

$$E_T (t, \theta_T) = E_1 (t, \theta_T) + E_2 (t, \theta_T)$$

$$= A_1 i \left( t - \frac{L_T}{c} - \frac{R - L_T \cos \theta_T}{c} \right) - A_1 i \left( t - \frac{R}{c} \right) + A_2 i \left( t - \frac{L_T}{c} - \frac{R + L_T \cos \theta_T}{c} \right),$$

where $A_1$ and $A_2$, respectively are the amplitude factors

$$A_1 = \frac{Z_0 \sin \theta_T}{4\pi R (\cos \theta_T - 1)}, \quad A_2 = \frac{Z_0 \sin \theta_T}{4\pi R (\cos \theta_T + 1)}.$$
Let us break each receiver dipole shoulder down into the infinite number of elementary sections $dL_R$.

The EMF induced in the elementary section is directly proportional to the projection of the vector $E_\Sigma(t, \theta_T)$ of the electric component of the field that is incident on this section: $dE = E_\Sigma(t, \theta_T) \sin \theta \cdot dL_R$. Upon the impact of this field, the elementary current begins to run through the elementary section:

$$dI = \frac{dE}{(Z_A + Z_L)}$$

where $Z_A$ — an antenna radiation resistance; $Z_L$ — a load resistance.

The elementary currents that appear in each section run both toward load and toward dipole ends. An assumption was made above that the dipole ends are matched. For this reason, a part of these currents running toward the dipole ends emits secondary radiation and is absorbed. The currents running toward the load produce a voltage drop on it. As in the case of radiative radiation, let us take into account the signal delay in the receiver dipole wire and in space (relative to the point $M$):

$$t_3 = \left(\frac{L_R + L_R \cos \theta_R}{c}\right).$$

The total current running through the load of receiver dipole will be as follows, considering this delay:

$$I_\Sigma(t, \theta_T, \theta_R, L_T, L_R) = \int_0^L \left\{A_1 \left(t - \frac{L_T}{c}\right) - \frac{R - L_T \cos \theta_T}{c} + \frac{L_R}{c} + \frac{L_R \cos \theta_R}{c}\right\} dL_R.$$

The voltage on receiver dipole load will be:

$$U_\Sigma(t, \theta_T, \theta_R, L_T, L_R) = I_\Sigma(t, \theta_T, \theta_R, L_T, L_R) Z_L.$$

The total current running through the load of receiver dipole will be as follows, considering this delay:

$$I_\Sigma(t, \theta_T, \theta_R, L_T, L_R) = \int_0^L \left\{A_1 \left(t - \frac{L_T}{c}\right) - \frac{R - L_T \cos \theta_T}{c} + \frac{L_R}{c} + \frac{L_R \cos \theta_R}{c}\right\} dL_R.$$

The voltage on receiver dipole load will be:

$$U_\Sigma(t, \theta_T, \theta_R, L_T, L_R) = I_\Sigma(t, \theta_T, \theta_R, L_T, L_R) Z_L.$$

Figure 16 shows an example of calculations of this voltage, designated from here on as $U(t, \theta)$ for different angles $\theta_R$. An assumption was made during the calculations that the radiating dipole should be excited by a current pulse with the shape (5), the
duration of which should be $\tau$ ($c\tau \leq L_T$), while $L_R = L_T$. The angle $\theta_TX$ is assumed to be 300°, since at this angle the pulse shape of the field which is incident on the receiver dipole comes closest to the shape of the derivative of the pulse of the current that excites the radiating dipole (Figure 10), and, in this way, the condition $c\tau \leq L_R$ is met for the receiver dipole, too.

From Figure 16 it is evident that the voltage in the receiver dipole load is a sum of two pulses, each of which follows the shape of the incident field pulse. Those pulses are mirror images of one another, since directions of the currents running to the load from different dipole shoulders are opposite. The pulses are especially noticeably divided at small angles $\theta_R$, when the incident field induces currents in the shoulders with a considerable delay. A change in the voltage pulse shape on dipole load, when the angle of the incident field pulse on dipole changes, is analogous to the change in the field pulse shape of radiating dipole, when the angle of surveillance has been changed. Quite similar to it is the behavior of the receiver dipole DP over field (determinable as the dependence on-load voltage on angular coordinates): whilst the incident field pulse is travelling across receiver dipole, this DP moves in space, i.e. to say it becomes transient vs. time. As a result, the receiver dipole requires computations of the energy DP, which is obtained via averaging of the power received from each angular direction in the course of travel of the dipole-incident field pulse across its aperture. The energy DP describes the density distribution of energy flow received by dipole from space.

$$W_R(\theta) = \frac{1}{Z_L} \int_{-\infty}^{+\infty} U^2(t, \theta) dt.$$  

Normalized energy DP:

$$W_{RN} = \frac{W_R(\theta)}{W_{R\max}}.$$  

where

$$W_{R\max} = \frac{1}{Z_L} \int_{-\infty}^{+\infty} U^2_{\max}(t, \theta) dt.$$  

Figure 17 shows a family of normalized energy DPs used in the above example at different values of the ratio $L_R/c\tau$. At the value of $L_R/c\tau$, which is close to unity, the energy DP coincides with the DP of half-wave dipole. With an increasing ratio $L_R/c\tau$, the shape of DP changes: the DP maximum bifurcates and becomes narrower.

The DG of receiving antenna $D_R$ is defined as a ratio of energy density flow received by the antenna from the direction of DP maximum $W_{R\max}$, to energy density flow received from space:

$$D_R = \frac{W_{R\max}}{W_{R0}}.$$  

Considering that for the matched antenna the radiation resistance would be $R_{cm} = Z_L$, the total energy of the field that is incident on antenna through sphere with the radius $R$ should be:

$$W = \frac{R^2}{Z_L} \int_0^{2\pi} \int_0^{+\infty} U^2(\theta, \varphi, t) \sin \theta d\theta d\varphi dt.$$  

By dividing this energy by surface area of the sphere surrounding the antenna, we shall obtain the energy density flow, as received by the equivalent isotropic antenna:

$$W_{R0} = \frac{W}{4\pi R^2} = \frac{1}{4\pi Z_L} \int_0^{2\pi} \int_0^{+\infty} U^2(\theta, \varphi, t) \sin \theta d\theta d\varphi dt.$$  

The obtained expressions allow us to determine the
7. Peculiarities of UWB Waveform Detection

The highly informational UWB waveforms allow for target discrimination, even “radio portrait” of the target when the high resolution is obtained. However, the process of target detection precedes that of target discrimination. While analyzing the process, it should be borne in mind that the shape of target-reflected signal remains unknown.

The optimum detector of unknown signal in the Gaussian noise is the square-law one (energy detector) [19] (Figure 18). Having stacked up a packet of M such signals (Figure 19), one employs non-coherent accumulation of the results of square-law detection. The gating is performed at the moments of completion of single signal integration and completion of the packet stacking-up from M signals. The signals found in the packet are coherent, although of unknown shape. The non-coherent processing of coherent signals brings about, obviously, some losses. In order to avoid those losses, one has to change the order of stacking-up and square-law detection operations. The setup of this energy detector is given in Figure 20.

A theoretical validation of the optimum detection algorithm for signal with unknown shape was performed by V.S. Chernyak [22]. For synthesis of the detection algorithm, a priori data were used concerning the probing pulse repetition period \( T_r \).

The target is illuminated with impulses with the duration \( \tau \), the impulse rep rate being so high that during a relatively short time interval the target can be regarded as motionless, while the returned signal \( u_s(t) \) can be visualized as a packet consisting of M similar impulses (Figure 19):

\[
U_s(t) = \sum_{k=0}^{M-1} u_0(t - kT_r),
\]

where \( u_0(t) \) — a returned signal (impulse) of unknown shape with the duration \( T = n\tau \) and energy \( E_s \),

\[
\int_{-\infty}^{\infty} u_0^2(t) \, dt = E_s.
\]

Individual impulses do not overlap in the packet, i.e.

\[
\int_{-\infty}^{\infty} u_0(t - kT_r) \, u_0(t - mT_r) \, dt = \begin{cases} E_s, & k = m, \\ 0, & k \neq m. \end{cases}
\] (7)

The noise shall be assumed as being white, Gaussian, with the zero mean value. The locked-on signal is a mean value of the sum of signal and noise.

For the known signal the logarithm of relation of the likelihood functional is as follows:

\[
\ln \Lambda = \ln \left[ \frac{W_{s/n}[u(t)]}{W_n[u(t)]} \right] = \sum_{k=0}^{M-1} \int_{-\infty}^{\infty} u(t) \, u_0(t - kT_r) \, dt
\]

\[
- \frac{1}{2} \sum_{k=0}^{M-1} \sum_{l=0}^{M-1} \int_{-\infty}^{\infty} u_0(t - kT_r) \, u_0(t - lT_r) \, dt. \quad (8)
\]

Here \( u(t) \) — a receivable realization of the signal-noise sum or just noise with the duration

\[
T_0 > (M - 1)T_r + T.
\]

Taking (7) into account, a simplification can be made (8):

\[
\ln \Lambda = \sum_{k=0}^{M-1} \int_{-\infty}^{\infty} u(t) \, u_0(t - kT_r) \, dt
\]

\[
- \frac{1}{2} \sum_{k=0}^{M-1} \int_{-\infty}^{\infty} u_0^2(t - kT_r) \, dt.
\]

Although \( u_0(t) \) is unknown (but not random), we shall employ the adaptive approach. The essence of the approach is to use an algorithm for a signal that is optimal at the known parameters, into which, replacing the unknown parameters, their maximum
likelihood evaluations are substituted [21]. In this event, one should consider the signal, as a whole, as a function of time, rather than individual parameters of the signal [22]. Let us consider the likelihood functional logarithm of the signal:

\[ \mathcal{R}_1 = -\frac{1}{2} \int_{-\infty}^{\infty} \left[ u(t) - \sum_{k=0}^{M-1} u_0 \left( t - kT_r \right) \right]^2 dt. \]

The maximum \( \mathcal{R}_1 \) as a function of \( \sum_{k=0}^{M-1} u_0 \left( t - kT_r \right) \) corresponds to the minimum of the expression found in the square brackets. Since duration of the receivable realization \( u_0(t) \) for the signal-noise sum embraces all incoming signals, then the minimum of the expression in the square brackets is fulfilled at the condition when, in each time interval where there is the signal, it is equal to the receivable realization:

\[ \hat{u}_0 \left( t - kT_r \right) = u(t) \text{ when } t \in (kT_r, kT_r + T). \]

Let us replace the variable in (7): \( t_1 = t - kT_r \). Then

\[ \hat{u}_0(t_1) = u(t_1 + kT_r) \text{ when } t_1 \in (0, T). \]

The evaluations of the same function \( u_0(t_1) \) obtained in the time intervals that are located over the repetition period \( T_r \), as regards one another, are statistically independent, since the Gaussian noise in those intervals is independent. This means that the optimum evaluation \( \hat{u}_0(t_1), t_1 \in (0, T) \) that is rendered accurate over \( M \) measurements looks like as follows:

\[ \overline{u_0(t_1)} = \frac{1}{M} \sum_{k=0}^{M-1} u(t_1 + kT_r) \text{ when } t_1 \in (0, T). \quad (9) \]

By substituting in (12) the variables \( t_1 = t - kT_r \), we shall obtain

\[ \ln \Lambda = \sum_{k=0}^{M-1} \int_{0}^{T} u(t_1 + kT_r)u_0(t_1)dt_1 \]

\[ - \frac{1}{2} \sum_{k=0}^{M-1} \overline{u_0^2(t_1)}dt_1. \quad (10) \]

By replacing \( u_0(t_1) \) with the obtained evaluation of \( \overline{u_0(t)} \) from (9), while skipping the index from \( t_1 \) and non-essential factor \( 1/M \), some simple transforms having been made, we obtain from (10) the optimum (according to the Neumann-Pierson criterion) adaptive detection algorithm in the form of:

\[ \mathcal{R} = U_{exit} = \left[ \sum_{k=1}^{M-1} u(t_kT_r) \right]^2 dt \geq U_{threshold}, \quad (11) \]

where \( U_{threshold} \) - a detection threshold.

Thus, the optimum algorithm boils down to the summation of the acceptable realization lengths (\( T \)-long each) in such time intervals where signals are expected; to computation of the energy of this sum and comparison of produced energy with the threshold that is determinable via the assigned false alarm probability. This is the known “energy” detector, with which one accounts for not only the energy of each impulse of the packet, but also for the energy of correlations between the impulses.

At a rather swift target movement relative to the radar, the impulses of the packet \( u_{r}(t) \) will begin to differ from one another. In this case, one can consider at least two adjacent impulses to be similar. For this reason, we shall now consider a case, which is important for applied purposes, of processing two impulses with a small repetition period \( (M = 2) \). The setup of the detector in this case is presented in Figure 21. The optimum algorithm (11) at \( M = 2 \) can be represented as:

\[ \mathcal{R} = U_{output} = \int_{0}^{T} u^2(t)dt + \int_{0}^{T} u^2(t + T_r)dt \]

\[ 2 \int_{0}^{T} u(t)u(t + T_r)dt \leq U_{threshold}. \quad (12) \]

The first and second integrals in (12) describe the energy computation algorithm of such signals that are received in two adjacent periods, the third integral determining the algorithm of their reciprocal correlation processing.

From (12) it is obvious that the detector under consideration can be represented (and realized) as three sub-optimal detectors operating jointly (Figure 22). The first two ones are well-known energy detectors of signals expected in the first and second period. The third one, dubbed as alternating-period correlation detector (APCD), was described earlier in reference [23]\(^5\). Schematics shown in Figures 21 and 22 permit to bring out the particulars of the optimal detector:

a) The noises coming from adjacent repetition periods to inputs of the detectors are not dependent; the integrator input receives either the square of the sum of independent segments of the normal processes (Figure 21), or the sum of two squares and product of these segments (Figure 22) corresponding to adjacent periods; the probability distribution at the integrator input substantially differs from the normal one in both cases.

b) The probability distribution at the integrator output approaches the normal one, as the

\(^5\)In reference [23] the alternating-period correlation detector is wrongly cited as optimal.
integration time increases; this time is determined not by the duration of radiated signal \( t \), but by the duration of signal returned from target \( T = n\tau \); in this way, the radial target length \( L = nce\tau /2 \) determines the integration time, level of normalization of process distribution at the integrator output, noise dissipation and, as a result, the detector threshold level.

Let us compare the efficiency of the optimal detector (12) and three sub-optimal ones.

Figure 23 presents the detection characteristics for the optimum algorithm (curve 2), APCD (curve 3) and sum of energies (curve 4). For the sake of comparison, the Figure shows the detection characteristic of the “classical” detector with precisely known (determined) signal (curve 1). At \( M = 2 \), curve 3 coincides with curve 4. The false alarm probability is \( f_a = 10^{-3} \). The characteristics are constructed for target with radial length upon which 16 resolution elements are laid. With such storage during the integration process, the output statistics of both optimal and sub-optimal detectors can be considered as Gaussian values. One can gather from those Figures that the ignorance of the signal shape leads to considerable losses. These losses are attributed to the absence in the receiver of an “outnoised” sample of the signal detected.

From curves shown in Figure 23 one can gather that the sub-optimal detectors’ efficiency is just slightly worse than that of the optimal one. That is why, in certain cases, a sufficiently unsophisticated APCD algorithm may find its way to practical applications.

Schemes shown in Figures 21 and 22 are optimum for detection of a motionless target with the fixed radial length. This hypothetical case was considered as standard reference enabling to evaluate the efficiency of detection under various conditions. To a certain extent, this approach is analogous to the problem from the classic theory of detection of a signal with exhaustively known parameters (determinant). In reality, when the target velocity and its extent are not known a priori, a multi-channel geometry of the system must be in order both in signal delay and in integration interval duration. The thus-obtained optimal and sub-optimal algorithms are realizable in digital form.

8. Determination of Target EDS while Employing UWB Waveforms

During target illumination with narrowband signal, when the radial target length \( L \) is considerably smaller than the spatial extent of the signal \( \tau \) and reflections from diverse brilliant specks from the target overlap in space, producing the added-up signal in the receiving point, in order to determine the EDS of a target located in the antenna far zone, the following well-
known expressions is used:

\[ \sigma = 4\pi r^2 E_2^2 / E_1^2, \]  

(13)

where \( r \) – a distance to target; \( \sigma_1 \) – an amplitude value of the probing signal field strength in the point of target; \( \sigma_2 \) – an amplitude value of field strength of a signal dissipated by target in the receiving point.

During target illumination with simple UWB waveform \((L \gg c \tau)\) the returns from brilliant spots of the target become divided in space, while the signal in the receiving point is represented as a sequence of pulses of variable amplitude, shape and polarity, shifted one relative to another to arbitrary time lengths (“target portrait”, see, for example, in Figure 5). The value of the amplitude \( \sigma \) in the receiving point in this event becomes vague [24–26].

To determine the EDS in this case [27], a notion of “generalized” EDS was introduced:

\[ \sigma = 2\pi \sigma_1 W_2 / W_1, \]  

(14)

where \( W_1 = \int \Pi_1(t)dt \) – an energy density flow of radar probing signal in the point of target; \( W_2 = \int \Pi_2(t)dt \) – an energy density flow of on-target dissipated signal in the receiving point; \( \Pi_1(t) \) – an instantaneous value of the Poynting vector of probing signal which exists during the time \( t_1 \) in the point of target; \( \Pi_2(t) \) – an instantaneous value of the Poynting vector of dissipated signal which exists during the time \( t_2 \) in the receiving point.

A similar situation with determination of the target EDS arises in those cases when UWB waveforms with chirp modulation are in use. In this event, the amplitude value of the uncompressed signal field in the receiving point \( E_2 \) and signal amplitude at the matched filter output, which is compared with threshold, differ (in relative units) by times the compression ratio. That is why using of the expression (13) to determine the target EDS would not be right. Since the signal energy prior to and after the compression (without consideration of the losses to processing) remains the same, the expression (14) enables to determine the target EDS on this score.

Now, we shall compare the values of the target EDS obtained during illumination of the target with narrowband and UWB signals. For that purpose, we shall consider as being target an ultimate group secondary radiator consisting of two reflectors (“dumbbells”). An assumption is made to the effect that the reflectors do not interfere with each other.

For the narrowband signal \((L \ll c \tau)\), this case is very well known [28]. The pulse of the field illuminating target has the high-frequency filling \( e_1(t) = E_1 \cos \omega t \). We shall designate the reflector-to-radar distance as \( r_{1,2} \) \((r_1 - r_2 = \Delta l \ll c \tau)\), while the moments of arrival of the reflected fields to the receiving point would be written as \( t_{1,2} = 2r_{1,2}/c \). The field dissipated by the secondary radiator in the receiving point would be as follows (the index in parentheses stands for the reflector number):

\[ e_2(t) = E_{2(1)} \cos \omega (t - t_1) + E_{2(2)} \cos \omega (t - t_2) = e_2 \cos (\omega t - \varphi), \]

where \( \varphi = \omega (t_1 - t_2) = (2\pi/\lambda) 2(r_1 - r_2) = 4\pi \Delta l/\lambda \).

The total field amplitude in the receiving point \( E_2 \):

\[ E_2^2 = E_{2(1)}^2 + E_{2(2)}^2 + 2E_{2(1)} E_{2(2)} \cos \varphi. \]

The target EDS in accordance with (1) now would be:

\[ \sigma_{\Sigma 2} = 4\pi r^2 E_2^2 / E_1^2 = \sigma_1 + \sigma_2 + 2\sqrt{\sigma_1 \sigma_2} \cos \varphi. \]

At the values of \( \varphi = 0 \) and \( \sigma_1 = \sigma_2 = \sigma \), that of the EDS will be maximal \( \sigma_{\Sigma 2} = 4\sigma \). Correspondingly, with \( N \) reflectors that have equal EDS \( \sigma \), if the reflected fields are synchronous in phase, the EDS of the group radiator will be \( \sigma_{\Sigma N} = N^2 \sigma \). In the general case, the phase \( \varphi \) is a random quantity that assumes any values from 0 to \( 2\pi \). In this case, the EDS is determined as mean value:

\[ \sigma_{\Sigma N} = N \sigma. \]  

(15)

When using UWB waveform \((L \gg \tau)\), the considered secondary radiator has a field pulse incident upon it with the duration \( \tau \) and energy density flow \( \Pi_1(t) \). When \( r_1 - r_2 = \Delta l > c \tau \), then from the two reflectors in the receiving point come two non-overlapping field pulses that have, in the general case, different energy density flows \( \Pi_{2(1)}(t) \) and \( \Pi_{2(2)}(t) \) and different durations \( t_1 \) and \( t_2 \).

In this event, the EDS of each secondary radiator is determined according to the formula (14):

\[ \sigma_{1,2} = 4\pi r^2 W_{2(1),2(2)} / W_1, \]
where \( W_1 = \int \Pi_1(t)dt \), \( W_2(1) = \int \Pi_{2(1)}(t)dt \), \( W_2(2) = \int \Pi_{2(2)}(t)dt \) — an energy density flow of the radar probing signal in the point of target and energy density flows of signals obtained in the receiving point from the first reflector at the moment \( t_1 \) and from the second reflector at the moment \( t_2 \).

Quite obviously, in this case of the EDS of the secondary radiator, the EDS of such reflector will be determined from which the energy flow has come in at the greatest density value \( \Pi_2(t) \). The energy flow arriving in the receiving point from the smaller reflector appears to miss being used, which can be regarded as a loss. With an increasing number of reflectors, those losses will increase, as well.

In order to avoid these losses and obtain the maximum EDS value of target during its illumination with UWB signal, it is necessary to "gather together", over one time interval, all of the energy received from different target reflectors over different time intervals. It can be done at the condition that radar should employ the above optimal signal processing.

Let us assume, as before, that the group secondary radiator reflectors should be equal. By substituting in (12) each two reflected pulses in each period (the index denotes the number of reflector), we shall obtain:

\[
U_{\text{output}} = \int_0^T u_1^2(t)dt + \int_0^T u_2^2(t)dt + \int_0^T u_1^2(t + T_r)dt + \int_0^T u_2^2(t + T_r)dt
\]

\[
+ 2 \int_0^T u_1(t)u_2(t)dt
\]

\[
= W_2(1,1) + W_2(1,2) + W_2(2,1) + W_2(2,2)
\]

\[
+ 2W_2(1,1-2,1) + 2W_2(1,2-2,2).
\]

If the energy of radar probing signal in the point of target is equal to \( W_1 \), then by substituting the obtained energy values in (14), we shall obtain the following:

\[
\sigma_{\Sigma^2} = 4\pi r^2(W_2(1,1) + W_2(1,2) + W_2(2,1) + W_2(2,2))\sigma
\]

\[
+ 2W_2(1,1-2,1) + 2W_2(1,2-2,2)/W_1
\]

\[
= \sigma_1 + \sigma_2 + \sigma_1 + 2\sigma_2 - 2\sigma_2 - 1.
\]

With the equality \( \sigma_1 = \sigma_2 = \sigma_1 - 2 = \sigma_2 - 1 = \sigma \), we shall produce \( \sigma_{\Sigma^2} = 8\sigma \). This EDS value has been obtained taking into account the target-reflected energy in two periods of probing. In one period of probing the ratio of energy in the receiving point to that in the point of target comes out with the EDS value \( \sigma_{\Sigma^2} = 4\sigma \). Respectively, with \( N \) secondary radiators that have the same EDS \( \sigma \), the total EDS of the group radiator will be as follows:

\[
\sigma_{\Sigma^2} = N^2\sigma.
\]  

While comparing the EDS values obtained after illumination of one and the same target consisting of \( N \) similar reflectors with UWB (17) and narrowband (15) signals (doing the matched processing of those signals in the receiving point), we can now see that the following inequality is fulfilled \( \sigma_{\Sigma^2} \geq \sigma_{\Sigma^N} \), that is \( \sigma_{\text{UWB}} \geq \sigma_{\text{AB}} \).

In the general case, with arbitrary EDS values of individual target reflectors the total target EDS value.

\[
\sigma_{\Sigma^2} = N^2\sigma.
\]
9. Passive Interference

Immunity

In UWB radar, immunity to natural and manmade passive interference has peculiar features of its own. We shall now consider those features by providing an example of employment in this kind of radar of a system of alternating-period compensation (SAPC) that takes measurements of variations of the phase difference of oscillations reflected from moving target and from clutter over one (or several) repetition period [29].

The short UWB impulse creates a very small impulse volume of radar beam over distance, which drastically reduces the EDS interference and simplifies target surveillance against its background (Fig. 24). However, the reduction of the radar impulse volume over distance does more than simply to decrease the power of interference. As a rule, a clutter producing the interference is movable (clouds, ground vegetation, artificial aerosols). Owing to the motion of the clutter inside the impulse volume during the repetition period, a part of this interference reaches outside this volume, while a part of new clutter comes in (those parts of interference are shown tentatively in Figure 24 as dark “belts” on the fringes of impulse volume). These changes act to disrupt the correlation of signals reflected from the clutter and received in contiguous repetition periods. The resulting violation of interpulse correlation acts to reduce the interference suppression factor in the SAPC system and worsens its efficiency. When the radar impulse volume is large enough (for example, at $\tau = 1$ microsecond, its reach over distance is 300 m), the shifting part of the clutter constitutes a small portion out of the entire impulse volume and does not exercise a considerable influence on the violation of inter-pulse correlation of the clutter. However, when the impulse volume becomes small (at $\tau = 1$ ns its reach over distance is 30 cm), this part of the clutter can constitute a considerable portion of the entire impulse volume, worsening substantially the inter-pulse correlation of signals returned from the clutter and decreasing the SNR at the output of the SAPC system.

Figure 25 presents the plots of SNR $Q$ relationships at the output of twofold SAPC (SAPC-2) vs. UWB waveform duration t, normalized to the average signal spectrum frequency period $T_{av}$, at different values of the repetition period $T_r$.

The plots show two extremums, the first of which (maximum) coincides with the impulse duration $t = 0.5_{av}$. Upon variation of the impulse duration from $t = 0.5_{av}$ to $t = 0.5_{av}$ the efficiency of the SAPC increases due to the reduction of impulse volume of the radar. At $\tau < 0.5_{av}$ an intensive clutter de-correlation influence begins, with the SAPC efficiency going down. With further decreasing of t the SAPC ceases to exert any influence on noise immunity on account of the total de-correlation of the clutter and only the first factor remains there, i.e. the impulse volume reduction. The noise level comes down again and $Q$ increases. In this way, the position of the second extremum (minimum) corresponds to the total clutter de-correlation and absence of velocity selection.

Note that, with an increasing impulse repetition
Fig. 26. Dependence of UWB radar impulse power on detection range.

Fig. 27. Allowable levels of UWB systems radiations for HF UWB systems.

period, the position of the second extremum shifts to the left in the direction of the shorter durations $\tau$. This means that, with a decreasing impulse repetition rate, the total clutter de-correlation comes about with shorter signal durations. Within the limits of non-fluctuating passive interference, the impulse duration reduction does not affect SAPC. Thus, the use of SAPC in UWB radar is the more efficient, the lesser is the clutter de-correlation and the higher is the repetition rate.

10. Range Equation. Particulars of Applications

Conventionally, the range equation connects the narrowband signal power coming to the input of receiver threshold device and the power of this signal emitted by transmitter. The signal energy determining the characteristics of radar detection is not, as a rule, included explicitly in the range equation. In UWB radar, the waveform variation during radar surveillance does not permit to employ such parameter as signal power. For this reason, the radar range determination $R$ is made using the energy parameters:

$$ R = \sqrt[4]{\frac{W D_T \sigma_{\Sigma} D_R}{(4\pi)^2 \rho q N_0} \tau}, $$

where: $W$ – an emitted signal energy; $D_T$ – the energy DG of transmitting antenna; $D_R$ – the energy DG of receiving antenna; $\sigma_{\Sigma}$ – the energy EDS of target; $\rho$ – losses in all radar systems; $q$ – the threshold SNR; $N_0$ – a noise power spectral density.

The high-resolution data content of UWB radars is achieved via employing the impulses of nano- and picosecond durations. If plain short waveforms are emitted, similar to those shown in Figure 5, then the energy needed for target detection can be produced only due to a high peak incident power $P_{\text{peak}}$. Let us assess the order of this power in a specific example. Considering that $W = P_{\text{peak}} \tau$, we shall determine the value of $P_{\text{peak}}$ using the range equation parameters:

$$ P_{\text{peak}} = \frac{(4\pi)^2 \rho q N_0 R^4}{D_T \sigma_{\Sigma} D_R \tau}. $$

The results of computations of the peak UWB generator power vs. radar range for several values of the impulse duration $\tau$ are given in Figure 26. The following numeric values were used in the computations: the energy DG of the antennas $D_T = D_R \approx 1000$ (circular mouth $\approx 3$ meters in diameter), $\sigma_{\Sigma} = 0.1$ m$^2$, the probabilities of true detection $P_D = 0.9$ and false alarm $P_F = 10^{-6}$.

The above example indicates that the use of plain UWB waveforms for target detection and discrimination over long ranges requires the employment of generators of nanosecond impulses at the peak power of several to tens giga-watts. These generators are unique devices warranting incorporation of elements with high electric strength, not being environmentally friendly both due to parasitic x-ray radiation and direct electromagnetic radiation.

A decreasing of the peak power in high-power UWB radars can be achieved in the conventional way: transition to complex, large-base signals and their subsequent matched filtration. In order to form and process these UWB waveforms with high correlation properties, the picosecond accuracy is required to exercise control over temporal position of the coded impulses. Acquiring of the necessary potential in these radars without involving the use of giga-watt generators can also be done, using active pulsed
antenna arrays. Yet, in this case as well, a high temporal stability of the system is needed for precise control over the array oscillators. However, the use in UWB radars of complex coded UWB waveforms with their subsequent matched processing and/or of active pulsed antenna arrays slots those radars in the category of sophisticated and costly systems, the technology of which is immature to date.

For all that, not even the complexity of engineering solutions is the main hindrance to creation of relatively high-power UWB radars.

The high-resolution data content of UWB systems mandates the use of broad portions of the bandwidth. On this account, the UWB radars may interfere with many other radio systems sharing the same portions of the spectrum, including such vitally important systems as the satellite navigational systems (GPS, GLONASS, Galileo), satellite mayday systems, air-traffic control systems (ATC), etc. For that matter, the electromagnetic compatibility of the UWB radars with other radio systems is of crucial importance for their prospective development.

Concerning low-power UWB radars, for the first time in the history of radio engineering, a normative document has been compiled permitting the simultaneous operation of UWB and narrowband radio systems across the same frequency range. This concerns the Regulations of US Federal Communications Commission (FCC) that went into effect in April of 2002, with supplements effected as of 2003 and 2004 [30,31]. This document represents a result of workout (relative to UWB radiative radiation systems) of Part 15, Article 47 of US Federal Regulations Code [32]. It formulates restrictions for UWB radiation levels in different frequency ranges, the so-called “mask” (an instance of this “mask” is given in Figure 27). Meeting those requirements, as well as observing some other restrictions, operation of UWB systems is authorized without licensing. At present, this document is used as reference in many countries all over the world.

In the meantime, the electromagnetic compatibility of UWB radars over long operation ranges, for example, for satellite-borne radars, with other radio systems operating in the same portions of the frequency spectrum, should, in all evidence, be very problematic.

All told, the most realistic way of using UWB technologies in radiolocation at present would be production of relatively low-power radars for operation over distances of several to tens of meters. Those radars are finding their applications in diverse areas of human activities and, as a consequence, have immense marketing prospects.

11. Some of Examples of UWB Radars Created by the Russian UWB Group of Moscow Aviation Institute

A simplified block scheme most of the radars’ examples shown below is represented at fig. 28. The oscillator with controlled pulse repetition frequency produces rectangular pulses with frequency of 0.05–30 MHz.

The transmitter consists of the short pulse shaper. The pulses from shaper output are delivered to transmitting antenna and make shock excitation of it. Transmitting antenna radiates short RF pulses.

The electromagnetic field’s pulses radiated are reflected from moving object. Here the modulation of pulse repetition frequency arises. The modulation percentage depends of the velocity and amplitude of target’s motion.

The radar works in conditions of high level of passive noise — the signals, reflected from stationary objects, which will have large amplitude and will disguise useful signal.

Time slots, opening the receiver at the moment of input of signal reflected from object at distance defined are formed in receiving path to eliminate interfering pulses. This task in radar design is executed.
by time discriminator, being gated. It consists of fast-acting electronic switches. The switching time is in order of 200–300 picoseconds. The switches connect receiving antenna to UWB amplifier at the moments of signals’ input. These moments are defined by delay magnitude of control signal at software-controlled delay line. All the rest time the receiver is shut. The signals received at time slots are detected and amplified in integrating amplifier and the signal, carrying data of target motion is selected at its output.

The gating unit is consists of software-controlled delay line and the shaper of short pulse. Time delay set by microprocessor-controlled unit defines the distance to object. Time constant of integration of integrating amplifier is chosen in dependence of the bandwidth of desired signal (dynamic characteristics of motion of object examined). For example under measuring of parameters of person vital activity the bandwidth of desired signal is near 40–50 Hz, that corresponds to accumulation of 10~30 thousands of pulses approximately. The accumulation permits to decrease average radiated power of transmitter and increase signal-to-noise ratio at the input of amplifier.

The selected and amplified low-frequency signal enters to analog-digital converter (ADC). The microprocessor-controlled unit directs the work of radar on given algorithms, monitors the state of major units and modules and provides data output for further digital processing in computer. The selection of moving targets, fast Fourier transform and digital filtration are software-programmable at the computer.

Physically all radars are built as modular hardware. All modules are implemented in shields, eliminated interference of each other. Antennas connection is carried out directly to output connectors of radar’s receiver and transmitter.

All examples of radars are created and tested with respect to principles pointed.

First example

Figure 29 demonstrates external and internal (printed circuit board) views of this UWB radar example.

The major specifications of example are given below:

- Range 0.1–3 m;
- Pulse power 0.4 W;
- Average power 240 µW;
- Repetition frequency 2 MHz;
- Duration of radiated radio pulses 4 ns.

First example is intended for medical researches. Figures 30 demonstrates the results of remote measuring physiological parameters. At the Figure 30(a) we demonstrate the time diagrams of summarized signals corresponding heart and respiratory beats, which were detected by the UWB radar. The amplitude of radar output signals is directly proportional to the amplitude of thorax and heat beats. At the Figure 30(b) shown time diagram
corresponding only heart beat signals detected by the radar when a patient holds his breathing. On top of Figures 30(a) and 30(b) is shown patient’s control electrocardiograms registered at the same time by a medical electrocardiograph.

To appreciate the accuracy of radar measurements and quality of information concerning patient’s heart activity obtained using UWB radar, we compared data collected with the radar and a medical cardiograph. The comparison was led in instantaneous (from beat to beat) variations of systole period (what is known as variability of cardiac rhythm). Fig. 31 demonstrates the measurement data depending on the number of a beat, which were obtained simultaneously with using radar (red line) and a cardiograph (blue line). The coefficient of correlation between these measurements was calculated. In this experiment, we have a correlation coefficient $\approx 0.91$.

**Second example**

Fig. 32 demonstrates views of the second UWB radar example.

Second example is intended for measurements of pulse. At Fig. 33 the output data of this radar are shown.

**Third example**

Fig. 34 demonstrates views of the third example of the UWB radar in process of the measurement of speed of
a pulse’s wave. This example consists of two radars - sensors which measure pulse on different vessels. At fig. 35 we show the output data of two sensors of the third example.

Speed of a pulse’s wave is the important diagnostic factor for cardiac and vascular diseases. Third prototype can simultaneously measure variability of an cardiac rhythm. It is the universal craniological tool.

**Fourth example**

This example of the UWB radar is intended for measurement of speed and a position of railway cars on a shunting-yard. The major specifications of example are given below:

- Range: 300 m
- Controllable speeds of cars: 10 – 0.2 m/s
- Pulse power: 10 W
- Average power: 8.4 mW
- Resolution on range: 41 cm
- Step of installation of a strobe of range: 3.75 cm

Fig. 36 demonstrates views of the fourth example of the UWB radar on the shunting-yard.

At the fig. 37 the example of an arrangement of cars and signals received from radar is shown. Signals from many radars located on towers is transferred to the control centre of the shunting-yard. After processing this information is used for formation of the trains.

**Fifth example**

The same radar has been used for detection of people in a forest. On fig. 38 we see a place of measurements and on fig.39 – the scheme of measurements. Detection was made on two persons moving on distance 50 meters from edge of a forest. Results of measurements are shown on fig. 40.

**Sixth example**

The radars described above have been tested for the detection of alive people concealed behind nontransparent barriers.

Using the prototype of the UWB radar, we have performed experiments on detection of moving and motionless people concealed behind optically nontransparent barriers. The radar was located at a distance of about 1m at one side of a brick wall of width 0.45 m. A person to be detected stood behind the wall.

At the beginning of experiment, a person was moving around the room, and then he became still and then began moving again. The time diagram of radar output signal is demonstrated in fig. 41(a). One can easily observe the signals corresponding person’s movement (large amplitude signals) and signals corresponding thorax movements of motionless person when respiration (periodical signal in the rectangular frame). Fig. 41(b) illustrates signal’s amplitude-frequency spectrum. The maximum correspondence a respiration frequency is clearly noticeable.

**12. Conclusions**

The Author is well aware of the inadequacy of his attempt to provide within the space of this article a certain generalization of the peculiar features of operation and construction of UWB radio systems. He finds his solace in a remark by the founder of this direction in radio engineering Dr. H. Harnuth made in 1981 [2], which has not lost its relevance to
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Fig. 37. Train and the signal reflected from it.

Fig. 40. Detection of two persons moving on distance 50 meters from edge of a forest.

Fig. 41.

(a) Radar output signal when recording person’s movement behind the barrier.

(b) Amplitude-frequency spectrum of radar output signals.

date, “A relative frequency band \( \eta \) ... may acquire any value in the interval \( 0 < \eta < 1 \). Our advanced engineering is based on the theory for the limiting case \( \eta \to 0 \). Both theory and technology, applicable across the entire interval of values \( 0 < \eta < 1 \), have to be more general and more complex than the theory and technology pertaining only to the limiting case \( \eta > 0 \). “Handbook of Radiolocation” by M. Skolnik, which concentrates on terse explanation of the problems related to the case \( \eta > 0 \), gives an idea about what to expect for the general case \( 0 < \eta < 1 \).

The compilation of “Handbook of Radiolocation” took forty years of the progress in radiolocation, including several technological revolutions. This fact is telltale about how much water shall have passed before a similar handbook sees the light for the general case”.

25 years since, the insufficiency of theoretical base for advancement of UWB technology and engineering both on the systems plane and creation of specific devices, especially antenna systems, remains the principal obstacle for their further development. This allows the Author to express Hope for arousing the interest of those specialists, who work in this area, in his stated point-of-view, all the more so, because the novel capabilities that UWB radio systems provide for a better quality and increased quantity of transmittable information, attracting to the R&D on them, in many countries the world over, ever-growing expenditures and intellectual resources.
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