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A CONCEPT OF A PASSIVE RADAR WITH QUADRATURE MICROWAVE PHASE DISCRIMINATORS

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Abstract

Passive radar does not have its own emitter. It uses so-called signals of opportunity emitted by non-cooperative illuminators. During the detection of reflected signals, a direct signal from a non-cooperative emitter is used as the reference signal. Detection of electromagnetic echoes is, in present day radars, performed by finding the maximum of the cross ambiguity function. This function is based on the multiplication of the received signal and the reference signal. Detection of echoes by means of a quadrature microwave phase discriminator QMPD was proposed in the work as an alternative solution for ambiguity function evaluation. This discriminator carries out vectorial summing of the received and the reference signals. The summing operations in QMPD are carried out with the aid of microwave elements and without the use of expensive digital signal processors. Definitions of the phase and phase difference of the so-called simple signals and noise signals were described. A proposal of a passive radar equipped with several independent quadrature microwave phase discriminators was presented. Ideas of algorithms of object detection and of the distance-to-object estimation designed for this radar have been also sketched.

Keywords: passive radar, instantaneous frequency measurement, IFM, microwave phase discriminator, phase of noise signal.

1. Introduction

An electric signal, including microwave signals, can be described by the well-known function of time in the form of:

$$S(t) = A \cdot \sin(2 \cdot \pi \cdot f \cdot t + \varphi) = A \cdot \sin(\omega \cdot t + \varphi), \tag{1}$$

where:
A – signal amplitude, f – signal frequency, ω – angular frequency (pulsation) of the signal, t – time, φ – signal initial phase.

The graphic representation of such a signal can be a vector with length A rotating with an angular frequency ω. The projection of the end of this vector on a straight line lying in the plane of orthogonal coordinates X0Y and parallel to the Y axis sketches the time figure of the signal $S(t)$.

If the length A and angular frequency ω are invariable in time then we have to deal with a so-called simple signal or, in other words, a monochromatic one. The time course of such a signal is the sinusoidal line shown in Fig. 1.

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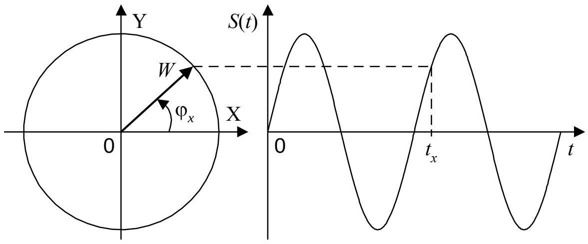

Fig. 1. Example of a monochromatic signal created by the end of rotating vector W (initial phase $\varphi = 0$).

Applying this model, the momentary phase φ_x which is the angle of the rotating vector orientation in the orthogonal co-ordinates XY at the point in time t_x, can be one of parameters characterizing the signal at any point in time t_x. According to this definition, the momentary phase φ_x achieves the values in the range $\pm\pi$ radians and is distinguished from the angle Φ_x which the vector W passed during the time t_x, this means from the beginning of the observation of the phenomenon ($t = 0$) until the moment $t = t_x$.

$$\Phi_x = \omega \cdot t_x, \tag{2}$$

$$\varphi_x = \Phi_x + \varphi - n \cdot 2 \cdot \pi. \tag{3}$$

The coefficient n in (3) is calculated with the use of the function *entier E* according to the expression (4). This takes only integer values and this is the number of complete turns which the vector W completed during time $t = t_x$.

$$n = E\left(\frac{\Phi_x}{2 \cdot \pi}\right). \tag{4}$$

In compliance to (2) and (3), because the angle Φ_x is a function of time so the momentary phase φ_x is dependent on time as well.

A signal with frequency modulation with a deviation Δf can be described, for example, by the relationship (5), in which the angular speed ω of the vector W rotation is a function of time.

$$S(t) = A \cdot \sin[\omega(t) \cdot t + \varphi] \quad \text{for} \quad 0 \le \omega(t) \le \omega_{\max}. \tag{5}$$

The momentary angular frequency $\omega(t)$ in the formula (5) is a non-negative number affecting the values in the range $\pm\, 2 \cdot \pi \cdot \Delta f$ around the angular frequency ω_o corresponding to the carrier frequency f_o of the signal.

A peculiar example of a signal applied in radiolocation is a noise type signal $N(t)$ with limited band ΔB and with finite energy. Apart from the Fourier distribution, at least two variants of the rotating vector model can be applied to its description. The first of them is expressed by the relationship (6). The shape of such a noise signal is presented in Fig. 2.

$$N(t) = A \cdot \sin[\omega(t) \cdot t + \varphi] \quad \text{for} \quad -\omega_{\max} \le \omega(t) \le \omega_{\max}. \tag{6}$$

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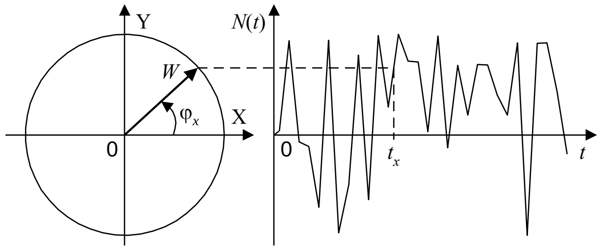

Fig. 2. Example of a noise signal created by the end of rotating vector *W* (initial phase φ = 0).

According to this model, the length of vector *W* is stationary and it is equal to the maximum value that the noise *N*(*t*) can have. However, the rotation of the vector can be clockwise or anticlockwise. This denotes that the angular frequency ω is a function of time and can have both positive values as well as negative and even zero. The extreme values of pulsation are determined by the width of the band ΔB.

One can also use the model expressed with formula (7), in which both the angular speed ω and the length *A* of the vector *W* are changing in time.

$$N(t) = A(t) \cdot \sin[\omega(t) \cdot t + \varphi] \quad \text{for} \quad -\omega_{max} \le \omega(t) \le \omega_{max} . \tag{7}$$

It should be noted that modeling the same noise signal *N*(*t*), angular frequencies $\omega(t)$ in expressions (6) and (7) will be described by different expressions.

In accordance with the abovementioned considerations, regardless of the signal structure and of the kind of models used for their description, every signal can be characterized with the use of a momentary phase. Therefore, one can compare phases φ_x of both simple signals (monochromatic) and so-called complex signals, and in this the noise signals as well. The momentary phase difference φ_w is the result of such a comparison.

$$\varphi_w = \varphi_{x1} - \varphi_{x2} , \tag{8}$$

where:
φ_{x1} – signal 1 momentary phase, φ_{x2} – signal 2 momentary phase.

If, for instance, simple signals with equal frequencies are compared, their phase difference φ_w is proportional, among others, to their delay but this will not change in the time domain. However, the phase difference of simple signals but with various frequencies, will always be changing in time without regard to their mutual location on the time axis.

The phase difference of exactly equal noise signals will also be time independent. If, however, the structures of noise signals are identical, but these signals are mutually delayed, then their phase difference φ_w varies in time. When the time shift of these signals will be larger then the frequency of these changes will also be larger. These features make it more effective to compare signals and to perform their identification even in a dense environment. Moreover, these features can be utilized in the process of target detection performed in radiolocation, especially in this using noise-like signals.
A standard active radar emits radio signals that are reflected off the target. The reflected signals are often called echoes. These echoes after being received are used to locate the targets in direction and at a range. The critical drawback is that the active radar should be equipped with its own transmitter. A passive radar system has no emitter and it does not transmit any signals [1-3]. It only receives electromagnetic (EM) signals and operates

independently without direct synchronization from the so called non-cooperative emitter or illuminator of opportunity.

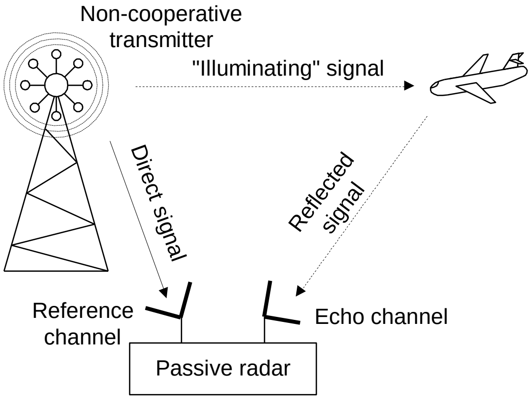

Fig. 3. Passive radar system scheme.

A passive radar has two or more receiving channels. One of them is used for capturing the so-called direct signal (Fig. 3). This signal is then utilized as a reference signal. Object detection and localization is performed by comparison of the reference signal and echo signal. Radio transmission systems suitable for passive radar functioning, are, for instance: Digital Audio Broadcast (DAB), Digital Video Broadcast (DVB) [4-6], Digital Communication System (DCS) and Global System for Mobile communication (GSM) [1, 7]. The structure of the signals emitted by these systems is similar to noise with a middle width band MWB.

Radiolocation, including passive, is based on searching in the received emissions for signals homothetic to the reference signal being delayed and frequency modified replica of the pattern, this is the signal illuminating the space [1, 2, 7, 8]. The delay τ of this pattern is proportional to the estimated distance R_X to the object, and the shift of the frequency f_d is determined by the Doppler phenomenon resulting from the so-called radial velocity ν of the object in relation to the radar [1, 2]. Saying this in the most general way, one can qualify functioning of the radar as the process of searching for a maximum of the ambiguity function $\chi(\tau,\nu)$ [6, 8, 9].

$$\chi(\tau,\nu) = \int_{-\infty}^{+\infty} S_E(t)\cdot S_R^*(t-\tau)\cdot \exp(-j\cdot 4\cdot\pi\cdot\nu/c\cdot f\cdot t)dt, \tag{9}$$

where:

$S_E(t)$ – received signal returned from the object (echo),

$S_R(t)$ – signal illuminating the space (pattern),

f – frequency of the illuminating signal,

c – velocity of the electromagnetic wave in vacuum,

$S_R^*(t-\tau)\cdot\exp(-j\cdot 4\cdot\pi\cdot\nu/c\cdot f\cdot t)$ – reference signal.

In the case of motionless objects or objects with a small radial velocity ($\nu\approx 0$), the value of the Doppler frequency f_d happens to be very small, so the exponent in the expression (9) can be neglected [1]. Then the ambiguity function reduces to a cross-correlation function $R_{ER}(\tau)$.

$$R_{ER}(\tau) = \int_{-\infty}^{+\infty} S_E(t)\cdot S_R^*(t-\tau)dt. \tag{10}$$

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In such conditions, the radar looks for the delay τ for which the function $R_{ER}(\tau)$ achieves its maximum, and the speed of the object ν is estimated by means of comparison of values τ corresponding to successive space illuminations by the radar.

In present day radars, the qualification of the degree of similarity of the echo signal and the reference signal is undertaken through the conversion of the frequency of these signals to the baseband (e.g. 70 MHz) and analog-to-digital conversion, and then calculating integrals (9) or (10) with the aid of digital signal processors [1, 4, 10-13].

As mentioned above, the difference of the phases of the signals can also be used as a measure of the similarity of these signals. In such a solution, those signals whose phase difference φ_w in a suitably long interval of time is invariable, will be recognized as homothetic (identical). In other words, the longer the interval of time that the phase difference of two signals stays invariable, the greater will be the extent of their similarity. That is why, among others, in active and passive radiolocation, arrangements possessing the ability of instantaneous measurement of the momentary value of the phase (phase difference) and momentary value of the frequency of microwave signals without frequency conversion (mixing), can be of great importance. These devices use quadrature microwave phase discriminators QMPD and microwave frequency discriminators MFD and they find use in so-called instantaneous frequency measurement receivers IFM [14-16]. Such receivers are used in warning and surveillance systems and for precise monitoring of short-lived changes of the parameters of microwave signals or of microwave devices [17].

2. The quadrature microwave phase discriminator

An example of the quadrature microwave phase discriminator is presented in Fig. 4. The construction, in the presented version, consists of three Lange directional couplers DC1 – DC3, one Wilkinson power divider PD1 and four microwave detectors D1 – D4 fulfilling the role of power measurement probes of microwave signals [14, 16].

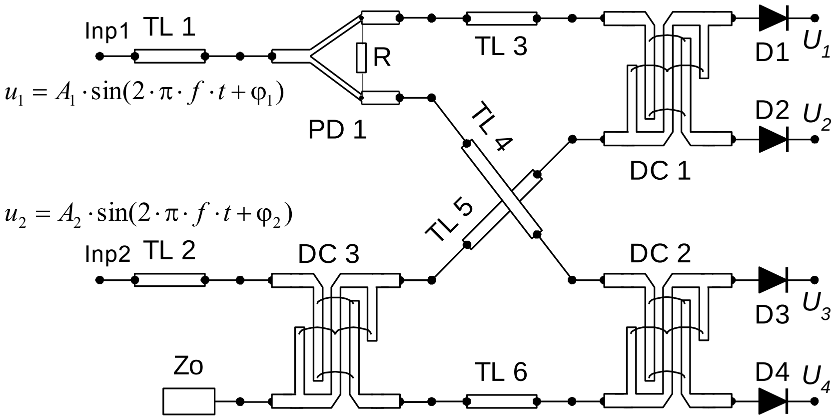

Fig. 4. An example of the quadrature microwave phase discriminator QMPD.

Directional couplers are the quadrature units of the vectorial splitting and adding up of microwave signals. Power divider is the in-phase element of signals splitting.

The reference signal is brought to one of the discriminator input ports Inp and the second port is excited by the signal being explored (echo signal). It is not essential, from the point of view of basic properties of the device, to which of the input ports the reference and tested signals are given.

The QMPD introduced in Fig. 4 does not multiply microwave signals u_1 and u_2 feeding ports Inp1 and Inp2, but only performs simultaneously four vectorial additions of these signals. This device was included to sum up quadrature microwave phase discriminators SQMPD because of this. In each of the additions, the phase of the individual signal is additionally suitably shifted [15]. As a consequence, signal powers in the directional couplers DC1 and DC2 output ports are functions of the amplitudes and phase differences of the input microwave signals. One can describe them, with closeness to the uniform factor, by the following expressions:

$$P_1 = 0.5 \cdot A_1^2 + 0.5 \cdot A_2^2 + A_1 \cdot A_2 \cdot \cos(\varphi_1 - \varphi_2), \tag{11}$$

$$P_2 = 0.5 \cdot A_1^2 + 0.5 \cdot A_2^2 - A_1 \cdot A_2 \cdot \cos(\varphi_1 - \varphi_2), \tag{12}$$

$$P_3 = 0.5 \cdot A_1^2 + 0.5 \cdot A_2^2 + A_1 \cdot A_2 \cdot \sin(\varphi_1 - \varphi_2), \tag{13}$$

$$P_4 = 0.5 \cdot A_1^2 + 0.5 \cdot A_2^2 - A_1 \cdot A_2 \cdot \sin(\varphi_1 - \varphi_2). \tag{14}$$

The values of these powers are measured with detectors D1 – D4 with voltage–power sensitivity γ. Therefore, it can be written:

$$U_1 = \gamma \cdot P_1, \tag{15}$$

$$U_2 = \gamma \cdot P_2, \tag{16}$$

$$U_3 = \gamma \cdot P_3, \tag{17}$$

$$U_4 = \gamma \cdot P_4. \tag{18}$$

Solving for the simplification that $\gamma = 1$, and subtracting these voltages in pairs, gives two components as $U_C(\varphi_T)$ and $U_S(\varphi_T)$. They contain information about the value of the phase difference φ_T of the signals u_1 and u_2 at the point in time t.

$$U_C(\varphi_T) = U_1 - U_2 = 2 \cdot A_1 \cdot A_2 \cdot \cos(\varphi_1 - \varphi_2), \tag{19}$$

$$U_S(\varphi_T) = U_3 - U_4 = 2 \cdot A_1 \cdot A_2 \cdot \sin(\varphi_1 - \varphi_2). \tag{20}$$

Both of these components are mutually quadrature in the phase difference φ_T domain. On this basis, one can estimate the momentary value of the phase difference of the signals incoming to QMPD according to the formula:

$$\varphi_T = \varphi_1 - \varphi_2 = \arctan\left[\frac{U_S(\varphi_T)}{U_C(\varphi_T)}\right]. \tag{21}$$

Information A_{12} about the amplitudes of these signals at the point in time t can also be extracted.

$$A_{12} = \sqrt{U_S^2(\varphi_T) + U_C^2(\varphi_T)} = 2 \cdot A_1 \cdot A_2. \tag{22}$$

Relation (22) is the equivalent of the quadrature amplitude detection that is usually performed utilizing the signals of the intermediate frequency obtained as the result of microwave signals' frequency conversion.

In radiolocation, the knowledge of the exact value of the power of received signals often is not required, but in noise or in disturbances only the presence of useful microwave signals is searched for. It was accepted therefore, for the purposes of simplification of the algorithms for detecting objects, that the momentary level of the received signal A_{12}^W can also be estimated basing on the sum of components $U_C(\varphi_T)$ and $U_S(\varphi_T)$ taken with the omission of their sign bits. We will then obtain:

$$A_{12}^{W}(\varphi_T)=|U_S(\varphi_T)|+|U_C(\varphi_T)|=2\cdot A_1\cdot A_2\cdot[|\sin(\varphi_T)|+|\cos(\varphi_T)|]. \tag{23}$$

Information about amplitude obtained in this way will be additionally a function of the phase difference φ_T, but the advantage of such a solution is that in comparable conditions, the value obtained from relation (23) can even be about 41% greater than that calculated according to formula (22). It will then be so when the phase difference φ_T comes up to ±45°. Generally, the relation will become true: $A_{12}^{W}(\varphi_T)\geq A_{12}$. Moreover, for the detection of the signals' amplitudes one can also use pairs of QMPD output voltages. This is shown by the formulae (24) and (25).

$$A_C=U_1+U_2=A_1^2+A_2^2, \tag{24}$$

$$A_S=U_3+U_4=A_1^2+A_2^2. \tag{25}$$

Volumes described by relationships (22) ÷ (25) differ in form, but they relate to these same amplitudes A_1 and A_2. In passive and active radiolocation, one of these amplitudes characterizes the reference signal, and the second refers to the echo signal, this means the microwave signal returned from the object being observed. Because the parameters of the reference signal are, with sufficiently good approximation, well-known and stable, so determination of the amplitude of the echo signal is possible using the described relations. One of the formulae (22) ÷ (25) can be used in radar for the estimation of momentary levels of the received signals or all four of them can be computed for the purpose of mutual verification.

The microwave phase discriminator, thanks to this, delivers voltages proportional to the momentary difference of the phase of the input signals, and can also be used to produce a signal with the frequency being equal to the difference of the frequencies of these signals. If two signals with frequencies f_1 and f_2 are given:

$$u_1=A_1\cdot\sin(2\cdot\pi\cdot f_1\cdot t+\varphi_1), \tag{26}$$

$$u_2=A_2\cdot\sin(2\cdot\pi\cdot f_2\cdot t+\varphi_2), \tag{27}$$

then, for example, the second of them can be considered to be the signal with the frequency of the first signal with additional phase changing with differential frequency f_X:

$$u_2=A_2\cdot\sin[2\cdot\pi\cdot f_1\cdot t+(2\cdot\pi\cdot S_{fx}\cdot f_X\cdot t+\varphi_2)], \tag{28}$$

where:

S_{fx} – sign of difference f_2-f_1,

$f_X=|f_2-f_1|$ – differential frequency. (29)

Therefore, the phase difference of two signals can be expressed generally by the universal relation given in the form:

$$\varphi_W=\varphi_1-\varphi_2-2\cdot\pi\cdot S_{fx}\cdot f_X\cdot t=\varphi_T-2\cdot\pi\cdot S_{fx}\cdot f_X\cdot t. \tag{30}$$

This means that if two signals with different frequency are feeding input ports of a quadrature microwave phase discriminator, the quadrature output components given by expressions (19) and (20) are a function of the current phase difference φ_T and they additionally change in time with the differential frequency f_X. Therefore, they can be described in a general form as: $U_C(t,\varphi_W)$ and $U_S(t,\varphi_W)$.

Because the average values of their changes with the frequency f_X are equal to zero, so the integration of components $U_C(t,\varphi_W)$ and $U_S(t,\varphi_W)$ will eliminate their parts which are caused by the differences of the frequency of signals u_1 and u_2. In this way, among others, useful signals can be extracted from a noisy environment. For example, as useful signals can be considered those that are identical to the reference signal passed to one of the quadrature microwave phase discriminator input ports. Values obtained from the integration of voltages $U_C(t,\varphi_W)$ and $U_S(t,\varphi_W)$ were assigned the symbols $I(\varphi_T)$ and $Q(\varphi_T)$.

$$I(\varphi_T) = \int_{t1}^{t2} U_C(t,\varphi_W)dt \tag{31}$$

$$Q(\varphi_T) = \int_{t1}^{t2} U_S(t,\varphi_W)dt \tag{32}$$

Components $I(\varphi_T)$ and $Q(\varphi_T)$ are also quadrature in the phase difference φ_T domain. If at least one of them is distinguished by a non-zero value, then it denotes that the summand similar to the reference signal is found in the explored signal. The phase φ_S and amplitude A_S of this summand as well as the phase φ_{Ref} and the amplitude A_{Ref} of the reference signal are related by the formulae:

$$\varphi_T = \varphi_S - \varphi_{Ref} = arctan\left[\frac{Q(\varphi_T)}{I(\varphi_T)}\right], \tag{33}$$

$$A_{12} = \sqrt{I^2(\varphi_T) + Q^2(\varphi_T)} = 2 \cdot A_S \cdot A_{Ref}\,. \tag{34}$$

The limits of integration $t1$ and $t2$ contained in expressions (31) and (32) are dependent on the characteristic features of the sought after useful signals. They are determined for instance, by pulse width or by width of the signal band.

3. Passive radar with the quadrature microwave phase discriminators

The simplified structure of the developed passive radar, built on the basis of the quadrature microwave phase discriminators QMPD is shown in Fig. 5. The antenna A1 of the reference channel connected to a low-noise amplifier LNA and bandpass filter BPF, receives the signal arriving from the occasional i.e. non-cooperative transmitter directly. This signal, after distribution with the aid of the power divider PD1, is used to produce the n number of the reference signals that are the delayed standards with diverse delays τ_i.

The signal received in the reference channel is also passed to the quadrature microwave frequency discriminator QMFD. This circuit evaluates the momentary value of the frequency and the envelope of the so-called signal of opportunity being used. In the echo channel, signals emitted by the non-cooperative transmitter are considered as useful, but only the ones returned from the object situated in the segment of the space observed by the passive radar. That is why the antenna system A2 and A3 of the echo channel is connected to the microwave beam forming network BFN. One of its main tasks is the cancellation of the direct signals arriving from the non-cooperative transmitter used.

The signal received in the echo channel, after amplification in the low-noise amplifier LNA is directed to the bandpass filter BPF and then is distributed by the power divider PD2 to all the microwave phase discriminators QMPD simultaneously. The individual reference signals numbered 1 – n are also applied to these discriminators. After making the integration at the outputs of each of the n QMPD, couples of voltages $I(\varphi_T)$ and $Q(\varphi_T)$ are obtained proportional to the constancy of the phase difference, and now showing the rate of the similarity of the received signal to the reference signal of number i.

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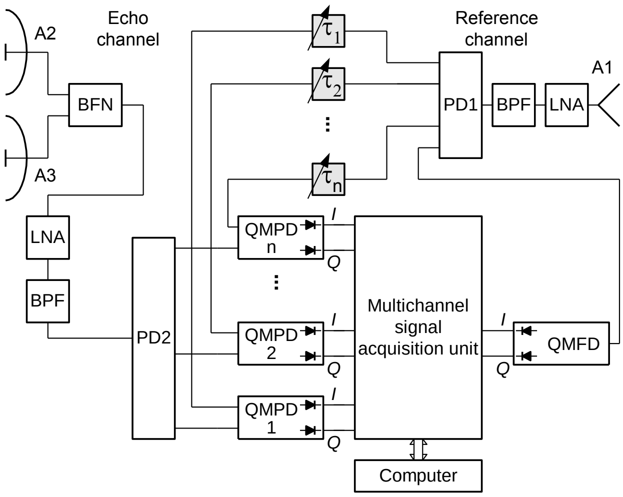

Fig. 5. Basic structure of the passive radar with the quadrature microwave phase discriminators.

The algorithms for object detection and distance estimation are processed by a computer. It decides that the object was detected when at least one of values $I(\varphi_T)$ or $Q(\varphi_T)$ is larger than the established threshold. The distance to the object is calculated in turn on the basis of three delays τ_j, τ_k, τ_m corresponding to the reference signals delivered to these phase discriminators QMPD, from whom the pairs of components $I(\varphi_T)$ and $Q(\varphi_T)$ give the largest values computed in accordance with formula (34).

Thanks to the use of several independently working QMPDs, the several operations of comparison of the received signal phase and the phase of reference signal delayed about several times τ_i, are processed simultaneously. This shortens the time of estimation of the delay of the signal returned (distance to the object) in comparison with the classic method of enumerating the ambiguity function $\chi(\tau, \nu)$ or the cross-correlation function $R_{ER}(\tau)$. The first approximate result of the estimation of the distance to the object is obtained just after the time to the largest delay τ_i (Fig. 5). In radars of 150 km range, this will be a time of 1 millisecond.

4. Conclusions

In passive radars using signals of opportunity emitted by non-cooperative transmitters, the pattern signal used to form the reference signal and to detect the echo signals has to be extracted from the surrounding space or reproduced on the basis of measurements of the received emission or on the basis of predictions or information coming from reconnaissance.

In the presented structure of the radar, to find the returned signals, instead of evaluating the cross-correlation function requiring the multiplication of signals, their parallel vectorial addition with simultaneous diverse phase shifting was proposed. In the presented version, the vectorial additions are processed using the microwave sixport which is part of the quadrature microwave phase discriminator QMPD and which consists of one power divider and three directional couplers. This discriminator can instantaneously compare phases of simple signals and phases of the noise signals too. Therefore, the passive radar based on the quadrature microwave phase discriminators is able to make use of pulsed or continuous narrowband and wideband signals as well.

Analogous operations of vectorial summing can also be done digitally after prior shifting of the microwave signal spectrum to the baseband and then digitization of the intermediate frequency signals being compared.

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