Coherent System of Synhronisation of Carrier Frequency Hopped Signals

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Abstract:**The paper presents a method and a device of an accurate and independent (coherent) reproduction of the form of carrier frequency hopped signals in the transmitter and receiver. The base of that process is a coherent device of synchronization serviceable with random delay fluctuations.**

Key words: **carrier frequency hopped signals, coherent device**

I. INTRODUCTION

The coherent formation of signals transmitted is a way to increase the noise immunity of telecommunication systems. Nowadays, due to their features the coherent lines allow to reach the conditions of optimal receiving closely by satisfying the requirements for optimally simple technical implementation. The development of coherent carrier frequency hopped communication systems is grounded on the fact that they give a possibility to provide a great amplification with signal processing in the receiver. It is of a special significance for the mobile telecommunications because of the limited power resources. However, the requirements for the synchronization system have been considerably increased, especially for the system of delay monitoring. The purpose of this paper is to propose a device of accurate and independent (coherent) reproduction of the form of carrier frequency hopped signals in the transmitter and receiver.

Two patents of coherent systems of delay monitoring have been known. They belong to Motorola and have found application to radioconection with fast-movable objects (airforce planes), i.e. under conditions of rapid fluctuations of delay [2,3]. In the delay monitoring system (DMS) proposed in [2], the signal in transmitter, intended to be transmitted in the band of modulation frequencies, is summed with the narrow-band pilot signal with angular frequency ω_p. At the same time the summed output signal is mixed with the carrier frequency hopped signal of the synthesizer generating a signal of continuous phase $\omega(t)$. The structural circuit of the coherent DMS used in the examined narrow-band transmitting system is presented in Fig.1**.** The obtained signal entering the DMS input is of the kind of:

$$S_1(t) = \mathrm{Re}\left\{\exp j\left[\omega_p t + \varphi_1(t)\right]\right\} \tag{1}$$

where $\varphi_1(t)$ is the integral phase of the carrier frequency hopped signal received.

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$$\varphi_1(t) = \int_0^T \omega(t')\,dt. \tag{2}$$

At the second input of the input mixer, reference signal $S_2(t)$ from the synthesizer frequency of carrier frequency hopped signals enters:

$$S_2(t) = \mathrm{Re}\left\{\exp\left[-j\varphi_2(t)\right]\right\} \tag{3}$$

where $\varphi_2(t)$ is the integral phase of the reference signal. The band filter (BF) separates the component corresponding to the frequency of pilot signal ω_p from the output signal. This component can be written in the kind of:

$$S_f(t) = \mathrm{Re}\left\{\exp j\left[\omega_p t + \varphi_1(t) - \varphi_2(t)\right]\right\} \tag{4}$$

Following the band filter, a circle for PPL (a phase-locked loop) is switched separating component S_f from the signal:

$$S_s(t) = K\left[\omega(t) - \varpi\right]\tau = K\left[\varphi_1(t) - \varphi_2(t)\right], \tag{5}$$

where K is the coefficient of the phase detector transmitting (PD); ϖ is the average angular frequency.

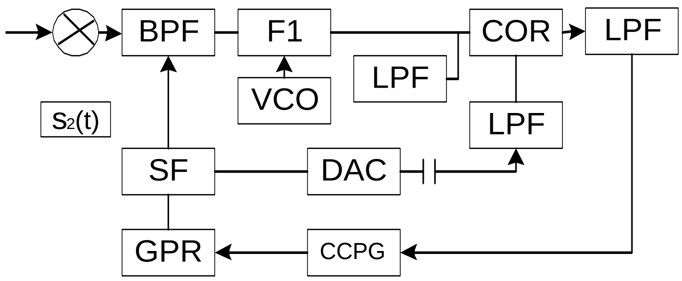

Fig1

The error signal formed is supplied to one of the correlator inputs made in the kind of multiplication and switched to a low pass filter (LPF) in series. At the other correlator input, a signal proportional to the reference frequency hopped signal is supplied. It is without a constant component, which is cut by a condenser. The second signal is:

$$S_s(t) = \omega(t) - \varpi \tag{6}$$

It should be underlined that to obtain the above ratio (5) it is necessary to provide a mode of operation in the linear section of the phase detector discrimination feature PD from the circle of PPL, which is satisfied for values $\tau << T$. Thus, at the corrector output signal

$$S_g(t) = c(t)\, f(t) = K\left[\omega(t) - \varpi\right]^2 \tau, \qquad (7)$$

is obtained, which after averaging in LPF has the kind of:

$$S_g(t) = K\overline{\left[\omega(T) - \varpi^2\right]^2}\tau \qquad (8)$$

The error signal obtained is used to final adjustment during the controllable clock pulse generator (CCPG) of DMS.

Another version of building a similar coherent delay monitoring system is described in [2]. The principal difference in comparison to the system examined in [1] consists in using a special device for phase synthesis (phase synthesizer) in the receiver in the part of a local generator-and-synthesizer carrying out the voltage to phase conversion. Thus the phase of the controllable signal for the phase synthesizer has to be equal to the value of the current synthesized frequency phase. The structure diagram of the DMS given is shown in Fig. 2. If the input signal in the receiver is presented with equation (1), then the signal at the output of the BPF (band pass filter) mixer intended to remove the frequency hopping will be described with equation (5).

In its general kind the phase of the received carrier frequency hopped signal can be written as:

$$\varphi_1(t) = \omega_0\tau + \int_0^t \omega(t')\,dt', \qquad (9)$$

where ω_0 is the central signal frequency.

$\omega(t')$ is the amplitude of the random oscillation with a rectangular form, which determines the value of the carrier frequency hopped signal and which is allocated evenly between ω_p and $-\omega_p$.

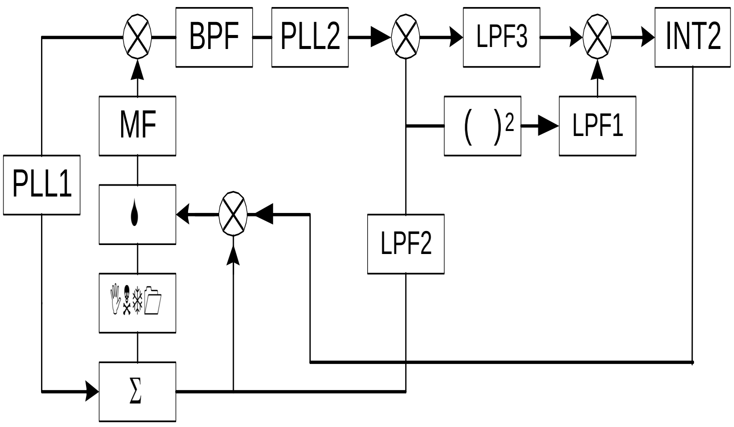

Fig.2

To accomplish the synchronization between the signal received and the reference signal in delay with a certain value to the phase, it is necessary to reduce the phase error to zero, i.e.:

$$\Delta\varphi = \varphi_2(t) - \varphi_1(t) \qquad (10)$$

For the case when $\tau << \tau_e$, an approximate equality has been accomplished:

$$\Delta\varphi \approx \tau \frac{d\varphi_1(t)}{dt}, \qquad (11)$$

or

$$\frac{d\varphi_1(t)}{dt} \approx \omega_0 + \omega(t). \qquad (12)$$

Hence, the assessment of the output signal phase can be presented in the kind of:

$$\hat{\varphi}(t) = \varphi_2(t) - \frac{\tau d\varphi_1(t)}{dt}. \qquad (13)$$

At that, the error of the assessment obtained can be evaluated by the expression:

$$\Delta\omega(t) = \omega(t+\tau) - \omega(t), \qquad (14)$$

with which $\left|\Delta\omega(t)\right| \le \left|\omega(t)\right|$.

The method proposed through the diagram in Fig.3 can be implemented in this way:

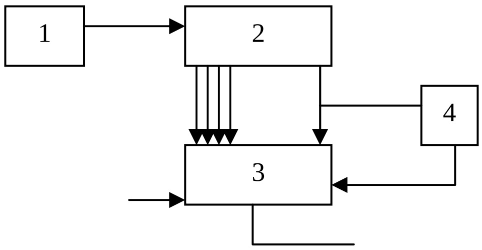

Fig.3

The signal determining the regularity of the phase change for the phase synthesizer (PS) is found at the first generator output and presents an integral of the sum of the phases of the central frequency received at the output of the circle for PPL1 (phase-locked loop 1) from the input radio signal and the signal entering from generator of pseudo random sequence (Gprs), which determines the value of $\omega(t)$. From the phase value, obtained in the removing device, the preliminary value of the evaluation of the phase error formed on the base of delay $\tau_з$ is deducted and the result obtained is used as a controlling effect for the PS. Thus PS forms a signal of the following kind:

$$S_2(t) = \mathrm{Re}\left[\exp j\left\{-\tau_з\omega(t) + \omega_0(t+\tau) + \int_0^{t+\tau} \omega(t')dt'\right\}\right]. \qquad (15)$$

The synthesizer examined, which is called a direct synthesizer in [2], presents an addressed storage device (SD), where the values of $\exp jx$ are kept, and where x is a number supplied at the SD input. At that, the digital-to-analog converter (DAC) operating according samples $\mathrm{Re}\{\exp jx\}$ is used at the SD output.

The signal at the mixer output is in the kind of:

$$S_m(t) = \mathrm{Re}\left\{\exp\left[(\tau - \tau_з) - \bar{\omega}(t)\right] - \tau_з \tau \dot{\omega}(t)\right\}. \quad (16)$$

Signal $S_m(t)$ passing through a band filter enters the input of the circle for PPL 2 (phase-locked loop 2). It divides the signal phase at the difference equal to the value of exponent argument $S_m(t)$. In the case examined, the circle for B PPL 2 in its structure and purpose is analogous to the circle for PPL 1 in DMS as described in [1]. . The signal at the output of the circle for PPL 2 is supplied to the input of the multiplier of the circuit for delay assessment, at the other end of which the value of $\omega(t)$ received by Gprs is averaged in PPL 1and LPF 1**.** The time constant of LPF 1 corresponds to the time constant of the BPF. In this way at the multiplication output the following signal is obtained:

$$S_D(t) = (\tau - \tau_з)\bar{\omega}^2(t). \quad (17)$$

The averaged value of $\bar{\omega}(t)$ is supplied to a squaring device and after averaging in LPF 2 with a time constant bigger than that of LPF 1, a signal of the following kind has been obtained:

$$S_d(t) = \bar{\omega}^2(t) \quad (18)$$

This signal is supplied as a divider to the input of a circuit for dividing. As a of dividend, a signal of the following kind is used:

$$\bar{S}_D(t) = (\tau - \tau_з)\overline{\bar{\omega}^2(t)}, \quad (19)$$

That is obtained from $S_\partial(t)$ by averaging in LPF 3. Thus a new evaluation of the delay is formed:

$$\tau'_з = \frac{S_\partial(t)}{S_d(t)} = \tau - \tau_з, \quad (20)$$

which at the integrator output is in the kind of:

$$\bar{\tau}'_з = \overline{\tau - \tau_з}. \quad (21)$$

The latter joins the multiplication where the next signal is formed for the frequency correction of the delay monitoring system DMS.

In comparison to the coherent system for monitoring described in [1], the circuit examined allows to remove the considerable phase distortions occurring at the moment of correlation. It is so because in that case the final adjustment is implemented not at the clock-pulse frequencies of the controllable clock-pulse generator CCPG, but directly in the generator phase on a signal copy, the phase synthesizer plays the part of which.

Another advantage of the given circuit stemming from its peculiarities is its capability to monitor Doppler's frequency displacement accurately. That is of a special significance in the carrier frequency hopped mobile radio telecommunications.

The disadvantages of these systems are the necessity of using a coherent synchronizing signal (pilot signal) and the possibility of operation only at minimal values of time delay, i.e. if the condition that the delay time is much less than the frequency element length of the carrier frequency hopped signal has been satisfied.

The purpose of the invention is to develop a method and a device for an accurate and independent (coherent) reproduction of the phase in the transmitter and receiver of carrier frequency hopped signals. On this basis, the task is to invent a coherent device of synchronization serviceable with random fluctuations of delays.

The problem has been solved developing a synthesizer frequency, which includes a source of standard (reference) signal, a generator of primary signals a switching circuit. The line of primary signals of an angular frequency corresponding to the reference signal is lead to the switching circuit output in a kind of a phase-displaced line. The switching system is synchronized with one of the primary signals and switching can be controlled by the choice of the number of primary signal cycles.

The advantage of the synthesizer frequency proposed is that the primary signals are generated at the same spectrum frequency as the reference signal source. As a sequence, if the switching circuit is synchronized with one of the primary signals, the chosen signal of the same frequency is obtained each time with the start of synthesis. That allows to control the synthesized signal phase and to reproduce it independently in the transmitter and receiver.

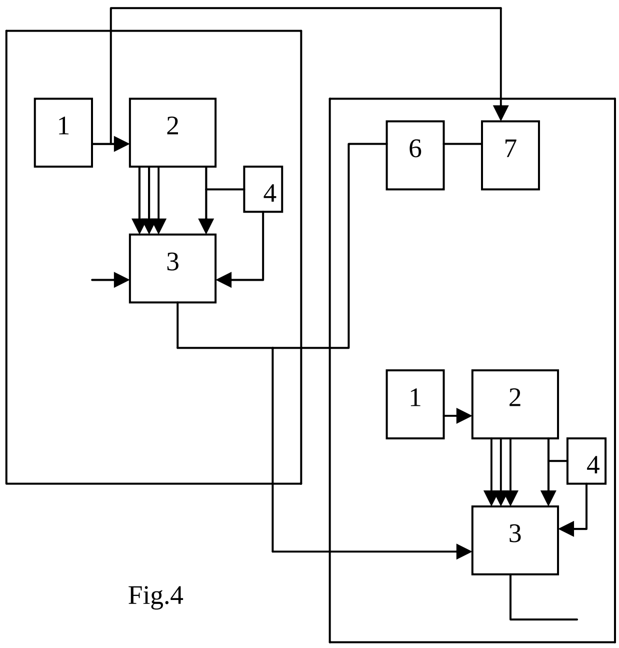

Fig.4

A sample of the proposed synthesizer frequency is shown in Fig. 1 presenting its block diagram. According to the invention the synthesizer frequency consists of a reference signal source, generator of primary simple signals made in the kind of a multi-pin delay line with N pins, a switching circuit, divider of frequencies synchronized with one of the generator simple signals and connected with the switching circuit pulse input.

The operation of the synthesizer is as follows: the signal of angular frequency ω_0 from the reference generator (1) is supplied to the input the generator of the signals (a multi-pin delay line with N pins). The switching circuit (3) is synchronized through the divider of frequencies (4) and switches at every k cycles of the primary signal. If the time interval between two switches is T_s , then the phase difference between the frequency synthesized and the reference signal increases with phase θ for time T_s . As $\theta = \frac{2\pi}{N}$, then the signal synthesized can be kept for an arbitrary part of time with a multiple use of N primary signals and the frequency of the signal synthesized ω_c will be $\omega_c = \omega_0 + \frac{k\theta}{T_s}$. Controlling the switching circuit (the choice of k), the output signal phase can be changed.

On the base of the synthesizer proposed, a device of carrier frequency hopped signals synchronization has been obtained.

It contains a transmitter that can be regulated (5), a receiver (8), a delay monitoring system (6), an extrapolator of the delay (7) and a reference signal source (), a generator of primary simple signals (2), a switching circuit (3), a divider of frequencies (4).

The advantage of the device proposed is that on the base of the delay assessment extrapolation, it determines the moment to start the synthesis of the desired signal in the correlator of receiver obtaining the same phase as that used in the transmitter.

A sample of the device of coherent synchronization according to the invention is shown in Fig. 4.

The operation of the device is as follows: In the standard delay monitoring circuit it is the delay current value is evaluated. For the delay compensation the extrapolator determines value $\tau^*(t)$ and hence the advance in time, i.e $\tau^*(t) = t - \tau(t)$ of transmitting in the transmitter, which is done by supplying the signal to the reference signals generator input (multi- pin delay line).

Let one of the n- number elements signals, which are due for those identically equal to zero out of the interval, is marked with $S_k[0,T]$. Then the signal transmitted can be described with the expression:

$$S(t) = \sum_{k=0}^{n} S_k(t - iT) \tag{22}$$

With the presence of random time lag, the useful signal received is:

$$S(t,\tau(t)) = S(t-\tau(t)) = \sum_{t=0}^{n} S(t - iT - \tau(T)) \tag{23}$$

Let present it in the kind of a know function of information discrete parameter $\theta(t)$ and random time lag $\tau(t)$, i.e.:

$$S(t,\theta,\tau) = S\left[t-\tau(t), \theta(t-\tau(t))\right] \tag{24}$$

The discrete parameter takes constant values on tact interval $\theta(t) = \theta_i$, $t \in [t_i, t_{i+1}]$. The values of the information parameter on the various tact intervals form a simple Markov's chain θ_i, $i = 0,1,...,n$ with n states and a known matrix of the transition from the i-th into the j-th state $\Pi = \Box_{i,j}$, and vector of the initial states $\Box = \Box_I$. The limits of the tact intervals are determined by a random time lag $\tau(t)$, i.e. $t_i = t_i(\tau)$. With the time lag realization specified, the limits of tact interval are:

$t_i = iT + \tau(t_i)$.

On tact interval t_i, t_{i+1}, signal $S(t,\theta,\tau)$ coincides with elementary signal $S_i(t - iT - \tau(t))$ if $\theta(t) = \theta_i$. Random time lag $\tau(t)$ corresponds to the signal lag caused by relative motion of the receiver and the transmitter and in the common case it can be examined a component of the diffusion Markov's process:

$$\lambda(t), \tau(t) = \lambda_1(t)$$

Process $\lambda(t)$ satisfies the system of stochastic differential equations:

$$\frac{\partial \lambda_i(t)}{\partial t} = f_i(t,\lambda) + n_i(t) \tag{25}$$

Here $f_i(t,\lambda)$ are functions satisfying the condition of Lipshits [1] and $n_i(t)$ are a Gauss noises with intensity $b_{ij}(t,\lambda)$. The a priori probable features of process $\lambda(t)$ are determined by the equation of Kolmogorov-Foker-Plank [2]:

$$\frac{\partial W}{\partial t} = -\sum_{\alpha=1}^{n} \frac{\partial}{\partial \lambda}\left[a_\alpha(t,\alpha)W\right] + \frac{1}{2}\sum_{\alpha=1}^{n}\sum_{\gamma=1}^{n} \frac{\partial^2 b_{\alpha\gamma}(t,\alpha)W}{\partial \lambda_\alpha \lambda_\gamma} \equiv L[W] \tag{26}$$

where $W = W(t,\lambda)$ is the a priori probable density of process $\lambda(t)$;

The observation on signal $S\left(t,\theta,\tau\right)$ is realized on the background of noise, i.e. it has kind of

$$\xi\left(t\right)=S\left(t,\theta,\tau\right)+n\left(t\right) \tag{27}$$

Where $n\left(t\right)$ is non- correlated with $\theta\left(\tau\right)$ and $\tau\left(t\right)$ is white noise of feature $\square\left\{n\left(t\right)\right\}=0.$

With a certain realization of lag $\tau\left(t\right)$, the optimal (according to the criterion of the error probability minimum) assessment of the information discrete parameter constant in interval $\left[t_i,t_{i+1}\right]$ is determined with expression [1]:

$$\theta_i^* = \max{}^{-1}\left\{P\left(\theta_i=i\right)\right\} = \max{}_i^{-1}\left\{P\left[\left(i+1\right)T+\tau,i\right]\right\} \tag{28}$$

With random values of tact interval t_{i+1}, when posterior probabilities $P=\left(\theta_i=i\right)$ have to be compared, in the moments of time the conditional probability has to be examined with fixed τ, i.e.

$$P\left(\theta_i=i\right)=P\left[t_{i+1}\left(\tau\right),i/\tau\right] \tag{29}$$

The optimal assessment of the discrete parameter can be examined as a non-conditional discrete parameter posterior probability at the end of the interval under observation. The probability can be determined by the expression:

$$P\left[t_{i+1}\left(\tau\right),i/\tau\right]=P\left[i/\tau,\xi^{t_{i+1}\left(\tau\right)}\right] \tag{30}$$

made average with weight corresponding to posteriori probable density of random lag $P\left(t,\tau\right)$. Then in the i-th tact interval the algorithm of assessment of the information discrete parameters takes the kind of:

$$\theta_i^* = \max{}_i^{-1}\left\{P\left(\theta_i=i\right)\right\} = \max{}_i^{-1}\left\{\int P\left[t_{i+1}\left(\tau\right),i/\tau\right]P\left[t_{i+1}\left(\tau\right),\tau\right]d\tau\right\} \tag{31}$$

II. CONCLUSION

The paper presents an algorithm for optimal receiving signals with hop-like change of frequency with random lag. The direct examination on the random lag allows obtaining an algorithm involving a wide range of tasks. For example, such tasks are receiving signals for which the random lag occurs not only in moving away the limits of the time tact intervals but also in receiving under the conditions of changing lag.

The algorithm obtained (31) presents a summary of an algorithm for assessment of a constant parameter in a certain interval with the presence of indeterminacy at the moment of signal appearing.

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