IMPROVING NOISE IMMUNITY OF DIGITAL RADIO SYSTEMS WITH ANGULAR MODULATION

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ABSTRACT

To double the noise immunity of digital radio communications of frequency-managed transport systems, it is proposed to use the method of coherent detection of frequency-manipulated signals (FMn) instead of envelope detection. At the same time, the reverse operation of the coherent detector of signals with absolute phase manipulation (PMn) at 180° is achieved due to the post-detector signal processing. Such processing is possible with participation of a frequency detector on frustrated circuits with a trigger connected to it.

Keywords: railway, radio communication, frequency, amplitude and phase manipulations, noise immunity, coherent detection, reference oscillation, absolute phase manipulation at 180°, reverse operation of a coherent detector and its elimination.

Background. *In railway radio communications [1], only narrowband frequency modulation is used, and the level of interference is substantially high, and the connection is not always satisfactory.*

Noise immunity and frequency efficiency are the most significant indicators of quality of radio communications, since noise immunity is related to safety of vehicles, and frequency efficiency helps to reduce the shortage of the frequency resource. The efficiency of a radio communication is characterized by criteria that follow from the well-known formula of Shannon [2, p. 308] for bandwidth C of the communication channel. The authors showed [3, p. 29] that the Shannon formula allows us to connect not two, as usual [4, p. 3; 5, p. 69; 6, p. 117; 7, p. 120], but three main criteria. For this, it is necessary to separate the left and right parts of the specified formula, as well as the power of signal P_s and noise P_n under its logarithm by the information transfer rate R, and not just replace C with R [3] with the division under the logarithm. As a result, we have:

$$C/R = F/R \cdot \log(1 + P_s/P_n \cdot R/R) \text{ or}$$

$$1 = \eta/\alpha \cdot \log(1 + \alpha/\beta) \text{ or } 2 = (1 + \alpha/\beta)^{\eta/\alpha}, \qquad (1)$$

where $\alpha = R/F$ – criterion of frequency efficiency;

$$\beta = \frac{R}{P_s / N_0}$$ – *criterion of energy efficiency;*

$\eta = R/C$ – criterion of channel bandwidth efficiency (information efficiency).

In this case, the third criterion (η-efficiency) is connected analytically with the criteria α and β according to (1). The β-efficiency includes noise immunity of railway radio communication.

Objective. *The objective of the authors is to consider the issue of improving noise immunity of digital radio systems with angular modulation.*

Methods. *The authors use general scientific and particular radio engineering methods, comparative analysis, mathematical methods, simulation.*

Results.

1. Coherent Detection

The transition from analog to digital modulating signal [10, p. 147; 11, p. 197; 12, p. 195], which is frequency modulation, often called frequency manipulation (FMn), provides a higher noise immunity, but the immunity provided is not the finite. The frequency-manipulated signal consists of two amplitude-manipulated (AMn) signals on its subcarriers. Therefore, the FMn signal detector consists of two parallel detectors along the envelope of the filtered AMn signals, the outputs of which are connected to a subtractor.

With coherent detection, the quadrature component of interference is eliminated, which is why this power is reduced on average by two times, and the noise immunity increases by the same number of times, which is not the case when AMn detects an envelope signal. It is difficult to apply coherent detection when the FMn oscillation is filtered out because of the probability of obtaining a carrier frequency oscillation. The authors proposed a different method [8, p. 2] for coherent detection of signals with angular modulation. The FMn signal is converted into an AMn signal, whose oscillation is carried by the FMn oscillation, i.e. converted to the AFMn oscillation. The latter is then multiplied with its own carrier frequency oscillation, the FMn oscillation.

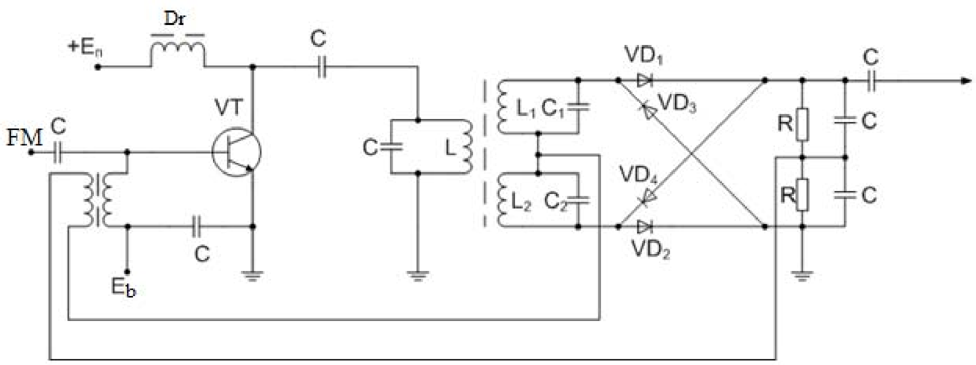
A diagram of the coherent detector of the FMn signals developed by [8, p. 2] is shown in Pic. 1. It consists of the final stage of the IFA receiver, the load of which is an oscillating circuit tuned in resonance to the intermediate frequency $f_{int.}$. This circuit is inductively coupled with two circuits, symmetrically and oppositely disturbed with respect to $f_{int.}$, as shown in Pic. 2. Both circuits are interconnected directly, and their outputs are connected to the inputs of a ring balanced multiplier (RBM). The outputs of RBM, in turn, are connected to the inputs of two RC circuits connected to each other. At the input of IFA, a transformer is turned on, the secondary winding of which is connected to the midpoints of RBM – the connection points of the disconnected circuits and RC chains. The amplitude-frequency characteristic (AFC) of the detuned circuits forms a linear region around f_{int}, three times larger than that of a single detuned circuit. This linear AFC section converts the input FMn signal to the AMn signal, whose oscillation is the input FMn oscillation, that is, converts the oscillation into the AFMn oscillation.

At the midpoints of RBM, an input FMn constant amplitude oscillation is applied, which is the carrier frequency oscillation for the AFMn signal. By multiplying AFMn and FMn signals in RBM and filtering in the RC-chains, the transmitted digital signal is obtained (Pic. 2). Analytically it is easier to show this process for the harmonic modulating signal $b(t) = U\sin\Omega t$.

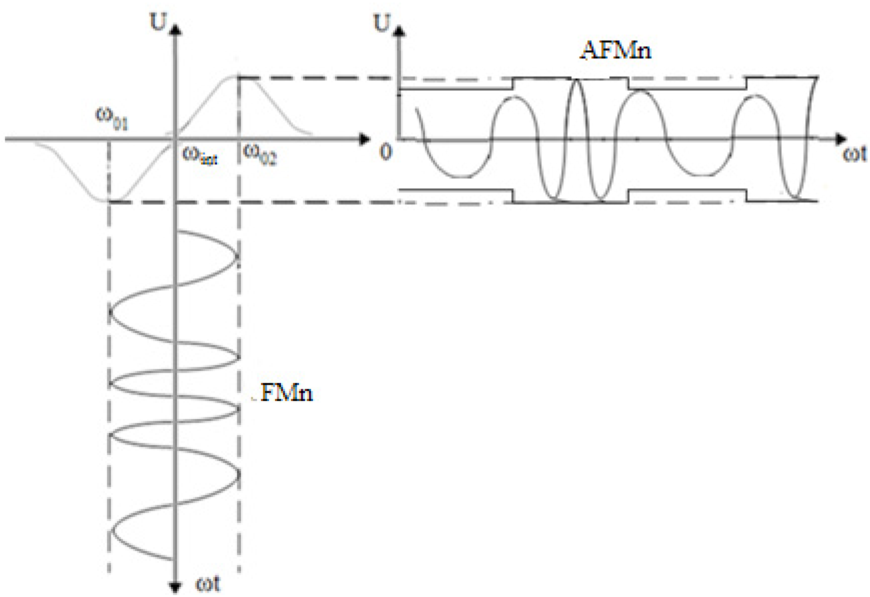
$$u_п(t) = u_{AFMn}(t) \cdot u_{FMn}(t) = U_m \cdot [1 + M\cos\Omega t] \cdot$$
$$\cdot \cos[\omega_{int} t + M\sin\Omega t] \cdot U_m \cos[\omega_{int} t + M\sin\Omega t] =$$
$$= U_m^2 \cdot [1 + M\sin\Omega t] \cdot \cos^2[\omega_{int} t + M\sin\Omega t] =$$
$$= U_m^2 \cdot [1 + M\sin\Omega t] \cdot (1 + \cos 2(\omega_{int} t + m\sin\Omega t)/2) =$$
$$= 0{,}5M \cdot U_m^2 \sin\Omega t + C + h.f, \qquad (2)$$

where h.f. – high-frequency component, which is eliminated by the filter (RC-chains), and C – the constant component, which is eliminated by the coupling capacitor, which is confirmed by the computer simulation.

Although the coherent detection of the FMn signal by the proposed method makes it possible to increase the noise immunity of railway radio communication twice, it is not, as already emphasized, the maximum that can be achieved. The maximum possible one,



Pic. 1. Schematic diagram of the developed frequency detector.



Pic. 2. Signal conversion: FMn to AFMn.

according to theoretical postulates, is provided by an absolute FMn at 180˚.

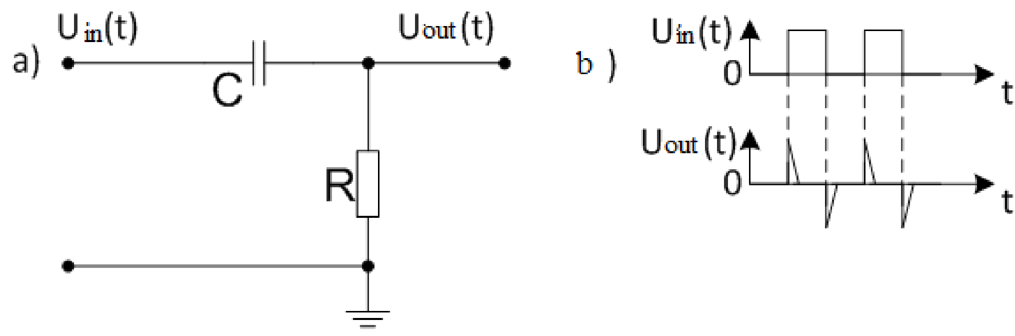
2. Maximum noise immunity

Absolute FMn at 180˚ is not used in practice because of the «reverse operation» of a coherent detector of its signals, when «1» of a digital signal is taken as «0», and «0» is taken as «1», i.e. vice versa. Instead of absolute, relative PMn (OPMn) at 180˚ is used, which is not only inferior to the absolute in noise immunity by half, but also requires more complex equipment. Therefore, it is of interest to develop a method for coherent detection of signals from the absolute PMn at 180˚ without reverse operation.

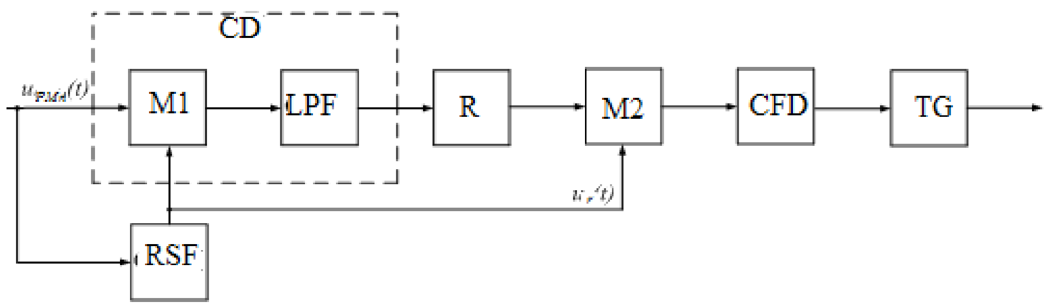
A coherent detector consists of a multiplier of signals with a low-pass filter at its output, in which the input signal $u_{in}(t)$ with the absolute PMn at 180˚ and the reference oscillation $u_r(t)$obtained from this input signal are multiplied together. The authors showed [9, p. 74] that in all coherent signal detectors with absolute PMn at 180˚, the resulting reference oscillation $u_r(t) = \pm\sqrt{u_{in}^2(t)}$ was obtained. Since the square root is preceded by a sign ±, then the reference oscillation $u_r(t)$ will have random phase jumps at 180˚, not only from the effects of interference, but also from other processes in the receiver. These phase jumps at 180˚ generate the reverse operation of a coherent detector of signals with the absolute PMn at 180˚.

Since $u_r(t) = \pm\sqrt{u_{in}^2(t)}$, the reverse operation of the coherent detector of signals with the absolute PMn at 180˚, consisting of a multiplier, a low-pass filter and a driver of the reference oscillation, is unavoidable, which has been proven in practice. However, the post-detector processing of its signals with additional blocks allows us to eliminate the reverse operation of the coherent detector [9, p. 73]. For example, in the Pistolkorsa detector, $u_r(t)$ is filtered by two resonant circuits of the first and second harmonics, and therefore the reference oscillation is obtained, in fact, without interference, but with random phase jumps at 180˚.

The authors proposed to use a regenerator at the output of the low-pass filter of the detector (a nonlinear



Pic. 3. Differentiating RC-chain (a), the pulse graph at the inputs and outputs of the chain (b).



Pic. 4. Scheme of the absolute PMn driver at 180˚.

device – say, an amplitude limiting amplifier), which restores distorted pulses and suppresses interference, but does not correct the reverse operation of the coherent detector. If, on the other hand, we multiply the alternating detected signal b(t) and the reference harmonic oscillation, we obtain a signal from the absolute PMn at 180˚ without phase jumps at 180˚ and without interference. The absence of phase jumps is easier to show with multiplication of harmonic signals:

$u_2(t) = b(t) \cdot u_r(t) = U_1(t) \cdot \cos(\Omega t + 180^o) \cdot U_2(t) \cdot$
$\cdot \cos(\omega t + 180^o) = U_1 U_2[\cos(\omega + \Omega)t + \cos(\omega - \Omega)t].$

Such PMn at 180˚ signal without interference and distortion can be coherently detected by the circuit shown in Pic. 1 by adding a trigger to the output of the low-pass filter.

Absolute PMn at 180˚ can be obtained by multiplying alternating discrete signals and harmonic carrier oscillations $u_c(t)=U_m\cos\omega t$, which allows it to be represented analytically:

$$u_{pmn}(t) = \gamma(t) \cdot u_c(t) = \gamma(t) \cdot U_m\cos\omega t =$$
$$U_m\cos(\omega t + \gamma(t)90^\circ - 90^\circ) =$$
$$= U_m\cos(\omega t + (\gamma(t) - 1) \cdot 90^\circ), \qquad (3)$$

where $\gamma(t) = \pm 1$ – alternating modulating signal.

It is known that PM is always accompanied by FM, since the circular frequency $\omega(t)$ and phase $\phi(t)$ are related by the relation $\omega(t) = d\phi(t)/dt$. The frequency $\omega(t)$ can be determined from the initial phase (1), i.e. with respect to $b(t) = \gamma(t) \cdot 90^\circ - 90^\circ$. For the analytical expression of FMn component (3), we expand the periodic modulating discrete rectangular signal into a Fourier series [6, p. 61] and obtain:

$$b(t) = \frac{1}{2} + \frac{1}{\pi}\sum_{k=1}^{\infty}\frac{1-\cos k\pi}{k}\sin k\Omega t. \qquad (4)$$

Hence

$$\frac{db(t)}{dt} = \frac{1}{\pi}\sum_{k=1}^{\infty}k\Omega\left(\frac{1-\cos k\pi}{k}\cos k\Omega t\right).$$

For clarity of this expression, let us skip the periodic sequence of rectangular pulses through the differentiating RC-chain presented in Pic. 3a, and head the output signal to the oscilloscope. In this case, the differentiating chain constant $\tau = RC << T$, where T is duration of a unit element (impulse or pause). As can be seen, at the output of the differentiator (Pic. 3b), only the front and section of each input rectangular pulse takes place. If a trigger (integrating device) is connected to the output of the differentiating chain, then at its output we will receive input rectangular pulses.

A scheme of the developed coherent detector of signals with the absolute PMn at 180˚ is shown in Pic. 4. It consists of the multiplier M1, the low-pass filter (LPF) at its output, the reference signal former (RSF), through which the first and second inputs of the multiplier M1 are connected, as well as the regenerator R connected in series to the LPF, the second multiplier M2, augmented by a coherent frequency detector CD, trigger TG. The second input of M2 is connected to the output of RSF.

In this case, the linear section of the PFC of the disturbed circuits (Pic. 2) receives a signal from the absolute PMn at 180˚ and its frequency component is converted to an amplitude-modulated oscillation, the carrier oscillation of which is the input PMn signal (3), i.e. is converted into the APMn oscillation:

$$u_{APM}(t) = \left[\begin{array}{l} 1 + M\frac{1}{\pi}\sum_{k=1}^{\infty}k\Omega\left(\frac{1-\cos k\pi}{k} + \cos k\Omega t\right)]\bullet \\ \bullet\cos(\omega t + \gamma(t)90^\circ - 90^\circ) \end{array}\right).$$

This oscillation is fed to the input of the RBM, and at the point of connection of the disconnected circuits and RC-chains, an PMn oscillation is from the secondary winding of the transformer connected to the input of the IFA. From this multiplication at the output of the RBM, an oscillation is formed:

$$U_{RBM}(t) = U_{APM}(t) \cdot U_{PM}(t) =$$

$$= [\left(\begin{array}{l} 1 + M\frac{1}{\pi}\sum_{k=1}^{\infty} k\Omega\left(\frac{1-cosk\pi}{k} + cosk\Omega t\right)] \cdot \\ \cdot \cos(\omega t + \gamma(t)90^{\circ} - 90^{\circ} \end{array}\right) \cdot$$

$$\cdot \cos\left(\omega t + \gamma(t)90^{\circ} + 90^{\circ}\right) =$$

$$= [(1 + M\frac{1}{\pi}\sum_{k=1}^{\infty} k\Omega\left(\frac{1-cosk\pi}{k} + cosk\Omega t\right)] \cdot$$

$$\cdot cos^2\left(\omega t + \gamma(t)90^{\circ} - 90^{\circ}\right) =$$

$$= [(1 + M\frac{1}{\pi}\sum_{k=1}^{\infty} k\Omega\left(\frac{1-cosk\pi}{k} + cosk\Omega t\right)]\frac{1}{2}\left(1 - cos2\omega t\right) =$$

$$= 0,5M\frac{1}{\pi}\sum_{k=1}^{\infty} k\Omega\left(\frac{1-cosk\pi}{k} + cosk\Omega t\right) + HF + C,$$

where M – depth of amplitude modulation ($M = kU_m$), and k – constant component.

The high-frequency components HF are eliminated by the RC chains (Pic. 1), and the constant component C is eliminated by a coupling capacitor.

As a result, the signal from the output of the RBM in the form of a front and a slice of a rectangular pulse (see Pic. 3b) is fed to the input of a trigger (integrator), at the output of which the transmitted rectangular pulses of a digital signal are formed. In fact:

$$U_{TG}(t) = \int 0,5M\frac{1}{\pi}\sum_{k=1}^{\infty} k\Omega\left(\frac{1-cosk\pi}{k} + cosk\Omega t\right)dt =$$

$$= 0,5M\frac{1}{\pi}\sum_{k=1}^{\infty}\frac{1-cosk\pi}{k} sink\Omega t + C_1.$$

Choosing the integration constant $C_1 = 0,5 \cdot (1 + M)$, we obtain the analytical expression of rectangular pulses (4).

The maximum possible interference immunity of the radio communication is provided not only by a single absolute PMn at 180°, considered above, but also by a double absolute PMn at 180°, when two digital signals are transmitted by oscillation of one carrier frequency shifted in phase by 90° using a single absolute PMn in each channel. In this case, the frequency band of only one channel is used, i.e. both noise immunity and frequency efficiency are doubled.

Conclusion.

1. It is analytically shown that the Shannon formula allows connecting three criteria of efficiency, and not two, as it has been previously.

2. The proposed coherent detection of FMn signals, which increases the noise immunity of the railway digital radio communication approximately twice.

3. It is analytically shown that with known coherent detection of signals with absolute PMn at 180°, a reference oscillation with random phase jumps at 180° is formed from them, due to which the «reverse operation» of the coherent detector occurs.

4. The post-detector signal processing of a coherent detector, which eliminates its reverse operation, is proposed. The novelty of the proposal is confirmed by a patent for an invention.

5. It is analytically shown that for the post-detection processing of the detected signal obtained from the absolute phase manipulation at 180°, according to point 4, it is advisable to use a coherent frequency detector on detuned circuits with a trigger connected to it.

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