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Improving Frequency Efficiency of Signals with Absolute Phase Shift Keying by 180º



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ABSTRACT

Communication with mobile objects in railway transport is carried out only by radio, therefore, it is of interest to maximize frequency efficiency and noise immunity of radio communication. This can be achieved using a double absolute phase-shift keying (DPSK) at 180º, but in practice it is still not used due to reverse operation of both coherent detectors of its signals.

The objective of this article is to develop a coherent detector without reverse operation, which allows to use it in practice without reducing its noise immunity. A filter-phase method of forming single-sideband with phase shift keying at 180º (SSB PSK) is proposed, which is equivalent to DPSK in frequency efficiency.

The proposed coherent detector of DPSK signals includes a reference waveform shaper (RWS) shaping the waveform from the input signal, two coherent detectors and introduced post-detector units that eliminate reverse operation of these coherent detectors. Single-lane PSK is formed using clipped speech signal and its quadrature.

Keywords: *transport, communication, coherent detector, reference waveform shaper, pulse regenerator, frequency detector, trigger, noise immunity, radio communication efficiency.*

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Background.

It is known [1, p. 181–187] that a double absolute phase-shift keying (PSK) (DPSK) by 180˚ provides not only the maximum possible noise immunity of radio communication, but also increases the frequency efficiency by 2 times [2] compared to a single one, since two quadrature signals can be transmitted through a single channel. This is equivalent to single-lane absolute PSK at 180˚. An increase in frequency efficiency reduces the main problem of radio communication: the shortage of frequency resources. Communication with mobile objects is carried out only by radio. Therefore, maximizing the immunity of radio communications contributes to improving their safety, in particular, increases train traffic safety. However, absolute DPSK at 180˚ is not used in practice due to the fact that positive discrete pulses are perceived as negative and vice versa. This phenomenon is called «reverse operation of coherent detectors» [2]. To eliminate it, new circuits of coherent detectors were developed that generate the carrier frequency oscillation from the input signal. The first coherent detector was A. A. Pistolkors's detector, developed in 1933, on which its reverse operation was discovered. In 1937 the coherent detector of V. I. Siforov was invented followed in 1945 by D. V. Ageev's detector. In 1954, the coherent detector of Costas, an American scientist in the field of radio communications, began to be used. None of the above detectors eliminated reverse operation, which is confirmed by the corresponding research sources quoted in this article [1–8]. Therefore, it was concluded that reverse operation is unavoidable [10]. In this regard, in 1954 the Russian scientist N. T. Petrovich proposed [1] a relative PSK (RPSK) at 180˚ [2; 10], which practically [1; 9] eliminated reverse operation of the coherent detector [9]. However, relative phase shift keying, from the point of view of noise immunity, is inferior to the absolute one.

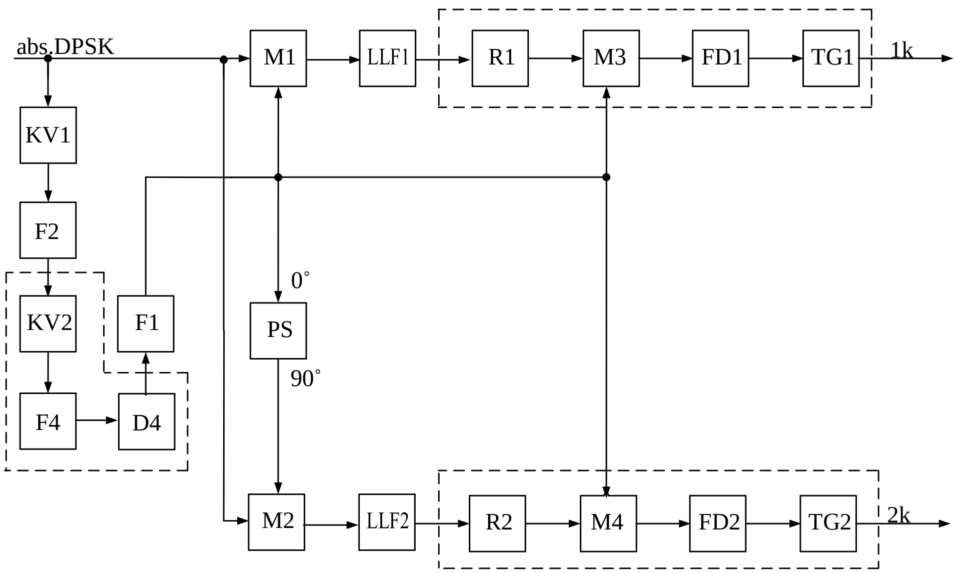
The *objective* of the study is to eliminate reverse operation using various methods.

1. a detector of signals without reverse operation with absolute PSK at 180˚ is proposed;
2. filter-phase method of forming single sideband with phase shift keying at 180˚ (SSB PSK).

1. Block diagram of the developed coherent detector

The block diagram of the developed coherent detector [3] is shown in Pic. 1.

In this picture it is indicated [14]: F – harmonic filters; KV – quadrator; D4 – frequency divider by 4; PS – phase shifter by 90˚; LFF – low frequency filters; M – signal multipliers; R – pulse regenerators; FD – frequency detectors; TG – triggers. Blocks KV1, F2, KV2, F4, D4, F1 form a reference



Pic. 1. Block diagram of a coherent detector.

Table 1

Variants of the difference of functions γ(t)

Variants $(\gamma_2(t) - \gamma_1(t))\bullet\pi/2$	Terms $\sin[(\gamma_2(t) - \gamma_1(t))\bullet\pi/2]$
$(+1 - (+1))\bullet\pi/2 = 0$	$\sin 0 = 0$
$(+1 - (-1))\bullet\pi/2 = \pi$	$\sin\pi = 0$
$(-1 + (+1))\bullet\pi/2 = 0$	$\sin 0 = 0$
$(-1 + (-1))\bullet\pi/2 = -\pi$	$\sin(-\pi) = 0$

Table 2

Variants of the sum of functions γ(t)

Variants $[\gamma_1(t) + \gamma_2(t) - 2]$	Terms $\sin\{2\omega_0 t+[\gamma_1(t)+ \gamma_2(t) - 2]\bullet\pi/2\}$
$(1 + 1-2)\bullet\pi/2=0$	$\sin 2\omega_0 t$
$(1-1 - 2)\bullet\pi/2 = -\pi$	$-\sin 2\omega_0 t$
$(-1 + 1-2)\bullet\pi/2 = -\pi$	$-\sin 2\omega_0 t$
$(-1-1 - 2)\bullet\pi/2 = -2\pi$	$\sin 2\omega_0 t$

waveform shaper (RWS), which eliminates PSK by 180°.

The input signal $u_{in}(t)$ with an absolute DPSK at 180° is fed to some inputs of both coherent detectors directly and in parallel to their other inputs through RWS. This input signal consists of the sum of two quadrature signals with 180° absolute single PSK. Each such signal with a single PSK $u_{psk}(t)$ is equal to the difference between the amplitude-manipulated signal $u_{am}(t)$ and its oscillation of the carrier frequency $u_c(t)$, the amplitude of which is 2 times less than the amplitude of the signal $u_{am}(t)$ [5], that is:

$$\begin{aligned} u_{fm}(t) &= u_{am}(t) - 0{,}5\bullet u_c(t) = \\ &= U_m(t)\bullet u(t)\bullet sin(\omega_0\bullet t+\varphi_0) - 0{,}5U_m\bullet sin(\omega_0\bullet t+\varphi_0) = \\ &= U_m\bullet sin(\omega_0\bullet t+\varphi_0)\bullet[u(t)-0{,}5] = \\ &= 0{,}5\bullet\gamma(t)\bullet U_m\bullet sin(\omega_0\bullet t+\varphi_0), \end{aligned}$$

where manipulating signals are:

$$u(t) = \begin{cases} 1 if\ 0 < t < \tau_0, \\ 0 if\ -\tau_0 < t < 0, \end{cases}.$$

$$\gamma(t) = \begin{cases} +1 if\ 0 < t < \tau_0; \\ -1 if\ -\tau_0 < t < 0. \end{cases}$$

This means that the signals from ASK and PSK [13] are equal to the product of the oscillation of the carrier frequency $u_c(t)$ and of the manipulating signals, respectively, $u(t)$ and $\gamma(t)$ [16]. In this case:

$$u_{psk1}(t) = \gamma_1(t)\bullet U_m sin(\omega_0 t+\varphi_0) =$$

$$= \pm U_m sin(\omega_0 t+\varphi_0) = U_m \sin\left[\omega_0 t + (\gamma_1(t)-1)\bullet\frac{\pi}{2}\right];$$

$$u_{psk2}(t) = \gamma_2(t)\bullet U_m sin\left(\omega_0 t+\varphi_0+\frac{\pi}{2}\right) = \pm U_m sin\left(\omega_0 t+\varphi_0+\frac{\pi}{2}\right) =$$

$$= U_m sin\left[\omega_0 t + (\gamma_2(t)-1)\bullet\frac{\pi}{2}+\frac{\pi}{2}\right] = U_m \cos\left[\omega_0 t + (\gamma_2(t)-1)\bullet\frac{\pi}{2}\right].$$

That is, input signal with absolute DPSK at 180 is described as follows:

$$u_{in}(t) = u_{psk1}(t) + u_{psk2}(t) =$$

$$= U_m\left\{\sin\left[\omega_0 t + (\gamma_1(t)-1)\bullet\frac{\pi}{2}\right] + \cos\left[\omega_0 t + (\gamma_2(t)-1)\bullet\frac{\pi}{2}\right]\right\}.$$

This input signal is squared in KV 1 block. Denoting the first term as *sin*α, and the second as *cos*β, we have at the output of KV signal block:

$$\begin{aligned} u_{kv1}(t) &= u_{in1}(t) + u_{in2}(t) = U_m^{\ 2}(cos á + cos\beta)^2 = \\ &= U_m^{\ 2}(sin^2 á + 2sin á\bullet cos\beta + cos^2\beta), \end{aligned}$$

where:

$$sin^2 á = \frac{1-cos2á}{2} =$$

$$= \frac{1-\cos\left[2\omega_0 t + (\gamma_1(t)-1)\bullet\pi\right]}{2} = \frac{1}{2} - \frac{1}{2}cos2\omega_0 t;$$

$$cos^2\beta = \frac{1+cos2\beta}{2} =$$

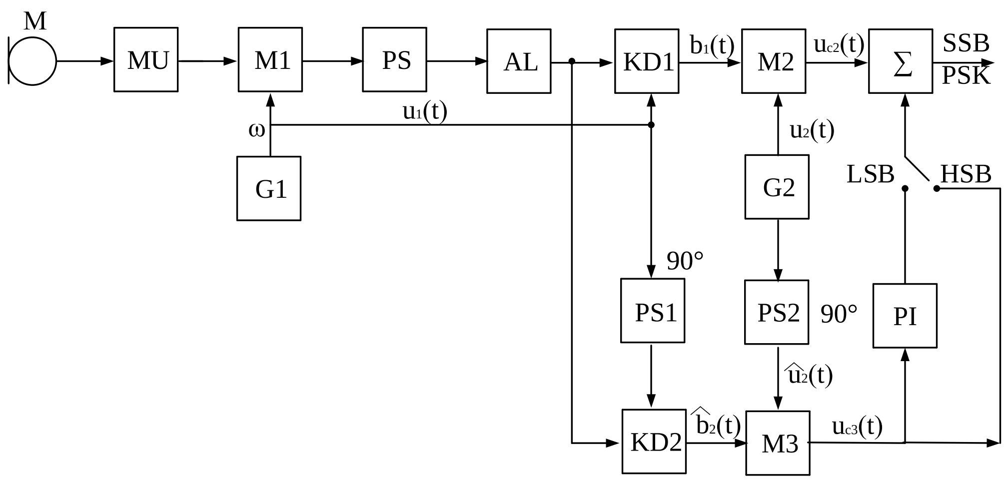
$$= \frac{1+\cos\left[2\omega_0 t + (\gamma_2(t)-1)\bullet\pi\right]}{2} = \frac{1}{2} + \frac{1}{2}cos2\omega_0 t;$$

$$2\sin\alpha\bullet\cos\beta = \sin(\beta-\alpha) + \sin(\beta-\alpha) =$$

$$= \sin\left[\gamma_2(t) - \gamma_1(t)\right]\bullet\frac{\pi}{2} + \sin\{2\omega_0 t + [\gamma_1(t)+\gamma_2(t)-2]\bullet\frac{\pi}{2}\}.$$

As a result at the output KV 1 we get an oscillation:

$$\begin{aligned} u_{kv1}(t) &= U_m^{\ 2}\{\frac{1}{2} - \frac{1}{2}cos2\omega_0 t + \frac{1}{2} + \frac{1}{2}cos2\omega_0 t + sin[\gamma_2(t) - \\ &-\gamma_1(t)]\frac{\pi}{2} + \sin\left\{2\omega_0 t + \left[\gamma_2(t)+\gamma_1(t)-2\right]\bullet\frac{\pi}{2}\right\}\} = \\ &= 1 + sin\left[\gamma_2(t) - \gamma_1(t)\right]\frac{\pi}{2} + \sin\left\{2\omega_0 t + \left[\gamma_2(t)+\gamma_1(t)-2\right]\bullet\frac{\pi}{2}\right\}. \end{aligned}$$



Pic. 2. Circuit of filter-phase method of RWS.

Although $\gamma_1(t) = \pm 1$ and $\gamma_2(t) = \pm 1$, but their sum and difference may be different. Therefore, we will compile tables considering the sums and differences $\gamma_1(t)$ and $\gamma_2(t)$.

According to Table 1 sine of the difference $(\gamma_2(t) - \gamma_1(t)) \bullet \pi/2$ is 0, and according to Table 2, the sum $[\gamma_1(t)+\gamma_2(t) - 2] \bullet \pi/2$ does not eliminate PSK by 180° (second column of Table 2). Therefore, the second quadrator KV2 is introduced, at the input of which the signal from the output of KV1 comes through a capacitor and a second harmonic filter F2. At the output of KV2 there is a signal:

$$u_{kv2}(t) = \left(\pm U_m^2 \sin(2\omega_0 t)\right)^2 = U_m^4 sin^2 2\omega_0 t =$$
$$= U_m^4 \bullet \frac{1 - cos4\omega_0 t}{2} = 0,5U_m^4 - 0,5U_m^4 cos4\omega_0 t.$$

It can be seen that the signal from KV2 does not contain the manipulating signals $\gamma_2(t)$ and $\gamma_1(t)$, and therefore it is fed to the frequency divider by a factor of 4, which is a double D-trigger with an amplitude-limiting amplifier at its input. From the direct output of the first D-trigger, an oscillation $u_0(t) = U_m cos\omega_0 t$ is fed through the resonant circuit of the first harmonic with a zero initial phase to the second input of the multiplier M1 of the coherent detector of the in-phase channel, and from the direct output of the second D-trigger, a quadrature oscillation

$$u_0(t) = U_m \cos(\omega_0 t + \frac{\pi}{2})$$

is supplied through the resonant circuit of the first harmonic to the second input of the multiplier M2 of the coherent detector of the quadrature channel. At the output of the multipliers M1 and M2, oscillations are obtained:

$$u_{c1}(t) = u_{in}(t) \bullet u_0(t) = U_m\{cos\left[\omega_0 t + (\gamma_1(t) - 1) \bullet \frac{\pi}{2}\right] +$$
$$+ cos\left[\omega_0 t + (\gamma_2(t) - 1)\frac{\pi}{2} + \frac{\pi}{2}\right]\} U_m cos\omega_0 t =$$
$$= 0,5U_m^2 \left\{cos(\gamma_1(t) - 1)\frac{\pi}{2} + cos\left[(\gamma_2(t) - 1) \bullet \frac{\pi}{2} + \frac{\pi}{2}\right]\right\} +$$
$$+ high\ frequency = 0,5U_m^2[\pm 1 + 0] + high\ frequency;$$

LFF (low frequency filter) at the output M1 and M2 eliminates high frequency oscillations, leaving only transmitted digital signals $\gamma_1(t) = \pm 1$ and $\gamma_2(t) = \pm 1$.

But for absolute DPSK by 180 we get the reference oscillation $u_0(t) = \pm\sqrt[4]{u_{in}^4(t)}$. The ± signs are equally probable and, therefore, interference or other processes in the receiver can cause a change in the signal phase by 180° for $u_0(t)$, which is the source of the reverse operation of coherent detectors, since the detected pulses are in phase with $u_0(t)$. The reference oscillation $u_0(t)$ passes through the resonant circuits F2, F4, F1, and therefore turns out to be practically free of interference, and the coherently detected pulses do not contain the quadrature component of the interference. They can be regenerated by amplification-limiting in amplitude and thereby additionally suppress the noise accompanying them. Since the reference oscillatio$u_0(t)$ and the detected pulses are in phase, their multiplication by each other excludes a change in the signal phase by 180°, as in the multiplication of harmonic oscillations:

$$u_n(t)=b(t)\bullet u_c(t)=U_\Omega\cos\left(\Omega t+180^\circ\right)U_c\cos\left(\omega t+180^\circ\right)=$$
$$=0{,}5U_\Omega U_c\left[\cos\left(\omega-\Omega\right)t+\cos\left(\omega+\Omega\right)t\right].$$

The multiplication of alternating pulses with oscillation of the carrier frequency, as shown at the beginning of the article, is an absolute one-time PSK by 180°, which in this case is practically free of noise and distortion. Since DPSK signals are already separated from each other, it is possible to multiply $\gamma_1(t)$ and $\gamma_2(t)$ only with $u_0(t)$ as shown in Pic. 1. Such a signal can be detected incoherently without changing the noise immunity of radio communication due to the absence of interference. In this case, it is proposed to use a trigger and a frequency detector at the output, since the phase $\varphi(t)$ and the angular frequency $\omega(t)$ are related by the ratio: $\omega(t) = d\varphi/dt$, and its reverse operation cannot be eliminated in principle.

Such a detector had already been proposed instead of the coherent Pistolkors detector back in 1951 [2], however, due to interference at its input, it did not find practical application. In the considered case, interference at its input is practically absent and therefore it can be used.

2. One single-sideband absolute PSK at 180°, equivalent in frequency efficiency to double PSK at 180°

This single-sideband absolute PSK (SSB PSK) at 180° was developed by the authors [17], its structural diagram is shown in Pic. 2.

The indices in the diagram [18] are: BF – band-pass filter; Σ – adder; M1, M2, M3 – signal multipliers; G1, G2 – generators; KD1, KD2 – coherent detectors; PS1, PS2–90° phase shifters; AL – amplitude limiter; PI – phase inverter.

At the output of KD1 there is a clipped [12] speech signal $b_1(t)$, and at the output of KD2 there is a quadrature speech signal. In this case, taking into account the fact that the signals are periodic, we have:

$$b_1(t)=\frac{2}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{sink\text{Ù}t}{k};$$

$$b_2(t)=\frac{2}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{cosk\text{Ù}t}{k}.$$

At the output M2 the function is formed:

$$U_{n2}(t)=b_1(t)\bullet u_{c2}(t)=\left(\frac{2}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{sink\text{Ù}t}{k}\right)\bullet U_c sin\omega t=$$
$$=\frac{U}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{\cos\left(\omega-k\text{Ù}\right)t}{k}-\frac{U}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{\cos\left(\omega+k\text{Ù}\right)t}{k}.$$

And at the output of M3 there is the oscillation:

$$U_{n3}(t)=\widehat{b_2}(t)\bullet\widehat{u_{c2}}(t)=\left(\frac{2}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{cosk\text{Ù}t}{k}\right)cos\omega t=$$
$$=\frac{U}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{\cos\left(\omega-k\text{Ù}\right)t}{k}+\frac{U}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{\cos\left(\omega+k\text{Ù}\right)t}{k}.$$

The adder Σ contains the signal:

$$u_{ó}(t)=u_{n2}(t)+u_{n3}(t)=\frac{2U}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{\cos\left(\omega-k\text{Ù}\right)t}{k},$$

which lower sideband (SSB PSK at 180°).

When subtracting (when $b_2(t)$ passes through PI block):

$$u_{ó}(t)=u_{n2}(t)-u_{n3}(t)=\frac{2U}{\pi}\sum_{k=1,3,5...}^{\infty}\frac{\cos\left(\omega+k\text{Ù}\right)t}{k},$$

There is higher sideband (SSB PSK at 180°).

With SSB PSK, the interference is 2 times less than with two sidebands, and therefore, with their equal amplitudes, the noise immunity of radio communication with SSB PSK will be 180° greater than with two side bands [7].

Conclusions.

1. A circuit of a coherent signal detector was proposed [11], with the exclusion of its reverse operation, which makes it possible to implement DPSK at 180° in practice. The novelty of the proposal is confirmed by the Russian invention patent [11].

2. It is shown that for formation of the reference oscillation $u_0(t) = U_m cos\omega_0 t$ according to the input signal with absolute DPSK at 180°, the second quadrator KV2 is required, and a frequency divider should need to divide frequency not by 2, but by 4.

3. It is shown [5] that the sources of reverse operation of a coherent detector with absolute DPSK, when +1 is converted to -1 and vice versa, are random and unremovable jumps in the phase of the reference

oscillation obtained from the input signal [14] with absolute DPSK at 180°, which are in phase with coherently detected pulses.

4. It is proposed to eliminate phase jumps by 180° for $u_0(t)$ and for detected alternating pulses by multiplying them by each other, which forms an absolute PSK by 180° practically without noise and distortion. There is no interference for $u_0(t)$ due to its filtering by resonant circuits of the second, fourth and first harmonics. There is no interference for detected pulses either due to coherent detection and their regeneration by amplification-limiting.

5. It is proposed to use a frequency detector with a trigger (integrator) at its output as a detector of multiplied signals according to clause 4, since the phase $\varphi(t)$ and the angular frequency $\omega(t)$ are related by the ratio: $\omega(t) = d\varphi/dt$. Such a detector had already been proposed [19] as the main one (instead of the Pistolkors detector) back in 1951, but did not find application due to interference at its input. In our case, there is practically no interference and distortion at its input.

6. To increase the frequency efficiency by 2 times, the authors proposed single-band absolute PSK at 180° operated with filter-phase method with clipping, which is equivalent by this indicator to absolute DPSK at 180°. The Russian invention patent was obtained [17].

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