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QUICK-ACTING, CONTROLLABLE PHASE SHIFTER FOR PHASE ANGLE ADJUSTMENT IN RADIO SIGNALS

Subject and Purpose. *The phase shifters intended for controlling the phase of radio signals are widely used in ultra-high frequency technology, communication systems, radar, and a variety of measuring and special-purpose radio equipment. Designers of phased array antennas face the need of providing for broad beam scanning angles and high antenna gains, which leads to the necessity of greatly increasing the number of array elements, each of which is to be controlled by a phase shifter. Therefore, the development and creation of quick-acting phase shifters is an urgent task. The purpose of this work is to develop high-speed, controllable phase shifters for performing phase angle adjustments and thus provide, at an acceptable cost, for desirable parameters of phased antenna arrays, frequency stabilizing systems of magnetrons, etc.*

Methods and Methodology. *The functional diagram of the proposed quick-acting, controllable phase shifter has been analyzed mathematically and modelled numerically.*

Results. *The controllable phase shifter can be successfully implemented through the use of two parallel-connected resonators at the input of a specific receiver. Analysis of the signal amplitude and phase at the output of the phase shifter in dependence on the values at the input confirms the possibility of adjusting the phase of the output signal over a wide range of angles.*

Conclusions. *A design concept of quick-acting, controllable phase shifters for producing adjustable phase angles has been developed. The device can be employed in phased antenna arrays or frequency stabilizing systems as a means of improving their operation parameters and reducing their cost at that.*

Keywords: *phased array antenna, controllable phase shifter, resonator, phase angle.*

Introduction

Currently, electronically steerable phased array antennas (PAA) are widely used in a variety of radar systems. The necessity of providing for a wide range of scanning angles and gain values leads to an increase in the number of antenna elements, from a few thousand to several tens of thousands. Each of the elements requires a high-speed controllable phase shifter to perform formation of the phase front of the wave emitted by the antenna. In the case of employing traditional beam control systems, like e.g., such based on *p–i–n* diode phase shifters, the cost of the phased arrays could reach such values that would make them uncompetitive compared with other types of antennas. Therefore, an important and urgent task is the search and creation of new types of

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phase shifters that may have the necessary parameters, while maintaining an acceptable cost level. [1].

The development of pulsed magnetron transmitters that would allow adjustment and stabilization of the frequency generated is another area of radar technology that requires quick-acting and controllable phase shifters. As is known [2], changes in the generated frequency and the pulse-to-pulse jumps of the initial phase of oscillations limit the efficiency of magnetron-based transmitters, and may limit the possibility of creating a radar with modern program control and high speed of information processing. This is especially true for magnetron generators with two output pins, i.e. the active and the reactive one. By changing the parameters of the reactive load, say, with the aid of a controllable phase shifter, it should be possible to effectively re-adjust the operating frequency of the magnetron within small limits [3]. In addition, to implement chirp modulation and apply a short-circuiting plug connected to the reactive energy output, it would equally be necessary to resort to a controllable high-speed phase shifter. The use of ferroelectric phase shifters faces the same problems as in the case of the PAA. In addition, their parameters are strongly temperature-dependent [4].

This work is aimed at developing a concept design of quick-acting controllable phase shifters for producing adjustable phase angles to ensure necessary values of operating parameters for phased array antennas or frequency stabilizing systems of magnetrons, while reducing their cost.

The concept of constructing a controllable phase shifter

The writers [5] considered the conditions for establishing electromagnetic compatibility of the passive and the active channel in complex radars through the use of a special filter. That involved two parallel-connected resonators at the input of the radar receiver. The resonators were to be adjusted in such a way that the signal from the transmitter (that jammed the radiometer) would be suppressed, and hence could not affect signals within the operating frequency bands of the radiometer. This was achieved by splitting the signal from the transmitter into two and feeding these through two parallel-connected resonators with corrective circuits. Next, the two channels were summed up in antiphase at the radiometer input (the resulting level of suppression depended upon the accuracy of settings and could reach values above 100 dB).

Apparently, such a filter can be used in other applications, like a quick-acting controllable phase shifter. To consider the phase shifter concept proposed, let us turn to Fig. 1.

Shown in the Figure is the vector diagram of signal addition for the voltages that have passed through the parallel-connected resonators. The initial state of the signal vectors $\overrightarrow{AB} = \vec{a}$ and $\overrightarrow{AC} = \vec{b}$ is that they are equal in length, while their initial phase angles are of opposite signs but equal magnitudes, $|\varphi_1| = |\varphi_2| = \varphi.$ Accordingly, the phase angle α of the total signal (i.e., of the vector $\overrightarrow{AD}$) is equal to zero.

The length of the total signal vector $\overrightarrow{AD}$ can be determined from the triangle ABD,

$$\left|\overrightarrow{AD}\right|^2 = \left|\overrightarrow{AB}\right|^2 + \left|\overrightarrow{BD}\right|^2 + 2\left|\overrightarrow{AB}\right|\cdot\left|\overrightarrow{BD}\right|\cdot\cos(\pi - 2\varphi).$$

Upon denoting $\overrightarrow{AD} = \vec{U}_\Sigma,$ $\overrightarrow{AB} = \vec{a},$ and $\overrightarrow{BD} = \vec{b}'$ we obtain

$$\left|\vec{U}_\Sigma\right|^2 = \left|\vec{a}\right|^2 + \left|\vec{b}\right|^2 + 2\left|\vec{a}\right|\left|\vec{b}\right|\cdot\cos 2\varphi. \qquad (1)$$

In view of $\left|\vec{a}\right| = \left|\vec{b}\right|$ this becomes

$$\left|\vec{U}_\Sigma\right|^2 = 2\left|\vec{a}\right|^2 (1 + \cos 2\varphi).$$

Now consider the case where one of the vectors, for example $\vec{b},$ will change its length without changing the angle $\varphi_2.$ Let us determine how does the total signal $\vec{U}_\Sigma$ change in this case. To determine numerically the dependence of the total signal $\vec{U}_\Sigma$ upon changes in the magnitude of the vector $\vec{b},$ we will change it by a factor k, where $k = 0 \div 1$. The value of the variable vector $\vec{b}$ will be denoted as $\vec{b}' = k\vec{b}.$ Substituting these values into Eq. (1) we obtain

$$\left|\vec{U}'_\Sigma\right|^2 = \left|\vec{a}\right|^2 + k^2\left|\vec{b}\right|^2 + 2\left|\vec{a}\right|k\left|\vec{b}\right|\cdot\cos 2\varphi,$$

or, with account of $\left|\vec{a}\right| = \left|\vec{b}\right|$

$$\left|\vec{U}_\Sigma\right|^2 = \left|\vec{a}\right|^2 + k^2\left|\vec{a}\right|^2 + 2\left|\vec{a}\right|^2 k\cdot\cos 2\varphi,$$

$$\left|\vec{U}_\Sigma\right|^2 = \left|\vec{a}\right|^2 (1 + k^2 + 2k\cdot\cos 2\varphi), \text{ whence}$$

$$\left|\vec{U}_\Sigma\right| = \left|\vec{a}\right|\sqrt{1 + k^2 + 2k\cdot\cos 2\varphi}. \qquad (2)$$

Making use of Eq. (2) we can define a functional dependency $\left|\vec{U}_\Sigma\right| = f(k)$ for $\varphi = const$. Fig. 2 shows the $\left|\vec{U}_\Sigma\right| = f(k)$ dependence, as given by Eq. (2), for φ = 15°, 30°, 45°, 60°, 70°, 80° and 90°.

As can be seen from Fig. 1, when the magnitude of one of the vectors $\vec{b}' = k\vec{b}$ changes, the phase shift angle α of the total vector $\vec{U}_\Sigma$ changes from $\alpha = 0$, which value is assumed with $k = 1$ and equal lengths a and b of the respective vectors, to some given $\varphi = const$ with $k = 0$.

Now let us follow the dependence of the phase angle α of the total voltage vector $\vec{U}_\Sigma$ on variations of one of the vectors $\left|\overrightarrow{ED}\right|$ i.e. the $\alpha = f(k)$ dependence. To do that, we appeal to Fig. 1 again. From the triangle AED, let us determine the length of $\left|\overrightarrow{ED}\right|$ – i.e., the quantity by which the length of $\overrightarrow{BD} = \vec{b}$ is reduced,

$$\left|\overrightarrow{ED}\right|^2 = \left|\overrightarrow{AE}\right|^2 + \left|\overrightarrow{AD}\right|^2 - 2\left|\overrightarrow{AE}\right| \cdot \left|\overrightarrow{AD}\right| \cdot \cos\alpha, \qquad (3)$$

whence

$$\cos\alpha = \frac{\left|\overrightarrow{AE}\right|^2 + \left|\overrightarrow{AD}\right|^2 - \left|\overrightarrow{ED}\right|^2}{2\left|\overrightarrow{AE}\right| \cdot \left|\overrightarrow{AD}\right|}.$$

$$\left|\overrightarrow{ED}\right| = \left|\vec{b}\right| - \left|\vec{b}'\right| = \left|\vec{b}\right| - k\left|\vec{b}\right| = \left|\vec{b}\right|(1-k),\ \left|\vec{b}\right| = \left|\vec{a}\right|,$$

and $\left|\overrightarrow{ED}\right|^2 = \left|\vec{b}\right|^2 (1-k)^2$.

From the triangle ABD we obtain

$$\left|\overrightarrow{AD}\right|^2 = \left|\overrightarrow{AB}\right|^2 + \left|\overrightarrow{BD}\right|^2 + 2\left|\overrightarrow{AB}\right| \cdot \left|\overrightarrow{BD}\right| \cdot \cos(\pi - 2\varphi).$$

Taking into account that $\left|\overrightarrow{AB}\right| = \left|\overrightarrow{BD}\right| = \left|\vec{a}\right|$, we can write

$$\left|\overrightarrow{AD}\right|^2 = 2\left|\vec{a}\right|^2 (1 + \cos 2\varphi),$$

and $\left|\overrightarrow{AD}\right| = \left|\vec{a}\right|\sqrt{2(1+\cos 2\varphi)}$.

Now let us evaluate the length of vector $\overrightarrow{AE}$, which is a sum of the vectors representing the signals that have passed through the resonators, with account of the reduced length of one of them, namely $\left|\overrightarrow{BE}\right| = \left|\vec{b}\right| k = \left|\vec{a}\right| k$.

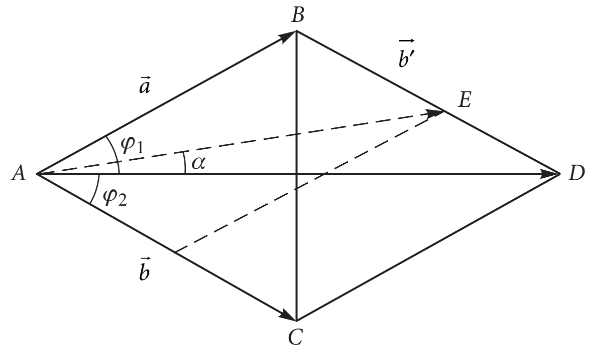

Fig. 1. Vector diagram of signal addition for the signals passing through the parallel-connected resonators

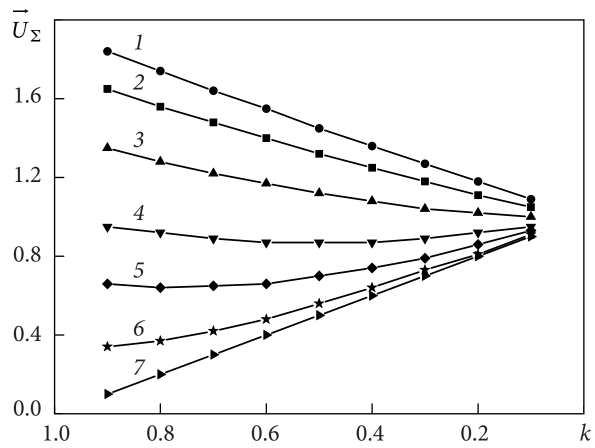

Fig. 2. The $\left|\vec{U}_\Sigma\right| = f(k)$ dependences plotted for φ = 15°, 30°, 45°, 60°, 70°, 80° and 90°, curves *1* to *7*, respectively

$$\left|\overrightarrow{AE}\right|^2 = \left|\overrightarrow{AB}\right|^2 + \left|\overrightarrow{BE}\right|^2 - 2\left|\overrightarrow{AB}\right| \cdot \left|\overrightarrow{BE}\right| \cdot \cos(\pi - 2\varphi),$$

$$\left|\overrightarrow{AE}\right|^2 = \left|\vec{a}\right|^2 + \left|\vec{a}\right|^2 k^2 + 2\left|\vec{a}\right|^2 k \cdot \cos 2\varphi,$$

$$\left|\overrightarrow{AE}\right|^2 = \left|\vec{a}\right|^2 (1 + k^2 + 2k \cdot \cos 2\varphi),$$

$$\left|\overrightarrow{AE}\right| = \left|\vec{a}\right|\sqrt{(1 + k^2 + 2k \cdot \cos 2\varphi)}.$$

Upon substituting the values obtained for $\left|\overrightarrow{AE}\right|$, $\left|\overrightarrow{AD}\right|$ and $\left|\overrightarrow{ED}\right|$ into Eq. (3) we arrive at

$$\cos\alpha = \frac{\sqrt{(1+k^2+2k\cdot\cos 2\varphi)}}{2\sqrt{2(1+\cos 2\varphi)}} + \frac{\sqrt{(1+\cos 2\varphi)}}{\sqrt{2(1+k^2+2k\cdot\cos 2\varphi)}} - \frac{(1-k)^2}{2\sqrt{2(1+k^2+2k\cdot\cos 2\varphi)(1+\cos 2\varphi)}}.$$

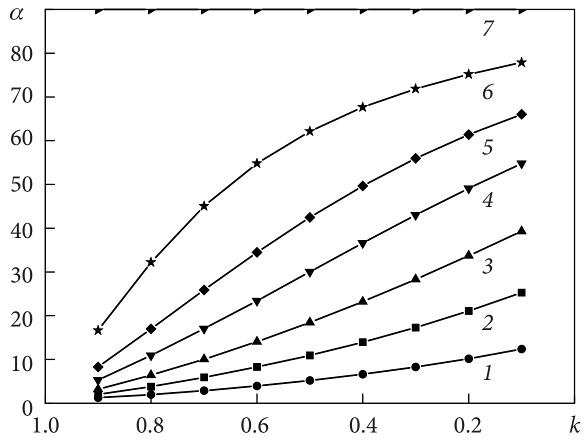

Fig. 3. The $\alpha = f(k)$ dependences plotted for φ = 15°, 30°, 45°, 60°, 70°, 80° and 90°, curves *1* to *7*, respectively

Bringing the latter expression to a simpler form we have

$$\cos\alpha = 0.707\frac{(k+1)\sqrt{1+\cos 2\varphi}}{\sqrt{(1+k^2+2k\cdot\cos 2\varphi)}},$$

wherefrom the adjustable phase shift angle α is

$$\alpha = \arccos\left(0.707\frac{(k+1)\sqrt{1+\cos 2\varphi}}{\sqrt{(1+k^2+2k\cdot\cos 2\varphi)}}\right), \qquad (4)$$

with $k = \left|\vec{b'}\right| / \left|\vec{b}\right|$ assumes values within $k = 0 \div 1$; $\varphi = |\varphi_1| = |\varphi_2|$ is the initial phase shift angle, which determines the variation range for the angle α (0°÷90°). The angles φ_1 and φ_2 are equal in magnitude and opposite in sign (see Fig. 1). This state is achieved through a proper adjustment of the resonators, such as to provide for a desired phase shift at the operating frequency. Fig. 3 shows the $\alpha = f(k)$ dependences (as obtained from Eq. (4) for a set of constant values of φ, namely φ = 15°, 30°, 45°, 60°, 70°, 80° and 90°. It should be noted that by changing the magnitude of another vector than $\vec{b}$, say for instance, vector $\vec{a}$, while leaving the vector $\vec{b}$ unchanged, the phase shift angle α can be made to have an opposite sign. This is a way to create an adjustable phase shift angle over the range from a specified $+\varphi$ to $-\varphi$, with a transition through zero, and to increase the dynamic range of phase control.

As a means for changing the level of signals at the resonator inputs and thus enabling control over the phase shift created, one might suggest some of the existing high-speed controllable attenuators, e.g., the field-effect transistor-based monolithic microwave attenuators, or other [6].

A change in the frequency of the input signal passing through the phase shifter (which involves the parallel-switched resonators) will lead to a change in the output signal amplitude U_Σ and magnitude of its phase shift angle α. Let us assume the parameters of the two resonators and their amplitude-frequency characteristics (AFC) to be identical and variations in the signal frequency reasonably small. In this case, should we change the frequency in one direction, one of the signals would increase in amplitude, since the frequency of that signal would get closer to the resonant frequency corresponding to the maximum transmission factor of the first resonator. Instead, the other signal would decrease in amplitude as its frequency would move away from the resonant frequency corresponding to the maximum transmission factor of the second resonator. Taking into account the previously accepted conditions, we assume the changes in amplitude of the two signals to be equal in magnitude and opposite in sign. ($\Delta_1 = \Delta_2 = \Delta$, where Δ_1 is the increment in the amplitude of the first signal, and Δ_2 is the decremental step in the amplitude of the second one). Thus, $\left|\vec{b'}\right| = \left|\vec{b}\right| - \Delta_2 = \left|\vec{b}\right| \cdot k$, $k < 1$, $\left|\vec{a'}\right| = \left|\vec{a}\right| - \Delta_1 = \left|\vec{a}\right| \cdot k_2$, $k_2 > 1$, $\Delta = 1 - k$; then

$$k_2 = 1 + \Delta = 1 + 1 - k = (2-k),$$
$$\text{and } \left|\vec{a'}\right| = \left|\vec{a}\right| \cdot (2-k). \qquad (5)$$

Now let us determine the total amplitude of the signals following a change in their frequency. At the same time, we will not take into account the changes in the phase shift angles ($\Delta\varphi_1$ and $\Delta\varphi_2$) of each of the signals that occur in accordance with the phase-frequency characteristics (PFC) of the resonators.

In accordance with Fig. 1 and taking into account Eq. (5) we may write

$$\left|\vec{U}_\Sigma\right|^2 = \left|\vec{a}\right|^2 (2-k)^2 + k^2 \left|\vec{b}\right|^2 + $$
$$+ 2\left|\vec{a}\right|\left|\vec{b}\right|(2-k)k\cos 2\varphi.$$

In the initial state we had $\left|\vec{a}\right| = \left|\vec{b}\right|$, hence

$$\left|\vec{U}_\Sigma\right|^2 = \left|\vec{a}\right|^2 \left[(2-k)^2 + k^2 + 2(2-k)k\cos 2\varphi\right]. \qquad (6)$$

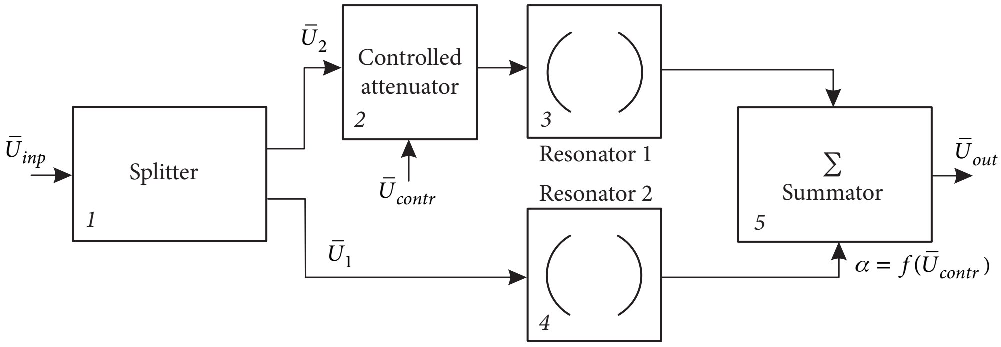

Fig. 4. Functional diagram of the high-speed phase shifter

Simplifying Eq. (6) we can obtain

$$U_\Sigma = 1.414\left|\vec{a}\right|\sqrt{2+(k^2-2k)(1-\cos 2\varphi)}.$$

Fig. 4 shows the functional circuit of a quick-acting phase shifter intended for creating an adjustable phase shift angle of radio signals: *1* stands for a signal splitter; *2* is a high-speed control-lable attenuator; *3* and *4* are Resonators 1 and 2, respectively, and *5* stands for a summator.

The controllable phase shifter operates as follows. The radio signal $\bar{U}_{inp}$ whose phase shift angle is to be adjusted is fed to the input of the signal splitter *1* which divides the input power by half. The signals in the output channels of the splitter are in phase and equal in amplitude. The signal splitter can be implemented with a waveguide, coaxial- or strip line, etc., depending on the frequency range and tasks to be solved. Next, the signal from one of the outputs of the splitter (position *1* in Fig. 4) is fed to the input of the controllable attenuator *2*. The device employed in the capacity of the high-speed controllable attenuator *2* can be, e.g., a *p–i–n* diode-based attenuator with either a smoothly or stepwise varied attenuation. The latter is controlled, depending on the specific task, by either a pulsed or analog voltage U_{contr}. The principal demands to the attenuator are the high operation speed and a relatively low cost. The output signal of the controllable attenuator *2* is fed to the first input of Resonator 1 (position *3* in the Figure), while Resonator 2 (position *4* in the Figure) receives at its input the d.c. voltage $\bar{U}_1$ from the second output of the Splitter (position *1*). The Resonators 1 and 2 (designated as positions *3* and *4* in the Figure) are identical and have similar amplitude-frequency and phase-frequency characteristics (the AFC and PFC, respectively). They can be implemented as lumped-element circuits for the frequencies corresponding to meter and decimeter wavelenghts, while at higher frequencies cavity resonators shall be used. Resonator 1 (position *3* in Fig. 4) is to be so tuned that its output signal at the operating frequency *f* should have, in accordance with its PFC, a phase shift angle $-\varphi_1 = \mathrm{arctg}\,\xi_0$, where $\xi_0 = Q_0 \dfrac{2(f-f_0)}{f_0} = Q_0 \dfrac{2\Delta f}{f_0}$ is the normalized frequency; f_0 the Resonator's resonance frequency, and Q_0 the Resonator's *Q* factor. Accordingly, Resonator 2 (position *4* in Fig. 4) must be tuned to the operating frequency *f*, so that its output signal would have the phase angle $\varphi_2 = |\varphi_1| = \mathrm{arctg}\,\xi_0$ of an opposite sign as compared with φ_1, but of same magnitude. In this scenario, the resulting phase shift angle of the total signal will be $\alpha = 0$, and always so, no matter which the individual values were. It is only important that the equality $|\varphi_1| = |\varphi_2|$ should hold, and the attenuator (position *2*) possessed zero attenuation, i.e. the levels of $\bar{U}_1$ and $\bar{U}_2$ were equal. The magnitude of the phase shift angle $|\varphi_1| = |\varphi_2|$ determines the dynamic range and maximum value of the phase angle $\alpha_{\max} = \varphi_1$ being produced. Further on, the signals that have passed through the resonators (positions *3* and *4* in Fig. 4) are fed into the Summator *5* to be subjected to vectorial addition (see Fig. 1) The phase shift angle created at the output of the phase shifter will be regulated by the control voltage U_{contr}, within specified ranges from 0 to φ, via the attenuator *2*. In case it would be necessary to obtain a phase shift angle of an opposite sign, the input of the attenuator *2* should be connected to the other output of the splitter *1*, while its output to the input of Resonator 2 (position *4* in the Figure). Then the phase shift angle α at the output of the phase shifter could be adjusted from 0 to $-\varphi$. That would allow one to expand the dynamic range of phase shift adjustment as the phase angle would be variable between $\alpha = \pm\varphi$.

There are other methods for obtaining the dynamic range like $\alpha = \pm\varphi$, which actually require special consideration which is beyond the scope of this work.

Conclusions

The concept of constructing a quick-acting, controllable phase shifter of radio signals that has been presented can be applied for creating phased antenna arrays of reduced cost and improved operating parameters. The experimental studies carried out with meter wavelength lumped-parameter resonators have confirmed correctness of the concept developed and of the theoretical assessments.

Another application of the proposed device could be an automatic frequency control system for pulsed magnetrons possessing two output channels and capable of changing reactive load parameters.

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ШВИДКОДІЮЧИЙ КЕРОВАНИЙ ФАЗООБЕРТАЧ
ДЛЯ РЕГУЛЮВАННЯ ФАЗОВОГО КУТА РАДІОСИГНАЛІВ

Предмет і мета роботи. Фазообертачі, що призначено для керування фазою радіосигналів, широко застосовуються в техніці надвисоких частот, системах зв'язку, радіолокації, різноманітній вимірювальній і спеціальній радіоапаратурі. Розробники фазованих антенних решіток (ФАР) стикаються з необхідністю забезпечення великих кутів коливання променя та коефіцієнту посилення антени. Це призводить до значного збільшення кількості елементів ФАР, кожен з яких має керуватися фазообертачем. З огляду на це створення швидкодіючих фазообертачів є актуальним завданням. Метою роботи є розробка швидкодіючих керованих фазообертачів для створення регульованого кута зсуву фаз, котрі забезпечували б, за прийнятною вартістю, необхідні параметри ФАР або систем стабілізації частоти магнетронів тощо.

Методи і методологія. Для дослідження функціональної схеми швидкодіючого фазообертача використано методи математичного аналізу та числового моделювання.

Результати. Керований фазообертач може бути реалізований шляхом паралельного підключення двох резонаторів на вході належного приймального пристрою. Аналіз залежностей амплітуди та фази сигналу на виході фазообертача від значень цих величин на вході підтвердив можливість регулювання фази вихідного сигналу в широкому діапазоні кутів.

Висновки. Розроблено концепцію побудови швидкодіючих керованих фазообертачів для створення регульованого кута зсуву фаз. Пропонований пристрій може бути використаний при створенні фазованих антенних решіток або систем стабілізації частоти як знаряддя, що поліпшує їхні параметри та знижує вартість.

***Ключові слова**: фазовані антенні решітки, керований фазообертач, резонатор, кут зсуву фаз.*