may be written down following the Fig. 5, whence the value

$$Z_{\text{ВЫХ.1}} = U_{\text{ВЫХ.1}} / I_0 = \frac{X_{12.1} + Z_{\text{ВХ.}\Sigma}}{1 + S_1 Z_{\text{ВХ.}\Sigma}}, \qquad (4)$$

is determined, where $Z_{\text{ВХ.}\Sigma}=Z_C \| X_{11.1}$ (Fig. 5).

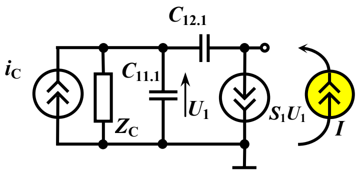

Fig. 5. *The part of the circuit of amplifier loaded by the second cascode step*

The value $Z_{\text{ВЫХ.1}}$ may be judge by (4). At high frequencies when $Z_{\text{ВХ.}\Sigma}$ decreases greatly, $Z_{\text{ВЫХ.1}}$ is comparatively high, about $|X_{12.1}|$. At low frequencies, it equals approximately $(1/S_1)(X_{12.1}/Z_{\text{ВХ.}\Sigma})$.

In the circuit in Fig. 3 the resistance of the elements connected in parallel denoted by $Z_{\text{ВЫХ.1}}$ and $C_{11.2}$ having rather high resistances is indicated by the variable $Z_\|$. Bridging the output resistance by the capacitance $C_{11.2}$ makes no difference. Value $Z_\|$ remains of the same order that $Z_{\text{ВЫХ.1}}$. Product $S_2Z_\|$, met in formulas (2) and (3), is much more than a unit while $S_2X_{12.1}>>1$, i.e. while frequency $f << S_2/2\pi C_{12.1}$. So, applying the field-effect transistors of the type KP341 this product can not be less than the ratio C_{11}/C_{12}, which equals approximately five.

Summary. The relative influence of active elements on noises of video amplifier cascode circuit at resistive load and application of active element of as the cascade dynamic load was considered. It was shown that in both cases the second transistor of the cascode circuit contributes insignificantly into the amplifier noises in comparison with the first one (<10 %). Contribution of noises of the dynamic load active element exceeds considerably the contribution of the traditional cascade resistive load and doubles practically in power the noises conditioned by the first active element at uniformity of all three active elements.

REFERENCES

1. Dobretsov A.I. Wide-band amplifier with a high gain / A.I. Dobretsov, V.A. Karzhavin, Yu.A. Tuflin // Devices and experimental technique. – 1978. – № 1. – P. 88–90.
2. Shustov M.A. Practical circuit technology. In 5 books. B. 1. 450 useful circuits for radio amateurs: collection. – Moscow: Alteks-A, 2003. – 352 p. – P. 44–53. – Bibliog.: p. 336-351. – ISBN 5-94271-002-3.
3. Van der Zil A. Noise. Sources, description, measurement / Translated from English by V.N. Kuleshov and D.P. Tsarapkin; ed. by A.K. Naryshkin. – Moscow: Sovetskoe Radio, 1973. – 228 p.

Received on 07.12.2006

UDC 621.372.57

ACTIVE FREQUENCY-DIVIDING FILTERS

V.P. Dovgun, V.V. Novikov

Siberian federal university, Krasnoyarsk
E-mail: vdovgun@emf.krgtu.ru

The method of synthesis of analog frequency-dividing devices on the basis of wave filters has been offered. Various variants of configurations of such filters are considered. It is shown that frequency-dividing filters synthesized according to the offered method, have the minimal order.

Processing and transmitting signals the necessity of using the frequency dividing devices intended for dividing signal spectrum into nonoverlapping parts often occurs. Such devices are called directional filters or multiplexers [1, 2]. The questions of directional filter application in radio systems and devices are examined in detail in monograph [1].

A particular case of multiplexer is two-channel frequency-dividing devices – diplexer. Diplexer implements two transfer functions meeting the condition

$$|H_1(j\omega)|^2 + |H_2(j\omega)|^2 \le 1.$$

Diplexers are the basic elements at construction of frequency-dividing devices with any amount of channels.

A significant number of works where various aspects of theory and design of the directional filters are discussed indicates the importance of the problems of their synthesis [1, 3–8]. The work [7], where the general technique of calculating the transfer functions of the directional filters with maximum flat gain-frequency characteristics (GFC) and controlled attenuation at frequency of GFC junction is considered, should be especially noted. The question of branching filter implementation on the basis of parallel or series connection of passive LC filters realizing separate transfer functions are examined in [4–6]. Such approach is not optimal as each transfer function is implemented by a certain filter and realizing circuit has a high order. The technique of synthesizing diplexer in the form of reactive six-pole is proposed in the article [8]. However, the examples given in [8] show that a number of reactive elements in synthesized network exceeds considerably the order of the implemented transfer functions.

In this paper the method of implementing diplexers based on the use of the analog wave filters (WF) is considered. The proposed approach allows obtaining the frequency-dividing filters of minimal order. It is suitable for designing both active (ARC) and passive (LC) frequency-dividing devices.

Wave filter represents an unbalanced network realizing simultaneously four transfer functions (Fig. 1). The connection between stresses on external terminals of wave filter is determined by the equations in transfer

$$\begin{bmatrix} U_{вых1} \\ U_{вых2} \end{bmatrix} = \begin{bmatrix} t_{11} & t_{12} \\ t_{21} & t_{22} \end{bmatrix} \begin{bmatrix} U_{вх1} \\ U_{вх2} \end{bmatrix}, \qquad (1)$$

or chain parameters:

$$\begin{bmatrix} U_{вых1} \\ U_{вх2} \end{bmatrix} = \begin{bmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{bmatrix} \begin{bmatrix} U_{вх1} \\ U_{вых2} \end{bmatrix}.$$

The elements of matrix of the transfer parameters $[T]$ are the transfer functions between separate inputs and outputs.

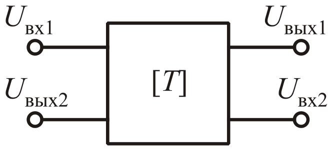

Fig. 1. *The wave filter*

Matrix of wave filter coefficients in eq. (1) is a bounded transfer matrix of spreading parameters meeting the para-unitarity condition [2, 3]:

$$[T]^{\bullet}[T] = [1]. \qquad (2)$$

Matrix $[T]^{\bullet}$ comes out of transposition $[T]$ and change of $j\omega$ by $-j\omega$. The Feldtkeller equalities [2]

$$|t_{11}(j\omega)|^2 + |t_{21}(j\omega)|^2 = 1, \qquad (3)$$

$$|t_{22}(j\omega)|^2 + |t_{12}(j\omega)|^2 = 1.$$

are the consequence of the (2).

It follows from the (3) that the elements $t_{11}(j\omega)$ and $t_{21}(j\omega)$ are complementary, i.e. the pass band $t_{11}(j\omega)$ corresponds to the stop band $t_{21}(j\omega)$. Thus, the wave filter implements automatically the branching filter. It was firstly noticed in the review [2].

A general theory of realizing the analog wave filters is examined in works [9, 10]. The synthesized structure represents the cascade connection of first-second order sections implementing zeros of the transfer function $H_1(j\omega)$ (Fig. 2). Section principle circuit depends on a type of realizable zeros.

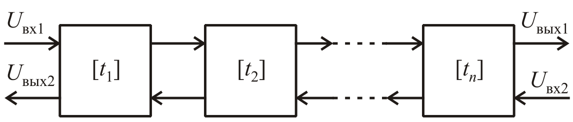

Fig. 2. *The structure of synthesized circuit*

For the filters of Butterworth and Chebyshev, zeros, the transfer of which are in infinity, the first order section are enough to be used. The matrix of section chain parameters realizing zero in infinity is of the form

$$[b^{(i)}] = \begin{bmatrix} 1 - c_i s & d_i s \\ d_i s & -1 - c_i s \end{bmatrix}.$$

Parameters c_i and d_i depend on the type of the realized transfer function $t_{11}(j\omega)$. The parameters of sections of Butterworth filters are given in the Table.

Table. *The parameters of sections of Butterworth filters. Order n=2–9*

The parameters of the recovered four-terminal networks c_i							
2	3	4	5	6	7	8	9
0,7071	0,5000	0,3827	0,3090	0,2588	0,2225	0,1951	0,1736
0,7071	1,0000	0,9239	0,8090	0,7071	0,6235	0,5556	0,5000
	0,5000	0,9239	1,0000	0,9659	0,9010	0,8315	0,7660
		0,3827	0,8090	0,9659	1,0000	0,9808	0,9397
			0,3090	0,7071	0,9010	0,9808	1,0000
				0,2588	0,6235	0,8315	0,9397
					0,2225	0,5556	0,7660
						0,1951	0,5000
							0,1736

Structure chart of the section implementing transfer zeros placed on imaginary axis is shown in Fig. 3. $T(s)=1/(1+sc_i)$ corresponds to transfer zeros in infinity. One of the variants of this structure chart realization in the base OU-RC is shown in Fig. 4.

The filter formed by the cascade connection of sections showed in Fig. 4 contains minimal quantity of reactive elements equal the order of the realized transfer functions. However, the number of active elements (operational amplifiers – OA) turns out to be high. For example, 12 OA are required for realization of the filter of the third order.

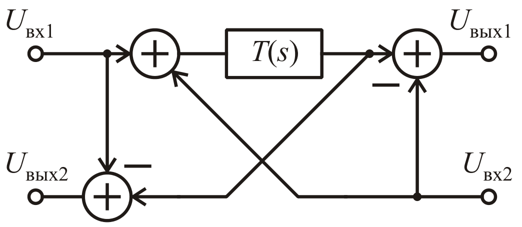

Fig. 3. *The structure chart of the section implementing zero on the imaginary axis*

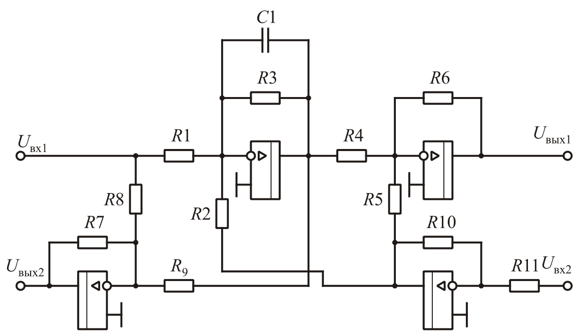

Fig. 4. *Principle circuit of the section implementing zero on the imaginary axis*

In order to obtain more saving implementation the method of equivalence conversions of chain parameter matrix is used [9]. Let us examine the conversion of similarity of the chain parameter matrix

$$[\tilde{B}^{(0)}] = [Q]^{-1}[B^{(0)}][Q]. \qquad (4)$$

Here $[Q]$ is the nonsingular transformation matrix. The variables on external terminals of four-terminal network are transformed to a new coordinate base:

$$\begin{bmatrix} \tilde{U}_{вых1} \\ \tilde{U}_{вх2} \end{bmatrix} = [Q]^{-1} \begin{bmatrix} U_{вых1} \\ U_{вх2} \end{bmatrix}, \quad \begin{bmatrix} \tilde{U}_{вх1} \\ \tilde{U}_{вых2} \end{bmatrix} = [Q]^{-1} \begin{bmatrix} U_{вх1} \\ U_{вых2} \end{bmatrix}. \qquad (5)$$

Structure chart shown in Fig. 5 corresponds to transformation (4). It is formed by the cascade connection of four-terminal network having a matrix of chain parameters $[\tilde{B}^{(0)}]$ and two sections of zero order having chain matrices $[Q]^{-1}$ and $[Q]$ respectively.

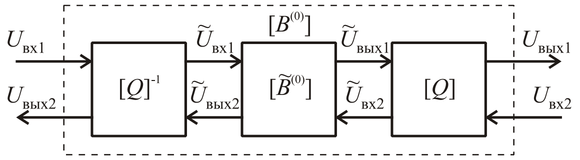

Fig. 5. Structure chart corresponding to transformation (5)

It is not difficult to show that transformation of the chain or transfer parameters of the whole circuit according to (4) is equivalent to transformation of parameters of each constituent four-terminal network separately.

Let us use the transfer matrix of the form

$$[Q] = \begin{bmatrix} 1 & \dfrac{1}{K(s)} \\ 1 & -\dfrac{1}{K(s)} \end{bmatrix}.$$

Here $K(s)$ is the fractional rational function of frequency in general case. Transforming the chain matrix $[B^{(0)}]$ according to (4), the structure chart shown in Fig. 6 is obtained.

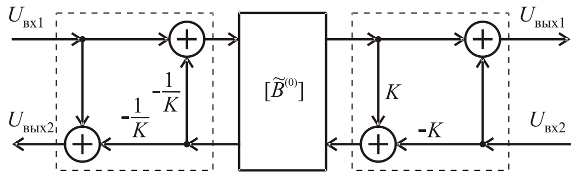

Fig. 6. The structure of the transformed wave filter

The transformed matrix of the transfer parameters is of the form:

$$[\tilde{t}^{(i)}] = \frac{1}{c_i + d_i} \begin{bmatrix} f(s) & \dfrac{f(s)}{K(s)} \\ -K(s)f(s) & f(s) \end{bmatrix}.$$

Function $f(s)$ depends on the type of the recovered zeros. At $s=0$ $f(s)=s$. Function $f(s)=1/s$ corresponds to zero in infinity.

Structure chart of a signal four-terminal network realizing the transfer matrix $[\tilde{t}^{(i)}]$ is introduced in Fig. 7. $K(s) = 1$ for the active RC-filters.

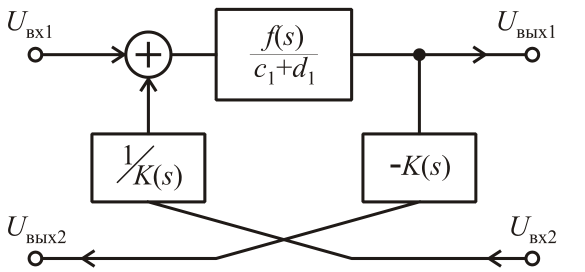

Fig. 7. Structure chart of the first order section

One of the variants of this structure chart realization is shown in Fig. 8.

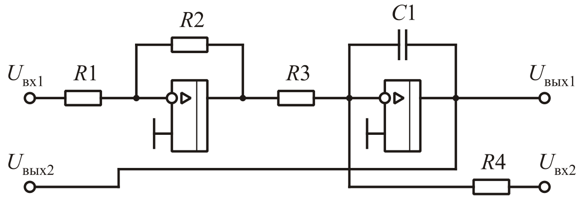

Fig. 8. Principle circuit of the section implementing zero in infinity

A more saving realization may be obtained connecting in turn the sections shown in Fig. 9, 10.

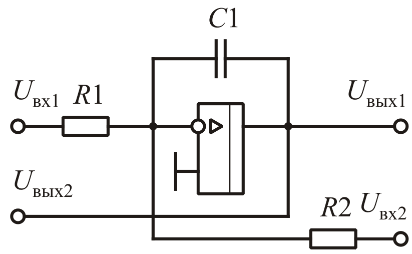

Fig. 9. An odd section of frequency-dividing filter

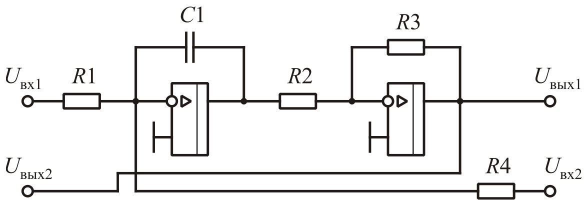

Fig. 10. An even section of frequency-dividing filter

The resistor voltages in circuits in Fig. 9, 10 are calculated by the formulas:

$$R_i = \frac{c_i}{\pi \cdot f_0 \cdot C},$$

here i=1,2,3,... – are the serial numbers of filter sections.

Integrating load sections with the first and the last links the configurations showed in Fig. 11, 12, respectively, are obtained.

So, for minimal realization of active frequency-dividing filter the section of four types:

- initial, Fig. 11;
- final, Fig. 12;
- intermediate even and odd, Fig. 9, 10 should be used.

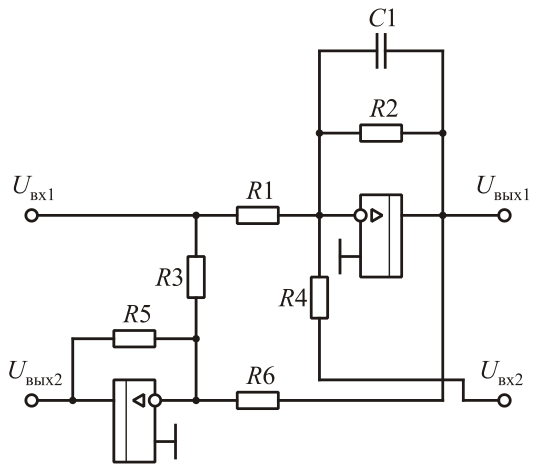

Fig. 11. *The initial section of frequency-dividing filter*

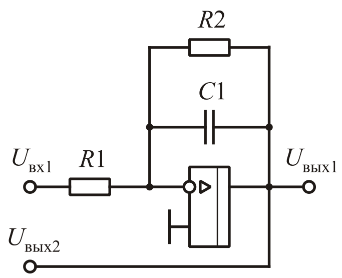

Fig. 12. *The final section of frequency-dividing filter*

Filters of various orders differ only in number of intermediate sections.

The original data at designing filter are the order and the form of the transfer function as well as the cutoff frequency f_0. The procedure of calculation consists in choice of suitable condenser capacities and resistor design by the ratio (10).

The diagram of diplexer of the third order designed according to the proposed approach is shown in Fig. 13. The cutoff frequency f_0=159 Hz. It is enough to realize the only input for the diplexer proper operation. The gain-frequency characteristics of diplexer are shown in Fig. 14.

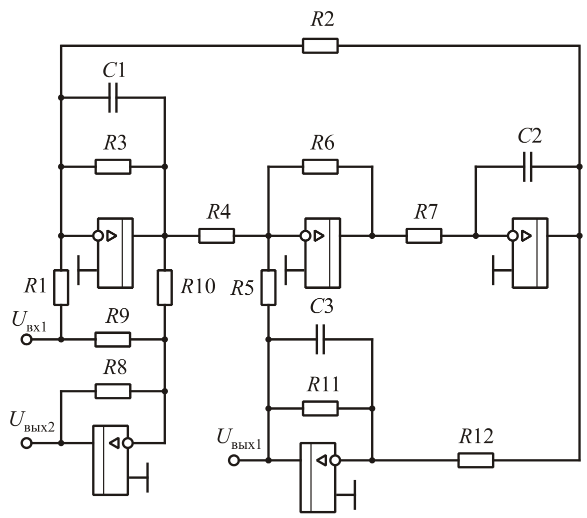

Fig. 13. *The frequency-dividing filter of the third order. R1=0,5 kOhm; R2–R12=1 kO; C1–C3=1 mkF*

Finally, let us note that the proposed approach allows designing the frequency-dividing filters of minimal order realizing the functions of various types (of Butterworth, Chebyshev etc.). It may be used as well for synthesis of passive diplexers.

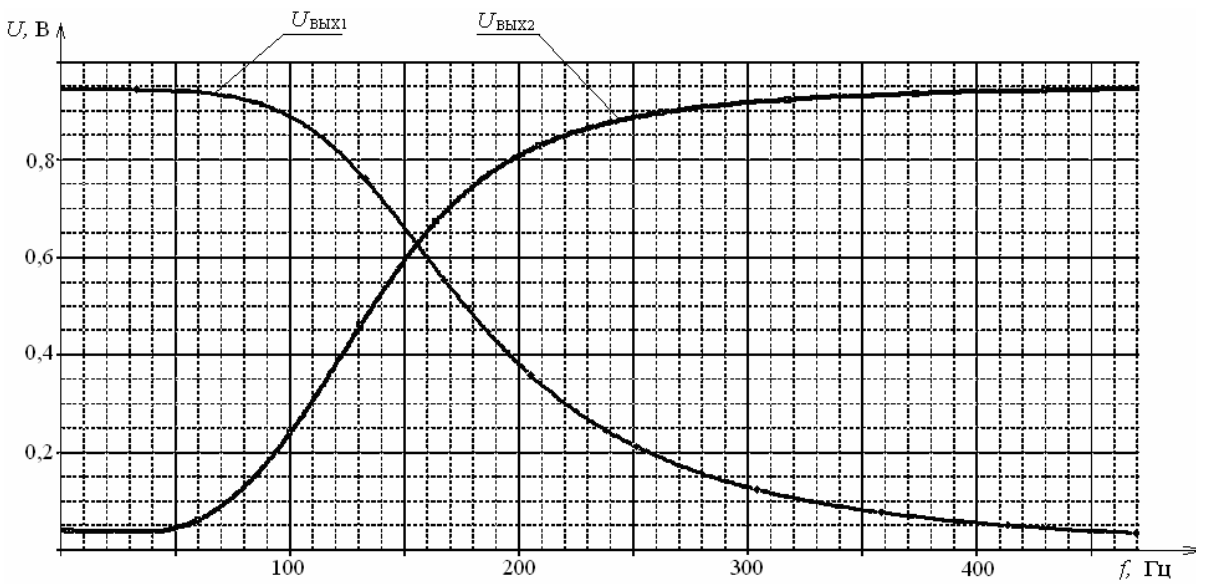

Fig. 14. *The gain-frequency characteristics of the synthesized filter*

REFERENCES

1. Alekseev O.V., Groshev G.A., Chavka G.G. Multichannel frequency-dividing devices and their application. – Moscow: Radio and communication, 1981. – 136 p.
2. Fettweis A. Wave digital filters // TIIER. – 1986. – № 2. – P. 35–99.
3. Neirynck J., Carlin C.-H. Synthesis of the lossless reciprocal three-port based on a canonic form of its scattering matrix // IEEE Transactions on Circuits and Systems. – 1981. – V. CAS-28. – № 7. – P. 736–744.
4. Youla D. Several key representations of circuit theory being a principle of filters calculation by the classical method of attenuation characteristic // TIIER. – 1971. – № 5. – P. 45–87.
5. Belevitch V. On filter pairs with Butterworth characteristics // International Journal of circuit Theory and Applications. – 1987. – V. 15. – P. 51–60.
6. Zhu Y.-S, Chen W.-K. On the design of a diplexer having Butterworth characteristics // IEEE Transactions on Circuits and Systems. – 1988. – V. CAS-35. – № 11. – P. 1455–1457.
7. Youla D. C., Pillai U., Winter F. Theory and design of maximally flat low-pass high-pass reactance-ladder diplexers // IEEE Transactions on Circuits and Systems. – 1992. – V. CAS-39. – № 5. – P. 337–349.
8. Neirynck J., Carlin C.-H. Synthesis of the lossless reciprocal three-port based on a canonic form of its scattering matrix // IEEE Transactions on Circuits and Systems. – 1981. – V. CAS-28. – № 7. – P. 736–744.
9. Bondarenko A.V., Dovgun V.P. Synthesis of analog wave filters // Electricity. – 2005. – № 5. – P. 52–56.
10. Dovgun V.P., Barybin P.A., Novikov V.V. The analog wave filters: main properties and procedure implementation // The Bulletin of association of KSTU graduates. Krasnoyarsk. – 2005. – Issue 12. – P. 156–164.

Received on14.05.2007

UDC 621.372.852

MODELING OF ATTENUATOR STRUCTURES ON FIELD EFFECT TRANSISTORS WITH MINIMAL PHASE SHIFT AT ATTENUATION REGULATION

O.V. Stukach

Tomsk Polytechnic University
E-mail: tomsk@ieee.org

Controlled absorbing attenuators on Schottky-gate FETs: T-circuit, T-shaped bridge circuit and transistor attenuator in the mode with controlled slope of voltage-current characteristic have been considered. Attenuator phase-frequency characteristics were modeled. The main difference of the circuits from the known ones consists in introduction of equalizers that conditions broadband feature and large attenuation range where minimum of the phase shift is achieved at regulation. As a result, the optimal parameters of adjusting circuits in attenuators are founded. It is shown that the least phase shift is provided in attenuators on transistor with controlled volt-ampere characteristic steepness. The comparative estimation of the considered base structures was given.

The requirement of phase shift constancy at transfer constant adjustment is made to modules of active phase lattices, systems of automatic phasing in transmitters, precise wide-band amplifiers, attenuators with smooth variation of attenuation and other devices with adjustable characteristics of signal transmission. Change of phase shift or group delay is conditioned by the influence of parasitic reactivity of elements with controlled resistance. There is a certain process limit in decreasing parasitic parameters. Therefore, one of the most important tasks is a balance of parasitic reactivities of the controlled elements by circuit solutions.

1. The problem of phase shift invariance

Electrically controlled attenuators (ECA) are intended for smooth change of signal level in a circuit. For a number of practical tasks, for example, CDMA of communication systems, phased arrays, surface radars etc., the excess requirement is made to attenuators; the requirement to phase shift stability of output signal relative to the input one at adjustment of transfer constant [1]. This aim is complicated at system operation in a wide band, in general case from zero to several GHz. Phase variation is conditioned by the influence of parasitic reactivities of the controlled elements. They may be decreased technologically only to a certain limit, especially in super-wide band. Therefore, the only method of supporting phase shift invariance to attenuation is the balance of parasitic reactivities of controlled elements by a circuit way. In particular, the equalizers included specially into the base structure find wide application in absorbing attenuators.

The methods of phase correction are developed best of all for ECA on *p-i-n* diodes [1]. Schottky-gate FETs (SFET) have a number of advantages although diodes gain in maximum power of controlled signal. In particular, the advantage is in switching time, decoupling between signal transmission circuit and control paths. Availability of using SFETs conditioned by low values of parasitic reactivities simplifies considerably the problem of constructing wide band ECA in microwave range.

The aim of the work is the investigation and simulation of circuit design characteristics for ECA on FETs, search for perspective circuit methods of phase shift correction and comparison of the obtained results with known characteristics of base diode and transistor structures.