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Features of the Signal System of Automated Design of Frequency Converters Systems - Ultrahigh Frequencies

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The article analyzes and outlines the modern principles of the theory of the signaling system of automated design with frequency transformations on a non-reciprocal electronic device - the transistor of ultrahigh frequencies. In addition, schematic variants of frequency converters on field and bipolar transistors, Schottky diodes (varicaps) are considered; the signal theory of transistor and diode frequency converters is described: resistive and capacitive. Schematic performances of frequency converters represent variants of the signaling system of automated designing at ultrahigh frequencies.

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Introduction

In the information path (IT) of telecommunication systems, the basic processing of signals is carried out in order to isolate the information contained therein (demodulation) and mitigate the effects of interference. At the same time, an important task is to select information with maximum reliability - the so-called optimal acceptanceFor this, in the presentation of IT, there is provided an optimal filter, links for post-detector processing, follow-up systems for frequency and phase automatic frequency adjustment, used to demodulate the signal, as well as to find and track it by frequency, phase, and delay. Here, frequency converters play a very important role in the reception or transmission of signals, which also determines the signal CAD, on the features of which we will dwell in this article.

I. Frequency converters and parametric amplifiers

Frequency converters (IFs) are designed to transfer the spectrum of radio signals from one region of the radio frequency range to another. Such a transfer of the spectrum must pass without changing the type and modulation parameters, that is, linearly (Fig. 1). Frequency conversion is possible as a result of multiplying two voltages. One of them is the received signal $U_c = U_c cos(\omega_c t + \varphi_c)$, a and the second - the voltage of the auxiliary generator, which is formed in the receiver $U_r = U_r cosw_r\ t$. Then, when multiplying the signal and the heterodyne, the following combinations of

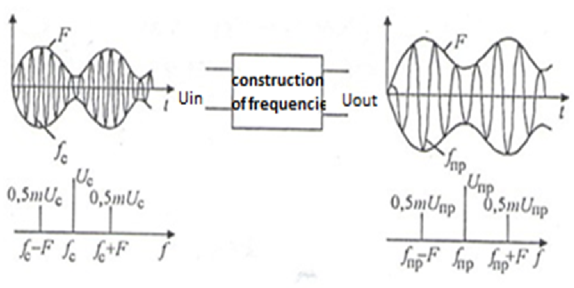

Fig. 1. The converter of the frequency and spectrum of the input signal in the output.

complex frequencies appear:

$$U_c U_r = 0.5 U_c U_r [(\omega_r \pm \omega_c)t \pm \varphi_r]$$

One of them is allocated by the filter and is called the voltage of the intermediate frequency $U_{\text{пр}} = U_{\text{пр}} cos(\omega_{\text{пр}} t + \varphi_{\text{пр}})$. As you can see, such a voltage receiver can be realized with nonlinear links from periodic variables under the action of the generator. In the role of nonlinear or parametric elements, which are called mixers, today use transistors in a discrete or integral performance and diodes.

The signal at the inlet of the mixer must be small so that the nonlinearity of its characteristics does not lead to a slight distortion of the received signal. The voltage of the heterodyne is initially large, therefore the conductivity of the mixer varies according to the law of changing the voltage of the heterodyne (Fig. 2).

Of course, you can submit it with several expressions:

$$g_{21}(t) = G_{21}^{(0)} + \sum_{R=1}^{\infty} G_{21}^{(R)} cos R \omega_r t, \qquad (1)$$

where $G_{21}^{(k)}$– the amplitude of the kth harmonic of the conductivity of the nonlinear element;$G_{21}^{(0)}$ – constant component of conductivity.

Current at the mixer outlet $i = g_{21}(t)U_c$. Substituting the meaning $g_{21}(t)$ i U_c get the following dependence:

$$i = G_{21}^{(0)} U_c cos(\omega_c t + \varphi_c) +$$
$$+ 0{,}5 \sum_{k=1}^{\infty} G_{21}^{(k)} U_c cos[(k\omega_r + \omega_c)t + \varphi_c] \quad (2)$$

From this expression it is clear that the combinational components$(k\omega_{\text{г}} \pm \omega_{\text{c}})$ appear as a result of the change in the conductivity of the nonlinear element (NO) under the action of the voltage of the heterodyne. They have the same structure as the output signal. Permanent component of conductivity $G^{(0)}_{21}$ does not

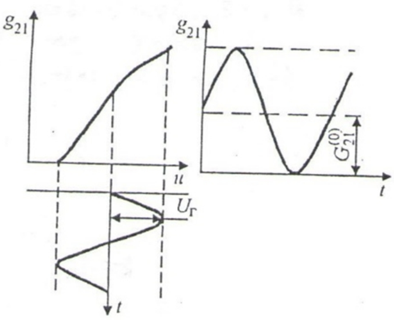

Fig. 2. Dependence of conductivity of the mixer on the voltage of the heterodyne.

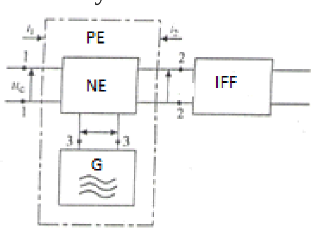

Fig. 3. Structural diagram of the frequency converter on the non-inverted transforming element.

allow frequency conversion. Similar results are obtained when changing the capacitor of the mixer under the action of the voltage of the heterodyne.

The main indicators of the quality of the IF are: the gain of voltage and power, the range of operating frequencies, selectivity, noise ratio, distortion, stability, reliability. They are similar indicators of resonant amplifiers, but some of them have features characteristic of the frequency conversion mode. For example, unlike amplifiers in the IF, there are unilateral receiving channels, which worsen their selective properties and cause some circuit design decisions.

II. The theory of frequency conversion on a non-interconnected electronic device

Generalized schematic diagram of the IF is given in Fig. 3. It consists of a non-linear element (NE) - a mixer, a filter of intermediate frequency (FPC) and a heterodyne (G). The mixer itself can be supplied with a 6-pole, which feeds the voltage of the transducer signal U_c and the heterodyne

$U_г$, and at the output voltage is already an intermediate frequency U_{np}. Subsequently, the presentation of nonlinear materials (NO), together with the heterodyne, will be called a transforming element (PE).

Voltages of the signal and the intermediate frequency are less than the voltage of the heterodyne, therefore we can assume that the conductivity does not change only under the action of the voltage of the heterodyne. This allows us to apply to the analysis of the IF a simple method of the theory of nonlinear circles: currents in the circles of the mixer as a function of the reduced stresses can be filed in the form of decay in the Taylor series in degrees of small voltages with the rejection of members of a series with high degrees.

Imagine the input and output currents of the PE in the form of a function

$i_1 = f_1(U_г, U_c, U_{np})$; $i_2 = f_2(U_г, U_c, U_{np})$; which are determined by the static characteristics of the mixer and the mode of its operation. To derive the equation of direct transformation we use expressions of currents and1 and and2 as functions of stresses and decompose in the Taylor series in degrees of small U_c i $U_{\text{пр}}$, limiting members to a schedule not exceeding the first order:

$$i_2 = f_2(U_r) + \frac{\partial f_2(U_r)}{\partial U_c} U_c + \frac{\partial f_2(U_r)}{\partial U_{\text{пр}}} U_{\text{пр}} + \cdots \qquad (3)$$

Here, the first term is a constituent of the mixer current under the action of the voltage of the heterodyne i2g = f2 (Ug). This current does not contain combinational components, but has only components with the frequency of the heterodyne and its harmonics.

Derivative $\frac{\partial f_2(U_r)}{\partial U_c}$ is the differential conductivity (steepness) of the direct action of the inverter for the signal voltage, which we denote by g_{21}. It periodically varies with the frequency of the heterodyne and can be given near Fourier. Second derivative

$\frac{\partial f_2(U_r)}{\partial U_{\text{пр}}}$ is already a differential output conductivity of the IF g_{22}, which can also be filed in the Fourier series:

$$g_{22} = G_{22}^{(0)} + \sum_{R=1}^{\infty} G_{22}^{(k)} cosk\omega_r t, \quad (4)$$

where $G^{(0)}_{22}$ – constant component of random mixing conductivity (average value

g_{22} in the period of heterodyne voltage);

$G^{(R)}_{22}$ – the amplitude of the R harmonic of the output conductivity. Taking into account the accepted symbols, the value of current i_2 can be given by the expression:

$$i_2=i_{2r}+q_{21}U_c+q_{22}U_{пр}, \quad (5)$$

Passing to complex amplitudes, the expression for the current $i_{пр}$ can be applied:

$$\acute{I}_{пр} = 0{,}5G_{21}^{(k)}\acute{U}_c + G_{22}^{(0)}\acute{U}_{пр} \quad (6')$$

$$\acute{I}_{пр} = 0{,}5G_{21}^{(k)}U_c + G_{22}^{(0)}U_{пр} \quad (6'')$$

Here $\acute{U}_c = U_c e^{i\varphi_c}$; $\acute{U}_{пр} = U_{пр} e^{i\varphi_{пр}}$ - complex amplitudes of signal voltage and intermediate frequency. By$U_c = U_c e^{-j\varphi_c}$ marked the complex-able amplitude Uc.

The expression (6 ') is the equation of the direct transformation of the inverting IF, in which the lateral strips in the transformation are replaced by places: the lower one becomes the upper one, and the upper lower one (Fig. 4, b).

The first term in (6 ') characterizes the process of frequency transformation, and the second is due to the load response. The coefficient of proportionality between the amplitude of the output current of the intermediate frequency and the amplitude of the voltage of the input signal at C.3 at the output is called the steepness of the transformation:

$$G_{21пр} = \frac{\frac{\acute{I}_{пр}}{\acute{U}_c}}{\acute{U}_{пр}} = 0 = 0{,}5G_{21}^{(k)} \quad (7)$$

As you can see, it is determined by half the amplitude of the R-th harmonic of direct conductivity. Initial conductivity of the IF at C3 at the input is determined by the constant component of the output conductivity of the mixer, varies under the action of the heterodyne:

$$G_{22пр} = G_{22}^{(0)} = \frac{\frac{\acute{I}_{пр}}{\acute{U}_{пр}}}{\acute{U}_c} = 0, \quad (7')$$

If the mixer has a nonlinear conduction of the inverse action, then in the IF, next to the straight line, will be the inverse transformation. It consists in the fact that if the intermediate frequency voltage is applied to the output clamps of the mixer, then at the action of the heterodyne voltage at the input there will be a current with the signal frequency. To output the equation of inverse transformation, the expression for current i_1 is also decomposed

Taylor by U_c i $U_{пр}$ limited to linear terms

$$i_1 = f_1(U_r) + \frac{\partial f_1(U_r)}{\partial U_c}U_c + \frac{\partial f_2(U_r)}{\partial U_{пр}}U_{пр} + \cdots \quad (7'')$$

Similarly enter the notation $i_1=f_1(U_r)$ – Current at the inlet of the mixer under the action of the voltage of the heterodyne; $g_{11} = \frac{\partial f_1(U_r)}{\partial U_c}$- differential input conductivity, $g_{12} = \partial f_c(U_r)/U_{пр}$ – differential conductivity of the internal inverse communication (OZ). Further, introducing the g_{12} and g_{11} the Fourier series is analogous to that of the current i_2 we obtain expressions for complex amplitudes:

$$\acute{I}_c = G_{11}^{(0)}\acute{U}_c + 0.5G_{12}^{(k)}\acute{U}_{пр}, (8')$$

at $\omega_c=k\omega_r + \omega_{пр}$ або $\omega_c = \omega_{пр} - k\omega_r$,

$$\acute{I}_c = G_{11}^{(0)}\acute{U}_c + 0.5G_{12}^{(k)}U_{пр}, \quad (8'')$$

at $\omega_c = k\omega_r - \omega_c$.

This is the inverse transformation equation for a non-inverting (8 ') and an inverting IF (8' '). Proportional

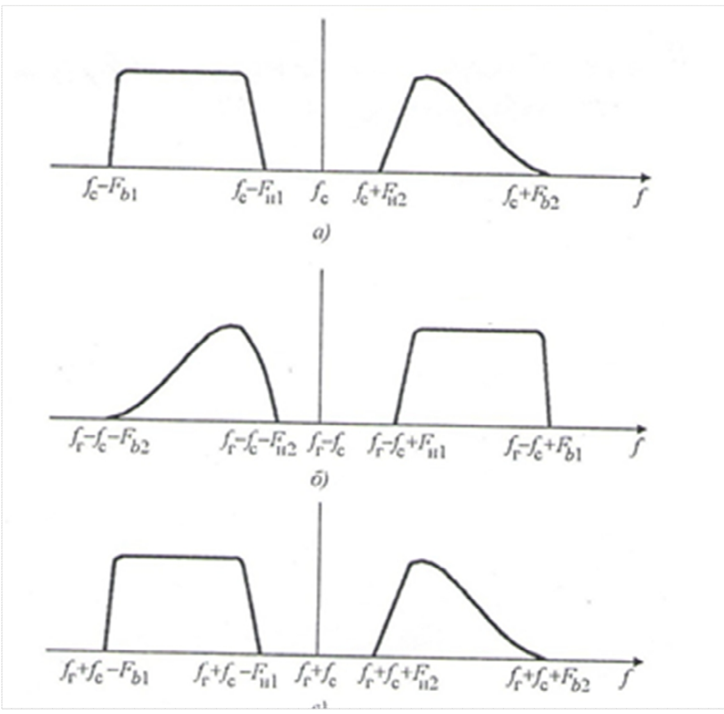

Fig. 4. Conversion spectra of non-inverting (a) and inverting (b) frequency converters.

coefficient between the amplitude of the current and the input signal frequency ω_c and the amplitude of the intermediate frequency voltage at the output the mixer U_{np} at K.3 at the input is called slope rotation is transformed:

$$G_{12\text{пр}} = \frac{\acute{I}_c/\acute{U}_{\text{пр}}}{\acute{U}_{\text{c}}} = 0.5G_{12}^{(k)}, \quad (8''')$$

The initial conductivity of the IF at K.3 at the output is determined by the constant component of the input conductivity of the mixer:

$$G_{11\text{пр}} = G_{11}^{(0)} = \frac{\acute{I}_c/\acute{U}_{\text{c}}}{\acute{U}_{\text{пр}}}, \qquad (8'''')$$

Therefore, in the general case, when using the inertial NO in the mixer, all parameters in the expressions (6) and (8) are complex, similar to the parameters of the amplifier (transistor), but taking into account the mode of frequency conversion and the action of the n-voltage of the heterodyne. Thus, the non-inverting IF can be described by the following system of two linear equations of the 4-pole

$$\begin{cases} I_c = Y_{11}U_C + Y_{12}U_{\text{пр}} \\ I_{\text{пр}} = Y_{21}U_C + Y_{22}U_{\text{пр}} \end{cases} \qquad (9)$$

The resulting expression (9) is valid only for the amplitudes, not the instantaneous values of currents and voltages, which differ in frequency at the input and output. They describe the result, not the mechanism of frequency conversion, and give the fact that for small signals, the transforming element (PE) can be considered as an active linear 4-pole, and the action of the heterodyne on NO is expressed in the transformation parameters (7) and (8). Since the equation (9) in the form coincides with the system of equations of the active 4-pole (transistor) using the known equations:

$\dot{I}_2 = Y_{21}\dot{U}_1 + Y_{22}\dot{U}_2$,

hen for the IF there is a fair equivalent circuit of the resonant amplifier taking into account the parameters of the transformation. This allows you to broaden the results of the considered IF

the theory of resonance and band amplifiers by replacing the amplifier parameters with the transformation parameters, using the theory of 4-pole, which allows to form a signal CAD for the frequency converters. For example, the resonant decomposition factor of the converter is determined by the expression:

$$K_{0\text{пр}} = \text{mn}|Y_{21\text{пр}}|R_c \qquad (10)$$

where $|Y_{21np}|$ - module of complex steepness of transformation, $k_c = \rho K_f$; ρ the resistance of the filter, K_f - filter transfer coefficient.The inverse transformation is analogous to the inverse communication (OZ) in the amplifier, but it is already nonlinear. In the IF on a non-reciprocal element, the OZ is more complex than that of an amplifier, but to rule the converter for the stability of the element to take appropriate measures to increase it.

III. Analysis of the side conversion channels in the IF

Unlike the resonant amplifier, the voltage at the output of the IF is already appearing, based on the theory of section 1, on different input signal frequencies$f_c = kf_r \pm f_{\text{пр}}$ depending on the number of the harmonic frequency of the heterodyne (5). At an intermediate frequency f_{np} is simply a slip-up amplifier $G_{21}^{(0)}$. This channel is a direct signal passing $U_c = U_c cos(\omega_c t + \varphi_c)$ without spectral transfer relative to the frequency of the heterodyne. Converting to the first harmonic the frequency of the heterodyne f_c is proportional$G_{21}^{(1)}$ at the output signal frequencies $f_г - f_{np}$ i $f_г + f_{np}$. Transformation on the second harmonic of the heterodyne $2f_г$ is already proportional$G_{21}^{(2)}$ at frequencies $2f_г - f_{np}$ i $2f_г + f_{np}$ and so on. Accordingly, as we see from Fig. 5 frequency response has already several maxima (1, 2, 3 ...). That is, the higher the order of transformation, so, as a rule, the slower conversion, and, the amplification factor of the mixer. In the bandwidth of the filter at the output of the IF, the products of the fluctuation of all channels are converted. One of these channels is the main one, and the rest are side-by-side that interfere with the transformation. For example, if the main channel 2 is selected with frequency $f_г$, then the secondary will be the channel, which is like a mirror image of the main channel 2, so it is called mirror (or symmetric), its frequency f_3 differs from the frequency of the main channel by $2f_{np}$. The suppression of the mirror channel is facilitated at a higher intermediate frequency. The solution to this contradiction can be achieved by double or triple transformation of the frequency. The scheme of double frequency conversion is given in Fig. 6. Here a frequency signal is applied f_c consistently converted to the first and frequency converters. The relatively high first intermediate frequency allows you to perform the necessary relaxation of the mirror channel in the filter pre-selector f_1. Converted signal at frequency f_{np1} is highlighted by a filter f_2 already in the tract of the first intermediate frequency, and then again lowered by the converter of the IF2. The relatively low second intermediate frequency facilitates the formation of an already required resonance characteristic and bandwidth of the path f_{np2} by a filter f_3.

The peculiarity of the double frequency conversion is the appearance of the second mirror channel $f_{дз2} = f_c - f_{np2}$, which is lagging behind the frequency f_{np1} on f_{np2} and placed symmetrically with respect to the frequency of the second heterodyne (Fig. 7). In the pre-selector, the second mirror channel is not significantly weakened, since the second intermediate frequency f_{np2} s relatively low and $f_{дз2}$ is placed sufficiently close to the receiving signal's frequency. At the output of the first IF, the frequency f_{dz2} is converted to the frequency $F_{g1} - f_{dz2} = f_{g1} - (f_c - 2f_{pr2}) = f_{pr1} + f_{pr2} = f_{pr3}$, which must be suppressed in the path of the first intermediate frequency by the f_2, filter, which is also intended for this. In the infradions (such superheterodyne receiver when $f_{np} > f_{cmax}$, which differs in that in its work in the frequency range only the heterodyne is tuned, and the pre-selector is not reconfigured), a frequency conversion is used in which the first intermediate frequency f_{np1} is selected above the maximum signal frequency. In the second frequency converter f_{np1} is already converted to f_{np2} and stands out by the filter f_2. Since the frequency f_{np1} s high, then for lowering it to the required value of one conversion of the IF may not be sufficient, therefore the third IF can be used in the

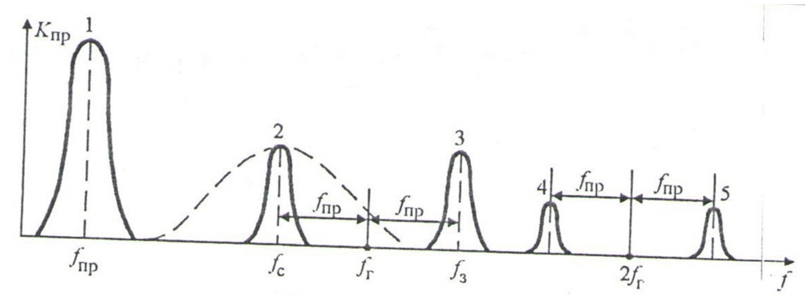

Fig. 5. Spectrum of output voltage in the frequency converter.

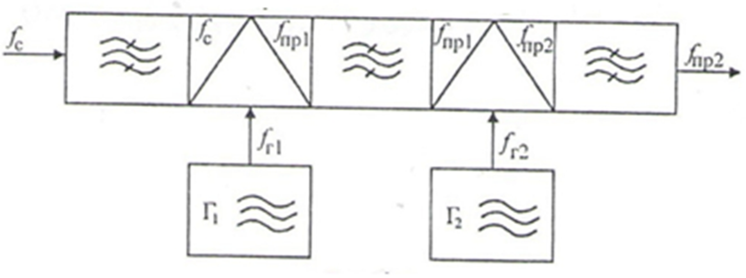

Fig. 6. Dual frequency conversion scheme.

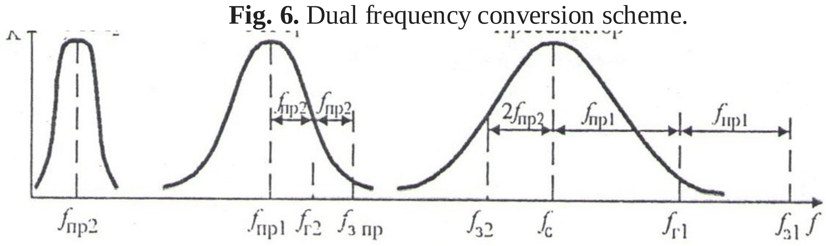

Fig. 7. Resonant characteristics of double freque

infradions. There is already added a third mirror channel, the suppression of which is provided by the filter f_3.

The advantage of infradin is to simplify the pre-selector (i.e. filter f1). In a variable frequency tuner with a wide range of frequencies, this filter is undesirable because it requires a smooth over-adjustment in the sub-band and the transfer of coils to change the sub-bands. Therefore, the mechanical switch is complex and unreliable and does not allow miniaturization. Touch switches are also complex and insufficiently reliable. Under the condition

$f_{пр1} > f_{cmax}$ the side channel of reception at an intermediate frequency is outside the receiver frequency range. The mirror channel frequency is also located at the upper limit of its range.

At $f_{г1} = f_c + f_{пр1}$ it lies in the range from

$(f_{cmin} + 2f_{пр1})$ to $(f_{cmax} + 2 f_{пр1})$. This allows using the filtered FPC filter as a filter f1, then passes the entire spectrum to the input of the IF1 with frequencies that are smaller than fcmax.

Frequency response of the IF supplied in Fig. 5 occurs at low levels of the input signal, when the nonlinearity of the signal is no longer manifested. At large levels of the input signal, the nonlinearity of the mixer in relation to the signal leads to an increase in the number of side channels of reception.

In this case, the intermediate frequency is formed in the form of a combination

$f_{пр} = kf_г \pm nf_c$, де k, n =1,2... When n = 1, there is a linear frequency transformation. The frequencies of the adjacent components of the admission are determined by the formula $f_{пр} = kf_г/n \pm nf_c/n$. For example, for R = n = 2, two side channels of the so-called semi-mirror frequency $f_{ндз} = f_г \pm 0.5f_{пр} = f_c \pm 0.5f_{пр}$. are created. They are formed by direct transformation of the second harmonics $|2f_{ндз} - 2f_г| = f_{ндз}$. The half-reflection channel

$f_{ндз} = f_c + 0{,}5 f_{пр}$ at $f_г > f_c$ or $f_{ндз} = f_c - 0{,}5 f_{пр}$ at $f_г < f_c$ is very close to the frequency of the useful signal fc in ε row near the mirror channel and its very hard to filter.

Then, due to additional side channels, there are interference distortions in the form of whistles (noise) not only at the intermediate frequency, but also in the harmonics and subharmonics fpr, as well as at frequencies far from $f_г \pm 0.5f_{пр}$, $\pm f_{пр}/3$ and so on. The converter is one of the first stages of the receiver, so its noise can significantly affect the overall noise ratio and, accordingly, the real sensitivity of the RPPP

IV. Schematic features of transistor IF

For frequency conversion, both field and bipolar transistors are used as nonlinear elements. In them, the frequency conversion is due to the change in the steepness of the characteristics of direct transmission under the action of the voltage signal and the heterodyne in the mixing elements. In fig. 8, a) the diagrams of the IF with a separate heterodyne on the PT and BT are given. In the first circuit, the signal voltage is applied to the gate circle, and the voltage of the heterodyne is in the leakage range. By signal, we get a schema from the SO, and for the heterodyne, the circuit with ZZ In the second scheme on BT the signal voltage is fed into the base, and the voltage of the heterodyne is in the circle of the emitter. This is achieved by eliminating the decoupling of the signal circles and the heterodyne.

The best interconnection between the signal and the heterodyne links is achieved in the circuit on the two-wound PT (Fig.1.8b), where the signal voltage and the heterodyne are fed to different gate terminals. In such a scheme, the frequency conversion is due to the change in the steepness of the characteristic of the signal shutter when the voltage on the heterodyne gate is changed. This is the original circuit diagram of the IF.

A very good signal circle and heterodyne decoupling and high self-excitation resistance in a wide range of frequencies are achieved in cascade (E-B) mixers. Figure 1.8 shows a version of the scheme in which the voltage signals and the heterodyne are fed to the base of different transistors (BTs).

The signals are a cascade ZE-ZB scheme, which provides high stability. The frequency conversion is due to the change in the steepness of the characteristics of the second BT, on which the mixer is assembled. Similarly, you can build a cascade mixer on a PT (ZV-ZZ).

The amplifier element of an IF is a transistor that can be used simultaneously to generate oscillations (autogenerator). In this case, the IFs are called generators or auto-parts. But the optimal modes for generating and converting the frequency are different. The stability of the frequency of generation is low, so these IF can only be realized in inexpensive receivers of low class. Therefore, the most common used is IF with a separate heterodyne.

Mode of DC mixers chooses to operate on a nonlinear section (area) in the transmission and, if possible, use a region with a linear change in the slope of this characteristic (Fig. 9).

From rice 9 a) the amplitude of the first harmonic of steepness$G_{21}^{(R)} \approx 0.5(g_{21max} - g_{21min})$, and the steepness of transformation by the first harmonic

$G_{21\text{пр}} = 0.5G_{21}^{(1)} = 0.5(g_{21max} - g_{21min})$ In the gain mode, you can accept the condition that $g_{21} = g_{21max}$, accordingly, the steepness in the mode of transformation is less steepness in the mode of amplification.

The analytical calculation of the parameters of the IF into the BT is possible using the exponential approximation of the characteristics$i_c = f(U_{\delta c})$, then approximated by the exponent$i_c = i_0(e^{bU_{e\delta}} - 1)$, where i_0 and $в$ – exponential parameters, which are determined from real transistor defects. Considering $i_к = h_{21}\delta ie$, Define the slope $g_{21} = \frac{dir}{dU\delta r} = nU\text{Б}\frac{dir}{dU\delta r} = ndU\delta\text{в}i_0e^{\text{в}U\delta e}$ (11)

In this expression $I_{k0} = n_{21\delta}I_{r0} = n_{21}\delta i_0e^{bU\delta 0}$ - collector current at the working point;$J_0(bU_г)$ i $J_k(bU_г)$ - modules of the functions of the imaginary argument of the first row of zero and R-th. Input and output supports of transistors in the mode of transformation are approximately 1.5 - 2 times greater than in the mode of amplification, and the input and output capacitances in the mode of amplification and transformation are the same.

The shutter characteristic in the PA is quadratic, therefore, the dependence of the slope of this characteristic g_{21} from $U_{зв}$ linear (Fig. 8, c). The output offset at the gate U30 is taken to be equal to half the cutoff voltage. $U_{30} = 0.5Ua_{\delta c}$, and the amplitude of the voltage of the heterodyne $U_г$ equal to U_{30}, to fully utilize the linear range of steepness change and not to enter the area of the shutter currents. In this case, the amplitude of the first harmonic of steepness = $0.5g_{21max}$ =, where is the steepness at the working point. The pace of transformation $G_{21\text{пр}} = 0.5$=0.5. From here we see that the steepness of the transformation $G_{21пр}$ in 2 times there is less steepness in the mode of amplification at the same voltage of displacement at the gate U_{30}. In the considered mode, without breaking off the amplitude of the higher harmonics of steepness $G_{21}^{(2)}$, $G_{21}^{(3)}$ nd above are equal to zero. Accordingly, here we will have only two side channels of reception: mirror and straight.

In an integral performance, the IFs are often performed on a balance or double-wavelength chart. Differential cascades are used here for frequency converters of frequency (Fig. 9).

Here is the collector voltage on the mixing transistors

VT_1 and VT_2 is fed through the midpoint of the inductance coil of the output resonance circuit, tuned to the intermediate frequency. Strums i_1 and i_2 The transistors of the pair VT1 and VT2 flow through the outgoing circuits, and the output voltage is proportional to their difference, as in the differential cascade. The voltage of the heterodyne on the mixing transistors is fed through the transistor VT3 in a phase, therefore the currents and1 and and2 with the frequency of the heterodyne, its harmonics and the components of the heterodyne noise currents have identical phases in both transistors, are mutually offset and do not generate voltages in the original circle. Under the action of the voltage of the heterodyne, the steepness of the characteristics of each of the differential VT1 and VT2 transistors varies.

In the balance compensated IFs, the frequency of the mirror channel is not suppressed. The converter with compensation of interferences with the mirror channel can be constructed according to the scheme (Fig. 9). The principle of operation of such a double-channel compensator with phase extension is that the receiving signal in different channels has the same phase and, when summed in the common path, doubles, and, accordingly, the mirror interferences are phase-off and compensate each other. The voltage from the heterodyne is fed to the mixers Cm1 and Cm2 in the role of which any of the above schemes can be used. Main channel signal voltage $U_c = U_c cos(w_c t + \varphi_c)$ and the mirror canal

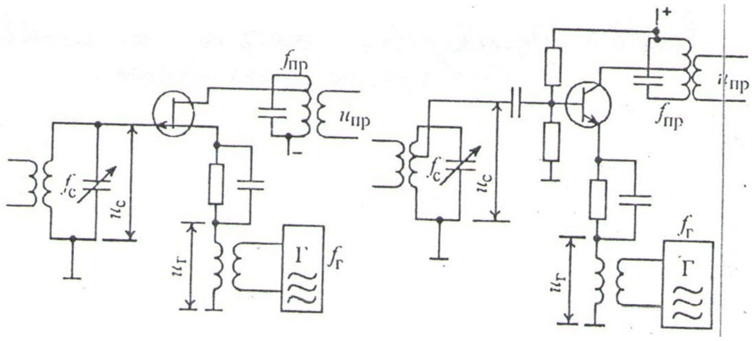

Fig. 8, a. Frequency converters on polar and bipolar transistors.

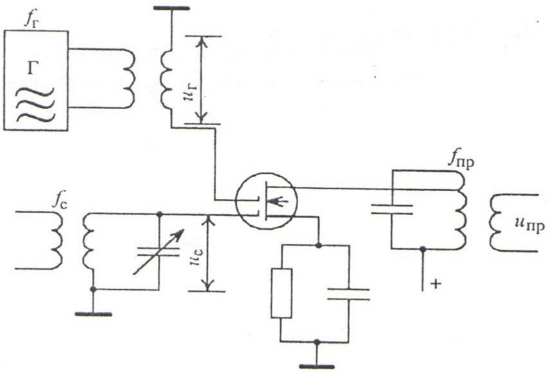

Fig. 8, b. Frequency converter on a two-wake field transistor.

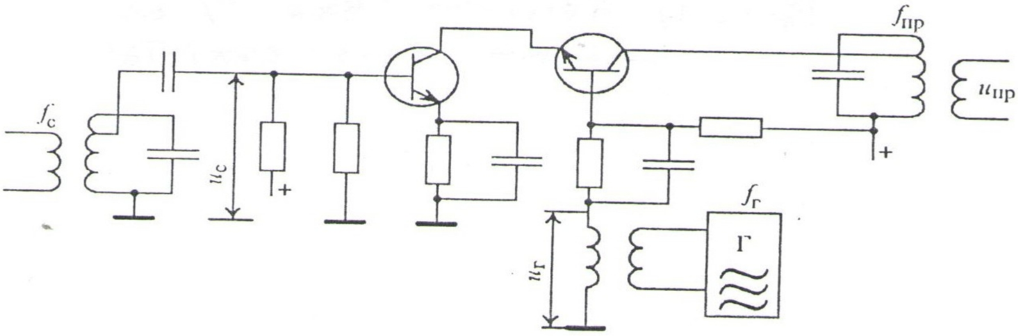

Fig. 8, c. Cascade frequency converter on bipolar transistors.

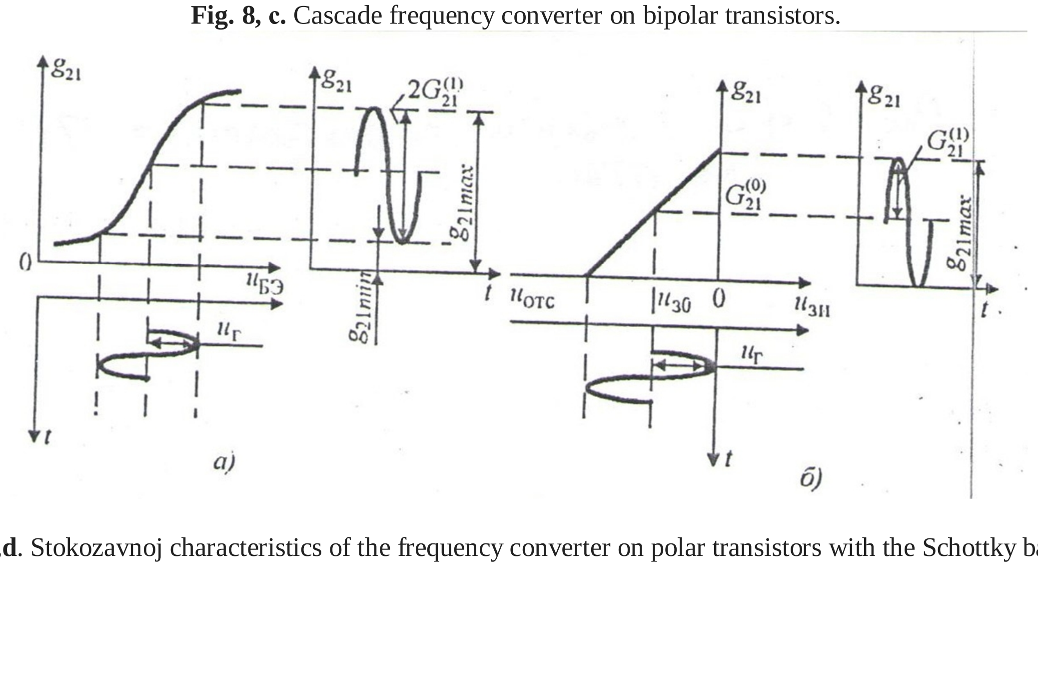

Fig. 8,d. Stokozavnoj characteristics of the frequency converter on polar transistors with the Schottky barrier.

are fed to the mixer C_{m1} directly and on the mixer C_{m2} via phase shifter fp with phase shift for π / 2. In the work of the main channel, a signal is received at the frequency

$f_c = f_г + f_{пр}$, and in the role of the mirror

$f_{зк} = f_c + f_{пр}$

Then, at the output of the mixer Sm1 after the FP filter, the voltage of the intermediate frequency of the main and mirror channels will be released:

$$U_{ак1} = U_{ак}K_n cos\left[\left(\omega_{ак} - \omega_g\right)t + \varphi_{ак} + \varphi_{с}\right]$$

where K_n – transfer coefficient of the mixer together with the FPCH filter. At the output of the second mixer Sm after the FPC (if the coefficient of phase shift is taken to be equal to one), respectively, will look like:

$$U_{ак2} = U_{ак}K_n cos\left[(\omega_{ак} - \omega_{г})t + \varphi_{ак} + \frac{\pi}{2} - \varphi_{с}\right]$$

After the phase converter $f_{в2}$ phase of the signal U_{c2} with the same with the phase U_{c1} ($\varphi_{c1} = \varphi_{c2} = \varphi_г - \varphi_c$),but the phase is upset $U_{зк2}$ differs from the phase $U_{зк1}$ by the value of π. With the same coefficients of transmission of the mixer voltages, the signal at the output of the adder has already doubled the amplitude, and the voltage of the mirror interference is accordingly compensated and the output is absent. In this and the originality of this circuitry mixer with compensation of interference.

V. Schematic features of microwave transistor converters

In bands of microwave bipolar transistors are used up to frequencies of 10 GHz at higher frequencies up to 100 GHz using PTSh in Fig. 10, a. Here the transistor is on the ZB circuit. Oscillations and the heterodyne are fed to the emitter via a directional separator (HB), which serves to decouple the signal circles and the heterodyne.

The circuit L1C2 in the circle of the emitter is tuned to the intermediate frequency and eliminates the inverse coupling of the current of this frequency.

The second loop of L2C4 in the output loop is also tuned to an intermediate frequency. Output offset to the base is fed through the R3R4 divider. Capacitors C1 and C6 are separate, and C1 is both concurrent.

The scheme of the mixer on the PTSH with the supply of signal fluctuations and the heterodyne to the gate through the directional separator is given in Fig. 10 b). In the role of the filter of the signal and the concurrent link, the segment of the microstrip line is used. (MSL) length $l_2 = \lambda/4$ and short-circuited spectrum length $l_1 = \lambda/4$. Through this spectrum, the bias voltage is applied to the gap (E3). In the output circuit of the PSTN transistor, in addition to the intermediate frequency filter (FPC), a high frequency filter (HF) is used, whose

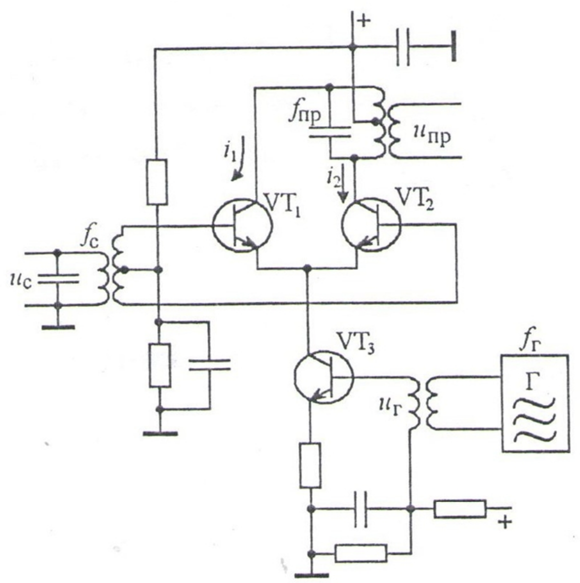

Fig. 9, a. Frequency converter, executed according to the balance scheme using the differential channel.

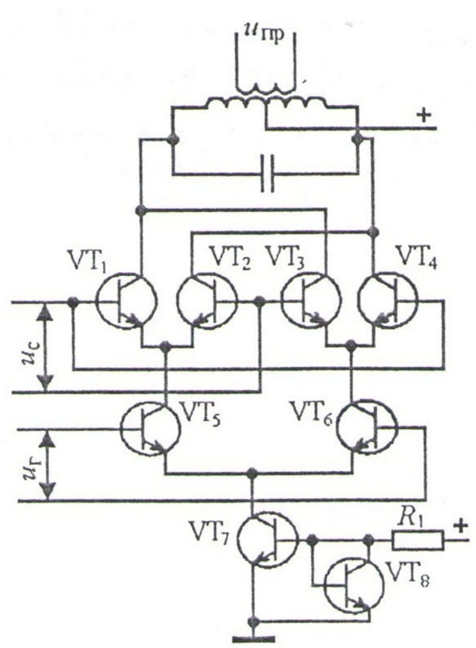

Fig. 9, b. Scheme of double balance mixer, which uses three differential pair of transistors

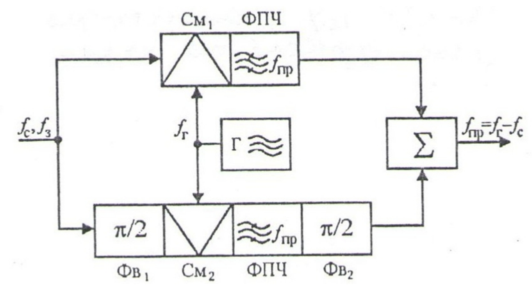

Fig. 9, c. Frequency converter with compensated interference to the mirror channel.

function is performed by an open segment length $l_3 = \lambda_г/4$,
then provides K.z at the frequency of the heterodyne frequency-weakening oscillations of the other combinational frequencies to eliminate the overload of the PPV by the voltage of the heterodyne. On the scheme C_p - separating capacitors $C_ф$, $R_ф$ – he compensator and the filter resistor in the power supply.

The disadvantage of the schemes given in Fig. 1.10, a), b) is the need for large power of the heterodyne because of its weakening in the guide deflector (AB). The scheme on the DZ PTSh, presented in Fig. 1.10, c), eliminates these disadvantages.

The signal fluctuations are brought to the first shutter through the segment of the microstrip line length $l_1 = \lambda_c/4$, and oscillations of the heterodyne through the length of the line segment $l_2 = \lambda_г/4$ to the second shutter. This ensures a qualitative resolution of the signal circles and the heterodyne without the use of bridges or directed illuminators (HB). Sections of microstrip lines in length l_1 i l_2 are the matching links between the signal source and the heterodyne and the transistor inputs. The initial displacement on the shutters is fed through the short-closed loops of length $l_{31} = \lambda c_m$ i $l_{31} = \lambda_m m$. Open loop length $l_3 = l_г/4$ provides K.S. flow for oscillation of the heterodyne. The conversion of microwave frequencies is characterized by multiple interactions of oscillations of combinational frequencies. In particular, there are effects that are caused by inverse and secondary transformations. Consider the manifestation these effects on the mixer device with a frequency difference $f_{пр} = |f_г - f_c|$. The spectra of the main frequencies of the mixer, when f below and above fc in Fig. 11. With the inverse transformation, the output voltage with the frequency fпр creates at the inlet of the mixer along with the voltage of the signal frequency $f_c = f_г \pm f_{пр}$. As a result of the secondary transformation, the input voltage of the mirror frequency forms at the output of the fluctuation mixer with the converted frequency $f_{пр}' = | f_{дз} - f_г| = f_{пр}$. The phases of this oscillation depend on many factors and, as a rule, differ from the phase of the intermediate frequency voltage, which is obtained at the basic

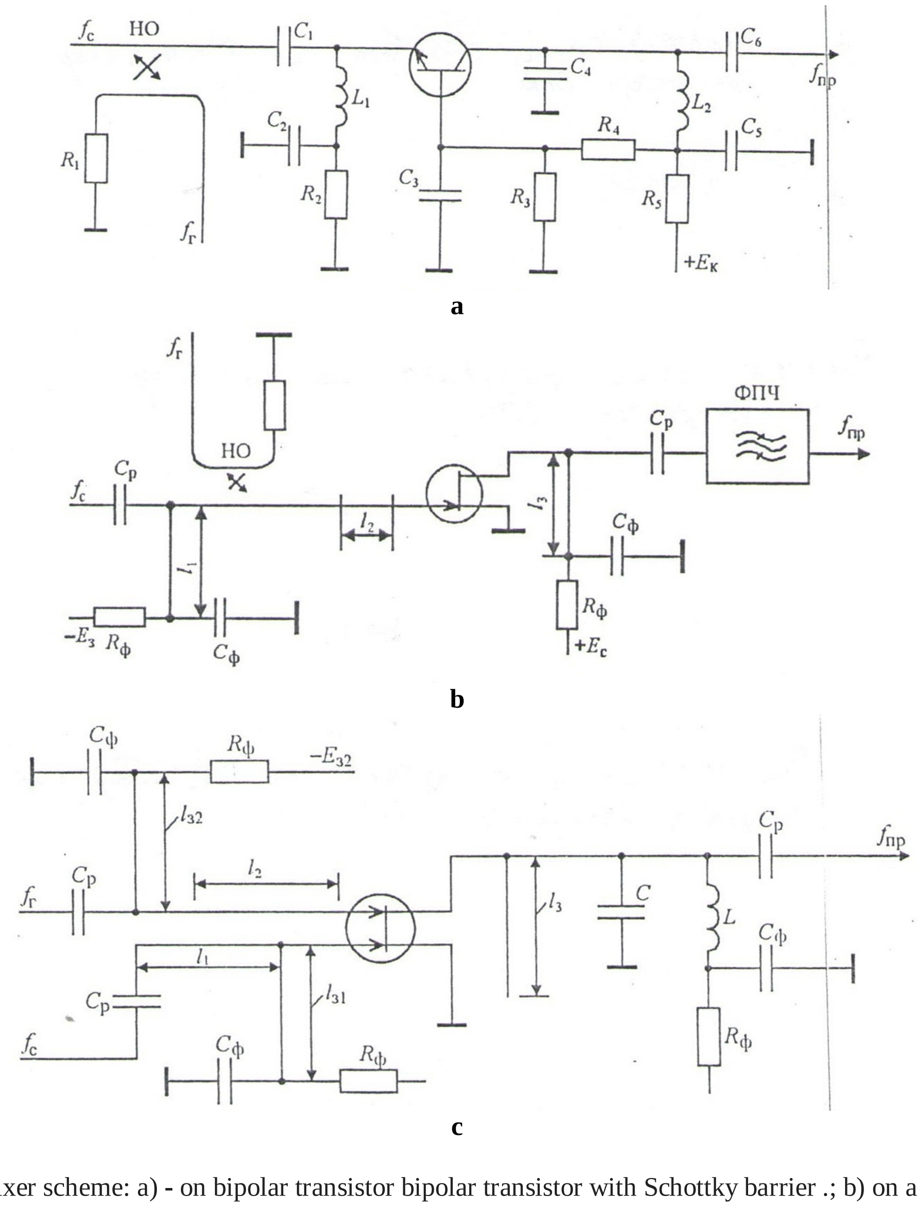

Fig. 10. Mixer scheme: a) - on bipolar transistor bipolar transistor with Schottky barrier .; b) on a field polar transistor with a Schottky barrier .; c) on a two-gate field transistor with a Schottky barrier.

transformation, which can occur before the occurrence of frequency and phase distortions of the signal.

The appearance of the oscillations of the mirror frequency in the mixer is possible and without inverse transformation due to the direct interaction of the oscillations of the signal frequency and the second harmonic of the heterodyne $2f_г - f_c = f_c \pm f_{пр} = f_{дз}$. In this ratio, the upper sign corresponds to the case $F_g < f_c$, and the bottom one – $f_g > f_c$.

As a result of direct transformation at the output for mirror frequency (DZCH), they can be significant and significantly dependent on the frequency.

To prevent frequency and phase distortion, due to the influence of fluctuations of the DPS and MF, mixers are used using the energy of combinational frequencies by reflecting or absorbing these fluctuations of special filter circuits. One of the variants of the construction of the mixer using the fluctuations of the DZCH and SM at $F_g < f_c$ shown in Fig. 12, a), where FS is a pass filter f_c and does not miss f_{dz}, FSF is a short-circuit filter on the output of the oscillation transistor $f_{сч}$.

Fluctuations in the mirror frequency (DF) occurring in the transistor and the variation of the total frequency (MF) at $f_g < f_c$ are reflected in accordance with the filters FS and FSF and return to the transistor. The phases of the reflected vibrations are selected by selecting the mixer at the same time as the difference f_{pr} there also appear oscillations of the total frequency (MF)$f_{сч} = f_g + f_{g1}$. With an inverse and a secondary direct transformation, it creates a voltage with the input frequency fc$(f_c'' = f_{сч} - f_г)$ and with frequency $f_{пр}$ at the outlet of the mixer, and with $f_г < f_c$ we have $f_{пр}'' = f_c'' - f_г$, at $f_r > f_c$ we get $f_{пр}'' = f_г - f_c''$. Phase shifts between the products of the main and secondary transformations, as well as distances l_1 and l_2 from these filters to the transistor. In the case of inhomogeneity of the oscillations of the main intermediate frequency and $f_{пр}$, obtained as a result of the transformation of reflected vibrations $f_{СЧ}$ i $f_{дз}$ the power properties of the mixer are noticeably improved help with an anti-phase bridge device (PMP). In the resulted scheme of Fig. 12, a) PMP at the output of the intermediate frequency on the drains of the transistor

Transistor mixers with absorption of mirror and total frequencies are more broadband, but somewhat inferior to the mixer reflecting tone for amplifying and noise characteristics. When constructing such mixers, FS and FC are commonly used. The scheme of such a mirror absorption mixer Frequency (DF) is shown in Fig. 11.

The ferrite circulator (FC) at the signal frequency works like a valve. Fluctuations of the mirror frequency (DF) coming from the transistor to the ange FC are absorbed in the coordinated load (OH). The use of the FF also reduces the emission of the heterodyne to the signal power circle.

Balanced transistor mixers (BTS) consist of two identical transistor double-acting cascades, in combination with bridge devices.

In addition, in contrast to balanced amplifiers in the BTZ, there is an additional possibility of setting the balance by selecting the phase of the heterodyte, and therefore, in the BTZ are also the use of both in-phase, and quadrature and bridge devices. In fig. 13, a) the circuit diagram of the balance mixer on the DZPTsh with the use of PMP is given. But the mixers with the reflection of the combination products are narrowband, since the inhospitality of the considered oscillations can only be achieved in a relatively small frequency range.

Here, the mixing transistors are excited by the signal antiphase, and the heterodyne is in phase. Output oscillations are phase-dependent, and therefore they must be overcome in the form of two phase shifters (f0) with a general phase shift and adder.

Balanced transistor mixers (BTS) significantly weaken the secondary receiving channels with the paired harmonics of the input signal. In particular, suppressed channels of the half-mirror frequency $f_{пдз} = f_c \pm 0{,}5f_{пр}$. Like balanced amplifiers, that is, differential, BTZ have a large dynamic range and a better static wave ratio compared to single-transistor amplifiers. The final transistor mixers (KTZs), shown in Fig. 13, b) are built on the basis of two balance mixers. They suppress adjacent channels of reception with pair harmonics and a signal, and a heterodyne. They have better characteristics in terms of saturation power and intermodulation distortion, which appear in response to several high-frequency interferences, compared with the BTZ. In the balance and ring transducers, channels of mirror frequency are not suppressed. They are weakened with

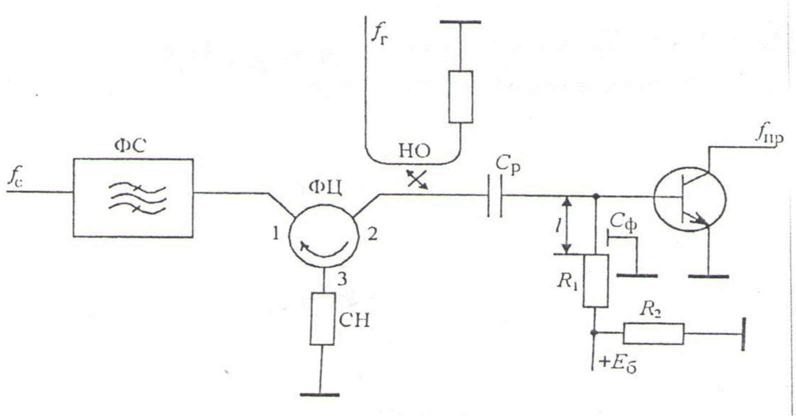

Fig. 11. Transistor mixer with absorption of mirror frequency and total frequency.

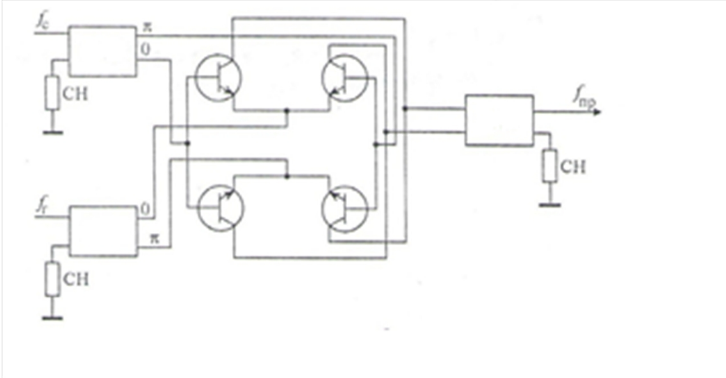

Fig. 12, a. Scheme of balance mixer.

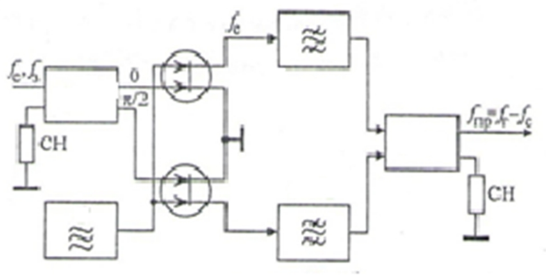

Fig. 12, b. Compensating mixer on two-wound polar transistor with Schottky barrier.

the help of a CD in the pre-selector, but the frequency requirements for the characteristics of such filters can not be performed, especially at low intermediate frequency(f_{np} << f_c). Two-channel compensation circuits, which consist of two mixers, in the role of which any of the considered ones can be used, are based on the resonance of the channels of the mirror frequency (DF) we have schemes in conjunction with phase shifters. An option for constructing a two-channel compensating mixer for two-wound PTs with DT feeds is shown in Figure 12 b).

Here in the role of the quadrature hanging signal at the input and quadrature oscillator adder f_{np}. At the output, a quadrature bridge device (KMP) is used to provide a phase shift between the output peaks equal to π / 2. The use of KMP facilitates coordination with signal source and load. The signal fluctuations are applied to the first shutter of the transistors with a shift π / 2, and oscillations of the heterodyte are weakened by means of a CD in the pre-selector, but the frequency requirements for the characteristics of such filters can not be performed, especially at low intermediate frequency (f_{np} << f_c). Two-channel compensation circuits, which consist of two mixers, in the role of which any of the schemes considered by us together with the phase rotators can be used, are based on the resonance of the channels of the mirror frequency (DF). Option for constructing a two-channel compensating mixer for two-wound PTs with a suppressor of the DF.

In the role of a quadrature signal divider on the input and a quadrature summator of the oscillation fpr on the output, a quadrature bridge device (KMP), which provides a phase shift between the output peaks equal to π / 2, is used. The use of KMP facilitates coordination with signal source and load. The signal fluctuations are applied to the first shutter of the transistors with a shift of π / 2, and the oscillations of the heterodyne are fed to the second gate of the same transistors in the same phase, that is, in-phase. After passing the output KMP, the output signals are composed in-phase, and the oscillations of the mirror frequency are extinguished in a coordinated load. The review analysis was carried out and the modern bases of the theory of signal CAD in the transformation of frequency on non-interconnected electronic devices - transistors and microwave diodes.

1. Schematic variants of frequency converters on field and bipolar transistors, Schottky diodes (varikaps) are considered.

2. The signaling theory of transistor and diode frequency converters is described: resistive and capacitive (parametric).

3. An original dot microwave diode on GaAs for mixers and frequency converters was developed.

4. The theory of frequency multipliers on microprocessor Fourier transforms for periodic, nonperiodic, discrete (digital) signals is developed and actually implemented, the circuitry of which will allow the manufacture of IFs in the integrated design.

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Abramova I.I. – student.

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Особливості сигнальної системи автоматизованого проектування перетворювачів частоти - систем надвисоких частот

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В даній статті проведено аналіз і викладено сучасні основи теорії сигнальної системи автоматизованого проектування при перетворенні частоти на невзаємному електронному приладі — транзисторі надвисоких частот. Крім того, розглянуті схемотехнічні варіанти перетворювачів частоти на польових та біполярних транзисторах, діодах Шотткі (варикапах); викладена сигнальна теорія транзисторних і діодних перетворювачів частоти: резистивних і ємнісних. Схемотехнічні виконання перетворювачів частоти представляють варіанти сигнальної системи автоматизованого проектування на надвисоких частотах.