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Theory of photoreactive effect in bipolar and MOSFET transistors

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ABSTRACT

The paper deals with the fundamentals of the theory of photoreactive effect in bipolar and field-effect transistor structures. Photoreactive properties of semiconductor devices are widely used in a variety of radio electronics devices. Therefore, the study of these phenomena in bipolar transistor structures with negative resistance, allows us to create new sensory devices, which have better parameters than existing ones. The method of construction of radiomeasuring microelectronic transducers is offered on the base of photoreactive effect in sensing bipolar and field transistor structures, that has established premises for embodying transducers of optical radiation with a frequency output signal.

Keywords: photoreactive effect, radiomeasuring microelectronic transducers, frequency optical transducer, negative differential resistance.

1. INTRODUCTION

The characteristics of transducers determine accuracy and reliability of systems of a radio control, instruments of monitoring of technological processes, environmental properties, safety of operation of kernel, thermal, chemical installations, flying apparatuses, sea plants, carrier etc. In regard to this matter the strict demands on transducers measuring the great variety of information are made. These devices should be cost-effective, noise-resistant to ensure high speed, sensitivity, measurement accuracy as well as to have whenever possible least overall dimensions and weight, be compatible with modern PCs and allow encoding the information during the transfer time on the long distances[1,2,3].

Therefore, one of perspective scientific directions to develop the transducers offered in the work, is usage of dependence of reactive properties and negative resistance of semiconductor devices on influence of external physical factors and making a new class of radiomeasuring microelectronic transducers of optical radiation on this base[4,5,6]. Thus, the need for development of qualitatively new theoretical approaches to making radiomeasuring microelectronic transducers has long been on the agenda as well as the development of their circuits and constructions, experimental research of their characteristics and metrological parameters[7,8].

2. THEORY OF PHOTOREACTIVE EFFECT IN BIPOLAR TRANSISTORS

Photoreactive properties of semiconductor devices are widely used in a variety of radio electronics devices. Therefore, the study of these phenomena in bipolar transistor structures with negative resistance, allows us to create new sensory devices, which have better parameters than existing one[9]. The complex physical processes related with the formation of an electric field in the base region of the transistoras and with the spatial distribution of photo-generated charge carriers in this region, occur under illumination. It causes an appearance of the photo-emf on the emitter and collector junctions, as well as the change in the resistance of the base region. All of these phenomena are superimposed on the injection processes of non-equilibrium carriers, both at the DC and AC voltage on the emitter junction. Therefore, it is necessary that a mathematical model of the photo-reactive effect taking into account these processes should be developed to calculate an impedance of the base region of the bipolar transistors. The calculation of impedance is needed for determination of the transformation function and sensitivity of optical frequency transducers.

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It is necessary to solve the equation of continuity and the Poisson's equation[10]. It would allow to determine the impedance of the base region of the bipolar transistor under illumination. The general one-dimensional continuity equation for holes and electrons is[10]:

$$\frac{\partial(p-p_n)}{\partial t} = G_p - \frac{1}{q}\frac{\partial j_p}{\partial x} - \frac{p-p_n}{\tau_p}, \quad \frac{\partial(n-n_p)}{\partial t} = G_n + \frac{1}{q}\frac{\partial j_n}{\partial x} - \frac{n-n_p}{\tau_n}. \tag{1}$$

where: G_n, G_p – the rate of carriers build up, j_p, j_n – the hole and electron current densities, τ_p, τ_n – the carrier lifetime, n, p are the non-equilibrium density for electrons and holes, p_n, n_p – the equilibrium density for electrons and holes, t – time, x – the direction in which the charge carrier density changes.

With the optical generation there is an photo-Dember electric field in the base region of the bipolar transistor, directed in such a way that it inhibits the diffusion of more mobile charge carriers and contributes to the diffusion of less mobile charge carriers. This phenomenon is called the bipolar diffusion. In view of that phenomenon and when substituting the values of the currents densities j_p and j_n the eqns. (1) is[10]:

$$\frac{\partial(p-p_n)}{\partial t} = D_p\frac{\partial^2 p}{\partial x^2} - \mu_p p\, div\vec{E} - \mu_p\left(\nabla p\vec{E}\right) + G_p - \frac{p-p_n}{\tau_p}, \; \frac{\partial(n-n_p)}{\partial t} = D_n\frac{\partial^2 n}{\partial x^2} + \mu_n n\, div\vec{E} + \mu_n\left(\nabla n\vec{E}\right) + G_n - \frac{n-n_0}{\tau_n}. \tag{2}$$

Expressions (2) are interconnected due to the action of an electric field $\vec{E}$ according to the Poisson's equation[10]

$$div\vec{E} = -\frac{4\pi}{\varepsilon}\rho = -\frac{4\pi q}{\varepsilon}\left[(n-n_p)+(p-p_n)\right],$$

where ρ – the space charge, formed by non-equilibrium charge carriers, ε – the dielectric constant of semiconductor.

Since there is electrical neutrality in the base region, then $n - n_p = p - p_n$, $\rho = 0$ and $div\vec{E} = 0$. It simplifies the eqns. (2). If we multiply each part of eqns. (2) to σ_{np}, σ_{pn} and add it one, then we can obtain[10]:

$$(\sigma_{p_n}D_n + \sigma_{n_p}D_p)\nabla^2 n + (\mu_n\sigma_{p_n} - \mu_p\sigma_{n_p})E\nabla n + G_n\sigma_{p_n} + G_p\sigma_{n_p} - \frac{n-n_p}{\tau}(\sigma_{p_n} + \sigma_{n_p}) = 0, \tag{3}$$

where σ_{p_n}, σ_{n_p} – the specific conductivities of the semiconductor determined by the corresponding charge carriers.

The considered one-dimensional case, as well as the stationary mode of operation of the transistor simplifies the appearance and solution of the eqn. (3). Taking these comments into account, eqn. (3) is expressed by[10]

$$\frac{D_n\sigma_{p_n} + D_p\sigma_{n_p}}{\sigma_{n_p} + \sigma_{p_n}} \cdot \frac{d^2(n-n_p)}{dx^2} + \frac{\mu_n\sigma_{p_n} - \mu_p\sigma_{n_p}}{\sigma_{n_p} + \sigma_{p_n}}E \cdot \frac{d(n-n_p)}{dx} + \frac{G_n\sigma_{p_n} + G_p\sigma_{n_p}}{\sigma_{n_p} + \sigma_{p_n}} - \frac{n-n_p}{\tau} = 0.$$

One can apply the following symbols

$$D = \frac{\left(D_n\sigma_{p_n} + D_p\sigma_{n_p}\right)}{\left(\sigma_{n_p} + \sigma_{p_n}\right)}, \tag{4}$$

with D – the bipolar diffusion coefficient,

$$\mu_E = \frac{\left(\mu_n\sigma_{p_n} - \mu_p\sigma_{n_p}\right)}{\left(\sigma_{n_p} + \sigma_{p_n}\right)}, \tag{5}$$

μ_E – the bipolar drift mobility.

Eqn. (3) based on eqn. (4) and eqn. (5) can be expressed as

$$D\frac{d^2(n-n_p)}{dx^2} + \mu_E E\frac{d(n-n_p)}{dx} - \frac{n-n_p}{\tau} = -G, \tag{6}$$

where $G = \frac{\left(G_n\sigma_{p_n} + G_p\sigma_{n_p}\right)}{\left(\sigma_{n_p} + \sigma_{p_n}\right)}$ – the bipolar generation rate.

Using the notations

$$\frac{\mu_E E}{D} = \frac{\mu_E E\tau}{D\tau} = \frac{l_E}{L^2}, \qquad L^2 = D\tau,$$

the eqn. (6) can be transformed:

$$\frac{d^2(n-n_p)}{dx^2}+\frac{l_E}{L^2}\frac{d(n-n_p)}{dx}-\frac{n-n_p}{L^2}=-\frac{G(x)}{D}, \tag{7}$$

When generating charge carriers under light takes place, the generation rate is described by the eqn[10].

$$G(x)=G(0)e^{-\alpha x},$$

with α – the absorption coefficient of light.

The solution of eqn. (7) consists of a general solution of a homogeneous equation and a partial solution of a nonhomogeneous equation. The general solution has the form

$$n(x)-n_p=A_1e^{K_1x}+A_2e^{K_2x},$$

where $K_1,\ K_2$ – roots of the quadratic equation

$$K^2+\frac{Kl_E}{L^2}-\frac{1}{L^2}=0,$$

which equal

$$K_{1,2}=-\frac{l_E}{2L^2}\pm\sqrt{\left(\frac{l_E}{2L^2}\right)^2+\frac{1}{L^2}}. \tag{8}$$

Then the general solution for the homogeneous eqn. (7), written in terms of eqn. (8), is:

$$(n(x)-n_p)=A_1\,exp\left(\left(\sqrt{\left(\frac{l_E}{2L^2}\right)^2+\frac{1}{L^2}}-\frac{l_E}{2L^2}\right)x+A_2\,exp\left(-\left(\frac{l_E}{2L^2}+\sqrt{\left(\frac{l_E}{2L^2}\right)^2+\frac{1}{L^2}}\right)x\right).\tag{9}$$

If the following notations are used,

$$\frac{1}{l_1}=\sqrt{\left(\frac{l_E}{2L^2}\right)^2+\frac{1}{L^2}}-\frac{l_E}{2L^2},\qquad \frac{1}{l_2}=-\left(\sqrt{\left(\frac{l_E}{2L^2}\right)^2+\frac{1}{L^2}}+\frac{l_E}{2L^2}\right),$$

then eqn. (9) can be expressed

$$n(x)-n_p=A_1e^{x/l_1}+A_2e^{-x/l_2}\ .$$

To determinate the coefficients $A_1,\ A_2$ one can use the following boundary conditions:

$$n'(x)|_{x=0}=n_p\left(exp\left(\frac{qU_{E0}}{kT}\right)-1\right),\qquad n'(x)\Big|_{x=W}=n_p\left(exp\left(\frac{qU_{K0}}{kT}\right)-1\right).$$

Making all the necessary transformations, gives value

$$A_1=\frac{(n'(W)-n'(0)e^{-W/l_2})}{(e^{W/l_1}-e^{-W/l_2})},\qquad A_2=\frac{(n'(0)e^{W/l_1}-n'(W))}{(e^{W/l_1}-e^{-W/l_2})},$$

Thus, the general solution of the homogeneous eqn. (9) has the following form

$$n(x)-n_p=\frac{(n'(W)-n'(0)e^{-W/l_2})}{(e^{W/l_1}-e^{-W/l_2})}e^{x/l_1}+\frac{(n'(0)e^{W/l_1}-n'(W))}{(e^{W/l_1}-e^{-W/l_2})}e^{-x/l_2}.$$

The partial solution of the nonhomogeneous eqn. (7) is found as $\bar{y}=R_k(x)e^{\alpha x}$, where $R_k(x)$ – the polynomial, which has a degree of k if right-hand side of equation has form $f(x)=Q_k(x)e^{\alpha x}$ [11,12]. Thus, the particular solution of eqn. (7) is described by the expression

$$n(x)-n_p=-\frac{(G(0)e^{-\alpha x})}{\left(D\left(\alpha^2-\alpha\frac{l_E}{L^2}-\frac{1}{L^2}\right)\right)}\ .$$

The general solution of eqn. (7) is given by

$$n(x)-n_p=A_1e^{x/l_1}+A_2e^{-x/l_2}-\frac{(G(0)e^{-\alpha x})}{\left(D\left(\alpha^2-\alpha\frac{l_E}{L^2}-\frac{1}{L^2}\right)\right)}. \tag{10}$$

In expression (10), the first two components describe the distribution of the charge carriers concentration in the base region of the bipolar transistor as a function of the action of the DC voltage, and the third component describes as function of the action of the optical radiation. Since the transistor operates under alternating voltages and currents, it is necessary that the distribution of the concentration of charge carriers for this case should be determined. The one-dimensional continuity equation for AC in stationary mode has the form

$$\frac{d^2(n_1)}{dx^2} + \frac{l_E}{L^2}\frac{d(n_1)}{dx} - \frac{n_1(1+j\omega\tau_n)}{L^2} = 0, \tag{11}$$

where n_1 – the concentration of injected carriers because of the action of the AC voltage at the emitter and collector junctions.

One can consider that the electron concentration consists of a AC component $n_{E0,K0}$ (function x) and DC component $n_1 e^{j\omega t}$ (functions x and t). If AC signals $U_{E1}(t)$ and $U_{K1}(t)$ are superimposed on the voltages U_{E0} and U_{K0} respectively, then the electron concentration at the emitter and collector junctions takes values

$$n_{E0} + n_{E1}(t) = n_p\, exp\left[\frac{q}{kT}\big(U_{E0} + U_{E1}(t)\big)\right], \quad n_{K0} + n_{K1}(t) = n_p\, exp\left[\frac{q}{kT}\big(U_{K0} + U_{K1}(t)\big)\right], \tag{12}$$

where $n_{E1}(t)$ and $n_{K1}(t)$ – the electron concentration, which are determined by the AC voltage imposed on the average electron concentrations, caused by DC voltage.

If the signal is small $\frac{qU_{E1}(t)}{kT} << 1$ and $\frac{qU_{K1}(t)}{kT} << 1$, eqns. (12) can be greatly simplified by decomposition into a number of exponential functions $exp\left(\frac{q}{kT}U_{E1}(t)\right)$ and $exp\left(\frac{q}{kT}U_{K1}(t)\right)$ with maintaining the first two components of the decomposition, yielding:

$$n_{E0} + n_{E1}(t) = n_p e^{\frac{q}{kT}(U_{E0})}\left[1 + \frac{qU_{E1}(t)}{kT}\right], \quad n_{K0} + n_{K1}(t) = n_p e^{\frac{q}{kT}(U_{K0})}\left[1 + \frac{qU_{K1}(t)}{kT}\right]. \tag{13}$$

The boundary conditions to use for solving the eqn. (9) are determined on the basis of (13). For the emitter junction

$$n_1^{'}(0,t) = n_p e^{\frac{q}{kT}(U_{E0})} + n_p e^{\frac{q}{kT}(U_{K0})}\frac{qU_{E1}(t)}{kT}, \tag{14}$$

and for the collector junction

$$n_1^{'}(W,t) = n_p e^{\frac{q}{kT}(U_{K0})} + n_p e^{\frac{q}{kT}(U_{K0})}\frac{qU_{K1}(t)}{kT}. \tag{15}$$

The eqn. (9) has a solution:

$$n_1(x,t) = A_3 e^{\frac{xC_B^*}{l_1}} + A_4 e^{-\frac{xC_B^*}{l_2}},$$

with $C_B^* = \sqrt{1 + j\omega\tau}$, τ – the electron lifetime in the base region, ω – the angular frequency.

The coefficients A_3 and A_4 are determined from the boundary conditions (14) and (15), hence

$$A_3 = \frac{n_1^{'}(W,t) - n_1^{'}(0,t)e^{-\frac{WC_B^*}{l_2}}}{e^{\frac{WC_B^*}{l_1}} - e^{\frac{WC_B^*}{l_2}}}, \qquad A_4 = \frac{n_1^{'}(0,t)e^{\frac{WC_B^*}{l_1}} - n_1^{'}(W,t)}{e^{\frac{WC_B^*}{l_1}} - e^{\frac{WC_B^*}{l_2}}}.$$

Thus, the general solution of eqn. (9) is given

$$n_1(x,t) = \left[\frac{\left(n_1^{'}(W,t) - n_1^{'}(0,t)e^{-\frac{WC_B^*}{l_2}}\right)}{\left(e^{\frac{WC_B^*}{l_1}} - e^{\frac{WC_B^*}{l_2}}\right)}\right] e^{\frac{xC_B^*}{l_1}} + \left[\frac{\left(n_1^{'}(0,t)e^{\frac{WC_B^*}{l_1}} - n_1^{'}(W,t)\right)}{\left(e^{\frac{WC_B^*}{l_1}} - e^{\frac{WC_B^*}{l_2}}\right)}\right] e^{-\frac{xC_B^*}{l_2}}.$$

It is necessary to estimate the electric field strength in the base region for determining the base region resistance of the bipolar transistor at the illumination and DC/AC voltages, using the equation

$$j_{gen} = q(\mu_n n + \mu_p p)E + q(D_n \nabla n - D_p \nabla p),$$

then

$$E = \frac{j_{gen} - q(D_n \nabla n - D_p \nabla p)}{q(\mu_n n + \mu_p p)}. \tag{16}$$

It is supposed that there is a high level of injection in the base region, if $n^{'} \geq p_p$, and the condition of neutrality is fulfilled. Accordingly, it follows that

$$\frac{\partial n}{\partial x} = \frac{\partial p}{\partial x} \text{ and } \frac{\partial n}{\partial t} = \frac{\partial p}{\partial t}.$$

Therefore, the eqn. (16) can be written

$$E = \frac{j_{gen}}{q\mu_p n(b+1)} - \frac{(D_n/\mu_p)\nabla n}{n(b+1)} + \frac{kT}{q}\frac{\nabla n}{n(b+1)},$$

where $b = \mu_n/\mu_p$. It should be noted that

$$j_{gen} = j_{cb} + j_{E0} + j_{E1}, \qquad \nabla n = \nabla n_{cb} + \nabla n_{E0} + \nabla n_{E1}, \qquad n = n_{cb} + n_{E0} + n_{E1},$$

then

$$E(x,\omega) = \frac{j_{cb}}{q\mu_p(b+1)(n_{cb}+n_{E0}+n_{E1})} + \frac{j_{E0}}{q\mu_p(b+1)(n_{cb}+n_{E0}+n_{E1})} +$$

$$+ \frac{j_{E1}}{q\mu_p(b+1)(n_{cb}+n_{E0}+n_{E1})} - \frac{(D_n/\mu_n)\nabla n}{(b+1)(n_{cb}+n_{E0}+n_{E1})} + \frac{kT\nabla n}{q(b+1)(n_{cb}+n_{E0}+n_{E1})}. \qquad (17)$$

Decreasing voltage in the base region is estimated by the expression

$$U_B = -\int_0^W E(x,\omega)dx. \qquad (18)$$

Substituting egn. (17) into egn. (18) gives

$$U_B = -\int_0^W \frac{j_{cb}}{q\mu_p(b+1)(n_{cb}+n_{E0}+n_{E1})}dx - \int_0^W \frac{j_{E0}}{q\mu_p(b+1)(n_{cb}+n_{E0}+n_{E1})}dx - \int_0^W \frac{j_{E1}}{q\mu_p(b+1)(n_{cb}+n_{E0}+n_{E1})}dx +$$

$$+ \int_0^W \frac{(D_n/\mu_p)\nabla n}{(b+1)(n_{cb}+n_{E0}+n_{E1})}dx - \frac{kT}{q}\int_0^W \frac{\nabla n}{q(b+1)(n_{cb}+n_{E0}+n_{E1})}dx\,. \qquad (19)$$

The impedance of base region is:

$$Z_B = \frac{U_{E1}}{Sj_{E1}},$$

where U_{E1} – the voltage at the base, determined by DC, S – the area of the base region.

Using components from eqn. (19) relating to the AC voltage one can write

$$Z_B = -\frac{2kT}{SqD_n(b+1)} \frac{\int_0^W \frac{A_3(C_{\text{б}}^*/l_1) - A_4(C_B^*/l_2)}{\frac{-(G(0)e^{-\alpha x})}{\left(D\left(\alpha^2 - \frac{\alpha l_E}{L^2} - \frac{1}{L^2}\right)\right)} + A_3 e^{xC_B^*/l_1} + A_4 e^{-xC_B^*/l_2}}dx}{A_3\frac{C_B^*}{l_1} - A_4\frac{C_B^*}{l_2}} + \frac{kT}{q(b+1)} \times$$

$$\times \int_0^W \frac{\alpha G(0)e^{-\alpha x}}{D\left(\alpha^2 - \frac{\alpha l_E}{L^2} - \frac{1}{L^2}\right)qD_nS\left[\frac{A_3C_B^*}{l_1} - \frac{A_4C_B^*}{l_1}\right]\left[A_3 e^{\frac{xC_B^*}{l_1}} + A_4 e^{\frac{-xC_B^*}{l_2}} - \frac{(G(0)e^{-\alpha x})}{\left(D\left(\alpha^2 - \frac{\alpha l_E}{L^2} - \frac{1}{L^2}\right)\right)}\right]}dx +$$

$$+ \frac{kT}{q(b+1)}\int_0^W \frac{A_3(C_B^*/l_1)e^{\frac{xC_B^*}{l_1}} - A_4(C_B^*/l_2)e^{-\frac{xC_B^*}{l_2}}}{qD_nS\left[\frac{A_3C_B^*}{l_1} - \frac{A_4C_B^*}{l_1}\right]\left[-\frac{(G(0)e^{-\alpha x})}{\left(D\left(\alpha^2 - \frac{\alpha l_E}{L^2} - \frac{1}{L^2}\right)\right)} + A_3 e^{xC_B^*/l_1} + A_4 e^{-xC_B^*/l_2}\right]}dx. \qquad (20)$$

The solution of integrals in expression (20) can be made by decomposition into a number of exponential functions and $\sqrt{1 + j\omega\tau}$ with saving two members of decomposition:

$$Z_B = \frac{2kT}{q(b+1)\left[I_{E0} + \frac{(G(0)\alpha Sq)}{\left(\alpha^2 - \alpha\frac{l_E}{L^2} - \frac{1}{L^2}\right)}\right]}\left[1 + \frac{\frac{((1+j\omega\tau)W^2)}{(2D\tau)}\left(\frac{qU_{sat}}{kT} - 1\right)}{1 + \frac{1}{2}(j\omega\tau)\frac{W^2}{D\tau}}\right], \qquad (21)$$

where

$$\frac{qU_{sat}}{kT} = ln\left[\frac{\left(n_{E0} + \frac{b}{b+1}(p_n - n_p)\right)}{\left(n_{K0} + \frac{b}{b+1}(p_n - n_p)\right)}\right].$$

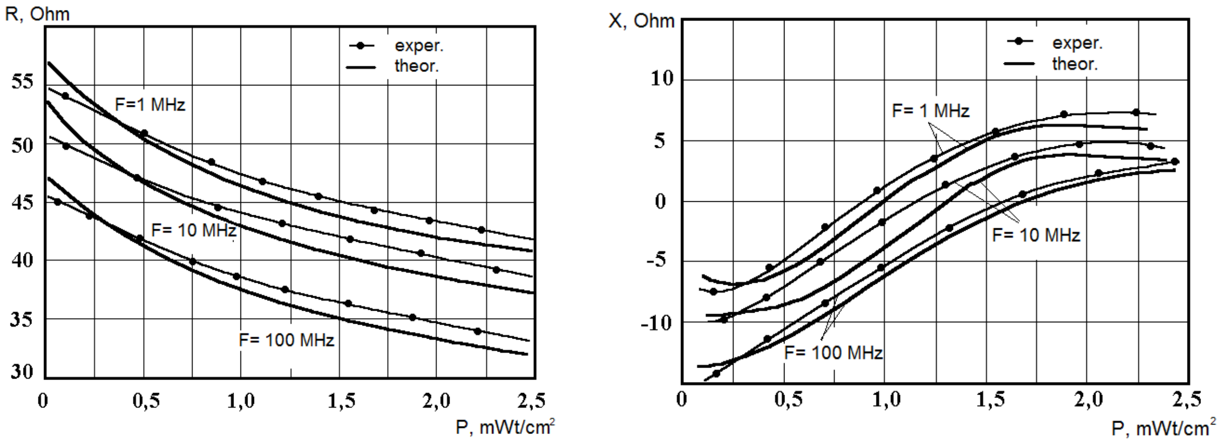
For operation of transistor at the frequencies $\omega\tau >> 1$, eqn. (21) can be simplified

$$Z_B = \frac{2kT}{q(b+1)\left[I_{E0}+\frac{(G(0)\alpha Sq)}{\left(\alpha^2-\alpha\frac{l_E}{L^2}-\frac{1}{L^2}\right)}\right]}\left[1+\frac{j\omega L_B/r_1}{1+j\omega L_B/r_2}\right], \tag{22}$$

with

$$r_1 = \frac{2kT}{q(b+1)\left[I_{E0}+\frac{G(0)\alpha Sq}{\left(\alpha^2-\alpha\frac{l_E}{L^2}-\frac{1}{L^2}\right)}\right]},\ r_2 = \frac{2kT\left(\frac{qU_{sat}}{kT}-1\right)}{q(b+1)\left[I_{E0}+\frac{G(0)\alpha Sq}{\left(\alpha^2-\alpha\frac{l_E}{L^2}-\frac{1}{L^2}\right)}\right]},\ L_B = \frac{kT\frac{W^2}{D\tau}\left(\frac{qU_{sat}}{kT}-1\right)}{q(b+1)\left[I_{E0}+\frac{G(0)\alpha Sq}{\left(\alpha^2-\alpha\frac{l_E}{L^2}-\frac{1}{L^2}\right)}\right]}.$$

The dependencies of the active and reactive components of the impedance of the bipolar transistor on the power of optical radiation are shown in Fig. 1.



a) b)

Fig. 1. Dependencies of active (a) and reactive (b) components of impedance of the bipolar transistor on the power of optical radiation

The active component decreases, and the reactive component changes from the capacitive into the inductive as shown in Fig. 1. The influence of optical radiation on the base region of the bipolar transistor is equivalent to the additional injection of electrons creating an excess negative charge near the emitter. This charge increases the electric field between the emitter and the collector. This field, in turn, causes the drift flow of the majority carriers (holes) in the direction towards the emitter that leads to an increasing of the hole density near the emitter, which tries to compensate for the charge of injected electrons and an electric field. The equilibrium is achieved when the increase of the charge density becomes sufficient to make an electric field having magnitude which able to create the electrons distribution almost similar in shape to the hole distribution. Under these conditions the applying of an AC signal leads to a change of the number of carriers over time in each local region of the base, which, in turn, implements rearrangement of the electric field in the base. Certainly, these processes are inertial towards the process of voltage variation on the emitter p-n junction. The voltage drop on the p-n junction is formed during the time interval, which is a lot less than the holes lifetime in the base (about an order of magnitude). The duration of the process of changing of the resistance of the base region, resulting from redistribution of charge carriers, is also smaller than their lifetime, but it is much longer than the time required to set the voltage on the p-n junction. The lag of the charge change in the base compared to the voltage is perceived at the external terminals as an inductive reaction.

The magnitude of the inductive resistance increases with increase of frequency until the transit time of the minority charge carriers through base region becomes commensurable with the period of the applied oscillations. A further increase of the frequency leads to the fact that the concentration of minority carriers in the base is not able to follow the change of the limiting values of the AC, that creates a decrease in inductance. Thus, obtaining of the dependence of the currents on the voltages and light in the static and dynamic modes makes possible the determining the equivalent circuit of a bipolar transistor with parameters depending on the optical radiation.

3. THEORY OF THE PHOTOREACTIVE EFFECT IN FIELD-EFFECT TRANSISTORS

The metal–oxide–semiconductor field-effect transistors (MOSFETs) are widely used in optical information receiving and processing systems. Theoretical questions of the optical radiation effect associated with the photo-conductivity effect have not been developed for case of influence of a small alternating signal and optical radiation on the conductive channel in dynamic mode. It is assumed that the interaction of optical radiation with a semiconductor crystal is based on penetration between electrodes and, as a result, the photo-voltaic effects prevail over the effects of photo-conductivity[13,14,15]. These issues acquire special relevance in the context of the development of optically controlled microwave devices and, in particular, microelectronic frequency sensory devices with optical rearrangement[16,17,18]. The common-source circuit provides effective control of the reactive component of the impedance of the transistor by controlling voltage, but the action of optical radiation on the crystal of the device does not exhibits sufficient activity. The circuit of reactive two-port network is realized on the basis of the use of the impedance of the distributed source-drain structure, whose small-signal parameters more depends on optical radiation.

To obtain the basic analytic relations, we use the common structure of a MOSFET with an induced p-channel[19], in which a harmonic voltage with a circular frequency ω a small amplitude of the signal $qU_1/kT << 1$ is applied between the source and the drain. At the same time, there are the uniformity and non-degeneracy of the semiconductor material as well as the steady mobility of the charge carriers in the channel, the absence of an optical stimulated recharge of surface states and photo-emission into the dielectric, the homogeneity of the lifetime of non-equilibrium charge carriers, and the absence of capture of excess charge carriers in the bulk and on the surface of the semiconductor. It is considered that the energy of the quantum of optical radiation exceeds the width of the energy bandgap of the semiconductor and the bipolar photogeneration takes place. The highest photosensitivity is inherent in the pre-threshold mode of operation of the MOSFET. Under the weak inversion, the main component of current is the diffuse one[20]. The one-dimensional continuity equation for the p-channel is:

$$\frac{\partial p}{\partial t} = \nabla(D\nabla p) - \left(\frac{\mu_E j}{\sigma \nabla p}\right) - \frac{p-p_0}{\tau_p} + G, \tag{23}$$

where

$$D = \frac{D_n D_p}{D_p p + D_n n} = \frac{2}{b+1} D_p \tag{24}$$

– the diffusion coefficient for bipolar drift,

$$\mu_E = \frac{\mu_n \mu_p (n-p)}{\mu_n n + \mu_p p} \tag{25}$$

– the bipolar drift mobility, j – the total current density, σ – the conductivity of the channel, τ_p – the hole lifetime, $p,\ n$ – the concentration of non-equilibrium holes and electrons in the channel, $\mu_p,\ \mu_n$ are the electron and hole mobility respectively, $b = \mu_n/\mu_p$ – the relation of the electron and hole mobilities, $D_n,\ D_p$ – the electron and hole diffusion constant respectively, G – the carrier generation rate.

In the stationary mode of the MOSFET operation, taking into consideration eqns. (24) and (25), eqn. (23) becomes:

$$D\frac{d^2(p-p_0)}{dx^2} - \mu_E E\frac{d(p-p_0)}{dx} - \frac{p-p_0}{\tau} = -G(x)\,. \tag{26}$$

Using the designations:

$$\frac{\mu_E E}{D} = \frac{\mu_E E\tau}{D\tau} = \frac{l_E}{{L_*}^2}, \qquad {L_*}^2 = D\tau,$$

eqn. (26) can be written as

$$\frac{d^2(p-p_0)}{dx^2} - \frac{l_E}{{L_*}^2}\frac{d(p-p_0)}{dx} - \frac{p-p_0}{{L_*}^2} = -\frac{G(x)}{D}\,. \tag{27}$$

When the charge carriers are generated under the illumination, the generation rate is described by the eqn.[23]:

$$G(x) = G(0)e^{-\alpha x},$$

with α – the absorption coefficient of light.

Assuming that the non-equilibrium hole concentration consists of two components

$$p = p_{01} + p_1 e^{j\omega t},$$

where p_{01} – the hole concentration defined by the DC voltage, a p_1 – the hole concentration defined by the AC voltage, then the eqn. (27) for the alternating component in the small signal mode has the form

$$\frac{d^2 p_1}{dx^2} - \frac{l_E}{{L_*}^2}\frac{dp_1}{dx} - \frac{p_1(1+j\omega t)}{{L_*}^2} = -\frac{G(x)}{D}\,. \tag{28}$$

The boundary conditions in the absence of a forward bias on the drain to solve eqn. (28) are

$$\begin{aligned} p_1(0,t) &= p_0\, exp[\beta(\psi_S + \varepsilon_b)] + p_0\, exp[\beta(\psi_S + \varepsilon_b)]\,\beta U_{1b}(t), \\ p_1(L,t) &= p_0\, exp[\beta(\psi_S + \varepsilon_C)] + p_0\, exp[\beta(\psi_S + \varepsilon_C)]\,\beta U_{1C}(t), \end{aligned} \tag{29}$$

where $\beta = kT/q$, ε_b, ε_C – the photo-emf on the p-n junctions of the drain and source of transistor, L – the channel length, ψ_S – the surface potential, which is related to the gate voltage and the effective level of photogeneration by the expression $\xi = \Delta n/n_i = \Delta p/n_i$ [24]

$$U_G - U_{FB} = \psi_S + \frac{\varepsilon_S \varepsilon_0 \varphi_S}{\beta C_0 L_{Di}^* |\varphi_S|} F(\varphi_S \varphi_F^*), \tag{30}$$

with U_{FB} – the potential of planar zones, φ_F – Fermi potential, C_0 – the dielectric capacitance, n_i – the intrinsic carrier concentration, ε_S, ε_0 – permittivities of semiconductor and vacuum respectively.

In the weak inversion mode the following approximation can be used[17]:

$$F(\varphi_S, \varphi_F^*) \approx \sqrt{(\varphi_S - 1)e^{\varphi_F^*}}\,. \tag{31}$$

Substituting eqn. (31) into eqn. (30) gives

$$\varphi_S = U_3 - U_{FB} - \sqrt{\frac{\varepsilon_S \varepsilon_0}{\beta^2 C_0 L_{Di}^*}(\beta U_3 - 1)e^{\varphi_F^*/2}}\,.$$

The solution of eqn. (27) consists of the general solution of the homogeneous equation and the particular solution of the nonhomogeneous equation. The general solution of the homogeneous equation can be written by

$$P_1(x,t) = A_1 e^{K_1 x} + A_2 e^{K_2 x},$$

where K_1 and K_2 are the roots of the quadratic equation:

$$K^2 - \frac{l_E}{{L_*}^2}K - \frac{1}{{L_*}^2} = 0\,.$$

Thus, we obtain

$$K_{1,2} = \frac{l_E}{2{L_*}^2} \pm \sqrt{\left(\frac{l_E}{2{L_*}^2}\right)^2 + \frac{1}{{L_*}^2}}\,.$$

The coefficients A_1 and A_2 are defined from the boundary conditions (28) and (29), yielding

$$A_1 = \frac{\left(P_1(L,t) - P_1(0,t)e^{-\frac{LC_B^*}{l_2}}\right)}{\left(exp\left(\frac{LC_B^*}{l_1}\right) - exp\left(-\frac{LC_B^*}{l_2}\right)\right)},\ A_2 = \frac{\left(P_1(0,t)e^{\frac{LC_B^*}{l_1}} - P_1(L,t)\right)}{\left(exp\left(\frac{LC_B^*}{l_1}\right) - exp\left(-\frac{LC_B^*}{l_2}\right)\right)},$$

where $C_B^* = \sqrt{1 + j\omega\tau}$.

Thus, the general solution of the homogeneous eqn. (28) is described by

$$P_1(x,t) = \left(\frac{P_1(L,t) - P_1(0,t) e^{-\frac{LC_B^*}{l_2}}}{e^{\frac{LC_B^*}{l_1}} - e^{-\frac{LC_B^*}{l_2}}} \right) e^{\frac{xC_B^*}{l_1}} + \left(\frac{P_1(0,t) e^{\frac{LC_B^*}{l_2}} - P_1(L,t)}{e^{\frac{LC_B^*}{l_1}} - e^{-\frac{LC_B^*}{l_2}}} \right) e^{-\frac{xC_B^*}{l_2}} .$$

A particular solution of the nonhomogeneous eqn. (27) can be found in the form $\bar{P}(x) = R_K e^{\alpha x}$, where R_K – the polynomial of degree k, if right-hand side of equation has form $f(x) = Q_K e^{\alpha x}$[10]. Thus, the particular solution of eqn. (27) is written as

$$\bar{P}(x) = \frac{(G(0) e^{-\alpha x})}{\left(D \left[\alpha^2 - \alpha \frac{l_E}{{L_*}^2} - \frac{1}{{L_*}^2} \right] \right)},$$

and so the general solution of eqn. (27) can be rewritten as

$$P_1(x,t) = \left(\frac{P_1(L,t) - P_1(0,t) e^{-\frac{LC_B^*}{l_2}}}{e^{\frac{LC_B^*}{l_1}} - e^{-\frac{LC_B^*}{l_2}}} \right) e^{\frac{xC_B^*}{l_1}} + \left(\frac{P_1(0,t) e^{\frac{LC_B^*}{l_2}} - P_1(L,t)}{e^{\frac{LC_B^*}{l_1}} - e^{-\frac{LC_B^*}{l_2}}} \right) e^{-\frac{xC_B^*}{l_2}} - \frac{G(0) e^{-\alpha x}}{D \left[\alpha^2 - \alpha \frac{l_E}{{L_*}^2} - \frac{1}{{L_*}^2} \right]}$$

To determine the resistance of the channel under the illumination and alternating voltage, it is necessary to determine the electric field in the channel. Using the following equation

$$j_{gen} = q(\mu_n n + \mu_p p)E + q(D_n \nabla n - D_p \nabla p),$$

we have

$$E = \frac{j_{gen} - q(D_n \nabla n - D_p \nabla p)}{q(\mu_n n + \mu_p p)}. \tag{32}$$

Since there is an electrical neutrality condition in the channel, then $\frac{\partial n}{\partial x} = \frac{\partial p}{\partial x}$ and $\frac{\partial n}{\partial t} = \frac{\partial p}{\partial t}$. Taking these remarks into account, eqn. (32) takes the form

$$E = \frac{j_{gen}}{q\mu_p p(b+1)} - \frac{kT \nabla p}{q(b+1)p} .$$

It should be noted that

$$j_{gen} = j_{cb} + j_1, \qquad \nabla p = \nabla p_{cb} + \nabla p_1, \qquad p = p_{cb} + p_1,$$

then

$$E(x,\omega) = \frac{j_{cb}}{q\mu_p(b+1)(p_{cb}+p_1)} + \frac{j_1}{q\mu_p(b+1)(p_{cb}+p_1)} - \frac{kT \nabla p_{cb}}{q(b+1)(p_{cb}+p_1)} - \frac{kT \nabla p_1}{q(b+1)(p_{cb}+p_1)}. \tag{33}$$

The voltage drop on the channel is described by

$$U_K = -\int_0^L E(x,\omega) dx. \tag{34}$$

Substituting eqn. (33) into eqn. (34) gives

$$U_K = -\int_0^L \frac{j_{cb}}{q\mu_p(b+1)(p_{cb}+p_1)} dx - \int_0^L \frac{j_1}{q\mu_p(b+1)(p_{cb}+p_1)} dx + \int_0^L \frac{kT \nabla p_{cb}}{q(b+1)(p_{cb}+p_1)} dx + \int_0^L \frac{kT \nabla p_1}{q(b+1)(p_{cb}+p_1)} dx . \tag{35}$$

The impedance of the MOSFET channel is defined by

$$Z_K = \frac{U_1}{S j_1},$$

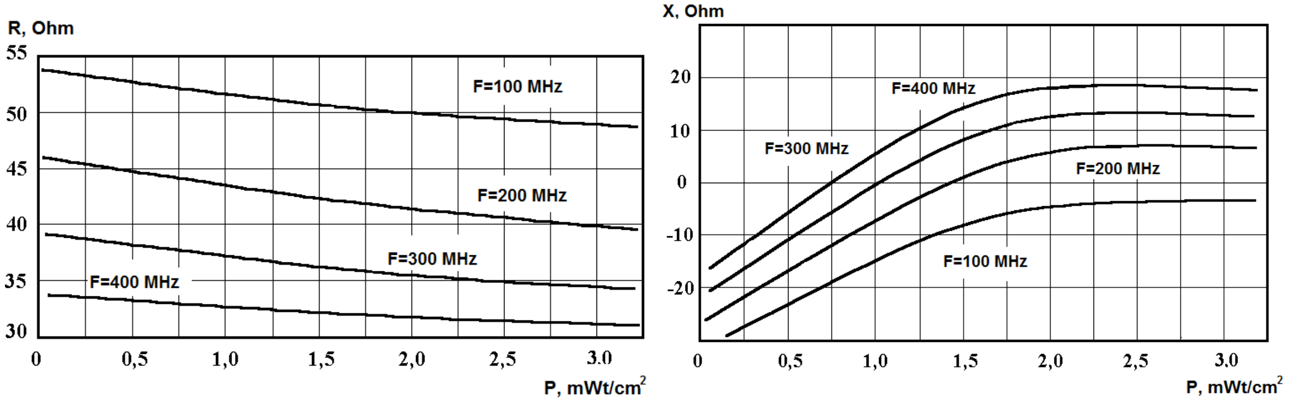
where U_1 – the AC voltage on the channel, S – the area of the channel.

Using the components from the expression (35) related to the AC voltage, one can write

$$Z_K = -\frac{1}{q\mu_p S(b+1)} \cdot \frac{\int_0^L \frac{A_1\frac{C_B^*}{l_1} - A_2\frac{C_B^*}{l_2}}{A_1\frac{C_B^*}{l_1}e^{\frac{xC_B^*}{l_1}} + A_2\frac{C_B^*}{l_2}e^{-\frac{xC_B^*}{l_2}} - \frac{(G(0)e^{-\alpha x})}{\left(D\left[\alpha^2 - \frac{\alpha l_E}{L_*{}^2} - \frac{1}{L_*{}^2}\right]\right)}} dx}{A_1\frac{C_B^*}{l_1} - A_2\frac{C_B^*}{l_2}} -$$

$$- \frac{kT}{q^2(b+1)SD_p} \cdot \frac{\int_0^L \frac{A_1\frac{C_B^*}{l_1}e^{\frac{xC_B^*}{l_1}} - A_2\frac{C_B^*}{l_2}e^{-\frac{xC_B^*}{l_2}}}{A_1\frac{C_B^*}{l_1}e^{\frac{xC_B^*}{l_1}} + A_2\frac{C_B^*}{l_2}e^{-\frac{xC_B^*}{l_2}} - \frac{(G(0)e^{-\alpha x})}{\left(D\left[\alpha^2 - \frac{\alpha l_E}{L_*{}^2} - \frac{1}{L_*{}^2}\right]\right)}} dx}{A_1\frac{C_B^*}{l_1} - A_2\frac{C_B^*}{l_2}} . \qquad (36)$$

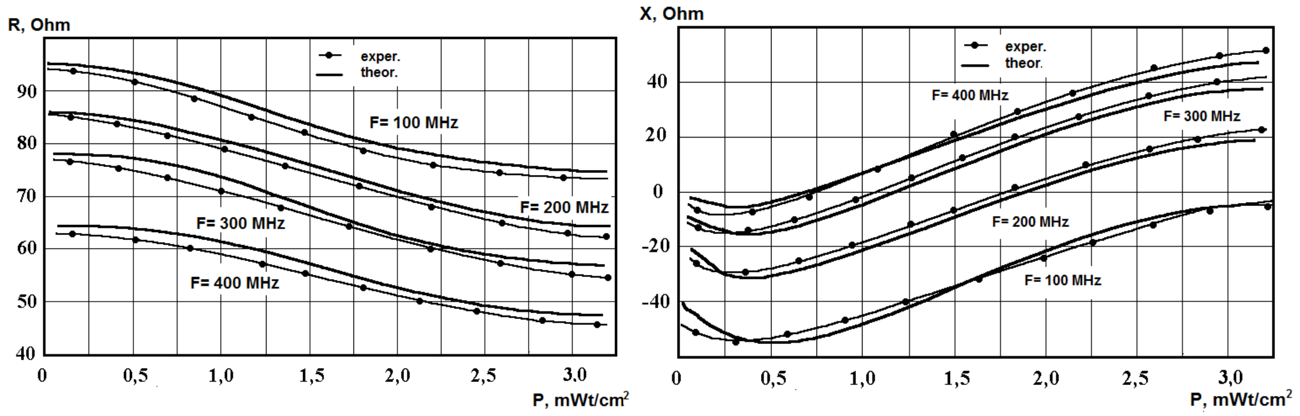
The solution of the integrals in eqn. (36) was obtained by a numerical method using PC. By separating expression (36) into the real and imaginary parts, the active and reactive components of the channel impedance can be defined. The dependencies of the active and reactive parts of the channel impedance on the power of optical radiation are shown in Fig. 2. The complete model of the photo-reactive MOSFET also requires consideration of the influence of the small-signal parameters of the crystal's active zone, the photo-diode structures of the source and drain, as well as the parasitic parameters of the hull.



a) b)

Fig. 2. Theoretical dependencies of active and reactive components of the impedance on power of optical radiation

The calculated and experimental dependences of the active and reactive components of the impedance of the MOSFET on the power of the optical radiation and the frequency of the signal are shown in Fig. 2. There was a decrease in the capacitive component, which is due to inertial processes of transfer of charge carriers through the channel of the transistor at frequencies exceeding the limiting ones. The phenomenon of phase delay is perceived as an inductive reaction, when current lags the voltage at the external electrodes of the device[10]. The physical nature of the photo-inductive effect in MOSFETs is analogous to the mechanism that occurs in photocells and bipolar transistors, which is confirmed by the qualitative coincidence of the obtained curves with the results of modeling the impedance of these structures[22]. Accuracy of developed mathematical model is ±5%.



a) b)

Fig. 3. Theoretical and experimental dependencies of active (a) and reactive (b) components of the impedance on power of optical radiation and frequency of signal

Fig. 3 shows the experimental dependencies for the transistors KP301B. At the increase in the frequency of the alternating signal, the point of inversion of the property of the reactance goes over to the region of lower optical power levels, as well as it occurs in bipolar photosensitive transistors[10]. As studies have shown, the use of the body in the circuit-making synthesis of an optically controlled reactive element leads to a negative effect, which is shown up in the almost total absence of a photo-induced increment of the impedance. Perhaps, the reason for this is the change of conditions for the photovoltaic distribution of charge carriers in the source (drain)-body structure.

4. CONCLUSIONS

The paper presents the basics of the theory of photoreactive effect in bipolar and field-effect transistor structures with negative differential resistance. The method of construction of radiomeasuring microelectronic transducers is offered on the base of photoreactive effect in sensing bipolar and field transistor structures, that has established premises for embodying transducers of optical radiation with a frequency output signal and microelectronic technology of manufacture.

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