FLUCTUATIONS IN LC OSCILLATORS

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Abstract

An analysis of the phase and amplitude fluctuations in oscillators with simple resonant circuit is presented. Negative feedback is used to minimize effect of the inherent noise produced by bipolar transistor on fluctuation characteristics.

Keywords:

noise in transistor, phase noise, amplitude noise, power spectrum characteristics

1. Introduction

Despite the large number of articles, describing the fluctuation characteristics of transistor oscillators, there is no systematic analysis of fluctuations in bipolar transistor oscillators for a wide frequency band with a bipolar transistor model that is accurate enough.

The purpose of this article is to explain the basic stages and results of fluctuation analysis, to determine their minimal values, and to set the requirements for real circuits. As a bipolar transistor model we will use a nonlinear charge-control model with piece-wise-linear approximation [10].

To explain the periodically modulated inherent noise sources, the structures mentioned in articles [11], [12] are used. To describe the $1/f$ noise, the model stated in [13] will be used. The influence of a negative feedback, phasing circuit, as well as the influence of the automatic bias circuit on lowering the fluctuation level of the oscillations are analysed here. A more detailed survey of these problems can be found in [14].

To write the basic equation for a three-terminal transistor oscillator we use substitution diagrams for high and low frequencies (Fig.1,Fig.2). The first diagram makes it possible to obtain equations for the amplitude and phase of oscillations, the second one is for automatic bias.

In Fig.1 there are current sources i_{b1n} , i_{c1n} containing spectral components of bipolar transistor calculated intrinsic noises, distributed close to oscillation frequencies. In Fig.2, current sources I_{b0n} , I_{c0n} contain the low frequency spectral components of calculated intrinsic noise currents. As shown in [11], [12], [14], their influence on all voltages and currents in external circuits is equivalent to the influence of all bipolar transistor intrinsic noise sources. Let us make an important assumption: in Fig.1 and Fig.2 there are all current sources calculated and due to the influence of emitter lead inductance L_E as shown in [12], [14].

Due to the existence of these components, the voltage-to-current transfer ratio between the base and terminal with of the transistor and cut off frequency is calculated. (The cut off frequency at $L_E = 0$, $R_E = 0$ will be refereed to as ω_{s0}) The influence of R_E will be reflected in changes of calculated transistor characteristics.

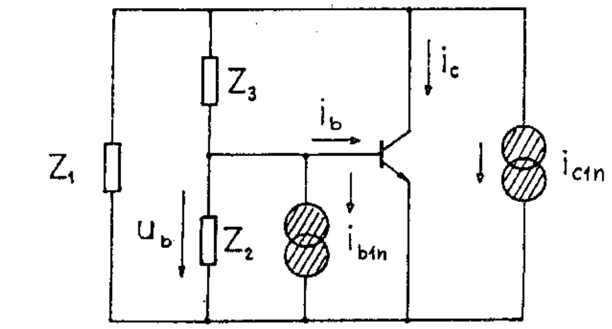

Fig.1
Equivalent circuit of the oscillator for high frequencies

2. Basic equations for nonlinear analysis with inherent noise

Let us write a symbolic equation for the diagram in Fig.1.

The derivation operator will be referred to as $p = d/dt$ and the symbolic controlled resistor of feedback circuit

$$Z_y(p) = -\frac{Z_1(p)Z_2(p)}{Z_1(p)+Z_2(p)+Z_3(p)}$$

together with equivalent transmission coefficient will be written

$$k_b(p) = Z_2(p)[Z_1^{-1}(p) + Z_y^{-1}(p)]$$

This coefficient expresses the influence of base current on circuit oscillations. Then the symbolic equation linking high frequency voltage on base u_b with controlled currents of collector i_c, base i_b and calculated noise of currents i_{c1n} , i_{b1n} will have the following form

$$u_b = Z_y(p)[i_c - k_b(p)i_b + i_{c1n} - k_b(p)i_{b1n}] \qquad (1)$$

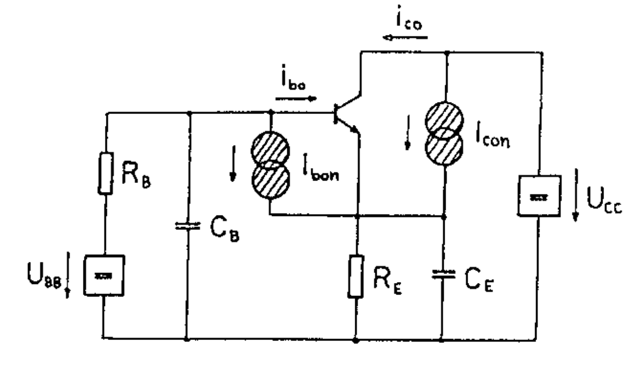

Fig.2
Bias equivalent circuit of the oscillator for low frequencies

The symbolic equation of the diagram shown in Fig.2 for low frequency components of currents I_{c0} , I_{b0} and voltage U_{b0} has the form

$$U_{b0} = U_{BB} - Z_B(I_{b0} + I_{b0n}) - Z_E(I_{c0} + I_{b0} + I_{c0n} + I_{b0n}) \qquad (2)$$

where

$$Z_B = \frac{R_B}{1 + pR_BC_B}$$

and

$$Z_E = \frac{R_E}{1 + pR_EC_E}$$

Let us assume that voltage $u_b(t)$ is almost harmonic, its frequency ω_0 is not exactly equal to resonant frequency ω_k of the resonant circuit consisting of impedances $Z_1(p)$, $Z_2(p)$, $Z_3(p)$. Relative frequency deviation

$$\Delta\omega/\omega_0 = (\omega_0 - \omega_k)/\omega$$

is comparable in order with damping factor of resonant circuit $\delta = 1/Q$. Then the following can be written

$$u_b = R_e[\mathbf{U}_b(t)]\exp(j\omega_k t)$$

where $\mathbf{U}_b(t)$ is the slowly changing complex amplitude.

Equation (1) can be rewritten in short form. In this equation, the relation between complex amplitude $\mathbf{U}_b$, complex amplitudes of first harmonic components of currents $\mathbf{I}_{c1}$, $\mathbf{I}_{b1}$ and complex amplitudes $\mathbf{I}_{c1n}$, $\mathbf{I}_{b1n}$ of calculated noise currents are shown as follows

$$(pT_Q + 1)\mathbf{U}_b = R(\mathbf{I}_{c1} - K\mathbf{I}_{b1}) + R(\mathbf{I}_{c1n} - K\mathbf{I}_{b1n}) \qquad (3)$$

where $R = KX_1^2Q/\rho$ is control resistor of feedback circuit on the resonant frequency, $T_Q = 2Q/\omega_0$ is time constant of the resonant circuit, ρ is characteristic resistor of the resonant circuit, $X_1 = R_eZ_1(p)$, K is transmission coefficient $k_b(j\omega_0)$ without δ order components.

Automatic bias equation (2), when condition $T_B = T_E = T_c$ is fulfilled, can be written as follows

$$(pT_c + 1)U_{b0} = U_{BB} - R_cI_{c0} - R_c(a_cI_{c0n} + a_bI_{b0n}) \qquad (4)$$

where

$$R_c = R_E + \frac{R_E + R_B}{h_{21E}},\ a_b = \frac{R_E + R_B}{R_c},\ a_c = \frac{R_E}{R_c}$$

h_{21E} - current gain of the transistor on low frequencies.

Equations (3), (4) describe stationary and transient processes in transistor oscillators with one resonant circuit and with automatic bias. We assume, however, that the dependence of I_{c0} , I_{b0} mean values (in oscillation period) and complex amplitudes $\mathbf{I}_{c1}$, $\mathbf{I}_{b1}$ on voltages $\mathbf{U}_b$, U_{b0} is known.The results of harmonic analysis of transistor currents (2) show that $\mathbf{I}_{c1}$ is a function of control voltage

$$\mathbf{U}_y = \mathbf{U}_b + j\omega_0\tau_k\mathbf{U}_c$$

where $\tau_k = r_b'C_{ka}$; r_b' - base resistor, C_{ka} - collector junction capacitance , i.e

$$\mathbf{I}_{c1} = \mathbf{I}_{c1}(\mathbf{U}_y, U_{b0}) \qquad (5)$$

Complex amplitudes $\mathbf{I}_{c1}$, $\mathbf{I}_{b1}$ in a bipolar transistor depend on the relation

$$\mathbf{I}_{b1}(\mathbf{U}_y, U_{b0}) = \frac{1 + \omega_0T_\beta}{h_{21E}}\mathbf{I}_{c1}(\mathbf{U}_y, U_{b0}) \qquad (6)$$

where $T_\beta = h_{21E}/\omega_T$, ω_T is transit frequency of the transistor.

Relation (3) can be modified by means of (5), (6), when we assume that $\mathbf{U}_b = K\mathbf{U}_c$, $\omega_0\tau_k/K << 1$, $K/h_{21E} << 1$. These assumption can find practical application. Then equation (3) can be written as

$$(pT_Q + 1)\mathbf{U}_y = R\left(1 - jK\frac{\omega_0}{\omega_T}\right)\mathbf{I}_{c1}(\mathbf{U}_y, U_{b0}) + R(\mathbf{I}_{c1n} - K\mathbf{I}_{b1n}) \qquad (7)$$

By means of equations (7) and (4) fluctuation characteristics of an oscillator in steady state, as well as in transient state, can be determined.

3. Steady state oscillations

Steady state equations can be obtained from (4), (7) for $\mathbf{I}_{c1n} = 0$, $\mathbf{I}_{b1n} = 0$ by substituting $p = j\Delta\omega$

$$(j\Delta\omega T_Q + 1)\overline{\mathbf{U}}_y = R\left(1 - jK\frac{\omega_0}{\omega_T}\right)\mathbf{I}_{c1}(\overline{\mathbf{U}}_y, \overline{U}_{b0}) \qquad (8)$$

$$\overline{U}_{b0} - E' = U_{BB} - E' - R_c I_{c0}\,(\overline{\mathbf{U}}_y, \overline{U}_{b0}) \tag{9}$$

To calculate steady state the following relation will be used

$$\overline{\mathbf{I}}_{c1}(\overline{\mathbf{U}}_y, \overline{U}_{bo}) = \frac{G_c\gamma_1(\Theta)}{1+j\omega_0/\omega_s}\overline{\mathbf{U}}_y$$

where G_c - voltage to current transfer ratio , ω_s - cut off frequency of G_c , Θ - high frequency conducting angle of collector current , $\gamma_1(\Theta)$ - decomposition coefficient of cosine current pulse.

The last relation is result of approximate harmonic current analysis $i_c(t)$ [10]. If we use it in equation (8), and compare its real and imaginary parts, we obtain

$$KR_kG_c\frac{1-K\omega_0^2/(\omega_s\omega_T)}{1+\omega_0^2/\omega_s^2} = \frac{1}{\gamma_1(\Theta)} \tag{10}$$

$$\Delta\omega T_Q = -\frac{K\omega_0/\omega_T + \omega_0/\omega_s}{1+K\omega_0^2/(\omega_T\omega_s)} \tag{11}$$

At the given angle Θ the feedback coefficient K can be obtained from the equation (10). The resonant resistor R_k is chosen from the condition of maximum amplitude of oscillations on collector of bipolar transistor. When solving (10) according to K it is necessary to fulfil the following inequality $K < \omega_T\omega_s/\omega_0^2$. At the known K the frequency correction $\Delta\omega$ is determined. After the basic parameters of steady state are calculated we can start tracking fluctuation in the oscillator.

4. Fluctuations in one-circuit oscillators without phasing

The noise currents give rise to small deviations of both amplitude and phase from the mean values

$$\mathbf{U}_y = \overline{\mathbf{U}}_y[1+m_y(t)]\exp[j\Delta\omega t + j\psi_y(t)]$$

$$\mathbf{U}_y \approx \overline{\mathbf{U}}_y[1+m_y(t)+j\psi_y(t)]\exp[j\Delta\omega t + j\varphi_y]$$

$$U_{b0} - E' = (\overline{U}_{bo} - E')[1+m_0(t)]$$

where $m_y(t)$ and $\psi_y(t)$ - amplitude and phase fluctuations of the control voltage , $m_0(t)$ - relative fluctuations of voltage $(U_{b0} - E')$.

With help of dependence linearization close to the steady state, we will obtain

$$\mathbf{I}_{c1}(\mathbf{U}_y, U_{b0}) = \mathbf{I}_{c1}(\overline{\mathbf{U}}_y, \overline{U}_{b0})[1+\sigma_{11}m_y + \sigma_{10}m_0 + j\psi_y] \tag{12}$$

$$\mathbf{I}_{c0}(\mathbf{U}_y, U_{b0}) = \mathbf{I}_{c0}(\overline{\mathbf{U}}_y, \overline{U}_{b0})[1+\sigma_{01}m_y + \sigma_{00}m_0] \tag{13}$$

where

$$\sigma_{00} = (\delta I_{c0}/\delta U_{b0})/(I_{c0}/(U_{b0}-E'))$$

$$\sigma_{01} = (\delta I_{c0}/\delta U_y)/(I_{c0}/U_y)$$

$$\sigma_{10} = (\delta I_{c1}/\delta U_{b0})/(I_{c1}/(U_{b0}-E'))$$

$$\sigma_{11} = (\delta I_{c1}/\delta U_y)/(I_{c1}/U_y)$$

are the local parameters calculated in the steady state.

Because of simplicity of notation of fluctuation equations, we will introduce the relative quantities of noise currents $\mathbf{I}_{b1n}$, $(\mathbf{I}_{c1n} - K\mathbf{I}_{b1n})$

$$\mu_{c1} = \mu_{c\|} + j\mu_{c\perp} = \frac{\mathbf{I}_{c1n} - K\mathbf{I}_{b1n}}{(1+jK\omega_0/\omega_T)\mathbf{I}_{c1}}$$

$$\mu_0 = \frac{a_c I_{c0n} + a_b I_{b0n}}{I_{c0}}$$

With help of the shown substitutions and by using of the time shift theorem we will obtain from the equations (4), (7) a system of complex fluctuation equations of the oscillator

$$(pT_Q + j\Delta\omega T_Q + 1)(m_y + j\psi_y) = A$$

$$(j\Delta\omega T_Q + 1)(\sigma_{11}m_y + \sigma_{10}m_0 + j\psi_y + \mu_{c\|} + j\mu_{c\perp}) = A$$

$$(pT_c + 1)m_0 = \lambda_c(\sigma_{01}m_y + \sigma_{00}m_0 + \mu_0)$$

where

$$\lambda_c = \frac{R_c I_{c0}}{U_{b0} - E'} = R_c G_c \gamma_0(\Theta)\Big[(-\cos\Theta_H)\sqrt{1+\omega_0/\omega_s}\,\Big]^{-1}$$

Θ_H - low frequency conducting angle [10].

We will introduce the vector $\vec{m} = (m_0, m, \psi)$ and the calculated vectors of noise currents

$$(\vec{I}_{cn})^T = (I_{c0n}\ ,\ I_{c1\|}/\sqrt{2}\ ,\ I_{c1\perp}/\sqrt{2}) \tag{14}$$

$$(\vec{I}_{bn})^T = (I_{b0n}\ ,\ I_{b1\|}/\sqrt{2}\ ,\ I_{b1\perp}/\sqrt{2}) \tag{15}$$

where I_{c0n} , I_{b0n} - the input and output noise currents,

$I_{c1\|}/\sqrt{2}$, $I_{b1\|}/\sqrt{2}$ - the components of the input and output noise current in phase with the collector current first harmonic,

$I_{c1\perp}/\sqrt{2}$, $I_{b1\perp}/\sqrt{2}$ - the quadrature part of the input and output noise with the collector current first harmonic.

Equations (14), (15) can be written in a matrix form

$$[f(p)]\vec{m} = I_{c0}^{-1}[A_c](\vec{I}_{cn} - [K]\vec{I}_{bn}) \tag{16}$$

where

$$[f(p)] = \begin{bmatrix} pT_c\lambda_c^{-1} + (\lambda_c^{-1} + \sigma_{00}) & -\sigma_{10} & 0 \\ -\sigma_{10} & pT_Q + 1 - \sigma_{11} & 0 \\ 0 & -\Delta\omega pT_Q^2 & pT_Q \end{bmatrix}$$

$$[K]=\begin{bmatrix} a_\delta/a_c & 0 & 0 \\ 0 & -K & 0 \\ 0 & 0 & -K \end{bmatrix}$$

The matrix $[A_c]$ in equation (16) is

$$[A_c]=\begin{bmatrix} a_c g_1(\Theta) & 0 & 0 \\ 0 & M\cos\varphi_s & -M\sin\varphi_s \\ 0 & M\begin{pmatrix} \sin\varphi_s - \\ \Delta\omega T_Q \cos\varphi_s \end{pmatrix} & \begin{matrix} M(\cos\varphi_s + \\ \Delta\omega T_Q \sin\varphi_s) \end{matrix} \end{bmatrix}$$

where

$$M=\sqrt{2\left[1+\Delta\omega^2 T_Q^2\right]/[1+K^2(\omega_0/\omega_T)^2]}\ .$$

To calculate the fluctuation power spectra of the $S_{m0}(\Omega)$ bias, $S_m(\Omega)$ amplitude and the $S_\psi(\Omega)$ phase it is necessary to represent them by means of matrices $[S^b(\Omega)]$, $[S^c(\Omega)]$, vectors $\vec{I}_{cn}$, $\vec{I}_{bn}$ and bipolar transistor noise current, calculated for input and output.

The determination method of these noise current spectral characteristics is explained in [14] and thus it is known.

The spectral matrix $[S(\Omega)]$ of vector $\vec{m}=(m_0, m, \psi)$ has the following representation

$$[S(\Omega)]=\begin{bmatrix} S_{m0}(\Omega) & S_{m0m}(\Omega) & S_{m0\psi}(\Omega) \\ S_{mm0}(\Omega) & S_m(\Omega) & S_{m\psi}(\Omega) \\ S_{\psi m0}(\Omega) & S_{\psi m}(\Omega) & S_\psi(\Omega) \end{bmatrix}$$

From the equation (16) we obtain the following representation of this matrix using $[S^b(\Omega)]$, $[S^c(\Omega)]$

$$[S(\Omega)]=(I_{c1}^2)^{-1}[J(j\Omega)]^{-1}\{[C]+[B]\}\{[J(-j\Omega)]^{-1}\}^T \tag{17}$$

$$[C]=[A_c][S^c(\Omega)][A_c]$$

$$[B]=[A_c][K][S^b(\Omega)][K]^T[A_c]$$

In expression (17) the mutual vector matrix is neglected. The correctness of this presumption is explained in the work [14].

5. Simplified equations of phase and amplitude noise spectrum

Besides the general relationship (17) the simplified versions are also of some interest for practice. We will bring an example where it is possible to neglect the influence of the noise current calculated for input ($\vec{I}_{bn}=0$), low frequency component of noise current calculated for output ($I_{c0n}=0$), inertia of automatic bias circuit ($T_c=0$) *and* $K \ll (\omega_T/\omega_0)^2$ from which follows

$$\Delta\omega T_Q=-\omega_0/\omega_s=-\tan\varphi_s\ ,\quad M=\sqrt{2}/\cos\varphi_s$$

Then, from the equation (16), the symbolic expression for relative amplitude fluctuations and oscillation phase can be obtained

$$m=\frac{1}{pT_Q+(1-\sigma)}\left[\frac{I_{c\parallel}}{I_{c1}}-\frac{I_{c\perp}}{I_{c1}}\tan\varphi_s\right] \tag{18}$$

$$pT_Q\,\psi=\frac{\Delta\omega p T_Q^2}{pT_Q+(1-\sigma)}\left[\frac{I_{c\parallel}}{I_{c1}}-\frac{I_{c\perp}}{I_{c1}}\tan\varphi_s\right]+\frac{1}{\cos^2\varphi_s}\frac{I_{c\perp}}{I_{c1}} \tag{19}$$

where

$$(1-\sigma)=1-\left[\sigma_{11}+\frac{\sigma_{01}\sigma_{10}}{\lambda_c^{-1}-\sigma_{00}}\right]$$

From the relations (18) and (19) we will arrive to the power spectra and we will neglect the mutual spectra

$$S_m(\Omega)=\frac{1}{\Omega^2 T_Q^2+(1-\sigma)^2}\left[\frac{S_{c\parallel}(\Omega)}{I_{c1}^2}+\frac{S_{c\perp}(\Omega)}{I_{c1}^2}\tan^2\varphi_s\right] \tag{20}$$

$$S_\psi(\Omega)=S_{\psi 1}(\Omega)+S_{\psi 2}(\Omega)+S_{\psi 3}(\Omega) \tag{21}$$

where

$$S_{\psi 1}(\Omega)=\frac{\tan^2\varphi_s}{\Omega^2 T_Q^2+(1-\sigma)^2}\frac{S_{c\parallel}(\Omega)}{I_{c1}^2}$$

$$S_{\psi 2}(\Omega)=\frac{(1-\sigma)^2}{1+\Omega^2 T_Q^2(1-\sigma)^2}\frac{S_{c\perp}(\Omega)}{I_{c1}^2}$$

$$S_{\psi 3}(\Omega)=\frac{1}{\Omega^2 T_Q^2}\frac{(1+\tan^2\varphi_s)^2}{1+\Omega^2 T_Q^2(1-\sigma)^2}\frac{S_{c\perp}(\Omega)}{I_{c1}^2}$$

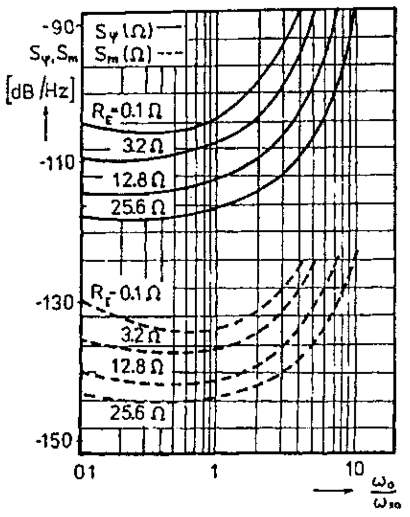

Fig.3
Amplitude and phase noise of the oscillator with the negative feedback

The equations (20) and (21) can be used for calculations of fluctuations if the shown above presuppositions are fulfilled.

The calculations according to (17) were carried out, in order to see how the working frequency, current feedback parameters and bias circuits influence the fluctuation characteristics of the oscillator with bipolar transistor close the carrier ($\Omega = 0.01/T_Q$) is small if compared to the resonance circuit bandwidth). Bipolar transistor was characterized by the parameters $h_{21E} = 70$, $\tau_k = 150ps$, $C_k = 1.5pF$, $I_{c\max} = 20mA$, $f_T = 0.8\,GHz$. The results are shown in Fig.3. In Fig.3 is shown the influence of amplitude and phase power spectrum fluctuations on scaled oscillation frequency ω_0/ω_{s0} for several values of R_E in current feedback. By increasing the phase fluctuations also increase which is caused mainly by decrease phase characteristic of feedback circuit and by increase of frequency correction.

As follows from Fig.3, increasing of resistance causes decrease of $S_\psi(\Omega)$ in the frequency bandwidth ω_0/ω_{s0} proportionally to the calculated noise current on output bipolar transistor. We should realize that with increasing of R_E , there is a decrease of power gain of bipolar transistor and the losses caused by the base circuit at the given form of collector current impulse also increase.

$S_m(\Omega)$, $S_\psi(\Omega)$ dependencies are minimal because increase of these spectra values in sphere of high values ω_0/ω_{s0} is caused by transformation of frequency fluctuations into the amplitude ones. This contribution of transformation starts to dominate over the influence of the same phase component of the noise current $I_{c||}$ calculated for output. The influence of R_E on $S_m(\Omega)$ is similar and it has the same causes.

6. References

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