Novel LNAs with Transformerless Positive Feedback

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Abstract - The prominent inventions of D.E. Norton [1,2] and also other similar schematics till now have a wide application in different circuits of amplifiers and mixers. However the presence of the broadband feedback transformer in these circuits still hampers their application in high-frequency integrated circuits (IC). We propose new configuration schematics of the transformer-less amplifiers with similar properties and admittance positive feedback (APF). Just now we understand that implicitly this opportunity was already contained in our previous patents and papers [3-8]. Here it is considered in detail the circuit of low noise amplifier (LNA) with a resistive positive feedback (RPF). RPF and CB amplifiers are compared. It is shown, that LNA RPF provides more than twofold lower noise temperature, good level of input and output matching, greater frequency band and high stability gain, despite of a positive feedback.

Key-Words: - noiseless amplifier, broadband amplifier, positive feedback, transformer-less feedback amplifier.

1 Introduction

Norton [1,2] for the first time have proposed the circuits of matched wideband amplifiers with a transformer negative feedback ideal by their simplicity, quality and application perspective. The successful combination of broadband transformers with distributed parameters and modern transistors has provided in these circuits an extreme sensitivity and linearity. The transfer function of such amplifier usually coincided with a voltage transfer function of the feedback transformer. Based on the similar principles, rather original circuits of amplifiers and mixers [3-12] were described and implemented. Abranin and Bruck [3,4] proposed amplifiers with the autotransformer feedback (AATF thereinafter). They differed in principle by implementing of both positive and negative feedback and by separation of amplification and input/output impedance matching functions. The analysis of these circuits [5] and their practical design [6,7] provided expanding by order the bandwidth in comparison with the transformer circuits and to design new distributed amplification system of the radio telescope UTR-2 [8]. Doubtlessly, the high quality of all similar circuits depends on the parameters of the transformers.

However for a long time we have met with a serious problem. Increase in frequency demands reduction of the size of transformers. It leads to technological difficulties and deterioration of their qualities and limits a scope especially in comparison with fast progress in technology of HF transistors. These factors hinder their wide usage in the high-frequency integrated circuits (IC). So the investigations of new circuits of LNA amplifiers without transformer in a feedback loop are important and especially perspective for IC technology.

2 Versions of schematics. General features.

The purpose of the present paper is to investigate the critical phenomenological features of the new circuits instead of detail comparison of theory and experiment. Therefore we will use the simplest equivalent schematics of transistors and also we will exclude from the circuit the elements, which are relevant for its normal operation, but accidental for analysis its basic features.

We propose and describe shortly two main versions of the APF schematics. They shown in Fig.1a,b.

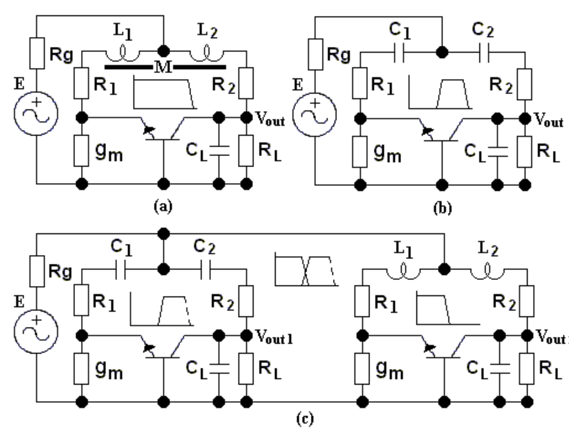
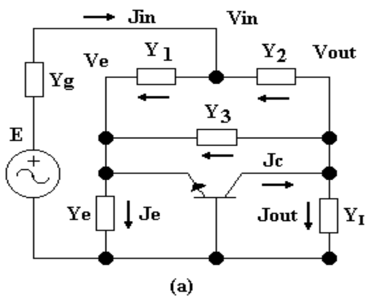


Fig.1 (a) RLPF; (b) RCPF; (c) DPF.

The first one is amplifier with resistive-inductive positive feedback (RLPF). At $M \cong \sqrt{L_1 L_2}$ RLPF converted into AATF [3,4], i.e. amplifier with negative current feedback. At $M < \sqrt{L_1 L_2}$ this amplifier is suitable for applying in IC technology. The second one is amplifier with resistive-capacitive positive feedback (RCPF) and it is the most interesting not only in IC. This RCPF amplifier provides high bandwidth and minimal noise level. RLPF and RCPF amplifiers converted into LPF and CPF noiseless ones at R_1, $R_2 \to 0$. At L_1, $L_2 \to 0$ and C_1, $C_2 \to \infty$ they converted into RPF amplifier. And at last Fig.1c demonstrates all matched active diplexer combined by RLPF and RCPF amplifiers.

The equivalent generalized circuit of the amplifiers proposed is shown in fig.2a.



$$Y \cdot \begin{pmatrix} n+1 & -1 \\ n \cdot a+1 & -1 \end{pmatrix}$$

Fig.2 (a) APF schematic; (b) its |Y| matrix.

It contains all essential elements of the real circuit: the bipolar transistor in a CB configuration, three elements of the feedback loop Y_1, Y_2, Y_3 and load Y_L. Note, that hereinafter all the labels such as *'Y'* or *'Z'* refer to the same admittance, for example, $Z_L = Y_L^{-1}$ and are applied for simplification in the subsequent text. For real values the labels R and R^{-1} are used. The circuit is convenient to insert in it any admittance of the 'π' equivalent circuit of transistor.

The directions of the current indicated in the circuit correspond to the direction of the generator current. According to directions of the current trough impedances Z_1 and Z_2 we will refer this circuit to the type of the circuits with a positive feedback. From this point of view in the AATF amplifier [3,5,6] negative feedback operates in the bandwidth of the autotransformer (current through Y_2 is guided in the opposite direction) and positive feedback – at the lower frequencies.

When the positive feedback is considered there are several suspicions about instability, sensitivity to change of parameters etc. Here we will examine some of them. It is worth notification, that implementation of the positive feedback in amplifiers is permanently expanded and becomes rather popular [13-17].

The equations set for unknown voltages *Vin, Vout, Ve* of the circuit is given by

$$Y_1 \cdot (Vin - Ve) + Y_3 \cdot (Vout - Ve) - Je = 0$$

$$Y_g \cdot (E - Vin) - Y_1 \cdot (Vin - Ve) + Y_2 \cdot (Vout - Vin) = 0 \qquad (1)$$

$$Jc - Y_L \cdot Vout - Y_2 \cdot (Vout - Vin) - Y_3 \cdot (Vout - Ve) = 0$$

$Je = Ye \cdot Ve$; $Jc = \alpha \cdot Je$; Ye-emitter admittance; α-current gain.

At first we explore the main features of the amplifier with the ideal transistor. It will give us an opportunity to demonstrate the basic features of the similar circuits.

We introduce the notations: $Z_2 = Z$, $Z_1 + Ze = Z/n$. Then the solution of the system (1) for $\alpha = 1$ gives the following four basic response functions of the input signal.

Response function of the input current *Gj*.

$$Gj = Jout / Jin = 1 . \qquad (2)$$

Voltage transfer function of the amplifier itself *Gv*.

$$Gv = Vout / Vin = (1+n) Z_L (Z_L + Z)^{-1} . \qquad (3)$$

Voltage transfer function of the input network *Gin*.

$$Gin = \frac{Vin}{0.5E} = \frac{2(Z_L + Z)}{Z_L + Z + (1+n) \cdot Z_g} . \qquad (4)$$

Here and hereinafter *Gin* is normalized by 0.5, which is transfer ratio in the impedance matching mode. On default the matching means an equality of complete but not conjugate impedances. For the RPF they coincide.

Since $Gj = 1$ (Eq.2), the function of the voltage amplifier gain *G* is

$$G = Gin \cdot Gv = \frac{2(1+n) \cdot Z_L}{Z_L + Z + (1+n) \cdot Z_g} . \qquad (5)$$

Therefore, the proposed circuit is a voltage amplifier. Its intrinsic transfer ratio (Eq.3) at $Z << Z_L$ is determined by division coefficient of the feedback loop and weakly depends on the load. The input network (Eq.4) behaves as a divider formed by impedance Z_g and $(Z_L + Z)/(1+n)$. If they are equal, the amplifier input is matched. This important feature requires more detail study. The solution of the system (1) gives input *Zin* and output *Zout* resistances of the amplifier.

$$Zin = (Z_L + Z) \cdot (1+n)^{-1} \qquad (6)$$

$$Zout = Z_g \cdot (1+n) + Z \qquad (7)$$

According to Eqs. 6 and 7 and also according to the shape of the [Y] matrix of the idealized amplifier circuit (Fig. 1b), it represents some kind of a voltage transformer with an internal resistance *Z* (output) or $Z/(1+n)$ (input) and transformation ratio of external impedances $1+n$. It is easy to provide the input ($Zin = Zg$) or output ($Zout = Z_L$) amplifier matching. However simultaneously - both on an input and on an output it is possible only at $Z = 0$. Indeed, from Eqs. 6 and 7

$$A \cdot B = (1 + Z/Z_L) \cdot [1 + Z/Z_g/(1+n)], \qquad (8)$$

where $A = Zin / Z_g$ and $B = Zout / Z_L$ is VSWRs of the input and output of the amplifier. Apparently, $A \cdot B > 1$ at $Z > 0$ and Eq.8 determines the criteria for the choice of parameters. Thus the better amplifier is matched, the less is the resistance of the feedback Z.

What are the criteria for the selection of Z? In this connection we will consider transfer function of the emitter current Q in a positive feedback loop. The solution of Eqs.1 gives

$$Q = Je / Jin = (1 - Le)^{-1} = n(1+n)^{-1}(1 + Z_L Y) + Y_3 Z_L \quad (9)$$

Where $Le = 1 - Q^{-1}$ - transfer function of the feedback loop. This loop is placed inside of the amplifier and it might seem that it does not influence on amplifier's properties. However it is not the case in general. Reduction of the resistance Z up to minimum value $n \cdot Ze$, which is useful from a point of view of reduction of the mismatch and amplifier noises, simultaneously leads to growth of the signal current in the emitter, i.e. $Q >> 1$. And it is accompanied by increasing of nonlinear distortions. Due to $\alpha = 1$ the admittance Y_3 does not enter in any of the equations Eqs.2-8 at all. This very important feature was used for example in the AATF circuit [3-8] for essential expansion of the bandwidth. Here we suppose $Y_3 = 0$.

And at last we can find the relationship between the parameters of amplification G and Q and the parameters of matching A and B. From Eqs.5-7 we have

$$B = [G(Q - A) - 2A] \cdot [GA(Q-1)]^{-1}. \quad (10)$$

We intend prove that all proposed amplifiers are LNA. Therefore in the next section we investigate the noise properties of the RPF amplifier.

3 Noises

An RPF amplifier by itself contains two source of noise: a resistive feedback loop and a transistor. The output noise includes also noises of a signal generator and a load resistors R_g and R_L. We will calculate and compare noise temperatures of RPF and CB circuits disregarding frequency dependences of its elements. We incorporate in series with any resistor R its EMF of the noise with the spectral density $E_R = \sqrt{4kT_0R} = \sqrt{sR}$ at $T_0 = 300^0$. From the system of equations (1) we will obtain the transfer ratio K_R from E_R to the load R_L and spectral density of the noise power $W_R = K_R{}^2 E_R{}^2 = sRK_R{}^2$. As usual, we will be interested in the relative spectral density $H_R = W_R / W_{Rg}$ and the temperature $T_R = T_0 \cdot H_R$, where $W_{Rg} = 0.25sG^2R_g$. For each resistor of the RPF circuit it gives the function of parameters G, Q, n.

The noise of the feedback resistors R_1 and R_2 are:

$$H_{R_1} = n^2 G(n+1)^{-2}[2n(Q-1) - Q(G-2)]^{-1}, \quad (11)$$

$$H_{R_2} = nG(n+1)^{-2}[2n(Q-1) - Q(G-2)]^{-1}. \quad (12)$$

From Eqs.11,12 we obtain the conclusion: $H_{R_1} = nH_{R_2}$ when $E_{R_1} = n^{-0.5}E_{R_2}$. This is due to difference between transfer ratios to the load, since $K_{R_1} = nK_{R_2}$.

The noise of the load resistor R_L is:

$$H_{R_L} = \frac{[n(Q-1)+Q] \cdot [2(n+1) - G]^2}{G(n+1)^2[2n(Q-1) - Q(G-2)]}. \quad (13)$$

Similarly, for comparison, the noise of the elementary CB circuit matched by the input is

$$H_{CB} = 1 + 4/G. \quad (14)$$

We analyse the obtained formulae. Since $n \propto G$ it's more convenient to represent these expressions as functions of independent variable n/G and parameters G and Q. In Fig.3a the output noise of the feedback divider for four pairs of values G (4 and 16) and Q (3 and 4) is shown.

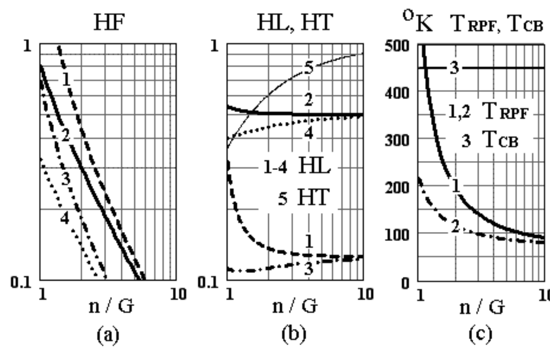


Fig.3 Families of the relative noise functions: (a) HF-resistive feedback noise, (b) HL-noise of the output load, HT-transistor noise, (c) total RPF and CB noises.

One can see from Fig.3a that the dependence is strong: the noise sharply decreases approximately as $\propto (n/G)^{-1}$ and $\propto Q^{-2}$ but weakly increases as $\propto \sqrt{G}$. In Fig.3b the output resistance noise of the load $HL = H_{R_L}$ for the same values n/G, G and Q is shown. Here the pattern is different. $HL(n/G,G,Q) \propto G^{-1}$ and almost does not depend on n/G and Q. The dependence HT is also shown here. It gives estimating roughly the contribution of the noise current of the transistor of the RPF circuit relative to the CB circuit. The calculation is made by the formula.

$$HT = [1 - 0.5G(n+1)^{-1}]^2. \quad (15)$$

In Fig.3c the noise temperature of the passive part RPF of the circuit $T_{RPF} = T_F + T_L$ and for comparison and estimation of benefits the noise temperature of the CB circuit T_{CB} - are given for the same values of G and Q. It is obvious that the noise is reduced by 2÷3 times.

The dependences lead to the important conclusions.

- The LNA noise in a RPF configuration is reduced more, than twice.
- For given G the reduction is reached by the choice of n and Q within the limits: Q = 2÷3 and $n = (1.5÷3)G$.

4 Amplifier with real transistor

We turn now to examination of changes of the main performances of the amplifier when the real transistor is used. We will estimate its bandwidth and influence of a limited gain taking into account only emitter and collector junctions. We suppose: Y_3=0, $Z = R$, $Z_L = R_L$, $Zg = Rg$ and, according to Eqs.6,7 $Rout = (1+n)Rg + R$, $Rin = (RL + R)(1+n)^{-1}$. We use the simplest frequency dependences of the parameters of the transistor and load of the amplifier in the form of:

$Ye(\omega) = g_m(1 + i\omega\tau_e)$ - admittance of the emitter; $YL(\omega) = RL^{-1}\cdot(1 + i\omega\tau_c)$ - admittance of the output; τ_e and τ_c – time constants of the emitter and collector junctions; ω - circular frequency. We have included admittance of the base-collector junction in the load.

$\alpha(\alpha_o, \omega) = \alpha_o / (1 + i\omega\tau_e)$; α_0 - static current gain coefficient; g_m - conductance of the emitter. Analogously to Eq.5, the substitution of these parameters in Eqs.1 gives

$$G(\alpha_o,\omega) = \frac{2[1+\alpha(\alpha_o,\omega)\cdot n]}{1+Rout\cdot YL(\omega)+[1-\alpha(\alpha_o,\omega)]\cdot n\cdot Y\cdot Rg}. \quad (16)$$

From Eqs.16 it follows, that the influence of the base-emitter junction on amplification and bandwidth of RPF is increased by Q times in comparison with a conventional CB circuit. Thereby at any $\alpha_0 < 1$ and $\omega \to \infty$, $G(\alpha_0,\omega) \to 0$, and the greater value of Q the more quickly $G(\alpha_0,\omega)$ tends to zero. At the same time due to the matching the influence of both junctions is reduced. Since RPF is the voltage amplifier, usually $\tau_c > \tau_e$ and for not too high values of Q the bandwidth of RPF can increase (for $\tau_e << \tau_c$ and $G \approx n + 1$ up to two times!). One needs to note that the described properties of the RPF circuit valid for all the circuits have generalized character. The actual characteristics depend both on the circuit parameters and type of the transistor and construction of the device.

It is confirmed by the results of the Multisim 7 simulation of the RPF and the CB circuits with the transistor MPSH10 (Jo = 10mA, $g_m = 2.6^{-1}$ S) and with elements: R_g = 50Ω, R_{CB} = 46Ω, R_L = 200Ω, R_1 = 21Ω, R_2 = 155Ω shown in Fig.4. It is clear, that $\omega_{RPF} = 1{,}4\omega_{CB}$.

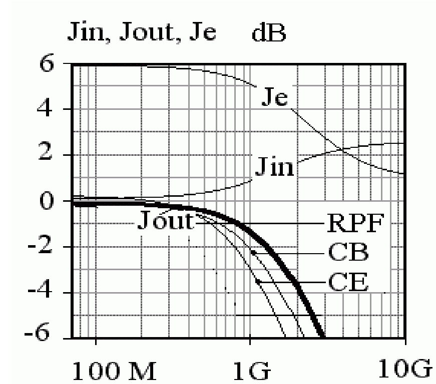


Fig.4 Comparison relative gain and cuurrents frequenciy responses for different circuit configurations.

To complete the picture the frequency characteristic of the reference CE stage matched by the input with the parallel negative feedback is given. Apparently the RPF amplifier has the maximum bandwidth. To confirm $Q \cong 2$ we show also in Fig.4 Jin, $Jout$ and Je currents.

We describe shortly optional matching procedures. Without going into details of this well-known problem we show in Fig.5 two principally different elementary circuits of the RPF matching.

Variant (a) – is a simple 'T'-shape low-pass filter. It improves the matching of the input ((c) curves 1,2) and output and gives increasing the frequency range ((d) curve 1).

Variant (b) – an all-pass filter of the first order, with $L = C_{in}R_{in}^2$, $R = R_{in}$ - gives the constant active input resistance R_{in} in the unlimited frequency band ((c) curves 3,4), but maintains the transfer function of the first order ((d) curve 2) of the mismatched circuit. Fig.5d illustrates the efficiency of the matching.

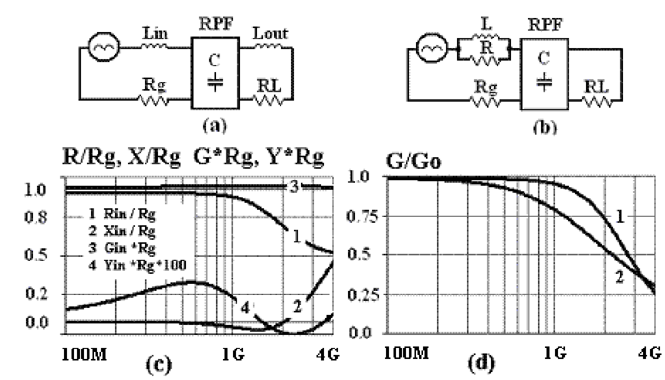


Fig.5. Matching: (a) first order "T"-shape filter, (b) all-pass filter, (c) results of the impedance matching, (d) relative gain frequency response.

5. Design.

Given analysis provides an opportunity to find all elements of the RPF circuit taking onto account any initial requirements. However the choice of parameters is not unambiguous and depends on many factors. We will give the optimum variant of a procedure of calculation.

Assuming known values of the parameters of the transistor, *Jo* and g_m and with given values of amplification *G, Q* and tolerance region of values *A* and *B* we will express all circuit elements by the resistance of the generator R_g and parameters of the amplifier. It gives

$$R_1 = R_g A Q^{-1} - g_m^{-1}, \tag{17}$$

$$R_2 = nR_g A Q^{-1}, \tag{18}$$

$$R_L = 0.5R_g G(1+A)\left[1 - Q(1-\alpha_0)\right], \text{ where} \tag{19}$$

$$A = 0.5GQ\left[n(Q-1) - Q(G-2)\right]^{-1}, \tag{20}$$

$$B = \left[2(n+1)G^{-1} - 1\right]\cdot\left[1 - Q(1-\alpha_0)\right]\cdot\left[1 - QA^{-1}(1-\alpha_0)\right]^{-1} \tag{21}$$

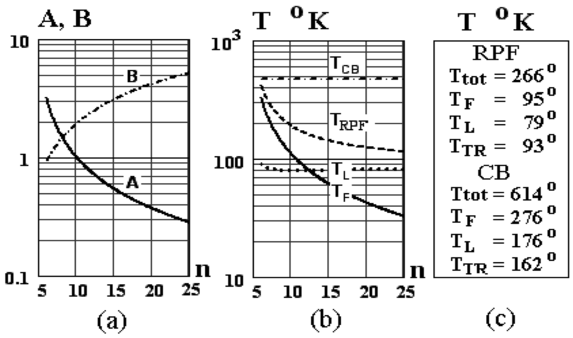


Fig.6 Optimal design process of the RPF amplifier: (a) and (b) choice of *n* parameter, (c) results of simulation.

One needs to choose the value of *n*, which provides reasonable values *A* and *B* (Eqs.20,21) and the noise temperature (Eqs.11-14). Most often one chooses $A \approx 1$ or $A \approx B$. The typical dependences of the parameters under consideration on *n* shown in Fig.6a,b.

Then by Eqs.17-19 we compute the values of resistors R_1, R_2, R_L and check up the bandwidth, linearity and noise of the amplifier.

6 Simulation

The carried out investigations are confirmed by the complete coincidence of results of calculation and simulation of a RPF circuit, resulted in Fig.6c and in the Table 1 (lines 2,5-calculation, lines 3,6-simulation). The simulation gives also the transistor noise T_{TR} and bandwidth: $Fmax_{RPF} = 1.0$GHz, $Fmax_{CB} = 0.70$GHz.

Table 1 Data of the first circuit RPF and CB amplifiers.

G	Q	Jo	n	A	B	Rg	Rin
7	2.5	10	10	1	1.94	50	50
6.9	2.5	10	10	0.94	1.89	50	47
R1	R2	RL	Rout	TF	TL	TRPF	TCB
17	200	368	714	94	76	**170**	**471**
17	200	370	699	95	79	**174**	**452**

This coincidence proves correctness of all relations and conclusions and also the appropriateness of accepted considerable simplifications, concerned with the model of transistor. The main conclusion concerned with the noise parameters. Since, $T_{CB} = 471^o$ the gain is $T_{CB}/T_{RPF} \approx 2.7$. And according to Eq.15 and Fig.6c, the noise of transistor (93° in RPF and 162° in CB) is also reduced by ≈ 1.7 times. Therefore LNA RPF is 2.3 times less "noisy". It's easy to estimate, that the bandwidth is increased by ≈ 1.4 times as compared to the CB circuit (see also Fig.4). We note, by the way, that linearity and noise of the CE circuit with a negative parallel feedback obtained by the simulation are worse, than those of CB.

Here we could complete our theoretical research, but we would like to do an additional test. We will carry out a simulation by professional simulator Spectra Cadence, with changed basic data and elements of the RPF and CB circuits, but with the same virtual origin of components. We choose a transistor with maximal frequency 30-40 GHz, utilized in IC technology.

Main circuit data and simulation results are given in Table 2 (lines 2,5-calculation, lines 3,6-simulation) and in Fig.7. Simulation gives also transistor noise and harmonics level.

Table 2 Data of the second circuit RPF and CB amplifiers.

G	Q	Jo	n	A	B	Rg	R1
4	2	10	6	1	2.4	50	21
3.9	1.95	10	6	0.98	2.4	50	22
R2	RL	TF	TL	TTR	TRPF	TTR	TCB
151	204	112	150	----	270	----	600
150	200	----	----	**100**	**370**	**160**	**760**

The first conclusion is concerned with the possibility and availability of LNA RPF with the bandwidth more than 10 GHz. Second – also important: the amplifier is stable and the values and results of simulation of the amplifier, active part of input and output conductivity, noise temperature coincide to a high degree of accuracy. Moreover, though the maximum bandwidth is increased by an order we observe surprising coincidence of view of all frequency dependences. Although the bandwidth is $B_{RPF} \le B_{CB}$. Before it was $B_{RPF} > B_{CB}$. The important feature of the present simulation is the possibility to define and compare nonlinear characteristics of the LNA RPF and CB. We do not know the limits of validity of the nonlinear model of transistor. Therefore the unique and careful enough conclusion which we are able to do is that the level of harmonics of RPF only by 4 dB grater than those of CB. And it is in spite of the fact that the signal current through the transistor RPF is increased by *Q* times (*Q* = 2). The examination of this difficult problem exceeds the frames of our short communication.

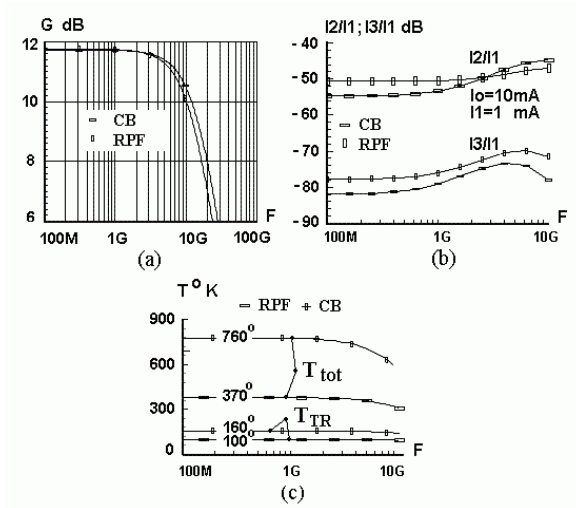


Fig.7 Simulated data of the RPF and CB amplifiers: (a) gain, (b) relative 2d and 3d harmonics, (c) noise.

We do not know the limits of validity of the nonlinear model of transistor. Therefore the unique and careful enough conclusion which we are able to do is that the level of harmonics of RPF only by 4 dB grater than those of CB. And it is in spite of the fact that the signal current through the transistor RPF is increased by Q times (Q = 2). The examination of this difficult problem exceeds the frames of our short communication.

7 Conclusions

So, we finished the investigation of a rather prospective transformerless amplifier circuit, with the parameters close to the one of the best circuits with a transformer negative feedback. It is shown, that LNA RPF provides more than twofold lower noise temperature, good level of input and output matching, greater frequency band. More details will be published elsewhere.

Within the accepted framework of simplifications of the transistor circuit the response functions both RPF and CB are the first order ones and consequently are absolutely stable. The simulation also confirms that. The practical implementation will show whether it is so.

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