Research Article

High-Frequency Sections of Active Filters of Mixed-Signal SoC Based on Current Amplifiers

Sergei Krutchinsky and Nikolai Prokopenko

South Russia State University of Economics and Service, Russia

Correspondence should be addressed to Nikolai Prokopenko, prokopenko@sssu.ru

The sufficient conditions for the efficient use of active elements are formulated by analyzing the basic structure of second-order unit. The expediency of current amplifiers usage in HF and SHF filters is shown. The examples of a methodical nature are given and conclusions of application importance are formulated.

1. Introduction

A modern mixed-signal systems on a chip (SoC) creating for communications and radio-technical systems are inextricably linked to the development of analog and analog-digital IP-blocks for a sufficiently accurate processing of HF and SHF range signals. Some of the basic components of these blocks are frequency filters, which are highly sensitive parametrically. As a rule, reduction of the parametric sensitivity causes the need to create fairly complex active elements, which consume much power. Exactly energy problems of precision filters reduce the attainable level of integration of the SoC and the area of practical usage. Typically, the frequency filter is designed for a specific task with given parametric and structural constraints that are defined by a specific technology. In HF and SHF ranges transimpedance amplifiers (voltage-current converters) are often used as the active elements for compensation of losses in the passive RC-circuits [1, 2]. However, the potentially achievable unity-gain frequencies of these elements give way to current amplifiers, thus ultimately limiting the product of the Q factor (Q) and the pole frequency (f_p) with the existing restrictions on power consumption. That is why the potential properties of the filter based on current amplifiers have a practical perspective and require further theoretical research. At least, the limiting theoretical estimations are not yet available and practical recommendations are not formulated.

2. Problem Statement

As a rule, to increase the stability of the frequency characteristics of high-order filters, additional intersection connections are used to ensure the minimization of sensitivity to the instability of the passive RC elements in the range of operating frequencies [3, 4]. In HF and SHF bands, this schematic approach is almost never used because of the high and usually not accessible requirements to a parasitic phase shifts of used active elements. In this regard, the cascading of the second-order sections remains the main direction of the implementation of the necessary poles and zeros of the desired transfer function [5, 6]. That is why improving of their parameters stability and reduction (minimization) of appropriate parametric sensitivities is the main way of the general problem solution. In addition, achievable through the implementation in modern technologies the precision of resistors and capacitors in frequency-defining elements is not sufficient to ensure admissible errors of transfer function poles of the main frequency filter types.

According to-above-mentioned things, solutions are necessary for the whole class of interrelated circuitry tasks which are aimed on the accuracy improving of the basic parameters realization in the second-order sections of HF and SHF ranges.

Firstly, taking into account the dominating influence of the pole frequency implementation error (δf_p) on the

capacitive characteristics of high-Q sections, it is necessary to create structures with a minimum parametric sensitivity to the instability of active elements and components, and first of all, current amplifiers [7]. Secondly, from the point of view of frequency-band extension, similar structures with maximally effective usage of these amplifiers resources (component and energy costs for the product of Q-factor and pole frequency) are needed [8]. At the same time possible and acceptable trade-off is the equality of contributions of attenuation instability ($\delta d_p = -\delta Q$) and pole frequency instability $-\delta f_p Q \approx \delta Q$. In terms of general problem solving it creates sufficient preconditions for energy consumption reduction or provides the structural degrees of freedom to increase the operating frequency range. Finally, and this is the most important from a technological point of view, the existing structural and hence the parametric degrees of freedom can be used for correction (adjustment) of the pole parameters (f_p, Q) by mode-defining and therefore low-frequency circuits. Specified property of such circuit solutions permits not only to integrate IP-blocks and SoC, but also to implement during their exploitation a temperature and temporal correction of plural transformation tracts through programmable cores of the system and simple digital-to-analog converters.

3. The Main Properties of the Basic Structure of the Second-Order Section

In general case the structure (Figure 1) consists of three autonomous units—the active element K, the frequency-dependent $\gamma(p)$, and frequency-independent β circuits.

That is why its transfer function

$$F(p) = \frac{K(p)[\gamma_1(p) - \beta_1]}{1 - K(p)[\gamma_2(p) - \beta_2]}, \quad \left(K(p) = \frac{K}{1 + p\tau_a} \right) \tag{1}$$

allows to realize any type of filter, amplitude, or phase corrector. In general case for the second-order section

$$\gamma_1(p) = \frac{N(p)}{p^2 + pD_p\omega_{pc} + \omega_{pc}^2}, \tag{2}$$

$$\gamma_2(p) = \gamma_0 \frac{a_2 p^2 + a_1 pD_0\omega_{pc} + a_0\omega_{pc}^2}{p^2 + pD_p\omega_{pc} + \omega_{pc}^2}, \tag{3}$$

where D_0, D_p are the attenuation of zero and the pole of frequency-dependent circuit; ω_{pc} is the frequency of its pole; γ_n is the scale transfer coefficient; a_i is binary (0, 1) coefficients; $N(p)$ is the polynomial that defines the type of implemented second-order section.

From (1)–(3) follows that for function in general form

$$F(p) = \frac{A(p)}{p^2 + pd_p\omega_p + \omega_p^2} \tag{4}$$

frequency ω_p and pole attenuation d_p for idealized schema ($\tau_a = 0$)

$$\omega_{(p)} = \omega_{pc}\sqrt{\frac{1 + \beta_2 K - K\gamma_0 a_0}{1 + \beta_2 K - K\gamma_0 a_2}}, \tag{5}$$

$$d_p = \frac{D_p(1 + \beta_2 K) - K\gamma_0 D_0 a_1}{\sqrt{(1 + K\beta_2 - K\gamma_0 a_2)(1 + K\beta_2 - K\gamma_0 a_0)}}. \tag{6}$$

Ratios (5) and (6) define structure of parameters influence of passive frequency-dependent (D_0, D_p, ω_{pc}, γ_0), frequency-independent (β_2) circuits, and active (K) elements and allow to receive the following practical conclusions. First, for exception of the most unstable parameter influence of the scheme K on frequency of a pole ω_p, it is necessary to use symmetric frequency-defining circuits ($a_2 = a_0$) when $\omega_p = \omega_{pc}$ and active element influence is directed on increase of Q-factor $Q = 1/d_p$. Secondly, in case of use of noninverting amplifier or the converter ($K > 0$) value d_p (6) is minimized at $a_0 = a_2 = 0$, $a_1 = 1$

$$d_p = D_p - \gamma_0 D_0 \frac{K}{1 + K\beta_2}, \tag{7}$$

and at application of an inverting active element ($K < 0$) this minimization is possible when $a_0 = a_1 = a_2 = 1$, $\beta_2 = 0$:

$$d_p = D_0 \frac{K\gamma_0}{1 + K\gamma_0} + \frac{D_p}{1 + K\gamma_0}. \tag{8}$$

Thus, in the first case ($K > 0$) in feedback loop of an active element (the amplifier and the converter) K it is necessary to use the frequency-defining circuits of band-pass type, and in the second case ($K < 0$)—rejection type. In this case frequency of a pole of the scheme will be always defined by frequency of passive circuit pole, and action of active element will be directed on increase of realized Q-factor.

For passive RC circuits of the second order [5] of band-pass type the condition $D_0\gamma_0 = D_p - D_0$ is characteristic, so at $\beta_2 = 0$

$$d_p = D_p(1 - K) + D_0 K. \tag{9}$$

Therefore, by the choice of passive elements ratios of these circuits (e.g., T-bridges) and structure of the amplifier K it is possible to provide realization of ω_p, and d_p with low parametrical sensitivity. Apparently from (8), realization of this condition for inverting amplifiers is connected with limiting ($K \rightarrow \infty$) increase of its gain.

As it is seen from (1), (2), the execution of the formulated conditions demands from passive frequency-defining circuits the additional condition realization of reproduction of the desired frequency characteristics in a circuit of direct transfer ($\gamma_1(p)$).

If to consider the influence of time constant of noninverting active elements (τ_a) on parameters of transfer function poles of basic structure then under condition of their low admissible relative changes ($\delta\omega_p, \delta d_p$) we receive

$$\delta\omega_p \approx \delta_{d_p} \approx -\frac{1}{2}\frac{\omega_p}{B_S}D_p\frac{K}{1 + \beta_2 K}, \tag{10}$$
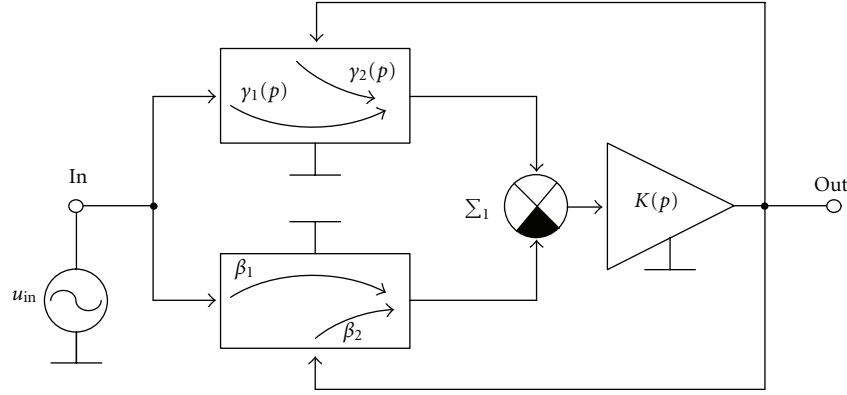
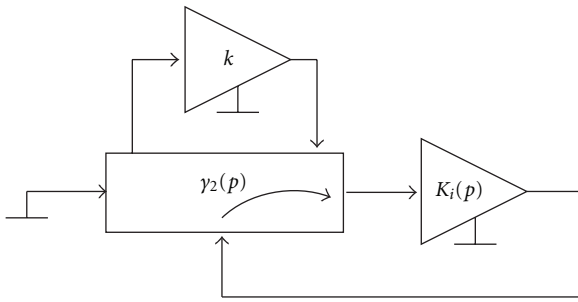FIGURE 1: Basic structure of second-order section.

FIGURE 2: The feature of formation of SHF section poles.

where $B_S = K/\tau_a$ is the gain area of active element K of basic structure.

In this case parametrical sensitivities

$$S_{B_S}^{\omega_p} = S_{B_S}^{d_p} = \frac{1}{2}\frac{\omega_p}{B_S}D_p\frac{K}{1 + \beta_2 K} \qquad (11)$$

are defined by numerical values of B_S and pole attenuation D_p of a passive chain and can not be minimized by frequency-independent feedback depth ($0 \leq \beta \leq 1$).

From the specified positions the most important qualitative parameter of passive frequency-defining circuits is their attenuation D_p which reduction favorably effects on a working frequency range and realized Q-factor. For RC circuits of the second order [5] $D_{p\,min} = 3$, thus $D_0 = 2$, and under condition of D_0 minimization (see (13)) $D_p = 2Q$ and, therefore, the influence of B_S on poles parameters increases. For reduction of D_p it is possible to insert into RC circuit the additional repeater k (Figure 2) that allows, in particular, to realize in schemes the additional useful structural and parametrical degrees of freedom.

4. Low-Sensitive SHF Sections of Band Type

The created above conclusions show that for active elements influence reduction on the frequency of high-Q pole it is expedient to use current repeaters (K_i) in which feedback loop the frequency-dependent circuit of band-pass type with the additional voltage buffer repeater k separating this circuit into elementary differentiating (integrating) circuits
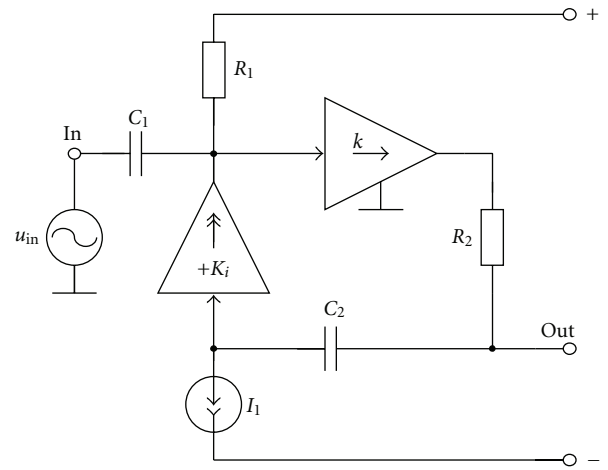

FIGURE 3: The structure of low-sensitive section of the band type filter.

is included. In this case for increasing of B_S and, therefore, according to ratios (10) and (11) ($\beta_2 = 0$), the reduction of sensitivity and parameters realization errors, the circuitry degrees of freedom are created or, at least, saved, and k usage allows to reduce attenuation of a passive circuit pole D_p. The structure of such section is shown on Figure 3.

Using a standard notation

$$F(p) = K_0 \frac{pd_p\omega_p}{p^2 + pd_p\omega_p + \omega_p^2}, \qquad (12)$$

it is easy to determine that

$$\omega_p = \frac{1}{\sqrt{\tau_1\tau_2}}; \quad (\tau_1 = R_1 C_1, \tau_2 = C_2(R_2 + r_i + r_u)), \quad (13)$$

$$d_p = \sqrt{\frac{\tau_1}{\tau_2}} + \sqrt{\frac{\tau_2}{\tau_1}}\left(1 - K_i\frac{R_1}{R_2 + r_i + r_u}\right), \qquad (14)$$

$$K_0 = Q\sqrt{\frac{\tau_1}{\tau_2}}, \qquad (15)$$

where r_i, r_u are input resistance of current amplifier and output resistance of the buffer voltage amplifier ($k = 1$).

Second-order circuit separation by means of the buffer k allows to get

$$D_0 = \sqrt{\frac{\tau_1}{\tau_2}}, \qquad D_p = \sqrt{\frac{\tau_2}{\tau_1}} + \sqrt{\frac{\tau_1}{\tau_2}}, \qquad (16)$$

therefore parametrical sensitivities of the scheme can be minimized.

From (13) follows that always

$$S_{R_1}^{\omega_p} = S_{C_1}^{\omega_p} = S_{C_2}^{\omega_p} = -\frac{1}{2}, \qquad S_{R_2}^{\omega_p} = -\frac{1}{2}\frac{R_2}{R_2 + r_i + r_u},$$

$$S_{r_i}^{\omega_p} = -\frac{1}{2}\frac{r_i}{R_2 + r_i + r_u}, \qquad S_{r_u}^{\omega_p} = -\frac{1}{2}\frac{r_u}{R_2 + r_i + r_u}. \qquad (17)$$

In addition, despite the differential realization principle of low pole attenuation (14) the minimization of its parametrical sensitivity to the changing of passive elements of discussed structure is possible. So, at execution of parametrical conditions

$$R_1 = R_2 + r_i + r_u, \qquad \tau_1 = \tau_2(1 - K_i) \qquad (18)$$

realized attenuation

$$d_p = 2\sqrt{1 - K_i} \qquad (19)$$

is accompanied by extremely low sensitivity

$$S_{C_1}^{d_p} = S_{C_2}^{d_p} = S_{R_1}^{d_p} = S_{R_2 + r_i + r_u}^{d_p} = 0. \qquad (20)$$

Thus, to ensure high qualitative indexes of schematics of band-pass sections it is necessary to develop special current amplifiers with transfer coefficient numerical values that following from (19). In this case, as it is evident from (15), the gain

$$K_0 = \frac{1}{2} \qquad (21)$$

is enough small and by multistage filter implementation with additional intersection connections can significantly affect the realizable dynamic range.

By such filter implementation it is essential according to (14), (15) to increase K_i and to optimize the ratio of resistive elements of the scheme. In the general case

$$K_1\frac{R_1}{R_2 + r_i + r_u} = \frac{K_0(K_0 - 1)}{Q^2} + 1. \qquad (22)$$

As seen from the given ratio the required values of K_0 and Q can be realized not only by the special structure of the current amplifier but also by selecting of the ratio of resistive elements R_1 and R_2. Particularly, this ratio can be used for designated parameters correction. In some cases the emerging degree of freedom can be used to optimize the sensitivity of the pole attenuation. Thus, to implement $K_0 = Q$ it is needed to meet the conditions

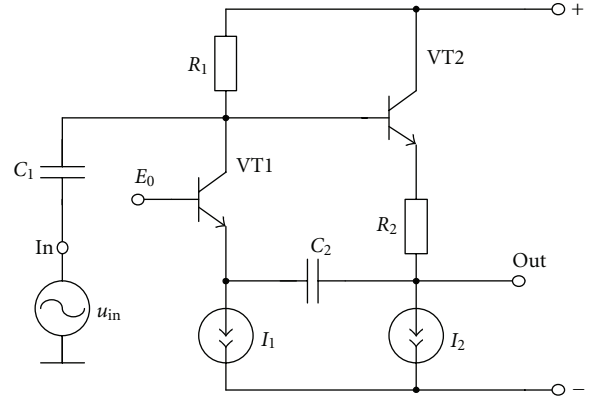$$\tau_1 = \tau_2, \qquad K_i\frac{R_1}{R_2 + r_i + r_u} = 2 - dp, \qquad (23)$$

FIGURE 4: The simplest low-sensitivity section schematic of the band-pass filter.

in this case the parametric sensitivity

$$S_{C_1}^{dp} = S_{R_1}^{dp} = -S_{C_1}^{dp} = -S_{R_2 + r_i + r_u}^{dp} = Q - \frac{1}{2} \qquad (24)$$

is practically defined by realizable Q-factor.

When designing the multisection filters with additional intersection connection the required value of K_0 is limited to values that are close to unity [3]. In this case, for sensitivity reduction it is advisable to use conditions

$$\tau_1 = \tau_2 d_p^2, \qquad K_i R_1 = R_2 + r_i + r_u, \qquad (25)$$

that ensure the equality of this sensitivity to unity.

5. Examples of Second-Order Band-Pass Sections

The simplest current amplifier ($K_i = \alpha$) and repeater (k) are the stages with common base and common collector that have the maximum B_s. Therefore, the simplest scheme of band-pass section is the scheme in Figure 4. Here

$$K_i = \alpha_1, \qquad r_i = h_{11.1}, \qquad r_u = h_{11.2}, \qquad (26)$$

where $h_{11.i} \approx \varphi_T/I_i$, α_i are input resistance and static gain of emitter current of transistor with i index.

According to (13) in this scheme the pole frequency adjustment is possible by changing of bipolar transistors operation modes (by sources I_1 and I_2). However, due to the influence of these parameters (expressions (14), (15)) on the pole attenuation d_p and gain K_0, it is feasible only for small $Q = 1/dp$ values, when conditions (19) and (20) are satisfied. Otherwise, as is evident from modeling results of this scheme (Table 1), this leads to significant changes of specified parameters.

To eliminate this effect in the scheme it is necessary to use a more complex current amplifier that eliminates restrictions on the ratios of resistive elements of the scheme (expressions (22), (23), (25)) and allows to change gain K_i regardless of the resistors r_i and r_u. One of the possible solutions of this problem is shown in Figure 5.

TABLE 1: The results of schematic modeling.

Scheme	f_p GHz	δf_p %	Q —	δQ %	K_0 dB	δK_0 %	I_0 mA	Conditions
Figure 4	1	0.8	8.6	72	13	22	3	$C_1 = 294$ fF, $C_2 = 800$ fF, $R_1 = 360$ $R_2 = 150$
Figure 5	1	0.6	5	3	13	1.4	6	$C_1 = 500$ fF, $C_2 = 800$ fF, $R_1 = 260$ $R_2 = 150$

Note: bipolar heterojunction transistors ($m = 1$, $l_e = 840*10^{-9}$, $W_e = 420*10^{-9}$, $x = 8$, $y = 2$) Supply voltage ± 2.5 V.
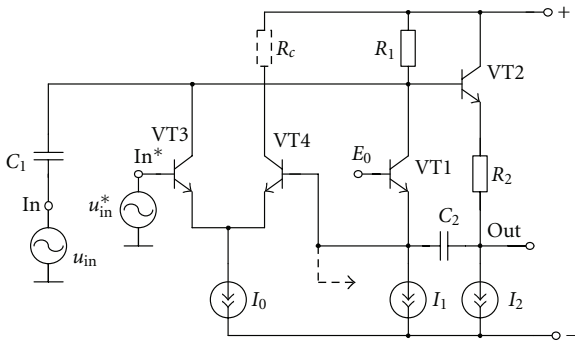

FIGURE 5: Schematic of band-pass section with adjustable pole parameters.

Here

$$K_i = \alpha_1 + \alpha_4 \frac{h_{11.1}}{h_{11.3} + h_{11.4}} = \alpha_1 + \alpha_4 \frac{I_0}{4I_1}. \qquad (27)$$

Thus, by the changing of the mode current I_0 it is possible at given values of $h_{11.1}$ and $h_{11.2}$ (currents I_1 and I_2) to change the K_i and, therefore, by retaining of any parametric condition (18), (19) (22), (23) to implement the required section parameters. The results of comparative modeling of reviewed schemes in software environment Cadence Virtuoso on the base of technological process SGB25VD components (technological inaccuracy of resistors and capacitors is $\pm 10\%$) are presented in Table 1.

The increasing of the current amplifier circuit complexity always generates additional structural degrees of freedom, which can be used to implement certain impedance conditions. So in reviewed scheme the input signal source can be connected to the base of VT4, and the load to the emitter of VT1. In this case, at pole parameters retention the gain at the pole frequency is determined by the following relation:

$$K_0 = -Q\alpha_4 \frac{h_{11.1}}{h_{11.3} + h_{11.4}} \sqrt{\frac{\tau_2}{\tau_1}} \frac{R_1}{R_2 + h_{11.1} + h_{11.2}}. \qquad (28)$$

This property of scheme is convenient to use by additional intersection connections implementation in high-order filters and by integration of band-pass section into different IP-blocks [3].

6. The Main Conclusions

In accordance with the stated task of the work and achieved results it is possible to formulate a number of theoretical conclusions. Firstly, the increasing of active elements use efficiency, in particular current amplifiers, is connected with the usage of symmetric frequency-defining chains in their feedback. This property ensures the independence of the pole frequency (5) from the active element gain and also maximal increasing of the Q-factor ($Q = 1/d_p$) (7), (8). Secondly, this property equalizes the impact of the active elements unity gain area B_s on the parameters of the poles (10), (11). In particular, the use of noninverting amplifiers ($K > 0$) is associated with the creation of passive frequency-defining chains of band-pass type, at the same time according to (9) the required value of gain is reduced and sufficient structural degrees of freedom for gain area B_s maximization are implemented [9]. Finally, the degree of influence of B_s is directly determined by own attenuation of passive circuit pole D_p. From this point of view the expediency of separated RC circuits using is justified and an area of parametric trade-off between different components of the overall sensitivity of schematics is created. This statement requires some comment.

As follows from the general relation (5) and its consequence (13), the pole frequency is determined mainly by the parameters of resistors and capacitors of passive circuit, hence the maximal deviation of amplitude-frequency characteristic and phase-frequency characteristic of sections is determined by the corresponding sensitivity and technological and temperature variations of resistors (Θ_R) and capacitors (Θ_C). That's why this contribution is proportional to the realizable Q

$$\left[\frac{\Delta |F(j\omega)|}{|F(j\omega)|} \right]_{\max} = Q(\Theta_R + \Theta_C),$$
$$\Delta \arg F(j\omega) = (\Theta_R + \Theta_C). \qquad (29)$$

On the other hand (expressions (9), (14)) the pole attenuation and Q-factor are determined by the accuracy of the implementation of these elements ratios ($\Delta\Theta_R$, $\Delta\Theta_C$), that are significantly less than Θ_R and Θ_C for all technologies.

Therefore, at the practical problems solution, the realization of higher parametric sensitivity of this parameter may be accepted (almost up to equal the contributions of Θ_R, $\Delta\Theta_R$ and Θ_C, $\Delta\Theta_C$). From (9) and the properties of symmetrical RC circuits of the second order (the functional dependence of D_p and D_0) follows

$$D_0 = 2m, \qquad D_p = 2m + \frac{1}{m},$$
$$S_m^{dp} = 3mQ, \qquad S_K^{dp} = -QK\frac{1}{m}, \tag{30}$$

where m is the ratio of single-type parameters of passive elements (R or C) with accuracy of implementation that is determined by $\Delta\Theta_R$ or $\Delta\Theta_C$.

Thus, the condition of equality of the contributions is associated with the implementation of conditions

$$\Theta_R = \Delta\Theta_R\frac{3m}{2}, \qquad \Theta_C = \Delta\Theta_C\frac{3m}{2}, \tag{31}$$

that define acceptable maximal value of

$$m \leq \frac{2\Theta_R}{3\Delta\Theta_R}, \qquad m \leq \frac{2\Theta_C}{3\Delta\Theta_C}. \tag{32}$$

As can be seen from (30), (10), (11) $m_{\text{opt}} = 1$ is the sufficient condition for the nondominant nature of the influence of attenuation on the instability of the frequency characteristics of filters. In this case, $D_p = D_{p\min} = 3$, $D_0 = 2$ and, as follows from (9), $K < 3$ for any Q. So

$$S_m^{dp} = -S_K^{dp} = 3Q; \qquad S_{BS}^{\omega p} = S_{BS}^{dp} < -\frac{9}{2}\frac{\omega_p}{B_S}, \tag{33}$$

and the implementation of the required quality factor is associated with the creation of qualitative current amplifier with $1 \leq K_i < 3$ [9].

Similarly, for the separated symmetric chains (example in Figure 4) can be obtained

$$D_0 = m, \qquad D_p = m + \frac{1}{m} \tag{34}$$

and shown that by $m_{\text{opt}} = 1$

$$S_m^{dp} = -S_K^{dp} = 2Q, \qquad S_{BS}^{\omega p} = S_{BS}^{dp} < 2\frac{\omega_p}{B_S} \tag{35}$$

and for any Q value it is required

$$1 \leq K_i < 2 \tag{36}$$

the lower value of the gain at the wider range of operating frequencies.

The last relation shows the benefits of schemes with separate RC circuits when the specified range of values K_i is obtained by bipolar transistors geometry and provides greater stability of the gain.

7. Conclusion

The achieved results show that the limiting estimations of the effective usage of active elements for localization of transfer functions poles (ω_p, d_p) of electrical filters create additional opportunities to optimize the structure and parameters of current amplifiers. In this case, as it follows from (35), the optimal relation of parameters of frequency-defining passive elements not only does not increase the overall instability of the frequency characteristics of filters, but also permits to reduce the influence of gain area (unity gain frequency) on the attainable operating frequency range of scheme. That is what ultimately allows to implement the total decomposition of simple procedures in active filters [10]. As seen from reviewed example of band-pass section, by this approach the functional adjustment of their parameters is easily realized via the mode dependence of input impedance and current amplifier gain. The designing of such amplifiers is possible by the use of common microcircuitry principles of self- and mutual compensation of the parasitic parameters influence of both field effect and bipolar transistors.

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