Запропоновано модифіковану макромодель операційного підсилювача струму, що є компромісним рішенням за критеріями простоти розрахунку, повноти відображення характеристик і точності амплітудно-частотного відгуку. Застосування степеневих поліномів дозволило покращити точність макромоделі без значного ускладнення її структури. Проведено порівняння продуктивності макромоделі з низькорівневою транзисторною архітектурою підсилювача, яке показало допустимість застосування макромоделі при проектуванні активних FBAR фільтрів

Ключові слова: макромодель, OTA, FBAR, активні фільтри, модель нелінійностей

Предложена модифицированная макромодель операционного усилителя тока, являющаяся компромиссным решением по критериям простоты расчета, полноты отражения характеристик и точности амплитудно-частотного отклика. Применение степенных полиномов позволило улучшить точность макромодели без значительного усложнения ее структуры. Проведено сравнение производительности макромодели с низкоуровневой транзисторной архитектурой усилителя, показавшее допустимость применения макромодели при проектировании активных FBAR фильтров

Ключевые слова: макромодель, OTA, FBAR, активные фильтры, модель нелинейностей

УДК 621.375.4

OPERATIONAL TRANSCONDUCTANCE AMPLIFIER MACROMODEL OPTIMIZATION FOR ACTIVE PIEZOELECTRIC FILTER DESIGN

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1. Introduction

Active filters are widely distributed today and perform various frequency processing functions in IF devices, audio equipment and DSP [1]. The success of active filters is explained, first of all by their integration capability and great theoretical basis for their design.

In recent years the progress in the field of thin film bulk acoustic resonators (FBAR) has led to the possibility of creating filters with features, which enable efficient use of frequency range up to 10 GHz [2]. High quality factor and electromechanical coupling coefficient of FBAR resonators allow designing devices with high selectivity, stability and low insertion loss in passband.

At the same time, high frequency architectures of operational transconductance amplifiers (OTA) are actively investigated, allowing overcoming the existing frequency barrier of several MHz for such active elements. Active filters based on OTA elements are simpler in design, have built-in tuning features, and together with FBAR frequency defining elements allows the realization of devices, which can meet modern requirements of operation frequency range, wideband and selectivity.

However, modern OTA are devices of high level integration and include a significant number of elements. Thus, 5-order filter consists of up to 10 active elements each of them includes 10-15 transistors. Simulation of such complex systems, including low-level OTA models, is inefficient in terms of complexity and calculation time.

Macromodel as a simplified analog of transistor architecture allows obtaining simulation results much faster. Thus, in case of transient analysis, calculation time may be reduced by 3 orders of magnitude with almost no loss in accuracy [3]. This opens up the possibility of application of direct numerical optimization techniques for filter structures regarding their nonlinearity, noise effects and dynamic range.

The purpose of this work is the development of the simplified macromodel, which imitate the properties of real amplifiers with sufficient accuracy.

2. Analysis of active elements macromodels and problem statement

In most cases, active elements macromodels are structured to sections (Fig. 1), each section of which takes into account different nonideal effects: finite input and output impedance, noise, frequency dependence of transconductance coefficient and its nonlinearity, systematic and random topology offsets, temperature effects etc. The first input section usually simulates the input characteristics of the amplifier, including differential and common-mode impedances, input nonlinearity. Intermediate sections define the finite gain and frequency response of the amplifier using poles and zeros. The output section simulates output impedance and nonlinearity, output voltage swing etc.

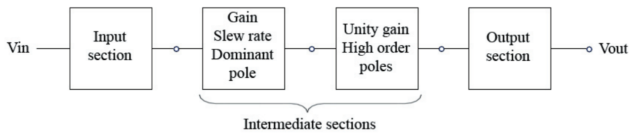


Fig. 1. The structure of the active element's macromodel

Three-section model, represented in Fig. 2, comprises the elements R_i and C_i, representing differential input resistance and capacitance. R_0 and C_0 describe respectively the output values of resistance and capacitance. Common-mode input conductance and capacitance are neglected, since in practice their values are usually much smaller than differential components. This applies to most filter architectures in which one of input terminals is grounded through a resistor or capacitor.

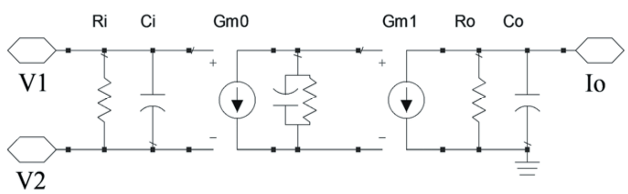


Fig. 2. Three-section macromodel of OTA

Frequency dependence of the OTA's transconductance gain can be described in various ways: a single-pole model [4], two-pole model [5] or pole-zero model [6]. In particular, single-pole model is described with the following equation

$$g_m(s) = \frac{g_{m0}}{1+s\tau}, \qquad (1)$$

where g_{m0} – transconductance DC gain coefficient,

$$\tau = \frac{1}{w_p} = R_p C_p, \qquad (2)$$

with w_p value, corresponding to finite frequency of OTA gain.

Three-section macromodels are often modified by introducing additional sections that take into account certain effects occurring in real active elements. Thus, parametric 7-section macromodel of CMOS OTA (Fig. 3) includes sections with high precision simulation of high input impedance of CMOS circuits, it also separates common-mode and differential gain with possibility to integrate dependencies on several input parameters (W/L, I_{bias}, etc.).

The disadvantage of the model may be the complexity of parameters extraction and the need to transform some links to meet the requirements of the specific CAD.

If there is a need to obtain accurate response of the scheme to changes of bias current Ibias, the third port may be added to the macromodel (Fig. 4). The input section of such a macromodel simulates differential (R_{id}, C_{id}) and common-mode (R_{ic}) input impedances, as well as common-mode rejection ratio (g_{1f}). The second section simulates the saturation mode of OTA, which occurs when the input signal is greater than 30 mV. OTA gain and dominant pole is simulated by the third section. The voltage source of intermediate section is controlled not only by input differential voltage, but also by bias current. The dominant pole is determined by the values of R_2 and C_1. Clarification of frequency response of amplifier is provided by second-order pole, which can be introduced by a special fourth section and R_3, C_2 values. The last section simulates the output impedance of the amplifier and determines the output voltage swing.

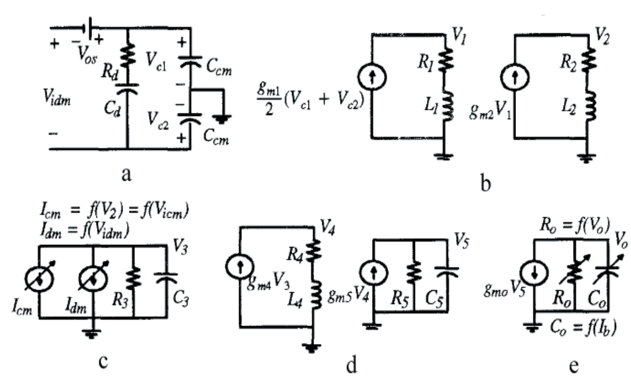


Fig. 3. Parametric 7-section macromodel [7]: a — input section; b — two common-mode sections; c — differential-mode and common-mode gain section with second pole; d — intermediate sections with zero and pole; e — output section and dominant pole

However, if the filter design does not involve tuning of its parameters by changing of I_{bias}, the structure of macromodel can be redundant.

It should be noted that the above macromodels have been developed for special applications, such as tunable OTA-C structure [7] or the quarter-square differential multiplier [8].

From the point of view of active continuous-time analog filter design, the most important characteristics that determine the performance of the whole filter circuit, are total harmonic distortion (THD) and signal-to-noise ratio (SNR). The main impact on THD is the inherent nonlinearity of OTA's transconductance because it induces the harmonic distortion.

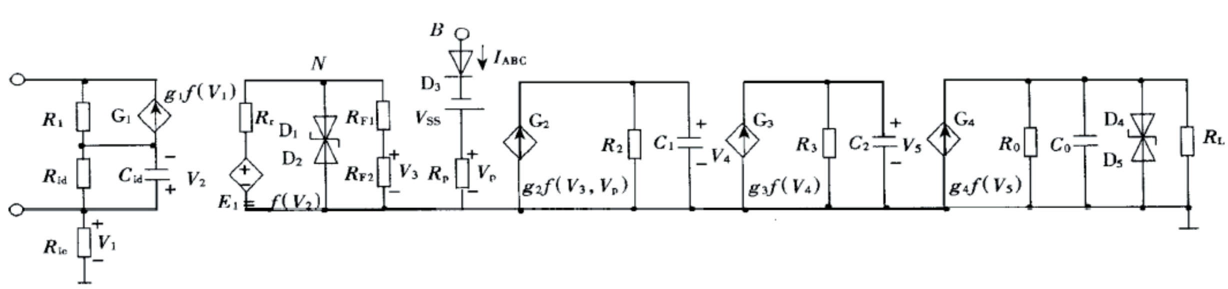


Fig. 4. 3-port OTA macromodel [8]

In turn, SNR is determined primarily by thermal and 1/f noise of OTA. It is important that nonlinearity and noise models may be incorporated into computer aided filter design systems.

So, the main requirements for such models are their versatility (which will provide an opportunity to adapt the same calculation methods for different topologies) and calculation speed, allowing the integration with numerical optimization algorithms.

Thus, the peculiarities of OTA application in high frequency FBAR active filters requires the development of a special macromodel, that takes into account all necessary, but excessive effects.

3. Optimal OTA macromodel design

The proposed OTA macromodel (Fig. 5) is based on a single-pole three-section macromodel with modification which take into account the nonlinear gain and noise characteristics.

As part of this paper we describe the nonlinear distortion model, based on an approximation of the nonlinear range by power polynomial series.

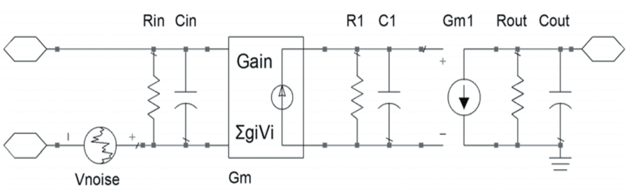


Fig. 5. OTA macromodel with nonlinearity and noise modifications

The transconductance gain gm of such scheme is nonlinear, frequency-dependent and is approximated by single-pole model, in which R_{in} and C_{in}, where C_{in} includes parasitic capacitances (C_{gs} and C_{gd}) of amplifier's differential input transistors. The output impedance is modeled by parallel grounded R_{out} и C_{out}, connected to output terminal.

Despite the fact that modern OTA architectures include various linearization techniques [9], introduced distortion is still quite significant and can be explained by the nonlinear nature of transistors.

The dependence of relative transconductance gain g_m/g_{m0} versus input differential voltage for the conventional OTA is shown in Fig. 6.

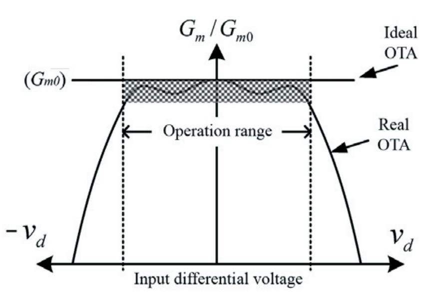


Fig. 6. A typical transconductance gain versus differential input voltage v_d

If the operation range is approximated by power series, the expression for transconductance g_m takes the form

$$g(v)=\sum_{i=0}^{N} g_i v^i = g_0 + g_1 v + g_2 v^2 + g_3 v^3 + \ldots + g_N v^N, \qquad (3)$$

where g_1 determines DC linear gain at $v_d=0$, and g_i are weigh coefficients of respective polynomial components.

The advantage of such an approach is the simplicity of circuit realization in most of specialized CAD systems, and downside is presence of the convergence radius of power series.

In addition, the absolute calculation error is distributed irregularly in the approximation range [10]. Nonlinearity model for the first 4 terms of the polynomial is shown in Fig. 7.

Approximation is applied to a narrow operation range around $v_d=0$. Outside the operation range the obtained g(v) results in significant cumulative errors and inability to use apply transient analysis in a wide range of input voltages. Adequate modeling of transconductance gain saturation was been provided by the introduction of ideal diodes D1 and D2, which cut off the influence of the polynomial series outside the operation range (Fig. 10, a).

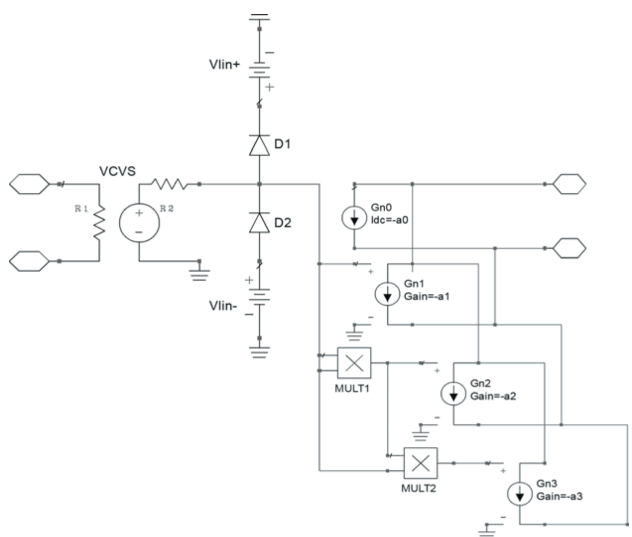


Fig. 7. Nonlinearity model of OTA

Macromodel shown in Fig. 5 also includes noise modeling components. The total noise introduced by OTA circuit is simulated by V_{noise} source, specifying the noise spectral density.

4. Macromodel parameters extraction and performance verification

Macromodel verification was made by the comparative analysis of its performance with low-level architecture of the balanced OTA, performed by 0.18-um CMOS technology (Fig. 8).

Circuit calculation was processed in Agilent ADS with the following conditions ensuring the clearly evident results: symmetric amplifier's power was ±5 V, control voltage -3,8 V,

DC bias current I_{bias} – 833 uA. The macromodel parameters, extracted from the transistor architecture shown in Table 1.

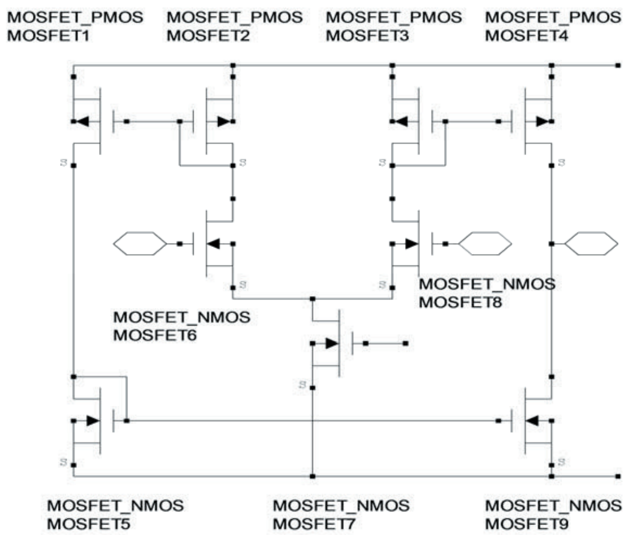


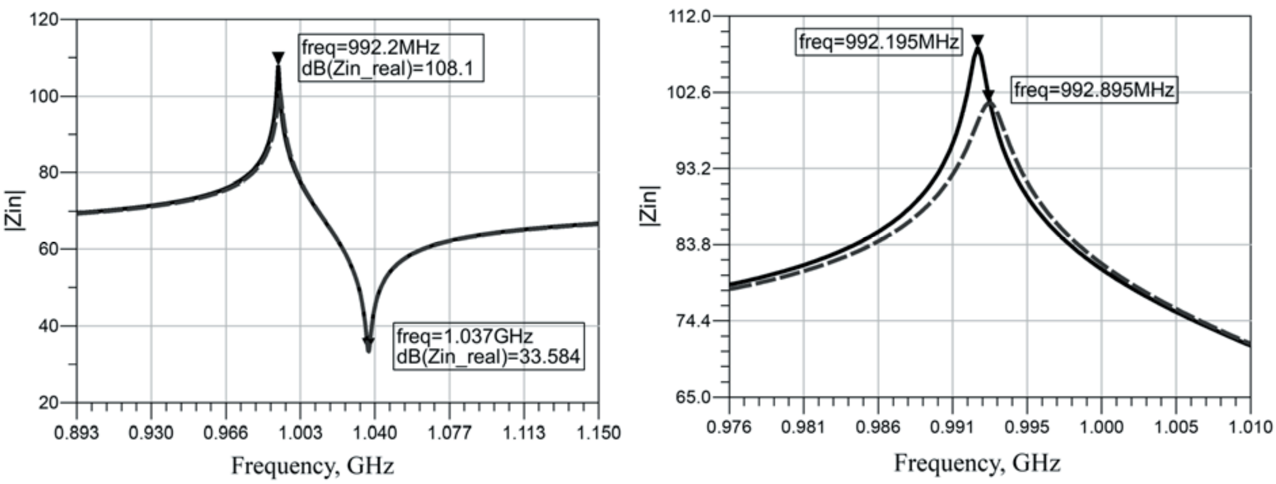
Fig. 8. Transistor architecture of the balanced OTA

Table 1

Extracted parameters of OTA

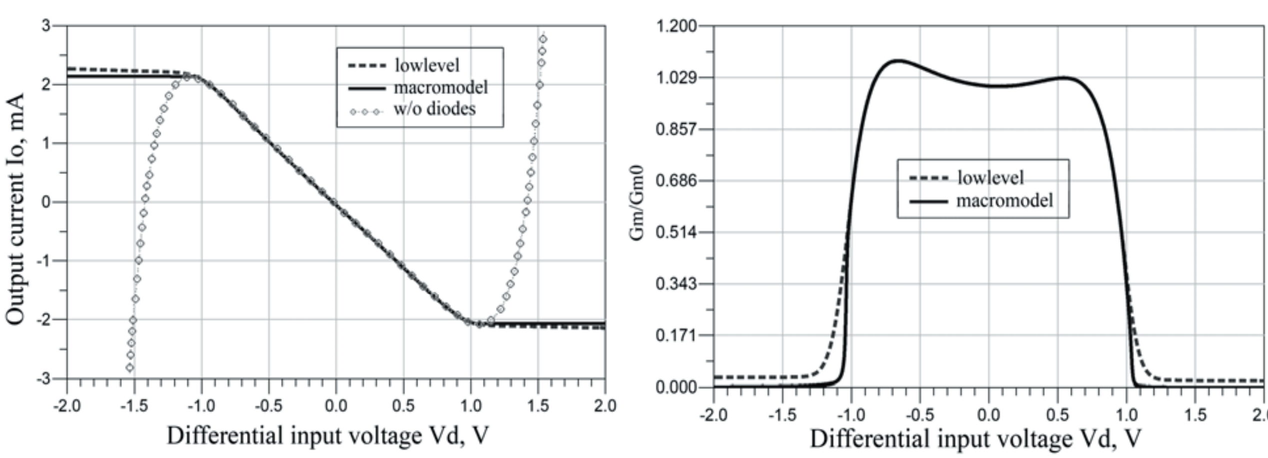
Parameters		Value
G_{m0}	DC transconductance gain	2.102 mS
ωp	dominant pole ($2\pi f_1$)	2π (4.45 GHz)
C_i	input differential capacitance	4.37 fF
R_i	input differential resistance	1.9 GOhm
C_0	output differential capacitance	6.91 fF
R_0	output differential resistance	26.09 kOhm
Vn	noise spectral density	1.7e-8 V/Hz

Input impedance calculation for the tested scheme, consisted of OTA-based gyrator and FBAR Butterworth-Van-Dyke model connected as a load was shown in Fig. 9, a.



a b

Fig. 9. Calculated input impedance Z_{in} versus frequency for transistor-level architecture and OTA macromodel: a — in a wide frequency band; b — near the resonance frequency



a b

Fig. 10. Comparison results: a — transmission characteristics $I_0(v_d)$ of tested structures; b — transconductance gain g_m/g_{m0} versus input differential voltage v_d

The frequency dependence of the input impedance of the transistor architecture displayed by the solid line, macromodel – dashed line.

In a wide frequency band macromodel's response imitates the response of transistor structure with high accuracy, without significant distortion in low-side and high-side frequency ranges.

When considering a narrow frequency band (Fig. 9, b) in presence of inverted series resonance, a small peak shift of 0.7 MHz became evident. With the possible bandwidth of 90 MHz of such a FBAR filter, the frequency error will not exceed 1. 5 %.

The weighting coefficients of polynomial series were extracted in Origin software using the dependence of output current versus differential input voltage of transistor architecture. There were been extracted following 9 coefficients: g_0 = -5.50541e-5, g_1 = -0.00211, g_2 = 3.09364e-5, g_3 = -1.43675e-4, g_4 = 2.02698E-7, g_5 = -2.70194e-7, g_6 = 5.00435e-5, g_7 = 9.66952e-5, g_8 =-7.87091e-6, g_9 = 9.43843e-5.

Transmission characteristics of transistor model and nonlinear macromodel were shown in Fig. 10. The application of diodes, which simulate the saturation of transconductance gain, eliminates the negative polynomial impact outside the operation range. Transconductance gain versus input differential voltage was shown in Fig. 10, b. In the operation range -1..1 V the residual sum of squares (RSS) was 4.05e-12, indicating the high accuracy of approximation of power series function.

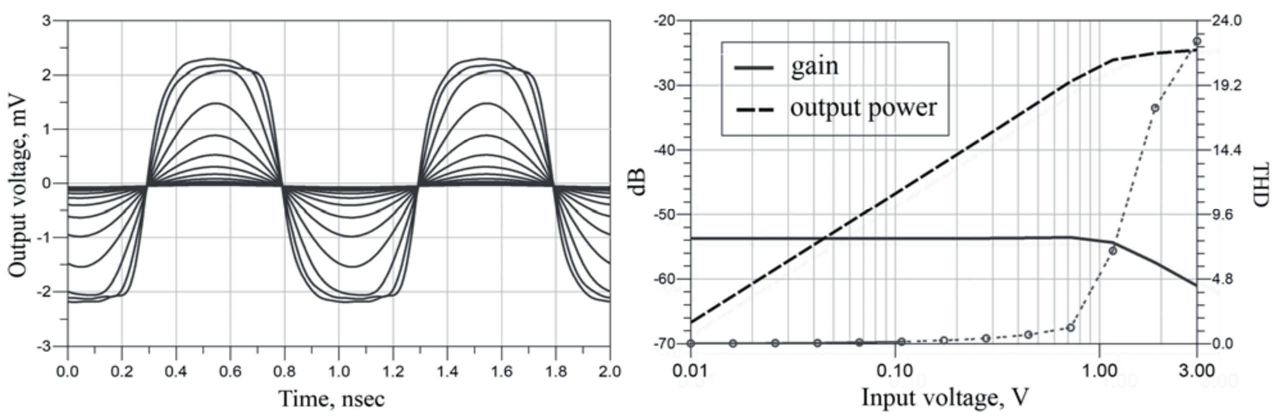
Transient analysis of nonlinear macromodel is shown in Fig. 11, a. The obvious result is significant distortion of the output voltage when the input differential voltages were greater than 0.9 V. A similar conclusion can be traced by examining the characteristics of output power and gain (Fig. 11, b). Total harmonic distortion in the operation range reaches values of 0.3-0.6, which is consequence of deliberately inflated power voltages to provide demonstrable results. Outside the operation range THD dramatically increases, as expected.

5. Conclusions

The nonlinear macromodel of operational transconductance amplifier was proposed for a defined application. The model was developed on the basis of three-section single-pole macromodel and takes into account the necessary amplifier nonidealities: finite values of input and output impedances, frequency dependence of transconductance gain and its nonlinear dependence on differential input voltage and noise characteristics.

Macromodel is versatile enough for application in most modern computer-aided filter design systems, and allows the effective and accurate evaluation of frequency response of the filter, its harmonic distortion and dynamic range. High-speed calculation of macromodel, caused by simple structure, allows the use of direct numerical optimization algorithms to optimize high order filter characteristics.

The verification of macromodel, including calculation of input impedance frequency dependence for the FBAR gyrator circuit and also transient analysis showed that proposed OTA macromodel accurately imitates the performance of transistor-level circuit without significant distortion of its characteristics. Proposed macromodel can be applied in the design of active FBAR filters.



a b

Fig. 11. Macromodel characteristics: a — transient analysis; b — output power, gain and THD performance versus input differential voltage

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ЗАСТОСУВАННЯ ОПЕРАЦІЙ БЕЗ ПЕРЕНОСУ У ВИСОКОШВИДКІСНИХ ОБЧИСЛЮВАЛЬНИХ СИНТЕЗАТОРАХ ЧАСТОТИ

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Розглянуто принципи побудови фазових акумуляторів у прямих цифрових синтезаторах частоти – DDS. Виконано аналіз виникнення затримок розповсюдження сигналу переносу у накопичувачах відліків фази. Розглянуто математичні оператори для побудови високошвидкісних фазових акумуляторів обчислювальних синтезаторів частоти. Запропоновані правила для створення операцій без переносу. Застосування запропонованих правил для побудови фазових акумуляторів дозволить зменшити енергоспоживання синтезаторів та покращити їх тактико-технічні характеристики

Ключові слова: синтезатор частоти, фазовий акумулятор, прямий цифровий синтезатор частоти, накопичувальний суматор

Рассмотрены принципы построения фазовых аккумуляторов в прямых цифровых синтезаторах частоты – DDS. Выполнен анализ возникновения задержек распространения сигнала переноса в накопителях отсчетов фазы. Рассмотрены математические операторы для построения высокоскоростных фазовых аккумуляторов вычислительных синтезаторов частоты. Предложены правила для создания операций без переноса. Применение предложенных правил для построения фазовых аккумуляторов позволит уменьшить энергопотребление синтезаторов и улучшить их тактико-технические характеристики

Ключевые слова: синтезатор частоты, фазовый аккумулятор, прямой цифровой синтезатор частоты, накопительный сумматор

1. Вступ

Прямі цифрові синтезатори частоти відіграють важливу роль у сучасних радіоелектронних пристроях. Це забезпечується багатьма значними перевагами: швидкість переналаштування частоти, висока розрізнювальна здатність, широка синтезована смуга частот. Багаторівневі DDS у силу своєї, технологічності, надійності, можливості мікромініатюризації та унікальності технічних характеристик (нерозривність фази під час перемикання з частоти на частоту, можливість формування сигналів складної форми, цифрове керування амплітудою, частотою та фазою вихідного коливання) на сьогодні знайшли застосування у системах зв'язку. Особливо перспективним є використання DDS у радіотехнічних системах передачі інформації з під-