UDC 621.372.54=111

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# A 180-nm CMOS HIGH-LINEAR COMPLEX $G_m$-$C$ FILTER FOR RECEIVERS OF SATELLITE NAVIGATION SYSTEMS

Realization of high-linear $G_m$-$C$ complex filter in 180-nm CMOS technology to be used in satellite navigation system receivers is covered. The filter with 25 MHz bandwidth was designed with THD of about 80 dB.

$G_m$-$C$ FILTERS; COMPLEX FILTERS; SATELLITE NAVIGATION SYSTEMS RECEIVERS; TRANSCONDUCTANCE AMPLIFIER; 180-nm CMOS.

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# КОМПЛЕКСНЫЙ $G_m$-$C$ ФИЛЬТР С НИЗКИМ УРОВНЕМ НЕЛИНЕЙНЫХ ИСКАЖЕНИЙ НА ОСНОВЕ КМОП-ТЕХНОЛОГИИ С РАЗРЕШЕНИЕМ 180 НМ ДЛЯ ПРИЕМНИКОВ СИСТЕМ СПУТНИКОВОЙ НАВИГАЦИИ

Рассмотрена реализация $G_m$-$C$ комплексного фильтра с низким уровнем нелинейных искажений для применения в приемниках систем спутниковой навигации. Рассчитан фильтр с полосой пропускания 25 МГц с уровнем третьей гармоники порядка −80 дБ.

$G_m$-$C$ ФИЛЬТРЫ; КОМПЛЕКСНЫЕ ФИЛЬТРЫ; ПРИЕМНИКИ СИСТЕМ СПУТНИКОВОЙ НАВИГАЦИИ; ТРАНСКОНДУКТИВНЫЙ УСИЛИТЕЛЬ; КМОП 180 НМ

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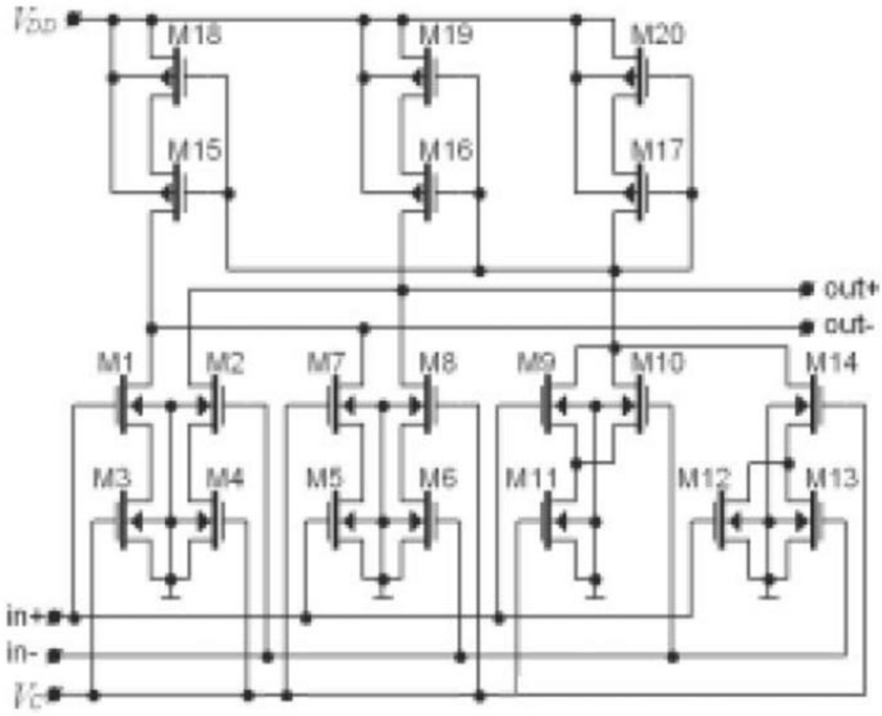
Complex $G_m$-$C$ filters are widely used for channel selection in the front-end of on-chip receivers. The advantages of $G_m$-$C$ filters, as compared to conventional active-RC filters and switched capacitor filters, include a higher operating rate and lower power consumption. However, high linearity is a problem for this type of filter structures. It is mostly caused by the characteristics of a transconductance amplifier (TA). If the linearity problem is solved, the advantage of a high operating rate could be of use in satellite navigation system receivers, e. g. GPS and GLONASS systems, where the channel passband in a broadband mode is about 25 MHz.

This paper presents the implementation of a well-known technique which realizes the $G_m$-$C$ complex filter, described in [1]. It was used for significantly low frequencies — Bluetooth and ZigBee applications. The technique is based on the frequency shift of low-pass filter amplitude response by implementing interconnections. It should be noted that there are just a few known TAs designed for a frequency range of satellite navigation systems. The first part of the paper is devoted to the problem of high-linear TA realization suitable for the proposed application. The second part describes the proposed filter realization. The third part discusses the problem of filter tuning.

### Transconductance Amplifier Realization

Linearization techniques for TAs were properly discussed in [2]. Most common are adaptive feedback technique, source degeneration technique and compensation technique. Even without considering the $G_m/I$ ratio, which is a factor of efficiency for TAs, it was shown that the compensation technique, first proposed by authors of [3], is optimal for the high-linear TA design in terms of THD level in equal conditions. Though lower THD could be achieved with a source degeneration technique, it is relevant in exchange for a lower $G_m/I$ ratio, which leads to higher consumption for the same $G_m$ value.

The proposed TA structure is shown in Fig. 1 *a*. It includes saturation region amplifier section, M1-M4, triode region amplifier section M5-M8, feedforward circuit M9-M14 *a*)




*b*)

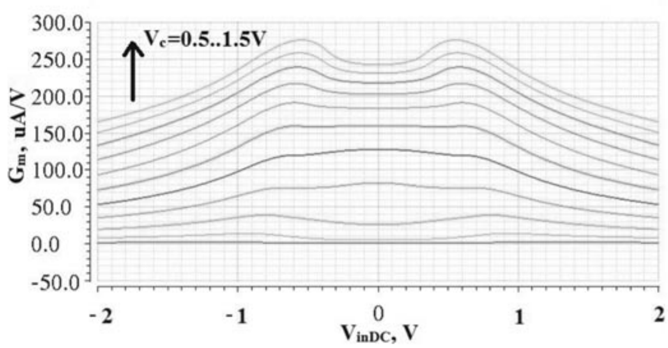



Fig. 1. Transconductance amplifier structure (*a*); DC characteristic of proposed TA (*b*)

Table 1

**Comparison of transconductance amplifiers**

| Transconductance amplifier | [3], 2002 | [4], 2010 | [5], 2012 (source degeneration) | This work |
|---|---|---|---|---|
| Technology, nm | 350 | 180 | 180 | 180 |
| Supply voltage, V | 2.5 | 1.8 | 1.8 | 1.8 |
| THD, dB | >54 (1 $V_{pp}$ 1 MHz) | 51 (1 $V_{pp}$ 50 MHz) | <50 (0.8 $V_{pp}$ 10 MHz) | 59.22 (1 $V_{pp}$ 50 MHz) |
| $G_m$ value, uA/V | 68 | 120–287* | 140–340* | 159 (130–200)* |

* Variation due to $V_c$ change for the fixed linear input amplitude range.

and current mirror as dynamic load M15-M20. Using a parametric analysis in Cadence IC 6.15 for UMC_180 library transistors P_18_MM and N_18_MM, supply voltage $V_{dd}$ of 1.8V and control voltage $V_c$ of 1V, the TA with $G_m$ value of 159 uA/V was designed. The transistor length was set to 240 nm for better operating performance. Resulting transistors sizes are the following: M1-M4, M11: W = 1.95 um, M9-M10: W = 0.975 um, M7-M8, M14: W = 240 nm, M5-M6: W = 2 um, M12-M13: W = 1 um, M15-M20: W = 14.8 um.

In Fig. 1 *b* DC characteristic $G_m = f(V_{inDC})$, where $V_{inDC}$ is the applied voltage difference at the input of the TA, is shown. 1 % $G_m$ variation range is ±672 mV, which is nearly 75 % of the supply voltage limit. Table 1 compares characteristics of the proposed TA versus familiar analogues. The $G_m/I$ ratio, which was mentioned only for [3], is 1.15 for the proposed amplifier. In [3] it was less than 1.

In [2], no information was given for the frequency at which THD was simulated. As it is, for lesser input amplitude of 0.5 $V_{pp}$ THD was certainly less than 59 dB. For better comparison the power consumption and $G_m/I$ ratio should be analysed, but most publications do not provide necessary information.

### Complex Filter

The proposed filter structure is shown in Fig. 2 *a*. Number 1 marks the input TA, which realizes voltage-to-current conversion; 2 marks TAs in a resistor mode; 3 marks a block of gyrators, which realizes inductance imitators; 4 marks blocks of grounded capacitors; 5 marks blocks of interconnection gyrators $G_{11}$-$G_{55}$ for complex reactance transform: each node of a frequency-dependent element, namely capacitor, is connected to its counterpart in other channel. The Frequency shift is defined by the equation:

$$\omega_0 = G_{ii} / C_i,$$

where $G_{ii}$ – value of interconnection TA (in gyrator); $C_i$ – value of the node capacitor.

Calculated values for the capacitors are the following:

$$C_1 = C_5 = 2.50 \text{ pF},\ C_2 = C_4 = 6.55 \text{ pF},$$
$$C_3 = 8.10 \text{ pF}.$$

Taking into account the effect of parasitic capacitance, after the first simulation the capacitor values were recalculated:

$$C_1 = C_5 = 1.89 \text{ pF},\ C_2 = C_4 = 4.96 \text{ pF},$$
$$C_3 = 6.13 \text{ pF}.$$

The values of the TAs needed for interconnection gyrators are the following:

$$G_{11} = G_{55} = 173 \text{ uA/V},\ G_{22} = G_{44} = 453 \text{ uA/V},$$
$$G_{33} = 560 \text{ uA/V}.$$

These TAs were designed by the same method as the main TA. The characteristics do not change linearly with a proportional increase (or decrease) of transistor sizes. Using a parametric analysis, amplifiers with nearly the same linearity and $G_m/I$ ratio were realized.

The complex filter amplitude response $|K(p)|$ is shown in Fig. 2 *b*. The dotted curve refers to a filter response before capacitors recalculation. The solid line refers to a final response of the designed filter. Table 2 compares complex filter characteristics. There are no available works where the bandwidth would be the same, but it is worth noting that even for lower frequencies and narrower bandwidth the linearity of the designed filter is significantly better.

### Tuning

It is known that in $G_m$-$C$ filters the tuning system requires the compensation of the technological variation of parameters. There are different approaches to the problem of building a tuning circuit, e. g. in accordance with [3], it could be a phase-locked loop with the $G_m$-based VCO and reference oscillator $f_{ref}$, as shown in Fig. 3 *a*. The idea is always in tuning by changing the control voltage $V_c$. Taking into account a significant difference in values of interconnection TAs, which could lead to a different change for the same control voltage change, it is imperative to check if the two tuning systems are required.

Fig. 3 *b* shows tuning characteristics of the filter. The control voltage $V_c$ was changed simultaneously for all TAs of the filter structure from 0.9 V (dotted curve) to 1.1 V (dotted-dashed curve). The solid line refers to an initial value of 1 V. The frequency shift is not the

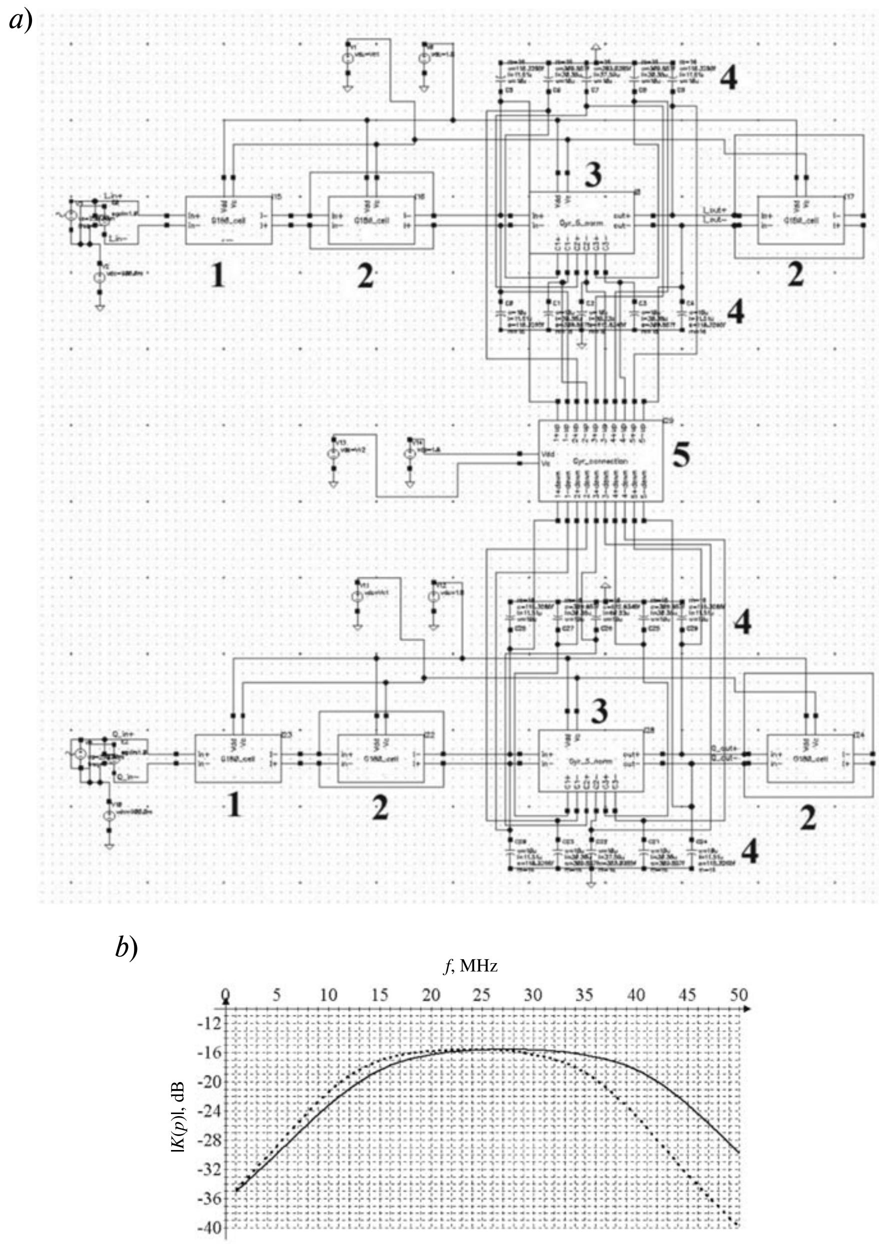



Fig. 2. $G_m$-$C$ complex filter structure (*a*); frequency response of complex filter (*b*)

Table 2

**Comparison of complex filter characteristics**

| Work | [6], 2006 | [7], 2013 | This work |
|---|---|---|---|
| Technology | $G_m$-$C$, 180 nm | ARC, 65 nm | $G_m$-$C$, 180 nm |
| Prototype | Butterworth, 6th order | 6th order | Butterworth, 5th order |
| Center frequency, MHz | 2 | 4–28 | 27.5 |
| Bandwidth, MHz | 3 | 4.2–18 | 25 |
| Image rejection ratio, dB | 55 | > 31 | > 47 |
| Linearity | THD 45 dB (0.2 $V_{pp}$ 2 MHz) | IIP3 25–28 dBm | THD >80 dB (1 $V_{pp}$ 25 MHz) IIP3 33 dBm (100 Ohm) |
| Power consumption, mW | 0.72 | 8.5–26 | ≈36 |

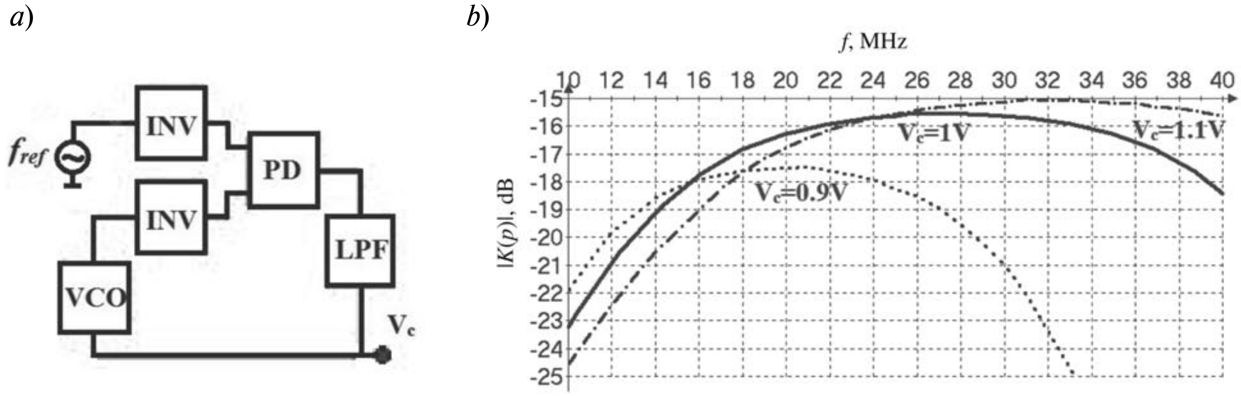



Fig. 3. Block diagram of tuning circuit [3] (*a*); tuning characteristic of proposed filter (*b*)

same for $V_c$ = 1.1 V and $V_c$ = 0.9 V, because of non-linear dependence of $G_m$ versus control voltage, as could be seen from Fig. 1 *b*. A more important thing is that the relative change of the center frequency and bandwidth is not the same (23.63 and 28 %, respectively for 0.9 V and 12.73 and 12 % for 1.1 V). Depending on the target range of the frequency shift and bandwidth change, both the second frequency tuning system and automatic gain control circuit could be required.

A complex 180-nm CMOS $G_m$-$C$ filter with high linearity was designed. The obtained linearity is more than 80 dB THD @1$V_{pp}$ 25 MHz, for 25 MHz bandwidth filter centered at 27.5 MHz. The power consumption is about 36 mW for 5+5 order Butterworth complex filter. The achieved linearity in terms of IIP3 is at least 5 dBm higher. The problem of filter tuning was discussed. It was confirmed that for a relatively high frequency shift more than one tuning system would be required.

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