Dynamic Range Potential of Monolithic Transconductance Amplifier Capacitor Filters

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Abstract

Noise models of CMOS operational transconductance amplifiers (OTA's) are developed. Analysis of the noise performance of monolithic Transconductance Amplifier Capacitor (TAC) filters based upon the derived OTA noise models are made. These noise models are used to characterize the performace of various tuning/adjusting strategies as well as to optimally establish the practical inherent dynamic range limitation for the TAC filters.

I. Introduction

Monolithic filters fabricated using Operational Transconductance Amplifiers (OTA) and Capacitors (termed TAC filters) offer potential for operation at higher frequencies than is attainable with alternative monolithic techniques such as switched capacitors or operational amplifier based active filter structures which use the MOS transistor as a resistor in the ohmic region[1,4]. The filter characterization parameters of interest, such as pole locations, resonant frequencies, Q factors, etc., are directly dependent upon the transconductance gains of the OTA's and the values of capacitors. As a result, linearity and the relationship between signal swing, distortion, the noise and, correspondingly, the dynamic range are of considerable interest.

A MOSFET transistor is essentially a squared law device. This nonlinear nature limits the linear input range of analog MOS circuits. A few techniques have been proposed to improve the linearity of the MOS OTA structures[2,3]. It has been shown that, by applying the right biasing current, these linearized OTA's can handle signals up to $4V_{pp}$ with less than 1% THD if $\pm 5V$ supplies are used. As such, the dominant limitation of the dynamic range is often the noise level.

It has been observed that the noise performance of TAC filters changes with g_m when tuning with the standard dc tail current source in the OTA. A TAC filter structure was recently reported in which two parameters in each OTA are available for adjusting the transconductance; the tail current bias and the mirror gain. The question naturally arises; can any improvements in noise performance be obtained through developing an optimal tuning algorithm. The related question, how should the OTA itself be designed to minimize noise throughout the tuning range also deserves consideration.

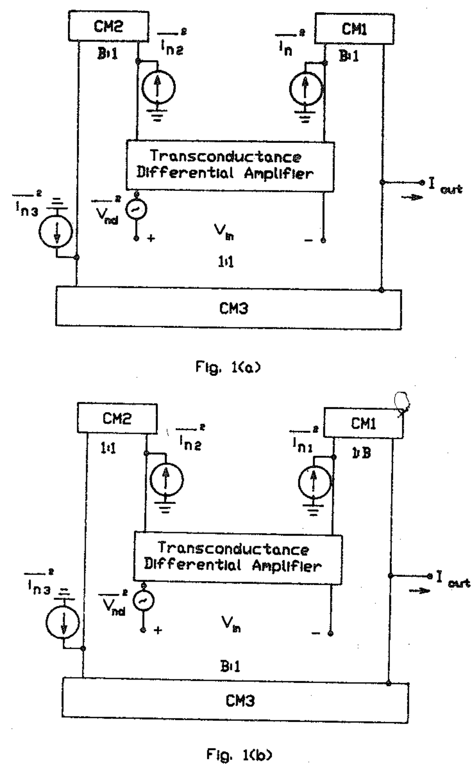

Fig.1 Basic 3-Mirror OTA Structures

In this paper, noise models for basic OTA structures are developed. The performance of two different OTA structures with two different tuning strategies are investigated. Noise models for biquadratic TAC filter structures based upon the derived OTA noise models are presented. Techniques for minimizing noise and their corresponding tradeoffs at both the design and tuning(adjustment) phases are discussed.

II. Noise Modeling of the Operational Transconductance Amplifiers

A common basic OTA structure is composed of a differential input stage and three current mirrors. In the basic OTA, each noisy current mirror block can be represented by a noiseless current mirror in parallel with an input-referred noise current source, and the noisy differential input stage can be modeled as a noiseless differential amplifier with an input-referred noise voltage source. Two variants of the basic OTA are shown along with the equivalent input-referred noise sources in Fig. 1. The total input-referred noise of the two OTA's can thus be derived :

$$\bar{v}_{nT1}^2 = \bar{v}_{nd}^2 + \frac{(\bar{i}_{n1}^2 + \bar{i}_{n2}^2)}{g_{md}^2} + \frac{1}{B^2}\frac{\bar{i}_{n3}^2}{g_{md}^2} \qquad (1a)$$

and

$$\bar{v}_{nT2}^2 = \bar{v}_{nd}^2 + \frac{(\bar{i}_{n1}^2 + \bar{i}_{n2}^2 + \bar{i}_{n3}^2)}{g_{md}^2} \qquad (1b)$$

for the OTA structure in Fig.1(a) and Fig.1(b), respectively, where g_{md} denotes the differential to single-ended transconductance gain of the differential amplifier and B denotes the designated mirror gain. The overall transconductance of the OTA is $g_m = 2g_{md}B$. In Fig.2, several basic MOS circuit elements which are used in OTA designs are presented with their corresponding input-referred noise sources. The input-referred noise spectral density of the simple OTA in Fig.3, which is considered as a direct implementation of Fig.1(a), can thus be obtained by

$$\bar{v}_{nT_1}^2 = 2\bar{v}_{n1}^2 + 2\left(\frac{g_{m3}}{g_{m1}}\right)^2(v_{n3}^2 + v_{n5}^2) + \left(\frac{g_{m7}g_{m3}}{g_{m5}g_{m1}}\right)^2(\bar{v}_{n7}^2 + \bar{v}_{n8}^2) = \bar{v}_{nw}^2 + \bar{v}_{nf}^2 \qquad (2a)$$

where

$$\bar{v}_{nw_1}^2 = \frac{2}{3}\cdot 4kT \cdot \frac{2}{g_{m1}B}\left(B\left(1 + \frac{g_{m3}}{g_{m1}}\right) + \frac{g_{m3}}{g_{m1}} + \frac{1}{B}\left(\frac{g_{m7}}{g_{m1}}\right)\right) \qquad (2b)$$

and

$$\bar{v}_{nf_1}^2 = \frac{2K_n}{W_1L_1C_{ox}f}\cdot\left(1 + \frac{\mu_pK_pL_1^2}{\mu_nK_nL_3^2} + \frac{1}{B}\cdot\left(\frac{\mu_pK_pL_1^2}{\mu_nK_nL_5^2} + \frac{L_1^2}{L_7^2}\right)\right) \qquad (2c)$$

$\bar{v}_{nw}^2$ and $\bar{v}_{nf}^2$ denote the spectral density for the white noise and the flicker($1/f$) noise, respectively. From eq.2, the following observations are made:

(1) A larger aspect ratio $(W/L)_1$ of the input transistors is desired for minimizing both white noise and flicker noise.

(2) Longer channel lengths of all mirror transistors and a higher current gain B introduce less flicker noise.

(3) If the transconductance gain Gm is controlled by adjusting the bias tail current, I_{tail}, in the differential amplifier,

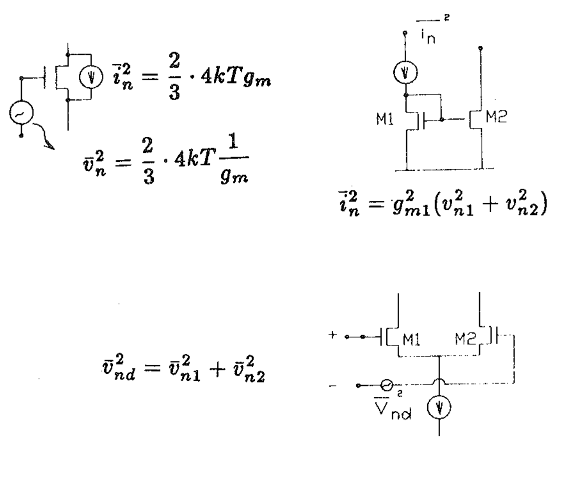

Fig.2 Noise Modeling of MOS Circuit Elements

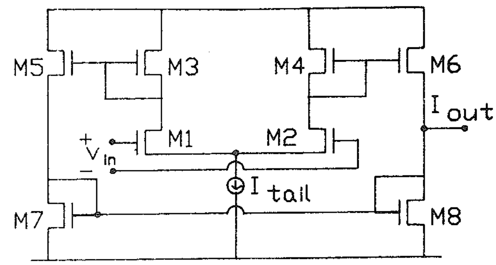

Fig.3 Simple OTA Circuits; the sizing of each transistor is given in Table 1.

Table 1
Device Sizes for OTA1 and OTA2 Structures of Fig.1 Using Topology in Fig. 3.

	M1	M2	M3	M4	M5	M6	M7	M8	Itail
OTA1	10/5	10/5	20/5	20/5	20/5	(6/5)-(63/5)	(6/5)-(63/5)	10/5	1.28-128 μA
OTA2	10/5	10/5	20/5	20/5	20/5	(6/5)-(63/5)	10/5	(3/5)-(32/5)	1.28-128 μA

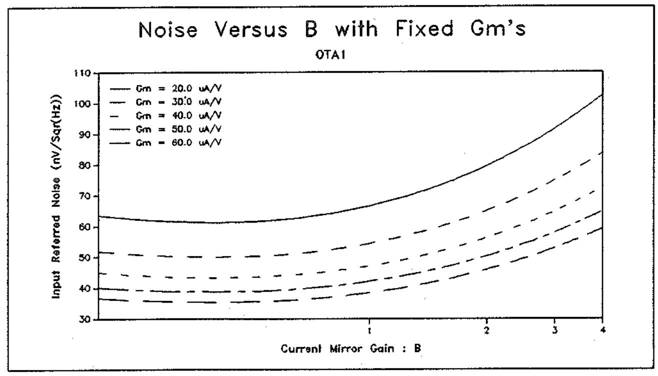

Fig.4 Noise vs. Current Mirror Gain B with Fixed G_m's for OTA1 Structure

then the white noise spectral density becomes inverse proportional to Gm and is given by

$$\bar{v}_{nw}^2 = \frac{K_1}{G_m} \qquad (3)$$

where

$$K_1 = \frac{16}{3}kT\left(B\left(1 + \sqrt{\frac{(W/L)_3}{(W/L)_1}}\sqrt{\frac{\mu_p}{\mu_n}}\right) + \sqrt{\frac{(W/L)_3}{(W/L)_1}}\sqrt{\frac{\mu_p}{\mu_n}} + \sqrt{\frac{(W/L)_7}{(W/L)_1}}\frac{1}{\sqrt{B}}\right) \qquad (4)$$

With tail current adjustment, the current mirror gain, B, can be optimally chosen by minimizing the white noise factor K_1 by letting:

$$B = \left(\frac{\sqrt{\frac{(W/L)_7}{(W/L)_1}}}{1 + \sqrt{\frac{(W/L)_3}{(W/L)_1}}\sqrt{\frac{\mu_p}{\mu_n}}}\right)^{\frac{2}{3}} \qquad (5)$$

In Fig.4 and 5, the relation between the input referred noise and the current mirror gain B is shown with several fixed trasconductance Gm's. Device sizes are shown in Table 1. The existence of an optimal B value is thus obvious. In this structure, the optimal value of B as given by (5) is indepen-

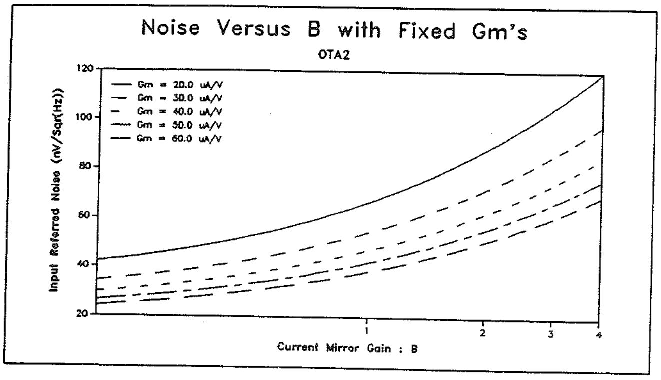

Fig.5 Noise vs. Current Mirror Gain B with Fixed G_m's for OTA2 Structure

dent of Gm. A plot of the noise versus g_m for the optimal value of B appears in Fig.4(c). It should also be noted, according to (eq.2), that changing tail currents does not affect the flicker noise in the OTA.

(4) An alternative way to change the transconductance Gm is by switching the current mirror gain B to various values. From a practical viewpoint, this might be achieved by switching to control the effective size of M5 and M6 in Fig.3(a). This technique has been shown to be capable of providing a larger Gm adjustment range than is attainable by changing the tail current[7].

III. Modeling Noise in Biquadratic TAC Filters

Several biquadratic TAC filter topologies have been proposed in the literature[4]. In this paper, the two typical band-pass structures shown in Fig.6 are analyzed. Structure BP1 consists of three OTA's and two capacitors. Its transfer function is given by:

$$H(s) = \frac{\frac{g_{m1}}{C_2} \cdot s}{s^2 + \frac{g_{m3}}{C_2} \cdot s + \frac{g_{m1} g_{m2}}{C_1 C_2}} = \frac{H_{max} \cdot (\frac{\omega_0}{Q}) \cdot s}{s^2 + (\frac{\omega_0}{Q}) \cdot s + \omega_0^2} \qquad (7)$$

The corresponding output noise spectral density is then obtained:

$$S_n = \frac{\bar{v}_{n2}^2 \omega_0^2 + (\bar{v}_{n1}^2 H_{max}^2 (\frac{\omega_0}{Q})^2 + \bar{v}_{n3}^2 (\frac{\omega_0}{Q})^2) \cdot \omega^2}{(\omega_0^2 - \omega^2)^2 + (\frac{\omega_0}{Q})^2 \cdot \omega^2} \qquad (8)$$

If C_1 and C_2 are fixed, the transconductance gains of the OTA can be uniquely determined as functions of the gain H_{max}, Q factor, and resonant frequency ω_0, which are given as follows:

$$g_{m1} = H_{max} \cdot C_2 \cdot (\frac{\omega_0}{Q}) \qquad (9a)$$

$$g_{m2} = \frac{C_1 \cdot \omega_0 \cdot Q}{H_{max}} \qquad (9b)$$

$$g_{m3} = C_2 \cdot (\frac{\omega_0}{Q}) \qquad (9c)$$

If all the OTA's are adjustable through changing the tail current I_{tail}, the input referred noise in each OTA can thus be modeled by (eq.3). Then the following output noise spectral density equation is established:

$$S_n = \frac{K_1 \cdot (\frac{H_{max} \omega_0^3}{C_1 Q} + \frac{\omega_0 \omega^2 (1 + H_{max})}{C_2 Q})}{(\omega_0^2 - \omega^2)^2 + (\frac{\omega_0}{Q})^2 \cdot \omega^2} = \frac{a_1}{C_1} + \frac{a_2}{C_2} \qquad (10)$$

It is therefore shown that the noise density can be reduced by using larger capacitors in (eq.10). If the total area of capacitance is limited to $C_T = C_1 + C_2$, then the optimal values of C_1 and C_2 can be determined to minimize the noise[5]. By integrating (eq.10) over the passband frequencies, the total in-band noise can be obtained. The relationships between the in-band noise voltage and the filter parameters are shown in Fig.7(a) and Fig.7(b). It is observed that the in-band noise is independent of the resonant frequency ω_0 for both OTA1 and OTA2. Moreover, the in-band noise voltage decreases with increasing Q factors. Unfortunately, large ω_0 or Q adjustment with I_{tail} is not practical with MOS OTA's because of the degradation in input signal swing with decreasing tail current.

When the tuning process is achieved by switching the current mirror gain of the OTA's, the input signal swing becomes essentially independent of g_m. The effect of this alternative scheme is also indicated in Figs.(7). The result shows that the in-band noise becomes ω_0 dependent, and the noise increases with increasing Q.

For the 4-OTA structure BP2, assume that $g_{m1} = g_{m2}$ With this constraint, the transconductance of all four OTA's can again be uniquely determined by filter parameters. The noise performance of this filter is characterized in Fig.8. The in-band noise of BP2 is in general higher than the BP1 structure since one more noisy source has been added.

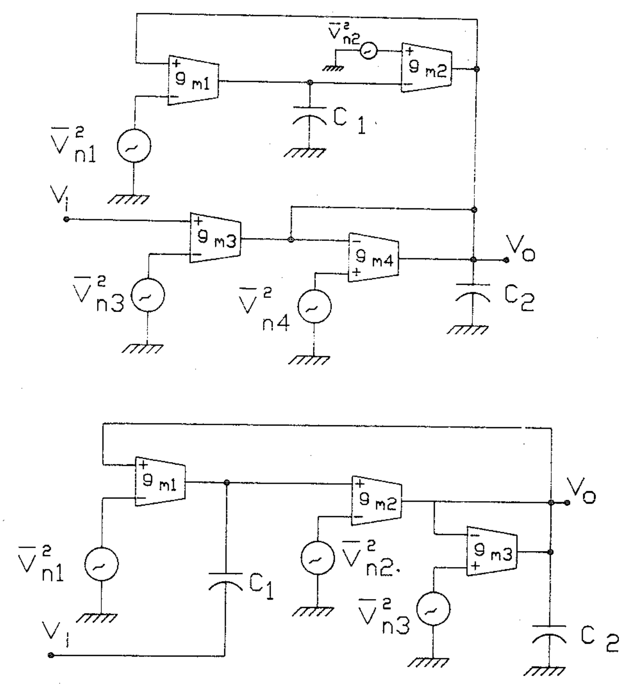

Fig.6 TAC Bandpass Biquad Structure:
(a)3-OTA Structure BP1
(b)4-OTA Structure BP2

IV. Modeling the Dynamic Range For both structures shown in Fig.1, distortions can be contributed by the differential transconductor and current mirror stages. If the current mirrors are relatively distortionless, then the following nonlinear model for both OTA's can be established:

$$i_d = B \cdot g_{m1} v_d \sqrt{1 - \frac{\beta_1 v_d^2}{4 I_{tail}}} \approx B \cdot g_{m1} v_d \left(1 - \frac{\beta_1 v_d^2}{8 I_{tail}}\right) \quad (11a)$$

where

$$\beta_1 = \mu_n C_{ox} (W/L)_1 \quad (11b)$$

provided that $\beta_1 v_d^2 / 8 I_{tail}$ is small. Since a THD less than -40 dB is usually required in most linear applications, the maximum differential input voltage can thus be determined by:

$$v_{dmax} = 0.4 \sqrt{\frac{I_{tail}}{\beta_1}} \quad (12)$$

From eq.(12) and (3), the following relation can be found accordingly:

$$DR \equiv \frac{v_{dmax}^2}{v_n^2} \propto \frac{0.16 B^2}{K_1} \cdot \frac{I_{tail}^2}{G_m} \quad (13)$$

where K_1 and B are given in eqs. (4) and (5). The above equation shows that the input dynamic range can be improved by increasing the tail current. However, the increase of tail current should still be limited to maintain output signal swing.

V. Conclusions

The models of the noise performance for two OTA's and two TAC filter structures have been established and investigated. It has been shown that both the topology and the mirror structure of the OTA impact the noise performance. The OTA tuning strategies for TAC filters also influence the noise performance. With modest tradeoffs between OTA structure, tuning strategy and topology, improvements in noise floor within a 10dB range are possible. The actual improvement in noise performance may be more significant since the noise optimization can be made according to various tuning ranges and strategies.

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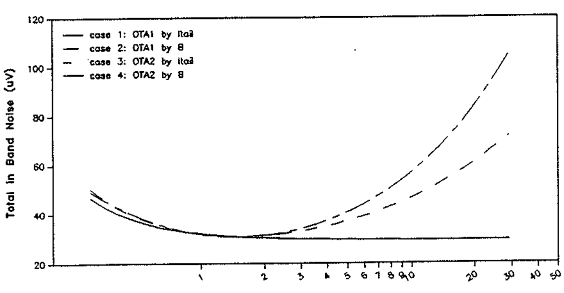

Fig.7(a) Noise vs. Q for the BP1 Structure in Fig.6(a), where case 3 behaves the same as case 1 for $B = 1$.

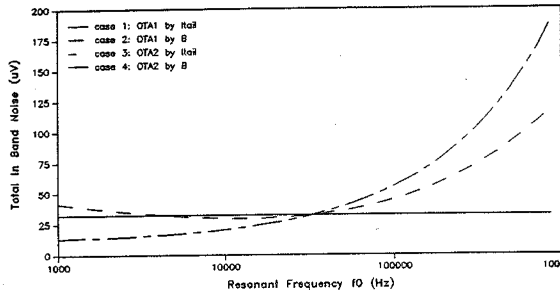

Fig.7(b) Noise vs. f_0 for the BP1 Structure

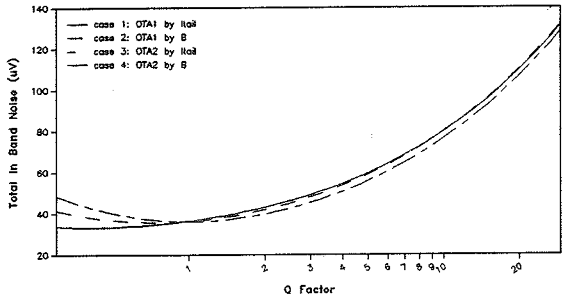

Fig.8(a) Noise vs. Q for the BP2 Structure in Fig.6(b)

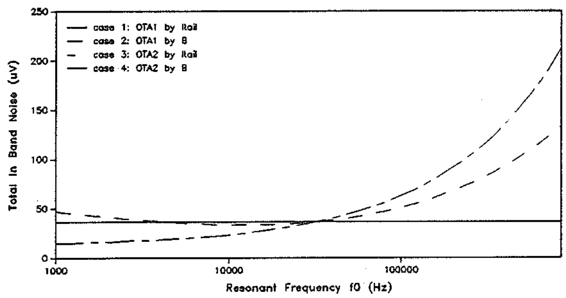

Fig.8(b) Noise vs. f_0 for the BP2 Structure

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