

A 1-MHZ VOLTAGE-CONTROLLED CONTINUOUS-TIME BANDPASS/LOWPASS FILTER USING LINEARIZED CMOS OTAs

A. P. NEDUNGADI and R.L. GEIGER

Dept. of Electrical Engineering, Texas A&M University
College Station, TX, 77843

ABSTRACT

The design and implementation of a continuous-time bandpass/lowpass filter with voltage-controlled center frequency and Q is presented. The circuit uses a linearised CMOS transconductance element as the basic integrating block. A voltage-controlled phase-adjusting scheme is employed in the integrator to compensate for excess phase in the transconductance at high frequencies. The fabricated filter readily achieves the nominal design-center frequency of 1 MHz with $Q = 10$, and provides output signal swings up to $4V_{p-p}$ with 1 % distortion.

INTRODUCTION

Recently, there has been considerable interest in the design and implementation of fully monolithic continuous-time analog signal-processing circuits [1-3]. Such circuits overcome certain inherent limitations of switched-capacitor networks arising due to their sampled-data nature [4]. In particular, the advantages of continuous-time processing become increasingly apparent as operating frequencies are raised well beyond the audio range. A basic requirement for realizing accurate and stable continuous-time filters is a filter block having time-constants that are controllable using a voltage or current, thereby allowing the realized filter to be tuned on chip against process and temperature variations. This paper describes the design and implementation of a voltage-tunable filter block capable of realizing cut-off frequencies in excess of 1 MHz. The circuit is a second-order bandpass/lowpass structure using a CMOS operational transconductance amplifier (OTA) as the basic active element. The OTA employs a linearized input stage which dynamically adjusts the bias current of a differential pair to cancel out nonlinearities of the MOSFETs over a wide input voltage range. A voltage-controlled phase adjusting scheme is employed in the OTA integrators to compensate for the excess phase of the OTAs at high frequencies.

PHASE COMPENSATED OTA INTEGRATOR

Fig. 1 shows the schematic diagram of a phase-compensated integrator using an OTA whose transconductance g_m is adjustable by a control voltage V_C . It is as-

sumed that, at frequencies of interest, the high-frequency behavior of the OTA can be modelled by a single-pole frequency dependent transconductance of the form

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

where g_{m0} is the transconductance value at low frequencies and τ is a high-frequency time constant. This model is fairly accurate at normal operating frequencies where $|s| \ll \tau$. The resulting transfer function of the integrator is obtained as

$$A_I(s) = \frac{g_{m0}}{sC} \left(\frac{1 + sC\tau_d}{1 + s\tau} \right) \quad (2)$$

where

$$\tau_d \approx \frac{\mu_n C_{ox} W}{L} (V_P - V_T) \quad (3)$$

is the small-signal drain resistance of MP in Fig. 1. In the above equation, W and L are, respectively, the channel width and length of device MP , while the parameters μ_n , C_{ox} and V_T have their usual significance [5]. Note that transistor MP introduces a high-frequency zero in the transfer function of the integrator. Since τ_d depends on the gate-source voltage V_P , the location of this zero is voltage-adjustable. The phase lead due to this zero can then be used to compensate for the excess phase lag within the OTA itself. That is, if V_P is adjusted to make $C\tau_d = \tau$, exact phase compensation is achieved. Assuming that the initial phase error is not large, the zero frequency required for compensation is much higher than the unity-gain bandwidth of the compensated integrator. Therefore, at frequencies of interest, the signal voltage across the compen-

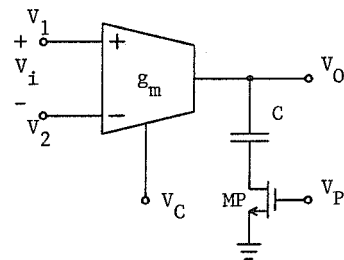


Fig. 1 Phase compensated OTA integrator

and

$$\hat{Q} = \frac{\sqrt{g_{m01}g_{m02}}}{g_{m03}} \sqrt{\frac{C_1}{C_2}} \quad (8)$$

are, respectively, the ideal center frequency and quality factor obtained without high-frequency parasitic effects ($\tau_1 = \tau_2 = 0$) and without phase compensation ($r_{d1} = r_{d2} = 0$). It is evident from (6) that, without compensation, the parasitic time-constants τ_1 and τ_2 can cause significant Q enhancement, especially for large values of \hat{Q} . Exact compensation ($Q = \hat{Q}$) is achieved if r_{d1} and r_{d2} are adjusted to satisfy the condition $C_1 r_{d1} + C_2 r_{d2} = \tau_1 + \tau_2$. Since these drain resistances depend on the phase control voltage V_P according to (3), this scheme provides a convenient method for voltage-tuning the realized Q of the filter. Note from (5) and (7) that ω_o is independent of this Q adjustment. The voltage-control of ω_o is achieved by adjusting V_C which modifies all the transconductance values according to (4). Since the same control voltage is applied to all OTAs, the ratio of transconductances in (8) is constant implying that \hat{Q} is independent of V_C . The realized Q in turn is also largely independent of changes in $\hat{\omega}_o$ since the corresponding term in the denominator of (6) is small. This feature greatly simplifies the task of tuning the overall response of the filter. In practice, however, due to effects neglected in the approximate relations derived here, a small degree of coupling between ω_o and Q is present.

EXPERIMENTAL RESULTS

The filter of Fig. 4 was designed to have a nominal bandpass center frequency $\hat{f}_o = \hat{\omega}_o/2\pi = 1$ MHz, and $\hat{Q} = 10$. In order to simplify design, identical integrators were used with $g_{m01} = g_{m02} = g_{m0}$, $C_1 = C_2$ and $r_{d1} = r_{d2}$. The required value for the third transconductance is then $g_{m03} = g_{m0}/\hat{Q} = g_{m0}/10$. This was obtained by scaling down the gains of the output current mirrors in the OTA (Fig. 2) by a factor of ten. In this way, the linearity of the input stage is preserved. Capacitances C_1 and C_2 had nominal values of 4.5 pF, while MP_1 and MP_2 had $W/L = 25$. The substrates of these devices were short-circuited to their sources. The entire filter, including on-chip output buffers, was fabricated using the process already mentioned [8], and occupies an area of approximately $1000 \times 700 \mu m^2$.

All measurements were made using ± 5 V power supplies with $V_B = -2.5V$. Fig. 5 shows the measured response of the filter at its bandpass output for three different values of V_C corresponding to f_o ranging from 500 KHz to 2 MHz. In each case, V_P was adjusted slightly to maintain the required value $Q = 10$. The relationship between center frequency and control voltage is essentially linear for values of f_o up to 1.7 MHz as expected from the linear dependence of g_{m0} on V_C in (4). Above this frequency, the large V_C required drives $M_8 - M_{11}$ in the OTAs into their ohmic regions of operation, resulting in a decreased

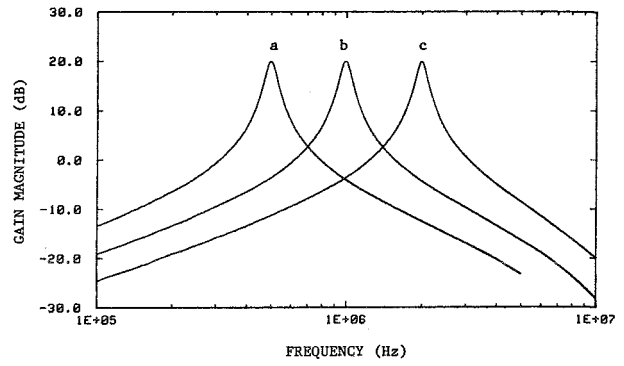


Fig. 5 Measured bandpass response; variation in f_o for $Q = 10$: (a) $f_o = 500$ KHz, $V_C = 1.270V$, $V_P = 1.137V$; (b) $f_o = 1$ MHz, $V_C = 1.510V$, $V_P = 1.300V$; (c) $f_o = 2$ MHz, $V_C = 2.226V$, $V_P = 1.713V$

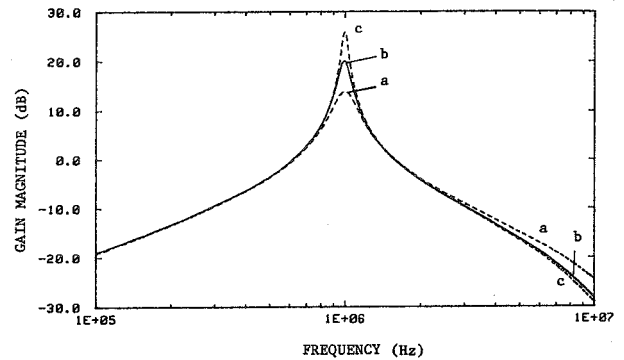


Fig. 6 Variation in Q for $f_o = 1$ MHz, $V_C = 1.51V$: (a) $Q = 5$, $V_P = 1.148V$; (b) $Q = 10$, $V_P = 1.300V$; (c) $Q = 20$, $V_P = 1.725V$

sensitivity of the OTA bias currents to V_C . This causes a nonlinear dependence of f_o on V_C and, for this particular design, limits the maximum obtainable value of f_o to 2.3 MHz. The effectiveness of the phase control scheme was tested by first tuning the filter to the desired specifications ($f_o = 1$ MHz, $Q = 10$) and then varying V_P while keeping V_C fixed. Fig. 6 shows the measured bandpass response for three different values of V_P corresponding to Q values of 5, 10 and 20. The maximum shift in f_o was less than 1% in this case. The distortion characteristics of the filter were studied. Since the filter has a passband gain equal to Q, signal swings are limited at the output of the filter. Fig. 7 shows the output spectrum obtained for the 1 MHz bandpass filter ($Q = 10$) for a $4V_{p-p}$ signal at its output. The total harmonic distortion is approximately 1%. This value is reduced to 0.2% for a $1.5V_{p-p}$ output. The measured inband noise at the bandpass output is approximately $70 \mu V_{rms}$.

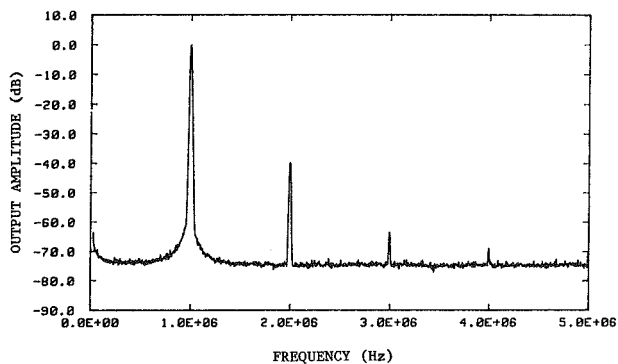


Fig. 7 Distortion spectrum of 1MHz bandpass filter for a $4V_{p-p}$ output at 1MHz

CONCLUSION

The design and implementation of a second-order continuous-time filter with voltage-controlled center frequency and Q has been presented. Experimental results demonstrate the viability of continuous-time processing in the MHz range using a standard 3μ CMOS technology. The voltage-control feature allows this filter to be tuned on-chip against process and temperature variations using known tuning schemes [3, 9]. The implementation of higher-order filters obtained by cascading the proposed second-order block is presently under investigation.

REFERENCES

- [1] M. Banu and Y. Tsvividis, "Fully integrated active RC filters in MOS technology," *IEEE J. Solid-State Circuits*, vol. SC-18, pp. 644-651, Dec. 1983.
- [2] H. Khorramabadi and P. R. Gray, "High-frequency CMOS continuous-time filters," *IEEE J. Solid-State Circuits*, vol. SC-19, pp. 939-948, Dec. 1984.
- [3] C. F. Chiou and R. Schaumann, "Design and performance of a fully integrated bipolar 10.7-MHz analog bandpass filter," *IEEE Trans. Circuits Syst.*, vol. CAS-33, pp. 116-124, Feb. 1986.
- [4] Y. Tsvividis, M. Banu and J. Khoury, "Active RC filters in VLSI: their implementation and advantages," *Proc. Int. Symp. Circuits Syst.*, pp. 239-242, 1985.
- [5] W. M. Penney and L. Lau (eds.), *MOS integrated circuits*. New York: Van Nostrand Reinhold, 1972.
- [6] A. P. Nedungadi and R. L. Geiger, "High-frequency voltage-controlled continuous-time lowpass filter using linearised CMOS integrators," *Electronics Letters*, vol. 22, pp. 729-731, June 1986.
- [7] A. P. Nedungadi and T. R. Viswanathan, "Design of linearized CMOS transconductance elements," *IEEE Trans. Circuits Syst.*, vol. CAS-31, pp. 891-894, Oct. 1984.
- [8] 'MOSIS User's Manual', USC Information Sciences Inst., Marina Del Rey, CA, 1985.
- [9] J. Fincher and R. Geiger, "Monolithic, frequency referenced, digitally self-tuned filters," *Proc. 26th Midwest Symp. Circuits Systems*, INAOE, Puebla, Mexico, 1983.