HIGH-FREQUENCY VOLTAGE-CONTROLLED CONTINUOUS-TIME LOWPASS FILTER USING LINEARISED CMOS INTEGRATORS

Indexing term: Filters

The design and implementation of a continuous-time lowpass filter with voltage-controlled cutoff frequency and passband ripple has been proposed. Most of these techniques realise filters in the audio frequency range, where switched-capacitor (SC) circuits have already been established as a viable approach. However, as operating frequencies are raised by an order of magnitude or more, the advantages of continuous-time processing become increasingly apparent. To date, only a few fully MOS realisations of high-frequency continuous-time filters have been reported. The technique proposed in Reference 2 realises a 500 kHz bandpass filter but has limited signal swing capability due to nonlinearities in the MOSFETs used. Recently, a low-frequency linearising scheme has been extended to high frequencies, but only simulated results using operational amplifiers with 200 MHz gain-bandwidth products are available. This letter discusses the implementation of a 1 MHz lowpass filter using linearised CMOS transconductance integrators. The results obtained from a fabricated test chip are reported.

Filter realisation: Fig. 1 shows the basic transconductance circuit used in the proposed filter. It employs a linearised input stage consisting of a simple source-coupled pair M1, M2 biased dynamically by a current component proportional to the square of the input voltage and the cross-coupled devices, the nonlinearities of the input stage can be largely cancelled out over a wide input voltage range. The remaining devices M3-M6 and coupled through a level-shifting device M7. By properly scaling the W/L ratios of the source-coupled pair and the cross-coupled devices, the nonlinearities of the input stage can be largely cancelled out over a wide input voltage range. The remaining devices M4-M31 are used to bias the input stage and to sum device currents to obtain the final output current. Assuming unity-gain current mirrors and a square-law model for the nonlinear transistors, the following linear i/v characteristic for the complete transconductor:

\[ I_o = g_m V = K(V_C - V_{in})V \]  \hspace{1cm} (1)

where \( K \) is a constant dependent on process parameters and the geometries of \( M_1-M_6 \) and \( M_3-M_{31} \), and \( V_{in} \) is the input voltage. The range of \( V \) over which eqn. 1 is valid and further design details for the basic input stage are discussed elsewhere and are not repeated here.

The complete transconductance circuit of Fig. 1 was fabricated using a standard 3 \( \mu \)m double-poly p-well CMOS process. The transconductance ranges of the geometries of \( M_1-M_6 \) and \( M_3-M_{31} \) were 10 \( \mu \)m/5 \( \mu \)m while \( M_7 \) had \( W/L = 20 \mu m/5 \mu m \). The substrates of these devices were connected to their respective sources. The remaining n-channel devices were in a common p-well connected to \( V_{SS} \). All p-channel devices shared a common substrate connected to \( V_{DD} \). The circuit occupies a total area of 220 x 700 \( \mu \)m\(^2\). Fig. 2 depicts the nonlinearity in the measured i/v characteristics of the fabricated circuit as a percentage of a 2 V (peak) full-scale value, using the nominal supply and bias voltage values indicated. The results are comparable to those for recently reported transconductor schemes.

However, the present circuit has the advantage of not requiring an accurately balanced input drive or a complicated output common-mode biasing circuit. The circuit consumes 10 mW with the nominal bias values, and exhibits a short-circuit 3 dB bandwidth of 15 MHz.

![Fig. 1 Linearised CMOS transconductance circuit](image1)

![Fig. 2 Measured nonlinear error of transconductor for \( V_{DD} = -V_{SS} = 5 \) \( V \), \( V_{in} = -2.5 \) \( V \), \( V_{C} = 1.75 \) \( V \), \( V_{in} = 0 \)

- a Input \( V_{in}, V_{in} = 0 \)
- b Input \( V_{in}, V_{in} = 0 \)

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![Fig. 3 3rd-order Chebyshev lowpass filter using phase-compensated transconductance integrators](image3)

Fig. 3 shows a 3rd-order Chebyshev lowpass ladder filter realised using the transconductor of Fig. 1 as a basic integrating building block. MOS capacitors \( C_{1}-C_{3} \) perform the required integration. The drain resistances of devices \( M_{1}-M_{3} \) in series with the capacitors introduce a high-frequency zero in the transfer function of each integrator. This zero is used to compensate for excess phase shift in the transconductance at high frequencies. The location of this zero is controllable using \( V_{in} \), and the distortion caused by the nonlinear phase-compensating device is mainly due to the second-harmonic component and is approximately given by

\[ HD \approx \frac{(g_m r_f)^2}{(V_p - V_{in} - 2V_{in})} \]  \hspace{1cm} (2)
where \( g_m \) is the integrator transconductance, \( C \) the integrating capacitance and \( r_s \) the small-signal resistance of the compensating transistor. Eqn. 2 has been derived assuming a first-order model of the compensating MOSFET operating in the ohmic region of its \( V/I \) characteristic. For typical values \( g_m, r_s = 0.02, V_A = 2 \text{ V} \) and \( \omega = g_m/C \), eqn. 2 gives \( HD < 0.05\% \).

**Fig. 4 Photomicrograph of experimental chip**

The complete filter of Fig. 3 was designed to have a nominal cutoff frequency \( f_c = 1 \text{ MHz} \) with a passband ripple of \( 1 \text{ dB} \) and was fabricated using the process already mentioned. MP\(_1\)–MP\(_3\) had \( W/L = 25 \) and their substrates were short-circuited to their sources. The entire filter, including the capacitances, phase-compensating devices and on-chip output buffers, occupies an area of \( 1500 \times 700 \mu\text{m}^2 \). A photomicrograph of an experimental chip including the proposed filter as well as single transconductor and buffer test cells is shown in Fig. 4. All measurements were made using \( \pm 5 \text{ V} \) supplies with \( V_B = -2.5 \text{ V} \). Fig. 5a shows the measured frequency response of the 1 MHz filter after adjusting \( V_{C1} = V_{C2} = V_{C3} = V_{C4} = V_{P} \) and \( V_A \) to obtain the specified cutoff frequency and passband ripple. Expanded passband characteristics are given in Fig. 4b for four different values of \( (V_A, V_P) \) corresponding to \( f_c \) ranging from 250 kHz to 2 MHz. In each case the control voltages were adjusted to make the measured passband response within 0.1 dB of the ideal Chebyshev response. For \( f_c \) as high as 1.5 MHz, the variation in cutoff frequency is essentially linear with respect to \( V_A \), as expected from the linear dependence of \( g_m \) in eqn. 1. Above this frequency, the increased value of \( V_A \) required drives \( M_8 - M_{11} \) into their ohmic regions of operation, resulting in a nonlinear dependence of \( f_c \) on \( V_A \). This also limits the maximum obtainable value of \( f_c \) to approximately 2.2 MHz. To test the effectiveness of the phase control scheme, the drain resistances of MP\(_1\)–MP\(_3\) were made very small by applying a large \( V_P \) (+10 V). This effectively disabled the phase compensation and resulted in a 2 dB peaking at the edge of the passband due to excess phase shifts in the integrators. The distortion characteristics of the filter were investigated. Fig. 6 shows the output spectrum obtained for the 1 MHz filter with a 4 V p-p input signal at 250 kHz. The total harmonic distortion in this case is within 1%. This distortion is reduced to 0.2% for a 2 V p-p input. The measured output noise spectral density above 10 kHz in the passband of the 1 MHz filter is fairly constant at 0.1 \( \mu\text{V}\sqrt{\text{Hz}} \). This value increases slightly at lower frequencies due to 1/f noise, rising to 1 \( \mu\text{V}\sqrt{\text{Hz}} \) at 100 Hz.

**Fig. 5 Measured frequency response of filter**

a) Cutoff frequency \( f_c = 1 \text{ MHz} \), \( V_A = 1.554 \text{ V} \), \( V_P = 1.845 \text{ V} \)
b) Expanded passband characteristics:

(i) \( f_c = 250 \text{ kHz} \), \( V_A = 1.554 \text{ V} \), \( V_P = 1.065 \text{ V} \)
(ii) \( f_c = 500 \text{ kHz} \), \( V_A = 1.288 \text{ V} \), \( V_P = 1.270 \text{ V} \)
(iii) \( f_c = 1 \text{ MHz} \), \( V_A = 1.554 \text{ V} \), \( V_P = 1.845 \text{ V} \)
(iv) \( f_c = 2 \text{ MHz} \), \( V_A = 2.562 \text{ V} \), \( V_P = 4.320 \text{ V} \)

**Fig. 6 1 MHz filter output distortion spectrum for \( V_A = 4 \text{ V p-p at 250 kHz} \)**

**Conclusion:** The design and implementation of a novel continuous-time filter technique using linearised high-frequency CMOS transconductors has been presented. A fabricated Chebyshev lowpass prototype exhibits a 1 MHz nominal cutoff frequency and is capable of handling input signals as high as 4 V p-p with less than 1% distortion using \( \pm 5 \text{ V} \) supplies. Using a voltage-controlled phase-compensation scheme, accurate response can be obtained for cutoff frequencies as high as 2 MHz. The voltage-control feature allows the filter to be tuned on-chip against process and temperature variations using known tuning schemes. Extension of this technique to higher frequencies is currently being investigated.

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References

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COMMENT

WIDEBAND FREQUENCY RESPONSE MEASUREMENT OF PHOTODETECTORS USING OPTICAL HETERODYNE DETECTION TECHNIQUE

Kawanishi and Saruwatari report some interesting results about the measurement of photodiode bandwidths using the heterodyne technique.

I do not think, however, that their statement, 'A novel wideband frequency-response measurement system . . . is absolutely correct. In fact, I believe that their method is similar to that reported in Reference B, the only differences being the operating wavelength (850 nm instead of 1.3 μm), the use of conventional single-mode Fabry-Perot laser diodes instead of DFB lasers, and the mixing of the two beams obtained by means of a beam splitter rather than by a fibre coupler.

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REPLY

We are grateful for Dr. Spano’s comment. It is true that our experiment was based on the optical heterodyne detection technique. The first experiment with GaAs/AlGaAs semiconductor lasers was made by Dr. L. Picarri and Dr. P. Spano.

In our letter it has been shown that the polarisation-maintaining single-mode fibre coupler and the isolator stabilised the shape of the beat signal and the DFB-LDs made it possible to continuously sweep the beat frequency up to some tens of gigahertz without longitudinal mode-hopping, resulting in wideband precise measurement of the frequency response of photodetectors.

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References


HIGH RETURN LOSS CONNECTOR DESIGN WITHOUT USING FIBRE CONTACT OR INDEX MATCHING

A noncontacting connector design which reduces the returned power from reflections is demonstrated. The design takes advantage of the poor coupling characteristics between tilted Gaussian beam modes of single-mode fibre. The connector’s mean return loss is −38 dB and its mean insertion loss is 0.7 dB.

Fig. 2 Experimental values and theoretical plot of return loss against reflection angle

Introduction: Optical feedback can degrade laser signal quality by inducing intensity fluctuations and frequency shifts. These effects will become increasingly penalising as fibre system bit rates increase. Since connector reflections are a major source of optical feedback in a system, the suppression of these reflections is particularly important for high bit rate systems. In contacting connectors, reflections are reduced by bringing the fibre faces into intimate contact. The reflections in this case are usually in the vicinity of −30 dB and are limited by the integrity of contact practically achievable. In this letter we present a design with which mean reflections of −38 dB are demonstrated without bringing the fibres into contact. This design has been incorporated into the biconic connector and designated as the SPA biconic connector.