A wide-range tunable BiCMOS transconductor

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This paper describes a fully differential BiCMOS operational transconductance amplifier (OTA) with a transconductance parameter G_m that is tunable over *1* decade. Another significant feature of this OTA is that the tuning scheme is based on current steering resulting in the maximum input signal level being independent of tuning and a reduction in bias dependencies. The circuit is suitable for implementing adaptive analog OTA-C filters for high-frequency data communication applications where moderate SNR is acceptable.

1. Introduction

The transmission of high data rates over twisted pairs has recently received much attention from both academia and industry. Making more efficient use of the existing copper channel is economically advantageous for the next two decades before global deployment of fibre telecommunications. Also, in a "fibre world" short connections to some terminals will remain copper, so improvements in twisted-pair transmission quality and speed are necessary. To support this, sophisticated filtering functions such as pulse shaping to limit signal emissions, echo cancellation to suppress transmission echoes or near-end cross-talk, matched filtering to suppress transmitter filter generated inter-symbol interference and equalization to compensate for cable amplitude and phase distortion are required. Three possible technologies to implement integrated filters exist: switched-capacitor (SC), digital, or continuous-time analog. Since SC filters are limited to a sampling frequency1 $\simeq f_t/50$ while digital filters are limited by analog-to-digital converters operating at $f_s \simeq f_t/30$ and dissipate power of the order of a few watts, analog filters are generally preferred for low-power (100s of mWs), high-speed (100s of MHz) applications. At these frequencies the filter technology receiving the most interest is OTA-C [1-8]. However, process variations and temperature cause OTA-C filters to deviate from their nominal design, while environmental changes such as humidity and different cable makeups cause the channel response to be variable. Furthermore, not all channels are known a *priori.* Thus adaptive filters that can compensate for process variations and track channel variability are essential. The underlying goal of this work is to determine the applicability of analog adaptive filtering for data communications.

This paper is concerned with the analog signal path of an analog adaptive OTA-C filter. A wide-range tunable OTA making use of current steering in a BiCMOS process is discussed. For this OTA the tuning mechanism does not affect input dynamic range and does not upset bias dependencies.

2. Operational transconductance amplifiers

An OTA or a voltage to current converter when loaded with a capacitor implements an integrator. By interconnecting these integrators, a filter of arbitrary transfer-function can be implemented [7], When optimizing for speed, OTAs are run open loop with the integration capacitor acting as a compensation capacitance. The filter passband frequency f_o becomes approximately the unity-gain frequency of the OTAs, f_u, which equals G_m/C, where G_m is the OTA transconductance parameter and C the integration capacitance. Tuning of the filter is generally done by changing the individual integrator's time constant accomplished by varying the bias dependent term G_m.

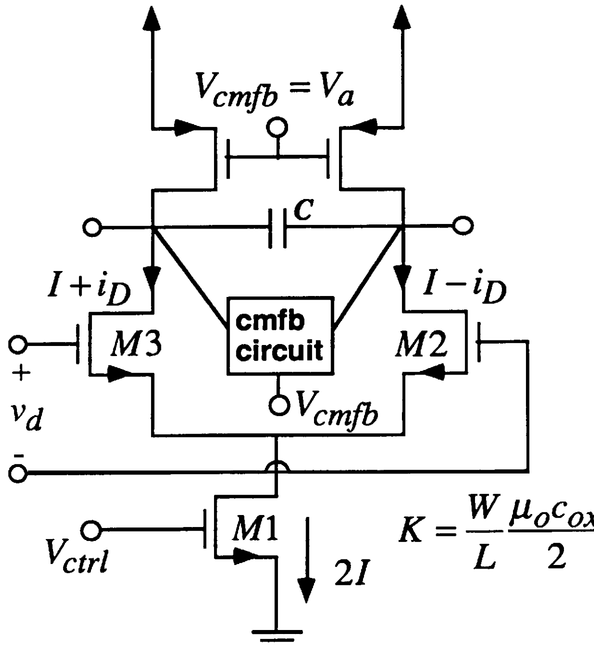

Fig. 1. A simple tunable transconductor integrator,

Figure 1 shows one possible style for an OTA or transconductor, loaded with a capacitor, for which tuning is achieved by varying tail bias current 2I. Other OTAs have also been implemented and achieve their tuning by varying the supply voltage V_{DD} [6], or the input common-mode voltage V_{CM} [7]. At high speeds OTAs are implemented using short-channel devices operating near device f_t. Consequently, process variations are large, of the order of ±33%. Common high-speed OTAs suffer from a low tuning range, of the order of 1-2, which may just cover for process variations. Here, a tuning range of 2 indicates the maximum G_m attainable is twice that of the minimum, while a tuning range of 1 is the limiting case where no tuning is possible. These low tuning ranges are a direct result of a tradeoff between input dynamic range and tuning [1, 7], velocity saturation, mobility degradation [7], and bias constraints. For example, the OTA in Fig. 1 must satisfy the condition $|v_d| \leqslant \sqrt{2I/K}$, hence tuning for lower transconductance or lower speed implies a lower input dynamic range. Similarly, for the OTA of ref [7], $|v_d| \leqslant 2(V_{CM} - V_t)$, while for the OTA in ref. [6], $|v_d| \leqslant 2(V_{DD} - V_t)$. When short-channel devices are used to implement the OTAs above, the MOSFET output current becomes more linear in the signal gate-source voltage due to velocity saturation and degradation of mobility. Consequently, G_m becomes a constant at high gate-source voltages reducing the tuning range to 1. A tuning range of about 2 was obtained for a $0{\cdot}9\,\mu$m implementation of the OTA in ref [7] and a $0{\cdot}8\,\mu$m implementation of the OTA in ref [6] was found experimentally to have a tuning range of $1{\cdot}3$ [9]. As for bias constraints, aside from the need to maintain channels, in the OTA of Fig. 1, for example, notice that as the OTA is tuned the tail bias current is changed resulting in an output offset [4]. This off set will of course cause the common-mode feedback circuit to produce a correction voltage at node V_a to stabilize the output common-mode. However, allowing the common-mode feedback circuitry to correct for tuning generated offset is not desirable, as it could leave its high gain linear range of operation. In the case of an inefficient common-mode feedback circuit, the filtering system would saturate. Tuning also upsets other bias dependencies such as f_o and g_o. Therefore, transconductor DC gain will be modulated by the tuning mechanism. Finally, notice that for all OTAs discussed above only positive or negative G_m can be obtained for a given interconnection of the OTA in a filter loop, not both.

3. The proposed OTA

The basic elements of the proposed transconductor are shown in Fig. 2. By varying control voltages V_{C1} and V_{C2} of the differential pairs Ql, Q2, Qll and Q12 it is possible to obtain both positive and negative values for transconductance. Mathematically:

$$
\begin{aligned}
i_{O1} &= I - (i_{C1} + i_{C11}) \\
&= I - \left[\frac{(I+i)\alpha}{1+\mathrm{e}^{-(V_{C1}-V_{C2})/V_T}} + \frac{(I-i)\alpha}{1+\mathrm{e}^{(V_{C1}-V_{C2})/V_T}}\right]
\end{aligned}
$$

$$
\begin{aligned}
i_{O2} &= I - (i_{C2} + i_{C12}) \\
&= I - \left[\frac{(I+i)\alpha}{1+\mathrm{e}^{(V_{C1}-V_{C2})/V_T}} + \frac{(I-i)\alpha}{1+\mathrm{e}^{-(V_{C1}-V_{C2})/V_T}}\right]
\end{aligned}
$$

For $\alpha \simeq 1$ we obtain

$$
\begin{aligned}
i_{O1} &= \text{-}i \tanh\left(\frac{V_{C1}-V_{C2}}{2V_T}\right) \\
&= -i_{O2} \\
&\triangleq i_o
\end{aligned}
$$

The transconductance is

$$
\begin{aligned}
G_m &= \frac{i_o}{v_d} \\
&= \frac{1}{2}\tanh\left(\frac{V_{C2}-V_{C1}}{2V_T}\right) g_m
\end{aligned}
$$

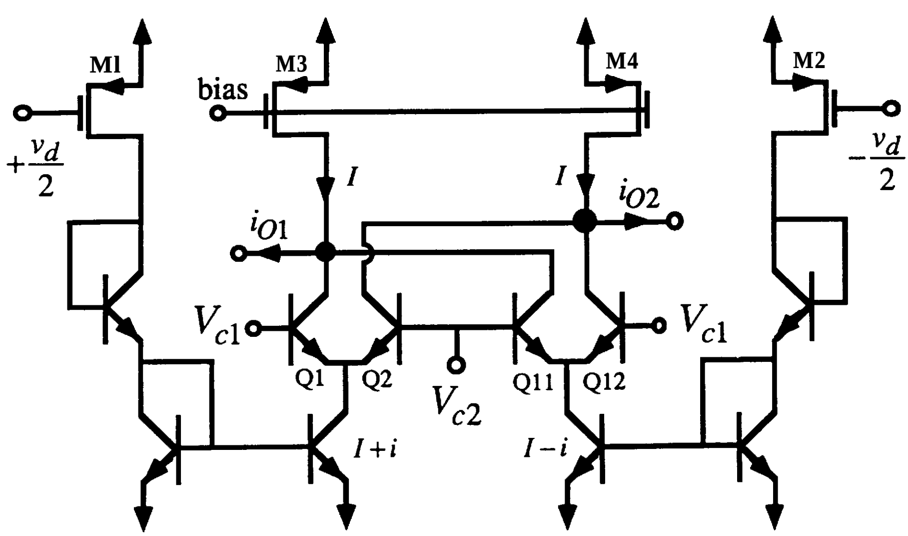

Fig. 2. The basic transconductor illustrating tuning mechanism.

where

$$g_m = \frac{\partial i_{D,M1}}{\partial v_{GS,M1}}$$

Observe that ideally an infinite tuning range can be obtained and that tuning control is independent of the signal voltage. Thus input dynamic range is not influenced by the tuning mechanism nor is the transconductor DC gain. However, unlike the previous circuits where tuning for lower frequencies results in lower power dissipation, this OTA will dissipate its maximum power throughout its entire tuning range. In addition, notice that tuning does not affect bias dependencies and so, ideally no output offset results from the tuning mechanism. Figure 4a shows this transconductance characteristics.

Choosing to linearize the input MOSFETS by operating at high overdrive (sources at signal ground [7]) results in a transconductor common-mode (CM) gain that is greater than 1 for this configuration. A common-mode feedback circuit will then have to compensate for an already large gain. When implementing filters by connecting transconductors in a loop, a CM gain of 1 or greater will lead to instability [7]. To alleviate this problem, the input signal v_d can be supplied to M3 and M4, as shown in Fig. 3. This OTA will have the transconductance characteristics shown in Fig. 4b. The common-mode gain will reduce to a mismatch level. Note that for similar bias conditions this configuration provides twice the G_m (hence twice the speed or twice the conversion efficiency) as compared to that of Fig. 2; however, only positive values for G_m are possible. To obtain positive and negative values for G_m (Fig. 4a), MOSFETS M3 and M4 can each be split to a pair of transistors (each being half the width of the original) with each input cross-coupled.

3.1 OTA frequency response

To support high frequency operation minimum length devices are generally used. A drawback of minimum length devices is that output conductance increases as feature size decreases. This limits this transconductor's DC gain to about 10;

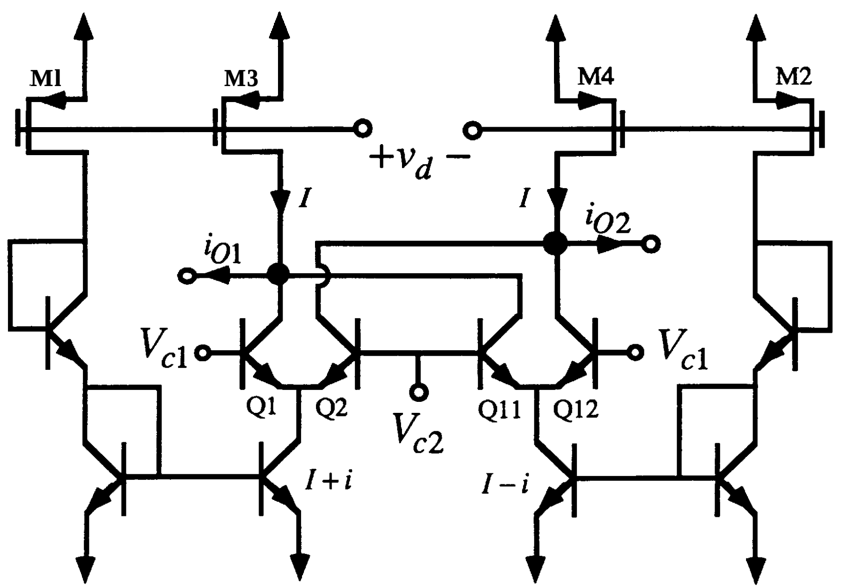

Fig. 3. The improved transconductor.

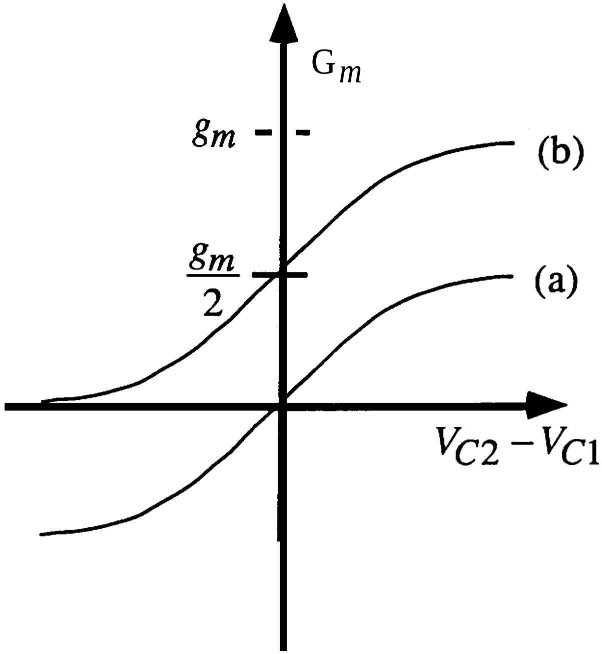

Fig. 4. Transconductance characteristics a, b for the transconductors in Figs. 2 and 3, respectively.

a common value also cited in the literature [6]. Consequently, the effective tuning range for this OTA can be no greater than 10 (limited by $G_m R_o = 1$).

A damped OTA-C integrator will cause filter poles to shift from their designed locations and filter zeros to shift from the jω-axis thus affecting filter response. To enhance DC gain, it is possible to load each OTA with an adjustable transconductor that can realize a negative G_m to cancel the effects of finite output conductance [6]. However, since this OTA will be used to implement an adaptive filter, we prefer to correct for this effect within the filter proper. For example, an adjustable feedforward term can be used to shift filter zeros back to the jω-axis, and a negative transconductance setting can be used to enhance loop Q. Naturally these adjustments will be adaptively controlled, These items will be illustrated in section 4.

Finite transconductor DC gain and parasitic poles and zeros will cause an OTA-C integrator to have a phase response that is not ideal [6-8]. For the OTA proposed in Fig. 3, summing the AC-signal from both input signal paths at the output results in an overall transconductor phase error that increases as the OTA is tuned for lower transconductance (lower speed). This effect will be illustrated graphically. Consider the signal path from M3/M4 to the output to be ideal in the sense that at the unity-gain frequency the phase shift is $-90°$. Let the Ml/M2 signal path have a slight phase error as shown in Fig. 5a. Summing both signals at the output, for a setting giving a maximum value for G_m, results in the sum signal phasor to have roughly twice the magnitude of each component, as expected, but with a slight phase error. Now consider a low value for G_m as in Fig. 5b. Observe that the resultant phasor has a large phase error. Although an adaptive system might correct for errors of this type, a thorough study in the particular application is required to determine system performance in the presence of non-idealities.

3.2 OTA simulation results

The extracted layout representation of the transconductor of Fig. 3 was simulated using HSPICE models for a $0{\cdot}8\,\mu$m BiCMOS process.

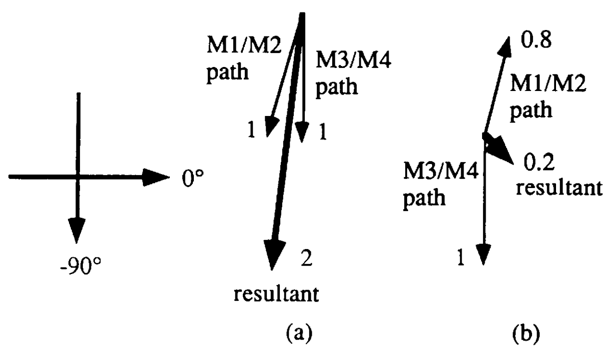

Fig. 5. Illustrating magnitude and phase of output signal for two possible G_m settings. (a) maximum G_m; (b) $0{\cdot}2G_{m,\max}$.

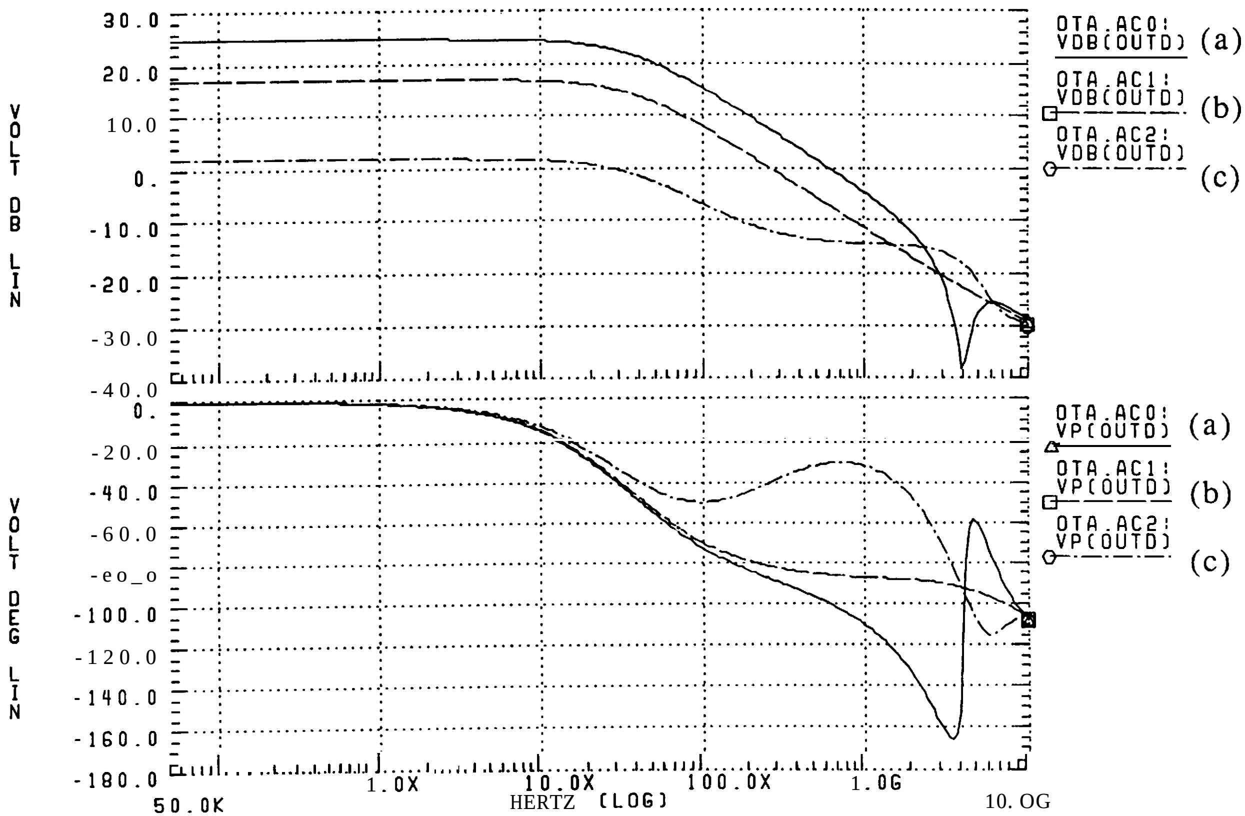

Fig. 6. OTA frequency response for a setting of (a) maximum G_m, (b) $0{\cdot}5G_m$, (c) a low value for G_m.

A common-mode feedback circuit (not shown) was devised to stabilize the output common-mode level. The DC bias current was 250 μA for transistors Ml-M4. This bias condition was chosen to keep the power dissipation of the entire OTA (including bias circuitry, control voltage generation circuitry, and common-mode feedback circuitry) to 10 mW at 5 V. The OTA was simulated with parasitic capacitances only (ie. no load capacitance) to determine maximum attainable frequency response.

The response is shown in Fig. 6 for three different tuning ranges. It is evident that the maximum speed is limited to about 6OOMHz due to the large collector-substrate capacitance of transistors Ql, Q2, Qll and Q12 which is about 30 fF per transistor. The effect of phase error is apparent and most notable at a low setting for G_m.

4. A biquad example

To investigate the concepts described above, the extracted layout representation of the biquad filter of Fig. 7 was simulated. The capacitor values (2C) were 8OfF. The state space representation [7] for this filter is

$$\boldsymbol{A} = \begin{bmatrix} -(g_{o12} + g_{ob}) & G_{m12} \\ -G_{m21} & -(G_{m22} + g_{o22} + g_{oi} + g_{o21}) \end{bmatrix}$$

$$\mathbf{b} = \begin{bmatrix} G_{mb} \\ G_{mi} \end{bmatrix}$$

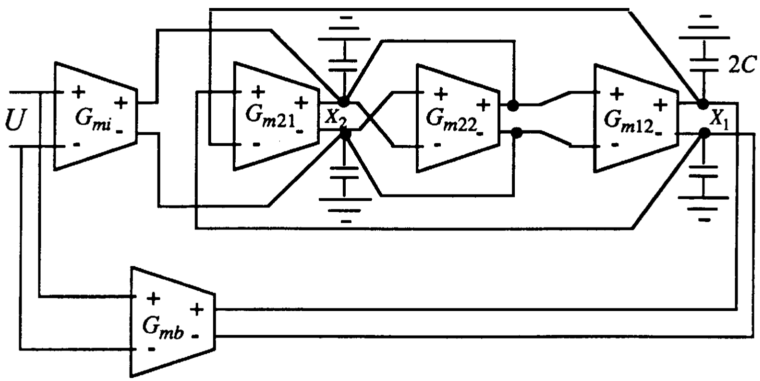

Fig. 7. **A** general biquad.

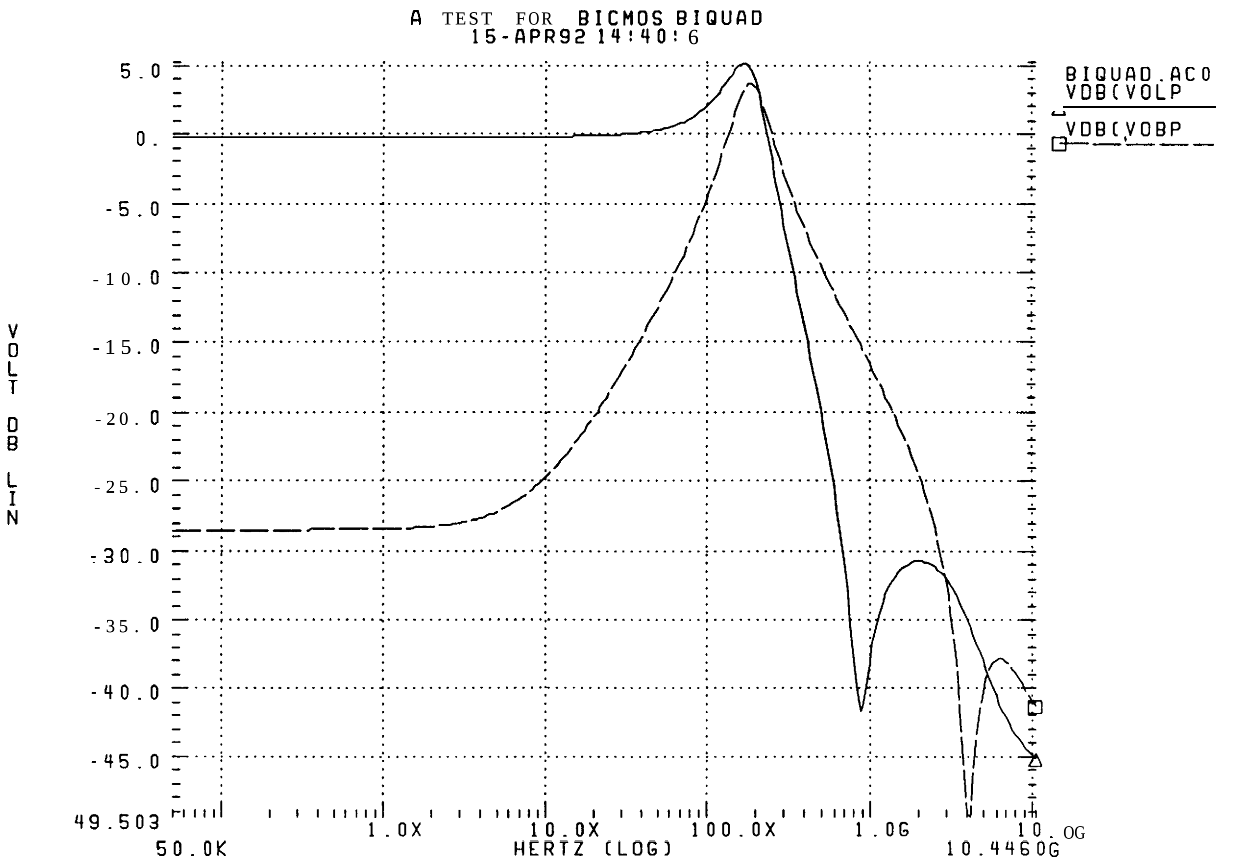

Fig. 8. Illustrating the effects of transconductor excess damping on biquad frequency response.

where the G_ms represent transconductance parameters as in Fig. 7, and the g_os represent respective OTAs output conductance. The transfer function for the bandpass function of this biquad is

$$\frac{X_2}{U} = \frac{\frac{G_{mi}}{C}s + \frac{G_{mi}}{C^2}(g_{o12}+g_{ob}) - \frac{1}{C^2}G_{m21}G_{mb}}{s^2 + \frac{1}{C}(G_{m22}+g_{o22}+g_{oi}+g_{ob}+g_{o12}+g_{o21})s + \left[\frac{(g_{o22}+g_{oi}+g_{o21}+G_{m22})(g_{o12}+g_{ob})+G_{m12}G_{m21}}{C^2}\right]}$$

Transconductor G_{m22} was configured in the $\pm G_m$ mode so that it can be used to enhance Q, while transconductor G_{mb} can be used to shift the mistuned bandpass transfer-function zero ($s = -g_{o12}/C$)* back to the origin. Transconductors G_{m12} and G_{m21} are used to tune the filter f_o. Note from the above transfer-function that adjusting Q via G_{m22} affects f_o. Hence f_o is not independent of Q adaptation, yet this dependency should not affect the operation of an adaptive filter.

Figure 8 shows the typical filter response without feedforward compensation and nominal design Q of 2. Note the effect of the left-half-plane zero at 10 MHz in the bandpass filter

*More correctly, with the G_{mb} block in place the zero is mistuned further to the left: $s = -\frac{1}{C}(g_{o12}+g_{ob})$.

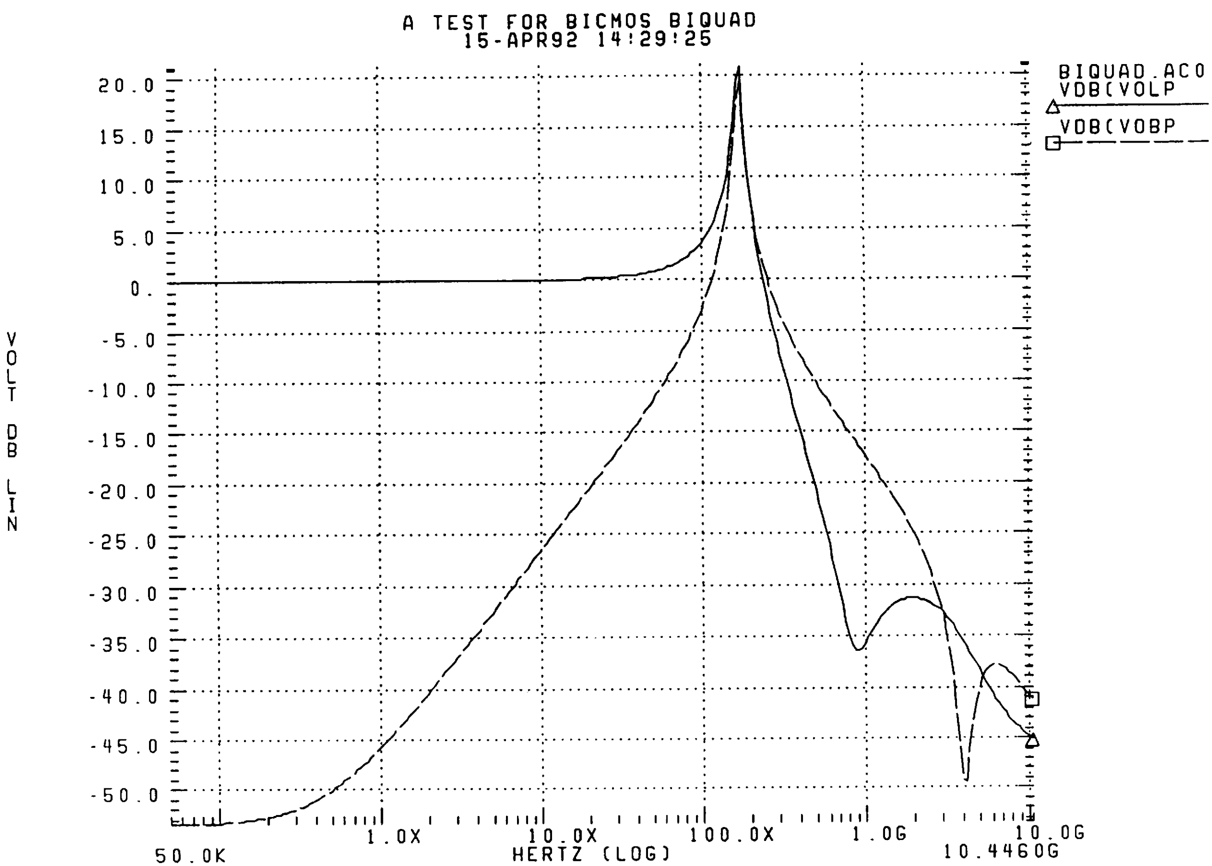

Fig. 9. Biquad frequency response showing the effects of feedforward and Q-enhancement.

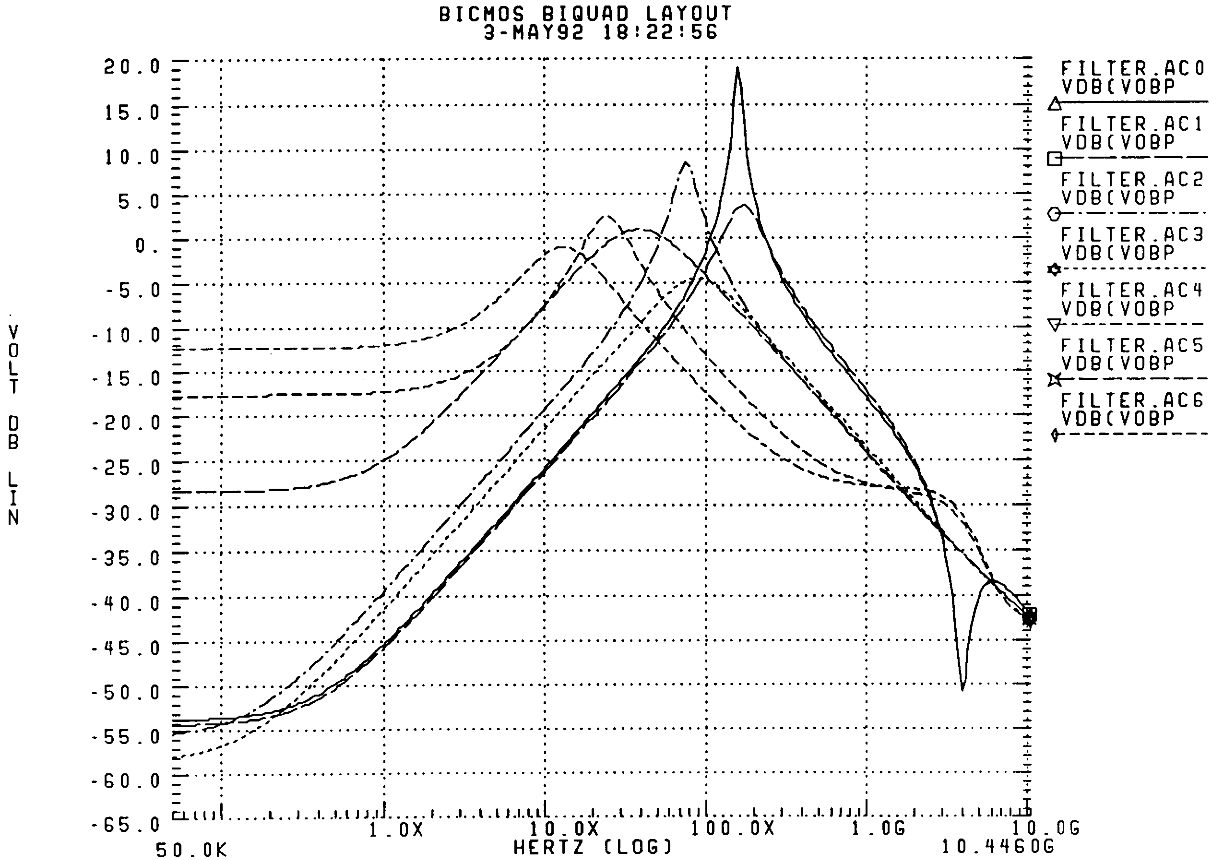

Fig. 10. Bandpass function frequency response for various f_o and Q settings.

response. Figure 9 shows the effect of feedforward compensation by tuning G_{mb} and Q enhancement by tuning for negative transconductance (G_{m22}). Finally, Fig. 10 is a superimposed plot showing the band-pass function at various f_o and Q settings. It can be seen that a continuous tuning range of about 10 is practical.

The state-space representation above is general and can be extended to higher order. With the style of tuning described here, an adaptive filter should therefore be able to realize practical transfer functions for data communication applications.

Conclusions

We have described a fully differential BiCMOS transconductor with a unique style of tuning such that G_m can be varied over 1 decade while DC operating conditions and dynamic range are not affected. We plan to use this transconductor to implement a preliminary adaptive analog filter for data communication applications such as pulse shaping, channel equalization, cross-talk cancellation or matched filtering in the frequency range 50–100 MHz. Through this work we plan to determine if adaptive analog filters have a niche in higher-frequency data communications.

A test biquad has been submitted for fabrication to provide preliminary experimental results.

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