A Wide-Range Tunable BiCMOS Transconductor

Ayal Shoval, David A. Johns and W. Martin Snelgrove
Dept. of Electrical Engineering
University of Toronto, Canada, M5S lA4
(416) 978-3381
shoval@ecg.toronto.edu

ABSTRACT

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.

I. INTRODUCTION

The transmission of high data rates over twisted pairs has recently been receiving 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 crosstalk, 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 frequency $f_s \doteq f_t / 50$ while digital filters are limited by analog-to-digital converters operating at $f_s \doteq f_t / 30$ and dissipate power on 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 [l-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.

II. 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.

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, on the order of ±33%. Common high-speed OTAs suffer from a low tuning range, on 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 [l], [7], velocity saturation, mobility degradation [7], and bias constraints For example, the OTA in Figure 1 must satisfy the condition $|v_d| \le \sqrt{2I/K}$, hence tuning for lower transconductance or lower speed implies a lower input dynamic range. Similarly for the OTA of [7], $|v_d| \le 2(V_{CM} - V_t)$, while for the OTA in [6], $|v_d| \le 2(V_{DD} - V_t)$. 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 0.9μm implementation of the OTA in [7] and a 0.8μm implementation of the OTA in [6] was found experimentally to have a tuning range of 1.3 [9]. As for bias constraints, aside from the need to maintain channels, in the OTA of Figure 1 for example, notice that as the OTA is tuned the tail bias current is changed resulting in an output offset [4] This offset 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.

III. THE PROPOSED OTA

The basic elements of the proposed transconductor are shown in Figure 2. By varying control voltages V_{C1} and V_{C2} of the differential pairs Ql, Q2, **Ql 1** and Q12 it is possible to obtain both positive and negative values for uansconductance.Mathematically:

$$\begin{aligned} i_{O1} &= I - (i_{C1} + i_{C11}) \\ &= I - \left[\frac{(I+i)\alpha}{1+e^{-(V_{C1}-V_{C2})/V_T}} + \frac{(I-i)\alpha}{1+e^{(V_{C1}-V_{C2})/V_T}} \right] \\ i_{O2} &= I - (i_{C2} + i_{C12}) \\ &= I - \left[\frac{(I+i)\alpha}{1+e^{(V_{C1}-V_{C2})/V_T}} + \frac{(I-i)\alpha}{1+e^{-(V_{C1}-V_{C2})/V_T}} \right] \end{aligned}$$

For **a** $\doteq$ 1 we obtain

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

The transconductance is

$$G_m = \frac{i_o}{v_d}$$

$$= \frac{1}{2}\tanh\left(\frac{V_{C2} - V_{C1}}{2V_T}\right) g_m$$

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 transconductancecharacteristics.

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 Figure 3. This OTA will have the transconductance characteristics shown in Figure 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 Figure 2, however only positive values for G_m are possible. To obtain positive and negative values for G_m (Figure 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.

3.2 *OTA Frequency Response*

To support high frequency operation minimum length devices am 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; 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 $j\omega$-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 adaptivefilter, 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 $j\omega$-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 IV.

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 Figure 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 Figure 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 with a slight phase error. Now consider a low value foG_m as in Figure 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.

3.3 *OTA Simulation Results*

The extracted layout representation of the transconductor of Figure 3 was simulated using HSPICE models for a 0.8μm BiCMOS process. A common-mode feedback circuit (not shown) was &vised to stabilize the output common-mode level. The DC bias current was 250μA 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 10mW at 5V. The OTA was simulated with parasitic capacitances only (i.e. no load capacitance) to determine maximum attainable frequency response.

The response is shown in Figure 6 for three different tuning ranges. It is evident that the maximum speed is limited to about 600MHz due to the large collector-substrate capacitance of transistors Ql, Q2, Qll and Q12 which is about 30fF per transistor. The effect of phase error is apparent and most notable at a low setting for G_m.

IV. A BIQUAD EXAMPLE

To investigate the concepts described above, the extracted layout representation of the biquad filter of Figure 7 was simulated. The capacitor values (2C) were 80fF. The state space representation [7] for this filter is

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

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

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

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

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. 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 10MHz in the bandpass filter response. Figure 9 shows the effect of feedforward compensation by tuning G_{mb} and Q enhancement by tuning for negative transconductance(G_{m22}). Finally, Figure 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.

---

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

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.

V. INDUSTRIAL INTERACTION

The research work was partially being done at Bell-Northern Research Ottawa, Ontario. This internship has allowed an early access to a 0.8μm BiCMOS process and an exposure to industrial application requirements, specifications and expertise. It is our hope to provide industry with useful circuit blocks and an alternative filter technology to support high-speed data communications. The internship was supported by the I.C. Department headed by Tom Kozelj. Discussions with department members, specifically Mike Altmann, are greatly appreciated.

VI. 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 50-1OOMHz. 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.

REFERENCES

[1] P. Visocchi et al, "Novel Tunable GaAs Mosfet OTA-C Integrators Suitable for High Precision Filtering Applications," 1992 ***IEEE Int. Symp. Circuits Syst., San-Diego, California*** vol. 1. pp. 212-215, May 10-13, 1992.

[2] J. Ramirez-Angulo and I. Grau, "Wide gm Adjustable Range, Highly Linear OTA with Linear Programmable Current Mirrors," 1992 ***IEEE Int. Symp. Circuits Syst., San-Diego, California*** vol. 3, pp. 1372-1375, May 10-13, 1992.

[3] A. Wyszynski et al, "Design of a2.7GHz Linear OTA in Bipolar Transistor-Array Technology with Lateral PNPs," 1992 ***IEEE Int. Symp. Circuits Syst., San-Diego, California,*** vol. 6, pp. 2844-2847, May 10-13, 1992.

[4] P. Wu, R. Schaumann and P. Latham, "Design Considerations for Common-Mode Circuits in Fully Differential Operational Transconductance Amplifiers with Tuning," 1991 ***IEEE Int. Symp. Circuits Syst.,*** pp. 1363-1366, May 1991.

[5] R. Castello et al, "A Very Linear BiCMOS Transconductance Cell for High-Frequency Filtering Applications," 1990 ***IEEE Int. Symp. Circuits Syst., New Orleans, Louisiana,*** vol. 2, pp. 1364-1367, May 1-3, 1990.

[6] B. Nauta, "A CMOS Transconductance-C Filter Technique for Very High-Frequencies," ***IEEE Journal of Solid-State Circuits,*** vol. SC-27, no. 2, pp. 142-153, Feb. 1992.

[7] M. Snelgrove and A. Shoval, "A Transconductance-C Biquad Tunable Over the VHF Range," ***IEEE Journal of Solid-State Circuits,*** vol. SC-27, no. 3, pp. 314-323, March, 1992.

[8] K. D. Peterson, A. P. Nedungadi and R. L. Geiger, "Amplifier Design Considerations for High Frequency Monolithic Filters," in ***Proc. ECCTD 87 (Circuit Theory and Design),*** R. Gerber, Ed. Amsterdam: Elvsier Science, 1987, pp. 321-326.

[9] David Ryan, private communications, BNR.

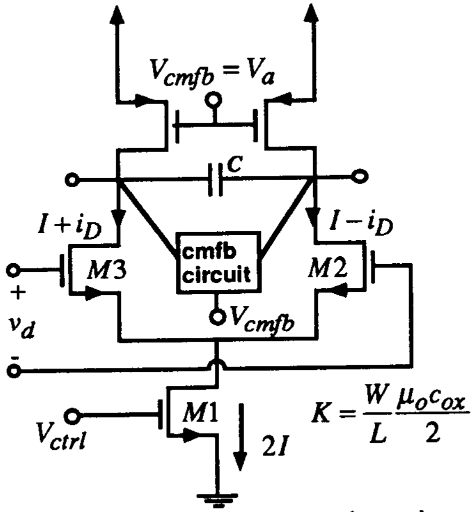

Figure 1: A simple tunable transconductor integrator.

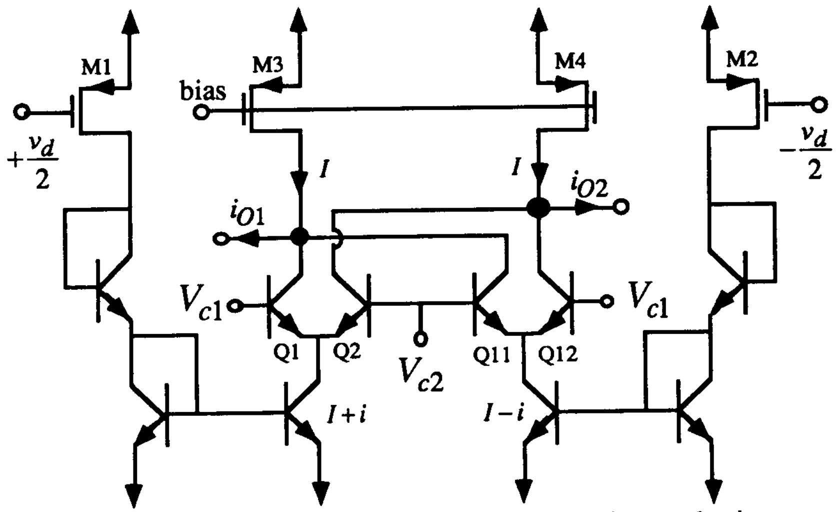

Figure 2: The basic transconductor illustrating tuning mechanism.

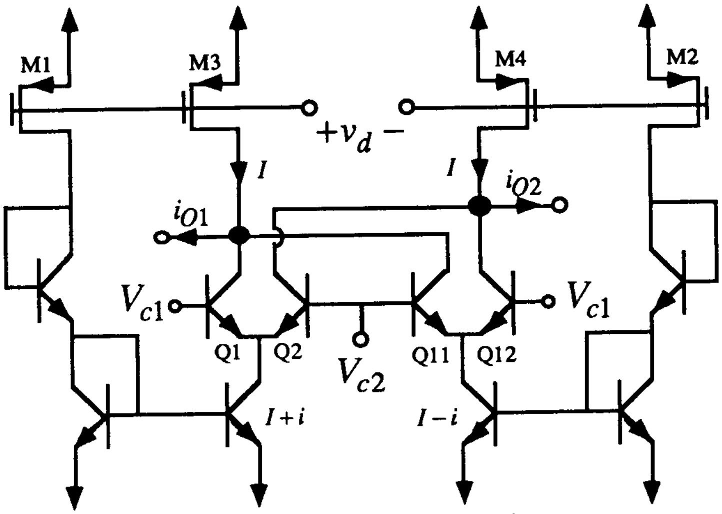

Figure 3: The improved transconductor.

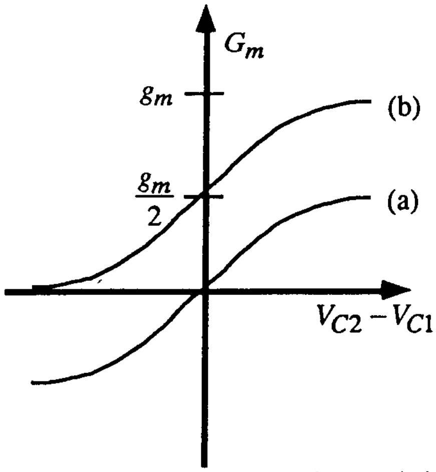

Figure 4: Transconductance characteristics a,b for the transconductors in Figure 2 and 3 respectively.

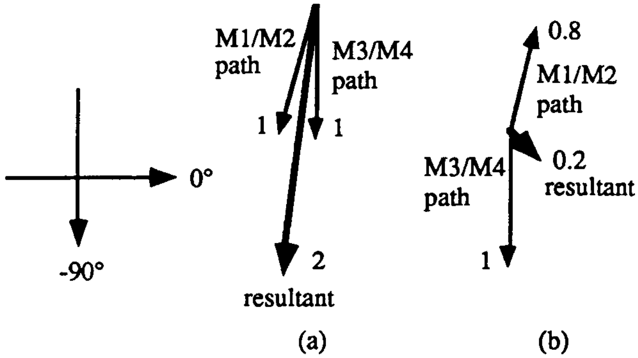

Figure 5: Illustrating magnitude and phase of output signal for two possible G_m settings; a) maximum G_m, b) $0.2G_{m,\mathrm{max}}$.

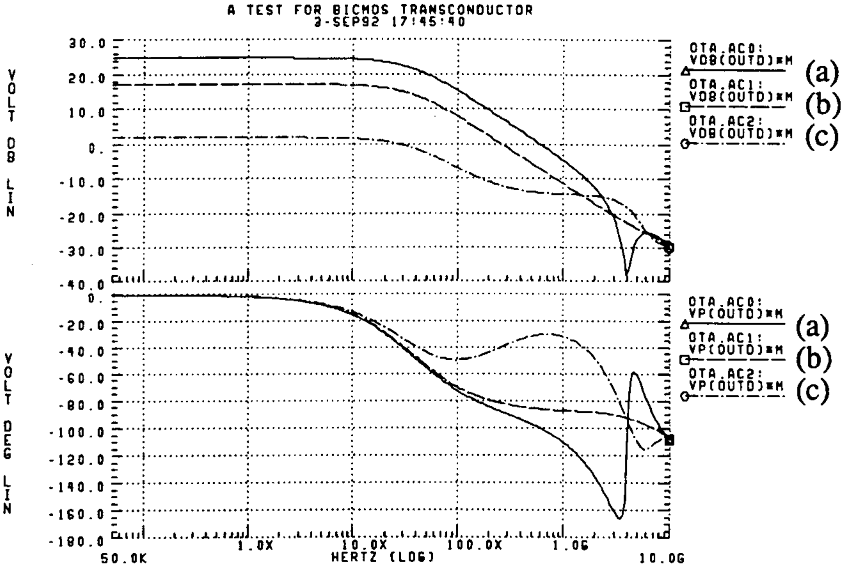

Figure 6: OTA frequency response for a setting of , a) maximum G_m, b) $0.5G_m$, c) a low value for G_m.

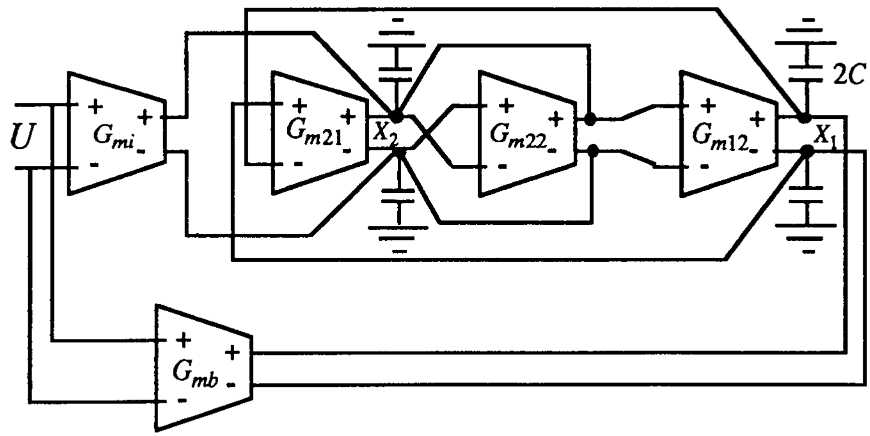

Figure 7: A general biquad.

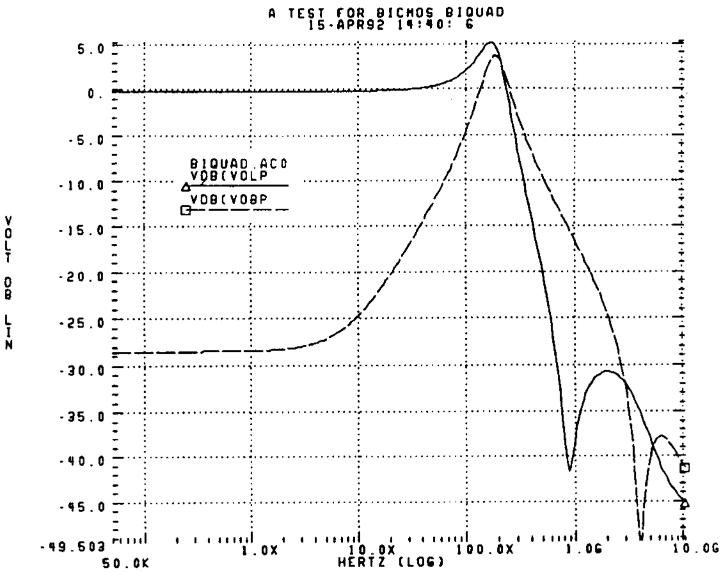

Figure 8: Illustrating the effects of transconductor excess damping on biquad frequency response.

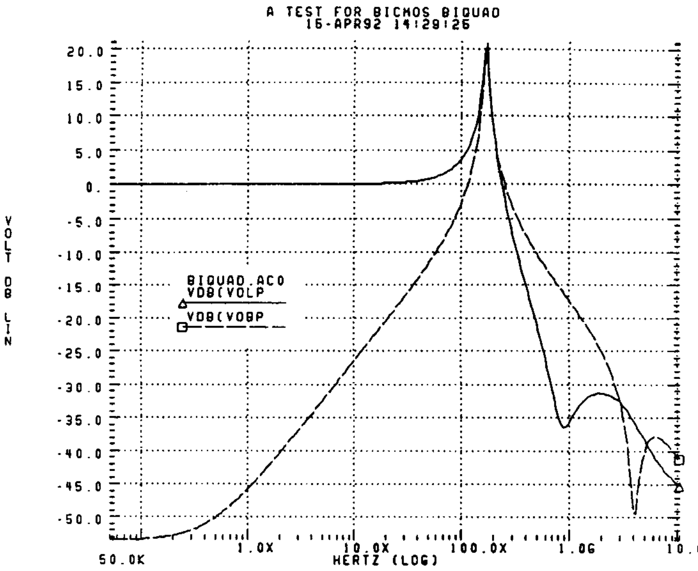

Figure 9: Biquad frequency response showing the effects of feedforward and Q-enhancement.

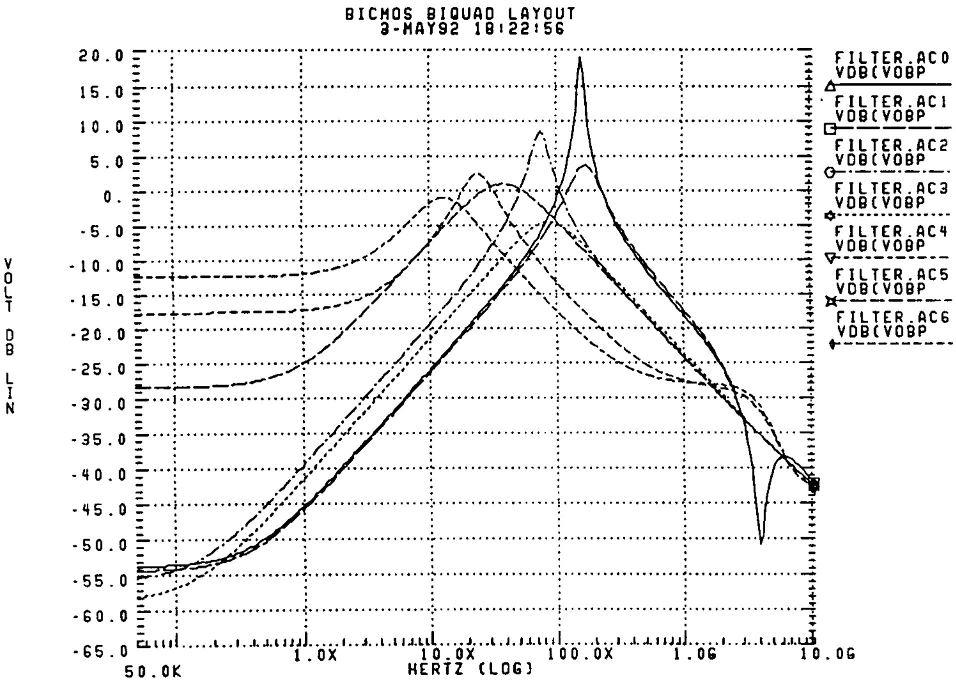

Figure 10: Bandpass function frequency response for various f_o and Q settings.