Voltage Controlled Filter Design

Using Operational Transconductance Amplifiers

Randall L. Geiger

John Ferrell

Department of Electrical Engineering
Texas A&M University
College Station, Texas 77843

ABSTRACT

A set of voltage controlled second-order filter structures is presented. These circuits have convenient external w_o and Q control features. The circuits use a minimum number of components. These second-order blocks can be used to build higher-order filters such as the Butterworth or Chebyschev type in which the cutoff frequency is controlled by a single dc voltage. The circuits also show promise for integrated applications.

INTRODUCTION

The operational transconductance amplifier (OTA) has been used as an active device in conventional active filter applications as well as for the controlling element in a host of voltage controlled filters, amplifiers, oscillators, etc. Although several authors [1-6] have proposed using the OTA in conventional filter applications, the resulting circuits are mostly impractical due to circuit complexity, sensitivity, and difficulty of tuning. Literature on OTA filter design must be considered very immature.

One reason for the complexity of some of the designs that have appeared is due to the attempt of the authors to merely modify existing op amp based structures rather than directly focus upon using the OTA as the active device. Others have attempted to make the filter characteristics relatively independent of the transconductance gain, g_m, of the OTA as was justifiably done in operational amplifier based filter designs.

We address the problem by using the OTA directly as the active device and establishing w_o and Q factors that are directly and inversely proportional to g_m respectively. The transconductance gain of commercially available OTAs, which is directly proportional to the input current of a current mirror, can typically be varied by several decades. The circuits presented thus have a comparable adjustment range for w_o and Q.

A class of lowpass, bandpass, highpass, and notch filters are presented which have independent w_o and Q adjustment via the externally controllable parameter g_m. Some of these configurations are shown to have a minimum number of components. When compared to existing OTA based structures, these circuits offer advantages in circuit simplicity, tunability, and performance. Applications include voltage controlled (e.g. w_o controllable) filters of the Butterworth, Chebyschev and Elliptic types.

Applications can also be found in monolithic active filter design where the tradeoffs between silicon area and effective RC time constants look promising and where control of the filter characteristics via the input current of a single input, multiple output current mirror offers potential for precision designs.

OTA MODEL

The OTA's considered on this paper are differential input single output devices with ideally infinite input and output impedances. The symbol for the device is shown in Fig. 1. The transconductance gain is given by

$$g_m = \frac{I_o}{V^+ - V^-} \tag{1}$$

For commercially available OTA's, such as the CA3080, the gain g_m is proportional to the external dc bias current I_B. This proportionality is typically maintained over several decades. Internally, this dc bias current serves as the input to a current mirror, the secondary current of which actually affects the transconductance gain g_m. Simultaneous adjustments of the g_m of several OTA's with a single dc bias current is practical in monolithic structures by using a single current mirror with multiple secondaries as shown in Fig. 2.

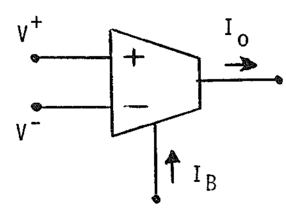

Fig. 1. OTA Model

CH1845-7/83/0000-0594 $1.00 © 1983 IEEE

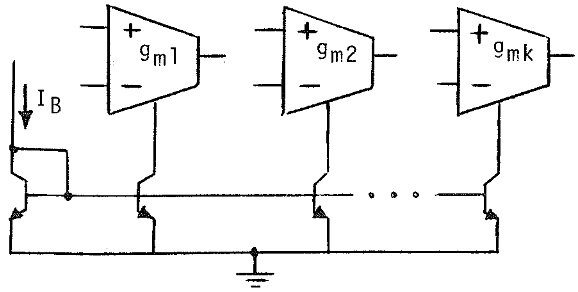

Fig. 2. Control of several OTA's in monolithic structure with single bias current

Filters designed in which the filter characteristics are dependent upon g_m can thus be "current controlled" by the externally adjustable dc bias current. Since standard techniques exist for generating a dc current which is proportional to a dc voltage, and since voltage control applications are more common, these filters are often termed "voltage controlled filters".

From a practical viewpoint, one of the major limitations of most existing commercially available OTA's is a very limited maximum differential input voltage. To minimize distortion, the maximum differential input voltage to these devices is often restricted to the 10-20 mv RMS range for commercial devices such as the CA3080 and LM13600.

g_m CONTROLLED FILTERS

The usefulness of g_m controlled filters is to a large extent dependent upon the relationship between the filter parameters of interest and the gain g_m. A large number of the OTA based filters that have appeared in the literature have <u>very</u> involved relationships between g_m and the common filter characteristics such as bandwidth, w_o, Q, etc. In addition, most of the biquadratic structures that have been presented to date require a large number of passive components. Most are not practical either as a fixed filter or as a voltage controlled device.

w_o and Q controllable circuits with transfer functions of the type listed in Table 1 will find considerable applications. The w_o and Q of these transfer functions are also given.

The usefulness of circuits that have the transfer functions allowing for control of the filter characteristics listed in Table 1 is apparent. Beyond their direct application as second-order filters, their application in higher-order structures is also promising provided all g_m's are simultaneously controllable. For example, cascaded lowpass w_o adjustable biquads can be used for the design of higher-order even-order Butterworth and Chebyschev filters with a voltage controlled cutoff frequency. The bandpass w_o adjustable types can be cascaded to obtain frequency adjustable higher-order even-order bandpass filters of the Butterworth and Chebyschev types. The w_o adjustable schemes also find applications in elliptic filter applications.

A collection of biquads which have transfer functions of the form indicated in Table 1 are shown in Fig. 3 along with their corresponding transfer functions in Table 2. The low component count and simplicity of the transfer function expressions is noteworthy.

First-order blocks which also are controllable should also be considered. For the sake of brevity discussion of these circuits will not be included. Suffice it to say that these circuits have been investigated and that the circuit topologies of the first-order structures are considerably simpler than the second-order circuits presented in this paper.

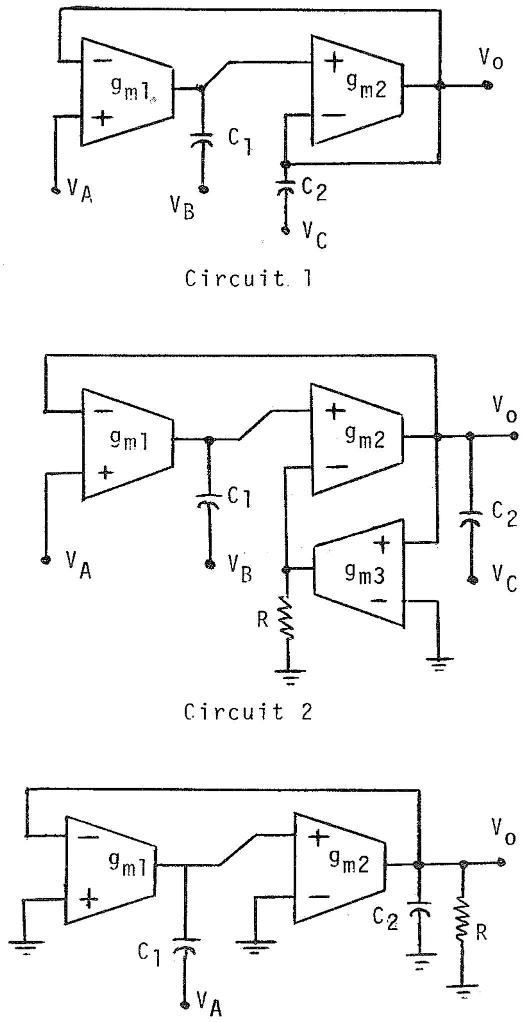

Fig. 3. g_m controlled second-order filters

Lowpass w_o adjustable	$T_{LPW}(s)$	$= \dfrac{\alpha_o g_m^2}{s^2+s\, g_m \beta_1 + g_m^2 \beta_o^2}$	$w_o=\beta_o g_m$	$Q=\dfrac{\beta_o}{\beta_1}$
Bandpass w_o adjustable	$T_{BPW}(s)$	$= \dfrac{\alpha_1 g_m/Q\ s}{s^2+s\, g_m \beta_1 + g_m^2 \beta_o^2}$	$w_o=\beta_o g_m$	$Q=\dfrac{\beta_o}{\beta_1}$
Highpass w_o adjustable	$T_{HPW}(s)$	$= \dfrac{\alpha_2 s^2}{s^2+s\, g_m \beta_1 + g_m^2 \beta_o^2}$	$w_o=\beta_o g_m$	$Q=\dfrac{\beta_o}{\beta_1}$
Notch w_o adjustable	$T_{NW}(s)$	$= 2\ \dfrac{s^2+\alpha_o^2 g_m^2}{s^2+s\, g_m \beta_1 + g_m^2 \beta_o^2}$	$w_{o,num}=g_m\beta_o$ $w_{o,den}=g_m\beta_o$	$Q=\dfrac{\beta_o}{\beta_1}$
Lowpass W_o and Q adjustable	$T_{LPWQ}(s)$	$= \left(\dfrac{\alpha_o g_{mA}^2}{s^2+s\, \beta_1 g_{mA} g_{mB} + g_{mA}^2 \beta_o^2}\right)$	$w_o=g_{mA}\beta_o$	$Q=\dfrac{\beta_o}{\beta_1 g_{mB}}$
Bandpass w_o&Q adjustable	$T_{BPWQ}(s)$	$= \dfrac{\alpha_1 g_{mA}/Q\ \ s}{s^2+s\, \beta_1 g_{mA} g_{mB} + g_{mA}^2 \beta_o^2}$	$w_o=g_{mA}\beta_o$	$Q=\dfrac{\beta_o}{\beta_1 g_{mB}}$
Highpass w_o&Q adjustable	$T_{HPWQ}(s)$	$= \dfrac{\alpha_2 s^2}{s^2+s\, \beta_1 g_{mA} g_{mB} + g_{mA}^2 \beta_o^2}$	$w_o=g_{mA}\beta_o$	$Q=\dfrac{\beta_o}{\beta_1 g_{mB}}$
Notch w_o&Q adjustable	$T_{NWQ}(s)$	$= \dfrac{s^2+\alpha_o^2 g_{mA}^2}{s^2+s\, \beta_1 g_{mA} g_{mB} + g_{mA}^2 \beta_o^2}$	$w_o=g_{mA}\beta_o$	$Q=\dfrac{\beta_o}{\beta_1 g_{mB}}$
Bandpass w_o adjustable const. BW	$T_{BPCBW}(s)$	$= \dfrac{\alpha_1 g_m\ s}{s^2+\beta_1\ s + g_m^2 \beta_o^2}$	$w_o=g_m\beta_o$	$BW=\beta_1$

TABLE 1. USEFUL CONTROLLED TRANSFER FUNCTIONS

Buffering of these stages with a unity gain buffer may be necessary for cascaded applications since the output impedance of the OTA is ideally very large. For experimental evaluation of these structures the input voltage must be kept small to prevent nonlinear distortion. A voltage attenuator (possibly buffered) is often used for this purpose with existing commercially available OTA's. Interstage attenuation in cascaded structures is not required provided the voltage gain of each stage is sufficiently small.

CONCLUSIONS

A set of voltage controlled second-order filter structures has been presented. These circuits are noted for their low component count and design simplicity as well as the practical relationship between g_m and the filter characteristics w_o and Q. Adjustment of w_o over several decades is possible by using commercially available OTA's. These topologies show promise in integrated applications because of the design simplicity, absence of resistors in most cases, and the ability to provide simultaneous adjustment of the g_m of several OTA's with a single bias current through the use of multiple secondary current mirrors.

Circuit # & Type	Input Conditions	Transfer Function	If $g_{m1}=g_{m2}=g_m$: w_o	If $g_{m1}=g_{m2}=g_m$: Q
1. w_o adjustable Lowpass	$V_i = V_A$; V_B and V_C grounded	$\frac{g_{m1}\, g_{m2}}{s^2C_1C_2+sC_1g_{m2}+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
1. w_o adjustable Bandpass	$V_i = V_B$; V_A and V_C grounded	$\frac{sC_1\, g_{m2}}{s^2C_1C_2+sC_1g_{m2}+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
1. w_o adjustable Highpass	$V_i = V_C$; V_A and V_B grounded	$\frac{s^2C_1C_2}{s^2C_1C_2+sC_1g_{m2}+G_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
1. w_o adjustable Notch	$V_i = V_A = V_C$; V_B grounded	$\frac{s^2C_1C_2 + g_{m1}g_{m2}}{s^2C_1C_2+sC_1g_{m2}+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\sqrt{\frac{C_2}{C_1}}$
2. w_o & Q adjust. Lowpass	$V_i = V_A$; V_B and V_C grounded	$\frac{g_{m1}\, g_{m2}}{s^2C_1C_2+sC_1g_{m2}Rg_{m3}+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\frac{1}{Rg_{m3}}\sqrt{\frac{C_2}{C_1}}$
2. w_o & Q adjust. Bandpass	$V_i = V_B$; V_A and V_C grounded	$\frac{s\,C_1\, g_{m2}}{s^2C_1C_2+sC_1g_{m2}Rg_{m3}+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\frac{1}{Rg_{m3}}\sqrt{\frac{C_2}{C_1}}$
2. w_o & Q adjust. Highpass	$V_i = V_C$; V_A and V_B grounded	$\frac{s^2C_1C_2}{s^2C_1C_2+sC_1g_{m2}g_{m3}R+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\frac{1}{Rg_{m3}}\sqrt{\frac{C_2}{C_1}}$
2. w_o & Q adjust. Notch	$V_i = V_A = V_C$; V_B grounded	$\frac{s^2C_1C_2 + g_{m1}g_{m2}}{s^2C_1C_2+sC_1g_{m2}g_{m3}R+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$\frac{1}{Rg_{m3}}\sqrt{\frac{C_2}{C_1}}$
3. Const. BW bandpass w_o adjustable	$V_i = V_A$	$\frac{sC_1g_{m2}}{s^2C_1C_2+s\frac{C_1}{R}+g_{m1}g_{m2}}$	$\frac{g_m}{\sqrt{C_1C_2}}$	$BW = \frac{1}{RC_2}$

Table 2. Gain Expressions for Circuits of Fig. 3

REFERENCES

[1] M. Bialko, W. Sienko and R. Newcomb. "Active Synthesis Using the DVCCS/DVCVS", International Journal of Circuit Theory and Applications, Vol. 2, pp. 23-28, 1974.

[2] F. Atiya, A. Soliman and T. Saadawi, Active RC Bandpass and Lowpass Filters Using the DVCCS/DVCVS", Electronics Letters, Vol. 12, pp. 360-361, July 1976.

[3] M. Bialko and R. Newcomb, "Generation of All Linear Circuits Using the Integrated DVCCS", IEEE Trans. on Circuit Theory, Vol. CT-18, pp. 733-736, 1974.

[4] R. Nandi, "New Ideal Active Inductance and Frequency-Dependent Negative Resistance using DVCCS/DVCVS: Applications in Sinusoidal-Oscillator Realization", Electronics Letters, Vol. 14, pp. 551-553, August 1978.

[5] D. Patranabis and A. Paul, "Floating Ideal Inductor with One DVCCS", Electronics Letters, Vol. 15, pp. 545-546, August 1979.

[6] F. Anday, "On the Analysis and Synthesis of Active Networks Containing DVCCS/DVCVS", Proc. of the IEEE, pp. 375-376, March 1976.