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A Voltage Tunable Active-R Filter*

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ABSTRACT

By using appropriate single-pole amplifiers, realized in any of a number of forms, and the voltage variable DVCCS of Fukahori, a basic building block for tunable filters suitable for VLSI is introduced. This is used to realize voltage tunable active-R filters. A canonical second order filter with all filter coefficients voltage adjustable is presented.

I. Introduction

Previously [1] it has been shown how the technique of active-R filter design of Soderstrand [2] can be improved upon by incorporating non-zero pole locations of the amplifiers. Here we combine the technique of [1] with some improvements on the idea of Fukahori [3] to allow the design of voltage tunable active-R filters.

Fukahori has introduced a very nice circuit for realizing voltage-variable gain in an integrated circuit (IC) voltage controlled current source. This is used by Fukahori to feed an op-amp with capacitive feedback to yield a voltage-variable integrator. Here we replace the capacitive feedback with resistive feedback and use instead the amplifier pole to obtain the filter dynamics. To make the idea practical we must go beyond a simple replacement and introduce a second resistive and adjustable feedback loop to achieve the basic integrator section. This has the advantage that it will allow the construction of second order filter sections where all coefficients can be independently tuned. The technique is similar to that of Brand & Schaumann [4] but differs in that the tuning is advantageously done via Fukahoris' circuit rather that through op-amps.

II. The Basic Components

Figure 1 shows the symbolism we will use for the differential voltage controlled current source (DVCCS) and voltage controlled voltage source (VCVS). The former follows the notation used with Bialko [5] but indicates two other controlling voltages V_T and V_o used to control the bias currents I_T and I_o in the circuit of Fukahori [3, Fig. 5]. In both parts of Fig. 1 we assume the input currents are zero while the output currents have a return to ground (omitted for simplicity of drawing). For the DVCCS, following Fukahori [3, p. 731], we can achieve

$$g_m = \frac{1}{R}\frac{I_T}{I_o} = \frac{1}{R}\frac{R_o}{R_T}\frac{V_T}{V_o} \quad \text{(II-1)}$$

for the control of g_m. Actually Fukahori sets only $I_T = V_T/R_T$, but using the same bias current generating circuit [3, Fig. 4] we can also set $I_o = V_o/R_o$, except that since two identical current sources I_o are used in the DVCCS, [3, Fig. 2], they are both to be controlled by one voltage V_o.

Concerning the VCVS, we shall assume that two types are available. One, the flat-gain type, will be such that $K(s) = -K_o$ for K_o a real constant. The other, the single-pole type, will have

$$K(s) = -\frac{K_o\omega_c}{s+\omega_c} \quad \text{(II-2)}$$

In practice these two types can be obtained over the frequency ranges of interest by various means, for example, in CMOS realizations by the circuits of Soderstrand [2]. Practically we will desire ω_c as small as feasible, which initiates a new kind of IC amplifier [6].

The basic integrating amplifier we use here is shown in Fig. 2. In Fig. 2 we will vary V_{Ta} for desired filter characteristics and V_{Tb} to obtain good integration. This latter is achieved by noting that for structural and Lyapunov stability we desire the integrator pole slightly in the left half plane. Thus in

$$T(s) = \frac{v_o}{v_+-v_-} = \frac{(K_o\omega_c)(R_a g_{ma})(R_b g_{mb})}{s+\omega_c[1+K_o(1-R_b g_{mb})]} \quad \text{(II-3)}$$

we set

$$\epsilon = 1 + K_o(1-R_b g_{mb}) \geq 0 \quad \text{(II-4)}$$

small; i.e. we choose, by adjusting V_{Tb},

$$g_{m_b}(V_{T_b}) = \frac{1}{R_b}\cdot\left(\frac{K_o+1-\epsilon}{K_o}\right) \quad \text{(II-5)}$$

which gives

$$T(s) = \frac{(K_o+1-\epsilon)\omega_c(R_a g_{ma})}{s+\epsilon\,\omega_c} \quad \text{(II-6)}$$

Since T(s) varies linearly with g_{ma} and the latter varies linearly with V_{Ta}, $T(s)$ varies linearly with V_{Ta}.

Since filters need addition and subtraction, Fig. 3 shows some examples of how these can be attained with the circuit elements on hand using the flat-gain amplifier.

III. The Filter Section

The filter configuration is that of Fig. 4. This uses the state-variable configuration of Fukahori [3, Fig. 6] to form the denominatior and the components of Fig. 3 to form an arbitrary (degree two) numerator. However, rather than use capacitive feedback to form the integrator, we use the active-R circuit of Fig. 2 for which we will take into account that the pole is not at zero. Thus we set, for eq. (II-6)

$$T(s) = \underline{T}(p) = \frac{k}{p} \qquad \text{(III-1a)}$$

$$p = s+\alpha,\ \alpha = \epsilon\omega_c,\ k = (K_o+1-\epsilon)\omega_c(R_a g_{ma}) \qquad \text{(III-1b)}$$

where we note that the gain k varies with V_{Ta} but not cut-off frequency α (which, however, is set by V_{Tb}).

Given any degree two transfer function

$$H(s) = \frac{v_o}{v_i} = \frac{b_2 s^2 + b_1 s + b_o}{s^2 + a_1 s + a_o} \qquad \text{(III-2)}$$

we predistort it via $s=p-\alpha$ to get

$$\underline{H}(p) = H(s) = \frac{\underline{b}_2 p^2 + \underline{b}_1 p + \underline{b}_o}{p^2 + \underline{a}_1 p + \underline{a}_o} \qquad \text{(III-3a)}$$

$$= \frac{b_2 p^2 + [b_1 - 2\alpha b_2]p + [b_o - \alpha b_1 + \alpha^2 b_2]}{p^2 + [a_1 - 2\alpha]p + [a_o - \alpha a_1 + \alpha^2]} \qquad \text{(III-3b)}$$

The philosophy is then to analyze Fig. 4 in p and equate terms of the result with (III-3b) in order to realize any given $H(s)$, (III-2). Any variable coefficents in $H(s)$ will become variable coefficients in (III-3b) and hence in the realized circuit. Some realizability constraints would be met along the way, the most interesting being that $\underline{a}_1$ & $\underline{a}_o$ should be positive (which can often be achieved by V_{Tb} to give small enough α). Classically, one wishes to vary the denominotor $\omega_o = \sqrt{a_o}$ without affecting its $Q = \sqrt{a_o}/a_1$, which we shall see is possible.

Setting

$$T_1 = k_1/p,\ T_2 = k_2/p,\ G_1 = 1/R_1,\ G_2 = 1/R_2 \qquad \text{(III-4a)}$$

$$\Delta(p) = \frac{k_1 k_2}{p^2} + c_1 \frac{k_1}{p} + 1 \qquad \text{(III-4b)}$$

$$c_1 = \frac{G_1}{G_1+G_2},\ c_2 = \frac{G_2}{G_1+G_2} \qquad \text{(III-4c)}$$

gives, on analyzing Fig. 4,

$$\frac{v_1}{v_i} = \frac{c_2\left(\frac{k_1 k_2}{P_2} + 1\right)}{\Delta},\quad \frac{v_2}{v_i} = -c_2 \frac{k_1/p}{\Delta}$$

$$\frac{v_3}{v_i} = c_2 \frac{k_1 k_2/p^2}{\Delta} \qquad \text{(III-5a)}$$

$$\frac{v_1 - v_3}{v_i} = \frac{c_2}{\Delta} \qquad \text{(III-5b)}$$

Therefore

$$\frac{v_1 - v_3}{v_i} = \frac{c_2 p^2}{p^2 + c_1 k_1 p + k_1 k_2},\quad \frac{v_2}{v_i} = \frac{-c_2 k_1 p}{p^2 + c_1 k_1 p + k_1 k_2},$$

$$\frac{v_3}{v_i} = \frac{c_2 k_1 k_2}{p^2 + c_1 k_1 p + k_1 k_2} \qquad \text{(III-6)}$$

Thus the output of Fig. 4 is

$$\frac{v_o}{vi} = \frac{\frac{K_o}{1+K_o} c_2 R_f [g_{m\gamma} p^2 - k_1 g_{m\beta} p + k_1 k_2 g_{m\alpha}]}{p^2 + c_1 k_1 p + k_1 k_2} \qquad \text{(III-7)}$$

Equating coefficients in (III-7) with (III-3b)

$$\underline{a}_1 = a_1 - 2\alpha = c_1 k_1 \qquad \text{(III-8a)}$$

$$\underline{a}_o = a_o - \alpha a_1 + \alpha^2 = k_1 k_2 \qquad \text{(III-8b)}$$

$$\underline{b}_2 = b_2 = \frac{K_o}{1+K_o} c_2 R_f g_{m\gamma} \qquad \text{(III-8c)}$$

$$\underline{b}_1 = b_1 - 2\alpha b_2 = -\frac{K_o}{1+K_o} c_2 R_f k_1 g_{m\beta} \qquad \text{(III-8d)}$$

$$\underline{b}_o = b_o - \alpha b_1 + \alpha^2 b_2 = \frac{K_o}{1+K_o} c_2 R_f k_1 k_2 g_{m\alpha} \qquad \text{(III-8e)}$$

From this we see that Fig. 4 as drawn is for $b_2>0$, $b_1<0$ and $b_o>0$; other signs can be achieved by reversing input leads in DVCCS's.

Therefore, given a_o, a_1, b_o, b_1, b_2 we choose $\alpha = \epsilon\omega_c$, by adjustment of V_{Tb} in T_1 & T_2, such that equations (III-8) can be met (perhaps also by sign changes in the g_m's). Then we solve for k_1, k_2, g_{m_α}, g_{m_β}, g_{m_γ}, eqs. (III-9) giving a unique solution. If any of a_o, a_1, b_o-b_2, vary, then variations in k_1-g_{m_γ} will be known and can be achieved by varying the appropriate V_T's & V_o's. For reference we have

$$k_1 = (a_1-2\alpha)/c_1 \qquad \text{(III-9a)}$$

$$k_2 = (a_o-\alpha a_1+\alpha^2)/k_1 \qquad \text{(III-9b)}$$

$$g_{m_\gamma} = [(1+K_o)b_2]/[K_o c_2 R_f] \qquad \text{(III-9d)}$$

$$g_{m_\beta} = -[(1+K_o)(b_1-2\alpha b_2)]/[K_o c_2 R_f k_1]$$

$$g_{m_\alpha} = [(1+K_o)(b_o-\alpha b_1+\alpha^2 b_2)]/[K_o c_2 R_f k_1 k_2] \qquad \text{(III-9e)}$$

If complicated adjustments are desired these could be microprocessor controlled.

IV. Discussion

Here we have presented an active-R second order filter based upon the voltage variable DVCCS of Fukahori. Because this latter is designed for IC fabrication, the filter structure becomes suitable for VLSI when suitable integrators and flat-gain amplifiers are available. The latter already appears in Fukahori and the former is under development. Thus the structure allows for on chip microprocessor control of filter characteristics.

Because, for stability purposes, ideal active-R integrators are not preferable to ones with nonzero poles, the theory has been set up to take into account nonzero poles of the integrators through predistortion. But the technique gives an added degree of flexibility in that the integrator pole position can electronically be adjusted.

Of course higher order and matrix-valued filter characteristics can be obtained by factorization of transfer functions and matrices into degree two (or one) portions [7] in which case the section presented here can serve as a universal building block for on-chip microprocessor controlled filters.

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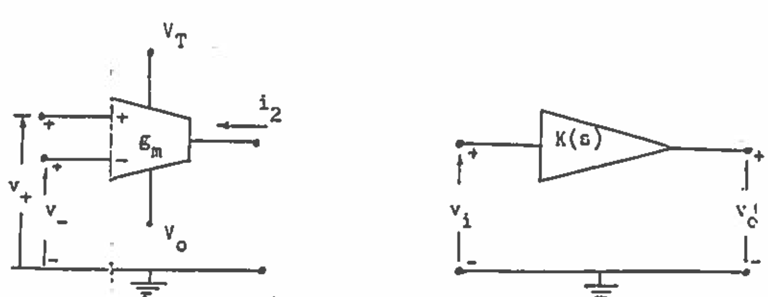


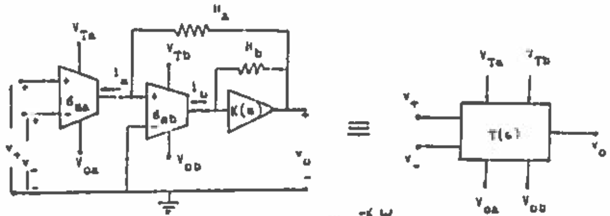
$i_2 = g_m(v_+-v_-)$, $g_m = \frac{1}{R}\frac{R_o V_T}{R_T V_o}$

a) DVCCS

$v_o = K(s)v_i$

b) VCVS

Figure 1



$$K(s) = \frac{-K_o \omega_c}{s + \omega_c}$$

$$\frac{v_o}{v_+ - v_-} = T(s) = \frac{(K_o \omega_c)(R_a g_{ma})(R_b g_{mb})}{s + \omega_c[1 + K_o(1 - R_b g_{mb})]}$$

Figure 2

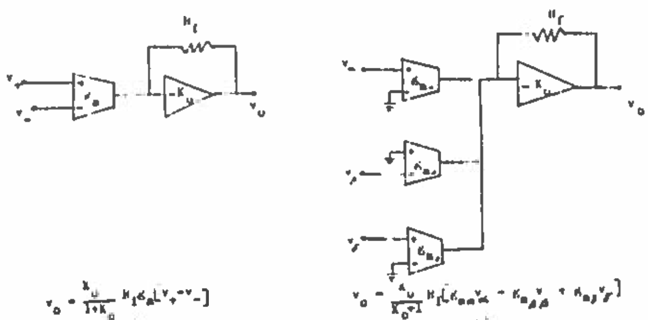


Figure 3

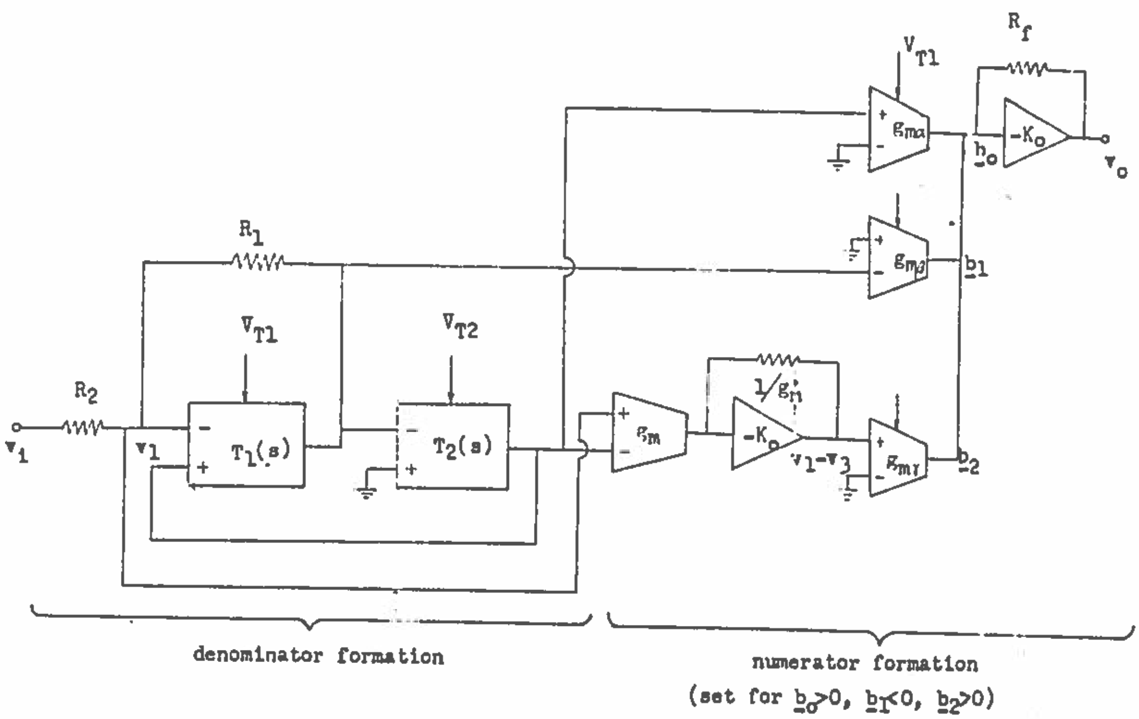


Figure 4

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