HIGH FREQUENCY, EXTENDED CMOS ACTIVE - R FILTERS*

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Abstract

By using a more exact integrating amplifier model Soderstrand's technique for the design of CMOS - active R filters is reinterpreted. The alternative designs obtained extend the performance of the realized filters through predistortion to accommodate inexact integration of the amplifiers used.

1. INTRODUCTION

There has been considerable interest in active filters for the obvious reason of circuit miniturization since the advent of solid-state devices [6,7]. And due to the relative ease of construction through the technology of MOS integrated circuits, active filters using the CMOS transistor pair have also attracted some attention in recent times. In fact Soderstrand has shown a fascinating and potentially very useful means of realizing active filters incorporating only resistors, CMOS devices, and bias sources [1, 2]. In doing this he approximates the CMOS pair voltage - gain by that of an integrator. Thus, although valid in a certain band of frequencies, the designed and actual performances do deviate.

Here using a previous technique of frequency variable transformation [3], the method of Soderstrand is extended to all frequencies for which a first order transfer function with a nonzero pole adequately describes the amplifier performance. That is, this paper deals with a more comprehensive approach to the design of CMOS - resistor building blocks which can be used for the construction of filter transfer functions. At the end an example is given for the design of a voltage transfer function which illustrates the improvement in circuit performance that can be obtained by using the design method of this paper.

2. IMPROVED CMOS-R FILTER REALIZATION

The small signal voltage gain of the unloaded CMOS transistor pair of Fig. 1(a) has been shown to be given by [1,2]

$$\frac{V_2}{V_1}(s) = A(s) = -GB/(s+\alpha) \qquad (1a)$$

where

$$\alpha = 2\pi f_c = \text{(radian) cut-off frequency} \qquad (1a_1)$$

$$-GB = A(0)\alpha = \text{Gain-bandwidth product} \qquad (1a_2)$$

$$s = \sigma + j\omega = \text{true complex frequency} \qquad (1a_3)$$

In this the cut-off frequency α is dependent upon the bias voltage V, of Fig. 1(a), as is GB, both increasing with V. Given such a CMOS pair, the parameters GB and α are readily determined experimentally, α being given by the 3db radian frequency and GB being the 0db radian frequency intercept. Also shown in Fig. 1(b) is Soderstrand's unity gain amplifier, which is simply the amplifier of Fig. 1(a) operated with bias voltage V large enough to move the cut-off frequency α above the range of interest (practically for experimentation V = 7.5v is used) and loaded in a conductance g_m to give unity gain, A(0) = -1.

In his studies Soderstrand [1] approximates the CMOS pair gain by

$$A(s) \approx -GB/s \qquad (1b)$$

consequently introducing this approximation into the design. However, this approximation can be completely avoided, while still making use of all other of Soderstrand's results, by transforming to

* This research was supported in part through NSF Grant ENG 75-03227.

a new variable

$$p = s + \alpha \tag{2}$$

for which the unloaded CMOS amplifier gain more exactly becomes

$$A(s) = \underline{A}(p) = -GB/p \tag{3}$$

In this, the p-plane is a shifted s-plane and "real" frequencies $s=j\omega$ show up as $p=\alpha+j\omega$, that is, translated α units to the left.

Now given a voltage transfer function

$$\frac{V_o}{V_i}(s) = T(s) = \frac{b_o + b_1 s + b_2 s^2}{a_o + a_1 s + a_2 s^2} \tag{4}$$

which is desired to be realized in the true frequency s, we can predistort by shifting the critical frequencies α units to the right, placing considerations in the p-plane, and then transforming back to the true frequency s-plane via (2). Thus

$$T(s) = \underline{T}(p) = \frac{[b_o+\alpha^2-\alpha b_1]+[b_1-2\alpha b_2]p+b_2p^2}{[a_o+\alpha^2-\alpha a_1]+[a_1-2\alpha a_2]p+a_2p^2} \tag{5a}$$

$$= \frac{\underline{b}_o + \underline{b}_1 p + \underline{b}_2 p^2}{\underline{a}_o + \underline{a}_1 p + \underline{a}_2 p^2} \tag{5b}$$

Because the amplifier gain in p, $\underline{A}(p) = -GB/p$, agrees in form with the amplifier gain, $A(s) \approx -GB/s$, used by Soderstrand, (5b) can be synthesized in the p-plane by Soderstrand's circuits. For example, Fig. 2 shows the configuration in the case that $b_1 \geq 0$. Soderstrand's formulas, and analysis of the circuit give

$$\underline{a}_o = (GB_1 x GB_2)/R_3 \tag{6a}$$

$$\underline{a}_1 = GB_1/R_2 \tag{6b}$$

$$\underline{a}_2 = (1/R_1) + (1/R_2) + (1/R_3) \tag{6c}$$

$$\underline{b}_o = (GB_1 x GB_2/R_1) x (R_7/[R_4+R_7]) \tag{6d}$$

$$\underline{b}_1 = (GB_1/R_1) x (R_7/[R_5+R_7]) \tag{6e}$$

$$\underline{b}_2 = (1/R_1) x (R_7/[R_6+R_7]) \tag{6f}$$

R_7 would usually be chosen to be much less than R_4, R_5 and R_6. A numerical design using this technique is carried out in the next section.

It is of course of interest to know the error caused by the approximation of (1b), that is $A(s) \approx -GB/s$. Thus, let us assume that we use $A(s) \equiv -GB/s$ in synthesizing the desired voltage transfer function $T(s)$, (4). Consequently, we are assuming in the synthesis $\alpha=0$ in which case synthesis occurs from (5) with $\underline{a}_i = a_i$, $\underline{b}_i = b_i$. Then use of real amplifiers having $A(s) = -GB/(s+\alpha)$, all with identical cut-off frequencies α, will shift the critical frequencies α units to the left giving the actually realized transfer function, $T_{real}(s)$, as [replace α by $-\alpha$ in (5a)]

$$T_{real}(s) = \frac{[b_o+\alpha^2+\alpha b_1]+[b_1+2\alpha b_2]s+b_2s^2}{[a_o+\alpha^2+\alpha a_1]+[a_1+2\alpha a_2]s+a_2s^2} \tag{7}$$

Comparison of the desired $T(s)$, (4), with the one actually realized by Soderstrand, $T_{real}(s)$ of (7), exhibits the error introduced by the approximation $s \approx s + \alpha$. Numerically, this is carried out in the design of the next section.

3. EXAMPLE

Experimental measurement, and a check by the SPICE routine, gives for the unloaded RCA type CA3600E CMOS pair biased at V = 2.5v

$$GB = 25.43 x 10^6 \text{rad/sec} \tag{8a}$$

$$f_c = 1.1 x 10^6 \text{Hz}, \ \alpha = 2\pi f_c \tag{8b}$$

while at V = 7.5v

$$1/g_m = 162\Omega \tag{8c}$$

Let it be desired to synthesize a bandpass filter with denominator specified by Q and $\omega_o = 2\pi f_o$. Thus, we desire, using (5),

$$T(s) = (\omega_o s)/[\omega_o^2 + (\omega_o/Q)s + s^2] \tag{9a}$$

$$= \underline{T}(p) = \frac{-(2\pi)^2(f_of_c) + (2\pi f_o)p}{(2\pi)^2[f_o^2+f_c^2-(f_of_c/Q)]+(2\pi)[(f_o/Q)-2f_c]p+p^2} \tag{9b}$$

For $Q = 2^{-\frac{1}{2}}$ (maximally flat denominator), f_o=2MHz, and the above amplifier, (9a) becomes

$$T(s) = \frac{4\pi x 10^6 s}{4(2\pi)^2 x 10^{12} + 4\,2\pi x 10^6 s + s^2} \tag{10}$$

and from (9b) we obtain

$$\underline{T}(p) = \frac{-(2.2)(2\pi)^2 x 10^{12} + 4\pi x 10^6 p}{(2\pi)^2[4+(1.1)^2-\sqrt{2}(2.2)] x 10^{12} + (4\pi)(\sqrt{2}-1.1) x 10^6 p + p^2} \tag{11}$$

Since $\underline{b}_2$ in the expression for $\underline{T}(p)$ is zero, for this example R_6 in Fig. 2 is infinite, that is, only R_4 and R_5 are connected to the summer. Applying (6a) - (6f) we determine the values of the resistances as

$$R_1 = 1.4, \ R_2 = 6.4, \ R_3 = 7.8 \tag{12a}$$

$$R_4/R_7 = 5.3, \ R_5/R_7 = 1.4 \tag{12b}$$

which on impedance scaling by multiplying by 10^3 give values in kilohms.

Using Soderstrand's approximation

$$T_{real}(s) = \frac{(2.2)(2\pi)^2 x10^{12} + 4\pi x10^6 s}{(2\pi)^2[4+(1.1)^2+ 2(2.2)]x10^{12}+(4\pi)(\sqrt{2}+1.1)x10^6 s+s^2} \quad (13)$$

Figure 3 shows the magnitude-frequency responses of both the desired transfer function T(s), realized by the method of this paper, and the transfer function $T_{real}(s)$, realized by Soderstrand's method. As can be seen, the curves agree from 5MHz up. But below 5MHz the curves differ considerably showing that the assumption of an ideal integrator introduces large distortion in ranges below about 5ω.

4. DISCUSSION

By making use of a more exact model for a CMOS integrating amplifier, we have shown how Soderstrand's technique for the design of CMOS active-R filters can be improved upon. Within frequency ranges which may be of interest, the resulting design gives a better performance in terms of its actual frequency response as Fig. 3 shows.

It is of interest to note that for a given transfer function to be realized, (5a) indicates the condition under which one or more of the coefficients of the denominator of the transformed transfer function will be positive and hence realizable by a stable predistorted circuit. This means that to ensure stability of the designed network, an integrating amplifier with appropriate cut-off frequency must be used. Compatible cut-off frequencies may be sought by varying the bias voltage. Of course the limit of this voltage "tuning" sets the limit for the practical applicability of the CMOS integrating amplifier in such a design.

The circuit configuration given for the second degree filter realization, Fig. 2, is for the case where the b_i of the transformed transfer function (5b) are positive. If the b_i are less than zero, the circuit changes only through the addition of a unity gain amplifier, Fig. 1(b), where appropriate, to provide for the negative sign.

If it is desired to realize a higher degree transfer function, the transfer function can be factored into degree two, or one, portions, the realizations of which can be put in cascade. Alternatively, any set of state - variable equations can be realized by working in the p-plane with the circuit of Fig. 1 (a) acting as a p-plane integrator. In the general state - variable realization case one does need to watch the practical resistive loading effect of the summers, but with large resistors possibly obtained by impedance scaling this is not of real concern for the cascade structures. Of course degree one functions are readily handled separately, though for theoretical purposes they are just special cases of degree two functions with some zero coefficients.

5. REFERENCES

1. M. A. Soderstrand, "An Improved CMOS Active-R Filter," Proc. IEEE, August 1977, pp. 1204 - 1206.
2. M. A. Soderstrand, "A Simple Synthesis Technique for Active-R Filters," Proc. IEEE Int. Symp. Circuits and Systems, Boston, MA, 1975, pp. 261 - 264.
3. B. D. Berman and R. W. Newcomb, "Transistor - Resistor Synthesis of Voltage Transfer Functions," IEEE Transactions on Circuit Theory, Vol. CT - 20, No. 5, Sept. 1973, pp. 691 - 593.
4. K. R. Bao and S. Srinisavan, "Low Sensitivity Active Filters Using the Operational Amplifier Pole," IEEE Trans. Circuits & Systems, Vol. CAS - 21, No. 2, March 1974, pp. 260 262.
5. R. Schaumann, "Active Filters Without External Capacitors," IEEE Transactions Circuits & Systems, Vol. CAS - 22, Jan. 1975, pp. 39 - 44.
6. R. W. Newcomb, "Active Integrated Circuit Synthesis," Prentice-Hall, Englewood Cliffs, New Jersey, 1968.
7. S. K. Mitra, "Analysis and Synthesis of Linear Active Networks," John Wiley, New York, 1969.

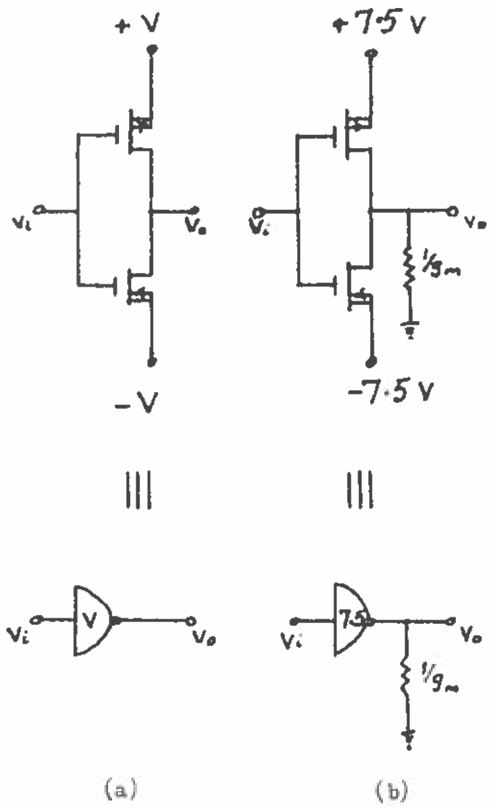

Figure 1. Basic CMOS Active-R Building Blocks
(a) Inverting integrator - circuit and symbol
(b) Inverting unity gain amplifier - circuit and symbol

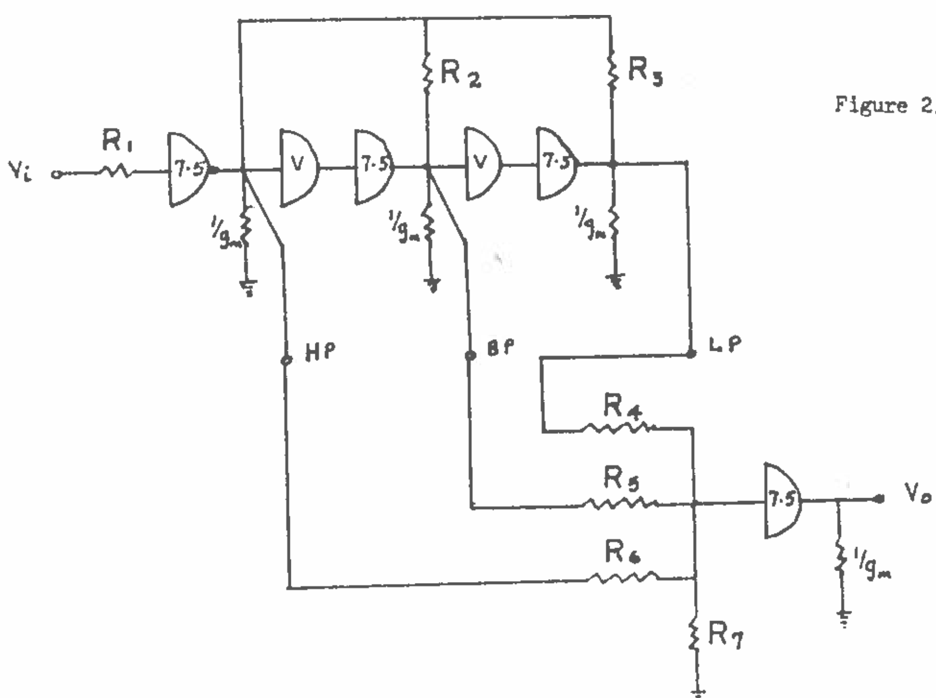

Figure 2. Second Degree Filter Realization Using CMOS Transistor Pair

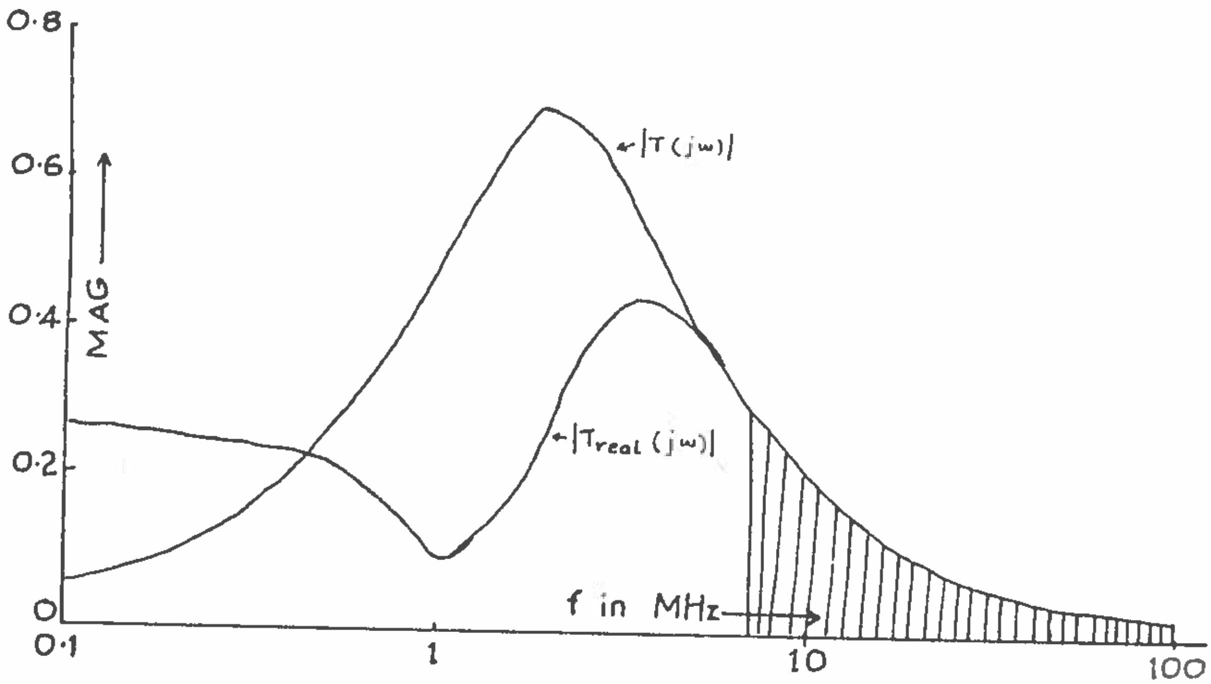

Figure 3. Magnitude-Frequency Response of Both the Desired Transfer Function T(s) and Soderstrand - Realized Transfer Function $T_{real}(s)$. Shaded region indicates area of invalidity of curves due to unity gain amplifier also becoming an integrator.

6. BIOGRAPHIES

Omotayo A. Seriki was born in Epe, Lagos State, Nigeria on the 10th of January, 1934. He received the B.Sc.(hons) degree in Electrical Engineering from the University of Manchester, England, and the Dr.-Ing. degree from the Technical University of Munich, West Germany, in 1964. At present he is Senior Lecturer in the Electrical Engineering Department, University of Lagos, Lagos, Nigeria. He was Minister of Public Works, Lagos State, and his research interests lie in nonlinear networks and systems and communication theory and circuits.

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Proceedings of the

Twenty-first MIDWEST SYMPOSIUM ON CIRCUITS AND SYSTEMS

August 14-15, 1978

Edited by
H. W. Hale/A. N. Michel

Iowa State University, Ames