distortion is less than 2 percent in the frequency range from 900 Hz to 3 kHz.

Fig. 3 shows the waveform at f_0 = 1 kHz. It was noticed that increasing C_1 increases the distortion, and at the same time allows faster growing up of oscillations.

It is noted that due to the fact that the operational amplifier is not ideal, the oscillator requires sophisticated AGC to maintain reasonable distortion levels and amplitude control. Otherwise, amplitude limiting results from amplifier saturation.

ACKNOWLEDGMENT

The authors would like to thank the reviewer for his useful comments.

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Active-*R* Resonator Realization

AHMED M. SOLIMAN AND MAHMOUD FAWZY

***Abstract*—A technique for simulating a series resonator is presented. Then, using only resistors and two internally compensated operational amplifiers, three circuits for realizing an active-*R* series resonator are introduced.**

INTRODUCTION

Active-*R* design using the 6 dB/octave rolloff characteristics of operational amplifiers has recently received considerable attention. The active-*R* synthesis of an impednace is interestingly a new subject which was attacked earlier by Allen and Means [1] who dealt with inductor simulation.

Here, a general procedure for simulating a driving-point impedance is presented. The synthesis of a specific impedance function imposes some restrictions on the network used for the realization.

Consequently, the realization of a series resonator employing only resistors and two operational amplifiers is illustrated by three circuit examples. Each of the proposed circuits resembles a series resonator containing a capacitor, a resistor, and a FDNR element which, according to Bruton [2] transformation, is equivalent to an *LCR* series resonance circuit.

DCR RESONATOR REALIZATION

From Fig. 1(a)

$$Z_{\text{in}} = \frac{V_1}{I_1} \tag{1}$$

$$I_1 = \frac{V_1 - V_2}{R_0} \tag{2}$$

$$\frac{V_2}{V_1} = T(s). \tag{3}$$

From the above equations, it is seen that

$$Z_{\text{in}} = \frac{R_0}{1 - T(s)}. \tag{4}$$

By taking $T(s)$ of the form

$$T(s) = \frac{a_1 s + a_2}{b_0 s^2 + b_1 s + b_2} \tag{5}$$

with

$$b_1 = a_1, b_2 = a_2 \tag{6}$$

Manuscript recieved May 23, 1977; revised October 18, 1977.
The authors are with the Electronics and Communications Engineering Department, Cairo University, Giza, Egypt.

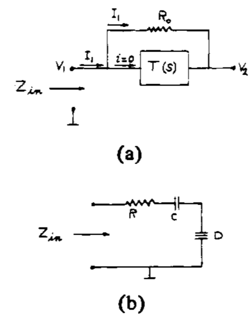

Fig. 1. (a) Basic block for generating a driving point impedance. (b) Series resonator equivalent circuit.

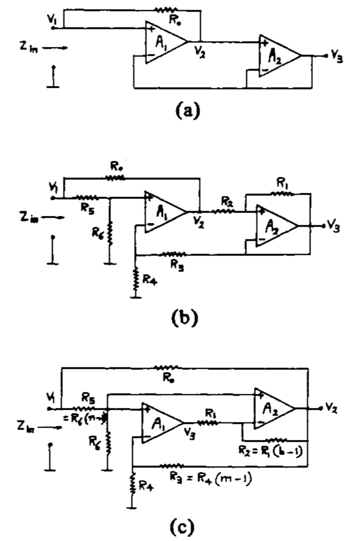

Fig. 2. Active-R series resonators. (a) Circuit 1. (b) Circuit 2. (c) Circuit 3.

and by applying (4), it follows that

$$Z_{\text{in}} = R + \frac{1}{Cs} + \frac{1}{Ds^2} \tag{7}$$

which represents a series *DCR* resonator whose equivalent circuit is shown in Fig. 1(b) with

$$R = R_0, C = \frac{b_0}{a_1 \cdot R_0}, D = \frac{b_0}{a_2 \cdot R_0}. \tag{8}$$

Equation (7) is that of Bruton [2]. With the aid of the equivalent *RLC* series resonance circuit, it can be proved that

$$W_0 = \frac{1}{\sqrt{RD}} \qquad Q = C\sqrt{\frac{R}{D}}. \tag{9}$$

ACTIVE-*R* SERIES RESONATORS

Let

$$A_i = \frac{\text{GB}_i}{s} = \frac{2\pi f_{ci}}{s}, \qquad i = 1, 2 \tag{10}$$

where GB is the gain-bandwidth product, or the unity gain crossover radian frequency of the operational amplifier.

By applying the previous equations on the circuits of Fig. 2 the results shown in Table I are obtained. A quick survey of the circuits under consideration reveals that:

1) It is too difficult for circuit 1 to guarantee $A_1 = A_2$ unless both operational amplifiers are integrated on the same chip. Also, for a given GB it is possible to separately specify only one performance factor, namely *R* or *C* or *D* while the other two factors are dependent.

TABLE I

Circuit No.	1	2	3
Circuit Parameters		$\alpha = R_6/(R_5 + R_6)$ $\gamma = R_4/(R_3 + R_4)$ $b = 1 + R_2/R_2$	$b = 1 + a$ $a = R_2/R_1$
Realizability Conditions	$A_1 = A_2$	$\alpha = \gamma$ $\alpha b \dfrac{GB_1}{GB_2} = 1$ $R_5, R_6 >>$	$m = n = b = K$ $R_5, R_6 >>$
R	R_0	R_0	R_0
C	$\dfrac{1}{R_0 \cdot GB}$	$\dfrac{b}{R_0 \cdot GB_2}$	$\dfrac{K}{R_0 \cdot GB_2}$
D	$\dfrac{1}{R_0 \cdot (GB)^2}$	$\dfrac{b}{\gamma \cdot R_0 \cdot GB_1 \cdot GB_2}$	$\dfrac{K^2}{(K-1)R_0 \cdot GB_1 \cdot GB_2}$
f_0	f_c	$\dfrac{f_{c2}}{b}$	$\dfrac{\sqrt{(K-1) f_{c1} \cdot f_{c2}}}{K}$
Q	1	1	$\sqrt{(K-1)\dfrac{f_{c1}}{f_{c2}}}$

2) It is to be noted that the resistors R_5 and R_6 in both cases of the circuits 2 and 3 should be chosen as large as possible. This enables us to consider that the grounded resistance shunting the series resonator as having an infinite value, which is a reasonable approximation.

3) As for circuit 2, it is has more advantages than circuit 1, in that it avoids the necessity of having identical operational amplifiers. Also, two performance factors out of three can be separately specified. The two circuits are only similar in having $Q = 1$.

4) Concerning the third circuit, it is possible to get $Q > 1$ which is an advantage over the previous two circuits. Since $f_0 \cdot Q \simeq f_{c_1}$, then either f_0 or Q is dependent on the other. For amplifiers having f_c = 1 MHz, and for a practically reasonable maximum resistor ratio K of 400, Q can reach a medium value of 20 with f_0 = 50 kHZ. As Q increases, the experimental results deviate from the theoretical ones.

Conclusions

A method for *DCR* series resonator realization is suggested which depends on a resistor R_0 and a specified block. The block represents an active network which incorporates only resistors and two internally compensated operational amplifiers. Using this method, three resonator realizations are presented.

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A Maximally Flat Group Delay Recursive Digital Filter with Chebyshev Stopband Attenuation

P. THAJCHAYAPONG AND P. LOMTONG

Abstract—**Due to the monotonically decaying nature of its magnitude response, an all-pole recursive digital filter, with its group delay approximating a prescribed constant value, does not exhibit a sharp transition band. Using transformation as the mathematical means, this correspondence describes a design technique for a recursive digital filter with a maximally flat group delay and a Chebyshev attenuation stopband.**

Manuscript received August 4, 1977; revised August 31, 1977.

The authors are with the Faculty of Engineering, King Mongkut's Institute of Technology, Ladkrabang Campus, Ladkrabang, Bangkok, Thailand.

I. Introduction

Due to the monotonically decaying nature of its magnitude response, an all-pole recursive digital filter, with its group delay approximating a prescribed constant value, does not exhibit a sharp transition band [1]-[3]. Recently a numerical technique has been used to assign a mirror-imaged numerator polynomial to the filter transfer function to improve the magnitude selectivity [4]. By observing the transformation technique in analogue design [5], this correspondence describes a design procedure for a recursive digital filter with a maximally flat group delay in the passband, and a Chebyshev stopband attenuation. The design will give all zeros on the unit circle. Hence, all these zeros are dedicated to attenuate the stopband magnitude while simultaneously satisfying the constant-group delay specifications.

II. Design Procedure

i) The transfer function of the required filter is defined to be

$$H(z^{-1}) = \frac{\prod_{i=1}^{m} (1 - 2\cos\theta_i z^{-1} + z^{-2})}{\sum_{i=0}^{n} a_i z^{-i}} \tag{1a}$$

for even n, and $m = n/2$, or

$$H(z^{-1}) = \frac{(1 + z^{-1}) \prod_{i=1}^{m} (1 - 2\cos\theta_i z^{-1} + z^{-2})}{\sum_{i=0}^{n} a_i z^{-i}} \tag{1b}$$

for odd n, and $m = (n-1)/2$. a_i's and θ_i's are to be determined so the filter gives a maximally flat group delay in the passband, and a Chebyshev stopband attenuation. Then, by the transformation $z = (1+s)/(1-s)$, $H(z^{-1})$ can be written as $H(s)$ in the s-plane where

$$H(s) = H(z^{-1})\big|_{z=(1+s)/(1-s)} = \frac{\prod_{i=1}^{m} (s^2 + \omega_i^2)}{\sum_{i=1}^{n} b_i s^i}. \tag{2}$$