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Active RC Applications of the Operational Transresistance Amplifier

Aktive RC-Schaltungen mit Operationsverstärkern als stromgesteuerte Spannungsquelle

Abstract
This paper presents several new active RC circuits employing the Operational Transresistance Amplifier (OTRA) as the active element. Applications of the OTRA in realizing a differential voltage amplifier, a differential integrator, continuous time filters and a quadrature oscillator are presented. The effectiveness of the proposed circuits is demonstrated using PSpice simulations.

Übersicht
Die Arbeit stellt einige neue aktive RC-Schaltungen vor, in denen Operationsverstärker mit stromgesteuerten Spannungsquellen (OTRA) als aktive Elemente zum Einsatz kommen. Anwendungen des OTRA als differentieller Spannungsverstärker, differentieller Integrator, biquadratische Filer und Quadratur-Oszillator werden gezeigt. Der Nutzen der vorgeschlagenen Schaltungen wird mit PSpice-Simulationen nachgewiesen.

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Für die Dokumentation
Operationsverstärker als stromgesteuerte Spannungsquellen / aktive Schaltungen

1. Introduction

The growing demand for mobile communications has led to high level of chip integration and directed research towards the field of low voltage and high frequency applications [1]. Recently new analog building blocks operating from low voltage supplies have been developed to overcome the finite gain-bandwidth product associated with traditional operational amplifiers (op amps) [2-5].

Although the Operational Transresistance Amplifier (OTRA) is commercially available from several manufacturers under the name current differencing amplifier [6], it has not gained attention until recently. These commercial realizations do not provide virtual ground at the input terminals and they allow the input current to flow in one direction only [7, 8]. The former disadvantage limited the functionality of the OTRA, whereas the later forced designers to use external DC bias circuits leading to complex and unattractive designs.

The OTRA is receiving increasing attention as a basic building block in analog circuit design and few recent realizations have been suggested to implement it [3, 9, 10]. The basic principle behind the design of the OTRA is to provide amplification of high frequency signals with the ease of using standard operational amplifiers.

Many op-amp applications depend on current processing properties where currents from different circuits are added at the input terminal that is virtually grounded as for example in the Tow Thomas biquad. On the other hand the output voltage can be distributed to several circuits connected in parallel at the output terminal. The OTRA can provide the previous advantages in addition to a constant bandwidth virtually independent of the gain.

2. Circuit Description

The Operational Transresistance Amplifier (OTRA) is a three terminal analog building block shown symbolically in Fig. 1 with a describing matrix in the form:

$$\begin{bmatrix} V_+ \\ V_- \\ V_o \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ R_m & -R_m & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_o \end{bmatrix}. \qquad (1)$$

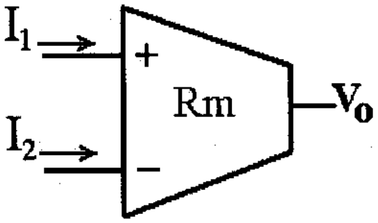

Fig. 1: Operational Transresistance Amplifier Symbol

Both the input and output terminals are characterized by low impedance, thereby eliminating response limitations incurred by capacitive time constants leading to circuits that are insensitive to the stray capacitances at the input terminals [9, 11]. For ideal operation, the transresistance gain, R_m, approaches infinity forcing the input currents to be equal. Thus the OTRA must be used in a negative feedback configuration in a way that is similar to the op amps. Practically the transresistance gain is finite and its effect should be considered. Also the frequency limitations associated with the OTRA should be considered.

Considering a single pole model for the transresistance gain, R_m, then

$$R_m(s) = \frac{R_o}{1 + \frac{s}{\omega_o}}. \qquad (2)$$

For high frequency applications, the transresistance gain, $R_m(s)$, can be expressed as

$$R_m(s) \approx \frac{1}{sC_p}, \qquad (3)$$

where

$$C_p = \frac{1}{R_o \omega_o}.$$

A CMOS realization of the OTRA is presented in **Fig. 2**. It is based on the cascaded connection of the Modified Differential Current Conveyor (MDCC) and a common source amplifier [10]. The MDCC provides the current differencing operation, whereas the common source amplifier provides the high gain stage.

The performance of the proposed CMOS OTRA circuit was ve-

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4. The offset voltage at the non-inverting terminal is shown in Fig. 5, whereas the systematic offset current is 50.9 nA.

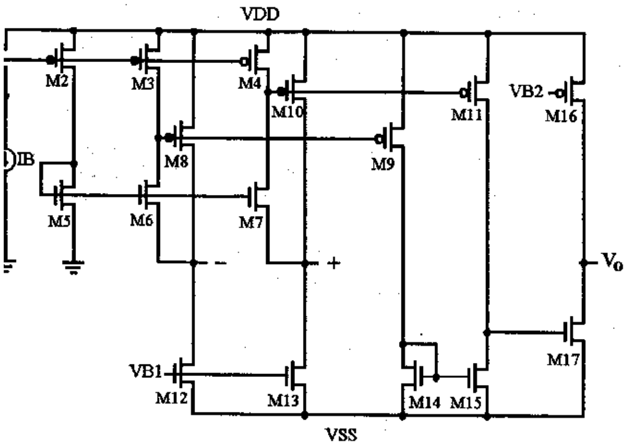


Fig. 2: CMOS realization of the OTRA

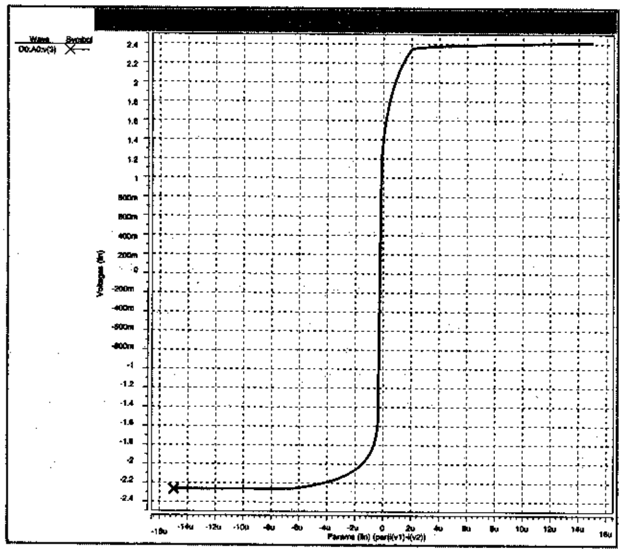


Fig. 3: Output voltage for open loop configuration

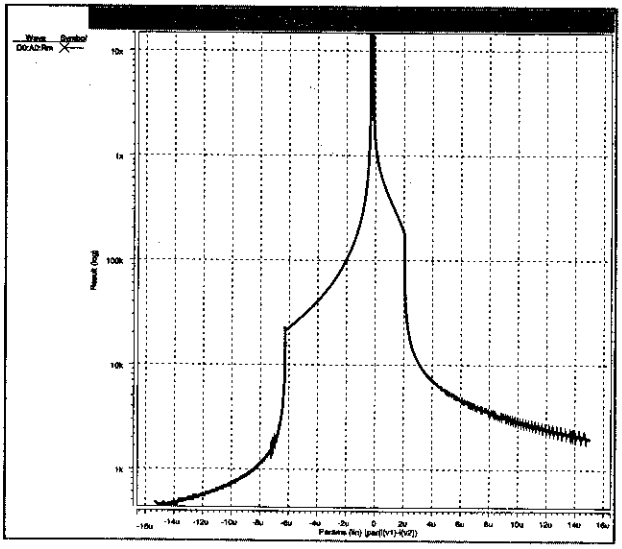


Fig. 4: Transresistance gain, R_m, at DC

mus (TT) biquad, a universal filter and a quadrature oscillator are presented.

3. The Differential amplifier

The OTRA can be used to realize different analog functions by using appropriate negative feedback circuits. The circuit shown in **Fig. 6** realizes a three port voltage controlled voltage source (VCVS). The output voltage is given by

$$V_o = K(V_2 - V_1) \tag{4}$$

where

$$K = \frac{R_1}{R}. \tag{5}$$

If the finite transresistance gain, R_m, was taken into consideration then

$$K = \frac{R_1}{R}\varepsilon(s) \tag{6}$$

where,

$$\varepsilon(s) = \frac{1}{(1 + \frac{R_1}{R_m(s)})} \tag{7}$$

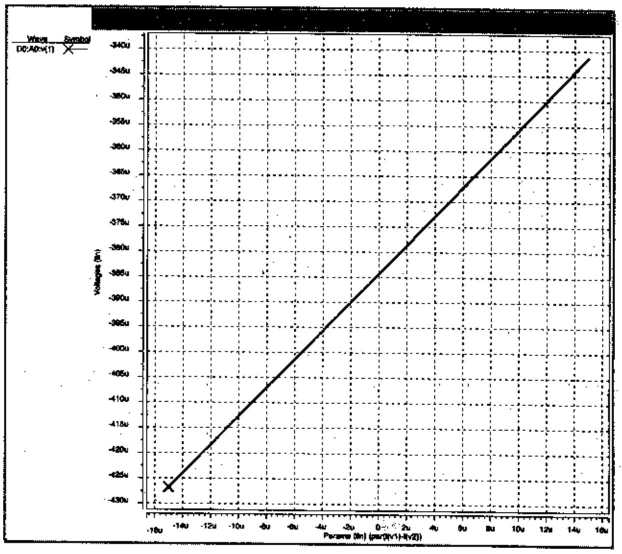


Fig. 5: Offset voltage at the non-inverting terminal of OTRA

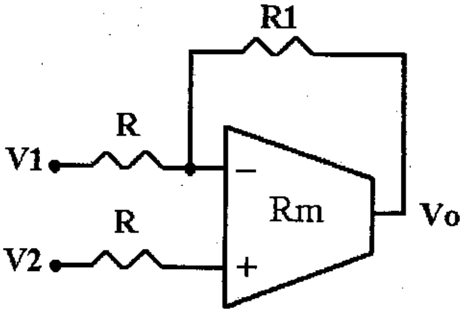


Fig. 6: OTRA used as a voltage controlled voltage source (VCVS)

By substitution from (3) into (7), the error function, $\varepsilon(s)$, reduces to

$$\varepsilon(s) = \frac{1}{(1+R_1C_ps)}. \tag{8}$$

Thus a single OTRA is capable of providing equal gain for both the inverting and the non inverting inputs a property unachievable with the VCVS using conventional op-amps in which the non inverting voltage gain equals the inverting gain plus one and the bandwidth depends on both the voltage gain and the bandwidth of the op amp. **Fig. 7** shows the frequency response of the output voltage for $K = 1$ and $K = 10$.

For high frequency applications, compensation methods must be employed in order to account for the error introduced in (8). Considering the circuit shown in **Fig. 8**.

$$\varepsilon(s) = \frac{1}{(1+R_1(C_ps - Y))}. \tag{9}$$

By choosing

$$Y1 = C_ps \tag{10}$$

$\varepsilon(s)$ reduces to its ideal value of unity. Therefore complete passive compensation of the VCVS can be achieved by using a single capacitor connected between the output terminal and the non inverting terminal.

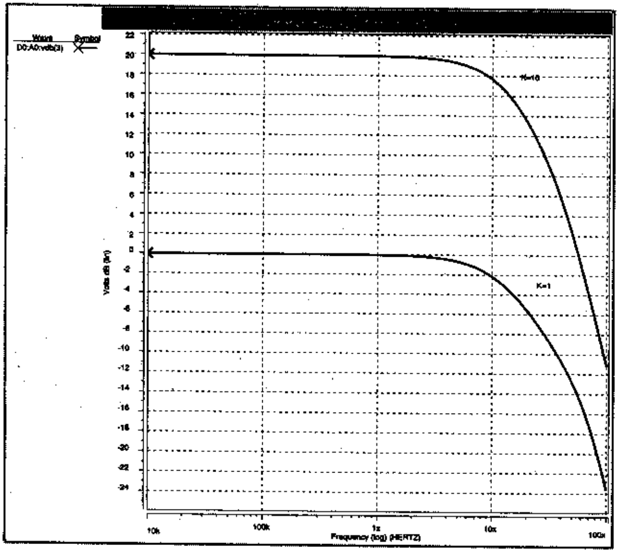
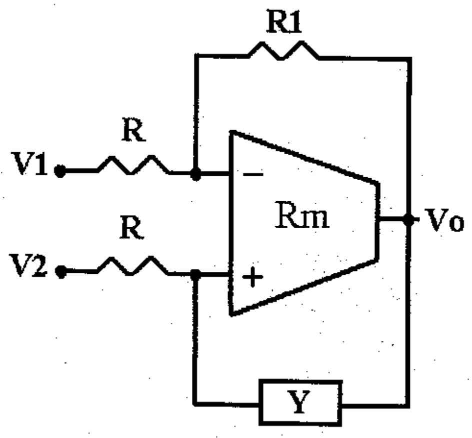


Fig. 7: Frequency response of the VCVS for $K = 1$ and $K = 10$



4. Differential integrator

A differential voltage integrator is shown in **Fig. 9**. The circuit has the advantage of using a single capacitor that is virtually grounded reducing the effect of stray capacitance [12]. The output voltage is given by

$$V_o = \frac{\omega_o}{s}(V_2 - V_1) \tag{11}$$

where

$$\omega_o = \frac{1}{CR}. \tag{12}$$

Fig. 10(a) and **Fig. 10(b)** represent the magnitude and phase responses of both the ideal and proposed non inverting integrator having $V_1 = 0$, $R = 50$ kΩ and $C = 200$ pF. Taking into consideration the effect of the finite transresistance gain given in (3) for the circuit shown in Fig. 9, (12) reduces to

$$\omega_o = \frac{1}{(C_p + C)R}. \tag{13}$$

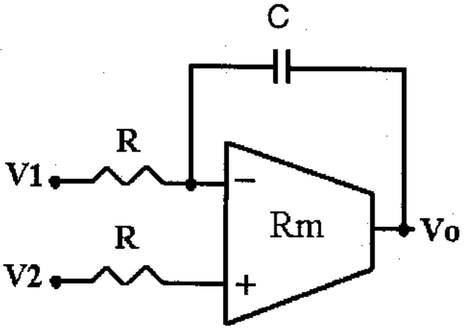


Fig. 9: Differential voltage integrator using OTRA

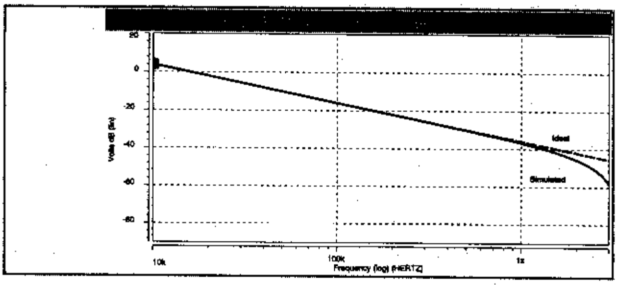
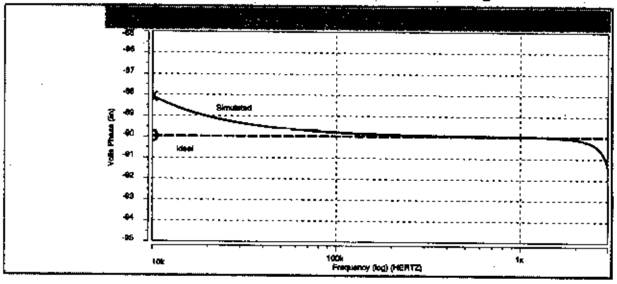


Fig. 10(a): Magnitude response of the OTRA based integrator



effect of C_p, by taking the design value of C equal to its theoretical value minus C_p. Self compensation can be achieved by using a capacitor, C', where

$$C' = C - C_p. \tag{14}$$

Thus the effect of C_p is absorbed in the integrating capacitance C' and no additional elements for compensation are needed.

5. Continuos Time Active Filters

Continuous time filters using conventional op amps, transconductors and switched capacitors are now widely accepted in industry where they are used in applications involving direct signal processing especially for medium dynamic range applications [2].

Usually two inverting integrators are cascaded and a third inverter allows closing the overall loop with the proper phase. This idea is behind many of the biquad filter structures available. The active elements count in these structures can be reduced, if a true non inverting integrator could be build with a single element [7]. Unfortunately this can't be done with conventional op amps but easily done using the OTRA. In addition op amp filters suffer from limited slew rate and are gain bandwidth limited.

Although gm-C filters can provide high frequency response, weighted sum integrators and weighted sum adders has to be performed with a number of gm-C cells with different gm factors. This is easily achieved using a single OTRA due to the presence of two virtually grounded terminals reducing the number of active elements leading to less power consumption and less noise to the whole system.

Switched-C filters are very accurate in the range of few kilohertz but become very power consuming in the range of Megahertz. Thus the applications of the OTRA will be useful in video applications which are in the range of few Megahertz.

The realization of the Tow Thomas biquad, a universal filter and a quadrature oscillator are introduced in the following subsections. The proposed circuits enjoy approximately the same structure of a cascade of an opposite polarity ideal and lossy integrators.

5.1. The Tow Thomas biquad and universal filter

Since the TT biquad depends on current summing by introducing a virtual ground at the input terminal of the conventional op amp and on the availability of performing only inverting integration, three op amps are needed. On the other hand since the OTRA doesn't suffer from these problems, four different realizations of the TT biquad can be implemented using only two OTRAs a feature that does not exist in the classical TT biquad [13]. Two different realizations of the TT biquad are shown in **Fig. 11**. The other two realizations can be derived by connecting the input resistance, R_4, to the positive terminal of the first OTRA leading to the inversion of polarities of both the bandpass and lowpass responses.

The transfer function of the bandpass and the lowpass outputs with their polarities determined based on the configuration used are given by

$$\frac{V_1}{V_i} = \pm \frac{\dfrac{s}{R_4 C_1}}{D(s)} \tag{15}$$

$$\frac{V_2}{V_i} = \pm \frac{\dfrac{1}{R_1 R_4 C_1 C_2}}{D(s)} \tag{16}$$

where $D(s)$ is given by

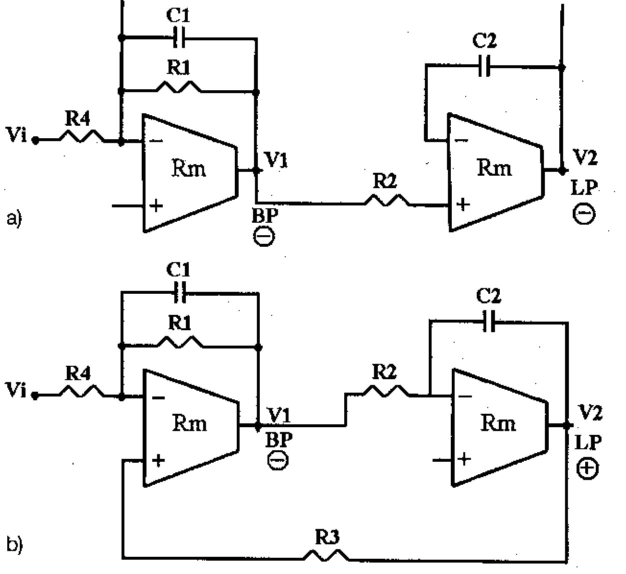


Fig. 11: Tow Thomas biquad

$$D(s) = s^2 + \frac{1}{R_1 C_1} s + \frac{1}{R_2 R_3 C_1 C_2}. \tag{17}$$

Thus ωo and Q are given by:

$$\omega_o = \sqrt{\frac{1}{R_2 R_3 C_1 C_2}} \qquad Q = R_1 \sqrt{\frac{C_1}{R_2 R_3 C_2}}. \tag{18}$$

For a lowpass response with a specified DC gain, K:

$$R_4 = \frac{R_3}{K}. \tag{19}$$

For a bandpass response with a specified center frequency gain, $|T(j\omega_o)|$

$$R_4 = \frac{R_1}{|T(j\omega_o)|} \tag{20}$$

It is clear from (18) that the quality factor Q can be independently controlled without affecting ω_o by varying R_1. Also from (19) and (20) the resistor R_4 controls the filter gain without affecting either ω_o or Q. The proposed filter has the advantage of operating in a mixed mode with input current and output voltage if the input resistance R_4 is short-circuited and a current source I_{in} replaces V_{in}.

The effect of the transresistance gain, $R_m(s)$, can be cancelled using the method described in (14). Thus the TT biquad filter is self compensated by absorbing the effect of the stray capacitance C_p presented in (3) in both C_1 and C_2 and no additional elements for compensation are needed.

Fig. 12 represents the bandpass response of the filter designed to give $Q = 50$. Both the ideal and the simulated results for the bandpass responses are presented where:

$C_1 = C_2 = 15$ pF, $R_2 = R_3 = 10$ kΩ, $R_1 = 500$ kΩ and $R_4 = 50$ kΩ.

The filter structure presented in **Fig. 11** can be slightly modified by applying feedthrough using a resistor and a capacitor from the input to the second OTRA to produce a novel universal filter that

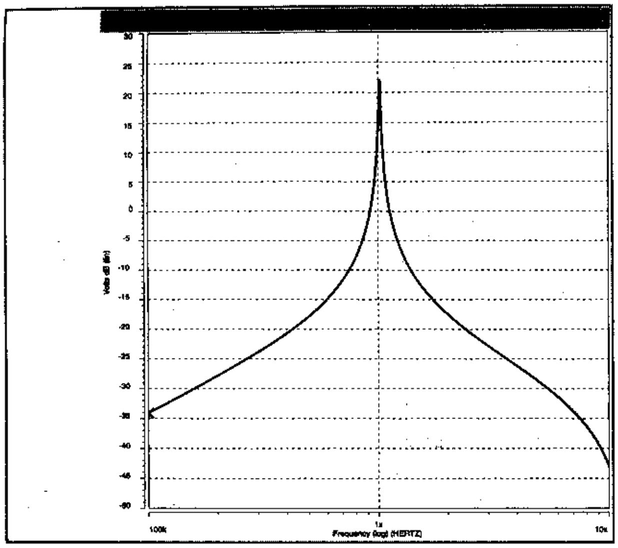


Fig. 12: Bandpass response of the TT biquad with Q = 50

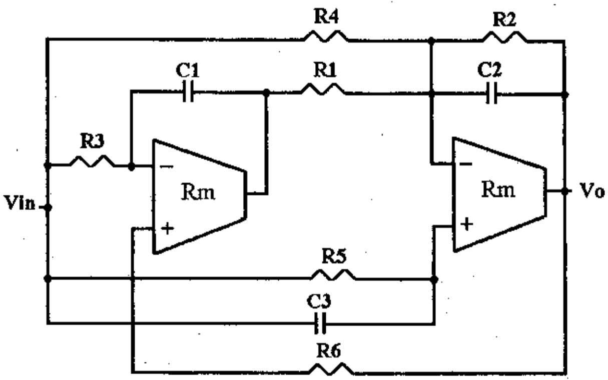


Fig. 13: Universal Filter

requires a minimum number of active elements as shown in **Fig. 13**. The output voltage is:

$$\frac{V_o}{V_i}=\frac{\frac{C_3}{C_2}s^2+(\frac{1}{R_5C_2}-\frac{1}{R_4C_2})s+\frac{1}{R_1R_3C_1C_2}}{D(s)} \tag{21}$$

where $D(s)$ is given by

$$D(s)=s^2+\frac{1}{R_2C_2}s+\frac{1}{R_1R_6C_1C_2}. \tag{22}$$

All possible outputs are summarized in **Table 1** with ω_o and Q given by:

$$\omega_o=\sqrt{\frac{1}{R_1R_6C_1C_2}} \qquad Q=R_2\sqrt{\frac{C_2}{R_1R_6C_1}}. \tag{23}$$

It is clear from (23) that the quality factor Q can be independently controlled by varying R_2 without affecting ω_o.

Again the proposed universal filter is self compensated since the effect of the stray capacitance C_p presented in (3) can be absorbed in both C_1 and C_2.

Fig. 14(a) and **Fig. 14(b)** represent the magnitude and phase responses of a notch filter having:

Table 1: Realizabilty conditions for the universal filter

Filter Response	**Realizability Conditions**	**Passive Elements**
HP	$R_3=R_4=R_5=\infty$	$3C$, $3R$
Non Inverting BP	$C_3=0$, $R_3=R_4=\infty$	$2C$, $4R$
Inverting BP	$C_3=0$, $R_3=R_5=\infty$	$2C$, $4R$
LP	$C_3=0$, $R_4=R_5=\infty$	$2C$, $4R$
Notch	$R_4=R_5=\infty$	$3C$, $4R$
AP	$C_2=C_3$, $R_6=R_3$ $R_4=R_2$, $R_5=\infty$	$3C$, $5R$

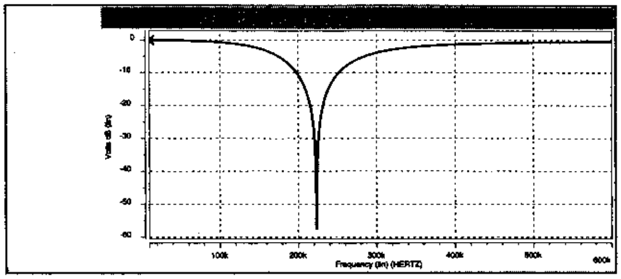


Fig. 14(a): Bandstop magnitude response of the universal filter

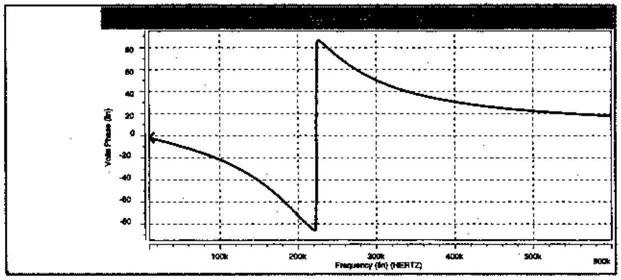


Fig. 14(b) Bandstop phase response of the universal filter

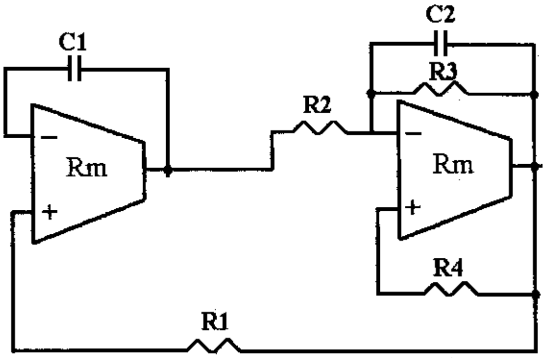


Fig. 15: Quadrature oscillator circuit

5.2. Quadrature Oscillator

A novel quadrature oscillator is presented in **Fig. 15**. Up to the authors best knowledge, this is the first **RC** oscillator using the OTRA. The characteristic equation is given by:

$$s^2+(\frac{1}{R_3}-\frac{1}{R_4})\frac{s}{C_2}+\frac{1}{C_1C_2R_1R_2}=0. \tag{24}$$

Thus the circuit represents a minimal component virtually grounded capacitor oscillator with independent control on the condition of oscillation as described by:

$$R_4=R_3. \tag{25}$$

It is seen that the resistances R_3 and R_4 control the condition of oscillation without affecting the radian frequency of oscillation

$$\omega_o = \sqrt{\frac{1}{C_1 C_2 R_1 R_2}}. \qquad (26)$$

The passive sensitivities of this oscillator are all low and are given by:

$$S_{R1}^{\omega_o} = S_{R2}^{\omega_o} = S_{C1}^{\omega_o} = S_{C2}^{\omega_o} = \frac{-1}{2}. \qquad (27)$$

If the effect of the finite transresistance gain, R_m, was considered, then (24) reduces to:

$$s^2 + \left(\frac{1}{R_3} - \frac{1}{R_4}\right)\frac{s}{(C_2 + C_p)} + \frac{1}{(C_1 + C_p)(C_2 + C_p)R_1 R_2} = 0. \qquad (28)$$

Thus the effect of C_p can be absorbed in C_1 and C_2 without increasing the order of the circuit.

6. Conclusions

The versatility of the Operational Transresistance Amplifier (OTRA) in realizing analog blocks is demonstrated. The OTRA provides constant bandwidth virtually independent of the gain making it suitable for continuous-time filters. Several filter structures that provide independent control over all filter design aspects have been presented. A novel quadrature oscillator with independent control on the frequency and condition of oscillation is introduced. Passive compensation and self compensation of the proposed circuits are presented. Simulation results which confirm theoretical analysis have been obtained.

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