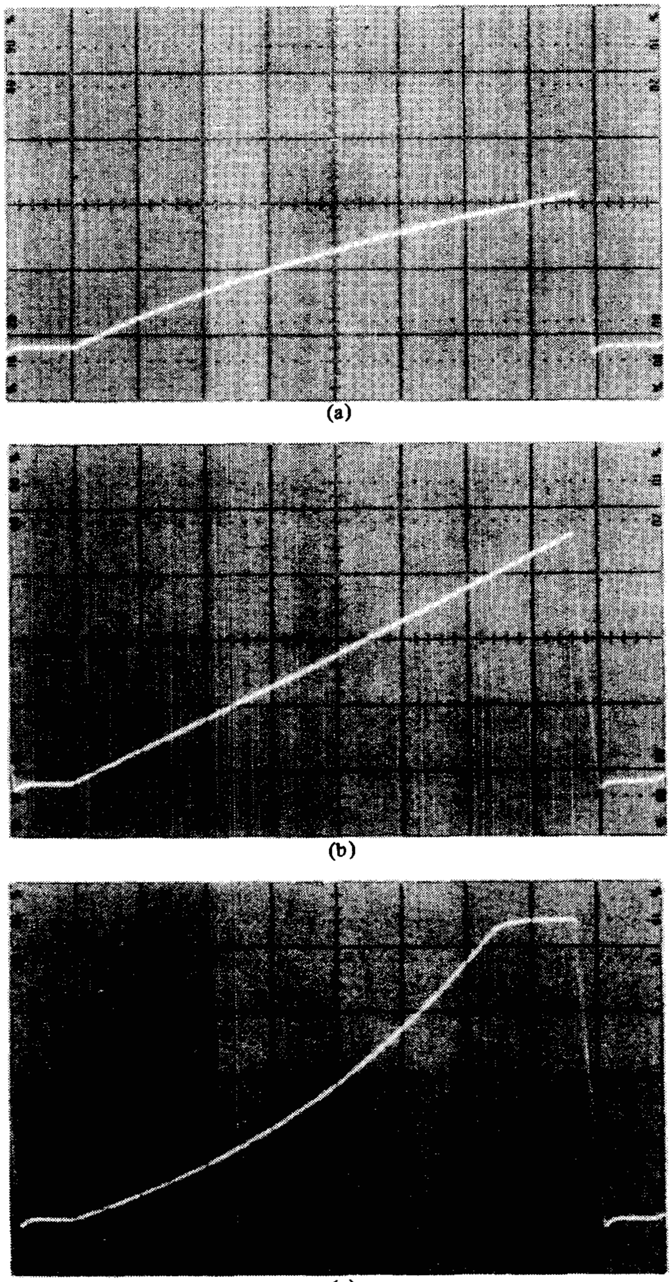

(a) (b) (c)

Fig. 3. (a) Uncompensated ramp generation. (b) Perfect compensation. (c) Overcompensation.

source. The transistor is used to periodically discharge the capacitor, so that the ramp voltage does not saturate. If we assume initial zero voltage across the capacitor, then it can be readily shown that

$$V_c(t) = (I\tau_c/C)\ \{1 - \exp(-t/\tau_c)\} \tag{1}$$

where $\tau_c (= R_{sh}C)$ is the leakage time constant of the capacitor.

Equation (1) is plotted in Fig. 1(c) against the ideal ramp voltage.

In order to correct for the exhibited nonlinearity, we will assume that a compensation current (I_c) is injected into the node d of Fig. 1(b) having a form given by

$$I_c(t) = V_c(t)/R_{sh} \tag{2}$$

where it is seen that $I_c(t)$ is equal to the leakage current. Therefore, the capacitor charging current is now given by

$$I_{charge} = I + I_c(t) = I + V_c(t)/R_{sh}. \tag{3}$$

It can then be easily shown that the resulting ramp voltage is given by

$$V_c(t) = (I/C)t. \tag{4}$$

In converting $V_c(t)$ to $I_c(t)$, an accurate and linear voltage to current converter will also be required. It can also be further shown that if the compensation current is larger than the required value, the slope of V_c with respect to time will be increasing with time rather than decreasing with time as in the case of Fig. 1(c).

Experimental Results and Discussions

Fig. 2 shows the experimental circuit used to verify the compensation principle. The OA_1 and Q_1 form an accurate voltage to current converter having R_1 as the conversion resistance. Q_2, Q_3, and Q_4 form a negative current mirror injecting the compensation current into point d. In order that the capacitance leakage effect can be conveniently shown on the oscilloscope screen, we have deliberately used a shunt resistor of 10 kΩ across the capacitor. The capacitance discharging transistor is not shown in Fig. 2.

Fig. 3(a) shows the waveform of V_c with no compensation. Fig. 3(b) shows the case of perfect compensation, where R_1 is very close to R, while Fig. 3(c) shows the case of overcompensation, where R_1 is significantly smaller than R.

It should be pointed out here that the compensation network looks like a negative resistance of magnitude R_1 from point d. Instability can be expected to occur when R_1 is smaller than R, which corresponds to the case of overcompensation, resulting in a positive feedback effect. This can be seen better if we suppose that $I = 0$; then any stray voltage across R_1 will result in an extra current to charge C, since the feedback current magnitude is greater than the current flowing through R. Thus it will lead to a buildup of voltage V_c. However, this instability does not introduce practical difficulty in ramp generation, since the capacitor will be discharged periodically. The instability simply manifests as an increasing slope of V_c with respect to time as shown in Fig. 3(c).

With slight modifications, the circuit can be used to compensate for the effect of the required parallel resistor in a sinusoidal integrator. However, care must now be taken not to overcompensate, which will result ultimately in the saturation of the voltage across the capacitor.

Conclusions

The proposed circuit technique should be useful in the generation of a very linear slow ramp voltage even with a rather lossy capacitor. It should also be useful in the generation of precision ramp voltages in general as required in electronic equipment.

On the Active Compensation of Noninverting Integrators

AHMED M. SOLIMAN AND MOHAMMED ISMAIL

***Abstract*—A new active compensated balanced time constant (BTC) noninverting integrator is proposed. The compensation is achieved by using a voltage follower in the feedback path, and this result in an extremely high Q-factor, namely $Q \simeq -|A|^3$. It is also shown, that the use of a voltage follower with the Deboo noninverting integrator will only double its Q-factor. A novel method is proposed for active compensation of the Deboo integrator, which results in an extremely high Q-factor.**

I. Introduction

The operational-amplifier (op-amp) integrators find wide use in many active RC filter biquads [1]-[4]. It is well known, that the finite and complex gain nature of the op-amp degrades significantly the performance of the integrator circuit. There are several methods for the compensation of the Miller (inverting) integrator [4]-[6]. On the other

Manuscript received December 6, 1978; revised January 8, 1979.
A. M. Soliman is with the Faculty of Engineering, Cairo University, Giza, Egypt.
M. Ismail was with the Faculty of Engineering, Cairo University, Giza, Egypt. He is now with the Arab Organization for Industrialization, Cairo, Egypt.

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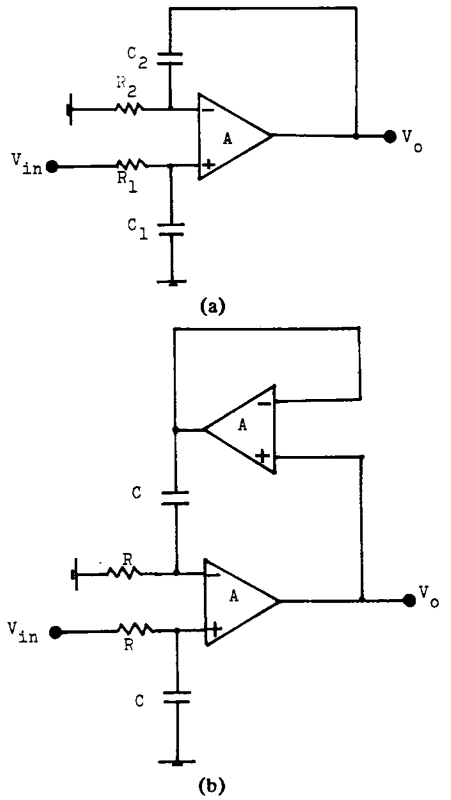

Fig. 1. (a) The BTC integrator. (b) The active compensated BTC integrator.

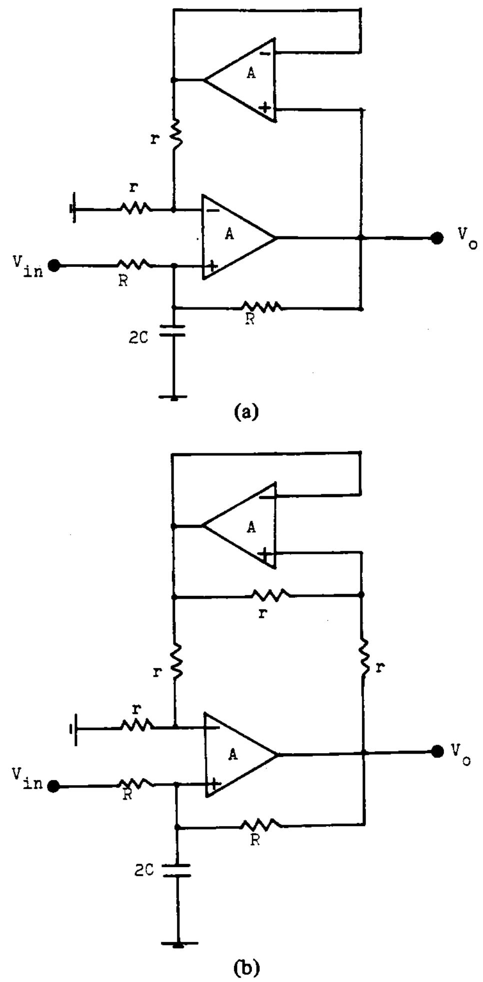

Fig. 2. (a) The active compensated Deboo integrator using a voltage follower. (b) The new high Q active compensated Deboo integrator.

hand, there are only few compensated noninverting integrators available in the literature. For example, Brackett and Sedra [4] have proposed a method for active compensation of the Miller inverter (noninverting) integrator. Their method is based on the feedforward-compensation technique and achieves a three-fold increase in the magnitude of the Q-factor by changing only one connection in the circuit. Two of the noninverting integrators that may be used in place of the Miller inverter cascade in the biquad circuits are the balanced time constant (BTC) integrator [2], and the Deboo integrator [7]. Recently, passive and active compensation methods have been proposed for the Deboo integrator [8]. As pointed out before, the passive compensation method is of limited practicality since the compensating capacitor must be adjusted according to the used op-amp, power-supply voltage, and temperature [8]. So if any of these conditions is changed, the passive compensation will no longer be satisfactory.

The purpose of this letter is to introduce an active compensation method for the BTC noninverting integrator. The modified BTC integrator uses a voltage follower in the feedback path in a similar manner as was used by Vogel [5], Brackett and Sedra [4], for active compensation of the Miller integrator. It is also shown that the use of a voltage follower is not very efficient with the Deboo integrator, as it will only double the Q-factor magnitude. A novel active compensation method for the Deboo integrator is introduced which results in an extremely high Q-factor. This proposed modified Deboo integrator is equivalent to the recently described circuit proposed by the authors [8].

II. The Integrator Q-Factor

Here the integrator Q-factor is defined as in [9]. Thus, taking the op-amp frequency response into account the transfer function of an integrator circuit may be expressed as:

$$T(j\omega) = \frac{1}{R(\omega) + jX(\omega)}$$

where $R(\omega)$ and $X(\omega)$ are real functions of the frequency variable ω. The integrator Q-factor is defined as: $Q = X(\omega)/R(\omega)$. If the op-amp is assumed to have a single-pole open-loop response with a unity gain bandwidth ω_t, then its gain is approximately given by:

$$A(s) \simeq \frac{\omega_t}{s}. \tag{1}$$

Taking the op-amp frequency response into account, the transfer function of an integrator circuit may be expressed as:

$$T(s) = \frac{\omega_0}{s} \cdot \epsilon(s) \tag{2}$$

where $\epsilon(s)$ is the error function contributed by the finite gain bandwidth of the op-amp, (in the ideal case $\epsilon(s) = 1$, and the integrator Q-factor is infinite). In general $\epsilon(s)$ may be expressed as

$$\epsilon(s) = \frac{1 + a_1 s}{b_0 + b_1 s + b_2 s^2} \tag{3}$$

where the coefficients a_1, b_0, b_1, and b_2 are functions of the op-amp unity gain bandwidth ω_t and the integrator time constant $1/\omega_0$. From (3) the following expression for the integrator Q-factor is easily obtained:

$$Q = \frac{b_0 + \omega^2(a_1 b_1 - b_2)}{\omega(a_1 b_0 - b_1) - \omega^3 a_1 b_2}. \tag{4}$$

It is very convenient to use the above expression for evaluating the integrator Q-factor in terms of the coefficients of the error function.

III. Active Compensated BTC Integrator

Fig. 1(a) shows the BTC noninverting integrator [2]. Assuming ideal op-amp its transfer function is

$$\frac{V_o}{V_{in}} = \frac{1}{sC_1R_1} \cdot \frac{s + 1/C_2R_2}{s + 1/C_1R_1}. \tag{5}$$

Hence the condition of perfect integration is:

$$C_1R_1 = C_2R_2 = CR. \tag{6}$$

And the above transfer function becomes:

$$\frac{V_o}{V_{in}} = \frac{1}{sCR} = \frac{\omega_0}{s}. \tag{7}$$

Taking the single pole model of the op-amp into account, the transfer function is expressed as:

$$\frac{V_o}{V_{in}} = \frac{\omega_0}{s} \cdot \epsilon(s) \tag{8}$$

where

$$\epsilon(s) = \frac{1}{(1 + \omega_0/\omega_t) + (1/\omega_t)\, s}. \tag{9}$$

The Q-factor is given by:

$$Q \simeq -\frac{\omega_t}{\omega} = -|A(j\omega)|, \qquad \omega_t \gg \omega_0. \tag{10}$$

An active compensation scheme for the BTC integrator is suggested in Fig. 1(b) by adding a voltage follower in the feedback path. Using matched op-amps the compensated BTC integrator will have the following error function:

$$\epsilon_c(s) = \frac{1 + s/\omega_t}{(1 + \omega_0/\omega_t) + (1 + \omega_0/\omega_t)\,(s/\omega_t) + (s^2/\omega_t^2)}. \tag{11}$$

Using (4) the compensated Q-factor is obtained as:

$$Q \simeq -\frac{\omega_t^3}{\omega^3} = -|A(j\omega)|^3, \qquad \omega_t \gg \omega_0. \tag{12}$$

It is seen that the proposed compensation method results in an extremely high Q-factor.

IV. Active Compensated Deboo Integrator

Taking the single pole model of the op-amp into account, the error function of the uncompensated Deboo noninverting integrator is given by:

$$\epsilon(s) = \frac{1}{(1 + 2\omega_0/\omega_t) + 2s/\omega_t} \tag{13}$$

and the integrator Q-factor is [4]

$$Q \simeq -\frac{\omega_t}{2\omega} = -\frac{1}{2}|A(j\omega)|, \qquad \omega_t \gg 2\omega_0. \tag{14}$$

Fig. 2(a) represents an active compensated Deboo integrator, where a voltage follower is used in the feedback path. Using matched op-amps the compensated error function is obtained as:

$$\epsilon_{c1}(s) = \frac{1 + s/\omega_t}{(1 + \omega_0/\omega_t) + (1 + \omega_0/\omega_t)(2s/\omega_t) + 2s^2/\omega_t^2}. \tag{15}$$

Thus using (4), the compensated Q-factor is:

$$Q \simeq -\frac{\omega_t}{\omega} = -|A(j\omega)|, \qquad \omega_t \gg \omega_0. \tag{16}$$

From the above result it is seen that the use of a voltage follower for compensation is not efficient with the Deboo integrator as it results only in doubling its Q-factor. That is, the improvement in the Q-factor is not as significant as in the case of the Miller integrator [4]-[5], or the BTC integrator.

A further improvement in the Deboo integrator Q-factor is possible by adding two more resistors to the circuit as shown in Fig. 2(b). For this novel circuit using matched op-amps, the error function is obtained as:

$$\epsilon_{c2}(s) = \frac{1 + 2s/\omega_t}{1 + (1 + 2\omega_0/\omega_t)(2s/\omega_t) + 4s^2/\omega_t^2} \tag{17}$$

In this case, the Q-factor is derived using (4) and is given by:

$$Q \simeq -\frac{\omega_t^2}{4\omega\omega_0} = -\frac{\omega}{4\omega_0}|A(j\omega)|^2, \qquad \omega_t \gg 2\omega_0 \tag{18}$$

It is worth noting that this modified high Q-factor Deboo integrator is equivalent to the recently described circuits proposed by the authors [8]. The expression for the Q-factor as given by (18) is more accurate than that given before [8], namely, $Q \approx -\frac{1}{8}|A(j\omega)|^3$ which was derived from (17), based on the approximation of neglecting $2\omega_0/\omega_t$ compared to (1), from the coefficient of the s term of the denominator of $\epsilon_{c2}(s)$.

V. Effect of Mismatched Op-Amps

Although dual op-amps having closely matched characteristics are now available at low cost, it is of interest to consider the effect of a small mismatch in the ω_t's of the op-amps. For brevity only the case of the compensated BTC integrator is discussed here.

Assuming mismatched op amps are used, the compensated BTC integrator will have the following modified error function:

$$\epsilon_{cm}(s) = \frac{1 + s/\omega_{t_2}}{(1 + \omega_0/\omega_{t_1}) + (1 + \omega_0/\omega_{t_2})(s/\omega_{t_1}) + s^2/\omega_{t_1}\omega_{t_2}} \tag{19}$$

where ω_{t_1} is the unity gain bandwidth of the integrator op-amp, and ω_{t_2} is the unity gain bandwidth of the voltage follower op-amp. The integrator Q-factor is given by:

$$Q_m \simeq -\frac{1}{\omega(1/\omega_{t_1} - 1/\omega_{t_2}) + \omega^3/\omega_{t_1}\omega_{t_2}^2}, \qquad \omega_{t_i} \gg \omega_0 \quad (i = 1, 2) \tag{20}$$

Assuming,

$$\omega_{t_1} = \omega_t, \omega_{t_2} = \omega_t(1 + \gamma)$$

where γ represents the normalized mismatch in the ω_t's which may be positive or negative.

$$Q_m \simeq -\frac{\omega_t^3}{\omega^3} \cdot \frac{1}{1 + \gamma\,\omega_t^2/\omega^2}, \qquad \omega_t \gg \omega_0, \gamma \ll 1. \tag{21}$$

Acknowledgment

The authors would like to thank the reviewer for his valuable comments to include Section V of this letter.

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A Generalized Active Compensated Noninverting VCVS with Reduced Phase Error and Wide Bandwidth

AHMED M. SOLIMAN

***Abstract*—A general circuit for the active compensation of the op-amp noninverting VCVS is given. The circuit has the same topology as the Geiger maximally flat magnitude circuit [1]. The proposed design has**

Manuscript received January 25, 1979.
The author is with the Faculty of Engineering, Cairo University, Giza, Egypt.

0018-9219/79/0600-0963$00.75