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

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

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***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**

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the advantages of a smaller phase error and a larger bandwidth than the design in [1]. Moreover, it relaxes the restriction on the relizable dc gain obtained using the design in [1]. The proposed modified design includes the recently described, active compensated, voltage follower [2] as a special case.

I. Introduction

The noninverting voltage-controlled voltage source (VCVS), which includes the voltage follower as a special case, is a very useful building block in active *RC* networks. It is well known that the finite and complex open-loop gain nature of the op-amp degrades significantly the performance of the VCVS [2]-[4].

Let the open-loop gain of the op-amp be represented in the form

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

where ω_t is the unity gain bandwidth. For the uncompensated VCVS of Fig. 1, the transfer function is given by

$$T(s) = (K+1)\epsilon(s) \tag{2}$$

where

$$\epsilon(s) = \frac{1}{1 + (K+1)\, s/\omega_t}. \tag{3}$$

From the above equation it is seen that the open-loop gain nature of the op-amp results in phase and magnitude errors given by

$$\phi \equiv \arg.\, \epsilon(j\omega) \simeq -\frac{\omega}{\omega_t}(K+1) \tag{4}$$

$$\gamma \equiv |\epsilon(j\omega)| - 1 \simeq -\frac{1}{2}\left(\frac{\omega}{\omega_t}\right)^2 (K+1)^2 \tag{5}$$

$$\omega_t >> (K+1)\,\omega.$$

The above expressions indicate that the phase error is of a first-order magnitude, whereas the magnitude error is of a second-order magnitude. The normalized 3-dB bandwidth of this uncompensated VCVS is given by

$$BW_n \equiv \frac{BW}{\omega_t} = \frac{1}{K+1}. \tag{6}$$

The above well-known facts indicate that the VCVS structure requires mainly phase compensation.

The purpose of this letter is to introduce an active compensation method for the noninverting VCVS in order to reduce its phase error to a negligible level and to provide for a very wide bandwidth operation.

II. The Active Compensated VCVS

The generalized, active compensated, noninverting VCVS is shown in Fig. 2. It is clear that this circuit has the same topology as the Geiger circuit [1], which is restricted to a dc gain $\geqslant \sqrt{2} + 1$. The design considered in [1] was directed towards a maximally flat magnitude response (thus providing no magnitude error) which is required in some applications of the VCVS. At the same time it ignored the phase error completely, which is an important factor to consider in most applications of the VCVS. Consider the generalized circuit shown in Fig. 2, by simple analysis, its transfer function is given by

$$T(s) \equiv \frac{V_o}{V_{\text{in}}} = (K_2 + 1)\,\epsilon(s) \tag{7}$$

where

$$\epsilon(s) = \frac{1 + \dfrac{s}{\omega_{t_1}}}{1 + [(K_2+1)/(K_1+1)](s/\omega_{t_2}) + (K_2+1)s^2/\omega_{t_1}\omega_{t_2}}. \tag{8}$$

The circuit in [1] is a special case from Fig. 2 with

$$K_1 = \frac{K+1}{\sqrt{2K+3}} - 1 \text{ and } K_2 = K. \tag{9}$$

Assuming matched op-amps are used, the transfer function of the circuit in [1] is given by

$$T_G(s) = (K+1)\,\epsilon_G(s) \tag{10}$$

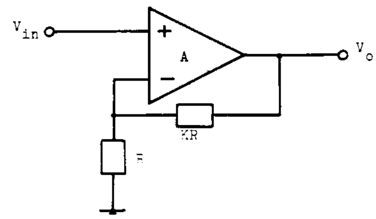

Fig. 1. The uncompensated, noninverting VCVS.

where

$$\epsilon_G(s) = \frac{1 + \dfrac{s}{\omega_t}}{1 + (\sqrt{2K+3})\dfrac{s}{\omega_t} + (K+1)\dfrac{s^2}{\omega_t^2}}. \tag{11}$$

which provides a maximally flat magnitude response with a dc gain of $(K+1) \geqslant \sqrt{2} + 1$.

Examining the above error function $\epsilon_G(s)$, it is seen that its phase error is of a first-order magnitude and is given by

$$\phi_G \simeq -\frac{\omega}{\omega_t}[\sqrt{2K+3} - 1], \quad \omega_t >> (K+1)\,\omega. \tag{12}$$

That is, if the VCVS is designed for a dc gain of 4 ($K = 3$), its phase error is only one-half that of the uncompensated VCVS of dc gain = 4, as can be seen from (4) and (12). It is clear that the design in [1] is not proper for phase compensation.

Now consider the generalized error function of (8). In order to design for a negligible phase error, choose

$$\frac{K_1 + 1}{\omega_{t_1}} = \frac{K_2 + 1}{\omega_{t_2}}. \tag{13}$$

Using matched op-amps (available now at low cost in dual packages), the above condition reduces to

$$\omega_{t_1} = \omega_{t_2} = \omega_t \text{ and } K_1 = K_2 = K \tag{14}$$

and the phase error is reduced to a negligible level given by

$$\phi_S \simeq -\left(\frac{\omega}{\omega_t}\right)^3 (K+1), \quad \omega_t >> (K+1)\,\omega. \tag{15}$$

From (13) it is obvious that any mismatch that may exist between the two opamps can be easily compensated for by adjusting one of the two resistors K_1R or K_2r. It is noted that although the two resistors R and r may be different, it is recommended to use equal resistors in order to minimize the spread in the circuit components. For the special case of interest, namely, when the conditions in (14) are satisfied, the generalized error function in (8) reduces to

$$\epsilon_S(s) = \frac{1 + s/\omega_t}{1 + s/\omega_t + (K+1)\, s^2/\omega_t^2}. \tag{16}$$

The magnitude error is given by

$$\gamma_S(s) \simeq \left(\frac{\omega}{\omega_t}\right)^2 (K+1), \quad \omega_t >> (K+1)\,\omega \tag{17}$$

which is still of a second-order magnitude as in the case of the uncompensated VCVS. For example, for a VCVS with a dc gain of 4, the proposed compensation reduces the magnitude error to one-half of its uncompensated value. However, the magnitude error is of a secondary importance since it is a second-order term, as mentioned earlier.

The 3-dB bandwidth of the proposed function in (16) is given by

$$BW_S = \frac{\omega_t}{(K+1)}\sqrt{\left(K + \frac{3}{2}\right) + \sqrt{\left(K + \frac{5}{4}\right) + 2(K+1)^2}}. \tag{18}$$

The corresponding bandwidth of the design in [1] is given by

$$BW_G = \frac{\omega_t}{(K+1)}\sqrt{\frac{1}{2} + \sqrt{\frac{1}{4} + (K+1)^2}}. \tag{19}$$

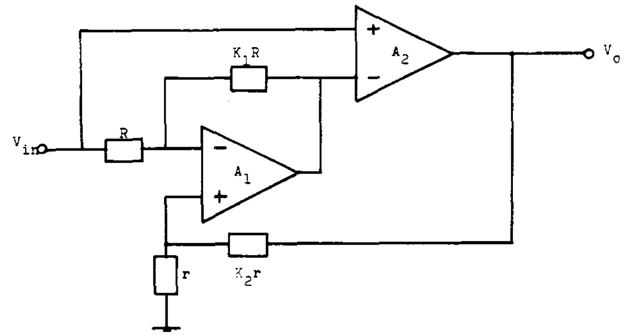

Fig. 2. The active compensated, noninverting VCVS.

From (18) and (19) it can be seen that the bandwidth improvement is given by

$$\frac{BW_S}{BW_G} = \left[\frac{\left(K + \frac{3}{2}\right) + \sqrt{\left(K + \frac{5}{4}\right) + 2(K+1)^2}}{\frac{1}{2} + \sqrt{\frac{1}{4} + (K+1)^2}} \right]^{1/2}. \tag{20}$$

As an example for $K = 3$, the bandwidth improvement $\simeq \sqrt{7/3}$.

It is worth noting that the compensated voltage follower may be obtained as a special case by setting $K = 0$ in (16). This corresponds to shorting the two resistors K_1R and K_2r in Fig. 2. Since the transfer function is now independent of the remaining two resistors r and R, they are taken as being open-circuited; this result in the same compensated voltage follower was described recently [2].

III. Conclusions

A generalized, active compensated, noninverting VCVS is given. As seen from (7) and (8), the resistor K_2r controls the dc gain, whereas the resistor K_1R controls the selectivity of the compensated response. It is found that the design in [1] (which is restricted to a dc gain $\geq \sqrt{2} + 1$) is not suitable for phase compensation. A modified design is proposed which is suitable for realizing a noninverting VCVS (having a dc gain $\geq$ 1) with a negligible phase error and a wide bandwidth. The recently proposed active compensated voltage follower [2] may be obtained as a special case from the modified circuit.

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Beamshaping and Polarization Control Properties of Flexible Hollow Metallic Rectangular Pipes in the Midinfrared

CHUNG YU

Abstract—**Radiation patterns of midinfrared (MIF) lead-salt tunable diode lasers (TDL) after transmission through a flexible hollow rectangular metallic pipe are studied. Near 100 percent transmission efficiency and polarization control properties are observed, leading to possible TDL waveguide encapsulation as an MIF source assembly with well-defined far-field radiation pattern and polarization.**

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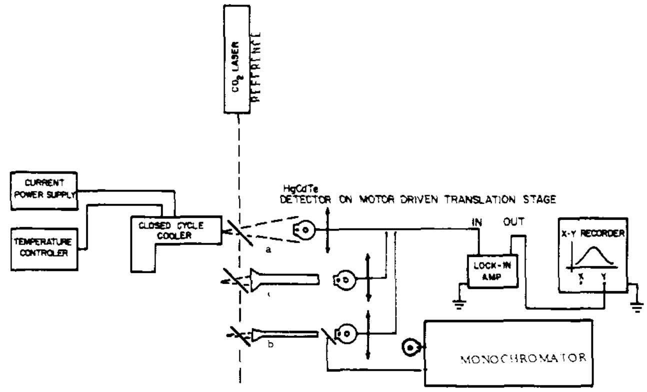

Fig. 1. Experimental setup and pipe orientations (b), (c).

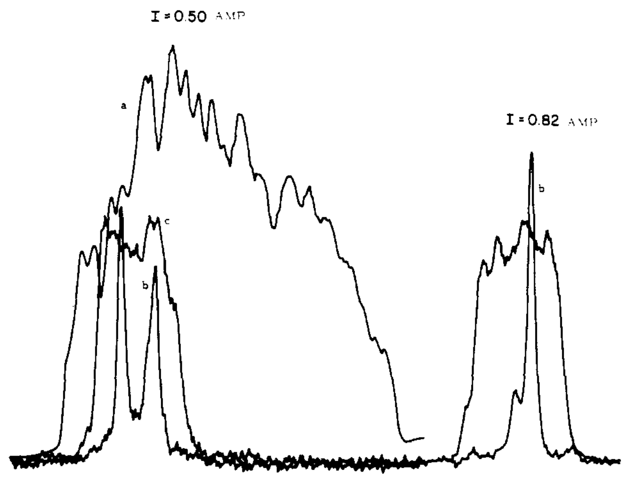

Fig. 2. (a) TDL radiation pattern. (b) TDL radiation pattern through pipe in one position. (c) TDL radiation pattern through pipe in orthogonal position.

Flexible hollow metal pipes have recently been proposed [1] as an efficient guiding medium for midinfrared (MIF) radiation in view of the nonavailability of optical fibers in this wavelength region. These pipes are reported to give 95 percent transmission in a one-meter length at 10.6 μm. This has prompted us to attempt to introduce such piping in our current and temperature tunable lead-salt diode laser (TDL) characterization setup to replace conventional lenses and mirrors.

Mirrors and KRS 5 or germanium lenses have hitherto been used to collect, shape, and guide MIF radiation. Manipulation of the MIF TDL beam through such optics has been found to be time consuming, and switching diodes means realignment of the entire guiding optics. A rapid diode testing scheme must therefore be comprised of relatively fixed receptive optics or guiding medium adaptable to a moving TDL source. Flexible hollow metal rectangular pipes are capable of shaping and controlling the polarization of radiation propagating through them independent of the nature of polarization and beamshape of the incident laser light. A number of such pipes were fabricated according to the design criteria of [1] using one-side-bright aluminum sheets. Typical pipe dimensions are $\frac{7}{16} \times \frac{1}{32} \times 13$ in. The spatial and spectral mode patterns are of interest. The rapidly divergent TDL beam at a nominal wavelength of 5 μm was studied on the experimental setup shown schematically in Fig. 1.

The TDL current drive was internally chopped so that the detected radiation could be fed to a lockin amplifier with the output traced out on an x-y recorder. The detector with a 1 mm × 1 mm aperture was mounted on a motor driven translation stage with a 2 in travel. The recorder patterns are given in Figs. 2–4.

Fig. 2(a) shows a complete TDL radiation pattern beyond the cold head window approximately 1 in from the diode emitting surface. The