Conclusions: Experimental investigations revealed that hygroscopic water increases both the real and imaginary parts of the dielectric constant of dust. The increase in the imaginary part is more remarkable. Previously published results on microwave attenuation due to dust storms are therefore valid only for dust with low water content, e.g. in desert and semidesert regions. For dust storms in tropical areas, microwave attenuation can be worse.

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0013-5194/80/100393-02\$1.50/0

NOVEL ACTIVE-COMPENSATED WEIGHTED SUMMER

Indexing terms: Active networks, Amplifiers

A summer amplifier with extended bandwidth is proposed. Compensation of the frequency characteristics is achieved by employing two operational amplifiers instead of external reactive components.

Introduction: In recent years, many authors have focused their attention on the development of active compensation schemes for the frequency characteristics of operational amplifiers. The availability of dual integrated amplifiers renders the approach very attractive. However, little effort has been directed towards the realisation of active-compensated summers which combine positive and negative weights. This is a serious limitation on the usefulness of active-compensated amplifiers, which are in practice reduced to the implementation of v.c.v.s.s.[1–3] The purpose of this letter is to report a new circuit configuration which can act as a compensated summer for both positive and negative gains.

Preliminaries: The output voltage of a conventional one-amplifier summer[4] is given by

$$V_o(s) = \left[\sum_{j=1}^{m} k_{j+} V_{j+} - \sum_{i=1}^{n} k_{i-} V_{i-}\right] \varepsilon(s) \tag{1}$$

The term $\varepsilon(s)$ is an error function[5] which represents the effect of the operational amplifier's dominant pole. It may be expressed by

$$\varepsilon(s) = \frac{1}{s\dfrac{k}{GB} + 1} \tag{2}$$

where

$$k = \max\left[\sum_{j=1}^{m} k_{j+},\ \left(1 + \sum_{i=1}^{n} k_{i-}\right)\right] \tag{3}$$

The low-frequency magnitude and phase errors are given, respectively, by

$$|\varepsilon(j\omega)| \simeq 1 - \frac{1}{2}\frac{\omega^2}{\omega_0^2} \tag{4a}$$

$$\arg|\varepsilon(j\omega)| \simeq -\frac{\omega}{\omega_0} \tag{4b}$$

where

$$\omega \ll \omega_0 = \frac{GB}{k} \tag{5}$$

Proposed summer: The new circuit is shown in Fig. 1. Routine analysis allows us to write the summer output voltage, by eqn. 1 whenever the following restriction applies:

$$1 + \frac{G_{0+}}{G_F} + \sum_{j=1}^{m} k_{j+} = \frac{G_{0-}}{G_F} + \sum_{i=1}^{n} k_{i-} \tag{6}$$

where

$$k_{j+} = G_{j+}/G_F; \qquad k_{i-} = G_{i-}/G_F \tag{7}$$

Assuming identical amplifiers, the error function for this circuit may be given by

$$\varepsilon(s) = \frac{s\left(\dfrac{2+a}{GB}\right) + 1}{s^2\dfrac{k}{GB^2}(2+a) + s\dfrac{k}{GB} + 1} \tag{8}$$

with

$$k = \max\left[\sum_{i=1}^{n} k_{i-},\ \left(1 + \sum_{j=1}^{m} k_{j+}\right)\right] \tag{9}$$

It is worth noting the difference between eqns. 3 and 9.

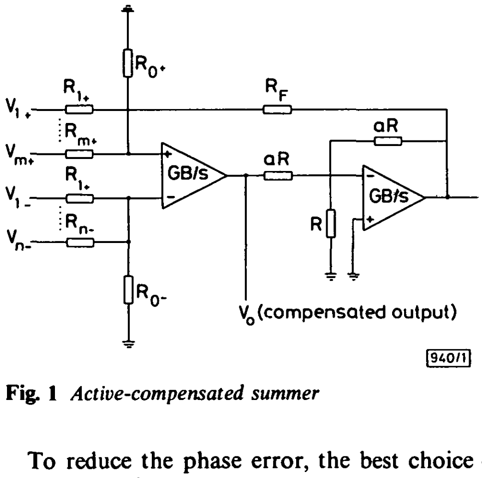

Fig. 1 *Active-compensated summer*

To reduce the phase error, the best choice of the auxiliary parameter a is

$$a = k - 2 \tag{10}$$

The low-frequency magnitude and phase errors are then expressed by

$$|\varepsilon(j\omega)| \simeq 1 + \omega^2/\omega_0^2 \tag{11a}$$

$$\arg[\varepsilon(j\omega)] \simeq -\omega^3/\omega_0^3 \tag{11b}$$

where eqn. 5 is still valid for the value of k given by eqn. 9.

Design considerations: The values of G_F and R can be arbitrarily fixed by the designer. Then G_{i-} and G_{j+} are determined via eqns. 7. Since the value of a is fixed by eqn. 10 the only resistor values which remain unknown are R_{0+} and R_{0-}. Their values can be obtained by the following closed formulas:

$$G_{0+} = \frac{G_F}{2} [1 + \text{sgn } M] \sum_{i=1}^{n} k_{i-}$$
$$G_{0-} = \frac{G_F}{2} [1 + \text{sgn } -M] \left(1 + \sum_{j=1}^{m} k_{j+}\right) \tag{12}$$

where

$$M = \sum_{i=1}^{n} k_{i-} - \left(1 + \sum_{j=1}^{m} k_{j+}\right)$$

and sgn $x = 1$ for $x > 0$ and sgn $x = -1$ otherwise.

It must be remarked that either G_{0+} or G_{0-} (or both) must be zero.

Discussion: An active-compensated summer has been reported which can be used for positive and negative gains. The proposed summer exhibits a second-order magnitude error as for an uncompensated one. However, the phase error, which is a significant factor in most applications, has been reduced to a negligible level, as can be seen by comparing eqns. 4*b* and 11*b*. Experimental data reveals good agreement between theoretical prediction and practical behaviour when using μA747 dual amplifiers. As an illustration, Fig. 2 shows the Bode plot corresponding to a summer with $n = m = 1$, $k_+ = 20$, $k_- = 10$ and the same input voltage for positive and negative inputs.

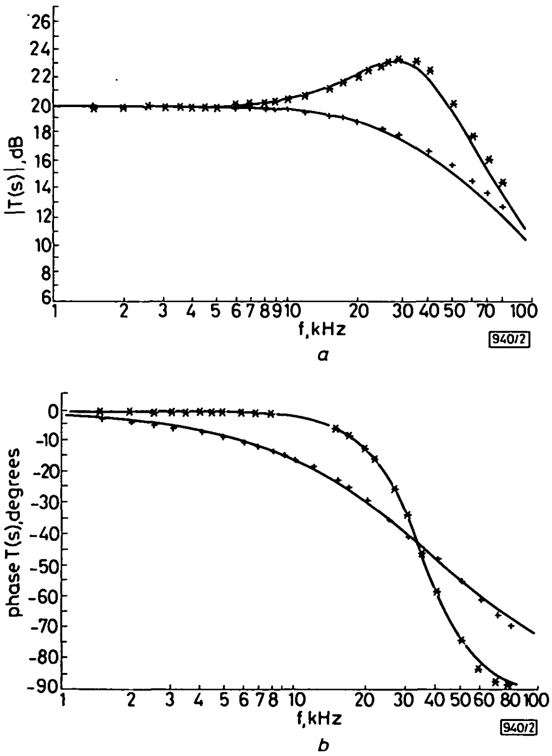

Fig. 2 *Bode plot for compensated summer with $n = m = 1$, $k_+ = 20$, $k_- = 10$ and $V_{1+} = V_{1-} = V_i$*

a Magnitude characteristics
b Phase characteristics
* compensated summer
\+ uncompensated summer
Continuous trace corresponds to theoretical Bode plots

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0013-5194/80/100394-02$1.50/0

NOVEL STRAY-INSENSITIVE SWITCHED-CAPACITOR INTEGRATOR

Indexing terms: Switched-capacitor networks, Filters

A stray-insensitive switched-capacitor integrator is introduced which enables the design of leapfrog filters suitable for high clock/signal frequency ratios without causing large capacitor spreads. A design example is given.

The well-known active-filter leapfrog technique[1] has been used for the design of switched-capacitor filters using Bruton's lossless discrete integrators (l.d.i.)[2] or bilinear integrators.[3] The advantage of the leapfrog technique is that it simulates *LC* ladder filter structures which are known to have low sensitivity to component variations. Problems arise, however, when for some reason the ratio of the clock frequency and the signal frequency must be made high, because the required capacitor spread may then increase to an unrealisable value. It can be shown,[2] for example, that the 3 dB bandwidth of a single-pole s.c. filter is given by

$$\omega_{3\,dB}^{SC} \simeq f_c \frac{C''}{C'} \tag{1}$$

where f_c is the sampling frequency and C', C'' are the two capacitors associated with the pole. Thus for a given cutoff frequency, the ratio C'/C'' increases with increasing clock frequency. A high clock frequency is often required because it simplifies the design of the antialiasing filter. It will be shown in this letter that by realising C'' as the difference between two capacitors, a high capacitive ratio can be obtained without the need for a large capacitor spread.

Consider the circuit shown in Fig. 1, which is an extension of the integrator introduced in Reference 4. Note that this circuit is completely insensitive to stray capacitances. The output voltages are readily obtained as

$$V_{out}^e = -\frac{C_1}{C_0}\frac{1}{1-z^{-2}} V_1^e + \frac{C_2}{C_0}\frac{z^{-1}}{1-z^{-2}} V_2^o + \frac{C_3}{C_0}\frac{z^{-2}}{1-z^{-2}} V_3^e - \frac{C_4}{C_0}\frac{z^{-1}}{1-z^{-2}} V_4^o \tag{2a}$$

$$V_{out}^o = -\frac{C_1}{C_0}\frac{z^{-1}}{1-z^{-2}} V_1^e + \frac{C_2}{C_0}\frac{z^{-2}}{1-z^{-2}} V_2^o + \frac{C_3}{C_0}\frac{z^{-1}}{1-z^{-2}} V_3^e - \frac{C_4}{C_0}\frac{1}{1-z^{-2}} V_4^o \tag{2b}$$

where $z = \exp(s/2f_c)$ and the open-loop gain A_0 of the amplifier is assumed to approach infinity.

New integrators can be derived using the morphological box[5] shown in Table 1. A solution S_{ik} designates an integrator obtained by connecting the input terminals i and k in Fig. 1, where $i, k = 1, 2, 3, 4$. Note that $S_{ik} = S_{ki}$. The superscript e or o denotes the time slot during which the output voltage is sampled.

Each solution in Table 1 must now be examined to see whether it provides the desired integration. Useful solutions