Exact cancellation of the two signals only occurs when the two receiving elements are precisely in line with the reflected principal ray. This direction will take into account the whole paraboloid surface and will vary with small distortions of that surface. The shortcoming with respect to the usual static split sensor is that the error signal obtained is axially symmetric. No directional information can be obtained to supply correcting drives, as is the case with monopulse tracking antennas. Since it is axially symmetric, however, the single added feed can give error signals in any direction, instead of the combination of four required in static-split receivers.

Focal fields of paraboloid: Minnett and Thomas[2] give the field equations for axial waves in the focal region of a paraboloid illuminated by a plane wave. For a wave in the direction of the paraboloid axis, these simplify to

$$E_{(z)} = -2k_0 f E \frac{\sin k_0 z}{k_0 z} e^{i(kz+\Delta)} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where E is the incident field polarised parallel to $E_{(z)}$, f is the focal length of the paraboloid, θ is the aperture halfangle, $k_0 = (2\pi/\lambda)\sin 2\theta/2$, $k = 2\pi/\lambda$, and Δ, the phase anomaly in optics, is $\pi/2 - k_0 z$. In this relation, z is measured positively from the focus along the axis away from the paraboloid, and λ is the wavelength.

The position of the first subsidiary maximum occurs therefore at z_1, where $k_0 z_1 = 4{\cdot}49$, and, since k_0 contains the factor $\sin 2\theta/2$, z_1 can be a multiple of the wavelength.

The magnitude of the first maximum is the value common to all $(\sin x)/x$ patterns, i.e. $-13{\cdot}2$ dB.

Experimental results: A paraboloid of aperture diameter 38 cm and focal length 26·5 cm (aperture halfangle approximately 70°) was illuminated with the near field of a larger paraboloid at a distance of 10 m, and the measurements were made at X band frequencies ($\lambda = 3{\cdot}1$ cm). A flanged open-ended waveguide was used as a receiving feed and was moved on a track along the nominal axis of the paraboloid to plot the field given by the above theory. The result was in good agreement, the only major difference being a slightly larger separation between the maxima than that predicted.

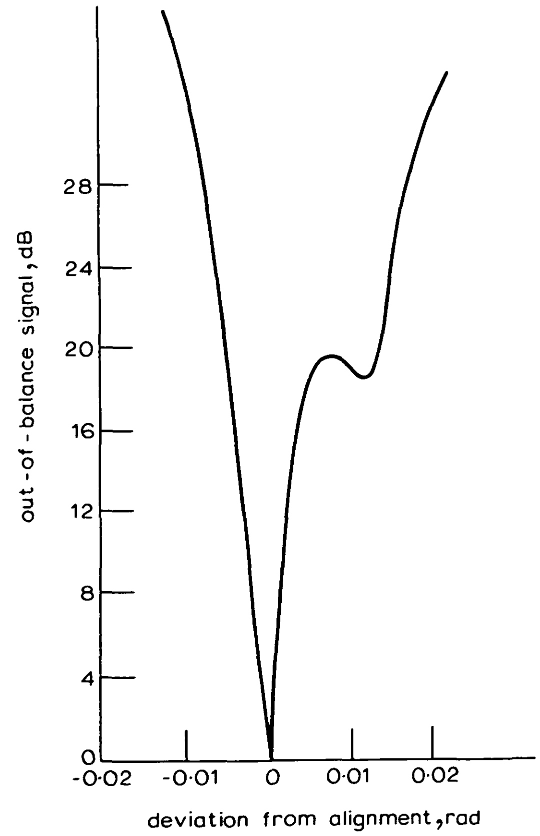

Fig. 2 *Error signal for small angular rotation*

A double feed and comparator was constructed with the separation required, and allowance for minor adjustment (see Fig. 1). The sample from the feed at the main focus was adjusted to the level of that at the secondary focus by a 10 dB directional coupler and level-setting attenuator. The phase of the secondary-feed signal was adjusted to cancel the sample, and it was found that complete cancellation could only be achieved when the two feeds were in the direct line of the paraboloid centres.

Small angular rotations of the receiving paraboloid were made and measured using a radius arm and a depth micrometer, which gave an accuracy of better than a milliradian. The response of the apparatus to rotations of less than 0·02 rad is shown in Fig. 2. During this experiment, it was observed that the galvanometer, when at zero in the null condition, was easily able to indicate vibrations of the paraboloid made by a light tap at the rim. An alignment procedure for large antennas can be proposed based on this principle. With a distant source and with the double feed in line with the nominal axis, the antenna is manoevred until the null is observed. Adjustments are made to the feed position and the procedure is repeated until the line of the double feed is exactly directed to the distant source. This line is then the precise electrical axis of the paraboloid.

Conclusions: The addition of a secondary feed to the usual front feed of a paraboloid system, spaced axially from the primary feed, provides a simple method of obtaining pointing accuracy of the same degree of sensitivity as a standard static-split system. A similar arrangement could be proposed for Cassegrain antennas, where the secondary feed could be positioned at the 'obscured' prime focus of the main paraboloid. The sensivitivity of the combination to small vibrations indicates that distortions of the paraboloid, such as the effects of gravity on large antennas in different elevation positions, could be detected and rectified. The combined feed could also be used for aperture-field determination where this uses the effects of beam pointing change.[2]

Acknowledgment: The apparatus was designed and constructed, and measurements were made, by S. G. Blacker, in his final year as a student in the department of physics at the University of Surrey.

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REALISATION OF OPERATIONAL-AMPLIFIER ALLPASS NETWORKS

Indexing terms: Operational amplifiers, Transfer functions, Network synthesis

With the effect of the nonideal operational amplifier taken into consideration, it is shown that the differential-input configuration for realising allpass transfer characteristics is superior to the dual-input configuration.

The realisation of 2nd-order allpass transfer characteristics with a single operational amplifier has received a great deal of attention recently.[1–8] The above realisations belong to either the differential-input configuration (also known as the operational-amplifier allpass circuit with earthed RC network), as shown[1–4] in Fig. 1, or the dual-input configuration,[5-8] which uses an earthed inverting operational amplifier, and is shown in Fig. 2. In all these References, it was assumed that the operational amplifier has infinite gain. On the basis of this assumption, it is well known that both configurations will have identical expressions for the open-circuit voltage transfer function, provided that the network N

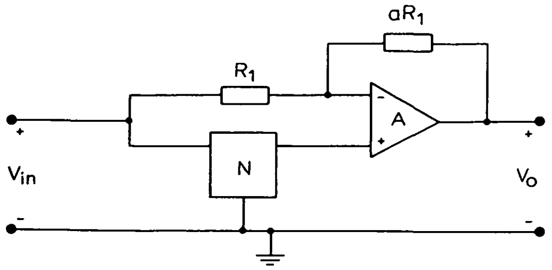

Fig. 1

is the same in both cases. The purpose of this letter is to compare the two configurations, taking into consideration the finite gain of the operational amplifier. For the differential-input operational-amplifier configuration, by direct analysis,

$$G(s) = \frac{V_0}{V_{in}} = \frac{1}{\left(1+\frac{a+1}{A}\right)}\{T(s)(a+1)-a\} \quad . \quad . \quad (1)$$

For the earthed-inverting-operational-amplifier configuration,

$$G(s) = \frac{1}{\left(1+\frac{a+1}{A}\right)}\left\{T(s)(a+1)\left(1+\frac{1}{A}\right)-a\right\} \quad (2)$$

where $T(s)$ is the transfer function of the RC network N, and A is the gain of the operational amplifier. In integrated circuits, operational amplifiers are designed with a dominant pole to assure stable operation. Thus[9]

$$A = \frac{A_0}{1+\frac{s}{\omega_1}} \simeq \frac{GB}{s} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

where

A_0 = open-loop d.c. gain,

ω_1 = open-loop 3 dB bandwidth, rad/s

$GB = A_0\,\omega_1$

From the above equations, it is seen that, for both configurations, the poles of $G(s)$ consist of the two real poles of $T(s)$, plus a simple real pole due to the operational amplifier at $s = -GB/(a+1)$. For the differential-input operational-amplifier configuration, the zeros of $G(s)$ are absolutely insensitive to the GB of the operational amplifier. On the other hand, for the earthed-inverting-operational-amplifier configuration, the zeros are sensitive to the amplifier GB. The sensitivity will depend on the passive RC network N.

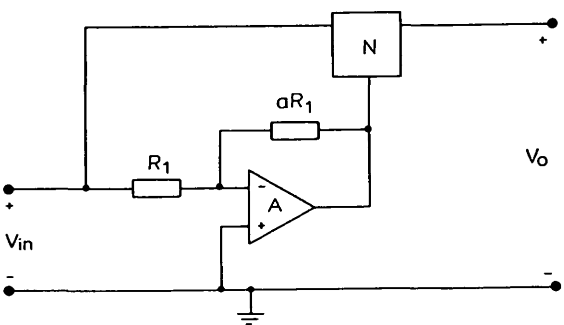

Fig. 2

As an example, consider the allpass network[5-6] with

$$T(s) = \frac{\frac{s}{CR}}{s^2+\frac{3s}{CR}+\left(\frac{1}{CR}\right)^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

From eqns. 4 and 2,

$$G(s) = \frac{-a}{1+\frac{a+1}{A}} \times \frac{s^2-\frac{s}{CR}\left\{\left(1+\frac{1}{a}\right)\left(1+\frac{1}{A}\right)-3\right\}+\left(\frac{1}{CR}\right)^2}{s^2+\frac{3s}{CR}+\left(\frac{1}{CR}\right)^2} \quad (5)$$

As $A \to \infty$, an allpass transfer characteristic with $\omega_z = 1/CR$ and $Q_z = \frac{1}{3}$ results if

$$a = \tfrac{1}{5} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

From eqns. 3 and 6 in eqn. 5,

$$G(s) = \frac{-1}{5\left(1+\frac{6s}{5GB}\right)} \times \frac{s^2\left(1-\frac{6}{CRGB}\right)-\frac{3}{CR}s+\left(\frac{1}{CR}\right)^2}{s^2+\frac{3}{CR}s+\left(\frac{1}{CR}\right)^2} \quad (7)$$

Therefore the actual values of ω_z and Q_z are given by

$$\omega_{z_a} = \frac{\omega_z}{\sqrt{\left(1-\frac{6\omega_z}{GB}\right)}} \simeq \omega_z\left(1+\frac{3\omega_z}{GB}\right) \quad \text{for } \omega_z \ll GB \quad (8)$$

$$Q_{z_a} = Q_z\sqrt{\left(1-\frac{6\omega_z}{GB}\right)} \simeq Q_z\left(1-\frac{3\omega_z}{GB}\right) \quad \text{for } \omega_z \ll GB \quad (9)$$

Hence

$$\frac{\Delta\omega_z}{\omega_z} = -\frac{\Delta Q_z}{Q_z} \simeq \frac{3\omega_z}{GB} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

The above sensitivities are significant at high frequencies. For example, for the μA 741 operational amplifier with $GB = 2\pi \times 10^6$ rad/s, if $\omega_z = \pi \times 10^5$ rad/s,

$$\frac{\Delta\omega_z}{\omega_z} = -\frac{\Delta Q_z}{Q_z} \simeq 15\%$$

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