IMPROVED ATTENUATED PROBE ELECTRONICS

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Plasma Studies

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Introduction

PLP 43 described a differential amplifier for use with attenuated probes. The amplifier discussed there had the following disadvantages:

(1) Driving the shields of the probe leads permits low impedance plasma to capacitively couple to the shield through the glass, producing an erroneous reading of the floating potential.

(2) The common mode rejection ratio was strongly dependent on frequency; and although it was possible to obtain a CMRR of 100:1 at any given frequency up to about 1 mc., the CMRR was much worse at other frequencies, particularly above 2 mc.

(3) The rise time of the amplifier was 0.2 μsec., corresponding to a high frequency cutoff of only 2 mc.

(4) The circuit was unnecessarily complicated. In particular, the nuvistors, which were originally used to give a high input impedance, are no longer necessary when the amplifier is used with an attenuated probe since, in that case, we desire a low input impedance.

(5) The filament battery lifetime was short and the amplifier gain was quite dependent on battery voltage.

This paper describes the steps which have been taken to overcome the above difficulties and suggests several improvements which might yet be made.

Coupling between Plasma and Shield

Whenever a shield is driven by a cathode follower, it ceases to be effective as a shield if its impedance to ground (usually the cathode resistance) is comparable to the coupling impedance between the plasma and the shield. Figure 1 shows an equivalent circuit for the plasma coupling. Vp is the floating potential, a rapidly changing function of time. Rp is the plasma impedance, which is purely resistive at low frequencies, and strongly dependent on plasma density. Cp is the coupling capacitance between the plasma at the surface of the glass and the shield inside. By direct calculation (concentric cylinders), an upper limit of 50 pf can be placed on Cp. This is in agreement with the value obtained by measuring the capacitance between the shield and a layer of aluminum foil wrapped around the glass. It is easily shown that for effective shielding, the following relationship must hold:

$$\sqrt{Rp^2 + \frac{1}{\omega^2 Cp^2}} \gg Rk .$$

Consider some typical values:

$$Rp = 10K = 10^4 \Omega$$
$$\omega = 2\pi f = 10^7 \text{ cps}$$
$$Cp = 10 \text{ pf} = 10^{-11} \text{ f}$$

Then $\sqrt{Rp^2 + \frac{1}{\omega^2 Cp^2}} = 14K.$ With the 12K cathode resistor previously used, the shielding would not be effective. To eliminate this problem, the shields were disconnected from the cathodes and grounded. The penalty for grounding them is that a higher attenuation factor is necessary, reducing the sensitivity of the system. With the shields grounded, the capacitive pickup was at most one part in 10^4 at all frequencies.

Common Mode Rejection Ratio

For a differential amplifier whose input consists of two identical cathode followers, the common mode rejection ratio critically depends upon accurately balancing both the gain and phase shift of the cathode followers for the following reasons:

(1) Any difference in gain or phase shift causes part of the common mode signal to appear at the output.

(2) The feedback and thus the capacitance between the center conductor and shields will differ and the attenuator will no longer be balanced.

Although it was possible to compensate for these effects at a given frequency, it was virtually impossible to obtain a high common mode rejection ratio over a large frequency spectrum with that amplifier.

Grounding the shields eliminates the second difficulty, since the capacitance at the output of the attenuator is a constant, dependent only on the cable capacitance and the input capacitance of the amplifier.

To overcome the other difficulty, a single-ended transistor amplifier was used. Since the whole circuit "floats" with the floating potential, there is no necessity of matching gains. The only requirement is that the capacitance and resistance from each input to ground must be equal, a condition which is easily satisfied with capacitive trimmers and pots.

Figure 3 shows the common mode rejection ratio as a function of frequency for a typical probe-amplifier combination. The amplifier can be adjusted so that the resonance can be moved to any frequency below about 1 mc. The limitations on the CMRR with the new system are twofold:

(1) The attenuators in the probe are not ideal. This effect will be discussed at some length in the next section.

(2) At high frequencies the CMRR is limited by the capacitive coupling between windings in the output pulse transformer. Figure 2 shows an equivalent circuit for this effect. By direct calculation, the CMRR of the pulse transformer is given by

$$\mathrm{CMRR} = \sqrt{1 + \frac{1}{\pi^2 f^2 R_L^2 C_c^2}} \approx \frac{1}{\pi f R_L C_c} .$$

Using the manufacturer's specification of C_c = 1.23 pf, the dotted curve in Figure 3 was obtained. This represents the maximum CMRR that one can achieve using a PE-5163 with a termination of $R_L = 500\Omega$. (500Ω was the smallest load that the amplifier was capable of driving without reducing the frequency response).

Frequency Response

The rise time of the amplifier in PLP 43 was only 0.2 μsec. The frequency response was limited by the leakage inductance of the output pulse transformer. In particular, the leakage inductance and capacitive load cause a resonance in the frequency response, above which the signal is greatly attenuated. By critically damping, the resonance can be eliminated, but the high frequency cutoff is given by $f = \frac{1}{2\pi\sqrt{L_L C_L}}$. For a capacitive load of 200 pf (180 for 6 ft. of RG 58 + 20 for the scope), the cutoff frequency is 4 mc., using the manufacturer's value of L_L = 14 μhy.

In the new amplifier, the frequency response was doubled by using a high impedance cable (RG 114AU, 6 pf / ft.) to couple the output to the scope. It was also found necessary to use a faster transistor (2N1132) since the cutoff frequency of the 2N404 and 2N1304 is rather low.

Figure 4 shows the frequency response of the amplifier with the above modifications. With the probe attached, the frequency response is reduced. The reason for the limited bandwidth of the probe is that the attenuators in the probe are non-ideal. Figure 5 shows an equivalent circuit for the attenuator. The capacitance C is $\sim$ 0.2 pf, assumed distributed evenly inside the resistor. With R_O adjusted so that the ideal attenuator relationship is satisfied ($RC = R_O C_O$), we would expect zero phase shift at all frequencies if the distributed capacitance, C', between the resistor and ground is neglected. When one includes C' (assumed to be distributed evenly along the resistor), the phase shift is given by the equation

$$\phi = \sqrt[4]{\frac{\omega^2 R^2 C'^2}{\omega^2 R^2 C^2 + 1}}$$

For low frequencies ($\omega \ll 2$ mc.), $\phi = \sqrt{\omega RC'}$, and for high frequencies ($\omega \gg 2$ mc.), $\phi = \sqrt{\frac{C'}{C}}$. In our case, it appears that C' is comparable to C, limiting the frequency response to ~ 2 mc.

Furthermore, any small difference in C' in the two attenuators puts the output signals out of phase with one another and reduces the common mode rejection ratio.

Amplifier Circuit

The necessity of a low input impedance and a relaxation of the requirement that the shields be driven allows a much simpler amplifier circuit to be used. A single-ended transistor amplifier is mounted inside a box so that it "floats" with the common mode potential. The capacitance between the + input and ground is $\sim$10 pf and the capacitance between the - input and ground is $\sim$50 pf (40 pf of which is the capacitance between the pulse transformer primary and the interwinding shield). When the amplifier is connected to a probe, the probe lead with greatest capacitance to its shield should be used as the + input. The 2-25 pf trimmers at the input allow the capacitances to be brought into balance. In some cases, additional capacitance must be added from the + input to ground.

The common mode is read off the + input, through an attenuator, with a Tektronix P6028 (X1) probe or any other cable with capacitance less that $\sim$100 pf. The attenuation when used unterminated with a 1 meg. scope input is 5:1. When the amplifier is used alone, the common mode frequency response is linear to $>$10 mc. When used with a probe, the gain falls off at high frequencies (see Figure 7).

Gain Linearity

The transistor was purposely biased so as to handle signals between -10 volts and +0.2 volts at the input in order to reduce drain on the battery. Output voltage vs. input is plotted in Figure 8. The curve remains linear to below -10 volts. The amplifier gain is ~ 2 (because of the 1:2 ratio at the pulse transformer). With an attenuated probe, the overall gain of the system is $\sim 5 \times 10^{-3}$. Therefore, one would expect the system to saturate at difference voltages of ~ 40 volts. By rotating the probe 180°, however, the saturation point is ~ 2 KV.

If one desires to extend the linear range further into the positive region, the value of the 100K resistor in Figure 6 should be lowered. For example, a 33K would extend the linear range by a factor of three, while also increasing the battery current by three.

The probe-amplifier gain for the common mode is $\sim 6 \times 10^{-4}$. When used with a Tektronix type L preamp., the CM sensitivity is about 10 volts/cm. on the most sensitive scale.

Battery Voltage

Neither the amplifier gain nor the frequency response is affected by the battery voltage. However, the signal level over which the amplifier is linear is directly proportional to the battery voltage.

The amplifier requires about 1 ma. at 22 volts. Using a Mallory RM-412 mercury battery, rated at 1600 ma.-hours, we would expect a lifetime of over one year, assuming four hours of operation per day. There is reason to believe that the battery life is even longer when used intermittently.

Future Improvements

The limitations of the present design which one would hope to overcome in the future are:

(1) Limited frequency response

(2) Low level signals at the output

(3) Poor CMRR at high frequencies

Three improvements immediately suggest themselves:

(1) Add a small capacitance ($\sim$ 1 pf) in parallel with each resistor at the end of the probe. This would make the attenuators more nearly ideal and thus help the frequency response and CMRR. It would also make possible a lower attenuation factor, giving more signal at the amplifier output. This change would necessitate an amplifier with a higher input impedance ($\sim$ 20K) and a larger signal handling capacity.

(2) Use a faster pulse transformer. A PE-5158 has a rated rise time of 20 nsec. (20 mc. cutoff). In addition, the inter-winding capacitance is about half that of a PE-5163 insuring a higher common mode rejection ratio. By terminating the output with a lower impedance, the CMRR could be further improved. This, however, would require an amplifier with a lower output impedance in order to drive the pulse transformer at low frequencies.

(3) Use a stage of amplification ahead of the emitter follower to amplify only the difference signal. The CMRR would be improved by a factor equal to the gain of the amplifier.

Each of these ideas are presently being studied, and it is hoped that a probe-amplifier system can be made with a bandwidth of at least 10 kc. - 20 mc.

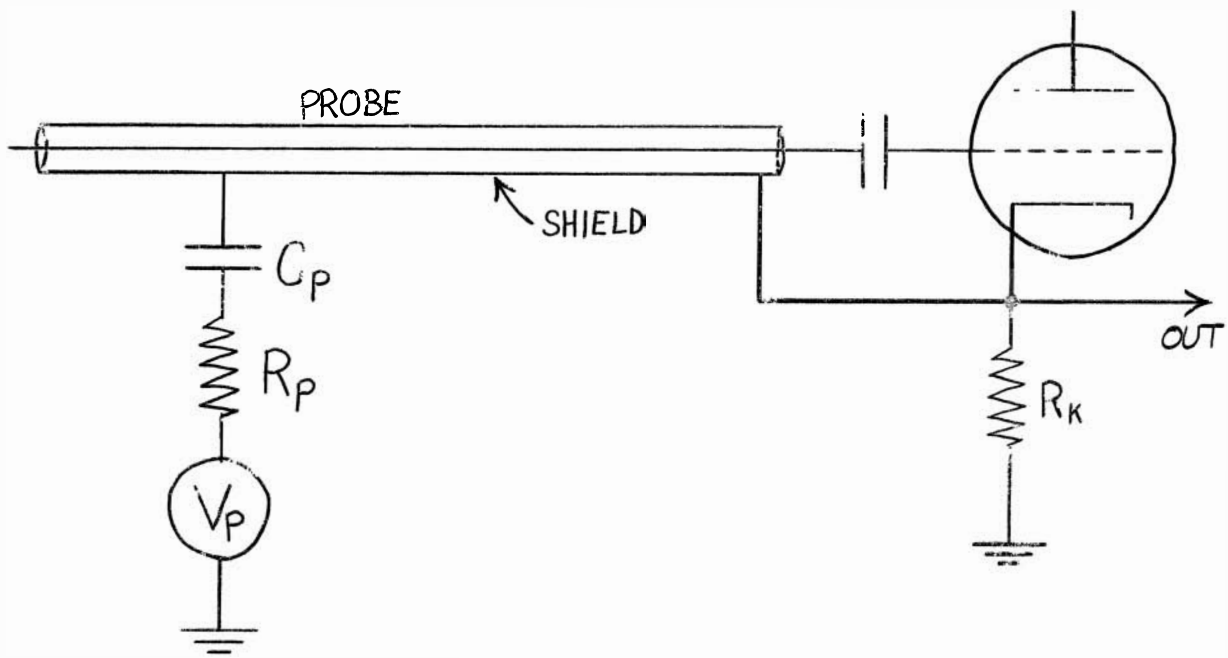

FIG 1

PLASMA COUPLING TO PROBE SHIELD

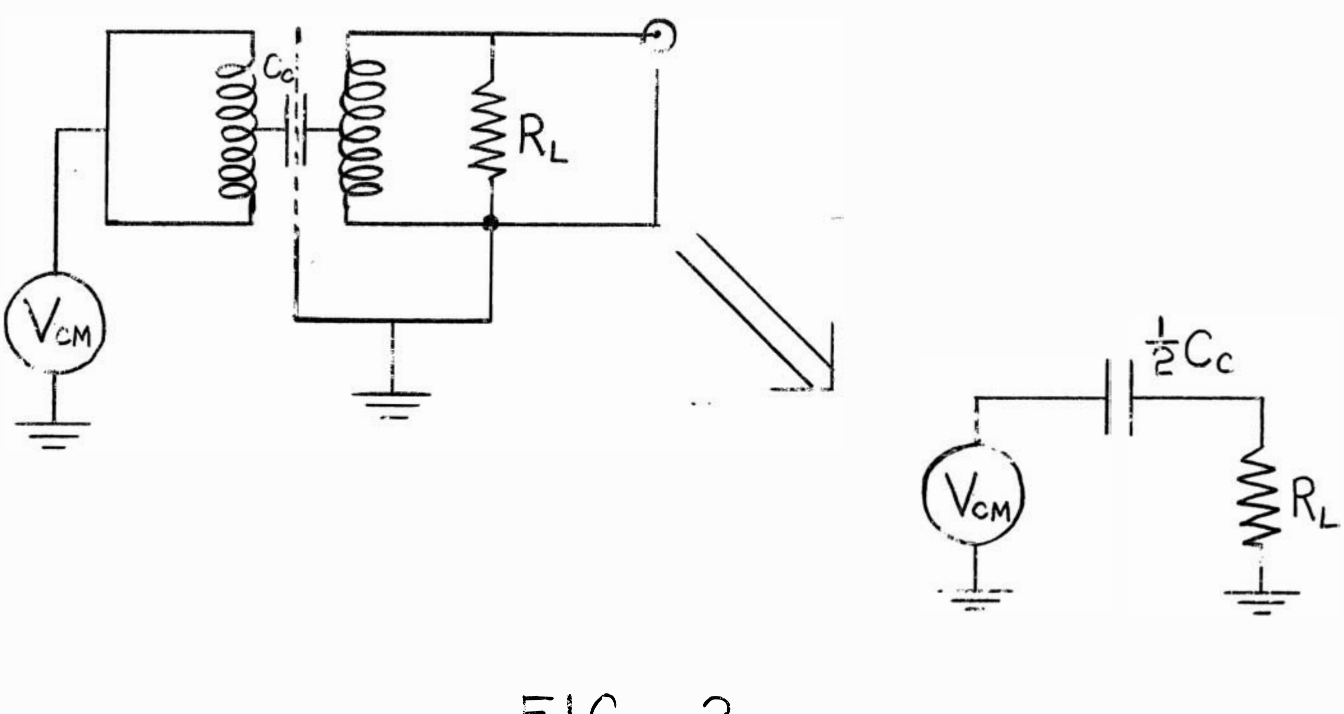

FIG 2

CAPACITIVE COUPLING BETWEEN TRANSFORMER WINDINGS

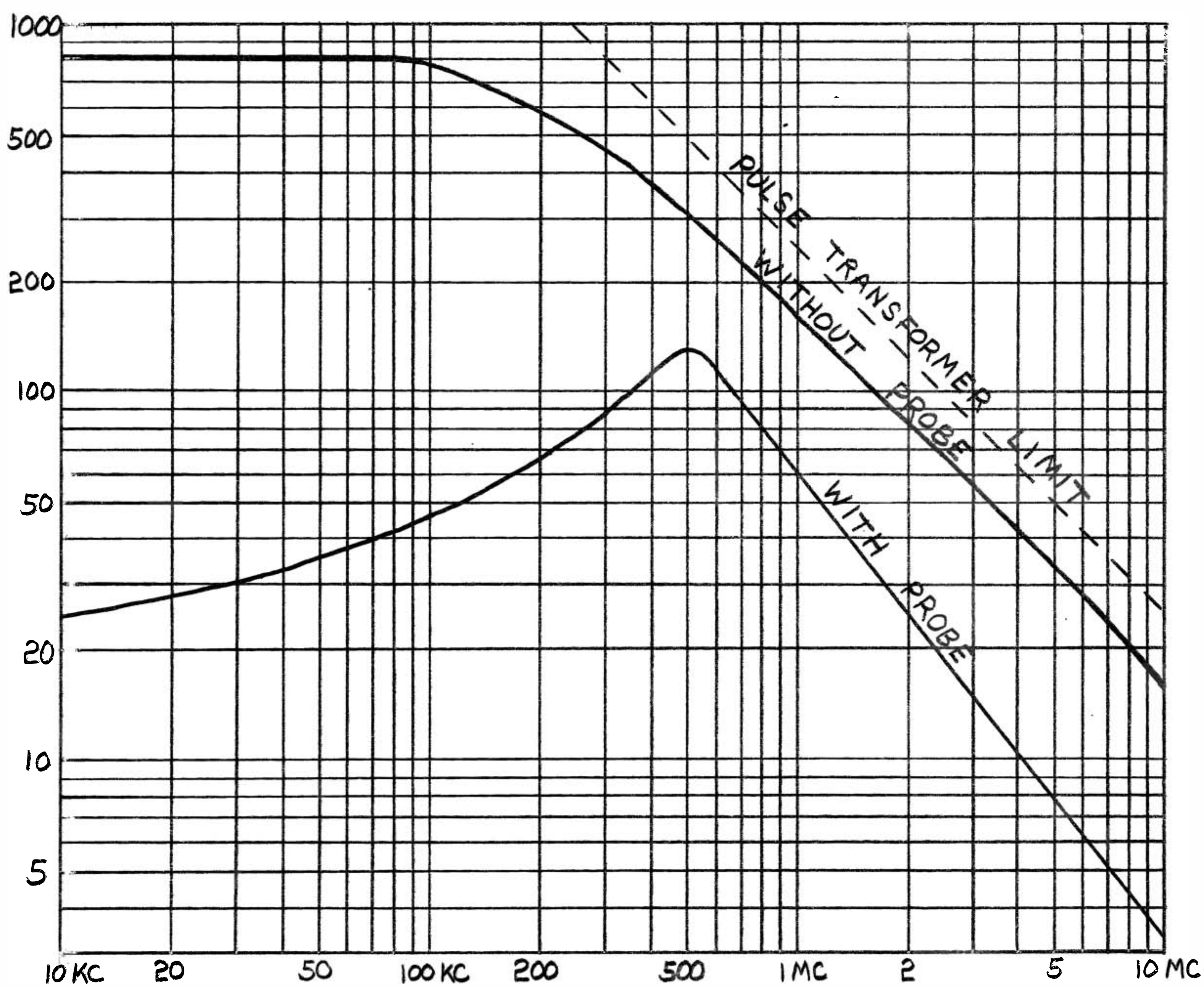

FIG 3 COMMON MODE REJECTION RATIO

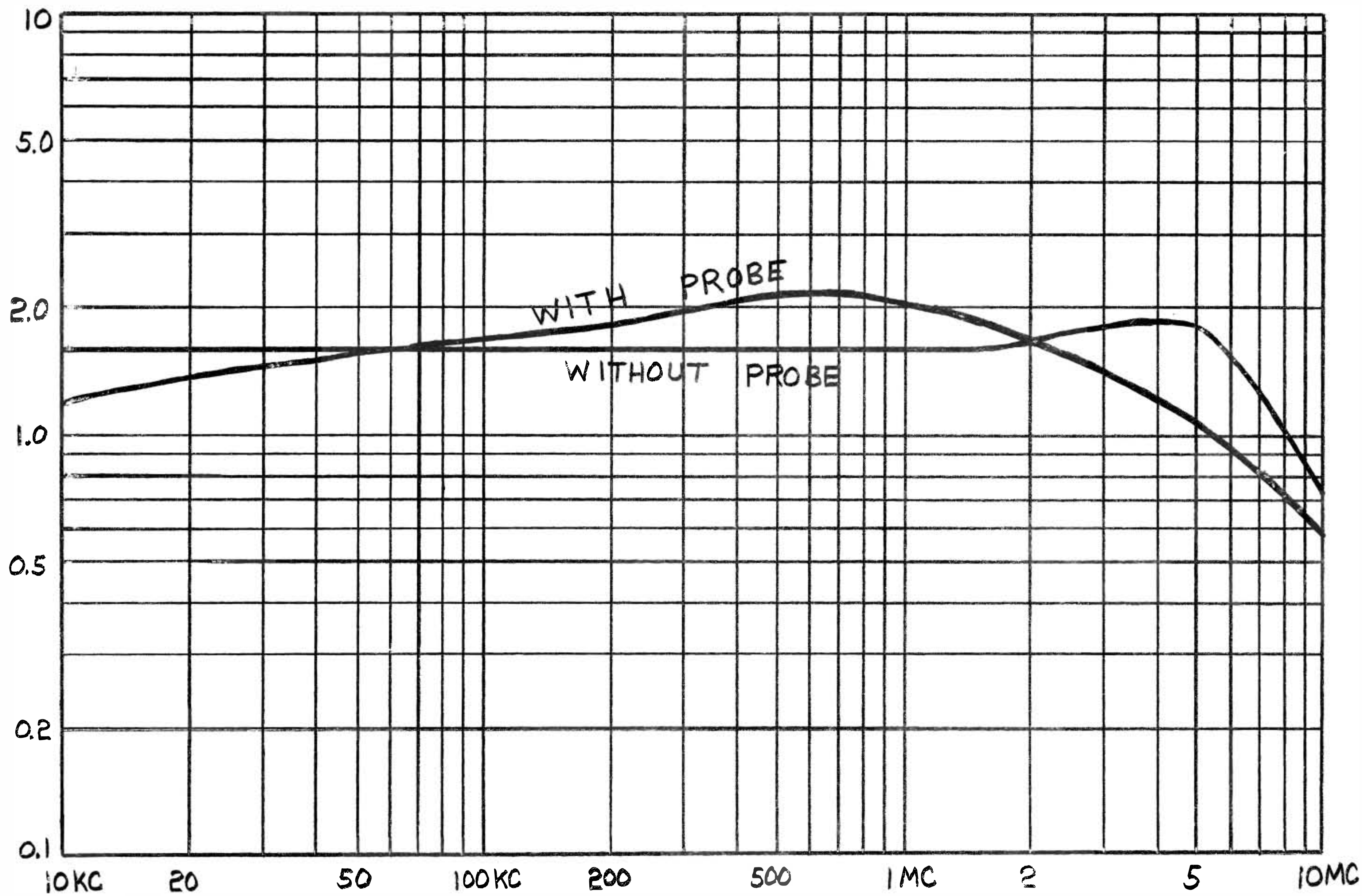

FIG 4 DIFFERENTIAL FREQUENCY RESPONSE

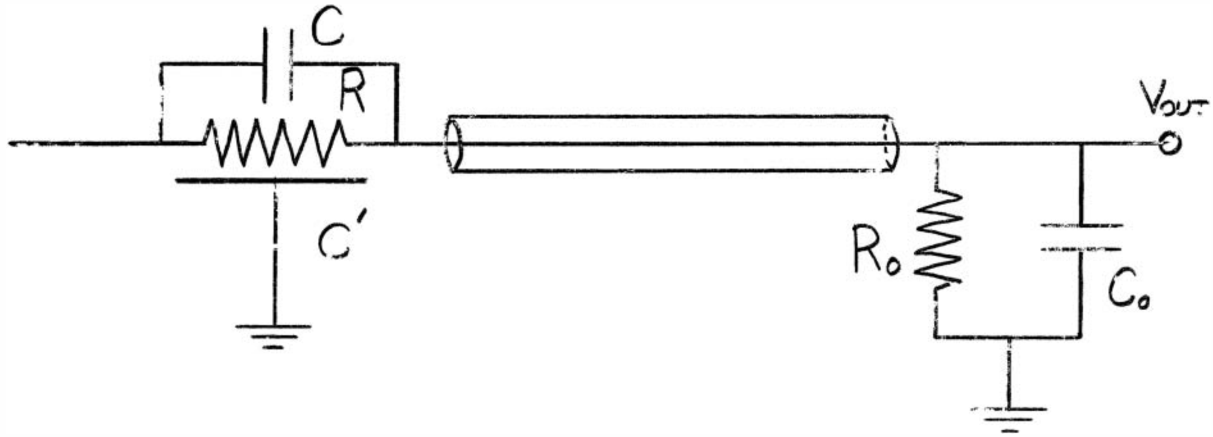

FIG 5

EQUIVALENT CIRCUIT FOR A NON-IDEAL ATTENUATOR

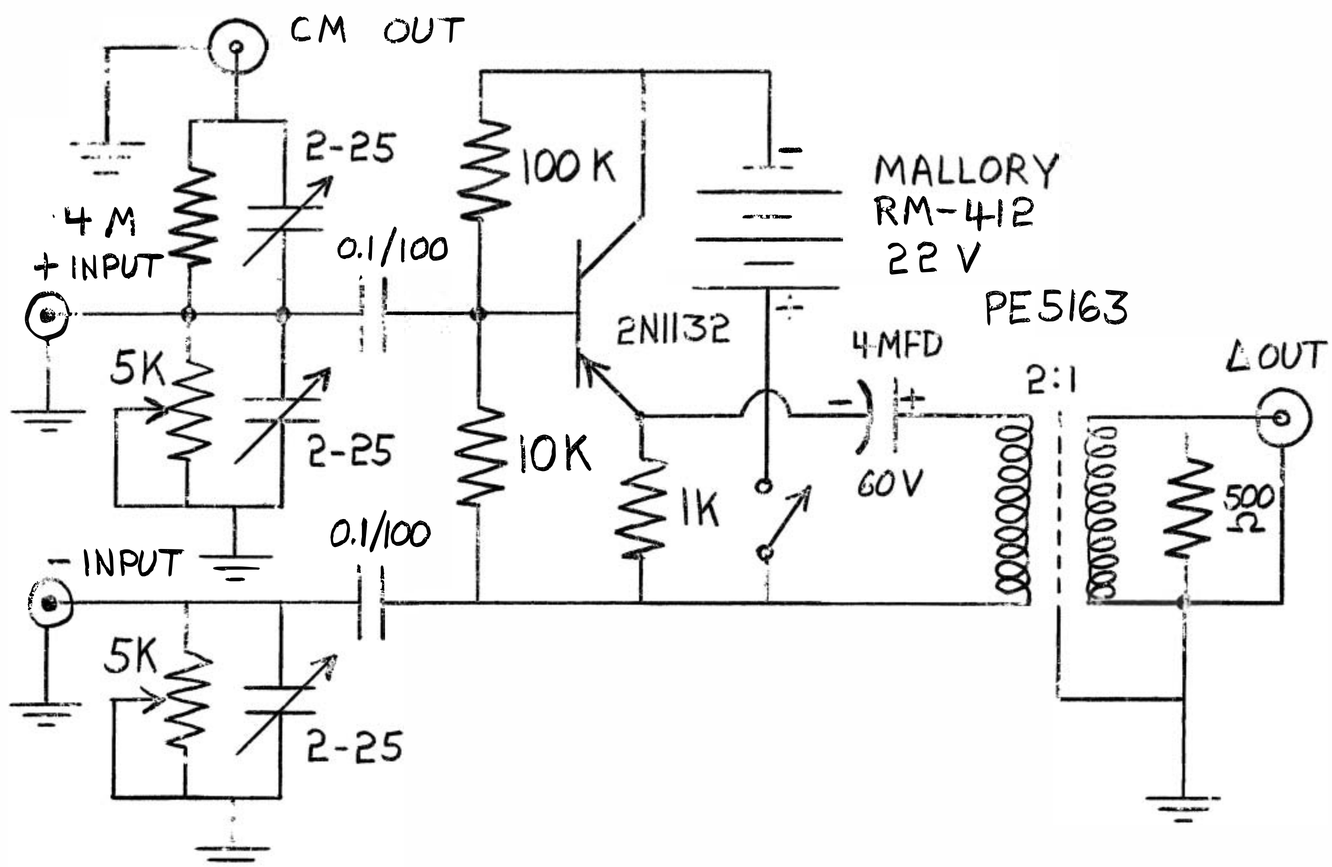

FIG 6

DIFFERENTIAL AMPLIFIER

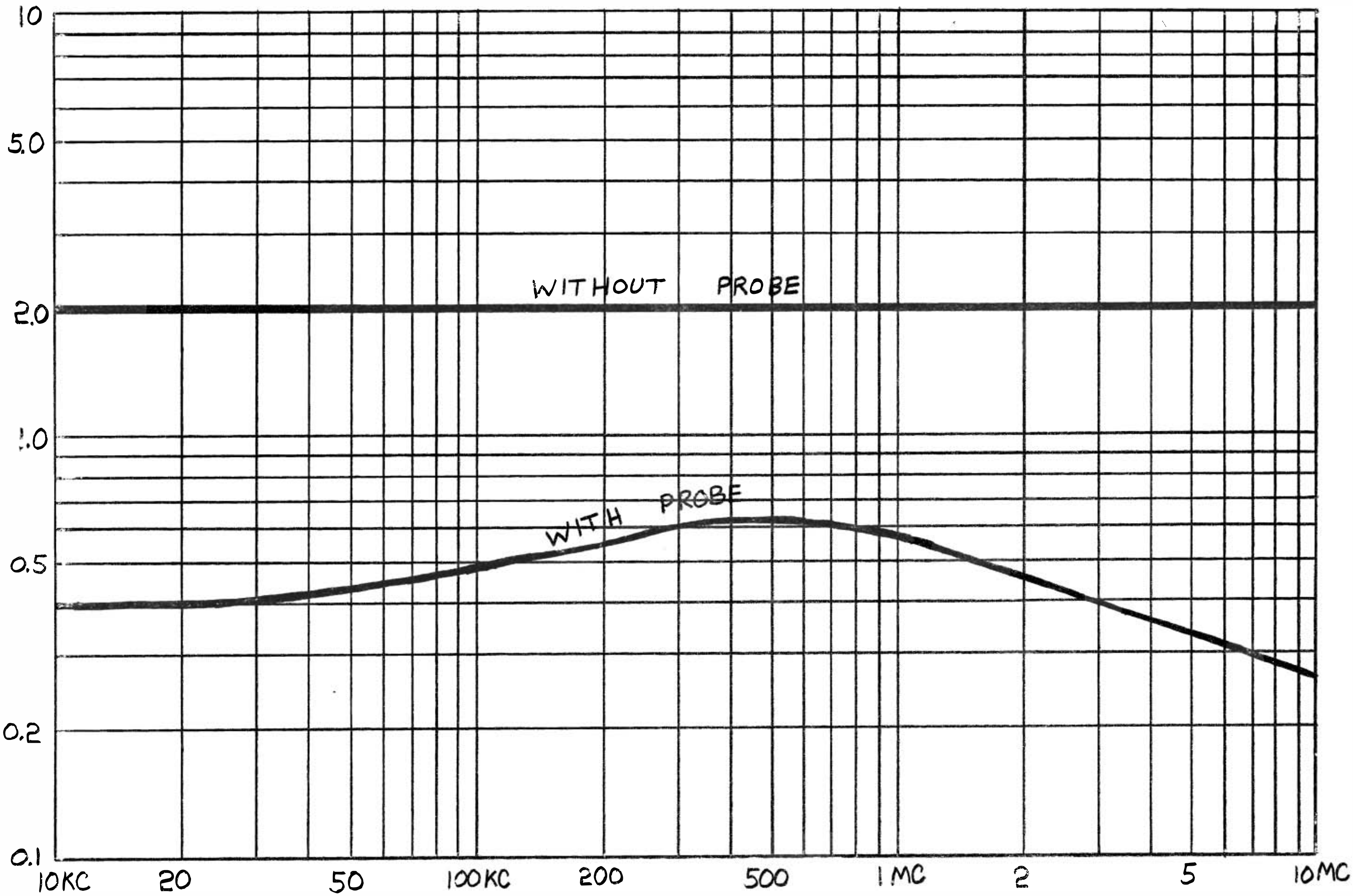

FIG 7 COMMON MODE FREQUENCY RESPONSE

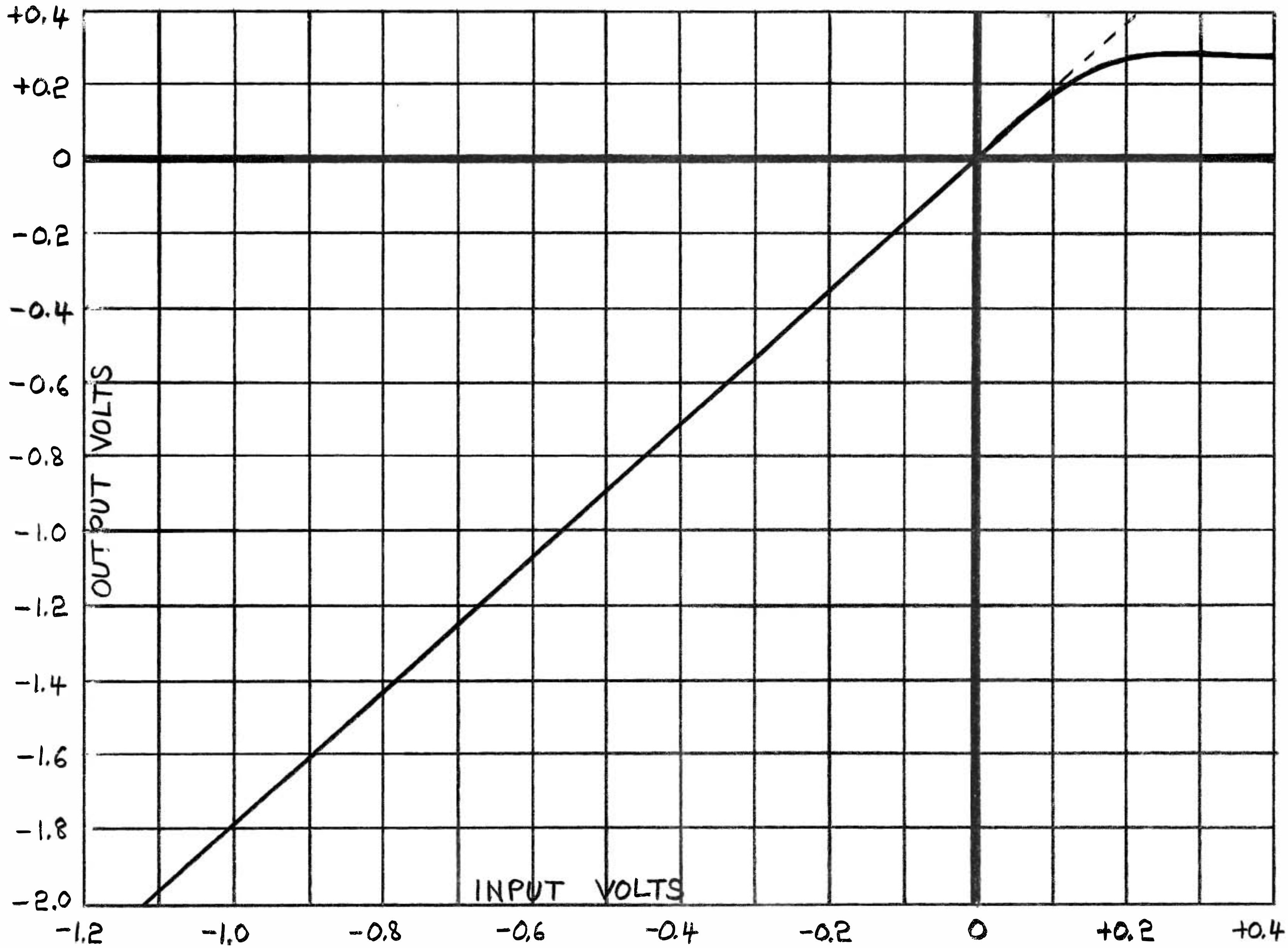

FIG 8 GAIN LINEARITY