

Now, the noise figure of any active 2-port network is given by

$$n.f. = n.f._{min} + \frac{R_n}{R_e(Y)} |Y' - Y'_{on}|^2 \quad (3)$$

By solving a set of four such equations corresponding to four different source admittances Y , the four noise parameters can be obtained.

P and R depend on the device technology and can be theoretically approximated.⁴ We preferred to optimise their values from measured values of $n.f._{min}$ and Y_{on} at 5 and 10 GHz. The values of P and R obtained were, respectively, 1.03 and 0.07. Introduction of the same noise generators in c.s. and c.g. equivalent circuits shows how the noise parameters

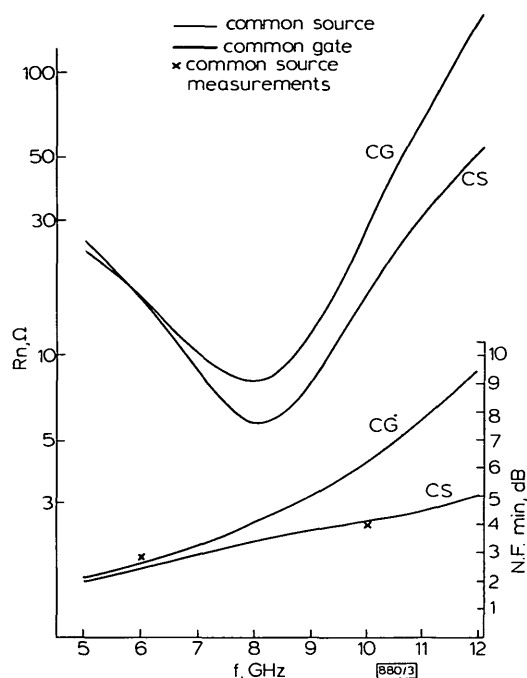


Fig. 3 Minimum noise figure $n.f._{min}$ and noise equivalent resistance R_n against frequency in common-source and common-gate configurations

NOVEL METHOD FOR AMPLITUDE MEASUREMENT OF SIGNALS UP TO MICROWAVE FREQUENCIES

Indexing terms: Microwave measurement, Rectifying circuits, Voltage measurement

The method gives very good input/output linearity down to input amplitudes of 100 mV; by using germanium diodes, this could probably be extended down to 10 mV or even lower.

There are many well known methods for precision rectification of alternating signals at low frequencies, notably those utilising the high gain of operational amplifiers to effectively eliminate the forward voltage drop of the rectifying diodes by enclosing them in a feedback loop.¹ While these methods can, in principle, be applied at high frequencies, it is impracticable to achieve accuracy beyond about 1 MHz. At frequencies above this it is usual to use a simple halfwave rectifier employing a low-capacitance germanium diode, this material being chosen for its low bandgap, resulting in acceptable detection linearity at low signal levels. Various methods are used to improve the linearity, including prebias and special linearisation circuits.² However, these can only achieve an approximation to the inexact amplitude response of the diode, and rarely take into consideration the effects of temperature variation.

The method described here requires only the availability of a pair of diodes having closely matched characteristics, and preferably sharing the same thermal environment. It might at first seem that a circuit such as shown in Fig. 1 might be used, the principle being that at small signal levels the low conductance of the diode is compensated by a matched device in the

R_n , $n.f._{min}$ and Y_{on} vary with frequency. They are plotted on Figs. 2a and 3. The main results are

(a) The minimum noise figure $n.f._{min}$ is greater in the c.g. configuration at frequencies above 5 GHz. The increase relative to the c.s. configuration is 3 dB at 11 GHz.

(b) The optimum noise source admittance Y_{on} is similar in both configurations and does not correspond to variations of $|S_{11}|$.

Explanation of results: The results obtained with this circuit model explain the deviations between experimental results and theoretical expectations based on low-frequency calculations. The low-frequency model¹ is based on the assumption that

$$|(1/R_0) + j\omega C_0| \ll |g_{om}| \quad (4)$$

i.e. $|Y_{as}| \ll |Y_{dg}|$ in notations of Reference 1. This is no longer valid above 6 GHz, owing to the influence of the capacitance C_0 . At 12 GHz, for example, the first term is $25 \times 10^{-3} S$, while the second is $18 \times 10^{-3} S$.

This model shows why the minimum noise figure of a GaAs f.e.t. is intrinsically higher in the common-gate than in the common-source configuration at microwave frequencies and why the decrease of $|S_{11}|$ does not correspond to a noticeable change in the optimum noise source admittance.

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feedback path of the operational amplifier. This certainly provides a measure of improvement in linearity and temperature stability, but the method is again inexact, because the rectifying diode is actually being cycled over a large range of its nonlinear characteristic, while the compensating diode operates at a constant current. The novel feature of the method now described is that both the diodes operate under the same waveform conditions; the 'reference' diode typically receives a variable-amplitude sinewave at 1 kHz, but any waveform and frequency (the latter invariably lower than the input signal) may be used.

Fig. 2 shows a typical embodiment of the method. The input signal (assumed here to be a sinusoid $E_1 \sin \omega_1 t$) is rectified by D_1 and its associated load R_1-C_1 and applied as one input to the error integrator A_1 , the other input being the output of the rectifier $D_2-R_2-C_2$. The multiplier M_1 receives as one input the reference carrier $E_2 \sin \omega_2 t$ and the integrated difference signal as the other. Only 2-quadrant operation is needed from this multiplier, but, in practice, it will be convenient to use a complete, fully trimmed 4-quadrant device, and the additional diode D_3 is included to prevent

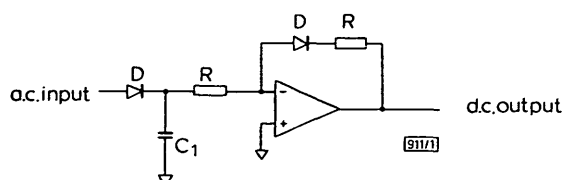


Fig. 1 Possible method for linearisation of detector

an incorrect operating mode which can arise owing to the phase-insensitive nature of the rectifiers.

This loop forces the output of the multiplier to be equal in amplitude to the input signal, and, in fact, it is a time-translated signal $E_1 \sin \omega_2 t$ which may now be handled by a precision low-frequency measurement system. Alternatively,

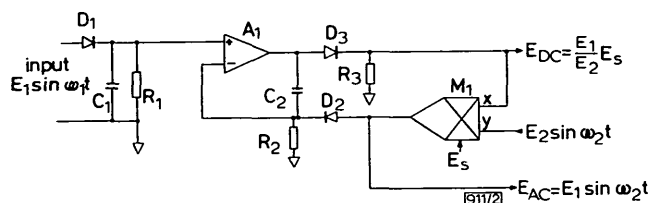


Fig. 2 Proposed method of linearisation

Both d.c. and a.c. outputs are provided with high accuracy

a d.c. output may be taken from the point shown, which provides a loadable, low-impedance node. A noteworthy feature of this output is that the amplitude of the carrier appears in the denominator of the expression

$$E_{DC} = \frac{E_1}{E_2} E_s$$

where E_s is the scaling voltage of the multiplier (typically 10 V). Thus, in some applications, a form of automatic sensitivity control is available.

The method is applicable to virtually any detector, but some care is needed in those cases where the diodes have a low breakdown voltage. If a large carrier amplitude is used, this can result in anomalous operation. When using diodes having a large forward voltage drop, the signal generated at the input to the error amplifier A_1 becomes very small at low signal levels, and for accurate operation this amplifier should have a low voltage offset. It is also desirable that the multiplier has small offset voltages.

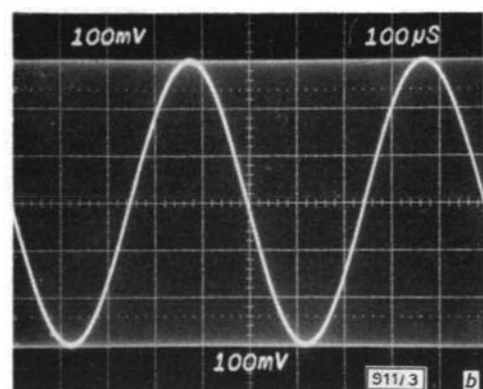
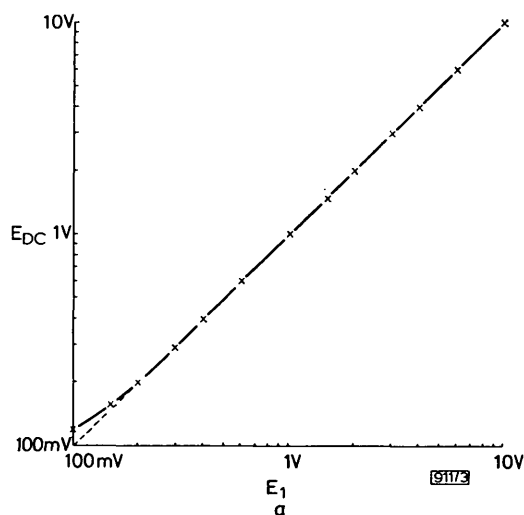


Fig. 3

a Measured amplitude response using hot-carrier diodes

b The a.c. output at 2 kHz is here shown to closely match the amplitude of the 5 MHz input

The performance has been verified experimentally, using Hewlett-Packard 5082-2826 hot-carrier diodes, an Analog Devices' AD510 laser-trimmed op-amp for A_1 and an AD534 laser-trimmed multiplier for M_1 , with $R_1 = R_2 = 10 \text{ k}\Omega$, $C_1 = 1000 \text{ pF}$, $C_2 = 0.1 \text{ }\mu\text{F}$, $D_3 = 1\text{N}4148$ and $R_3 = 5 \text{ k}\Omega$. A carrier amplitude of 10 V at 2 kHz was used. Very good linearity down to input amplitudes of 100 mV was possible, at which voltage the diodes are only slightly conducting ($I_F \approx 0.1 \text{ }\mu\text{A}$). Fig. 3a shows the input/output amplitude response over two decades; using germanium diodes, this could probably be extended down to 10 mV or even lower. Fig. 3b shows the good fit of the low-frequency signal to the incoming 5 MHz signal.

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LOW-LOSS PROPAGATION IN THE DIPOLE-MODE ALONG A COAXIAL WAVEGUIDE

Indexing terms: Waveguides, Waveguide theory

A theoretical investigation has been made as a step towards defining more precisely the operating conditions of a fully screened dipole-mode waveguide, comprising a coaxial structure whose inner consists of a cylindrical grid of parallel longitudinal wires separated by air and whose outer is represented by a concentric tube of continuous metal.

In the absence of the outer, the wire-grid structure (Fig. 1) has been shown¹ to support dipole-mode propagation and to exhibit falling axial attenuation α accompanied by rising radial attenuation a when the number of conductors forming the wire grid is increased and their spacing reduced. In these circumstances, detailed examination shows that, for a radial propagation coefficient $u = a - jb$, we have $a \gg b$ (typically, a may be 4 or 5 nepers/m) and, in general, the field outside the surface of the guide decays rapidly while transmission losses

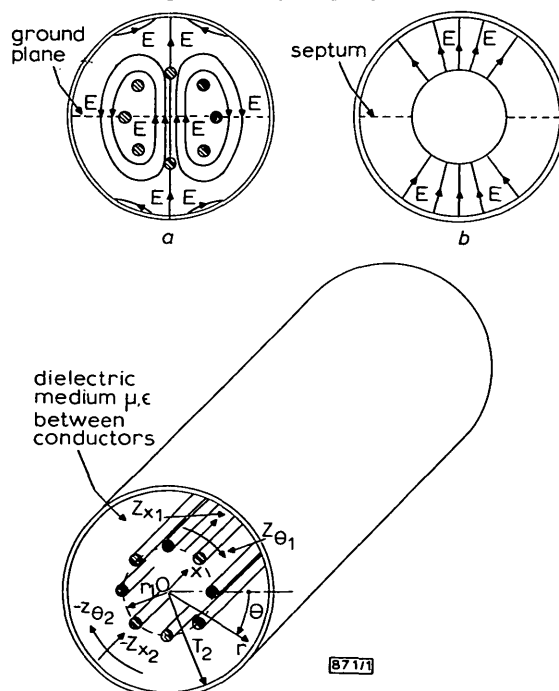


Fig. 1 Fully screened dipole-mode waveguide with wire-grid inner

Approximate transverse electric field in coaxial guide supporting (a) dipole mode and (b) septum mode