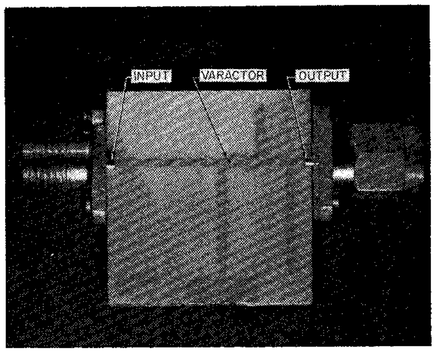
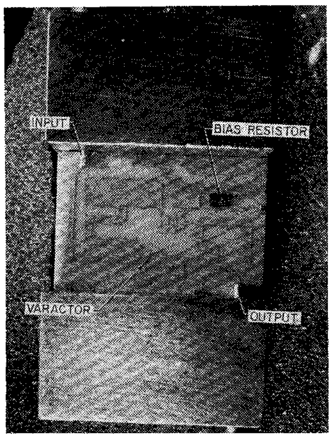
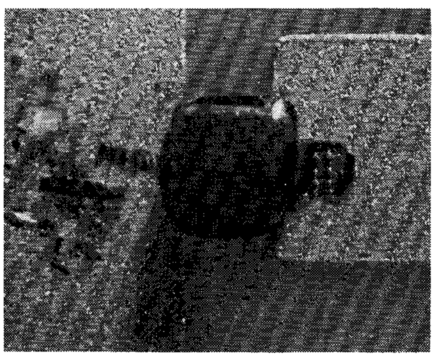


(a)



(b)



(c)

Fig. 3. (a) The 2.125 to 8.500 GHz quadrupler—unfolded model. (b) The 2.125 to 8.500 GHz quadrupler—folded model. (c) The beam lead varactor mounted in microstrip.

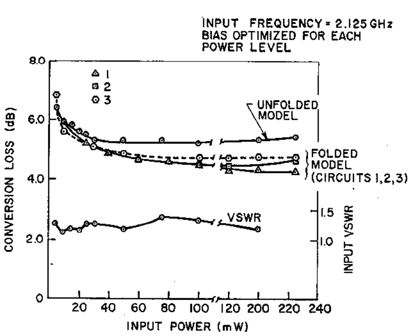


Fig. 4. Input VSWR and conversion loss versus input power for four quadrupler circuits.

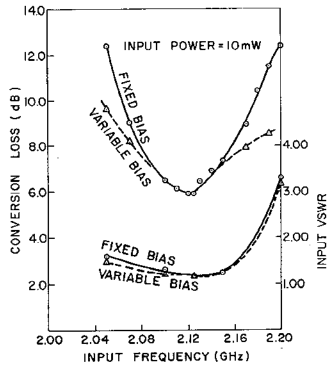


Fig. 5. Input VSWR and conversion loss versus frequency for 10 mW input power.

narrow bandwidth, the feasibility of building quadruplers with reasonable conversion efficiency using microstrip integrated circuit fabrication techniques has been demonstrated. The use of beam lead varactors has contributed significantly to this success.

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Cooling of Avalanche Diodes

This correspondence presents the experimental results of cooling reverse-biased p-n junctions operating in an avalanche transit time mode, and is an extension of previous efforts where the generation and amplification of microwave energy was obtained from conventional varactor diodes.[1]–[4] These are commercially available epitaxial GaAs and Si varactors. The diodes were originally designed for operation as either low-noise parametric amplifiers or harmonic generators. Typically, the diodes exhibit breakdown voltages in excess of 25 volts and junction capacitances less than 1 pF at zero bias.

Basically, the experiment consists of immersing a cold finger attached to the diode through the waveguide mount into a reservoir of liquid nitrogen. Performance was then measured in a *Ku*-band system.

The microwave mount used for these experiments is shown in Fig. 1. It consists of a split guide design insulated at both corners and flange faces when connected to the waveguide system.[5] The diode is mounted on a variable diameter post to which a copper cold finger is attached. The dc bias is directly applied to the diode by means of the bias posts, eliminating the need of coaxial circuitry featuring folded chokes and bypass capacitors to prevent RF leakage. This design has been used extensively as a waveguide circuit for CW operated avalanche transit time devices at this Laboratory.

The dependence of RF output power on reverse-bias current is illustrated in Fig. 2. These curves represent performance of these devices when cooled and uncooled. Because of poor thermal resistance and junction heat sinking, the bias levels of uncooled GaAs units were limited to 10 mA (typically 500 A/cm^2) to prevent burnout. Cooling had the effect of lowering the junction temperature, thus extending the range of operation in this case by 4 mA. With this increase in operating level, a corresponding increase of RF power from 30 to 50 mW was obtained. The operating range of the Si devices was extended from 30 to 40 mA with a corresponding increase in RF power from 28 to 50 mW. In both the GaAs and Si devices the output power did not appear to saturate at the above levels of operation. It is interesting to note that when the cooled devices were operated at the uncooled current levels, the GaAs oscillator showed a slight decrease in RF power output, and the Si devices showed a slight increase. These results were typical for many devices tested.

A plot of the dc to RF efficiency as a function of reverse-bias current is shown in Fig. 3 for the same operating conditions as Fig. 2. The GaAs oscillator shows a net increase of 20 percent in efficiency from 6.3 to 7.5 percent for the 4 mA increase in current. For less efficient GaAs oscillators, the net increase in efficiency is more dramatic; typically an uncooled diode which is 3 percent efficient at 10 mA will be 6 percent efficient at 15 mA. The Si oscillator showed a 50 percent net increase in efficiency from 1.1 percent at 30 mA to 1.7 percent at 40 mA. Saturation effects are dominant for both devices although the RF power out versus the reverse current of Fig. 2 did not indicate any saturation mechanism.

Cooling also improved the performance of these diodes when operated as avalanche amplifiers. Negative resistance amplification was obtained from both the GaAs and Si diodes when used in a circulator-coupled network which formed a one-port reflection type amplifier. The RF power input versus RF power output is shown in Fig. 4 which illustrates the performance of a cooled and uncooled GaAs device. Although amplification was also obtained from Si devices, the results in terms of noise figure, gain, tunability, and stability were not comparable to the GaAs results. In general, the overall response for the cooled amplifier was superior to the uncooled operation. Several points are of interest on this curve. First, gains in excess of 10 dB are obtainable at 0 dBm input level. The room temperature amplifier yielded 17.8 mW output power for 1 mW input when operated at a reverse current of 9.0 mA. The amplifier was then cooled and the reverse current was raised to 15 mA. Under these new operating conditions the output power increased to 38.0 mW for 1 mW of input power. The second point of interest for local oscillator applications is the input level at which 1 mW of output power can be obtained. For the uncooled amplifier this value corresponds to -38 dBm, and for the cooled device it is -52 dBm. The effects

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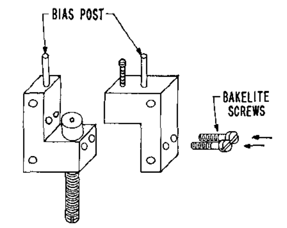

Fig. 1. Split guide mount featuring cold finger.

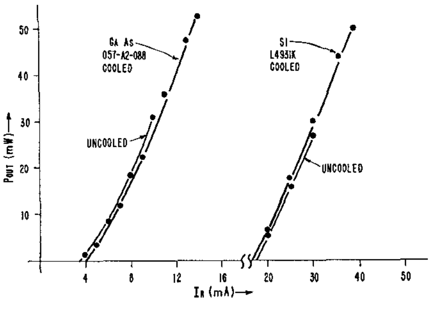

Fig. 2. Variation of RF output power with bias current for both cooled and uncooled operation.

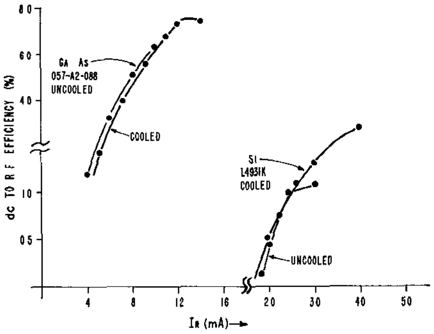

Fig. 3. Oscillator efficiency as a function of reverse-bias current for cooled and uncooled operation.

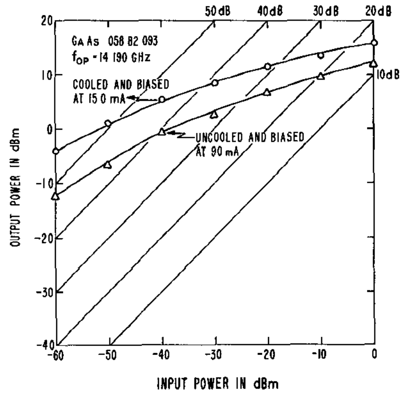

Fig. 4. Gain as a function of input power for GaAs avalanche amplifier.

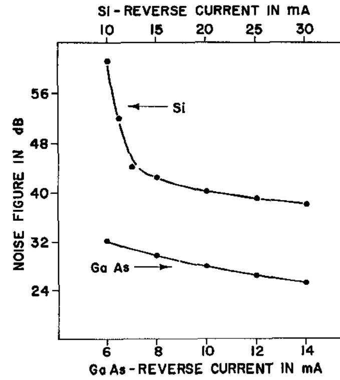

Fig. 5. High-gain noise figure as a function of reverse current.

of cooling resulted in a 14 dB increase in gain at this point. The bandwidth was measured at the 0 dBm input level and was found to be 69 MHz.

Reflection amplifier noise figure measurements were also made under high-gain conditions (10 to 30 dB). Fig. 5 shows the dependence of the noise figure on avalanche current as measured in identical circuits for a Si and GaAs varactor diode having identical avalanche frequencies and the same magnitude of negative conductance. The GaAs reverse current was varied from 6 to 14 mA, and the corresponding noise figure varied from 32 to 25.5 dB. The Si amplifier had a 61 dB noise figure at 10 mA and 38 dB at 30 mA. Between 10 and 12.5 mA of reverse current, the Si device was near the threshold of amplification, giving an excessively high-noise figure. In general, the GaAs device is 13.0 dB less noisy than the equivalent Si device.

Significant improvement in the operation of avalanche transit time amplifiers and oscillators is obtainable by cooling.[6] At room temperature the effects of excessive junction heating limits the performance of these devices. Cooling has the effect of extending the operating range of these devices which resulted in improved oscillator output efficiencies, and improved amplifier gain and noise figure. The need for good thermal heat sinking and good thermal resistance of junctions seems to be apparent. Then, with these improvements, optimum operation comparable to the cooled results should be derived from uncooled devices.

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Broadband Fixed-Tuned Acoustic Delay Lines

This correspondence is concerned with the design of broadband fixed-tuned delay lines to operate at *L*- and *S*-band frequencies. These delay lines utilize layer transducers incorporating CdS thin films of the type used by deKlerk,[1] sapphire delay media, and suitable electrical matching networks in either coaxial or in strip transmission line.

The impedance seen by an electrical network terminated in the transducer shown in Fig. 1(a) has been obtained in the past[2]–[6] by solving the piezoelectric equations subject to the indicated boundary conditions. The impedance in Evans' notation neglecting dielectric losses in the transducer is

$$Z = \frac{-jL}{\omega A \epsilon} + \frac{k^2 V}{\omega^2 A \epsilon} \cdot \left[\frac{\frac{2\xi_0}{-\xi_2}\left(1 - \frac{1}{\cos \beta L}\right) + j \tan \beta L}{1 + j \frac{\xi_0}{\xi_2} \tan \beta L} \right]$$

which may be written, by breaking into real and imaginary parts, as

$$Z = R(\omega) - jX_c - jX(\omega)$$

where the real part of the impedance $R(\omega)$ may be termed the acoustic radiation resistance, X_c is the normal capacitive reactance, and $X(\omega)$ is a small additional reactance. For many transducer configurations using CdS in the gigahertz range, $R(\omega)$ may be as small as 0.01 ohm making possible only loose coupling from electromagnetic to acoustic energy. Correspondingly large mismatches are obtained. This situation may be altered by adjusting the ratio ξ_0/ξ_2. Since ξ_2 is the acoustic impedance seen looking into the right electrode of Fig. 1(a), suitable metallic layers may be used to adjust the value of ξ_2. Fig. 1(b) is a plot of values of $R(\omega)$ and $X(\omega)$ versus normalized frequency for several values of ξ_0/ξ_2. Here a normalizing factor $(\pi/4k^2)(f/f_0 X_c)$ has been used to allow for general transducer materials. The relationships of Fig. 1(b) are useful in bandwidth considerations indicating that a device having a large $R(\omega)$ yields a narrowband device. For cases in which ξ_2 is obtained from multilayer electrode arrangements, the ratio ξ_0/ξ_2 will exhibit a frequency dependence and the relationships will be modified.

In order to design broadband delay devices at *L*- and *S*-band frequencies, a computer program was prepared in which either a two- or three-step Chebyshev electrical transformer is terminated by the acoustic transducer of Fig. 1(a). A VSWR ripple of 1.70 to 1 was selected when the network is terminated by the real part of the impedance $R(\omega)$, and provision was made in programming where the impedance ξ_2 would represent a general impedance formed by the acoustic delay media and suitable metallic or dielectric layers.

Multilayer electrode configurations have

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