Tunnel diodes provide a means of low-noise microwave amplification, with the amplifiers using the negative resistance of the tunnel diode to achieve amplification by reflection. The tunnel diode and its assumed equivalent circuit are discussed. The concept of negative-resistance reflection amplifiers is discussed from the standpoints of stability, gain, and noise performance. Two amplifier configurations are shown, of which the circulator-coupled type is carried further into a design for a C-band amplifier. The result is an amplifier at 6000 mc/s with a 5.5-db noise figure over 380 mc/s. An X-band amplifier is also reported.

G. T. Munsterman

Tunnel-Diode Microwave Amplifiers

Recent advances in tunnel-diode fabrication techniques have made the tunnel diode a practical, low-noise, microwave amplifier. Small size, low power requirements, and reliability make these devices attractive for missile application, especially since receiver sensitivity is significantly improved, with resulting increased homing time. Work undertaken at APL over the past year has resulted in the unique design techniques and hardware discussed in this paper.*

Low-Noise Microwave Amplification

The need for low-noise microwave amplification can be seen when the system noise figure is considered. The overall noise figure, F, of a system consisting of n stages is given by the following expression:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots = F_1 + \sum_{i=2}^{n} \frac{F_i - 1}{\prod_{j=1}^{i-1} G_j},$$

where the gain of the ith stage is denoted by G_i and its noise figure by F_i. This equation shows that stages without gain $(G < 1)$ contribute greatly to the overall system noise figure, especially if they are not preceded by some source of gain. If a low-noise-amplification device can be located near the source of the signal, the contribution from the successive stages can be minimized by making G_1 sufficiently large, and the overall noise figure is then that of the amplifier F_1. If low-noise amplification is not available until the second or third stages, then a high-noise figure results. This latter case pertains to many microwave systems in which down conversion is necessary before amplification can take place. The inclusion of tunnel-diode amplifiers in systems formerly using down conversion and then amplification results in noise figure improvements generally greater than 6 db. When tunnel-diode amplifiers are used instead of traveling wave tubes, they simplify greatly the required power supplies, reduce power dissipation, and make warm-up time negligible. It should be pointed out, however, that traveling wave tubes are superior in available gain and maximum output power. Thus, tunnel-diode amplifiers are usually considered only for receiver applications where signal levels are well below tunnel-diode saturation.

* The author wishes to acknowledge the assistance of E. E. Skelton and S. W. May of APL in the fabrication and testing of the tunnel diode amplifiers.

The Tunnel Diode

The discovery of the tunnel diode was made in the 1950's during research on back diodes, that is, on diodes whose reverse conduction is greater than forward conduction.[1] High reverse conduction can best be pictured as the limiting case of the zener breakdown. As theory had predicted, this reverse breakdown voltage could be decreased by higher doping concentrations of the semiconductor material. The result of these high impurity levels was not only an immediate reverse conduction, but unexpected negative resistance.[2] Figure 1A shows a typical tunnel-diode voltage-current characteristic. The high reverse conduction and negative resistance region can be compared to the conventional diode curve shown by the broken line.

Design Considerations

The concept of a negative resistance producing gain can best be shown by using simple transmission line theory. A transmission line of characteristic impedance Z_0 is terminated in an impedance of value Z_d. The loss of power because of reflection is given by the square of the voltage reflection coefficient, Γ. This coefficient is defined in terms of the impedances by the relationship

$$\Gamma = \frac{Z_d - Z_0}{Z_d + Z_0}\,.$$

For positive impedance this is less than unity, as expected. But if Z_d is a negative impedance, as a tunnel diode could be, a voltage reflection coefficient greater than unity can result. The power gain, or voltage reflection coefficient squared, is then greater than one (Z_d is now assumed negative):

$$\text{Gain} = |\Gamma|^2 = \left|\frac{-Z_d - Z_0}{-Z_d + Z_0}\right|^2 = \left|\frac{1 + Z_0/Z_d}{1 - Z_0/Z_d}\right|^2\,.$$

The ratio of Z_0/Z_d for high gain is close to one.

For a perfect match ($Z_d = Z_0$), oscillation results. The necessity of impedance control is then the main problem of designing tunnel-diode amplifiers. For a certain desired gain a certain impedance ratio must exist. For a given bandwidth this ratio must be constant over the band; for stability it cannot be unity since unity match represents an unstable condition. Tunnel-diode amplifiers are usually designed for less than 20-db gain to assure stable operation with expected diode resistance variation.

A low-impedance DC source is required for stable biasing; this is illustrated in Fig. 1A. The two load lines shown represent the circuit resistance shunting the tunnel diode. The high-impedance load line intersects the voltage-versus-current curve of the tunnel diode in three places. The two positive

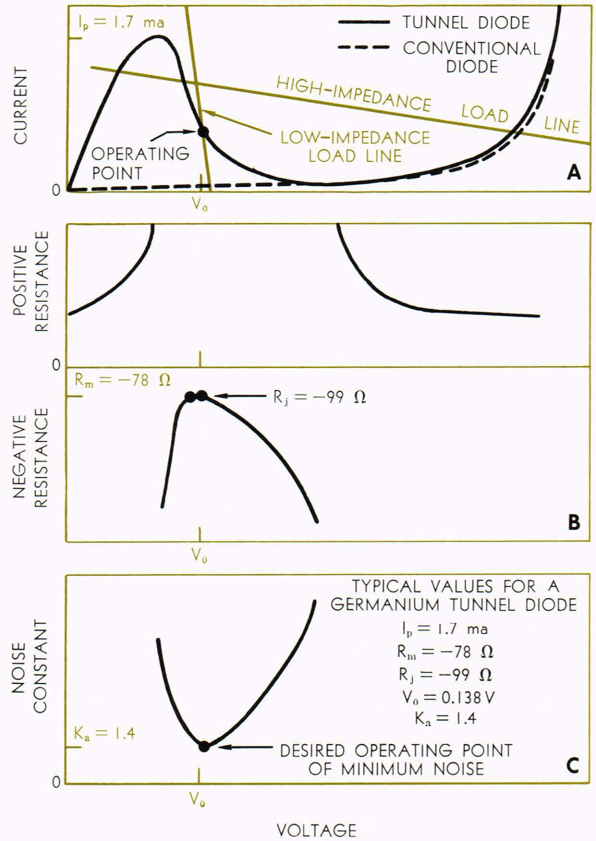


Fig. 1—Tunnel-diode bias voltage plotted against (A) current, (B) resistance, and (C) noise constant.

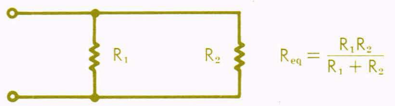
resistance intercepts are stable and are utilized in computer applications of the tunnel diode. For amplification a stable negative resistance is desired. A load line that results in a single valued bias point within the negative resistance region is also shown in Fig. 1A. The requirement for this load line is that the external circuit resistance (load line) be less than the magnitude of the minimum negative resistance, R_m, available from the tunnel diode. This minimum resistance is indicated in Fig. 1B.

[1] L. Esaki, "New Phenomenon in Narrow Germanium p-n Junctions," *Phys. Rev.*, **109**, 1958, 603-604.

[2] L. Esaki, "Fundamentals of Esaki Tunnel Diode in Circuit Applications," Monograph on Radio Waves and Circuits (ed. S. Silver) Elsevier Publishing Co., Amsterdam, 1963, 359-373.

The low-resistance shunt on the tunnel diode required for biasing introduces two problems. The first is that noise currents from this resistor must be isolated from the tunnel diode in order to preserve low-noise performance. Secondly, this resistance must be isolated from the tunnel diode at desired frequencies of amplification in such a way that a negative resistance is available for amplification.

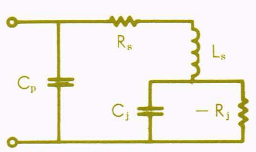
This second problem can be illustrated by considering the equivalent resistance, R_{eq}, of two parallel resistors, R_1 and R_2.



If R_2 is a negative resistance, R_{eq} will be negative provided R_1 is greater than $|R_2|$: If these resistors are isolated by an inductance, AC stability becomes a problem.

Tunnel-Diode Equivalent Circuit

The accepted equivalent circuit of a tunnel diode is shown below:



R_s is the resistance of the bulk material and contacting plates. It is minimized by using thin semiconductor pellets, high-conductivity packaging materials, and high semiconductor mobility. It is frequency-dependent because of the "skin effect."

R_j is the negative resistance of the tunnel diode. It is bias-dependent, as shown in Fig. 1B, and frequency-dependent only when carrier lifetime is approached; this frequency is about 10^6 mc/s. R_j is the predominant temperature-sensitive parameter.

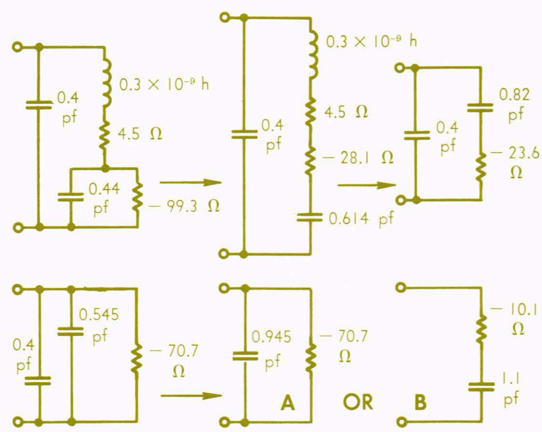
L_s is the parasitic series inductance. This is caused by the ribbon material used to connect the contact plate with the junction material. A typical minimum value for this inductance is 0.3×10^{-9} h. Recently announced packaging techniques indicate a marked reduction in L_s, but these claims remain to be evaluated.

C_p is the distributed capacity of the case. It is minimized by using small package areas and low-dielectric package materials.

C_j is the inherent capacitance of the junction. The shunting effect of this capacitance on the negative resistance limits the frequency range. This is easily seen if the parallel combination of R_j and C_j is put in a series-equivalent circuit. This is the first step in reducing the equivalent circuit to a two-element shunt network that is used in tunnel-diode amplifier design. Further reductions can be shown with an example. A typical tunnel diode for amplification at C-band, 5000 to 6000 mc/s, would have the following parameters:

$C_j = 0.44$ pf	$C_p = 0.4$ pf
$L_s = 0.3 \times 10^{-9}$ h	$-R_j = 99.3\Omega$†
$R_s = 4.5\Omega$	$-R_m = 78.3\Omega$

The equivalent circuit reduction at 5800 mc/s, is shown below:



Circuit B is appropriate for use when series tuning is employed. The design of a tunnel-diode amplifier using shunt tuning, for which circuit A is best, will be discussed in a later section.

Two frequencies, resistive and reactive cutoff, can be used in defining the practical frequency range of a tunnel diode and, because they are figures of merit, in tunnel-diode selection.

RESISTIVE CUTOFF FREQUENCY—The resistive cutoff frequency, f_{ro}, is that frequency at which the real part of the input impedance, neglecting C_p, is zero. This frequency, at which the parallel combination of R_m and C_j has a negative resistance equal to R_s, is the frequency limit of the negative resistance; C_p is neglected since it is assumed that this reactance is resonated by a shunt inductor;

† R_j is a negative resistance at a bias chosen for minimum noise figure.

R_m is the previously defined minimum negative resistance.

$$f_{ro} = \frac{1}{2\pi R_m C_j}\sqrt{\frac{R_m}{R_s} - 1}\ .$$

Reactive Cutoff Frequency—The reactive cutoff frequency, f_{xo}, is the frequency at which the input reactance to the diode is zero.

$$f_{xo} = \frac{1}{2\pi R_m C_j}\sqrt{\frac{R_m^2 C_j}{L_s} - 1}\ .$$

Tunnel diodes are usually operated below f_{xo} where the tunnel diode is capacitive.

The example of a diode given above has the following cutoff frequencies:

$$f_{ro} = 18{,}500 \text{ mc/s}$$

$$f_{xo} = 13{,}100 \text{ mc/s}$$

This diode, if used at 5000 to 6000 mc/s, would be operating well below reactive cutoff and would appear capacitive. The usual rule for operation of a tunnel diode at one-third resistive cutoff frequency is to insure adequate negative resistance as well as minimum degradation of the noise figure. Work at APL has shown that the one-third factor may be too conservative. Useful results have been obtained with ratios of 1/2. With the state-of-the-art in tunnel-diode manufacture at f_{ro} of 40,000 to 50,000 mc/s, the upper frequency for tunnel-diode amplification is 20,000 to 25,000 mc/s. Problems exist at higher frequencies since waveguide, with much higher line impedances resulting, must be used.

It has been postulated that operation of a tunnel diode above its resistive cutoff frequency is feasible.[3] A possible explanation is that the L_s and C_p of the diode equivalent could be considered a portion of the transmission line. This idea has not yet been proven practical.

Noise Considerations

One of the advantages of a tunnel-diode amplifier is its low noise figure. The sources of noise in a tunnel diode amplifier are:

1. Circuit thermal noise.
2. Noise from the bias resistor.
3. Noise from the diode series resistance.
4. Shot noise associated with tunneling.

Circuit impedances and mismatches are minimized by careful selection of components and design of the supporting structures.

As has already been explained, the bias resistor, if not isolated from the tunnel diode, may shunt the negative resistance sufficiently to make the combination a positive resistance. This effect is reduced by placing a capacitor across the resistor, which provides a low impedance at operating frequencies. This combination is then located a quarter wavelength away from the diode. The noise from the bias resistor is also reduced by this capacitance.

The series resistance of the diode bulk material is kept down to several ohms by precise fabrication techniques and by proper selection of materials.

The process of tunneling produces shot noise that is proportional to the negative resistance per unit current; that is, the greater the peak current the lower the shot noise. The equivalent circuit of the tunnel diode, when modified to include the two noise sources, can be reduced to a parallel equivalent of shunt conductances and current generators. These can be used to find the total noise contribution.[4] For high gain the following equation results:

$$F = \frac{1 + K_a}{\left[1 - \frac{R_s}{|R_j|}\right]\left[1 - \left(\frac{f}{f_{ro}}\right)^2\right]}\ ,$$

where $R_s / |R_j|$ is usually much less than 0.1. If the diode is operated well below f_{ro}, then the term $(f/f_{ro})^2$ contributes very little to the overall noise figure. The factor K_a (noise constant) then becomes the property of prime interest. This parameter is dependent on both negative resistance and current.[5] Germanium tunnel diodes have noise constants from 1.2 to 1.4, while other materials are capable of lower noise performance but are more temperature-sensitive than germanium. The noise constant versus bias voltage is shown in Fig. 1C. In this figure it is paralleled with the current characteristic and resistance characteristic. The desired operating point is determined by finding the minimum noise figure. The amplifier design is then based on the negative resistance available for this minimum noise constant. Fortunately several manufacturers supply information on the mini-

[3] K. Ishii and D. D. Hoffins, "X-Band Operation of S-Band Esaki Diodes," *Proc. IRE,* **50**, July 1962, p. 1698.

[4] A. Yariv and J. S. Cook, "A Noise Investigation of Tunnel Diode Microwave Amplifiers," *Proc. IRE,* **49**, Apr. 1961, p. 739.

[5] L. D. Armstrong, "Tunnel Diodes for Low Noise Microwave Amplifications," *Microwave Journal,* **V**, Aug. 1962, 99-102.

mum noise constant and on the corresponding negative resistance for ease in amplifier design.

High-Frequency Stability

To determine stability consider the circuit of Fig. 2A where $Z_d(s)$ represents the diode and $Z_a(s)$ the associated circuitry, both in Laplace transform notation. This circuit has the characteristic equation $Z_a(s) + Z_d(s) = 0$, which must be examined for root locations. This method has been covered in the literature and applied to several diode configurations.[6, 7]

A simple result is found if the circuit of Fig. 2B is considered. Here actual lumped elements are assumed for Z_a and Z_d. Since the diode is operated at a frequency below its reactive cut off frequency, it can be considered to be capacitive and will require an inductive circuit to resonate it. The associated circuitry shown in Fig. 2B is considered inductive by an amount L_E, with resistance R_E, and driven by a voltage E. The diode equivalent is used with the package capacitance resonated. If the resulting circuit, also shown in Fig. 2B is examined, the following inequality is derived:[8]

$$\frac{L_T}{|R_j|C_j} < R_T < |R_j|.$$

This is the condition for stability.

[6] B. W. Nelson and R. Masens, "Stability Criteria for Tunnel-Diode Circuits," *Electro-Technology*, **73**, June 1964, 52-58; see also Aug. 1964, p. 37, for correction.

[7] L. I. Smilen and D. C. Youla, "Stability Criteria for Tunnel Diodes," *Proc. IRE*, **49**, July 1961, p. 1206.

[8] L. E. Dickens, "The Tunnel Diode, A New Microwave Device," *The Johns Hopkins University Radiation Laboratory*, Technical Report 92, May 1962.

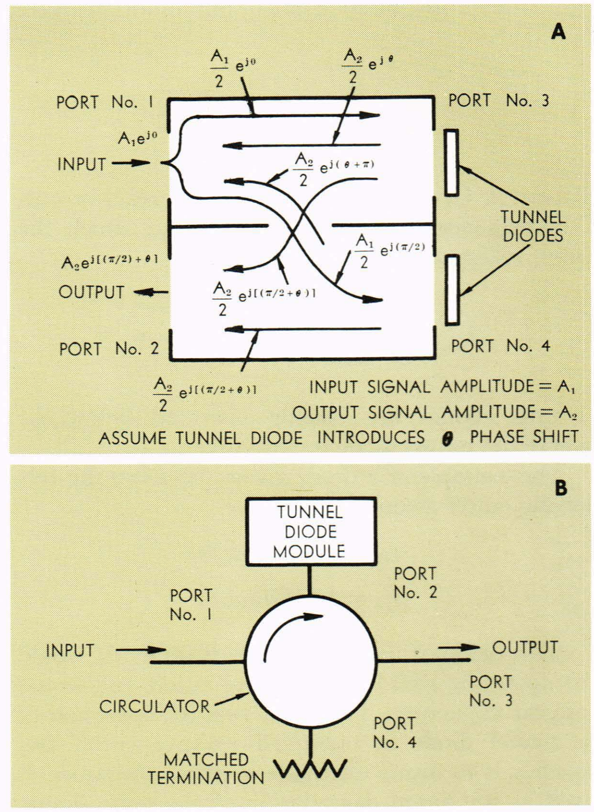


Fig. 3—Two configurations of tunnel-diode amplifiers; (A) is the hybrid coupled amplifier, and (B) is the circulator type.

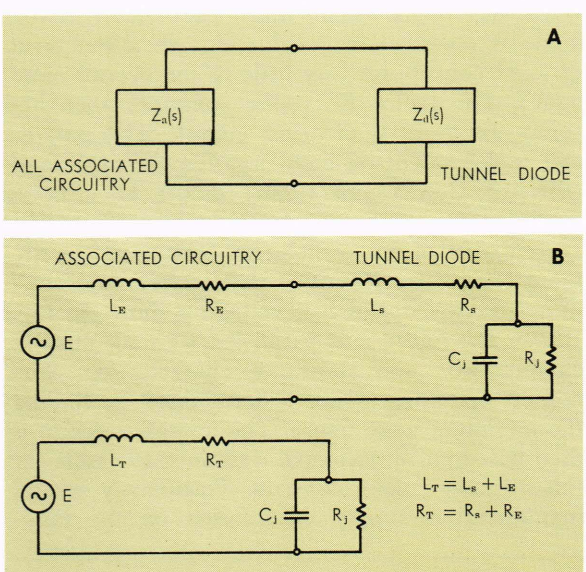


Fig. 2—Tunnel diode with (A) associated circuitry, and (B) associated circuit lumped elements.

Amplifier Configurations

While many configurations exist for stable, low-noise, tunnel-diode amplifiers, most are of only theoretical interest at the present time. A basic requirement of an amplifier configuration is a means of isolating the input and output, which is necessary because the tunnel diode is a one-port device. The majority of current research and development in tunnel-diode amplifiers has been restricted to two configurations, the difference between them being the method of separating the incident (input) and reflected (output) signals.

The first of these is the hybrid coupled type, with signal flow as shown in Fig. 3A. As indicated, signals introduced at port 1 are transmitted to ports 3 and 4, with a 90° phase difference introduced into the output of port 4. Tunnel diodes are located at ports 3 and 4. It is assumed that ports 1 and 2 are properly matched so that there are no reflections from them. Figure 3A shows that if both tunnel diodes introduce the same phase shift and gain, then the reflected signals at port 1 are 180° out of phase while the signals at port 2 are in phase.

The analysis of a signal incident at port 2 would result in a similar solution, that is, the hybrid coupled amplifier has bilateral gain. This property is seldom used since it requires additional devices at ports 1 and 2 to separate the two signals. Some means of isolation is usually used at port 2 to insure unilateral operation. The main advantages of the hybrid coupled tunnel diode amplifier are the wide bandwidths achievable and increased stability.[9] Octave bandwidths to 1000 mc/s are possible, and 20% bandwidths at microwave frequencies are available. These wide bandwidths are possible because of the broadband match that is available with hybrids. The disadvantage of the hybrid coupled tunnel-diode amplifier is that the tunnel diodes must be closely matched so that the phase shift and gain introduced by each will be equal.

The second type of amplifier is the circulator type shown in Fig. 3B. The means of separating the input and output is a ferrite circulator. Signals incident at port 1 are directed to port 2; any reflections from port 2 are directed to port 3, etc. Four- or five-port circulators are used to gain additional isolation between input and output. Operation of the circulator type of tunnel-diode amplifier is as follows: The input signal at port 1 is directed to the tunnel-diode module at port 2; the amplified signal reflected from the tunnel diode enters port 2 and is directed to port 3, which is the output. If port 3 is not perfectly matched, reflections from the terminal are absorbed in the load on port 4. If a three-port circulator were used, reflections from the output would appear at the input of the amplifier. If these reflections were such that the gain around the circulator (circulator isolations times amplifier gain) were greater than unity, oscillation would result. Work at APL has been limited to the circulator type of amplifier because of its simplicity and because it also provides good gain-bandwidth characteristics.

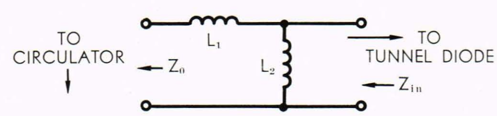
The interface between the tunnel-diode module and the circulator is critical since this match is closely related to the match within the tunnel-diode module that determines gain, bandwidth, and stability. The requirement of stability of mismatch at such an interface is that the power reflected back from the mismatched circulator port to the tunnel diode must be small enough that oscillation cannot result.

Tunnel-Diode Amplifier Design Example

The method we have developed for tunnel-diode amplifier design can be shown with a design for a C-band amplifier. The conditions below must be fulfilled, as previously explained.

1. The diode must be matched to resonate its reactive elements as well as to be shunted across a transmission line having the characteristic impedance for the desired gain.
2. The low-resistance biasing resistor must shunt the diode at DC, be stable with the rest of the circuit at all frequencies, and be isolated from the tunnel diode at frequencies of amplification.
3. The matching circuit and circulator must not give reflections that could cause instability.

The method we are currently using to tune the diode is a combination of series and shunt tuning. The matching circuit shown below is a lumped parameter representation.



The input impedance to this network is:

$$Z_{in} = Z_0 \left[\frac{\omega^2 L_2^2}{Z_0^2 + \omega^2 (L_1 + L_2)^2} \right] + j \left[\frac{\omega L_2 Z_0^2 + \omega^3 L_1 L_2^2 (L_1 + L_2)}{Z_0^2 + \omega^2 (L_1 + L_2)^2} \right].$$

This network presents a reactance to resonate the diode and allows a control on the real part of the impedance for closer matching to achieve the

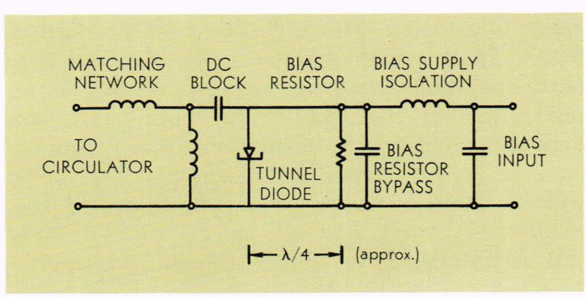


Fig. 4—Equivalent circuit of the APL-designed tunnel-diode amplifier.

[9] J. J. Sie, "Absolutely Stable Hybrid Coupled Tunnel Diode Amplifier," *Proc. IRE,* **48**, July 1960, p. 1321; see also Oct. 1960, p. 1783, for correction.

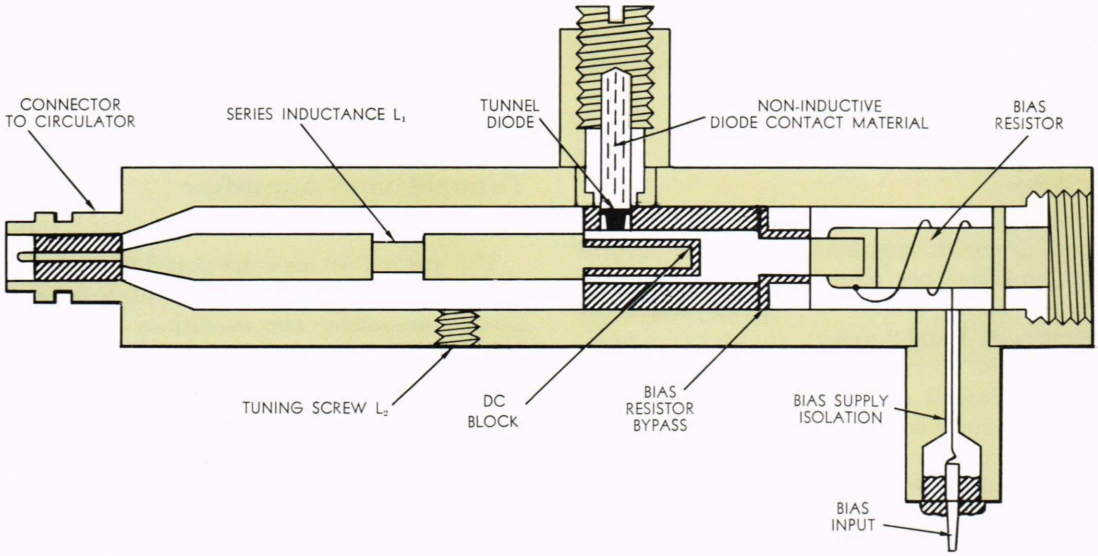


Fig. 5—Cross-section diagram of the C-band tunnel-diode amplifier. All shaded areas are of solid brass, and crosshatched areas are Teflon.

desired gain. For ease of calculation with a Smith Chart, this network is transformed into shunt equivalents.

The lumped equivalent of the APL-designed tunnel-diode amplifier is shown in Fig. 4. A DC block is necessary to isolate the diode bias. The RF isolation network prevents radiation as well as signal coupling into the amplifier. The bias resistor and its bypass are located approximately a quarter wave from the diode. This lumped equivalent can be compared to the cross section of the actual amplifier shown in Fig. 5.

To determine gain and stability it is necessary to relate to a common point all the elements presented by the lumped equivalent. An obvious choice of reference plane would be the location of the tunnel diode.

The Smith Chart can be used with negative impedances. However, this requires transformations since the left half of the complex plane is then represented. We have found it to be far easier to associate with each point on the normal Smith Chart a gain that would be obtained if a particular diode were shunted by the impedance represented by that point. The shunt equivalent derived in the section on the tunnel-diode equivalent circuit will be used as an example. The diode will be shunt-resonated. The admittance of the diode is $Y_d = -1/70.7 + j\,\omega\, 0.95 \times 10^{-12}$. This can be normalized by multiplying by 70.7 to give $Y_d = -1 + j\,2.45$ at 5800 mc/s.

Constant-gain curves are shown on a Smith Chart in Fig. 6. These consist of circles around the point $1 - j\,2.45$. The locus of constant gain can be seen to be a circle. This is expected since gain is an indication of reflected power, as are standard wave ratio (SWR) and the voltage reflection coefficient that are both concentric circles on the Smith Chart.[§] The areas of unstable operation are bounded by the dotted lines. The design example will be based on an 18-db amplifier centered at 5800 mc/s. From Fig. 6 a desired line admittance is chosen for this gain. This was chosen as $Y = 1.3 - j\,1.96$. All admittances must now be related to the diode plane. The remaining admittance necessary to obtain the value determined above is calculated. This value is compared to that admittance that is known to be obtainable with a radial tuning screw, and tuning-screw placement is determined to obtain the desired tuning admittance. The measured impedance of the actual amplifier is also indicated in Fig. 6 and shows very close agreement to the desired value.

The design steps are then as follows:

1. Line admittance chosen for desired gain is $1.3 - j\,1.96$.
2. Admittance of bias resistor and bypass capacitor is $2.94 + j\,3.23$. It should be noted at this point that if this conductance, 2.94,

§ It has been shown[10] that if ρ is the voltage reflection coefficient resulting from the termination of a line by positive resistance R, then $1/\rho$ is the gain if that same line were terminated in $-R$. This is especially important for wide bandwidth designs since bandwidths of positive resistance networks are easily defined.

[10] E. S. Kuh and J. D. Patterson, "Design Theory of Optimum Negative Resistance Amplifiers," *Proc. IRE*, **49**, June 1961, 1043-1050.

were at the diode plane, a negative tuning admittance would be required for the desired gain. As will be shown, this admittance at the diode plane is less than 1.3 and is, therefore, a realistic value.

3. Bias resistor and bypass admittance at the diode is $0.16 - j\,0.1$.
4. The tuning admittance still required is equal to step 1 minus step 3, which is $1.14 - j\,1.86$.
5. The admittance of the circulator and L_1 is now calculated and related to the diode, $0.6 + j\,0.57$.
6. Thus, an admittance of $0.54 - j\,1.29$ is still needed.
7. Based on physical considerations, a point in the line to the circulator is chosen at which this impedance will appear as a value within the range of adjustment of a tuning screw or screws.
8. The chosen point needs an admittance of $0.5 - j\,0.23$, which requires the use of several tuning screws.

This procedure has fulfilled the requirement for center frequency gain. Selection of L_1 is derived from the admittance necessary to make tuning possible with the radial screws; thus several trials of the above procedure are necessary. Three sets of tuning screws are used in the APL-designed tunnel-diode amplifier, and they are necessary in order to achieve correct conductance and susceptance. Exact data relating coaxial probe position and resulting impedance do not exist, with the result that tuning-screw position is not precisely defined.

There are several disadvantages to this inexact engineering. One is that not all line impedances are readily achieved with single screws. Also, tuning is difficult because resulting dimensions between the center conductor and the tuning screw are very small (several thousandths of an inch) and very critical. An investigation of the exact impedance presented by these tuning screws is being undertaken by the author.

The consideration for stability will now be examined. The circles of constant gain shown in Fig. 6 are indicated only in the stable areas. The previously derived condition that $L_T/(|R_j|C_d) < R_T < |R_j|$ is divided into the two areas shown. It can be seen that only a small range of line impedance is usable and that these impedances are surrounded by unstable areas. As the parasitic inductance L_s becomes larger, or as larger inductances are required for tuning, this region of stable gain is reduced.

In order to derive these limits of stability, the equivalent of the tunnel diode is used in its basic form. The required line susceptance is reduced by

[11] T. Moreno, *Microwave Transmission Design Data,* McGraw-Hill Book Co., New York, 1948, p. 102.

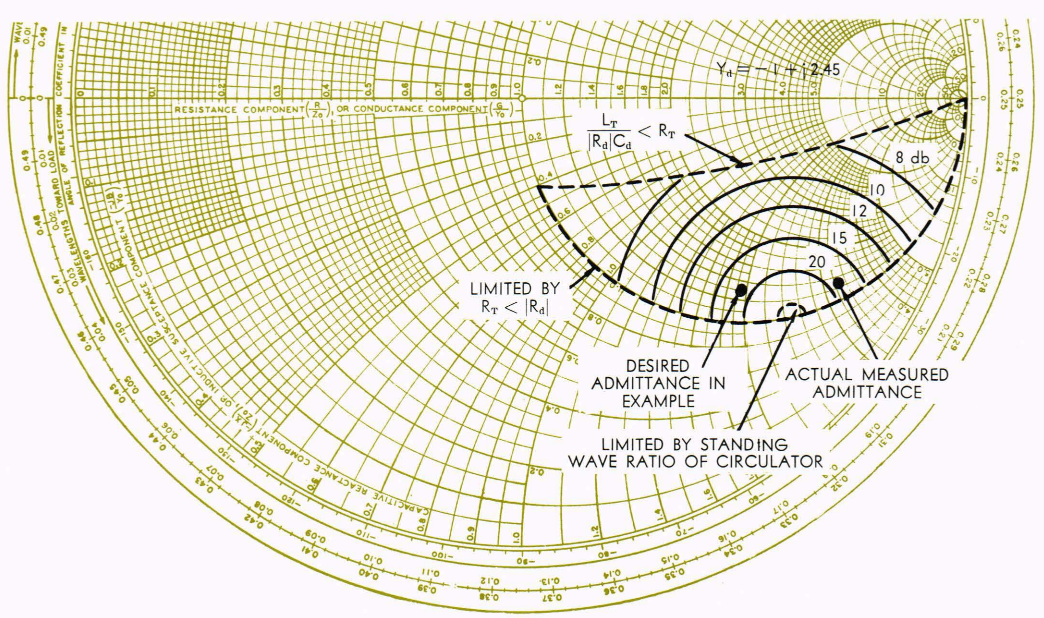


Fig. 6—Constant-gain curves of a tunnel-diode amplifier for all line admittances. Curves are shown in the stable areas only. (For convenience, only the essential portion of the Smith Chart is shown.)

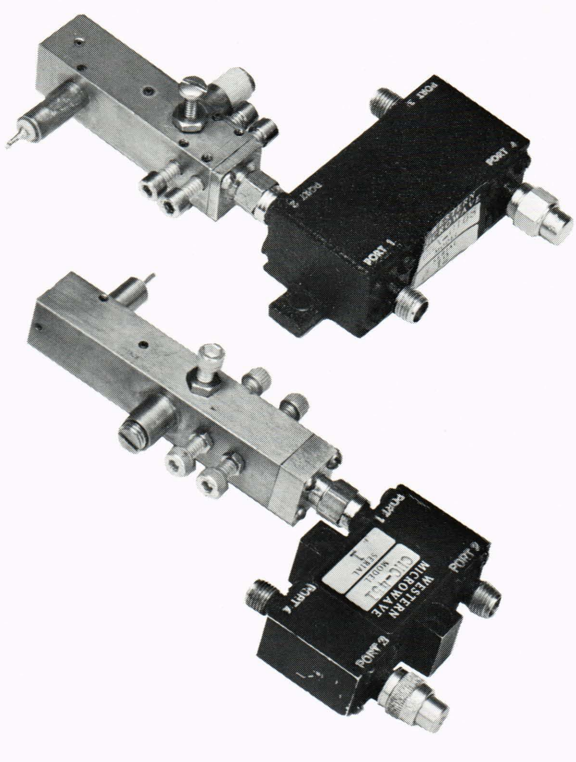


Fig. 7—Photograph showing an APL-designed X-band tunnel-diode amplifier (top) and a C-band amplifier (bottom.) Both units use circulators that are produced by Western Microwave Laboratories.

the amount necessary to resonate the package capacitance. The line admittance can then be converted to a series equivalent so that Fig. 2B now applies to the configuration. The inductances and positive resistances are combined to form R_T and L_T. The values of C_j and R_j are then used to compute $L_T/(|R_j|C_j)$. For this design example, $L_T/(|R_j|C_j) = 23.8$ and $R_T = 29.3$. Therefore, the stability criterion of $L_T/(|R_j|C_j) < R_T$ is met.

The selection of a 24-Ω characteristic impedance at the diode was based on the bias resistor chosen

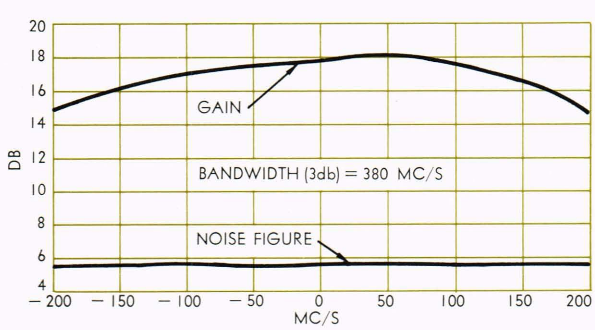


Fig. 8—Gain and noise plotted against frequency for the C-band tunnel-diode amplifier.

for stable biasing and the impedance necessary to keep $R_T > L_T/(|R_j|C_j)$. Thus, a 24-Ω section and termination provides an adequately low impedance for biasing, and it maintains stability.

Our measurements show that R_j is less than −70.7 Ω. This was determined by using several center conductors and comparing gain for equal bandwidth. Each gain was then related to the negative resistance based on the impedance at the diode interface. One of the sources of design error is that the high frequency equivalent is not exactly as given; to correct for this, high-frequency-parameter variations should be included. The methods of diode measurements are being investigated, and initial published results indicate at least a 10% parameter variation.[12]

The shunting of the bias resistor by the bypass capacitance produces a roll-off characteristic of the noise output from the combination. The corner frequency should be below operating frequency. In the design example this is 5000 mc/s, although a lower figure would be desirable and could be achieved by means of a larger capacitor.

The tunnel-diode amplifier described in the previous example is shown in Fig. 7. It is capable of 18-db gain over a 380-mc/s bandwidth at C-band (5000 to 6000 mc/s). Its noise figure is 5.5 db. The gain and noise figures versus frequency are shown in Fig. 8. An X-band amplifier has also been built, capable of 15-db gain over 480 mc/s, and is also shown in Fig. 7. Its noise figure is also 5.5 db. The tighter tolerances required by these high frequencies (X-band) make tuning very difficult. Waveguide amplifier configurations are being investigated to allow higher frequency operation where coaxial construction cannot be used. The amplifiers described above compare favorably with those obtainable commercially. Wider bandwidths are available but have not been stressed in the work at APL because excessive bandwidth increases interference problems.

We have thus designed and built useful tunnel-diode amplifiers with stable gain and phase characteristics. Environmental problems have been minimized by careful mechanical design of the amplifier and by choice of the tunnel-diode material (germanium) that is the least temperature-sensitive. We have also shown that bias variation can be used as a method of temperature compensation to preserve noise figure and bandwidth. Further studies are being conducted to determine exact impedance requirements so that wide latitudes in design frequency can be realized with a minimum of tuning adjustments.

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[12] C. S. Kim and C. W. Lee, "Microwave Measurement of Tunnel Diodes," *Microwaves*, 3, Nov. 1964, 18-21.