although wet oxygen shows large discrepancies at around 400 mc. In fact, wet oxygen appears to go through a minimum in this frequency range, a fact which is not presently understood.

If one considers the surface resistance R_{s1} to be the same as the dynamic low-frequency barrier resistance, a serious objection to the above model is raised when one calculates the values of R_{s1} from the data (see Table II), for it is seen that R_{s1} is about two to three orders of magnitude lower than the low frequency dynamic resistance (approximately 10 megohms at -1 volt bias). Furthermore, the expected inverse proportionality between R_{s1} and the reverse current in different ambients is not apparent (if anything they seem to be directly proportional).

One way out of this difficulty is to consider a more complete equivalent circuit for the surface. Fig. 3 shows a schematic representation of a gold bonded diode, where R_{s1} is the resistance along the surface which shunts the transition capacitance C_t, and C_s is the capacitance associated with the surface space-charge region. In addition, we shall introduce a second surface resistance R_{s2} which expresses the dependence of the reverse current on the surface generation rate, and is analogous to the diffusion resistance at the bulk junction (omitted from the diagram because it is very large compared to the reactance of C_t). It is clear that R_{s2} is in parallel with C_s in order to provide a dc current path. Although R_{s1}, R_{s2}, and C_s are represented as lumped constants, they are actually distributed over the surface and may vary from point to point.

Experimental information indicates that $C_s \gg C_t$, otherwise the measured equivalent capacitance would differ from the transition capacitance C_t. At microwave frequencies, C_s short-circuits R_{s2} so that the equivalent series resistance is still expressed by (2). However, at dc and low frequencies, the main contribution to the dynamic barrier resistance is given by R_{s2} which may be much larger than R_{s1}.

TABLE II
CALCULATED LEAKAGE RESISTANCE IN DIFFERENT AMBIENTS

Ambient	Reverse Bias	R_{s1} (K Ohm)
Ozone	0 Volt	4.7
Ozone	−1	18
Dry O_2	0	6
Dry O_2	−1	47
Wet O_2	0	18
Wet O_2	−1	156

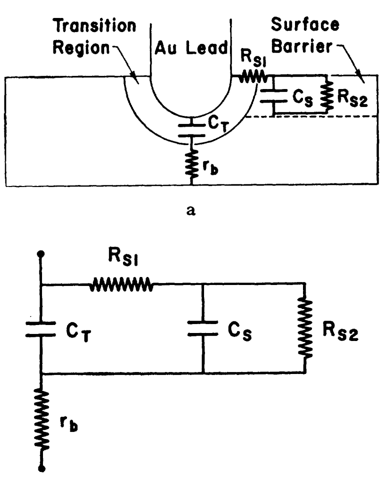

Fig. 3—(a) Schematic representation of gold-bonded diode. (b) Equivalent circuit of gold-bonded diode.

While this work was in progress, D. E. Sawyer of Lincoln Laboratories, Lexington, Mass., reported a $1/f$ frequency dependence (in the range 10 to 250 mc) for the equivalent series resistance of a variable capacity diode.[2] While our experimental conditions differ from his in several important instances (different diode structure, higher frequency range, etc.,) there is nevertheless no adequate explanation for this difference.

S. T. Eng
R. Solomon
Semiconductor Div.
Hughes Products
Newport Beach, Calif.

[2] D. E. Sawyer, Device Research Conf., Ithaca, N. Y., June, 1959.

WWV Standard Frequency Transmissions*

Since October 9, 1957, the National Bureau of Standards radio stations WWV and WWVH have been maintained as constant as possible with respect to atomic frequency standards maintained and operated by the Boulder Laboratories, National Bureau of Standards. On October 9, 1957, the U.S.A. Frequency Standard was 1.4 parts in 10^9 high with respect to the frequency derived from the UT 2 second (provisional value) as determined by the U. S. Naval Observatory. The atomic frequency standards remain constant and are known to be constant to 1 part in 10^9 or better. The broadcast frequency can be further corrected with respect to the U.S.A. Frequency Standard, as indicated in the table; values are given as parts in 10^{10}. This correction is *not* with respect to the current value of frequency based on UT 2. A minus sign indicates that the broadcast frequency was low.

The WWV and WWVH time signals are synchronized; however, they may gradually depart from UT 2 (mean solar time corrected for polar variation and annual fluctuation in the rotation of the earth). Corrections are determined and published by the U. S. Naval Observatory.

WWV and WWVH time signals are maintained in close agreement with UT 2 by making step adjustments in time of precisely plus or minus twenty milliseconds on Wednesdays at 1900 UT when necessary; a retarding time adjustment was made at WWV and WWVH on December 16, 1959.

* Received by the IRE, January 25, 1960.

WWV FREQUENCY
WITH RESPECT TO U. S. FREQUENCY STANDARD

1959 1600 UT	Parts in 10^{10}†
December 1	−30
2	−29
3	−29
4	−29
5	−29
6	−28
7	−28
8	−28
9	−28
10	−28
11	−28
12	−28
13	−28
14	−28
15	−28
16	−28
17	−28
18	−28
19	−28
20	−28
21	−28
22	−28
23	−28
24	−28
25	−28
26	−28
27	−28
28	−28
29	−28
30	−28
31	−28

† 30-day moving average seconds pulses at 15 mc. Method of averaging is such that an adjustment of frequency of the control oscillator appears on the day it is made. No change or adjustment in the control oscillator was made during December, 1959.

Note: Beginning January 1, 1960, the value of the USFS has been arbitrarily increased by 74 parts in 10^{10} to bring it into agreement with a cesium resonator frequency of 9192, 631, 770 cps. See "National standards of time and frequency in the United States," National Bureau of Standards, PROC. IRE, vol. 48, pp. 105–106; January, 1960.

National Bureau of Standards
Boulder, Colo.

The Tunnel-Emission Amplifier*

During the recent tumult caused by the "tunnel diode," this author had cause to reflect upon just what significant statements might be made concerning this device. The more important conclusions may be summarized as follows:

1) The device employs a *controlled source of majority carriers.*
2) Its frequency response is essentially limited by the number of majority carriers available.
3) In times past many negative resistance devices have been introduced, but in the course of time have given way to stable three-terminal amplifying devices.

Once interest in a negative resistance is abandoned, it becomes clear that semiconductors are of questionable value, since their carrier densities are inherently quite low. Metals with very large electron densities may be used as a carrier source, and insulators provide the necessary forbidden regions.

* Received by the IRE, December 28, 1959.

Surface Tunneling

Let us consider the behavior of a metal-insulator-metal layer structure with a voltage impressed between the two metal layers. A simplified energy band picture of such a structure is shown in Fig. 1(a). Under the bias condition shown, electrons near the Fermi level in the negative metal may "tunnel" through the forbidden region into the insulator conduction band and thus make their way to the positive metal. In order to find the total tunnel current we must find the number of electrons per second incident on the junction as a function of energy, multiply by the tunneling probability which will also depend upon the energy, and integrate over the available electron energies. An approximate expression may be obtained by assuming a "hard" Fermi sphere in the metal, an insulator whose forbidden region is centered on the metal Fermi level, and neglecting the contribution of electrons appreciably below the Fermi level. The results of such calculations are

$$J \approx J_0 \frac{\pi V_1}{2V_f} \frac{E}{E_0} e^{-E_0/E} \qquad (1)$$

where

$$J_0 = \frac{3qN_0\hbar k_f}{4m},$$

$$E_0 = \frac{qmaV_1^2}{\hbar^2},$$

V_1 is half the insulator forbidden band gap,
V_f is the metal Fermi level,
k_f is the wave number of an electron at the metal Fermi level,
N_0 is the electron density in the metal,
m is the electronic mass, and
q is the electronic charge.

Since the tunneling distance for electrons in the metal is much shorter than for insulator valence electrons, tunneling from the metal would be expected to dominate the process by a considerable margin. The values for aluminum and aluminum oxide are

$$J_0 \approx 10^{12} \text{ amp/cm}^2$$

$$E_0 \approx 1.5 \times 10^8 \text{ V/cm}.$$

The tunnel current as a function of electric field is shown in Fig. 2.

Tunnel Emission

Suppose we make the positive metal layer thin compared with an electronic mean free path. Electrons tunneling from the negative metal "emitter" on the left into the insulator conduction band may continue through the thin metal base. Those which possess sufficient momentum to overcome the metallic work function will be emitted from the surface to the right.

This structure may be useful as a "tunnel cathode" for conventional or microwave tubes, or it may be used as the emitter of a transistor-like three-terminal amplifier.

Let us place to the right of the thin metal "base" region another insulating layer and then a metal "collector." The simplified energy band picture for such a configuration is shown in Fig. 1(b). Electrons tunneling through the thin "base" region now find themselves in the conduction band of the second insulator, and provided the collector is biased positively, will be accelerated to the right and hence collected. Such a device may be characterized in much the same way as a transistor, and the same considerations concerning gains and impedances apply. In the common base connection, power gain is obtained as a result of the impedance transformation between input and output.

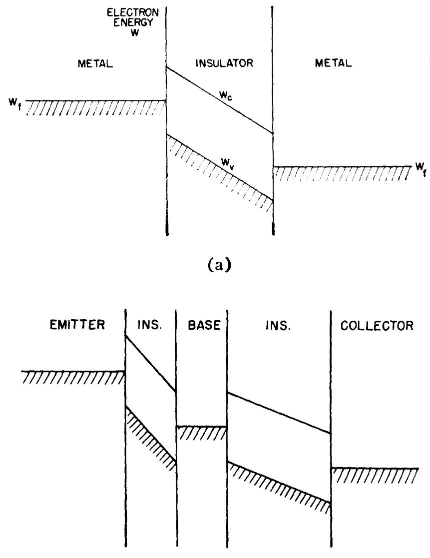

Fig. 1—Simplified energy band models for (a) diode, and (b) triode.

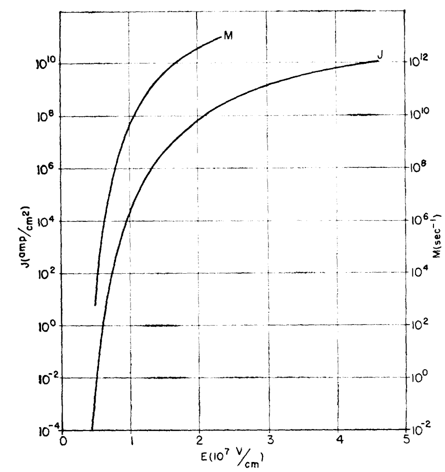

Fig. 2—Tunnel current density and figure of merit as function of electric field.

Frequency Limitations

Since the tunneling itself occurs in an extremely short time, we should expect the operating frequency to be limited by the rather large input capacitance. Using as a figure of merit

$$M = \frac{1}{RC} \qquad (2)$$

where R is the incremental input resistance and C is the input capacitance, we may estimate the frequency limitation from (1);

$$M \approx \frac{J_0 E_0 e^{-E_0/E}}{\epsilon E}.$$

This equation has been evaluated in terms of the electric field and plotted along with the current density in Fig. 2. At a field of 1.5×10^7 v/cm the current density would be of the order of 4×10^6 amp/cm² and M would be nearly 10^{12} cps. The real limiting factor appears to be just how small the devices may be built in order to keep the current density at as high a value as possible. Since there will always be some lateral current in the thin base region, current crowding problems will exist as in the transistor. Also such high currents will cause local heating of the metal and place a limit on the operating densities. The space charge limitation for typical geometries is of the order of 10^9 to 10^{10} amp/cm² and hence should not be a problem. It is clear that the input circuit must work at very low impedance levels, just as the negative resistance "tunnel diode."

Experimental Results

The first observations of surface tunneling were made on diodes constructed as follows: A relatively thick film of aluminum was evaporated on a glass substrate and subsequently anodized in a nonsolvent electrolyte. This method produced Al_2O_3 films whose thickness was calculated from the known oxide growth rate to be between 60 and 80 Angstroms. Many small aluminum squares were then evaporated over the oxide. Tunnel current was observed to begin very abruptly at fields of one to two times 10^7 volts/cm. The results were quite reproducible between units on one anodized film, but varied by as much as a factor of two between different films.

Triodes were made in the same manner, except the second aluminum evaporation was carefully controlled to give a thin base layer (about 100 Å). Silicon monoxide was then evaporated over the entire area to a thickness of several hundred Angstroms. A thick aluminum collector was evaporated over the insulating film. Tunnel current at the collector was observed to begin very abruptly at emitter-base fields of the same magnitude as in the diode. For the units tested, the emitter-collector current transport factor (α) was between 0.1 and 0.3, increasing markedly as the current density increased. This behavior is to be expected from such a device where surfaces may contribute a large number of trapping states.

All units were destroyed at rather small currents (less than 1 ma) indicating that the tunneling had occurred over one very small area. Again this behavior is to be expected from the rather crude techniques employed.

Conclusion

The devices described here are inherently capable of extremely high frequency performance. To realize their theoretical capabilities, however, a large amount of development effort will undoubtedly be necessary. The purity of materials and refinement of technique presently applied to transistors will be very valuable. With the advances in understanding and technique that accompany any new device, the tunnel-emission amplifier may be expected to extend the useful frequency range of solid state devices beyond that now attainable by a considerable margin.

Acknowledgment

The author wishes to extend special thanks to D. G. Dow for many helpful discussions and suggestions, and R. A. Baugh for his assistance in the experimental work.

C. A. Mead
California Inst. Tech.
Pasadena, Calif.

A Low-Level Pulse-Height Standard*

There has been an absence of circuits which can be used to produce calibrated, short, low-level pulses. The circuit in Fig. 1 shows a method of producing calibrated low-level pulses in the range zero to two volts to accuracy of better than one-half per cent of full range. The accuracy is limited on short pulses by the overshoot of the transistor; however, new high frequency transistors can be made to operate very well in this respect. A 2N502 transistor can be used for pulses down to 0.2 μsec with rise and fall times of 0.08 μsec.

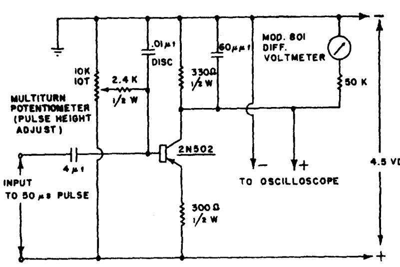

Fig. 1—Standard pulse generator.

The circuit operates by turning on the transistor to a level of conduction determined by the bias setting, and then turning off the transistor during the pulse. The dc collector voltage is measured by a high impedance differential voltmeter. The circuit drive is adjusted for clean "saturation" with no overshoot, so that the dc collector voltage is equal to the height of the output pulse. This may be checked by applying a variable pulse width to the input and observing overshoot while adjusting drive. The multiturn potentiometer used in the base circuit adjusts the amplitude of the output pulse. The 60 $\mu\mu$f capacitor is a compensating capacitor to improve the wave shape. The 50K resistor in series with the voltmeter is used to isolate the circuit from the voltmeter input capacitance.

The circuit is designed to operate with low-pulse duty cycles which contribute very little error to pulse height measurement. The 2N502 should be selected for low leakage, dependent on the pulse height accuracy required in the circuit.

T. E. Lommasson
W. W. Grannemann
Engrg. Experiment Station
University of New Mexico
Albuquerque, N. M.

* Received by the IRE, September 14, 1959.

On the Generality of "Near-Zone Power Transmission Formulas"*

In his correspondence,[1] J. Robieux claimed that my paper[2] gives some results of work on energy transmission from a transmitter to a receiver, *previously* published by him in a *more general* statement. This is misleading. For the purpose of clarification, I would like to present a few facts which show that his result was not obtained or published previous to mine, and also that my result, in my opinion, is really *more* general.

First, I would like to make it clear that I obtained my result without any knowledge of Mr. Robieux's work on a similar formula. This is supported by the following dated written and published records. My work on near-zone transmission relations was a part of a research project on "Study of Near-Zone Fields," sponsored by Rome Air Development Center, Rome, N. Y., under Contract No. AF30(602)-928. This particular result was obtained in the later part of 1955. Statements concerning this result can be found in the letter progress reports, each in a number of ditto copies, to the sponsor in March, 1956, and June, 1956. The final report of this project, including two parts, was written in April, 1957, and these transmission relations were included as one chapter in part two.[3] In October, 1957, an abstract and a summary, based upon this chapter, were sent to the IRE, and were presented at the 1958 IRE National Convention in March and published in the 1958 IRE National Convention Record. By comparing the dates given here and those given by him, the word *independently* may be more appropriate than the word *previously* for the description of the situation.

Secondly, he proved the formula under a number of unnecessary assumptions,[4,5] such as the following.

1) Each antenna has a directivity such that it radiates into, or receives from, only one hemisphere.

2) The two antennas are far apart so that the radiation of one is unaffected by the other.

3) The two antennas are sufficiently far apart that the $\overline{E}$ and $\overline{H}$ radiated by the same antenna are in phase.

The use of these assumptions in his proof restricts the possible applications of his result and therefore makes his result less general. I did not use any one of these assumptions except for special approximate forms. On the other hand, I used the condition that the two antennas are simultaneously matched.[2] In my opinion, this condition is essential in the proof of the form given, but it was not stated explicitly in Robieux' two papers.[4,5] However, a matched condition of special form which is applicable under his restrictive assumptions did appear in his correspondence.

Finally, I do agree with him completely that these formulas can be considered as very general ones in physics. They can be generalized to include anisotropic medium and can also be extended to fields other than electromagnetic theory.

Ming-Kuei Hu
Dept. of Electrical Engrg.
Syracuse University
Syracuse, N. Y.

* Received by the IRE, July 10, 1959.

[1] J. Robieux, "Near-zone power transmission formulas," Proc. IRE, vol. 47, pp. 1161; June, 1959.

[2] M. K. Hu, "Near-zone power transmission formulas," 1958 IRE National Convention Record, pt. 8, pp. 128–138.

[3] M. K. Hu, "Study of Near-Zone Fields of Large Aperture Antennas," Syracuse University Res. Inst., Syracuse, N. Y., Final Rept., pt. 2, pp. 72–84; April, 1957.

[4] J. Robieux, "Interaction entre onde rayonnée et onde de surface et son application à la propagation lointaine," *L'onde Electrique*, vol. 37, pp. 1089–1097; December, 1957.

[5] J. Robieux, "Interaction entre deux aériens," *Compt. rend. Acad. Sci.*, vol. 245, pp. 793–796; August, 1957.

A Microwave Adler Tube*

Adler *et al.*[1] have recently described an electron beam parametric amplifier with a noise figure of 1.4 db (for double channel operation) working in frequency bands between 400 and 800 mc. This amplifier makes use of "fast" transverse cyclotron waves excited on the electron beam in the presence of a longitudinal magnetic field.

We have now built an amplifier, identical in principle but operating in the microwave region at 4140 mc. First experiments have yielded a noise figure of 2.5 db for double channel operation (noise temperature of 225°K). Gains of up to 24 db with a bandwidth of 67 mc to 3 db points have been achieved.

Fig. 1 shows the arrangement of the important parts of the tube, which was operated in a continuously pumped demountable system. (Also see Table I.) Instead of the lumped circuits of Adler's tube, resonant cavities are used. The input and output cavities are identical and are arranged to produce an RF electric field transverse to the beam.[2] They are resonant at the signal frequency (4140 mc) which is approximately equal to the cyclotron frequency of the longi-

* Received by the IRE, October 30, 1959; revised November 12, 1959.

[1] R. Adler, G. Hrbek, and G. Wade, "The quadrupole amplifier, a low noise parametric device," Proc. IRE, vol. 47, pp. 1713–1723; October, 1959.

[2] C. L. Cuccia, "The electron coupler," *RCA Rev.*, vol. 10, pp. 270–303; June, 1949.