**OBJECTIVES**

*To disseminate to RCA engineers technical information of professional value.*

●

*To publish in an appropriate manner important technical developments at RCA, and the role of the engineer.*

●

*To serve as a medium of interchange of technical information between various engineering groups at RCA.*

●

*To create a community of engineering interest within the company by stressing the interrelated nature of all technical contributions.*

●

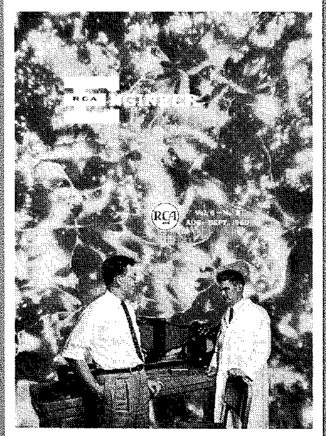
*To help publicize engineering achievements in a manner that will promote the interests and reputation of RCA in the engineering field.*

●

*To provide a convenient means by which the RCA engineer may review his professional work before associates and engineering management.*

●

*To announce outstanding and unusual achievements of RCA engineers in a manner most likely to enhance their prestige and professional status.*



**OUR COVER**

The pattern displayed is a highly magnified structure of germanium crystal material detected during investigations of materials for tunnel diodes. Norman Ditrick (left) and Al Wheeler of the Semiconductor and Materials Division, Somerville, N. J., both authors in this issue, are shown with photomicrographic equipment used to observe and record minute variations in the structure and composition of materials.

# THE ELECTRONICS REVOLUTION

The Industrial Revolution, an important milestone in the development of civilization, marked the extension of man's muscle power by the harnessing of mechanical energy. Today, we in RCA are participating in an even more profound revolution, the *Electronics Revolution,* which marks the extension of man's brain power by the harnessing of the swift, obedient electron.

The first phase of the Electronics Revolution has been the extension of the information-gathering senses — sight and hearing — of the brain. This phase started about fifty years ago and is already well developed. Radio signals span the globe and have extended man's senses across the enormous distances of outer space.

The second phase of the revolution, the extension of the brain's reasoning powers, is just starting. This phase is the most exciting of all and will undoubtedly have the most profound effect upon civilization. Even the most vivid imagination is incapable of foreseeing the extent of the impact to be made by future electronic computers.

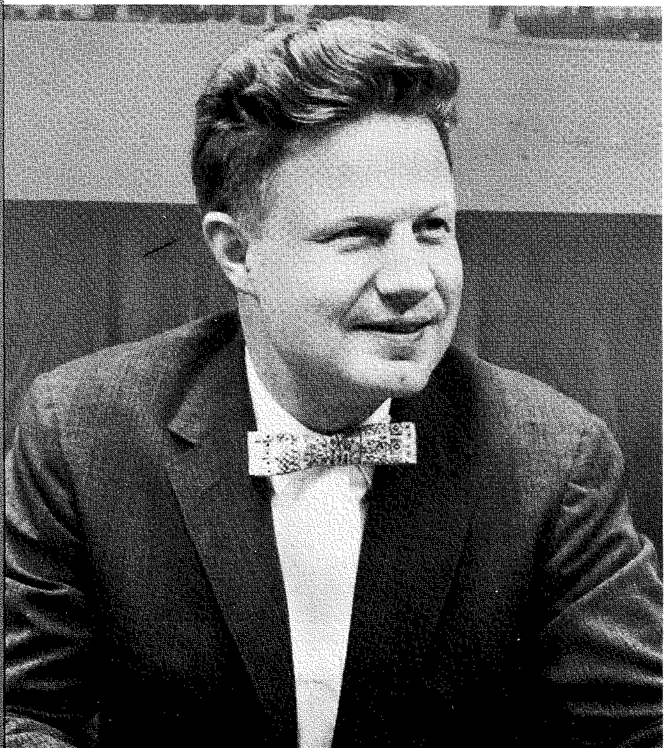
The first electronic computer was made shortly after World War II, and RCA engineers were among the first to develop the basic electronic circuits. More recently, RCA has pioneered in developing and marketing a machine which uses the highly reliable and efficient transistor.

This marvelous little semiconductor device is revolutionary in itself. It and the rapidly growing family of semiconductor devices are, in fact, the adrenalin that makes the pulse of electronics beat so rapidly today.

Indeed, we are growing to realize that semiconductor devices are even more potent than we first imagined—they now promise to perform almost all of the circuit and device functions in a computer. Furthermore, they ultimately will be able to perform all these functions in an unbelievably small space, at high speed, with high reliability, and at very low cost. The amazing compactness and capability of the human brain itself is the measure of what is yet to be achieved through electronics.

The years ahead will see a merging of biology, physics, and electronic information-handling techniques. Devices, circuits, and systems will be integrated to an extent far beyond anything we know today. All of us will be hard pressed to learn new disciplines and to keep pace with new inventions and developments.

The Electronics Revolution will provide an increasingly interesting and challenging future for us in RCA. We are indeed in the mainstream of human progress. To measure up to the challenge will be very difficult, but the rewards for a good job will be great.



Edward O. Johnson

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*Semiconductor and Materials Division*
*Radio Corporation of America*

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RCA ENGINEER

**VOL. 6, NO. 2 • AUGUST-SEPTEMBER, 1960**

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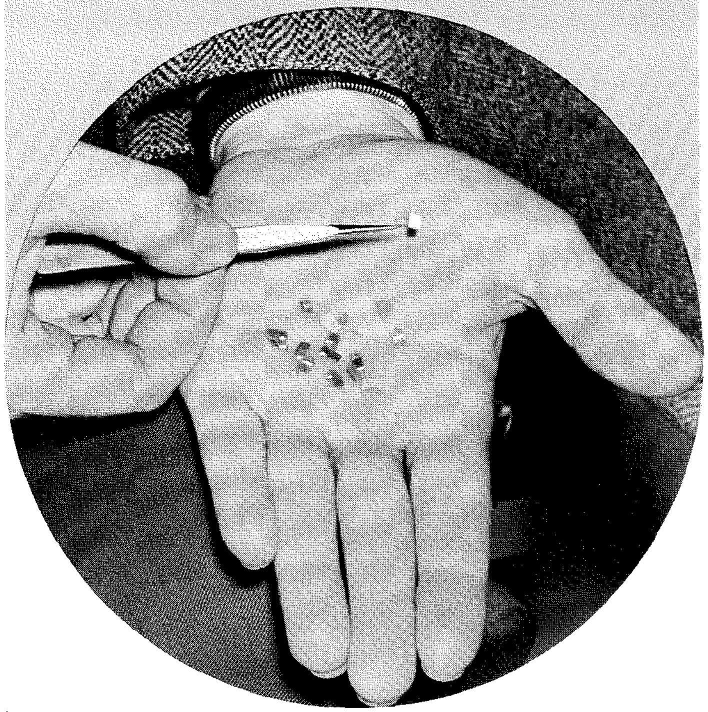
# Introducing . . .
# THE TUNNEL DIODE



The remarkable characteristics of **tunnel-diode** devices and circuits indicate that they will assume an important role among electronically active devices. The articles herein describe some of the many contributions of RCA engineers and scientists to tunnel-diode development and indicate the future course of this work.

Following this introductory description by **Dr. Donahue,** the basic principles of the tunnel diode are described by **Dr. H. S. Sommers** of the RCA Laboratories, who pioneered tunnel-diode work at RCA and has played a large part in the success of the program. Following this, **J. B. Schultz** and **H. B. Yin** of Home Instruments, Cherry Hill, discuss tunnel diodes in linear circuits, reflecting their considerable effort towards potential applications in radio and television. Then, **R. H. Bergman** and **M. M. Kaufman** of the Electronic Data Processing Division, IEP, describe tunnel-diode applications for computer logic and memories. They have done extensive work on the promising applications to very-high-speed computers. **A. J. Wheeler,** who was instrumental in fabrication of the first good gallium arsenide tunnel diodes at Somerville, describes that work in the next article. Finally, **H. Nelson** of the RCA Laboratories and **N. Ditrick** of the Semiconductor and Materials Division discuss the design and fabrication of germanium tunnel diodes. Nelson developed a novel junction fabrication technique and was responsible for the rapid progress made at the Laboratories on the device fabrication. Ditrick played a key role in development of germanium tunnel diodes at Somerville.

As these articles show, the tunnel diode—first brought to the attention by Japan's Esaki in 1958—is a newcomer that although in an early stage of development, holds great promise as ". . . the latest midget prodigy."

*by* **Dr. D. J. DONAHUE, Mgr.**
*Advanced Development*
*Semiconductor and Materials Division*
*Somerville, N. J.*

THE UNPARALLELED progress of the electronics field has been characterized by the discovery and development of the electronically active devices used for amplification, oscillation, and switching. The first such discovery of lasting importance was that of the vacuum electron tube in 1907. For many years, vacuum tubes completely dominated the active-device field and provided the basis for the tremendous electronic industry which exists today.

Another event of comparable significance was the invention of the transistor in 1948. The past decade has witnessed the development of a great number of transistor types and related semiconductor devices employing the p-n junction. In this brief 10-year period, transistors have found use in almost every conceivable type of electronic equipment and have taken their place alongside the electron tube as a permanent member of the active-device family.

Now, before the transistor industry has even begun to reach full maturity, another new electronically active device showing outstanding potential has been discovered. This new device, the tunnel diode, was first reported by L. Esaki of Japan in 1958. Although still in an early state of development, this newcomer promises to assume an important place in the electronics field.

### KEY FEATURES

The most outstanding of the tunnel diode's many features is speed of response. After only one year of development, tunnel diodes can "outrun" the fastest transistor, and in a few years it is expected that they will challenge the "fastest" electron tubes.

Tunnel diodes require much less energy than transistors. Semiconductor surface variability, a knotty problem in obtaining good transistor production yields and stability on life, is almost totally without effect in tunnel diodes. This fact, together with the small size and simple structure of the tunnel diode, promises a low-cost device having excellent reliability. In addition to these advantages, tunnel diodes produce less noise than most other amplifying devices, can operate over greater extremes of temperature, and are orders-of-magnitude more resistant to nuclear radiation than are transistors.

### OPERATING PRINCIPLE

Although the tunnel diode uses the p-n junction, as do most other semiconductor devices, its operating principle is radically different. Electrons cross the junction by "tunneling" through, rather than "climbing over," as in a transistor or ordinary semiconductor diode. The tunneling is made possible by a p-n junction having an extremely thin depletion region 50 to 100 angstroms in width.

*Quantum mechanical tunneling,* from which the new device received its name, is not new, even to semiconductor p-n junctions, since Zener breakdown of junctions is a tunneling process. However, Esaki was the first to report tunneling that increased and then decreased with increasing forward bias of a p-n junction. It is this behavior which produces the very interesting and useful negative-resistance characteristic of tunnel diodes (Fig. 1). This negative-resist-

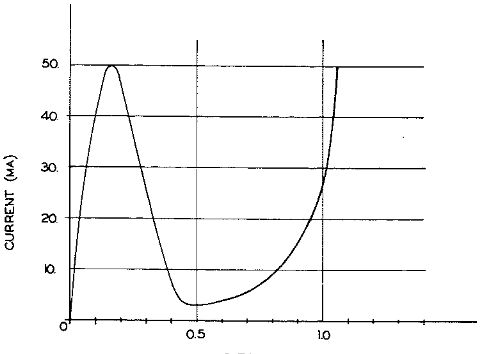


Fig. 1—Characteristic of gallium arsenide tunnel diode.

ance characteristic provides the tunnel diode with its remarkable ability to amplify, oscillate, and switch.

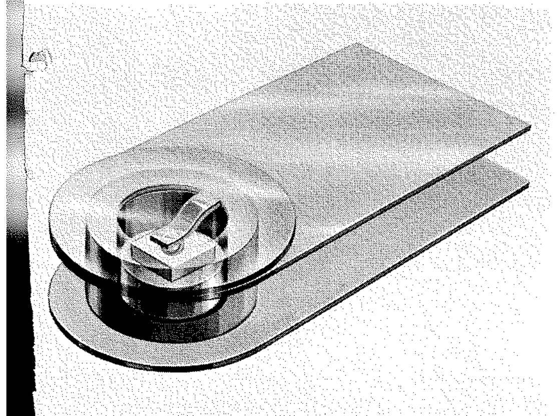
### STRUCTURE

In outward appearance and basic construction, the tunnel diode is virtually the same as any other simple semiconductor junction diode. The junction, for example, can be formed by standard alloying processes. The important internal difference—the one that produces the thin depletion region and, consequently, the tunneling—is the much higher impurity density in the semiconductor material on each side of the p-n junction.

The tunneling current and maximum speed of the device are exponential functions of the impurity density. In addition, it is generally desirable to keep the junction area as small as possible to provide a low junction transition capacitance. Another important point of difference is the nature of the case used to enclose high-frequency tunnel diodes. Because of the low-impedance nature of tunnel diodes, the lead inductance becomes very troublesome at high frequencies. RCA's low-inductance case for high-frequency tunnel diodes is shown in Fig. 2.

The semiconductor from which the diodes are constructed also has a pronounced effect on their characteristics. The techniques of junction formation and device fabrication are most advanced with germanium, and the characteristics of germanium tunnel diodes are generally good. Silicon is more difficult to use and produces diodes having characteristics which are marginal in a number of respects. Gallium arsenide, on the other hand, is not too difficult to use and produces diodes superior in many respects to both germanium and silicon units. The outstanding characteristics of gallium arsenide devices are their greater voltage range, their reduced temperature dependence, and their improved current ratio.

Fig. 2—RCA low-inductance case for high-frequency tunnel diodes.



### DEVELOPMENT

The inherent simplicity of construction of tunnel diodes, together with their freedom from troublesome surface effects, promises much faster development than was possible for the transistor, particularly since the necessary technological and philosophical groundwork has already been so well prepared in the design and development of the transistor. Some idea of the rapid development of the tunnel diode can be gained by a study of its history.

Widespread development started early in 1958, shortly after Esaki's paper was published. By early 1959, a variety of techniques for the fabrication of the devices had been developed to the point where large numbers of diodes were made. By mid-1959, the fabrication and application of tunnel diodes had proceeded to the point where an entire session of the Solid-State Device Research Conference was devoted to tunnel diodes.

Tunnel diodes were placed on the market for sale in the fall of 1959. RCA was one of the first companies to announce large-scale commercial sampling of developmental tunnel diodes. Thus, in a matter of a little over 1½ years, the tunnel diode had moved from early research to pilot production.

Now, in mid-1960, tunnel diodes are being sampled by almost every major semiconductor-device manufacturer in the country, as well as by several companies abroad. Germanium, silicon, and gallium arsenide units are now available in a number of different packages and a variety of electrical characteristics.

### APPLICATIONS

The effort directed toward possible applications of tunnel diodes in various types of circuits has been as intensive as the development of the device itself. It has already been demonstrated that tunnel diodes can perform well as amplifiers, frequency down converters, oscillators, and switches. The outstanding problem associated with the use of tunnel diodes, however, is the two-terminal nature of the device. The fact that the signals must be put in and taken out of the same terminals leads to a bilateral condition which must be overcome by some external means of isolation.

The most promising application of tunnel diodes is as monostable or bistable switching devices. Switching speeds of 0.1 millimicrosecond have already been reached. This high speed, together with very low power requirements and simple circuits, makes tunnel diodes attractive for computer logic circuits. They can also serve as memory elements in very-high-speed computers. The use of tunnel diodes now permits the construction of computers which are 1000 times faster than present commercial models.

As frequency down converters, tunnel diodes not only exhibit very low noise figures, but also provide a power gain instead of the usual loss. Used in the front end of a radio or television receiver, tunnel diodes might provide high sensitivity and low noise without the use of an r-f amplifier.

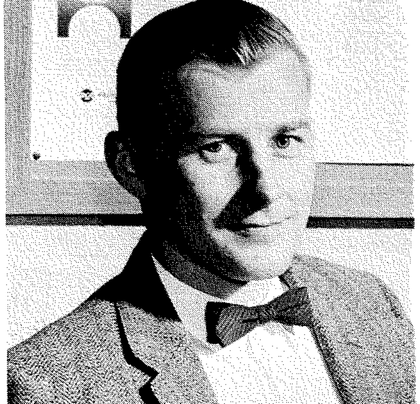
The first use of tunnel diode amplifiers will probably be in the microwave field, where isolation is not much of a problem and where low noise and wide bandwidth are very important. After the remaining circuitry problems are overcome, tunnel diodes will probably find use in all types of amplifiers.

Tunnel diodes function very well as oscillators; however, power output is limited to a milliwatt or so per diode unit. In this application the higher voltage, and consequently the higher power, of gallium arsenide units is very important. Again, tunnel-diode oscillators will first find use in the microwave field at frequencies beyond transistor capabilities. When diode prices become lower, they will also find use in lower-frequency oscillator applications.

### CONCLUSIONS

The discovery and development of tunnel-diode devices and circuits represents an outstanding advance in the electronics field. Because of its remarkable characteristics — very high frequency, low noise, low power, and wide bandwidth — the tunnel diode should find widespread use.

For Dr. Donahue's biography prior to his appointment as Mgr., Advanced Development, see Vol. 5, No. 4, page 44.



# TUNNEL DIODES, THE LATEST MIDGET PRODIGIES

*by* **Dr. H. S. SOMMERS, Jr.**
*RCA Laboratories*
*Princeton, N. J.*

THE TUNNEL DIODE is a semiconductor junction diode requiring only a d-c bias supply to become an active device of extreme versatility. In this sense, it is a contrast to the maser and the variable-capacitance diode, both of which are negative-resistance diodes with a high potential in a restricted range of applications as low-noise amplifiers or high-frequency oscillators.

The tunnel diode should be regarded as *a general-purpose device*, in the same sense as the transistor. It can perform a great range of functions quite well, but in general falls short of the ultimate performance achieved by the more specialized devices.

### WHY ANOTHER GENERAL-PURPOSE DEVICE

The transistor is indeed a very useful and powerful element. But, it has certain limitations to overcome, a fact which accounts for the great current interest in the tunnel diode as another general-purpose device. Among these limitations are moderately high power dissipation per unit, which becomes a serious handicap in equipment designed for space vehicles, battery-operated systems, and complex computers; other drawbacks are rather large temperature sensitivity and limited speed. This speed limitation is the fundamental difficulty with the bipolar transistor and warrants further discussion, in order to better understand the tunnel diode.

### TRANSIT TIME LIMITATION OF BIPOLAR TRANSISTORS

The bipolar transistor is a charge-control device in which the flow of current by one sign of carriers (the majority carriers) is controlled by injecting a few carriers of the opposite sign (minority carriers). The speed of the device is ultimately limited by the rapidity with which these oppositely charged carriers can diffuse or drift through the sea of majority carriers.

Fig. 1 illustrates the transistor speed problem by a crude but useful mechanical analogy. The speed of operation of the transistor, shown schematically as *P-N-P* at the top of the sketch, is limited by the time it takes a signal to travel from the input at the left to the output at the right. In the mechanical analogy, shown in the middle, the input is the hammer which falls onto the *p* region, represented by the racked row of billiard balls. When the hammer hits the first ball, the impulse is transmitted as an elastic wave to the last ball, propelling it into the middle region. Across the middle, the ball rolls with a fixed velocity which corresponds to the drift of the minority carrier through the base region. The analogy emphasizes the contrast between *majority-carrier* signalling by an elastic wave, and *minority-carrier*, which requires the transport of a discrete charge. Finally, the minority carrier reaches the collector, and the impulse is transmitted to the output as an elastic wave —again, majority-carrier transport.

The bottom section in Fig. 1 gives an idea of the relative times involved. Since the signal propagates as an electromagnetic disturbance through the emitter and collector, the overall time is limited by the minority-carrier drift. This drift time is always relatively long, for the drift velocity is relatively slow. The speed of a transistor is thus limited by how short the base region can be made. For a 300-mc transistor, the thickness is already reduced to around 0.1 mil; trying to reduce it much further is a losing business, unless fleas can be trained to do the fabrication. It is hard to believe that a 1000-mc bipolar transistor can ever be sold at anything but premium prices.

This analogy illustrates one other point: while the diffusion region slows the device, it also gives isolation between input and output. This permits triode action for the transistor. The tunnel diode, having no diffusion region (as will be discussed further), sacrifices the circuit isolation.

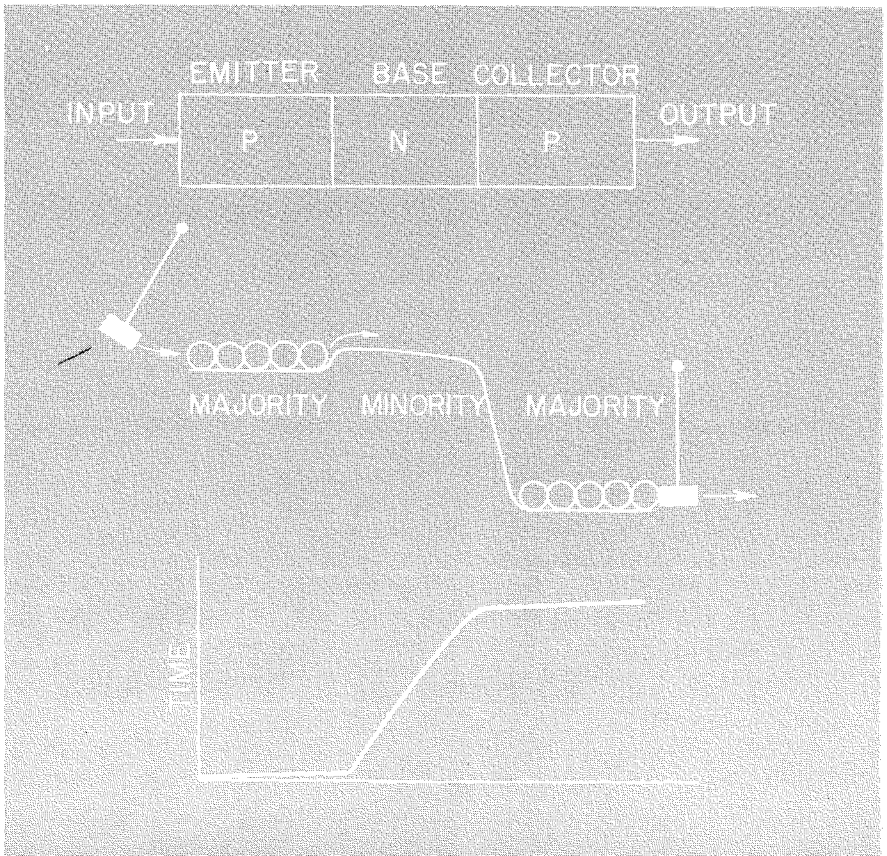
### TUNNEL-DIODE THEORY AND CHARACTERISTICS

Some idea of the nature of the tunnel diode can be gained from the mechanical analogy just discussed by imagining how the barrier to the injection of minority carriers becomes higher and higher until there is no noticeable flow of the balls over the top of the center region. At the same time, the middle region becomes thinner and thinner. When it becomes thin enough, there is a small probability (from quantum-mechanics theory) that the ball can go from the one majority-carrier band to the other without going over the top. This process is given the descriptive name of *tunneling;* it is the basis of charge transport across the junction in a tunnel diode.

#### Basic Concept

Just over two years ago, Esaki published a letter in the Physical Review entitled, "A New Phenomenon in Narrow P/N Junctions," which described a junction diode with an interesting voltage-controlled negative resistance. The construction of the diode has an appealing simplicity, illustrated in Fig. 2. Onto a thin 2-mil layer of highly conducting germanium (0.001 ohm-

Fig. 1—Mechanical analogy showing the transistor signalling-time problem.



cm), a 2-mil dot is placed. The unit is then heated at 400°C for half a minute, forming the junction. It is mounted by soldering the wafer to one lead and the dot to another.

The I-V characteristic of such a diode is shown in Fig. 3, where it is compared with a more normal type of rectifier diode. The dashed line is the rectifier. In the reverse direction, it is blocking until breakdown; while in the forward, it starts to conduct by injection of minority carriers at around 300 mv. The tunnel diode (solid line) is highly conducting for all reverse biases. In the forward direction, the current rises rapidly to a sharp maximum, drops to a deep minimum, and then goes over into the typical minority-carrier injection curve. The drop in current with increasing voltage for modest forward bias is the *Esaki effect.*

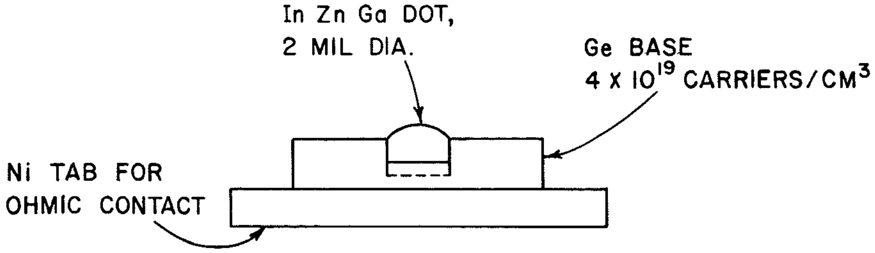


Fig. 2—Sketch showing the constructional simplicity of the tunnel diode.

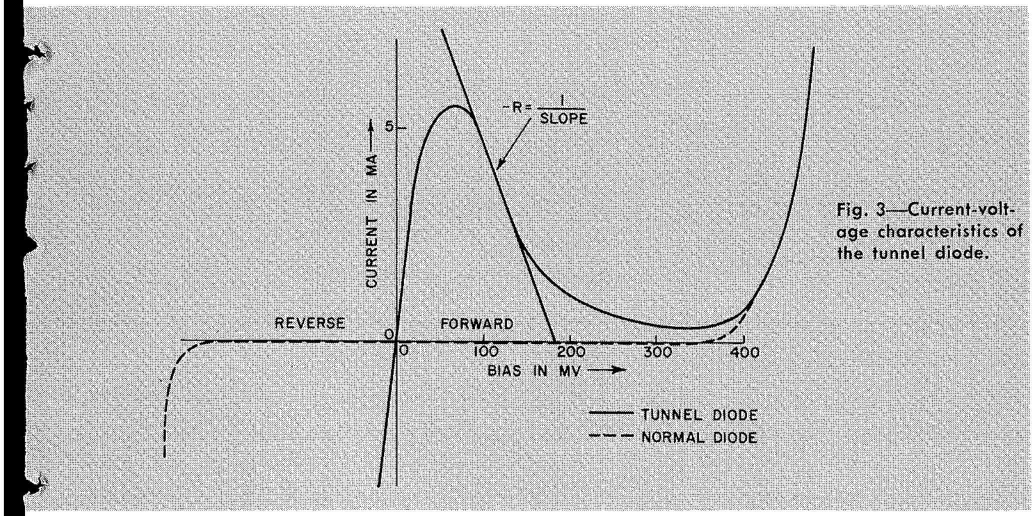


Fig. 3—Current-voltage characteristics of the tunnel diode.

### Esaki Effect

The Esaki effect combines the phenomenon of electrons tunneling through a region where their momentum becomes imaginary with the displacement of the energy bands of the crystal under bias voltage. A fundamental discussion of the effect would require going far more deeply into wave mechanics and the band picture of solids than is warranted here. The following simple analogy treats the Esaki effect in terms of transmission through a waveguide. This is not too far-fetched a comparison, for in wave mechanics the electrons are treated as waves and the bands of the semiconductor are associated with the pass band of a waveguide.

Fig. 4 is a waveguide representation of the Esaki effect. The *n* and *p* sides of the semiconductor diode are waveguide sections *A* and *B*. The electrons in the semiconductor are electromagnetic waves in the waveguide. The transition region of the diode becomes a connecting piece of waveguide whose dimensions are too small for the impinging wave; i.e., section *D* is a waveguide beyond cutoff. As a result, the wave leaving *A* is attenuated exponentially as it traverses section *D*, and only the most minute amount trickles into *B*. This is equivalent to the tunneling of electrons through the transition region.

The Esaki effect is represented as a diaphragm at *C* in the connecting waveguide. This diaphragm has the property that its transmission coefficient changes with the bias voltage across the diode. In particular, for certain regions of forward bias, the transmission drops with increasing bias.

### Transit Time

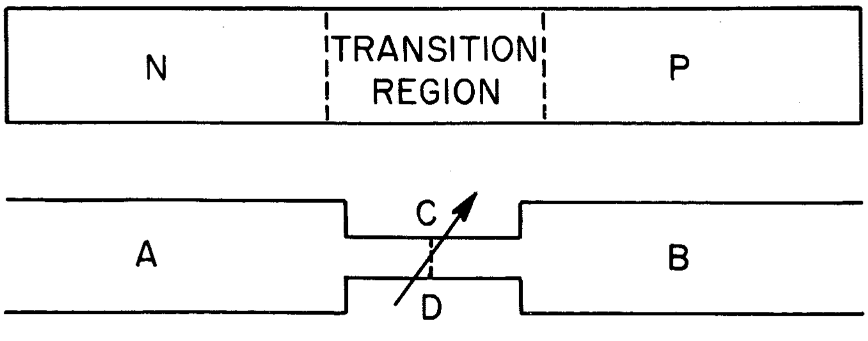
The waveguide analogy is also very useful as an illustration of the transit time. Through *A* and *B*, the signal travels as an electromagnetic wave; i.e., the current is conducted by an elastic wave of majority carriers. Whether or not the only limitation through the transition region *D* is the wave velocity is not certain yet, but certainly the time delay here will be very short because the thickness is so tiny, around $10^{-6}$ cm. Hence, the tunnel diode does indeed give promise of speeds *far beyond* the region where bipolar transistors fail.

### Circuit Interpretation

The active character of the tunnel diode (see Fig. 3) is associated with a drop in current for a rise in voltage, which is the negative conductance, or its reciprocal, the negative resistance. Its circuit interpretation is an element with a negative power absorption, i.e., a power generator. The current scale is arbitrary, since the actual current depends on the junction area as well as the transmission coefficient of the barrier. The voltage, however, is roughly representative of the Esaki effect in any material.

The current maximum occurs at a small fraction of a volt, between 40 and 100 mv. The current minimum also occurs at small voltages, between ¼ and ⅔ of a volt depending on the material and the processing. To a first approximation, the voltages of the maximum and minimum, as well as the

Fig. 4—A simple waveguide analogy of the "Esaki" effect of the tunnel diode.



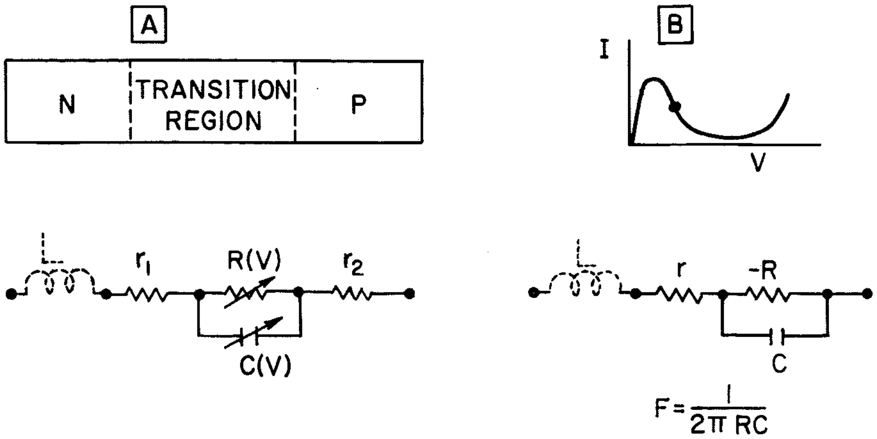


Fig. 5—Tunnel-diode equivalent circuit, including parasitic inductance due to the mounting.

available voltage swing as a device, can be pictured as being constant. This is a great convenience, since the magnitude of the negative resistance can be considered as being determined solely by the peak current.

**Equivalent Circuit**

The equivalent circuit of the tunnel diode is shown in Fig. 5. The *n* and *p* regions act as pure resistances. The transition region is represented as a voltage-sensitive resistance, since tunneling is a function of voltage and junction capacitance. This capacitance is just that of a plane-parallel condenser with plates separated by the transition region. Fig. 5-B is the form of the equivalent circuit when the diode is biased to have an operating point in the negative resistance region. The dynamic resistance is negative, and for small signals both it and the capacitance are constant. In Fig. 5, a parasitic inductance due to the mounting is indicated (dashed portion). For low-frequency diodes this is unimportant, but at higher frequencies (above 100 mc) the inductance looms as an increasingly important parameter, as will be described later.

The figure of merit of a tunnel diode is given by $F = 1/(2\pi RC)$. This product has two very useful interpretations: it is the diode gain-bandwidth product for linear circuits, and its reciprocal is the diode switching time as a logic element.

This figure of merit gives an inkling of why the tunnel diode is such a flexible element. In most devices, the RC product is practically constant for a given type of structure. For instance, with change in area the capacitance increases at essentially the same rate that the resistance drops; so the product becomes independent of area. The same is true when any one of the linear dimensions is scaled.

For the tunnel diode, however, a new result may be expected. True, the product should again be independent of area; nevertheless, since the capacitance is a classical quantity while the negative resistance is due to tunneling—a strictly quantum-mechanical concept—R and C may be expected to vary independently of each other. This proves true; the RC product can be changed indefinitely by varying the doping of the semiconductor. As a result, the figure of merit of one structure can be changed from cycles per second to many kilo-megacycles per second simply by increasing the free-carrier concentration in the bulk material.

## SWITCHING-SPEED APPLICATIONS

As a first example of the applications of tunnel diodes, the switching speed of a diode is graphically analyzed as shown in the equivalent circuit of Fig. 6. This indicates how currents divide inside the diode when a constant-current step is supplied to the diode. It is assumed that the diode, originally in the low-voltage state, is pulsed with a current $I_0$. Of course, the static curve of the tunnel diode, the ordinary characteristic curve, gives the resistive component through the diode at any voltage. This is the tunneling current. The rest of the current is used in charging the junction capacitance, and the larger this component of current, the faster the diode will switch.

Thus, there will be two distinct regions in the switching transient. Near the current peak is the region where the charging current is quite small and depends very critically on the amount that $I_0$ exceeds the peak current of the diode. Beyond the peak is the region where the charging current is high and the diode voltage builds up rapidly. The rate of build-up in the switching region will depend on both the slope of the diode characteristic curve and on the capacitance of the junction; hence, the switching time should be determined by the RC time constant of the diode. Accurate analysis shows that for an appreciable overdrive, the delay in getting over the hump is small and the switching time becomes approximately $2\pi RC$. A rather curious factor in this switching transient is that the capacitive component of the current *increases* with time over the mid-region of the response.

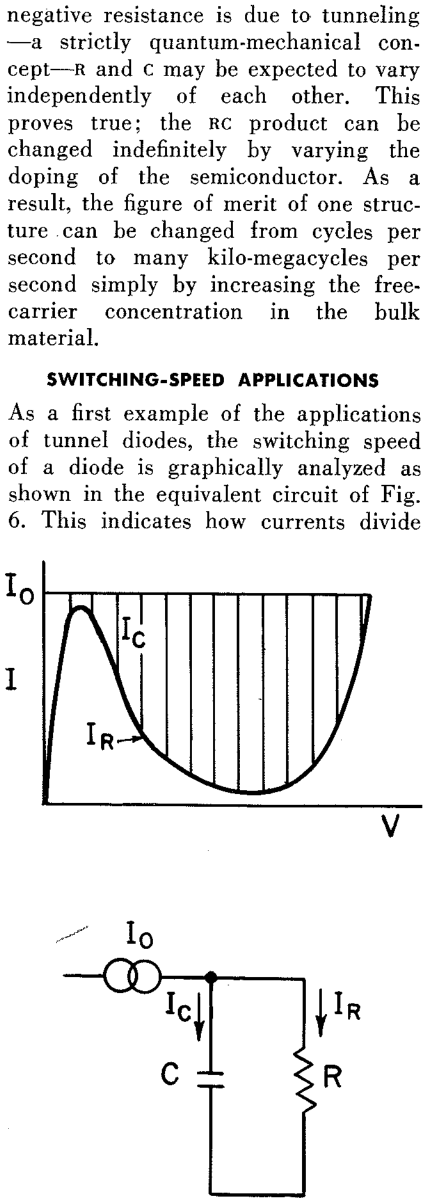


Fig. 6—Equivalent circuit of the tunnel diode switch showing how currents divide inside the device.

Fig. 7 is a switching test of a moderate-speed diode. An unbiased diode was driven with a current pulse, and the voltage across the diode monitored on a sampling scope. In the test example, the overdrive was kept small to exhibit the delay, the current pulse exceeding the peak current by perhaps 10 percent. The delay after the application of the pulse is about 8 nanoseconds and the final switching time after passing the hump about 4, giving a total of 12 nanoseconds. On removal of the pulse, the delay was only 4 nanoseconds plus 4 more to get back to zero. From such tests, for a 10-percent overdrive the total elapsed time is only slightly greater than $2\pi RC$ for the diode.

On application of the pulse, there is a voltage spike caused by the inductance of the diode mount (in this case a transistor stem.) This inductance must be kept *very small* on high-speed diodes to prevent serious transient effects. For microwave applications, special packaging of the diodes will be required.

Fig. 8 gives a sampling-scope picture of a faster diode in a lower-inductance mount. The overdrive is sufficient so the

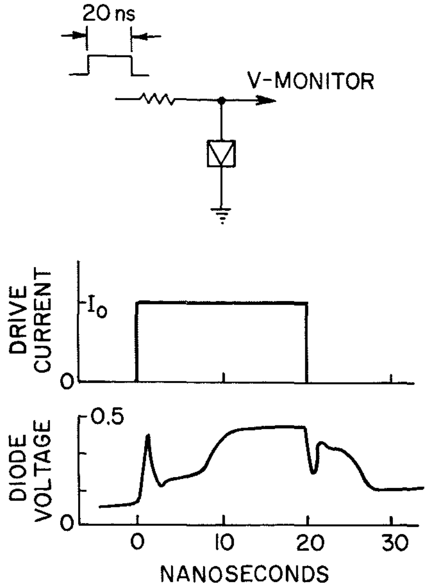


Fig. 7—Switching-test waveforms showing the total elapsed time for an unbiased, moderate-speed diode.

switching delay is negligible. The scale is $10^{-9}$ second/division; notice there is no inductive transient for this diode, and the rise time is that of the sampling scope.

## USE IN LINEAR CIRCUITS

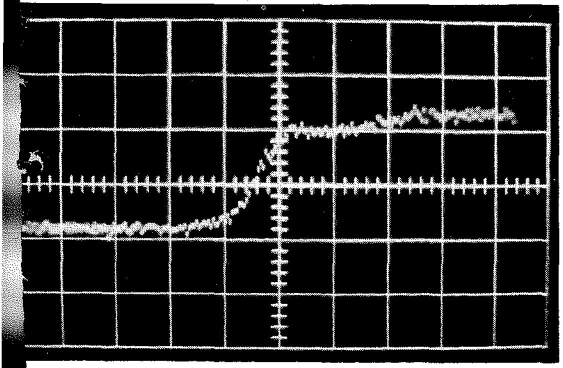
The application of tunnel diodes to linear circuits requires that an operating point be established in the negative-resistance region. The d-c load line in Fig. 9 must be very steep so as to intersect the static characteristic at only one point; thus, the d-c source must have a resistance $r_1 < R$. This still permits two possibilities for the a-c load line: either steep with one intersection as the load line for an amplifier (a highly-damped circuit), or flat with three intersections, giving oscillation.

### Oscillators

Fig. 10 is the circuit schematic of a tunnel-diode oscillator. The left side is the bias supply, the right the tank circuit. The tank is an inductance in parallel with the diode capacitance; the capacitor $C_0$ is a d-c blocking reactor which offers no impedance at the tuned frequency. From *D*, the tunnel diode, there are two circuits—the tuned circuit *B* and the input circuit *A*.

Since the bias supply must have a very low internal resistance $r_1$ to establish the operating point, the diode actually looks into two tank circuits and can choose between them. In general, it chooses the battery circuit. It takes intelligent circuit design to make the diode work into the tank circuit, of which there are a number of examples. They all involve terminating the battery circuit in such a way that from the battery the effective resistance is positive while from the tank circuit it is negative.

Fig. 8—Sampling-scope picture of a faster diode than that used in Fig. 7, and in a lower-inductance mount.



### Amplifiers

The circuit diagram of a lumped-circuit amplifier made by Chang to demonstrate what can be done with a tunnel diode is given in Fig. 11. The battery circuit is connected across the blocking capacitance. Provided $C_0$ is large enough to satisfy the relation $C_0 > L_0/Rr_0$, the battery looks into a positive resistance at all frequencies and so is stable. The autotransformer *T* performs the double purpose of tuning the diode capacitance to give a bandpass at the design frequency and of matching the diode to the input and output. The condition for amplification is that the shunt conductance be higher than $1/R$, the magnitude of the diode conductance. The circuit becomes a *video* rather than an i-f amplifier when the inductance T is omitted.

The gain of such an amplifier, as well as the gain-bandwidth product, is readily calculated. Since the noise of the diode comes from the shot noise of the bias current, the noise figure can also be calculated. Table I shows some early circuit measurements made by Chang. The expression for power gain at the bottom of the table shows that high gain can be achieved by making the parallel conductances cancel each other. Of course, the better the cancellation, the more subject to drift the amplifier gain becomes. Some studies have shown that with passive elements for the input and output, 20 db of gain is a usable level. The next expression is the voltage gain times the bandwidth, which is the usual figure of merit of a circuit. This is determined solely by the constants of the diode for this single-tuned circuit. Thus the diode figure of merit, $F = 1/(2\pi RC)$, can be interpreted as the gain-bandwidth product of the device.

The third and fourth columns of Table I compare the measured values of gain and bandwidth with theory. With proper circuit constants, a gain-bandwidth was obtained equal to the diode figure of merit. The last column shows the comparison of the measured noise figure of the amplifier with the calculated factor, assuming all the noise

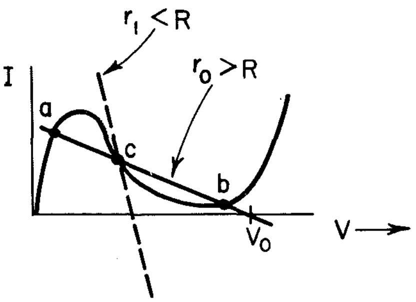


Fig. 9—Tunnel-diode load lines.

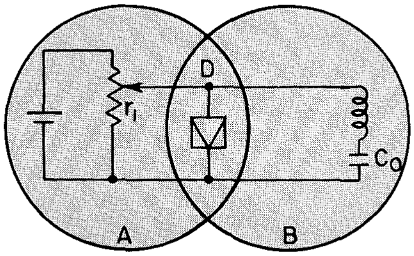


Fig. 10—Tunnel-diode oscillator, including bias.

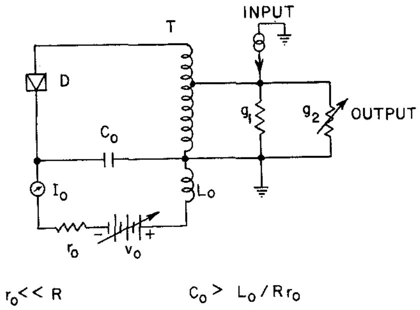


Fig. 11—Lumped-circuit amplifier.

TABLE I—AMPLIFIER RESPONSE AT 30 MC

| Diode Current, μa | Diode Conductance, milliohm | Power Gain, db | | Bandwidth, mc | | Noise Factor, db | |
|---|---|---|---|---|---|---|---|
| | | Meas. | Comp. | Meas. | Comp. | Meas. | Comp. |
| 250 | −2.7 | 20 | 23 | 0.20 | 0.30 | 4.5 | 4.7 |
| 300 | −3.2 | 40 | 36 | 0.19 | 0.16 | 6.3 | 5.5 |
| 350 | −4.8 | 27 | 26 | 0.8 | 1.05 | 8.0 | 6.8 |

Power Gain $\equiv G_p = 4g_1g_2/[g_1 + g_2 - g]^2$

Gain-Bandwidth $\equiv 2G_p^{1/2}\Delta f = \frac{g}{2\pi C}$

Noise Figure $= 1 + \frac{1/g}{2KT/e}$

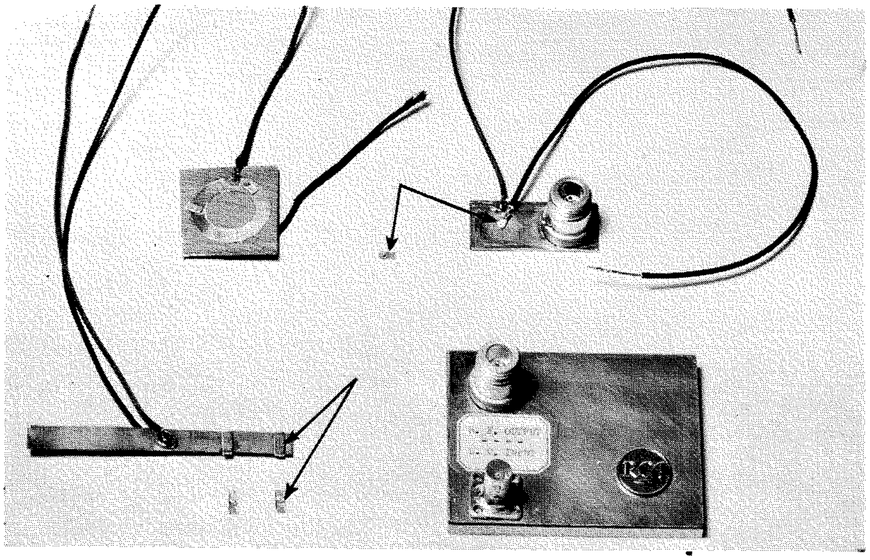


Fig. 12—An assortment of stripline oscillators for low-microwave frequencies. Arrows indicate tunnel diodes for size comparison. Top left, ring oscillator; top right, waveguide; bottom left, quarter-wave oscillator; bottom right, packaged stripline oscillator.

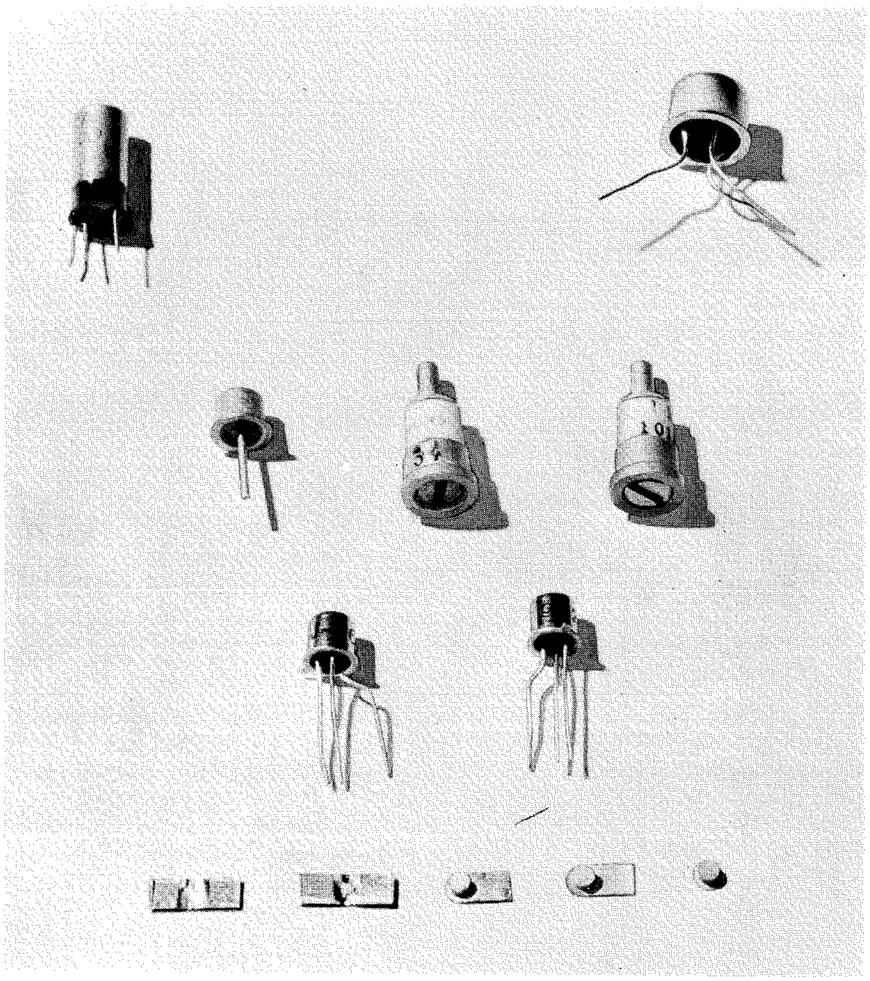


Fig. 13—Assortment of tunnel diodes produced by various electronic equipment manufacturers. An RCA 2N109 transistor is included at top left to provide a size comparison. Across the middle are the units from Sony; the microwave cartridge holds experimental samples and the other diode mount is a preproduction sample having rather high inductance but a junction capacitance of 5 $\mu\mu f$. The GE units below these are commercial, low-capacitance diodes similar in character to the Sony units. RCA tunnel diodes are at the right bottom, with stripline units shown at left.

of the diode is due to the white noise of the bias current.

The agreement is satisfactory, meaning both gain-bandwidth and noise figure can be predicted from the static characteristic of the diode. For germanium diodes at room temperature, a noise figure of about 3 db exists; stated another way, the germanium diode has a noise temperature of about 300°K, independent of frequency up to the diode limit.

A variety of amplifiers have been made that give performance in agreement with predictions. Among these are i-f amplifiers with 3-db noise figure and 10-percent bandwidth at 20-db gain for frequencies up to nearly 1 kmc, and video amplifiers with 20-db gain and 70-mc bandwidth. The utility of such amplifiers has not been established as yet, however, because of their sensitivity to the input and output circuits—a sensitivity enhanced by the fact that they are bilateral rather than unidirectional. At present they can only be used successfully when they are inserted between isolating stages rather than with other tunnel-diode amplifiers in cascade.

### Circuit-Design Mechanics

Again, consider the tunnel-diode oscillator, which is a simpler use because there is no question of cascading stages. Fig. 12 shows an assortment of stripline oscillator circuits for low-microwave frequencies—a very tempting and easy way to make breadboard circuits. The starting point is a printed-circuit board, with copper sheathing on both sides of a 10- to 20-mil sheet of insulation. The circuit is cut out with a pair of scissors and a razor blade. The only additional elements in each circuit are a tunnel diode and a resistor; these are inserted by cutting away the insulation and sandwiching the element between the two copper foils.

The simplest is the quarter-wave stripline oscillator shown in Fig. 12. This has the geometry of a plane-parallel transmission line a quarter-wave long. Looking down on the upper conductor, the left end is the open end of the line; thus it is the voltage-maximum in the standing-wave pattern. One-quarter wavelength away, a noninductive resistor with small resistance is inserted, establishing a voltage null at this point and fixing the mode of oscillation. Just beyond this point, the diode is inserted at the extreme right end of the line, so that the diode capacitance is tuned by the wave reflected from the open end of the line. The parallel combination of diode and resistor presents a positive resistance to the input leads, suppressing oscillation in the biasing circuit. When 100 mv, d-c, is supplied to the leads shown, the circuit oscillates at 500 mc. It can be tuned over a considerable band by snipping off segments from the open end with a pair of scissors, or splicing on sections with Scotch tape.

To the right in Fig. 12 is a packaged oscillator of this same type. The d-c leads are brought in through a BNC connector and the output is transformed from the 10-ohm level of the strip to a 50-ohm output at the type N connector with a quarter-wave transformer. When this is coupled out through a reactive element in the 50-ohm line, by feeding through a tuning stub, the circuit can be tuned from 1000 to 1500 mc. The output power is somewhat in excess of one milliwatt.

A modification of this same circuit is shown at the left center of Fig. 12. This is a ring oscillator derived from the quarter-wave circuit by closing it around onto itself. The top plate is the ring and the under copper foil is a ground plane. A diode and resistor are inserted, as before, at appropriate points. A radial oscillator is at the top, with the resistor in the center of the disk and the diode an appropriate distance out along a radius. This oscillator has worked to 2300 mc, but with no output.

The final circuit, at the center-right in Fig. 12, is a waveguide made from a stripline by closing up all edges. The input circuit incorporates the diode and resistor, which are placed between the end of the top lead and the copper sheet; the output is again matched to the 50-ohm type N connector. This unit has delivered power to the 50-ohm line but is not yet fully tested; it is designed for operation from 3000 to 6000 mc. At present, the full possibility of this higher-frequency range has not been achieved because the inductance of the diode case is too high to permit stabilization of the d-c circuit. To get a milliwatt at 6 kmc, the diode case inductance must be reduced to $10^{-10}$ henry.

The simplicity and compactness of these circuits are outstanding. The small size can be appreciated by scaling them against the output connectors, which cover a large share of the entire construction. Obviously, miniature connectors are going to be in demand.

### PACKAGING TUNNEL DIODES

For the kind of circuitry in Fig. 12, the tunnel diode must of course be in a flat package, the most desirable geometry in microwave applications, for several reasons: First, the tunnel diode is inherently a low-impedance unit, since it has a large capacitance ranging from 10 to 100 $\mu\mu$f for commercially available units. The flat packaging reduces the series inductance and still does not introduce appreciable capacitance. Also, this type of mount is naturally very small and permits high-packing density.

Fig. 13 is a random assortment of tunnel diodes obtained from various electronic equipment manufacturers in exchange for RCA sample units.

Several RCA tunnel diodes are shown in the bottom row of Fig. 13, with experimental stripline mounts at the left; the others are three "mount" variations of the high-frequency package now in production. Probably some version of the mount with pigtail leads will be the cheapest and so the popular one for low frequencies. In the uhf region, however, the stripline mount becomes desirable because of its inherently smaller inductance. Some form of this will probably be used for microwave devices.

### INHERENT DIFFICULTIES

Along with the attractive features of the tunnel diode, the circuit engineer should be aware of certain application limitations. Negative-resistance diodes are in a class different from ordinary devices and generally cannot be directly substituted into existing circuits.

The tunnel diode suffers with all diodes in being bilateral for small signals—it has nothing to define *in* from *out*. In general, the engineer must build into his circuits the function that the extra electrode in the triode serves—isolation between input and output. This is a relatively new problem and one for which no general solution yet exists.

Also, the tunnel diode is by its nature a low-impedance, low-power device. Since its voltage swing is less than a volt, power output comes only from large current, which means low impedance. For germanium diodes, an approximate relationship for available power is $P = (2 \times 10^{-3})/R$ watts, where $R$ is the resistance of the circuit. From a 10-ohm diode, 0.2 milliwatt of power is obtained. While it is conceivable that tunnel diodes can be made with a milliohm resistance, it is not so obvious that milliohm circuits can be readily made; if they can be produced, then 2-watt diodes are possible.

### CONCLUSIONS

The tunnel diode is a simple, low-power, low-noise device which is usable from d-c to the centimeter range. It is smaller, more rugged, and more stable against changes in ambient conditions and radiation than other active semiconductor devices. However, it is of lower impedance than present devices, and it is bilateral.

These factors mean general application of tunnel diodes will require much circuit research. Probably the first uses on any scale will be in applications where a single operation is performed in a manner analogous to present functions, such as oscillators, discriminators, or detectors. The real potential of the device will only be disclosed as the new circuits and the device characteristics become adapted to each other.

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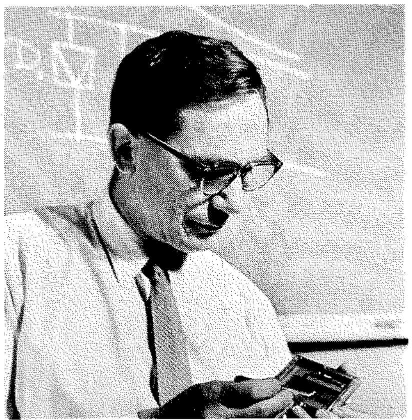

# TUNNEL DIODES IN LINEAR CIRCUITS

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THE FUNDAMENTAL reason for the great interest in the tunnel diode is its negative-resistance characteristic at certain forward-bias voltages.[1] The tunnel diode makes possible negative-resistance amplifiers having low noise and high gain at high frequencies,[2] allowing features that are particularly attractive for AM, FM, and television receivers—small size, low power consumption, and potentially low cost.

Possible applications of tunnel diodes in communication-receiver circuits discussed herein include amplifiers, oscillators, converters, and detectors. There are both advantages and problems in such applications, and results have been measured to indicate the performance of various linear circuits using tunnel diodes.

### GENERAL CHARACTERISTICS

The general characteristics of the tunnel diode (a two-terminal device) are shown in Figs. 1a, b, and c. The d-c current-voltage characteristics are shown in Fig. 1a; between the bias voltages $v_1$ and $v_2$, the tunnel diode exhibits a negative-resistance or negative-conductance characteristic. In this region, the negative resistance is a function of bias (see Fig. 1b) and has a definite inflection point where the negative resistance is minimum or the negative conductance is maximum, as shown in Fig. 1c.

When a tunnel diode is biased in this negative-resistance region with a d-c load line having a slope steeper than that of the negative resistance of the diode (Fig. 1a), many linear applications can be achieved. A d-c load line having a slope flatter than the negative-resistance slope of Fig. 1a results in bistable operation of the tunnel diode; this is only desirable for large-signal applications such as relaxation oscillators and switching circuits.

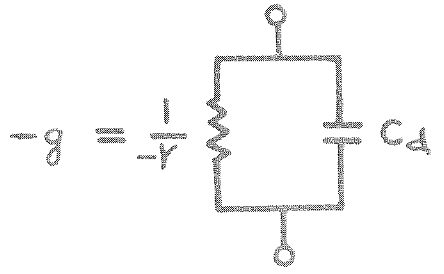

Fig. 2—Tunnel-diode approximate equivalent circuit.

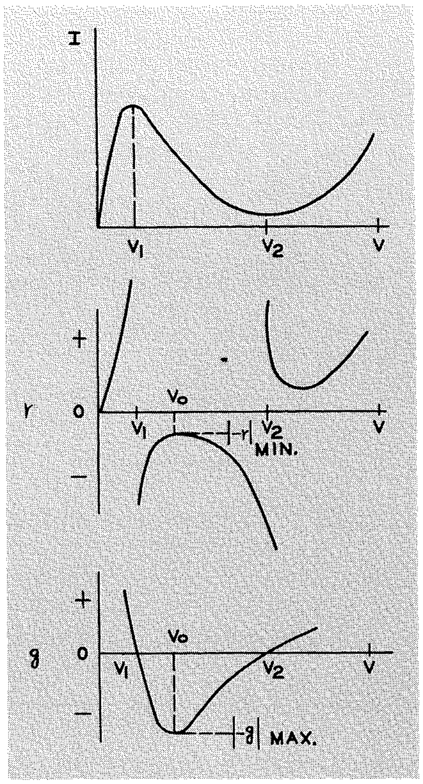

Fig. 1—General characteristic wave shapes for the tunnel diode.

Fig. 2 shows the approximate equivalent circuit for the tunnel diode, neglecting spreading resistance and series inductance. The figure of merit of this device can be expressed[3] as

$$\text{Figure of Merit} = \frac{1}{2\pi \, r \, C_d} \qquad (1)$$

At the present time, figures of merit as high as a few kilomegacycles have been realized, indicating that the useful frequency range extends into the microwave region.

### AMPLIFIERS

The shunt circuit shown in Fig. 3 can be used to operate the tunnel diode as an amplifier. (For simplicity, d-c bias circuits are not included in the circuit diagrams shown in the remainder of this paper.) The negative conductance of the diode cancels a portion of the positive conductance in the circuit. For stable operation, the total conductance, as viewed from the current-generator source into the circuit, must be positive, or $g_s + g_L > | -g|$ In order to maintain this relationship for all bias voltages, the inequality $g_s + g_L > | -g|$ must be fulfilled when $| -g|$ is at its maximum value (Fig. 1c). Since $| -g|$ can only decrease with changes in bias, the stability of the circuit is ensured, and the criteria for stability is:

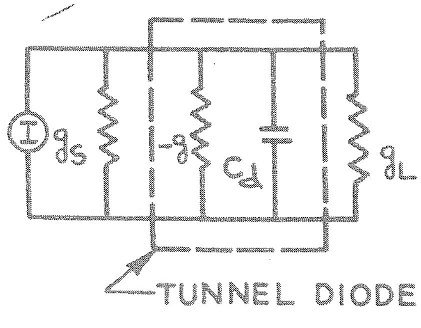

Fig. 3—Low-pass tunnel-diode amplifier. $g_s$=source conductance, $g_L$=load conductance.

$$g_s + g_L > | -g_{\text{max}}| \qquad (2)$$

The available power gain of the circuit shown in Fig. 3 is:

$$\text{P.G.} = \frac{4 g_s \, g_L}{(g_s + g_L - g)^2 + \omega^2 C_d^2} \qquad (3)$$

At low frequencies, the power gain becomes:

$$\text{P.G.} \, |_{d\text{-}c} = \frac{4 g_s \, g_L}{(g_s + g_L - g)^2} \qquad (4)$$

Fig. 4 is a plot of Equation 4 showing the effect on the power gain of a mismatch between the source and load for a given negative conductance. For a given power gain, the greater the ratio $| -g| \; / g_s$, the less critical the circuit adjustment becomes for stable operations; e.g., with a 20-db gain, with $g_s = g_L$, and $| -g| \; / g_s = 1.8$, the terminal conductance $g_s$ or $g_L$ can be decreased 20 percent before oscillation occurs. When $g_s$ and $g_L$ vary simultaneously in the same direction, the tolerable variance for $g_s$ and $g_L$ reduces to 10 percent.

For the same gain of 20 db, with $g_L = 0.1 \; g_s$, and $| -g| \; / g_s = 1.04$, then $g_s$ can decrease about 6 percent and $g_L$ about 60 percent. Therefore, the wide variation of terminations presents a stability problem in using tunnel diodes in amplifier circuits.

The cut-off frequency $f_c$ at which the power gain is 3-db down is given by:

$$f_c = \frac{g_s + g_L - g}{2\pi C_d} \qquad (5)$$

By adding an appropriate inductance, a bandpass tunnel-diode amplifier is possible, as shown in Fig. 5. Under such conditions, the power gain is:

$$\text{P.G.} = \frac{4 g_s \, g_L}{(g_s + g_L - g)^2 + (\omega C_d - 1/\omega L)^2} \qquad (6)$$

At resonance, Equation 6 reduces to Equation 4, and the 3-db bandwidth, $2\Delta f$, is (assuming the resonant frequency $f_o >> 2\Delta f$):

$$2\Delta f = \frac{g_s + g_L - g}{2\pi C_d} \qquad (7)$$

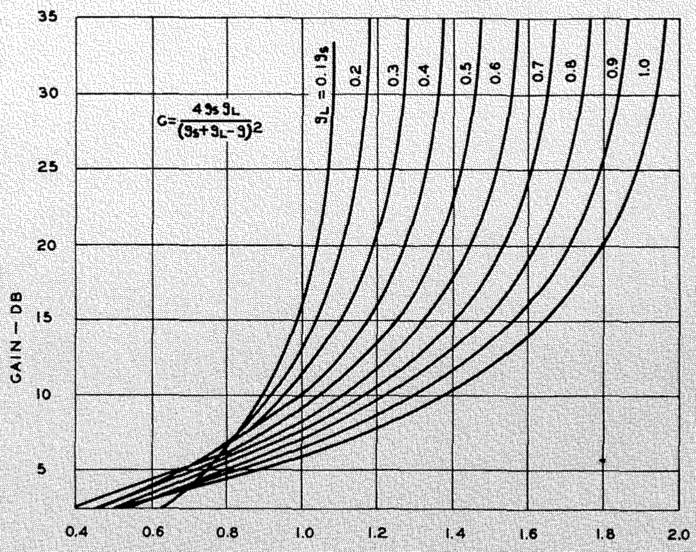

Fig. 4—Gain versus mismatch of terminal conductances.

Equations 5 and 7 are identical, and the voltage-gain–bandwidth product in both the low-pass and band pass amplifier is:

$$\sqrt{\text{P.G.}} \times \text{B.W.} = \frac{\sqrt{g_s\, g_L}}{\pi C_d} \qquad (8)$$

Equation 8 shows that for a given negative conductance (with $g_s$ and $g_L$ chosen, and $C_d$ a constant) the voltage-gain–bandwidth product is independent of the negative conductance of the tunnel diode. If value of $C_d$ of the tunnel diode can be lowered, then the voltage-gain–bandwidth product of Equation 8 can be improved. Equation 8 also states that the voltage-gain–bandwidth product is maximum when the $g_s = g_L$. For a very-high-gain amplifier, where $g_s + g_L$ is very close to $|-g|$, the product approaches the figure of merit of the device as expressed in Equation 1. Therefore, the voltage-gain–bandwidth product of an amplifier is always less than the figure of merit of the device. The measured performance data of several tunnel-diode amplifiers are shown in Table I.

Another important consideration in amplifier design is noise factor. K. K. N. Chang of the RCA Laboratories has derived the following expression for the noise factor of a tunnel-diode amplifier[2]:

$$\text{N.F.} = 1 + \frac{T}{T_o}\left(\frac{g_l}{g_s} + \frac{g_L}{g_s} + \frac{g_e}{g_s}\right) \qquad (9)$$

Where, $T_o$ = reference temperature of the source conductance $g_s$, $T$ = ambient temperature, $g_e$ = equivalent shot noise conductance of the diode, and $g_l$ = circuit loss conductance.

In accordance with Equation 9, tunnel-diode amplifiers have potentially low noise factors; noise factors as low as 4 db have been measured.

For a particular amplifier application where the load and source conductances are fixed, almost any desired power gain can be achieved by modifying the tunnel diode conductance either by transformer-coupling[2], or by resistance-coupling (as shown in Fig. 6). The circuit in Fig. 6a shows how the effective negative conductance of the tunnel diode is lowered by an external conductance $g_o$. However, when amplification is desired, $g_o$ cannot be made equal to or greater than $|-g|$ of the diode. Fig. 6b indicates that the effective negative conductance is raised by the presence of a series conductance. To ensure amplification, the condition $|-1/g| > 1/g_o$ must be satisfied.

**TABLE I—PERFORMANCE OF TUNNEL-DIODE AMPLIFIERS**

| Diode | Low-Pass Amplifier | | | Band-Pass Amplifier | | | |
|---|---|---|---|---|---|---|---|
| | d-c gain *db* | $f_c$ *mc* | $\sqrt{\text{P.G.}} \times f_c$ *mc* | $f_o$ *mc* | P.G. *db* | $2\Delta f$ | $\sqrt{\text{P.G.}} \times 2\Delta f$ *mc* |
| 1 | 12.5 | 85 | 357 | 75 | 29.5 | 4.6 | 138 |
| 2 | 14.1 | 32 | 163 | 110 | 28 | 4.6 | 116 |
| 3 | 14.5 | 32.5 | 172 | 165 | 25 | 5.5 | 98 |

### OSCILLATORS

The negative-resistance characteristic of the tunnel diode also makes it very suitable for oscillator applications; Fig. 7 shows a typical tunnel-diode oscillator circuit. Either harmonic (sinusoidal) or relaxation oscillation can be attained with the appropriate selection of the circuit values and tunnel-diode parameters shown in Fig. 7. Harmonic oscillation can be obtained when the following conditions are fulfilled:

$$\frac{R_L}{L} < \frac{|-g|}{C_d}$$

$$\frac{1}{LC_d} > \left(\frac{R_L}{2L} + \frac{|-g|}{2C_d}\right)^2 \qquad (10)$$

Similarly, relaxation oscillation can be achieved when these conditions are satisfied:

$$\frac{R_L}{L} < \frac{|-g|}{C_d}$$

$$\frac{1}{LC_d} < \left(\frac{R_L}{2L} + \frac{|-g|}{2C_d}\right)^2 \qquad (11)$$

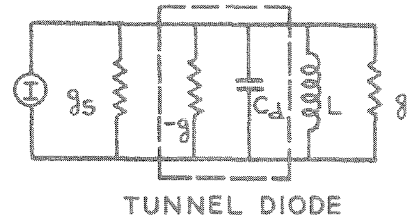

Fig. 5—Band-pass tunnel-diode amplifier.

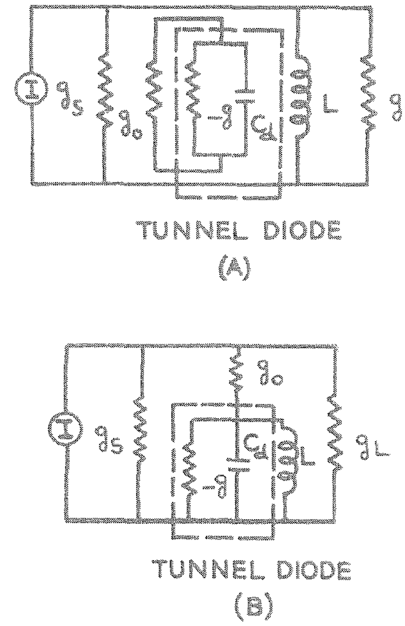

Fig. 6—Various band-pass amplifier configurations.

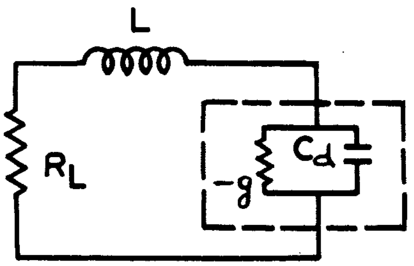

Fig. 7—Tunnel-diode oscillators.

For both oscillator applications, $R_L$ can also act as the d-c bias resistor. In communication receivers, the harmonic oscillation content is an important requirement for local-oscillator applications. Several tunnel-diode harmonic oscillators were designed and built; some of the measured results are summarized in Table II.

**TABLE II PERFORMANCE OF TUNNEL-DIODE HARMONIC OSCILLATORS**

| Description | Frequency of Operation, mc* | Power Output, mw |
|---|---|---|
| Lumped constant | 200 | 0.5 |
| Distributed | 300 | 0.5 |

* Frequency of oscillation as high as 3K mc has been reported.[4]

### MIXER CONVERTER

As a frequency-converting device, the tunnel diode offers two major advantages over a conventional diode: first, the tunnel diode provides conversion gain, while the conventional diode does not; and second, since the tunnel diode is a negative-conductance device—it can be used as a self-oscillating converter.

As a mixer, two modes of operation can be achieved: operation in the negative-resistance region to cancel portions of positive resistance of the circuit and attain conversion gain, and operation in the positive-resistance region where bias can be anywhere from 0 to $v_1$ of Fig. 1. Conversion energy is obtained from the pump power. However, the pump power required is greater because it is necessary to swing into the negative-resistance region to attain conversion gain. In this operation, low noise and high conversion gain have been reported.[5] Fig. 8 illustrates the measured performance of a mixer circuit in which the tunnel diode is biased in the negative-resistance region.

The self-oscillating converter shown in Fig. 9 uses a diode that oscillates at a frequency determined primarily by the elements $C_2$ and $L_2$, with the value of $C_2$ being made smaller than that of $C_d$. The r-f signal is coupled to the diode through the $L_1$ and $C_1$. Converted i-f signal is obtained from the inductance $L_3$; the frequency bandwidth is very narrow, yet adequate for FM receiver applications.

### DETECTORS

The ability of the tunnel diode to sustain oscillations at very low levels results in several advantages when the device is used in synchronous-detector circuits. Such a detector is shown in Fig. 10. Linear detection can be maintained at extremely low signal levels, and by controlling the mutual coupling of the transformer, either AM or FM detection is possible.

An AM detector which uses a combination of a transistor envelope detector and a tunnel-diode synchronous detector is shown in Fig. 11. This circuit makes use of the facts that the synchronous detector operates at very low power levels and the transistor envelope-detector operates at large signal levels. Thus, linear detection is possible over a much wider range than that of standard detector circuits.

### POTENTIAL PROBLEMS

Although tunnel diodes can be used for many linear-circuit functions, there are several potential drawbacks confronting circuit designs for communication receivers:

1) Since the tunnel diode is a two-terminal device, the input and output terminals are identical and therefore have no isolation. Attempts to cascade two or more such devices presents a severe problem.
2) The presently available tunnel diodes are small-signal devices.

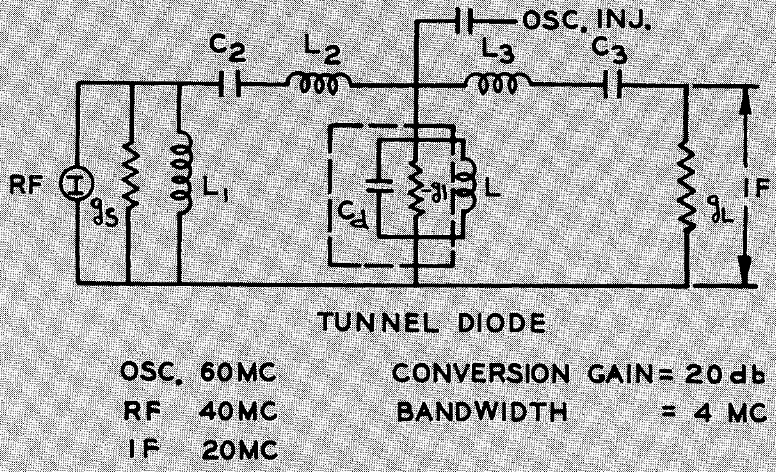

Fig. 8—Tunnel-diode mixer circuit.

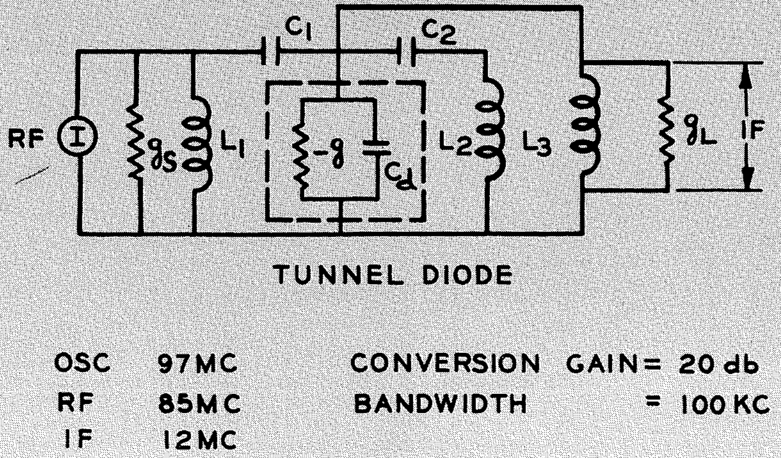

Fig. 9—Tunnel-diode converter.

When a large signal is applied, the tunnel diode acts as a limiter and introduces nonlinear distortion which cannot be tolerated for most linear applications.

3) Changes of terminal impedances cause instability.
4) Spurious oscillation may be caused by the stray inductive reactance of the tunnel diode structure.

Nevertheless, circuit designers are optimistically attacking all these problems at the present state-of-the-art, since solutions must be discovered before the tunnel diode can be widely employed in the communication field.

## CONCLUSION

Based on preliminary investigations and measurements, the use of tunnel diodes in different communication circuit functions has been accomplished with varying degrees of success.

In some applications—where linear operation at large signals is important—the presently available tunnel diodes cannot be used. In other functions—such as low-level r-f amplifiers, converters, and mixers—the high-frequency and low-noise properties of the tunnel diode offer distinct advantages. In these applications, however, maintaining constant source and load impedances and the cascading of amplifiers are problems requiring solution.

In switching circuits, oscillators, and detectors, the high-speed properties of the tunnel diode can be used with significant advantage.

## ACKNOWLEDGMENT

The authors wish to acknowledge the contributions to this paper by their colleagues, particularly E. Miller, (noise factor measurements), D. J. Carlson and W. Gilmore (converter circuits), G. E. Theriault (detector circuits), M. Cooperman (oscillators), and O. Ramanis (device measurements).

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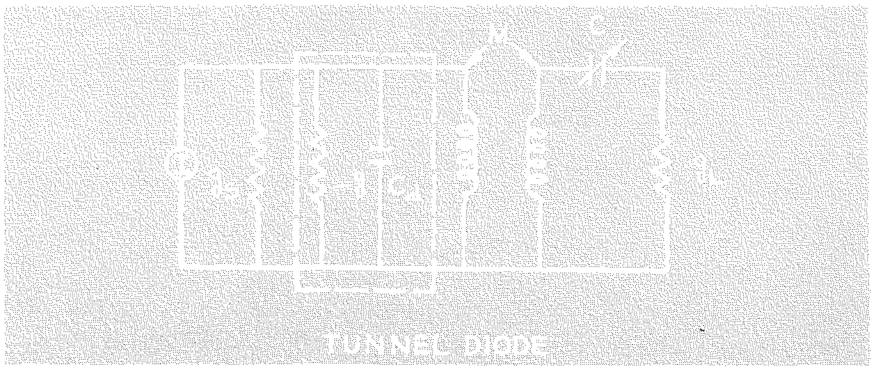

Fig. 10—Tunnel-diode synchronous detector.

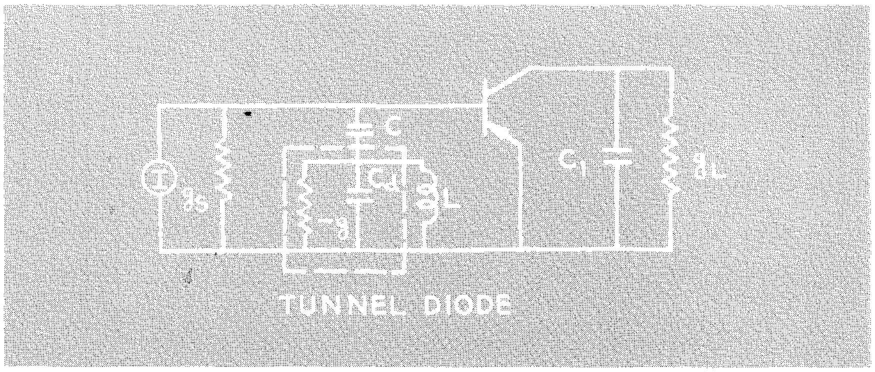

Fig. 11—Combination transistor and tunnel-diode detector.

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# TUNNEL DIODES IN DIGITAL COMPUTERS[*]

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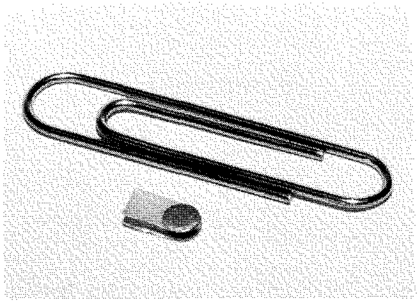

Fig. 1—RCA tunnel diode.

With the coming of age of electronic computers during the past decade, efforts have been made to constantly seek new applications of these valuable tools. Electronic computers have ranged from a desk-top package to giant systems occupying thousands of square feet. In addition to diversity of application and variation in size and complexity, the computer industry has continually increased the operating speed of its equipment. This was caused in part by scientific necessity, in part by economic conditions. Various means have been employed to achieve increases in speed as computers evolved from those using relays, then tubes, and recently transistors. Today the computer industry appears to be on the threshold of another significant step-up in speed by the use of tunnel diodes (Fig. 1).

Tunnel diodes are characterized by their singular property of quantum-mechanical tunneling. It was the discovery of this tunneling effect in heavily-doped semiconductor junctions that triggered the considerable tunnel-diode development effort. Very high cut-off frequencies have been obtained, and predictions of even higher frequencies have suggested their use as elements in ultra-high-speed computer circuits. This breakthrough in high-speed semiconductor technology, coupled with the simplicity of the device and its stability, make it a prime candidate for computer application.

## BASIC TUNNEL-DIODE COMPUTER CIRCUITS

In order to satisfy the requirements for the logical and control portion of a computer, a set of logical building blocks is required. Normally associated with the logical building block is some means of signal amplification or fan-out, and signal standardization. In addition to the logical circuits, a compatible storage device is required.

The most critical requirement in the logical building block is that of amplification, since the other functions may be performed by passive elements. As the speed requirements are increased, the number of active elements which have an adequate gain-bandwidth product becomes the limiting factor. Two methods of amplification are possible with the tunnel diode: *linear* and *regenerative*. Both methods are illustrated in Fig. 2 by the static characteristic of a typical tunnel diode.

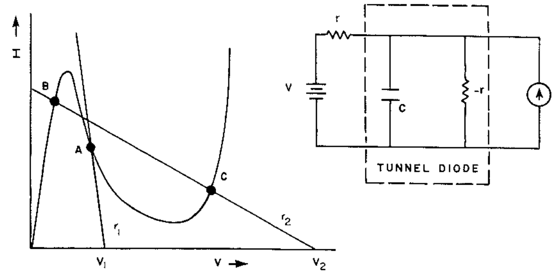

Fig. 2—Operating modes with tunnel diode.

### Linear Amplification

For linear amplification, an operating point is established in the most linear region of the negative resistance (Point A) by the load line $r_1$ and the d-c bias voltage $V_1$. The resulting circuit is also shown in Fig. 2.

If a small signal is applied from a constant current source, it is amplified by the factor:

$$\frac{|-r|}{|-r|\,(1+j\omega C r_1)-r_1}$$

Where, $r_1 < |-r|$, in order to establish a single stable operating point. As the signal amplitude is increased and the operating point swings into the higher negative resistance regions of the tunnel diode, the gain decreases, thus providing signal limiting. Examination of stability requirements indicates that series lead inductance must be extremely small, i.e.:

$$L_s < C\,r_1\,|-r|$$

Where, $C =$ capacity of the tunnel diode plus any additional shunting capacity. A further disadvantage is the requirement for a linear and stable region in the negative portion of the diode characteristic. This requirement arises from the sensitivity of the gain of the circuit to changes in negative resistance.

### Regenerative (Switching) Mode

The regenerative, or switching, mode of operation is obtained by increasing the load line to $r_2$ and the voltage source to $V_2$, thus intersecting the diode characteristic in three places; however, only points B and C are stable operating points. If the circuit is operating at point B and a positive current step of sufficient amplitude is applied, the operating point will switch to point C. Correspondingly, a negative input signal will switch it back to point B. Only input signals greater than the threshold will cause switching, a nonreversible action; that is, if the threshold is exceeded, the circuit will switch to the high state and remain there even when the input signal is removed. Therefore, it has memory, an application which will be discussed later.

An advantage of the switching mode is its nonsensitivity to the exact linearity or shape of the negative-resistance region of the tunnel-diode characteristic. Slight irregularities in the negative characteristic will have a negligible effect in the switching action. Amplitude standardization is also obtained in the switching mode by the diode characteristic in the positive-resistance regions, since the circuit will switch to the relatively constant-voltage high state regardless of the amplitude of the input signal.

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[*] This program has been sponsored by the Dept. of the Navy, Bureau of Ships, NObsr 77523.

Gain may be obtained in the switching circuit by adjusting point B so that it is near the current threshold of the diode characteristic. Then, a relatively small input current can be used for triggering. Once switching to point C has occurred, considerably greater current is available to the load.

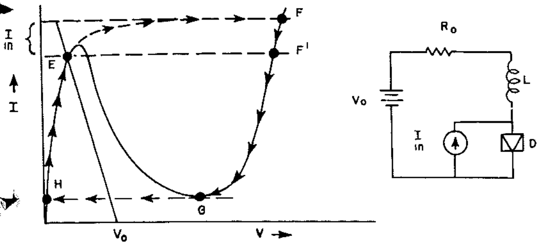


Fig. 3—Monostable circuit.

Inasmuch as static storage is not desired in a basic logical element, some means of returning to the original operating point is desirable so that repetitive logical functions may be performed. A circuit accomplishing this and other logical functions is described in the following sections.

### Basic Monostable Circuit

The basic monostable circuit (Fig. 3) using tunnel diodes was chosen for its low power-dissipation properties, the low d-c power requirement of the circuit, and the gain available. The latter results from the storage characteristics, which cause the device to switch on an almost horizontal path to the high state, thus retaining all the current available for useful output.

In the monostable circuit of Fig. 3, the static load line is determined by resistance $R_o$ and voltage $V_o$. If $R_o$ is less than the minimum dynamic negative resistance of the diode, only a single operating point will exist. The gate will be stable in the low state if $V_o$ is adjusted with its operating point at E. The dynamic load line is determined by the inductive time constant $L/R_o$. When $L/R_o$ is long compared to $T_s$ (switching time), the load line becomes that of constant current.

If a small step of current $I_{in}$ is applied to the diode, the operating point will switch to the high voltage point F, along the constant-current path shown. Removal of the input will cause the operating point to move to F′. At this point, the energy stored in the inductor must be dissipated before the circuit can return to its original operating point. As the energy in the inductor decreases, the operating point moves along the diode characteristic to the point of minimum current at G. Upon reaching this point, switching occurs once again along a constant-current path to point H. The cycle of operation is completed by a recovery region where the energy in the inductor builds up to its original level. During this time, the operating point moves up the diode characteristic to the starting point. The waveform resulting from this cycle of operation is shown in Fig. 4. Because the circuit operation is controlled by its inductance, the output waveform contains essentially no d-c components when d-c source impedance is low, and may be a-c coupled without requiring d-c restoration.

The monostable circuit is then the analog of a one-shot circuit. It requires an input of sufficient amplitude and duration to drive the tunnel diode into the region where it exhibits a negative resistance and then will continue its cycle unaided.

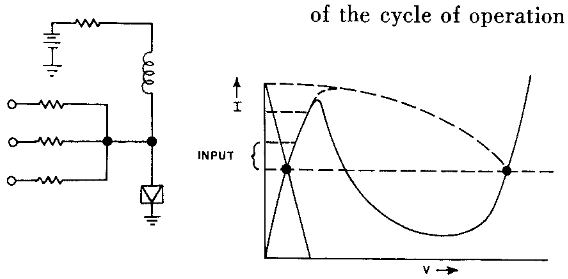
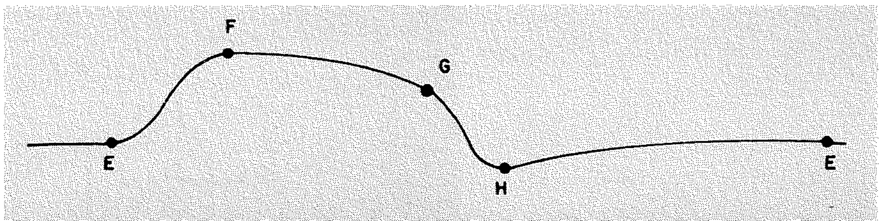


Fig. 5—Three-input *AND* gate.

### Monostable Circuit Applications

Since the shape of the characteristic at the current knee indicates a very abrupt change in dynamic resistance from a low positive to a low negative resistance, a rather well defined threshold is established. Only inputs exceeding this threshold will trigger the circuit. If a standard input current to a circuit is adopted by assuming constant voltage swings for triggered gates and fixed coupling impedances between gates, a logical threshold function can be performed. For example, if the static operating bias is adjusted so that one input is required to exceed the threshold, an *Or* function is performed. Similarly, if all inputs are required to exceed the threshold, an *And* function is performed. Since the coupling impedance is high compared to the diode impedance, the inputs can be considered as current sources during the region that the thresholding operation is being performed. Fig. 5 shows a three-input *And* gate with proper biasing point and current inputs. By a slight increase in the operating point bias, the circuit may be made to trigger on two of its three inputs. The logical function now performed is that of a *Majority* gate.

Fig. 4—Monostable circuit waveform.



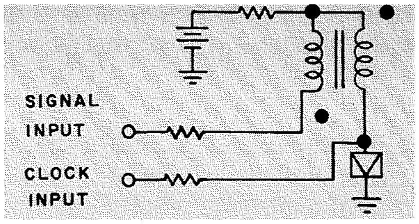


Fig. 6—Inverter using a transformer.

To complete a set of logical functions, negation or inhibition is required. The basic circuit operation that provides these functions is signal inversion. In vacuum-tube and transistor circuitry, the inversion is inherent to the basic operation of the devices. Since the tunnel diode is a two-terminal device, no inversion is provided in the device itself, and inversion must be performed by the circuit. The use of a transformer to perform this function is shown in Fig. 6. The transformer replaces the inductance of the original circuit and provides a 1-to-1 inversion. In order for the timing of the cycle of operation to remain constant, the transformer has an additional requirement of controlled primary inductance of the same value as the inductor it replaces.

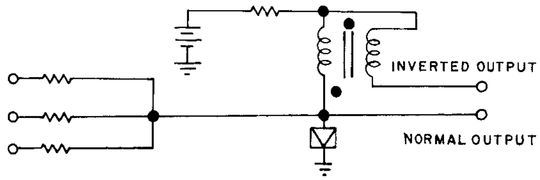


Fig. 7—Logical building block.

Therefore, there are no additional recovery problems involved due to variations in pulse rates through the transformer. Negation is performed by inverting the signal through a transformer and using this inverted output to cancel or inhibit an *Or* gate driven by a clock pulse with appropriate timing. Any logical signal may also be inhibited in a similar manner. A single logical building block may be constructed as shown in Fig. 7. The inputs to the gate form an *And*, and either normal or inverted outputs are available. All logical functions may be generated with such a circuit.

For example, a combination of two bistable tunnel-diode circuits, using the transformer in place of the inductors, leads to an analog of the Eccles-Jordan vacuum-tube circuit (Fig. 8). If a positive-going pulse is fed into both inputs simultaneously, the circuit will assume a condition where one is stabilized in the high state and the other in the low. The diode assuming the high state will be the one having the lowest knee current, since it will trigger first and inhibit the other. If the initial condition assumes diode A to be high, the following action is expected: a trigger pulse is applied equally to the two diodes, whereupon the current in A is relatively undisturbed, since it remains in the high state momentarily. However, B is triggered through to its high state causing a high rate of current change to induce a pulse across the transformer in B side which is inverted in the A winding, thus causing the combination of inhibiting the supplied trigger as well as a negative-going pulse switching the diode back to the low state. This circuit element provides both *1* and *0* outputs as well as *set*, *reset*, and *count* inputs.

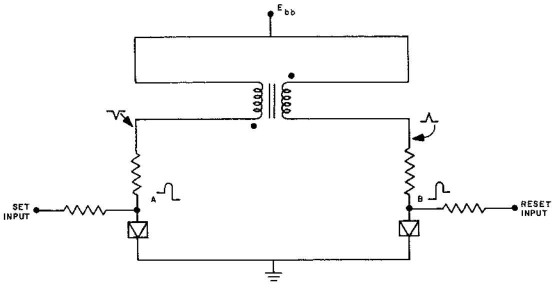


Fig. 8—Tunnel-diode flip-flop.

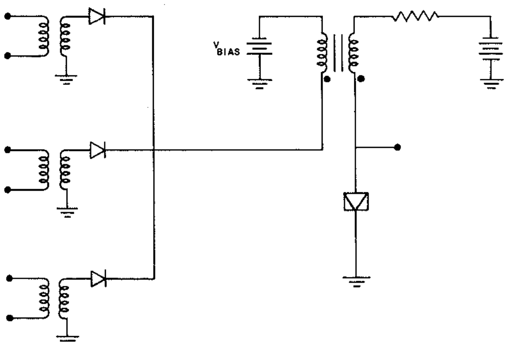


Fig. 9—Biased *OR* gate.

**Directionality Requirements**

Investigation of the operation of an *Or* (i.e., a threshold) gate reveals that all inputs need not be high to trigger the gate. Since inputs and outputs are common to the gate, the inputs which are not high during this triggering action will act as loads if the coupling elements between gates are bilateral. In this situation, false triggering of some gates is possible. To prevent such action, a directional coupling element is required. The use of a diode rectifier as the coupling element satisfies the basic requirement of providing directional isolation because of the difference in forward and reverse impedance. An additional requirement for the coupling element is that it must present the proper forward impedance required by the characteristics of the tunnel diodes used. Also, this forward impedance must be obtained with the signal-voltage swings available from the tunnel diode.

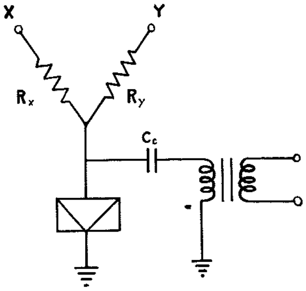


Fig. 10—Memory cell.

If the signal swings are not adequate, a forward-biasing scheme (Fig. 9) may be employed. This circuit requires one additional bias voltage, but provides the buffering function required to give these devices directionality.

In situations where all inputs to a gate must be in the high state for triggering to occur, such as in an *And* gate, directional coupling is not required and resistance coupling may be employed. In this case, one must consider the effects of the inputs upon the total fan-out capabilities of the gate. Examination of the sequence of events in the triggering of an *And* gate indicates that loading caused by the inputs to the gate does not occur at the same time as the loading caused by the driven elements.

## TUNNEL DIODES IN MEMORIES

To achieve a bistable element using a tunnel diode, it is only necessary to place a resistive load across the diode. If this resistive load is then divided into an X-selection resistor and a Y-selection resistor, the element may be selected in an X-Y matrix. Each resistor as shown in Fig. 10 will then deliver half the switching current. The peak and valley currents ($I_p$ and $I_v$ in Fig. 11) of the tunnel diode represent adequate thresholds (coercive currents), allowing the current-driven memory cell to be selected out of a random-access coincident matrix.

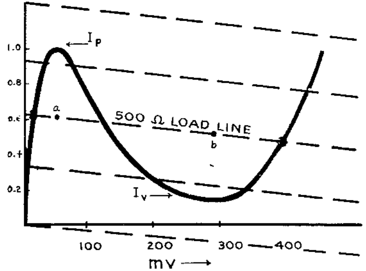


Fig. 11—Tunnel-diode d-c characteristic and load line.

To sense a memory location's stored information, the location is read destructively. That is, the selection current direction is referenced with respect to the tunnel diode's *0* state, always leaving the tunnel diode in the *0* state; upon selection, a switching tunnel diode or nonswitching tunnel diode is said to have stored a *1* or *0*, respectively. Therefore, if the memory cell is storing a *1* when selected and read, the information is destroyed and must be regenerated. Destructive read-out is used because it results in the greatest voltage change across the tunnel diode. The transformer coupling in Fig. 10 gives the largest corresponding signal output. The large output is required to reduce the over-all gain-bandwidth in the regeneration loop.

The basic method for packaging the memory is shown in Fig. 12. It involves strips with components which connect to the X-selection lines, strips with components which connect to the Y-selection lines, and strips with components which connect to the ground plane. All connections can be made by dip-soldering.

An experimental 8 x 8 bit memory plane (Fig. 13) using this basic packaging method was constructed for evaluation in an ultra-high-speed–computer research program and was successfully tested at tenth-microsecond memory speeds. The 8 x 8 bit plane was built to simulate a 32 x 32 bit plane with the same over-all dimensions, and the dimensions per memory cell were scaled by a factor of four. To make the memory cell practical, a printed-circuit transformer was used.

## SUBSYSTEM USING TUNNEL-DIODE CIRCUITS

After thorough evaluation of individual logic circuits, problems peculiar to combinations of these circuits were

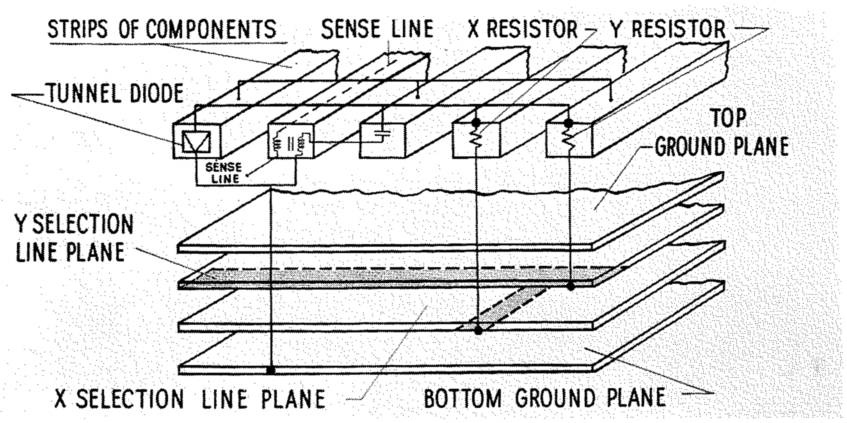


Fig. 12—Basic method for packaging memory cells.

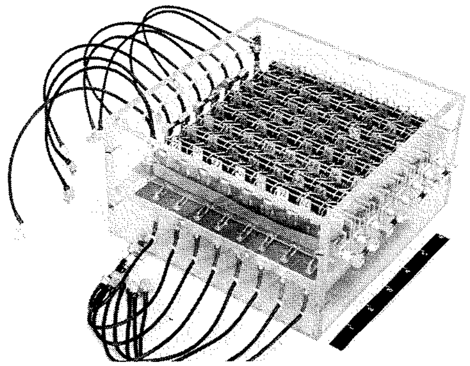


Fig. 13—An experimental 8 x 8 bit memory plane.

investigated. These problems were power supply impedance requirements, ground currents, synchronism in circuits, and effects of tolerances of devices and circuits. Circuits were arranged such that all forms of gates would be utilized and that the machine would check its operation and errors. Reset functions were included to make it self-correcting. The subsystem is shown in Fig. 14. The tunnel diodes used are of the 20-ma, 150-$\mu\mu$f type which set the stage delay to about 5 nanoseconds (5 x $10^{-9}$ second). Operating from a signal source of 4 mc, the actual information rate within the bit loops is equivalent to that of an 80-mc clock rate. Experience with the subsystem of 71 tunnel diodes shows that a very-low-impedance power supply is required. Ground-current difficulties were eliminated through good engineering precautions. The effects of tolerances on both device and circuitry have been controlled to the extent that wide parameter variance may be tolerated. However, synchronism requirements and sensitivity to variations in gate switch times and wiring delays will demand development of a more asynchronous logical building block.

## CONCLUSION

Application of the tunnel diode will undoubtedly revolutionize the computer industry because of its many favorable characteristics; but, to overcome its few drawbacks will still require imagination and inventiveness by the circuit-design engineer.

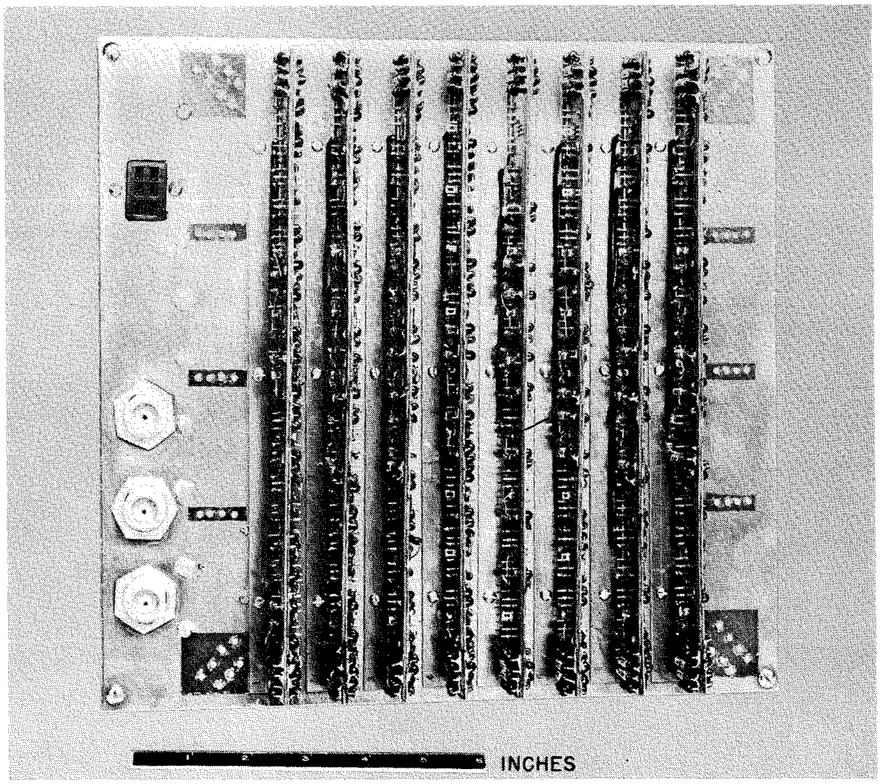
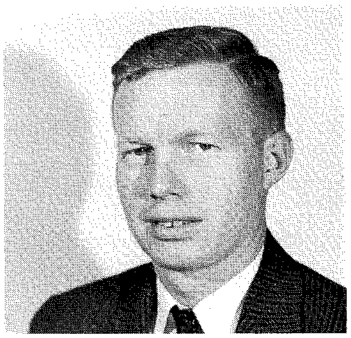
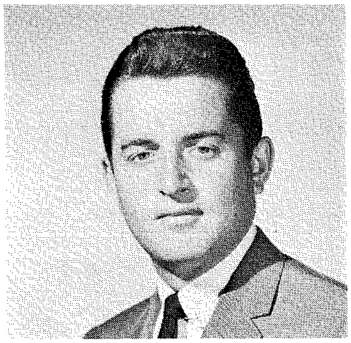


Fig. 14—Tunnel-diode subsystem.



**RICHARD H. BERGMAN** received his B.S. degree in Electrical Engineering from Rutgers University in 1953. He joined RCA in 1953, working primarily on pulse and digital circuitry, analog-to-digital conversion, ultrasonic measurements, and the application of tunnel diodes to computers. He served in the U.S. Army Signal Corps from 1954 to 1956, returning to RCA upon completing active duty. Mr. Bergman is a member of IRE, Tau Beta Pi and Eta Kappa Nu.



**MELVIN M. KAUFMAN** received his B.E.E. degree from the City College of New York in February, 1956, and his M.S.E.E. degree from the University of Pennsylvania in February, 1960. He is presently continuing his studies at the University of Pennsylvania. Mr. Kaufman started with RCA in 1956 as an Engineering Trainee, and after a short survey period, came to the Electronic Data Processing Division where he was assigned to the investigation of resistor transistor gates. He then participated in the development of a transistorized partial-switching ferrite plate memory. At the advent of an ultra-high-speed computer project, he became associated with development and construction of millimicrosecond memories. He first worked on diode-capacitor storage techniques, which resulted in a laboratory model with ten-millimicrosecond cycle time. He is presently working on tunnel-diode storage techniques. He is a member of Eta Kappa Nu.

# GALLIUM ARSENIDE TUNNEL DIODES

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Tunnel diodes made from gallium arsenide have several advantages over germanium and silicon units. For example, gallium arsenide has a relatively large band-gap energy which makes possible a higher operating temperature than can be attained with similar devices made from germanium or silicon. It also exhibits a high ratio of peak current to valley current—ratios of ten or twenty to one are common, and much higher values are occasionally observed.

In addition, gallium arsenide tunnel diodes have a voltage-current characteristic that permits a greater voltage swing and, consequently, a higher power output than can be obtained from germanium or silicon units.

### EQUIVALENT CIRCUIT

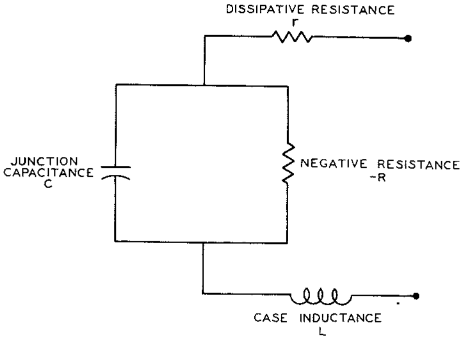
The equivalent circuit of the tunnel diode[1] is shown in Fig. 1. The negative-resistance component $R$ is paralleled by a fairly large junction capacitance $C$ which for a diode having a peak current of 50 milliamperes, may be about 20 $\mu\mu$f. The magnitude of the negative resistance $R$ is inversely proportional to the peak current of the device. For a typical diode having a peak current of 50 milliamperes, this resistance is about 4 ohms. From these values of capacitance and negative resistance, the gain-bandwidth product[1] $G\Delta f = 1/2\pi RC = 2000$ mc.

The series-inductance component $L$ results primarily from the package and is typically 0.4 m$\mu$h. The dissipative resistance $r$, about 0.7 ohm, establishes an upper limit to the maximum frequency of self-excited oscillation. This upper frequency limit $f_o$ is:

$$f_o = G\Delta f \left(\frac{R}{r_t} - 1\right)^{1/2}$$

Where: $r_t$ = the total dissipative resistance of the loop. For the diode discussed above, $f_o = 4300$ mc.

Fig. 1—Equivalent circuit of tunnel diode.



### TUNNEL CURRENT

The peak current of a tunnel diode[1] is proportional to exp $[-E_g(\epsilon M/N)^{1/2}]$, where: $E_g$ = the band gap in volts, $\epsilon$ = dielectric constant, $M$ = the electron-reduced effective mass in grams, and $N$ = the impurity density per cubic centimeter. This relation shows that the peak current is sensitive to variations in impurity density $N$. Preliminary measurements for gallium arsenide indicate about a five-fold increase in peak current for an increase in impurity concentration from $3 \times 10^{19}$ to $7 \times 10^{19}$ per cubic centimeter.

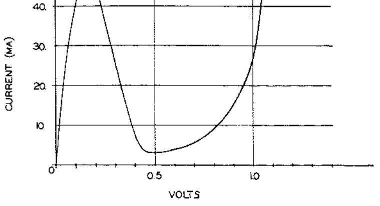
### CHARACTERISTICS

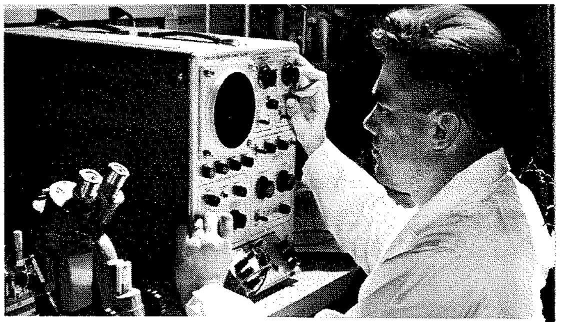
The voltage-current relationship of a gallium arsenide tunnel diode is shown in Fig. 2. The region of negative resistance occurs for bias voltages in the range from 0.1 to 0.45 volt. This range is fixed by the kind of semiconductor material used and does not depend on the geometry or processing of the device. The peak current and electrical capacitance of the diode are, however, largely dependent on the resistivity of the crystal and the area of the junction. These parameters can be varied to obtain the desired values of peak current and capacitance. Tunnel diodes presently being made have a capacitance of the order of 1 $\mu\mu$f for each milliampere of peak current. The peak current can be closely controlled during fabrication of the diode to within a few percent of the desired value.

### DEVICE FABRICATION

Fabrication of the tunnel diode starts with the diffusion of a p-type impurity, such as zinc, throughout a wafer of gallium arsenide under such conditions that the average resistivity of the wafer becomes about 0.002 ohm-centimeter. This diffusion occurs when the wafer and a small amount of zinc are sealed into an evacuated quartz tube about 1.5 cubic centimeters in volume. The tube is heated to a temperature of 1000°C for 16 hours so that the zinc diffuses through the wafer to convert it to a low-resistivity, p-type crystal.

Fig. 2—Voltage-current characteristic of gallium arsenide tunnel diode.





**ALFRED J. WHEELER** received the B.S. degree in Physics from Worcester Polytechnic Institute in 1951. From 1951 to 1955, he was employed by the Research and Development Laboratories of the Army Engineer Corps at Fort Belvoir, Virginia, where he worked on spectrophotometric measurements of camouflage materials. He joined the Semiconductor and Materials Division in 1955 as an applications engineer on transistor circuitry. A year and a half later, he transferred to the Advanced Development activity, and has since worked on the development of silicon transistors and the development of gallium arsenide rectifiers and tunnel diodes. While working at RCA, he has attended Stevens Institute of Technology, receiving his M.S. in Mathematics in June, 1960. He is a member of the IRE.

After one side of the wafer is lapped to the desired thickness, a thin film of gold is evaporated onto the lapped surface and alloyed to it to provide a low-resistance electrical contact. Nickel is plated over the alloyed surface. The opposite side of the wafer is etched slightly and then scribed so that the wafer can be broken into square pellets about 30 mils on each side.

A tin dot alloyed to each pellet forms the n-type side of the p-n alloyed-junction diode. The diodes are soldered into low-inductance packages designed to make effective use of the tunnel diode's high-frequency capabilities. Before being sealed, the units are etched to obtain the desired peak current.

The electrolytic etching is carried out in a solution of potassium hydroxide. The current characteristic of the diode is continuously observed while it is being etched by use of a curve tracer as the source of voltage. As a result, the peak current of each tunnel diode can be very close to the desired value.

### STATUS

The gain-bandwidth product of the best units at the present time (May 1960) is about 13,000 mc. Although this value represents very great progress, there are still problems to be solved in making such units stable for use in amplifier and oscillator circuits. Both series dissipative resistance and case inductance must be reduced. Device characteristics affecting noise in down-converter circuits must be studied and an improvement made in the voltage-current characteristic for optimum use in high-speed switching circuits. When these ends have been reached, the gallium arsenide tunnel diode will be a welcome addition to microwave work.

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1. H. S. Sommers, Jr., "Tunnel Diodes as High Frequency Devices," *Proc. IRE 47*, 67, (1959).

# DESIGN AND FABRICATION OF GERMANIUM TUNNEL DIODES

*by*

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THE TUNNEL DIODE has two unusual properties not normally found in transistors or vacuum tubes: it is naturally bistable, and it can be made to exhibit both stable states with no circuitry other than a battery and one resistor. Because the two stable states, or stable portions of the voltage-current characteristic, are joined by a region of negative resistance, the tunnel diode is capable of power gain or amplification. Even more important, however, is that the negative resistance is a *pure resistance* and is in no way associated with minority carriers, mobility, diffusion, drift, or any other frequency-limiting process.

### DESIGN CONSIDERATIONS

The RC time constant of the equivalent circuit (Fig. 1) is the fundamental frequency limitation of the tunnel diode. The gain-bandwidth product $G\Delta f$ is:

$$G\Delta f = \frac{1}{2\pi RC}$$

The figure of merit $f_o$ (the highest frequency at which the diode is capable of self-sustained oscillation) is:

$$f_o = \left(\frac{R}{r}\right)^{1/2}\left(\frac{1}{2\pi RC}\right)$$

where $r$ is the total positive resistance in the signal circuit. These equations show that for high-frequency operation, $R$, $C$, and $r$ should all be very small. The inductance in Fig. 1 is primarily due to the package into which the diode is placed, rather than to the diode itself. This inductance must be small, of course, if the tunnel diode is to operate at high frequencies.

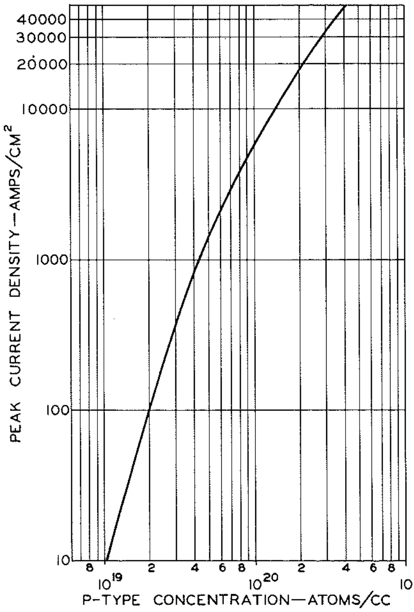


Fig. 2a—Variation of peak current $I_p$.

In addition to the equivalent-circuit parameters, other important characteristics of the diode include the voltages and currents at the peak and valley points of the characteristic curve and, in particular, the ratio between the peak current, $I_p$, and the valley current, $I_v$. For tunnel diodes to be really practical for such applications as computers, the characteristics of the diodes must be uniform from one to the next so that they are completely interchangeable. This uniformity is a primary factor in considering the possible methods for making tunnel diodes.

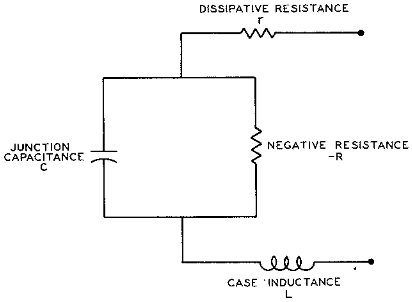


Fig. 1—Equivalent circuit of tunnel diode.

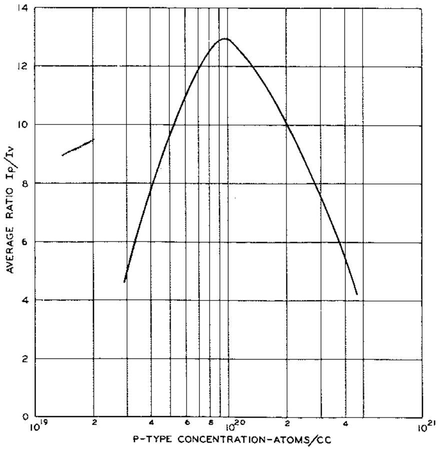


Fig. 2b—Peak-to-valley current ratio $I_p/I_v$ with impurity concentration.

### DESIGN PARAMETERS

Basically, only the following three design parameters of the tunnel diode affect its characteristics: 1) the doping in the p-type germanium (both the type of impurity present and its concentration); 2) the doping in the n-type germanium (also including both type and concentration); and 3) the p-n junction area or, for nonuniform doping, the effective area. Because there are seven parameters to be controlled (in addition to the inductance, a packaging problem) and just three independent variables, only certain specific combinations of parameters are possible, and in general, compromises have to be made.

In the design of tunnel diodes, it is necessary to know the relationship between the three design parameters just mentioned and the circuit and characteristic-curve parameters. The p-n junction capacitance, $C$, is probably the simplest parameter to control. It is directly proportional to the junction area and also to the square root of the impurity density in the more lightly doped side of the junction. However, because the n-type region is the more lightly doped and the impurity density in this region is approximately constant, the junction-capacitance factor can be neglected. The capacitance is equal to about 2.2 $\mu f/\text{cm}^2$ for the "standard" n-type layer to be described later.

The negative resistance $R$ is (to a first approximation) inversely proportional to the peak current $I_p$ and, thus, is not an independent variable. The series resistance $r$ is composed of two components, one inversely proportional to the junction diameter and the other strictly due to the ohmic resistance of the diode, the package, and the various soldered connections.

The peak and valley voltages $E_p$ and $E_v$ depend on the doping levels on the two sides of the p-n junction, and both $E_p$ and $E_v$ increase with increased doping. The only remaining parameters are $I_p$ and the ratio $I_p/I_v$, since $I_v$ itself is rather meaningless. The variations of $I_p$ and $I_p/I_v$ with p-type-doping density are shown in Fig. 2 for a constant, standard, n-type doping.

With these relationships, it is possible to design tunnel diodes to have any consistent set of parameters within the limits set by present technology. The desired values for the various parameters will depend on intended use, in that some parameters must

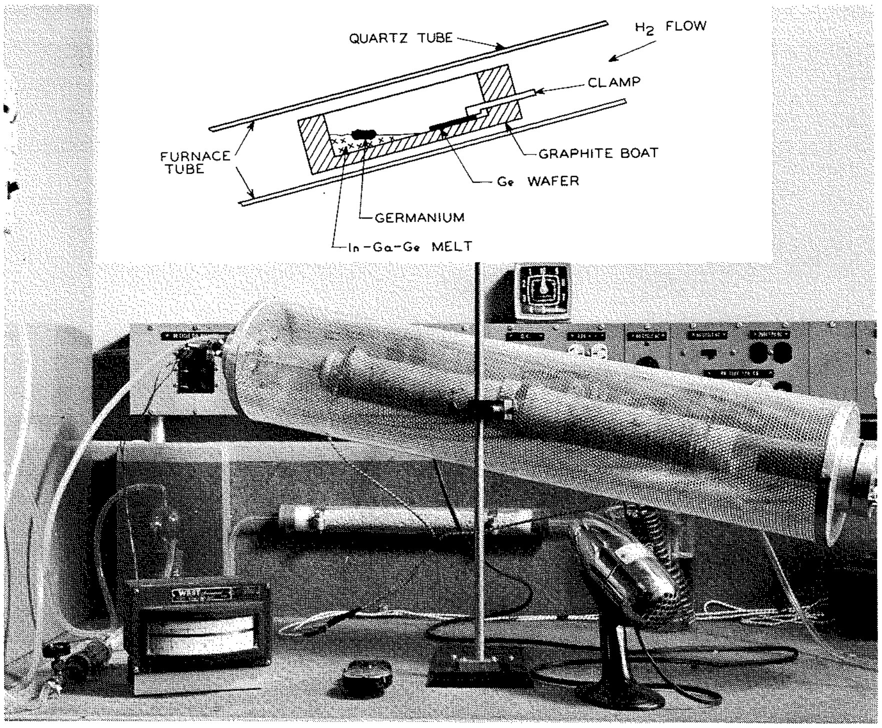


Fig. 3—Furnace used in solution-growth fabrication process.

usually be comprised to obtain desired values for others.

### FABRICATION TECHNIQUES

Several variations of two basic approaches for fabrication have been investigated. One uses the *dot-alloy* process, and the other a new *solution-growth* process for the formation of the p-n junction.

#### Dot-Alloy Method

The first tunnel diodes fabricated at the RCA Laboratories were prepared by the conventional dot-alloy technique. The base pellets were cut from germanium single crystals doped with arsenic to a concentration of about $3 \times 10^{19}$ atoms/cm$^3$. The major surfaces of these pellets were (111) planes. A 3-mil indium dot containing about one percent of gallium by weight was alloyed onto each pellet. Alloying schedules involving different maximum temperatures and heating and cooling rates were investigated. Several good tunnel diodes were made, but results varied greatly from lot to lot even at constant processing conditions. Neither were attempts to fabricate units having low negative resistance successful. Consequently, the conventional dot-alloy technique was discontinued when a new solution-growth process showed greater promise.

#### Solution-Growth Approach

The new solution-growth process provides a large-area semiconductor wafer having a surface layer of opposite type and permits the fabrication of numerous devices having precisely predetermined geometry and electrical characteristics. When this approach is employed in the fabrication of tunnel diodes, a wafer of degenerate n-type germanium ($n_d > 1 \times 10^{19}$ atoms/cm$^3$) cut perpendicular to the (111) axis is used as the starting material. This wafer is lapped until its major surfaces are parallel planes and until it is about 12 mils thick. A highly doped p-type germanium layer is then grown onto one of its major surfaces in a furnace (Fig. 3).

As shown in Fig. 3, the germanium wafer is held tightly against the flat bottom of a graphite boat by the graphite clamping arrangement. The graphite boat is fixed in position at the center of a constant-temperature zone of the quartz furnace tube. With the furnace tube tipped and with a flow of hydrogen through the tube, the graphite boat is brought to a temperature of about 500°C. It is maintained at this temperature until the mixture of indium, gallium, and germanium at the lower end of the graphite boat is brought into solution equilibrium, with excess solid germanium floating on the surface of the melt. The furnace tube is then tipped back so that the melt covers the exposed surface of the germanium wafer. The furnace temperature is raised a few degrees to cause further dissolution of the germanium floating on the melt, as well as on the surface of the wafer.

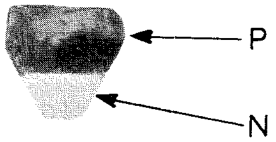
The furnace is then allowed to cool at a rate of about 10°C per minute until a temperature of about 400°C is reached. Then, during this cooling period, germanium precipitated from the melt grows onto the exposed surface of the wafer. At 400°C the furnace tube is tipped to its original position, and the melt is decanted from the wafer surface. The graphite boat is removed from the surface tube at 290°C, and the remaining melt is wiped from the surface of the wafer before it is removed from the boat.

Fig. 4 shows a germanium wafer processed by the solution-growth technique. A portion of the new-growth surface along the edge has been taper-lapped at an angle of 3 degrees with respect to the surface plane and stained to expose the p-type region at the surface and a portion of n-type base material. Measurements on this specimen show that the p-n junction is a flat surface about 12 microns below the original wafer surface, indicating that "old" germanium is dissolved from the exposed surface of the wafer before new growth begins. The new growth has proceeded to a level about 20 microns above the original wafer surface.

This new growth is a single-crystal extension of the original single-crystal body of the base wafer and is heavily doped with gallium and indium. The doping, or acceptor concentration, can be controlled through a wide range by variations in the ratio of gallium to indium in the melt. At 450°C, the saturation solubility[1] of indium in germanium is $1.7 \times 10^{18}$ atoms/cm$^3$; that of gallium in germanium is $4 \times 10^{20}$ atoms/cm$^3$. It is possible, therefore, to provide for acceptor concentrations from about $1.8 \times 10^{18}$ to $1 \times 10^{20}$ atoms/cm$^3$ by varying the percentage of gallium in the melt. In the example of solution growth described previously, the charge contained one percent by weight of gallium. Chemical and electrical determinations have shown an acceptor concentration of approximately $8 \times 10^{19}$ atoms/cm$^3$ in surface layers grown from this melt.

Tunnel diodes can be prepared from solution-grown germanium wafers by use of a masking and etching technique or by a lapping technique.[2] In both cases, a portion of the p-type surface layer is removed from the wafer to leave an array of tiny p-type mesas which rise

Fig. 4—Germanium base wafer processed by solution-growth technique. P: regrown p-type surface layer. N: n-type base wafer region.



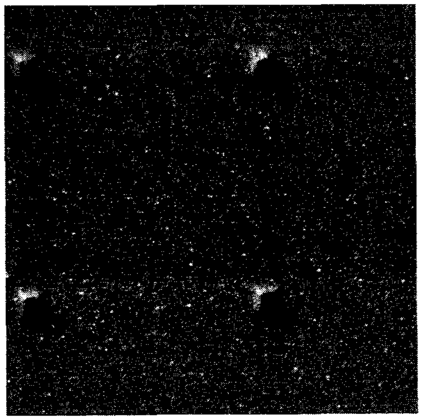


Fig. 5—Photomicrograph of portion of tunnel diode array.

from an n-type plain. A photomicrograph of a portion of one of these arrays is shown in Fig. 5. The wafer is cut into pellets, each having a mesa at the center, and the units are then mounted and encapsulated.

The forward-bias characteristics of tunnel diodes prepared by the etching or lapping technique are shown in Fig. 6. Curve A is typical of the results obtained when donor concentration in the base wafer is approximately 2.2 × $10^{19}$ atoms/cm$^3$ and the acceptor concentration in the p-region is approximately 8 × $10^{19}$ atoms/cm$^3$. Curve B shows typical results when these concentrations are approximately 2.2 × $10^{19}$ and 3.0 × $10^{20}$ atoms /cm$^3$, respectively. Curve C is an "envelope" of the oscillograph traces for 10 tunnel diodes of a 25-unit array. The forward-bias characteristics of these units are nearly identical, indicating the great uniformity from unit to unit through the solution-growth approach.

A comparison of Curves A and B of Fig. 6 shows that, for constant n-region doping, the negative resistance of tunnel diodes decreases with increased doping of the p-region; however, the ratio of maximum to minimum tunnel current decreases also. Thus, attempts to push the tunnel diode toward higher and higher frequency performance (by use of germanium doped to the highest possible concentration of arsenic, and by use of high acceptor doping) led to prohibitively low ratios of $I_p$ to $I_v$.

An investigation of the factors influencing this ratio led to changes in the processing procedures which resulted in a spectacular increase of the ratio of $I_p$ to $I_v$ at high tunnel-current densities. These new procedures are identical to the original solution-growth approach, except that p-type instead of n-type germanium is used for the base-wafer, and the surface layer is grown at about 410°C from a melt composed of lead, tin, germanium, and arsenic. The new procedure apparently permits higher arsenic concentrations in the n-region of the tunnel diode and more nearly equal concentrations of acceptors and donors at the high doping levels. This theory at least partially accounts for the improvement observed. The curves in Fig. 7 show the forward-bias characteristic of high-speed tunnel diodes fabricated by the original and by the improved solution-growth approach.

### FACTORS AFFECTING CHARACTERISTICS

Although a complete discussion of the factors affecting tunnel-diode characteristics is beyond the scope of this article, some of the more important results of several studies can be described. The curves in Fig. 2, for example, show data obtained to determine the peak tunnel-current density and the ratio of $I_p$ to $I_v$ as a function of acceptor concentration in the base wafer when the n-type surface is grown from a melt composed of (by weight) 34 percent lead, 58 percent tin, 6 percent germanium, and 2 percent arsenic. Fig. 8 shows the acceptor concentration in a solution-grown germanium surface layer as a function of the gallium concentration in the indium melt from which the layer is grown. Table I shows that exposure to the heating cycle of the alloy process leads to a decrease in the carrier concentration $n_d$ of degenerate arsenic-doped base wafers.

**TABLE I—EFFECT OF HEATING CYCLE ON CRYSTAL T2065**

| Characteristic | Seed End | | Tail End | |
|---|---|---|---|---|
| | Before Heating | After Heating | Before Heating | After Heating |
| $R_H$, cm$^3$/coul | 0.191 | 0.202 | 0.136 | 0.148 |
| $n_d$, cm$^{-3}$ | 3.3 × $10^{19}$ | 3.1 × $10^{19}$ | 4.6 × $10^{19}$ | 3.4 × $10^{19}$ |
| $s$, ohm-cm | 7.59 × $10^{-4}$ | 7.91 × $10^{-4}$ | 6.15 × $10^{-4}$ | 7.20 × $10^{-4}$ |
| $\mu_H$, cm$^3$/v-sec | 252 | 282 | 216 | 280 |

$R_H$ = Hall coefficient; $n_d$ = donor concentration; $s$ = resistivity; $\mu_H$ = Hall mobility.

### PACKAGING

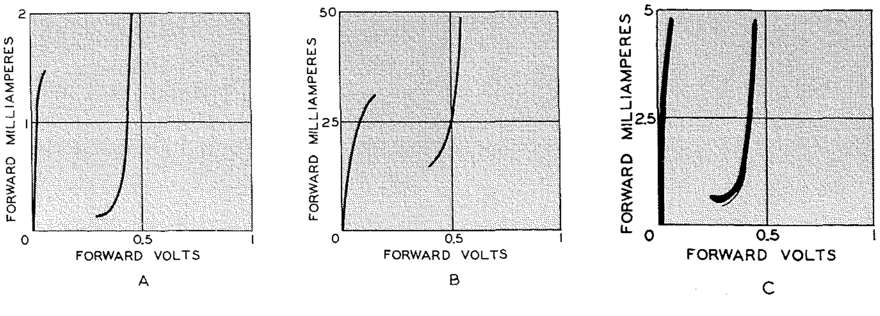
Although tunnel-diode junctions usually must be very small to provide low capacitance, the package must have large-area contacts to provide low inductance. These two requirements immediately eliminate a large number of possible mounting and packaging schemes, including all the currently common methods for transistors and conventional diodes. The package shown in Fig. 9 was chosen for the tunnel diode as offering a good compromise; the inductance is about 5 to 6 × $10^{-10}$ henry. This inductance is suitable for very-high-speed switching circuits, but is a serious limitation to the use of tunnel diodes as high-frequency amplifiers and oscillators.

The tunnel diode package consists of a ceramic ring, which holds the diode itself, brazed between two metal tabs. The germanium pellet containing the diode is soldered into the ceramic ring, and connections are soldered on, as shown in Fig. 9. The pedestal shown is optional; the germanium can be soldered directly to the lower tab to reduce the thickness of the package.

Because there are a number of possible methods for fabricating tunnel diodes, there are also variations in the packaging procedure. For the TD-100 to TD-111 series, the process was made as simple and straightforward as possible at the expense of a slight decrease in yield (i.e., some diodes have peak currents $I_p$ which fall outside the desired range).

The TD-100 to TD-111 series are made in wafer form on germanium wafers containing solution-grown layers. An array of diodes is produced when a pattern of 3-mil-diameter lead dots is evaporated onto a 30-mil-square grid. The solution-grown layer is then etched away everywhere except under the lead dots. A drop of epoxy resin is placed on the side of the wafer containing the mesas, and the wafer is clamped between two flat Teflon plates. When the resin hardens, the wafer is removed and

Fig. 6—Forward-bias characteristics of germanium tunnel diodes.



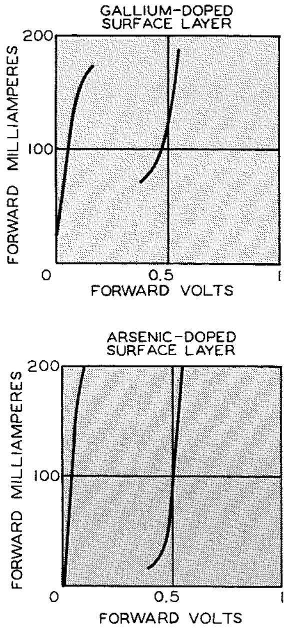
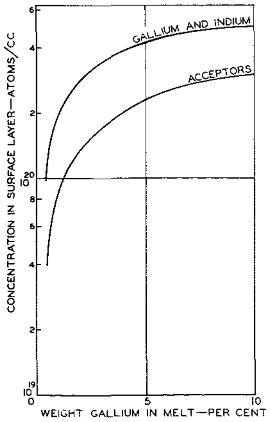


Fig. 7—Forward-bias characteristics of high-speed germanium tunnel diodes.

lapped very lightly to remove any resin from the mesa tops. Large-area copper contacts are evaporated over each mesa and on the entire back surface of the germanium wafer. The wafer is scribed with a diamond point and broken into individual germanium pellets 30 mils square, each containing a tunnel diode. The diode is then put into the package and soldered in, with the epoxy resin acting as an insulator to prevent shorting of the p-n junction. Excess liquid flux is removed and a circular cap is welded onto the package. The outstanding feature of this process is that small-area tunnel diodes are made without the use of any small, separate device parts and without the use of any precision assembly equipment. The one disadvantage to the process is that the value of the peak current $I_p$ follows a random distribution, and thus it is impossible to make large numbers of diodes having identical values of $I_p$.

For the manufacture of tunnel diodes having accurately predetermined values of peak current, it is necessary to process each diode individually. Electrolytic etching provides the most accurate control process because the etching that takes place is directly proportional to the current passed through the circuit. Furthermore, it is possible to monitor the peak current of the diode while it is being etched and to stop the etching at the desired value of $I_p$. The actual etching process used is almost identical to that used for the gallium arsenide units described in the paper by A. J. Wheeler in this issue.

Fig. 8—Acceptor concentration in solution-growth surface layer as a function of gallium concentration in the melt.



Tunnel diodes which are to be electrolytically etched must be mounted in the package first so that the electrical contacts necessary for etching are established, and so that the diode need not be disturbed once it is etched to the proper value of peak current. There are two methods, basically the same except that a solution-grown layer is used in one case. If a solution-grown layer is used, the germanium is cut into squares and soldered into the package. A small dot of suitable material is soldered to the germanium to serve as the contact to the grown layer. If the solution-grown layer is not used, the dot forms the p-n junction. This dot is placed on the germanium before the pellet is soldered into the package. A contact is soldered to the dot in either case, and the unit is etched electrolytically. At the present time, the grown-layer process seems more promising, mainly because the problems of junction formation and contacting can be treated as two separate processes and each can be individually controlled more easily.

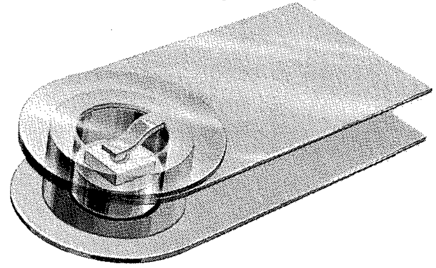


Fig. 9—Low-inductance package for tunnel diodes.

## STATUS

The best performance obtained to date with germanium tunnel diodes includes (1) a gain-bandwidth product of about 16,000 mc, (2) a total capacitance of less than 1 $\mu\mu$f, and (3) a current-to-capacitance ratio of greater than 10 ma/$\mu\mu$f. Capacitance values of 3 to 5 $\mu\mu$f can be obtained reproducibly and values less than 3 $\mu\mu$f obtained with some difficulty.

The best germanium tunnel diodes made to date can be used in microwave amplifiers and oscillators, and in switching circuits having switching times of less than 1 nanosecond.

**HERBERT NELSON** received his BS from Hamline University in 1927 and MS in Physics from the University of Minnesota in 1929. From 1929 to 1930 he was with the Westinghouse Lamp Company in Bloomfield, N. J. He transferred to RCA in 1930, engaging in research and development in thermionic emission, secondary electron emission and solid-state devices at the RCA Manufacturing Company at Harrison, N. J. until 1953. Since then, he has been a member of the technical staff of RCA Laboratories in Princeton, doing research on silicon, solid-state devices, and germanium tunnel diodes. He has published several papers and holds numerous patents. He has twice received an RCA Incentive Award for outstanding work in research, and is a member of Sigma Xi and the American Physical Society.

## ACKNOWLEDGMENT

The authors wish to thank their many co-workers both at Princeton and Somerville for their support and cooperation. Particular thanks are due to Sigurd Bragnum, Eugene Jordan, and Arthur Herd for gathering and tabulating much of the data reported.

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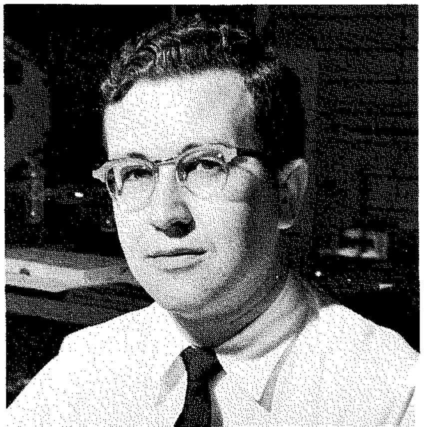
**NORMAN H. DITRICK** received his BS and MS degrees in Physics from Ohio State University in 1952. He joined RCA as a specialized trainee in June of 1952 and was later assigned to the Receiving-Tube Advanced Development Activity of the Electron Tube Division. He was transferred to the Semiconductor Advanced Devices Development Activity in April of 1953, and since that time has been working principally on high-frequency transistor development and recently on tunnel-diode development. Mr. Ditrick is a member of Sigma Pi Sigma, Tau Beta Pi, and the IRE.



# TELEVISION IMAGERY SIMULATOR

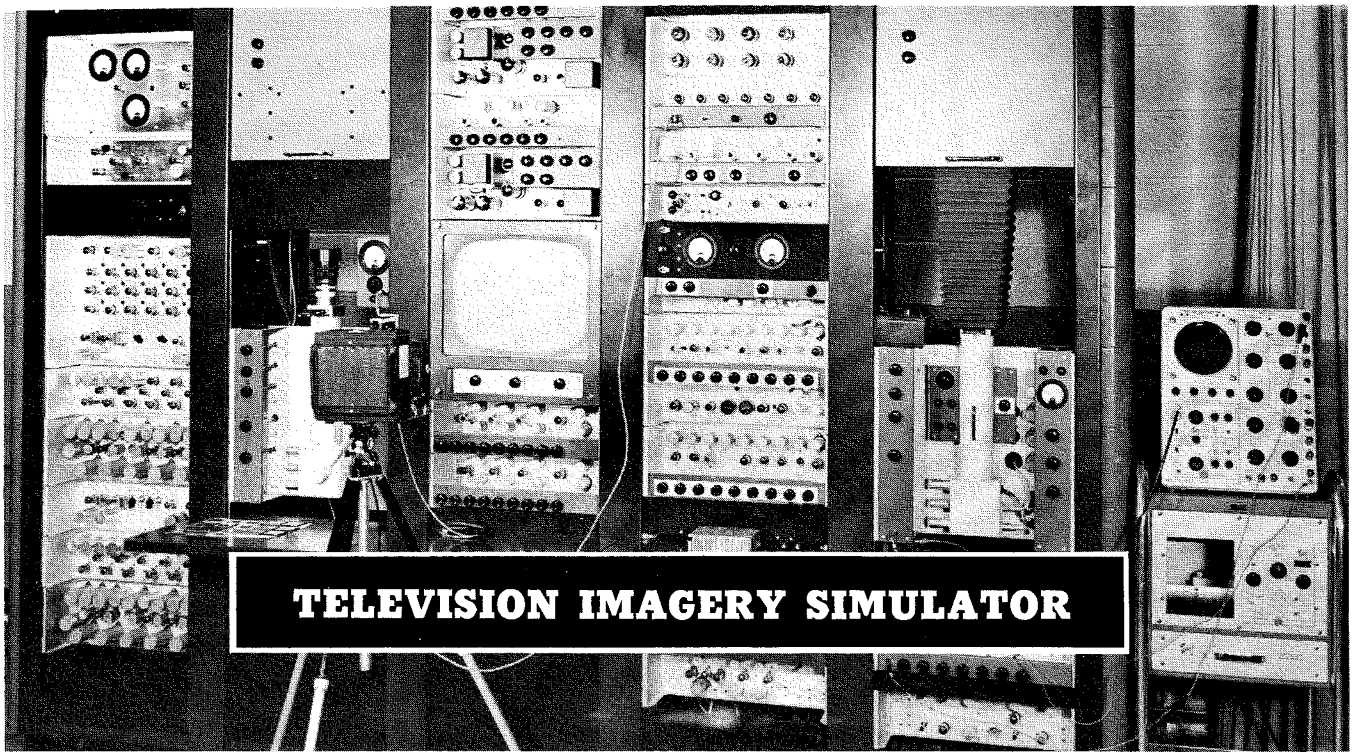


Fig. 1—TV Imagery Simulator.

*by* **J. F. BAUMUNK, R. L. HALLOWS,* and J. P. SMITH**
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*DEP, Princeton, New Jersey*

For evaluating TV systems and studying image-enhancement techniques, simulation equipment with adjustable and measurable parameters is a valuable tool. Such an equipment, the **TV Imagery Simulator,** has been built and utilized by AED. It was useful in design of the Tiros I satellite and is adaptable for studying other TV problems.

The Television Imagery Simulator is a tool for setting up and observing the effects of given TV parameters. It consists of a high-quality, high-resolution black-and-white TV chain, with input pictures in the form of 35-mm transparencies and the resulting TV image presented on a TV monitor. This imagery, which may consist of photographs of ground objects, clouds, sky objects, or of pictures taken with infrared light, ultraviolet light, or radar, may be observed directly or photographed and studied in a remote location. The equipment is shown in Fig. 1.

This simulator was useful in the early stages of the design of the Tiros I satellite in determining the area of cloud cover to be scanned and in choosing the shutter speed of its TV cameras. Before the Tiros launch, it supplied signals having the proper parameters for testing the ground-station equipment. Pictures from the orbiting Tiros I can be studied on the simulator to see how enhancement techniques may help meteorological interpretation.

## DESIGN FEATURES

Because results must be duplicated from day to day and from month to month, reliability of the equipment is of particular importance. When evaluating systems of high resolution, the resolution capabilities of the evaluating means must be better than the system under evaluation. Finally, the mechanical and electrical design must be flexible so that when new techniques present themselves, they may be developed and added to the existing equipment.

The simulator can vary the parameters in a very wide range. Thus, referring to the sync generator, (Block 1, Fig. 2), it is possible to vary the horizontal sync pulses from 30,720 cps to 240 cps in binary steps and the vertical sync pulses from 60 per second to 1 in 4.266 seconds, also in binary steps. Noninterlaced frequencies are generated, but interlacing is accomplished by adding a square-wave current of half the vertical scanning rate to both the flying-spot scanner and monitor yokes in phase (Blocks 2 and 4). The monitor may be interlaced independently of the flying-spot scanner to minimize line structure without changing resolution.

### Scanning Rates

Vertical scanning rates of 60, 30, 15 and 7.5 per second may be selected by switches, and horizontal scanning rates may be changed by other switches that also maintain the picture geometry. The number of scanning lines resulting, which may be selected by switches, are 1024, 512, 256, and 128. Subtracting the lines lost in the vertical blanking interval, the useful line numbers are 960, 480, 240, and 120. Line numbers between these values may be obtained by over-scanning the 35-mm transparency uniformly in both directions or by optical reduction of the image on the transparency.

If more than 1024 scanning lines should be required, the vertical rate may be reduced. Thus, if a vertical rate of 30 per second, interlaced, is used (the horizontal rate remaining the same) the number of lines is doubled to 2048. If interlacing is not desired, a

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* Now with RCA Laboratories.

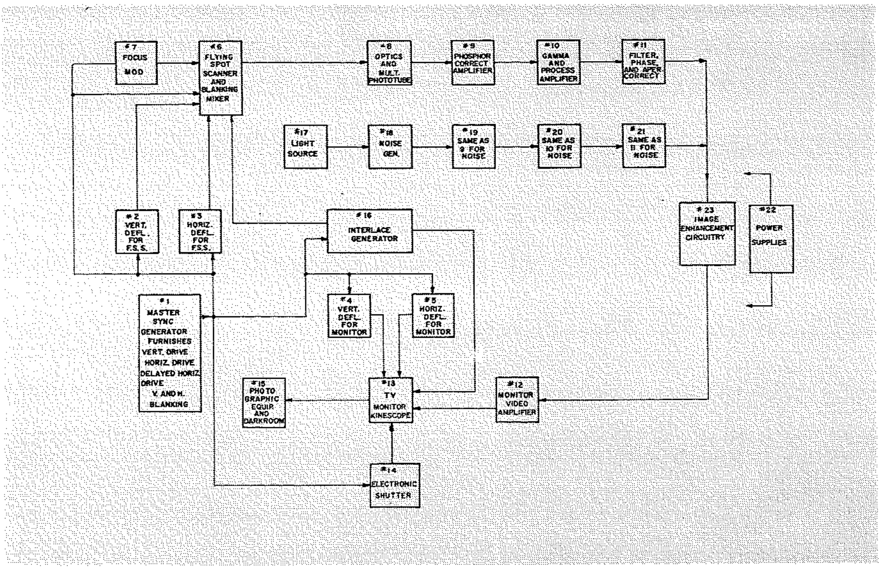


Fig. 2—Block diagram of the TV Imagery Simulator.

vertical rate of 15 may be used, making available 2048 straight scanned lines for this horizontal rate. If more horizontal resolution is required, the horizontal rate may be reduced.

The resolution limit depends on the spot and phosphor characteristics of the kinescope used in the flying spot scanner. Kinescopes of the type used in the simulator have a TV limiting resolution of 1300 lines in the center, with an ultor voltage of 30 kv and ultor current of 25 microamperes. The monitor kinescope is not a limiting factor in the system.

As the number of the scanning lines is changed, the horizontal resolution may be made equal to the vertical resolution by plugging in the proper filter and phase corrector (Block 11). Without the band-limiting filters, the overall system is flat to 18 mc and down 6 db at 25 mc. This bandwidth and an aspect ratio of 3 by 4 gives a horizontal limiting resolution of 740 TV lines for the highest horizontal rate generated, 30,720 scans/sec.

### Noise Generation

As may also be seen in Fig. 2, a separate video channel is used for noise generation and mixing with the video signals. "Flat" or "peaked" noise may be mixed with the video signals and the ratio measured. Signal-to-noise ratio is considered here as the ratio of half the peak-to-peak video signal to the root-mean-square value of the noise, measured separately. The signal-to-noise ratio in the photographic print is equal to this ratio divided by the gamma of the system between the point of noise insertion and the print. It is dependent, also, upon the number of frames exposed photographically, improving as the square root of the increased-exposure factor. Thus, if the exposure time is increased by a factor of 2, the signal-to-noise ratio will improve by $\sqrt{2}$, or 3 db.

An electronic shutter is used (Block 14) for controlling the signal-to-noise ratio in the print and for ensuring that one complete field or frame, or a whole number of fields or frames, is exposed on the film. Binary counters are triggered from the vertical drive pulses to blank the monitor kinescope except for 1, 2, 4, or 8 vertical scans. The number of scans may be preselected and the camera-shutter contacts used to actuate the electronic shutter.

### Complementary Equipment

Photographic equipment (Block 15) is at present a Burke and James 4 x 5 Speed Press camera with a Wollensak 210-mm, f/4.5 lens in an Alphax shutter. For making 35-mm slides and for close-up work, an Exacta VX with a 50-mm Schneider Xenon, f/1.9 lens is used.

The remainder of the blocks in Fig. 2, with the exception of Block 23, will not be described in detail, as they contain circuitry common to the television art. Block 23, covering the image-enhancement circuitry, will be described later.

In Fig. 1, the rack in the simulator with the extended bellows over the lens to keep out stray light is a picture generator for low-speed scanning rates. It also uses a flying spot scanner with 35-mm slides as input information. The scanning rates brought out on switches are: vertical, 0.9375 and 0.46875 frames per second; horizontal, 480 and 240 per second.

The bandwidth of the video amplifiers may also be controlled with filters. The numbers of scanning lines available on switches are 256, 512, and 1024. No monitor is provided except for the Tektronix 545 oscilloscope, because the signals are used for testing other equipment, where pictures may be viewed.

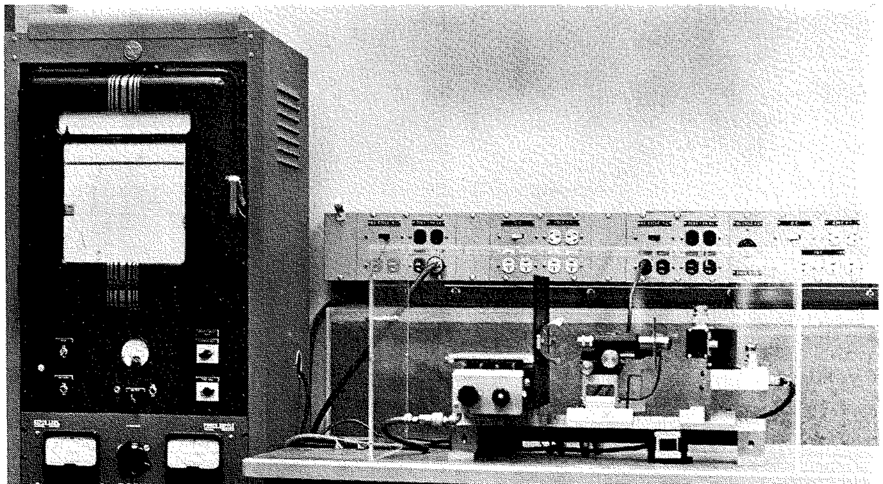
The measurement of the transmission characteristics of a complete TV chain is a fairly complex process. To supplement the simulator and to perform this testing and calibrating function, a microphotometer is needed. With this facility, transparencies or opaques taken from the TV monitor may be scanned and plotted on a recorder. By using a number of test slides, the microphotometer and its associated equipment (Fig. 3) will provide information for calculating the effective line number, or $N_e$, of the complete system. It will also measure the signal-to-noise ratio in a print or transparency as a check on the simulator calibration. With this facility, the user may more completely specify the parameters of the TV system being simulated.

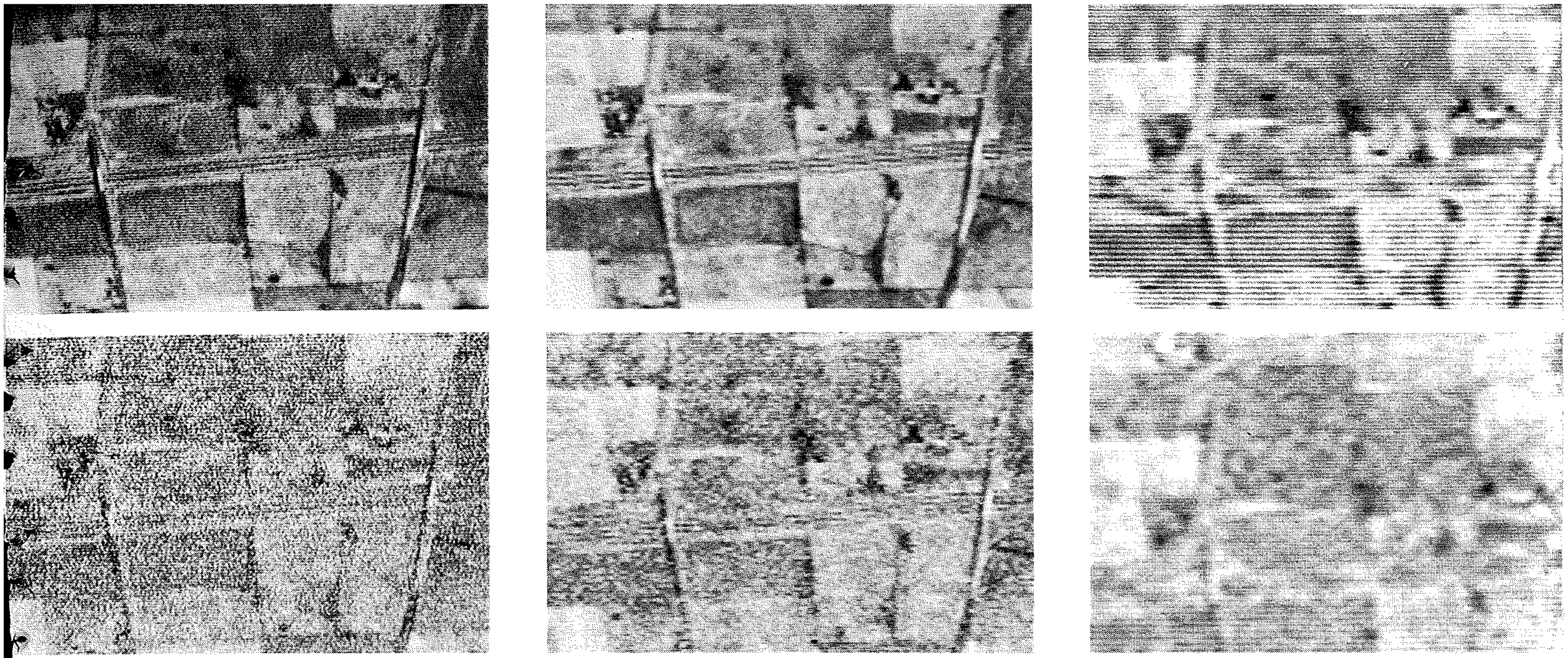
## APPLICATION TO SYSTEMS EVALUATION

The most important use of the simulator up to the present time has been in TV systems evaluation studies, although interest in the image-enhancement field is increasing. In some cases, the two are inter-related. For example, it may be necessary to know how much noise can be tolerated before enhancement; or, if enhancement is used, what maximum bandwidth is tolerable before noise becomes objectionable. These questions have a bearing on the feasibility of adding enhancement circuitry in a satellite before radio transmission. Herein, however, these factors will be considered separately, insofar as is possible.

Imagery prepared by the simulator has been used in several experiments[1]

Fig. 3—Microphotometer equipment.





a) Top: 960 useful lines, 18-mc bandwidth, 27-db s:n ratio. Resolution 13.75 ft/line in both directions.

b) Bottom: Same as 4a, except 15-db s:n ratio.

c) Top: 480 useful lines, 30-db s:n ratio, resolution 27.5 ft/line in both directions. Band-limiting filter and phase corrector added to maintain equal vertical and horizontal resolution, and no noise was added.

d) Bottom: Same as 4c, except 15-db s:n ratio.

e) Top: 240 useful lines, 33-db s:n ratio, resolution 55 ft/line. As in 4c and 4d, band-limiting filter and phase corrector added to maintain equal vertical and horizontal resolutions, and no noise was added.

f) Bottom: Same as 4e, except 15-db s:n ratio.

Fig. 4—Aerial photograph of a rural area presented on the simulator, with TV parameters varied as noted. NOTE: The printing process (halftoning) necessary to present these here has made some of their varying characteristics less distinguishable than they actually are on the simulator. Attention should be directed to the strip through their center (a turnpike) for a comparison of the effects of the varied parameters.

in which experienced photo-interpreters extracted meaningful information under controlled conditions. The evaluations of the various imagery were then based on extracted information. It is interesting to note that the quality and quantity of information extracted by the experienced photo-interpreters is often greater than the layman might think possible from the appearance of the imagery.

Fig. 4 is an aerial photograph of a rural area with varied resolutions and signal-to-noise ratios introduced by the simulator. Many pictures of a variety of subjects with line numbers ranging from 60 to 960 have been utilized in the simulator, but Fig. 4 is sufficient to illustrate its flexibility. For example, a series of cloud pictures, covering an area 700 x 700 miles (the area covered by the Tiros wide-angle camera) have been observed with the simulator adjusted for parameters similar to Fig. 4.

## IMAGE-ENHANCEMENT CAPABILITIES

In addition to the flexibility of standard TV parameter control, facilities for special treatment of the video signal have been provided in the simulator for evaluation. These effects, or *image-enhancing* techniques, are of value in performing the specialized function of image interpretation and information extraction from the standpoint of emphasizing certain features of the pictorial input while de-emphasizing others that may not be significant.

### Slicing

The effect of contrast stretch[1], or *video slicing*, is one of providing increased contrast in a desired portion of the gray-scale transfer characteristic. It is an extension of the philosophy of gamma correction, which provides greater gain in a certain portion of the video signal, by controlled nonlinear amplification. The contrast in that region of the characteristic is thus increased, rendering certain details more vividly. *Gamma* may be defined as the ratio of the log of the subject

a) Original photo.

b) 5-to-1 black stretch.

c) 5-to-1 white stretch.

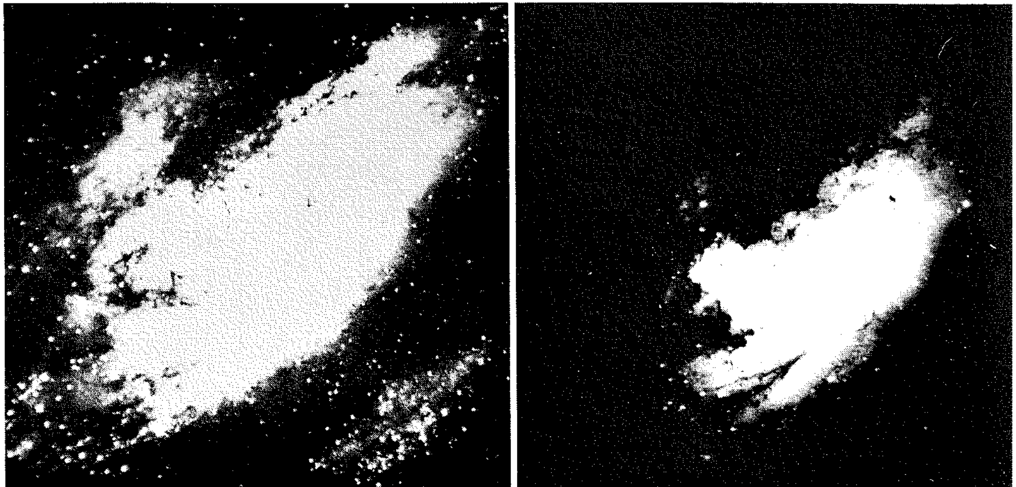
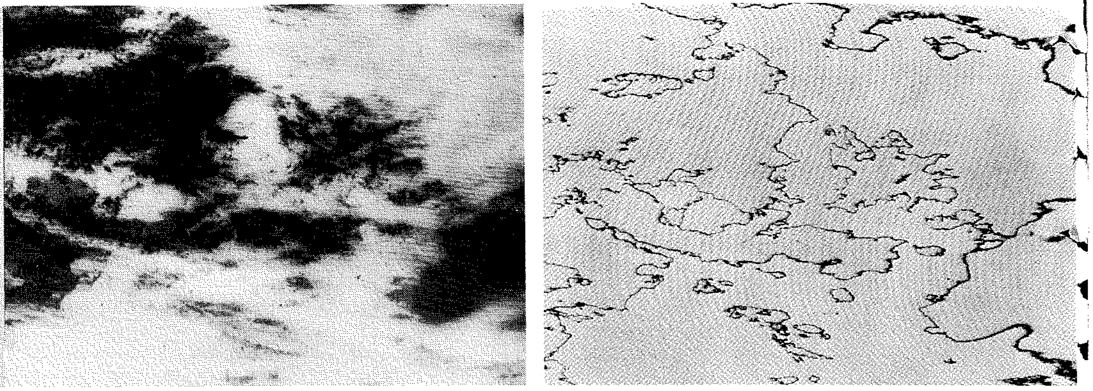


Fig. 5—Contrast stretch, or "slicing", illustrated by photo of nebula. Note contrasts in lower right of each picture; i.e., trade-off of white intensity for detail, and vice-versa.

Fig. 6—Outlining applied to photo of cloud formations. The outlining effect is very useful in delineating such nebulous objects, as these pictures illustrate.



a) Original photo, 960 lines and 27-db s:n ratio, (no outlining).

b) Derived outline signal.

incremental brightness to the log of the corresponding incremental image brightness. It is equal, in a general imaging system, to the slope of the transfer characteristic plotted logarithmically.

The concept of gamma correction is usually limited to the compensation for the nonlinear drive characteristic of the viewing kinescope. Slicing allows the gamma of the already-corrected signal to be increased about any desired gray level from black to white to emphasize the difference between nearly equal gray areas, within the limits of usability imposed by granularity of the input picture and electrical system noise.

The slicer circuitry is essentially a double-clipper which passes only a small portion (about 10 percent) of its maximum signal input. The signal portion lying between the clipping levels is controlled by a variably-biased clamp, thus providing for "slice-position" adjustment. The slice "thickness" is controlled by adjustment of the input signal amplitude to the double-clipper, so that a thinner slice of the total video signal is passed when the input amplitude is increased.

Fig. 5a to 5c represent an intermediate slice thickness of approximately one-fifth the peak-to-peak input level. They have been reproduced with none of the original signal super-imposed, and at full contrast.

#### Outlining

A circuit has been developed for producing an outline, or contour of constant intensity and of either polarity, along the loci of a selected video gray-level. This circuit has the ability to convert into sharp pulses of fixed amplitude, variations, or edge transitions lying between the pass limits of the slicer circuitry. The gray-level to be outlined is therefore set by the slicer position and thickness control.

In some portions of an outlined image, the subject of which contains only two gray-levels, outlining occurs along a contour which is precisely parallel to the scanning lines of the raster. This effect occurs when the input signal level exists between the two levels of the slicer output. The circuit is activated by the first excursion into the region of slice and it will remain on until an excursion out of the slice occurs in either direction. The excellent speed of response to momentary transitions through the slice region is evidenced by the narrowness of the outline spikes produced when a steep vertically-oriented transition is scanned.

The outlining effect (illustrated in Fig. 6) has found its greatest usefulness in delineating nebulous objects, such as cloud formations. It is also expected to find use in quick evaluation of relative-intensity distributions in optical or astronomical "star images" when used with rough calibration techniques.

If video slicing is used and presented with full amplitude in the absence of normal video, a condition approaching two-level quantization is obtained. This condition, in superposition with the outliner just described, may serve as a means of making shaded outline map effects which may be quickly recorded photographically, as seen in Fig. 6e.

#### Differentiation

The interesting bas-relief rendition of the cloud streets of Figs. 7a and 7b is achieved by the simple expedient of providing a short RC time-constant of differentiation as a coupling network in the video amplifier. This provides practically no response at low frequencies and increasing response to the higher frequencies, which correspond to steeper transitions.

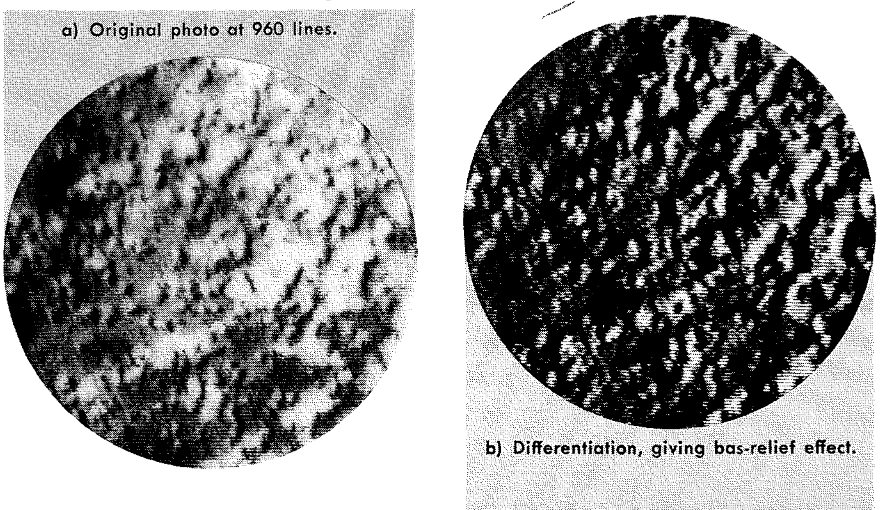
This technique is akin to the well-known television technique of aperture compensation, in which the normally flat video amplitude response is peaked, or made to rise from beyond the mid-frequency range, but with a linear phase characteristic. The nonlinear phase function of a simple differentiation network causes the three-dimensional light-and-shadow effect.

### CONCLUSIONS

The TV Imagery Simulator has proven to be a flexible research tool for evaluating TV systems and for studying the effects of image enhancement. Parameters may be varied to simulate most existing TV systems, and signal-to-noise ratio may be adjusted and measured.

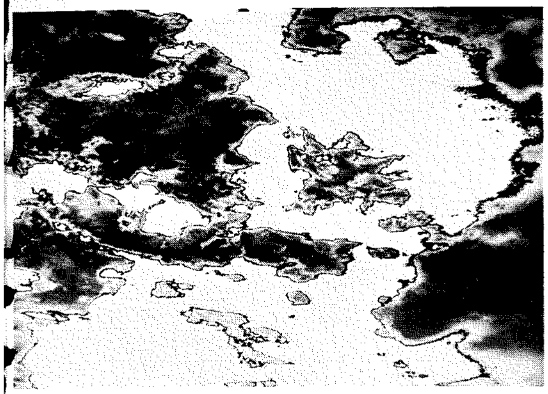
Contrast stretch, or slicing, a form of visual pre-emphasis, is an enhancement technique useful in object recognition when used before noise is encountered. Constant brightness contouring, or outlining, is useful in showing up lines or

Fig. 7—Cumulus cloud streets.

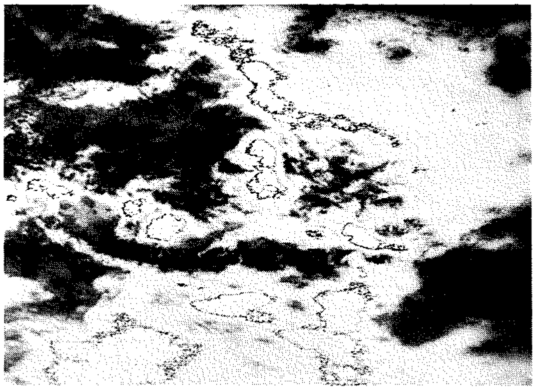


a) Original photo at 960 lines.

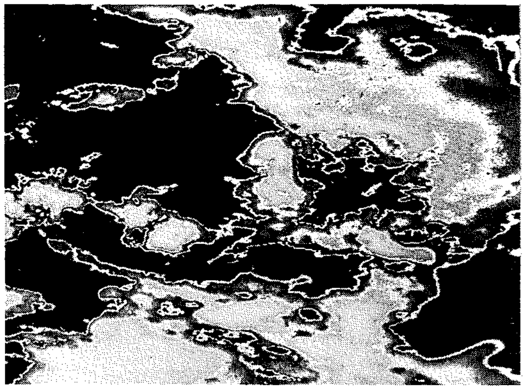
b) Differentiation, giving bas-relief effect.



c) Video superposition of original photo and derived brightness contour.



d) Contouring further into the white region of the cloud.



e) Two-level quantization and outlining. Video slicing presented at full amplitude in absence of normal video, in superposition with outline signal.

areas of equal brightness quickly. Differentiation may be useful in synoptic observations.

Because of its mechanical and electrical design, the simulator may be adapted for studying other television problems.

#### ACKNOWLEDGMENTS

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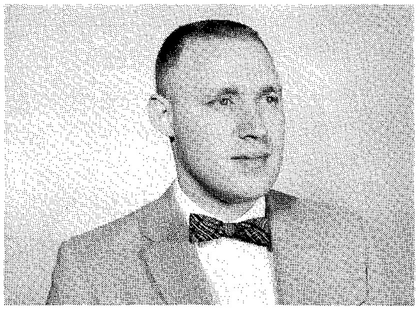

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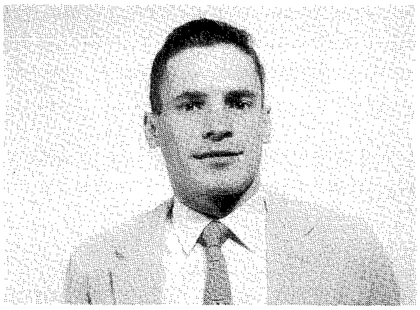


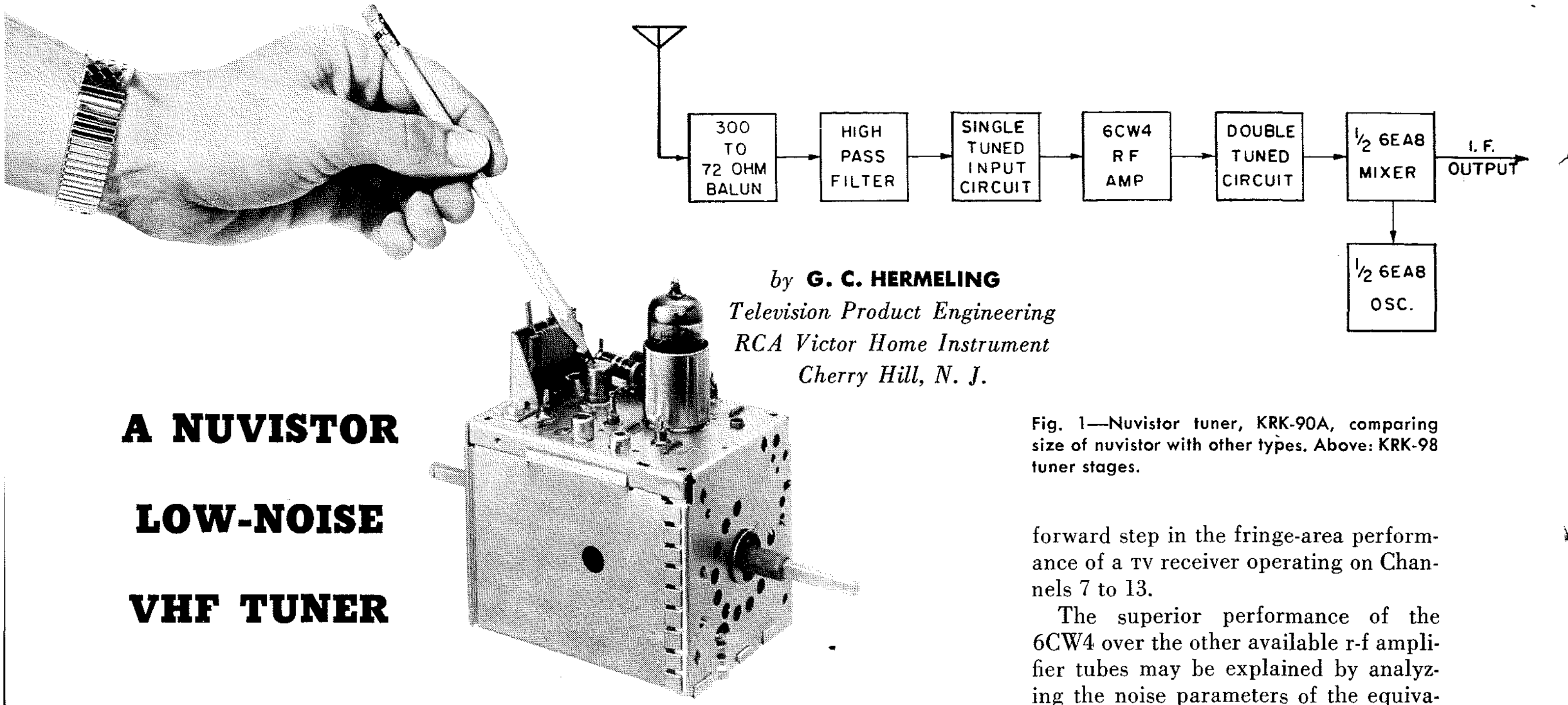
R. L. Hallows

J. P. Smith

**JOHN F. BAUMUNK** received his BSEE from Kansas State University in 1955. He has since taken advanced courses in electrical engineering and solid-state physics at both the University of Maryland and the Department of Agriculture, Washington, D.C. Mr. Baumunk served as a First Lieutenant in the U.S. Air Force from 1955 to 1957. During this time, he was project engineer in a transmission system group doing research and development on radio and wire digital transmission systems. He joined the Special Systems and Development Department of RCA at Princeton, N. J., and transferred to the Astro-Electronic Products Division when it was formed. He has been engaged in the design and construction of a television image simulator, the development and design of the ground television equipment for a satellite system, and most recently in satellite communications.

**RAYMOND L. HALLOWS** received the BSEE in 1952 from the Missouri School of Mines. He joined the RCA Specialized Training Program in June 1952 and was assigned to the Advanced Development Group of the Electron Tube Division, working in special studies on television systems. His activities included design and development on optical sine-wave response measuring apparatus and on a high-definition closed-circuit color television chain. Mr. Hallows transferred to the Astro-Electronic Products Division in 1958, where he worked on the development of special television systems and on measuring and specifying parameters affecting the reproduction of images in photographic and video systems. In 1959, he transferred to the RCA laboratories, where he has participated in studies concerning infrared systems and components for the detection and tracking of space targets. He is a member of the Society of Motion Picture and Television Engineers and the IRE.

**JOHN P. SMITH** received his BSEE from Texas A. and M. College in 1927. From 1927 to 1930, he was a test engineer with the General Electric Co., Schenectady, N. Y., working on carrier current and early TV circuits. He joined RCA Victor Co., Camden, N. J. in 1930 as a TV development engineer, where he worked on studio and terminal equipment, specializing in sync generators. In 1942 he joined the RCA Laboratories, Princeton, N. J. Here he did research and development on storage tube applications, high-voltage power supplies, 500 mc transmitters, TV carrier synchronization, wide band oscilloscopes, color-TV camera systems, and lenticular film recording. In 1957 he joined the Special Systems and Development Group in DEP which later became part of the Astro-Electronic Products Division. At AEP he worked on TV system and imagery simulation until 1960. He is presently working on equipment problems in connection with electrostatic TV-tape research. Mr. Smith is a senior member of the IRE and Sigma Xi.



# A NUVISTOR LOW-NOISE VHF TUNER

*by* **G. C. HERMELING**
*Television Product Engineering*
*RCA Victor Home Instrument*
*Cherry Hill, N. J.*

Fig. 1—Nuvistor tuner, KRK-90A, comparing size of nuvistor with other types. Above: KRK-98 tuner stages.

From the standpoint of improved television receiver performance, the RCA, nuvistor[1] electron tube represents a major breakthrough—particularly with respect to features of small size, low noise, low power drain, and greater reliability. The space-saving feature of the nuvistor offers interesting possibilities to the receiver designer who is considering compactness (see Fig. 1). The cylindrical electrode structure[1] permits the critical tube elements to be assembled in a simple fixture, thus making possible a close-spaced electrode assembly with accuracy and uniformity.

### PERFORMANCE TESTS

The application of the new RCA 6CW4 nuvistor triode in r-f amplifier circuits of vhf tuners for monochrome or color TV receivers has produced improvements in noise figures, gain characteristics, and reductions in power drain. For example, frame-grid tube 6ER5, when used as an r-f amplifier, draws 0.18 ampere filament current compared with the 0.13 ampere for the nuvistor. The d-c requirements for the nuvistor are 10 ma at 80 volt, while the 6ER5 requires 10 ma at 200 volts.

Channel 13, first-stage noise figures as a function of source resistance $R_s$ were measured (Fig. 2) for the 6CW4 nuvistor and several other currently used head-end tubes. A grounded-cathode circuit was used, and a 5722 noise diode with means for changing $R_s$ was mounted adjacent to the tube under test to minimize the effects of lead length and stray capacity. All triodes were neutralized by resonating grid-to-plate capacitance at the test frequency and the input-circuit capacitance was tuned for maximum signal. A correction factor of 0.3 db was added to the indicated noise figures to correct for the internal resonance of the 5722 noise diode.[2]

Several nominal nuvistor triodes and conventional tubes were measured; the curves shown in Fig. 2 are typical of the data obtained. The improvement in noise figure (approximately 1.5 db) of the 6CW4 nuvistor over the recent 6ER5 frame-grid triode and the 6FH5 conventional-grid triode is a significant forward step in the fringe-area performance of a TV receiver operating on Channels 7 to 13.

The superior performance of the 6CW4 over the other available r-f amplifier tubes may be explained by analyzing the noise parameters of the equivalent circuit (Fig. 3) for noise in triodes. Tube noise is completely described by the four noise parameters $R_n$, $G_n$, $R_z$, and $C_z$. The equivalent noise resistance $R_n$ accounts for the shot noise actually occurring at the plate, the noise conductance $G_n$ accounts for the uncorrelated induced grid noise due to transit-time loading, and $Y_z = (1/R_z) + jwc_z$ is the well known correlation admittance. Rothe and Dahlke,[3] and van der Ziel[4] have shown that to a good approximation these noise parameters can be found by simple laboratory measurements. The saturated noise diode current required to double the output noise power is plotted as a function of the input-circuit capacitance (with $R_s = \infty$). The terms $R_n$ and $G_n$ can be calculated from this curve; and $1/R_z$ (which is very small) is assumed equal to zero.

Fig. 3 shows a typical graph and simple equations for calculating $R_n$ and $G_n$. For a test frequency of approximately 100 mc, the values of $R_n$ and $G_n$ so obtained can be used to represent the noise performance of triodes over a wide frequency range. It is useful to

Fig. 2—First-stage noise figures versus source resistance.

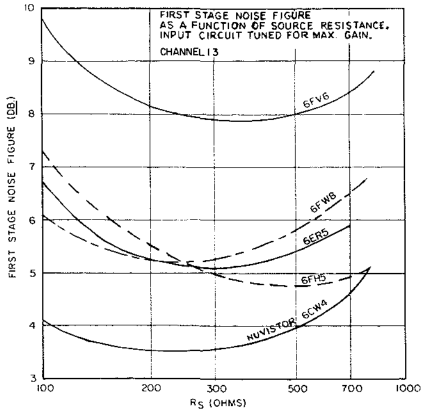
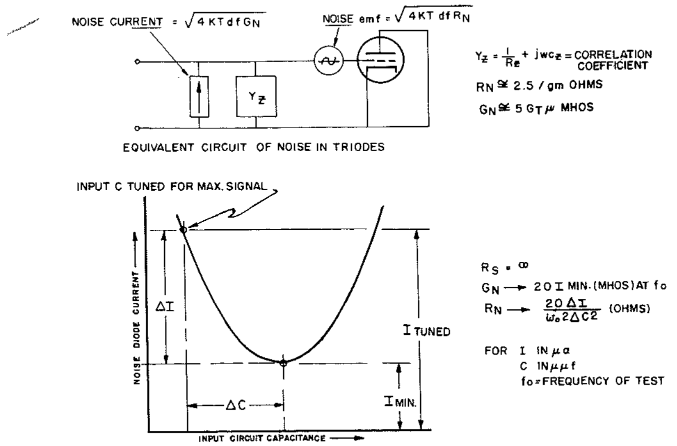


Fig. 3—Noise diode current required to double the output noise power as a function of input capacitance.



plot optimum $R_s$ and minimum $F_1$ as a function of frequency from:

$$R_{s_{opt}} = \frac{f_o}{f}\left(\frac{R_n}{G_n}\right)^{1/2} \qquad (1)$$

$$F_{1_{min}} = 1 + \frac{2f}{f_o}\ (R_n G_n)^{1/2} \qquad (2)$$

Where: $f$ = any frequency, and $f_o$ = frequency of test.

The minimum first-stage noise figure $F_{1_{min}}$ assumes noise matching $(R_{s_{opt}})$, noise tuning $(C_{in} = -C_z)$, and no input-circuit losses.

With a test set-up similar to that used by Metelmann,[5] the graphs of noise diode current as a function of input-circuit capacitance for several nuvistor 6CW4, 6ER5, and 6FH5 tubes were obtained. The nominal low-frequency $G_m$ and the values of $G_n$ and $R_n$ calculated from these curves are as follows:

| Tube | $G_n$ | $R_n$ | $G_m$ |
|---|---|---|---|
| 6CW4 | 330 | 181 | 15000 |
| 6ER5 | 680 | 200 | 10500 |
| 6FH5 | 670 | 235 | 9500 |

Using these values of $G_n$ and $R_n$ and and Equations 1 and 2, $R_{s_{opt}}$ and $F_{1_{min}}$ were plotted as a function of frequency. Fig. 4a shows that the optimum source resistance for the 6CW4 is about 20 percent higher than for the 6ER5 or 6FH5.

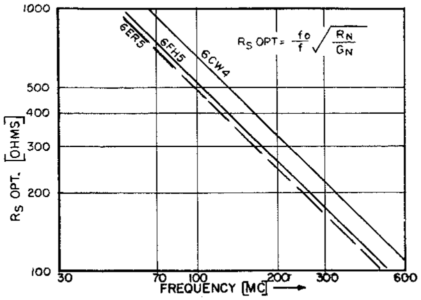


Fig. 4a—Optimum source resistance as a function of frequency.

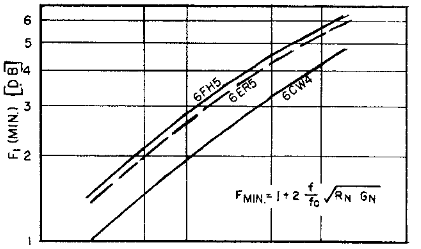


Fig. 4b—First-stage minimum noise figure as a function of frequency.

The 3.25-db noise figure at 200 mc for the 6CW4, as shown in Fig. 4b, is approximately 1.0 db better than the 6ER5 or 6FH5. The exceptionally high $g_m$ of the 6CW4 (15,000 $\mu$mhos) assures more first-stage gain than that obtained with the other two tubes. Therefore, when using the 6CW4, the mixer contributes less noise to the over-all system noise figure.

### CROSS MODULATION

Perhaps the next most important parameter of the r-f tube is its cross-modulation characteristic. It has been found experimentally that 1-percent cross modulation produces interference just perceptible in a TV picture. For that reason, it is convenient to plot the interfering signal level at the r-f grid that produces 1-percent cross modulation as a function of agc bias or gain reduction. (A schematic diagram and step-by-step procedure of the test set-up are available from the author.)

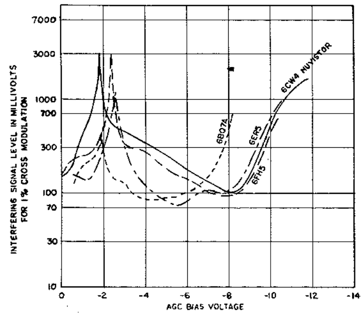


Fig. 5—Interfering signal level as a function of bias voltage for nuvistor and other types.

Fig. 5 shows the interfering signal level for 1-percent cross modulation as a function of bias voltage for the 6CW4, 6ER5, 6FH5, and 6BQ7. If gain reduction were used instead of bias voltage for the abscissa, all the curves would converge at cut-off. The 6CW4 nuvistor is somewhat better than the other tubes measured; it produces 1-percent cross modulation for a 100-millivolt interfering signal at the grid when —8 volts of bias is applied.

The actual signal level required at the input terminals of the TV tuner to produce 1-percent cross modulation is also a function of the transformation ratio and selectivity of the input circuit. The tube is "on its own," so to speak, in regions where two very strong high-channel TV station signals are present. In Union City (a favorite field-test location), for instance, there are 100,000-$\mu$v-plus signals from Channels 9 and 11. The input circuit offers little rejection to the undesired signal in this case, and the bias could be at the most unfavorable point for cross-modulation interference.

### THE PRODUCT-DESIGN TUNER

On the basis of engineering evaluation of the 6CW4 tube and in view of the excellent performance achieved in the early prototype nuvistor tuners, a product design was scheduled and completed. Inset in Fig. 1 is a functional block diagram of the RCA KRK-98 four-circuit, switch-type tuner including the input-balun, the high-pass filter assembly, the nuvistor r-f amplifier tube, and 6EA8 oscillator mixer tube. The signal from the antenna passes through the balun, high-pass filter and single-tuned input circuit before reaching the 6CW4 grid. Circuits represented by these three blocks must perform their transformation function and provide good selectivity without excessive power loss so that the excellent noise figure promised from earlier device measurements may be achieved. Mutual magnetic coupling is used on all channels in the double-tuned circuit following the nuvistor r-f amplifier; excellent skirt selectivity both above and below the signal frequency is assured.

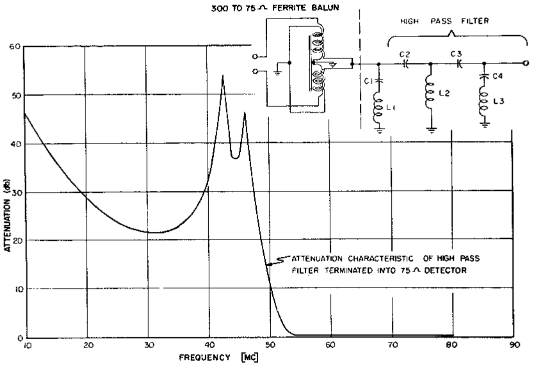


Fig. 6—Typical attenuation characteristics of the balun and high-pass filter.

The input balun (see Fig. 6) developed by L. A. Harwood is wound with miniature 150-ohm twin lead on a flat ferrite, 1 by ½ inches by ⅛ inch thick. It has very low loss (0.2 db) at 200 mc and a superior balanced-to-unbalanced ratio. The unit lends itself to automated machine winding and requires very little mounting space. The 300-ohm-balanced to 75-ohm-unbalanced connection is used principally because it makes possible a lower-loss, more-compact filter than can be achieved with a 300-ohm-balanced to 300-ohm-unbalanced connection. Typical attenuation characteristics of the balun and high-pass filter are shown in Fig. 6. Measurements were taken with the signal generator transformed from 50-ohm unbalanced to 300-ohm balanced to match the nominal input impedance of the balun. The high-pass filter is terminated in a 75-ohm resistive load.

The input tuned circuit of a TV receiver represents a perplexing compromise and has challenged the ability and imagination of tuner engineers from the inception of television. The

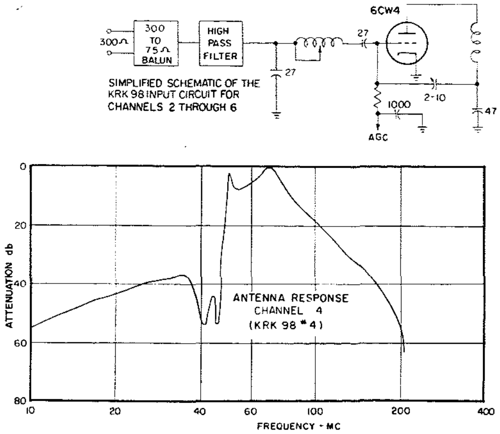


Fig. 7—Low-channel input circuit and attenuation versus frequency.

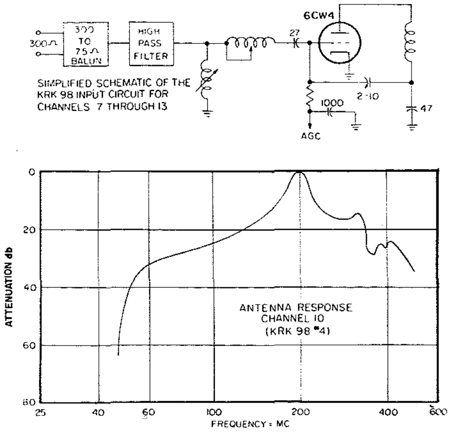


Fig. 8—High-channel tapped-inductance circuit and attenuation versus frequency.

KRK 98 circuit was chosen for the following reasons. A low-pass circuit offers protection to the r-f grid when receiving a low-channel TV station in the presence of a strong high-channel signal. On the other hand, it would give relatively little protection when tuned to a high channel in the presence of a strong FM or low-channel TV station. In either case, the high-pass filter of Fig. 6 will give a high degree of protection for signals at i-f frequencies and below (0 to 50 mc). These considerations and the fact that the necessary switching can be achieved on one side of a switch wafer led to the use of a low-pass circuit on the low channels (Channels 2 to 6) and a tapped-inductance, band-pass circuit on the high channels (Channels 7 to 13).

Fig. 7 shows a simplified diagram of the KRK 98 low-channel input circuit and the attenuation-versus-frequency characteristic taken with the grid and cathode acting as a high-impedance diode detector. The 2-to-10-$\mu\mu$f trimmer and the 47-$\mu\mu$f plate-circuit bypass combine with the grid-plate and plate-cathode capacitances to form a conventional bridge-neutralizing circuit. Neutralization is achieved at bridge balance; or, in other words, when a change in voltage across the plate inductance does not give rise to a change in voltage between grid and cathode. By combining the information available in the selectivity curve of Fig. 7 and the previously presented cross-modulation data (Fig. 5), it is possible to forecast the resultant cross modulation from any known level of interfering signal.

Fig. 8 shows the simplified schematic of the high-channel, tapped-inductance circuit. The inductance on both sides of the tap is made to vary (necessary for constant step-up) with simple "one-side" switching by taking advantage of the rotor and stator inductance changes.

### TYPICAL PERFORMANCE DATA

Table I shows important performance data for a typical KRK 98 tuner with average tubes. The noise figures were measured with a Polytechnic Research Development (PRD) Model 904 coaxial noise diode. Using the same noise diode and test set-up, tuners using the 6ER5 and 6FH5 measure 1½ to 2 db poorer. This is slightly better than expectations based on some of the earlier device measurements. Noise figures measured with a 5722 noise diode would average about 1½ db more optimistic on the high channels; the tests indicated that the PRD is more realistic at these higher frequencies. For instance, the PRD 904 used for these measurements correlated to within 0.2 db of the new Kay Thermanode at 200 mc. In both cases, a Linear Equipment Co. balun was used to match to the 300-ohm balanced input.

Over-all tuner gains are about 2 db better than the 6FH5 and 6ER5 tuners and considerably better than older tuners (cascode) with the same bandwidth. Image and i-f attenuations are seen to be excellent, and as previously covered, the r-f and mixer grids are adequately protected for cross modulation.

**TABLE I—PERFORMANCE DATA OF NUVISTOR TUNER**

Tubes: 6CW4, r-f; 6EA8 oscillator mixer. + B: 270 v at 20 ma.
Filament: 6.3 v a-c. Input: 300 ohm balanced.

| Channel | Voltage Gain, db | Noise Figure, db | Image Atten., db | I.F. Atten., db |
|---|---|---|---|---|
| 2 | 45 | 5.1 | >75 | 66 |
| 3 | 44 | 5.1 | >75 | 73 |
| 4 | 44.5 | 5.1 | >75 | >75 |
| 5 | 44.0 | 4.9 | >75 | >75 |
| 6 | 43.5 | 5.0 | >75 | >75 |
| 7 | 44.5 | 4.7 | >60 | >75 |
| 8 | 44.5 | 4.9 | >60 | >75 |
| 9 | 44.7 | 4.9 | >60 | >75 |
| 10 | 44.7 | 5.0 | >60 | >75 |
| 11 | 44.2 | 5.3 | >60 | >75 |
| 12 | 44.2 | 5.3 | >60 | >75 |
| 13 | 45 | 5.3 | >60 | >75 |

### CONCLUSION

Only one application of the nuvistor triode has been described, yet its use may be extended to computers, communications receivers, and other types of electronic equipment calling for a rugged and reliable tube of compact design. Eventually, a whole family of nuvistors may be developed, as signalled by the successful introduction of the developmental tetrode a short time ago.

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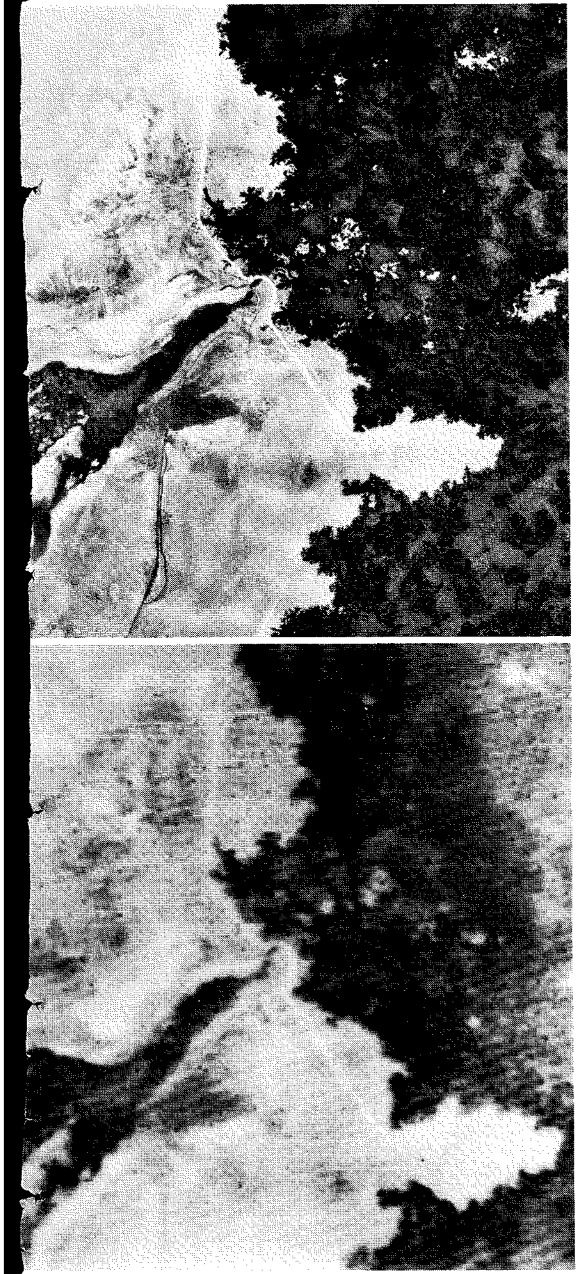
**G. C. HERMELING** received the BSEE from Washington University, St. Louis, Missouri in June, 1949, and joined RCA immediately upon graduation. His engineering work has been in television receiver design, principally in UHF and VHF tuners. His work has included the design of continuously tuned UHF converters, color and black-and-white television tuners, and developmental transistor tuners. Mr. Hermeling is a Senior Member of the IRE, and a member of Tau Beta Pi and Sigma Xi.

# MISSILE-BORNE TELEVISION

*by* **F. F. MARTIN, Mgr.,**

*TV Equipment Development*
*Airborne Systems Division*
*DEP, Camden, N. J.*



Comparison of an aerial photograph (top) and picture from the TV capsule.

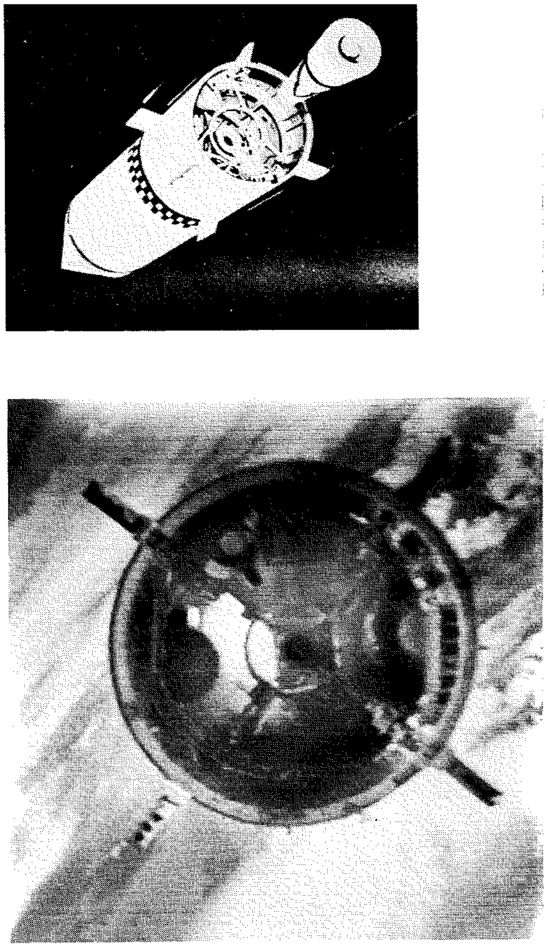
In future warfare, since time will be of paramount importance and expensive weapons will be employed over long distances, it will be essential for a field commander to know positively and quickly whether or not his weapon has performed as planned. Particularly, he must know if he has to launch yet another weapon to destroy the target. Ideally, in order to perform this evaluation of weapon effectiveness, he would like to be at the target area. Since this is not practical, a unique method to extend his senses to the target area has been provided by RCA and proved, through extensive field tests including actual missile firings, to be feasible and practical. This program has been called TFD—*Television Feasibility Demonstration.*

For this application, a small capsule containing a television camera and transmitter is mounted in a ballistic missile. This TV capsule is ejected from the missile at a prescribed point on the downward trajectory. Because of its high drag configuration, the capsule falls slower than the missile and, using slow-scan television techniques, obtains and transmits to a mobile ground station a visual record of the target area before and after missile impact.

In order to prove the feasibility of this concept, RCA built, under sponsorship of the Army Ballistic Missile Agency and the United States Army Research and Development Laboratory, a number of airborne units with a set of associated ground-monitoring and blockhouse checkout equipments. These units were installed in aerodynamic capsules furnished by the Chrysler Corporation Missile Division and were used in actual missile-borne tests at White Sands Missile Range, using a Redstone Missile as the carrier vehicle.

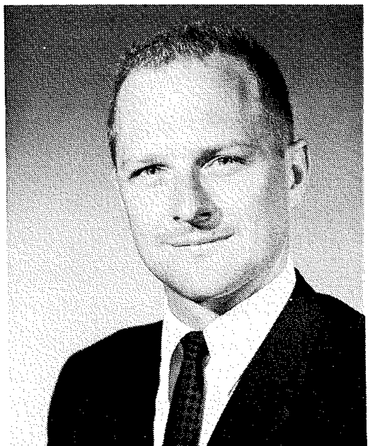
The TV pictures shown on this page are representative of pictures obtained by the ground-monitoring station during the April 19 firing. This station is equipped to display the capsule output in real time, or, as shown here, to make a reproducible permanent record of the results.

As a result of these missile tests and other tests performed with the system, the feasibility of the TFD concept has been established.



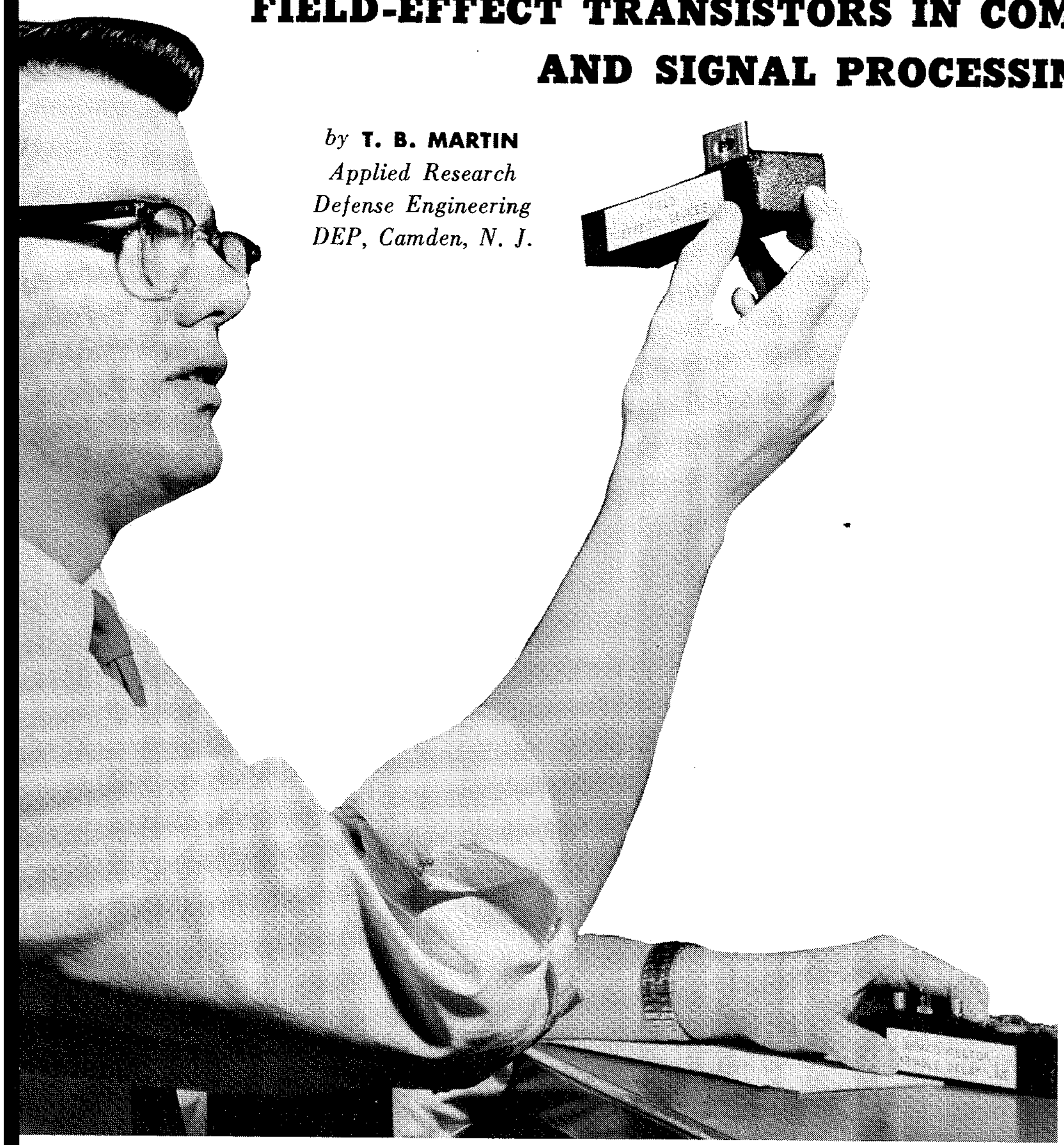
TV picture of missile just after ejection of capsule. Sketch at top shows relative missile and capsule positions at this time.

**F. F. MARTIN** received the BSEE from Newark College of Engineering in 1951 and the MSEE from Columbia University in 1955. Upon graduation from Newark College of Engineering, he became a Second Lieutenant in the USAF. After attending Guided Missile School, he was asigned to Patrick AFB for eighteen months as a Guidance and Control Officer for the Matador missile. After receiving his MS from Columbia, he was assigned to the Wright Air Development Center where his initial assignment was Project Engineer for development of interceptor-type infrared equipment. Later he acted as Project Engineer for advanced interceptor fire control systems, including the F-108. He directed development of the complete systems for these aircraft. He attained the rank of Captain before leaving the Air Force in 1958. That year he joined the Airborne Systems Division of RCA as staff assistant to the manager of the Airborne Radar and Missile Engineering Section. He was employed on the ASTRA Program for the development of the electronic system for the Canadian CF-105 Interceptor. Later, he became Systems Leader for the Television Feasibility Project. Mr. Martin is now Mgr., TV Equipment Development.



# FIELD-EFFECT TRANSISTORS IN COMMUNICATIONS AND SIGNAL PROCESSING

*by* **T. B. MARTIN**
*Applied Research*
*Defense Engineering*
*DEP, Camden, N. J.*



The author and a field-effect device.

NEW INTEREST has been focused on the field-effect transistor because of the ease with which integrated semiconductor circuitry can be achieved with this device. J. T. Wallmark and S. M. Marcus[1] have developed a whole group of such integrated circuits for use in digital computers. This article, however, describes some of the non-digital applications of the field-effect transistor with emphasis on achieving circuit functions that are difficult to realize with ordinary bipolar transistors; these applications include communications and signal-processing circuits.

### THE FIELD-EFFECT TRANSISTOR

Possessing several unique and useful features, the field-effect transistor (Fig. 1) is a three-terminal device that operates on a different principle than that of ordinary junction transistors. The three terminals are the drain, source, and gate; a channel of p- or n-type material is spaced between the drain and source. A p-n junction with a gate contact is formed along one or both sides of the channel region which consists of high-resistivity material. The gate material is of much lower resistivity. Between the gate and the body of the field-effect transistor is the junction depletion layer whose thickness can be varied by applying a reverse voltage to the gate terminal. Current flowing in the channel can be controlled by varying the width of the depletion region. This modulation of the current flowing in the channel requires very little power, since the junction is reverse-biased.

Ideal normalized drain characteristics shown in Fig. 2 are very similar to those of vacuum-tube pentodes. The characteristics are normalized to pinch-off voltage, $V_p$, and to the channel resistance with zero gate and drain voltages, $R_o$. These characteristics can be explained by observing the manner in which the width of the depletion region changes. Large drain currents, $I_D$, increase the potential difference between the gate and the drain end of the channel, $V_D—V_G$; consequently, the drain end of the channel becomes constricted. Since the source is grounded, the gate-source potential difference, $V_G$, remains constant as $I_D$ and $V_D$ are increased.

Consider the case in which the gate terminal is shorted to the source and drain current is flowing. The drain end of the channel becomes constricted and the point is reached at which the depletion layer extends across the width of the channel. Thereafter, the drain current remains constant as shown on the $V_G = O$ curve of Fig. 2. The drain voltage at which the current becomes constant is known as the pinch-off voltage, $V_p$. Also shown on Fig. 2 is the locus of $(V_D—V_G) = V_p$; the current is constant for drain voltages beyond this locus.

### SMALL-SIGNAL AMPLIFIER CIRCUITS

The field-effect transistor can be quite useful in applications requiring a high input impedance. High input impedance is presently achieved with semiconductors by the emitter follower and variations thereof. These circuits normally have low cut-off frequencies and do not simply achieve the input impedance of a reverse-biased junction. Both the grounded-source and grounded-

**T. B. MARTIN** graduated *magna cum laude* from the University of Notre Dame in 1957 with a BS in EE. He received his MS in EE in 1960 on the RCA Graduate Study Program at the University of Pennsylvania. He has done applied research work in the field of semiconductors, involving temperature stabilization of d-c and low noise video amplifiers, composite transistor structures, Hall Effect devices, parametric oscillators, semiconductor delay lines, and applications of the field-effect transistor. The field-effect transistor investigation resulted in applications such as analog multipliers electronically variable resistors, variable gain amplifiers, and high input impedance amplifiers. He has contributed significantly to a project concerned with a long-range communications system at vlf-lf frequencies. The areas investigated were small, low-frequency antennas, propagation, coding and decoding, low-frequency receiver design, and binary modulation processes. Mr. Martin presented a paper on d-c amplifiers at the AIEE Mid-Winter Conference in February 1959. He has one patent pending on the subject of field-effect transistor applications.

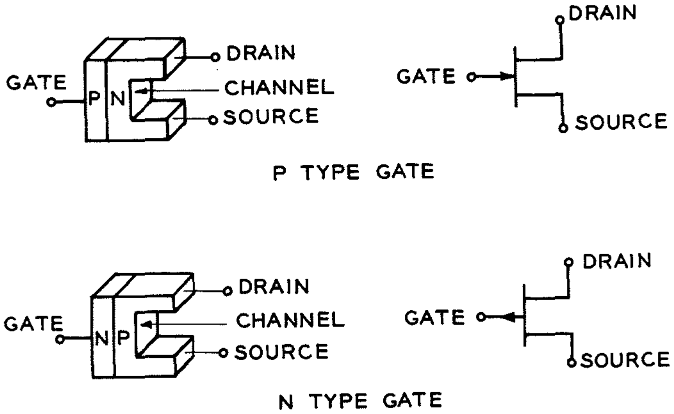


Fig. 1—Construction of field-effect transistors.

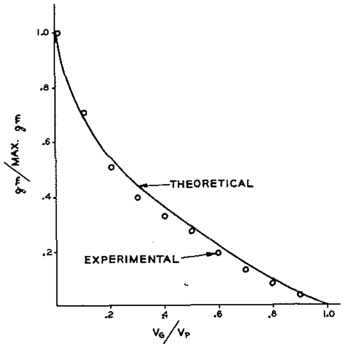


Fig 3—Transconductance as a function of gate voltage.

drain circuits are useful as high input impedance amplifiers. The grounded-source circuit has the property that the transconductance, $g_m$, depends upon d-c gate voltage (Fig. 3). For small input signals, optimum gain is achieved by operating as close as possible to zero gate bias. This effect is illustrated in Fig. 2 by the unequal spacing of the characteristics for equal increments in gate voltage. It is possible to vary the transconductance from the maximum value to essentially zero, with the greatest rate of change occurring near the maximum value. Typical units have maximum values of $g_m$ from 100 to 1000 micromhos.

A more useful application of the field-effect transistor is in the grounded-drain circuit, or source follower. Both high input impedance and low output impedance are achieved simply with one stage of amplification. It can be used as an impedance transformer between a high-impedance transducer and a low-impedance circuit such as the input to a bipolar-transistor amplifier.

Since the low-frequency equivalent circuit is identical to that of a vacuum tube, the operation is very similar to the cathode follower. An important property of this circuit is that the voltage gain is relatively independent of variations in $g_m$. The percent-gain variations caused by percent changes in $g_m$ are reduced approximately by a factor of $(1 + g_m R_L)$. This is the same factor by which the load resistance, as seen from the output terminals, is reduced.

The availability of a semiconductor device with the input impedance of a reverse-biased junction makes possible further extensions of the applications of semiconductor devices.

## ELECTRONICALLY VARIABLE RESISTANCE

The preceding descriptions deal mainly with conventional applications of the field-effect transistor; however, the unique features of the field-effect transistor may be used to obtain an electronically variable resistance.

For small drain currents, the grounded-source connection may be used as a variable resistance. To do this, the device must be operated very near the origin of the characteristics shown in Fig. 2. Also, operation must be in the first quadrant only, since the junction becomes forward-biased in the third quadrant, resulting in very nonlinear characteristics. A circuit which will overcome these disadvantages is shown in Fig. 4.

The gate bias supply, $V_B$, is symmetrical with respect to the drain and the source. The value of $R_G$ must be lower than the leakage resistance between the gate and either end of the channel, but higher than the maximum useful channel resistance. These two conditions ensure that the full value of $V_B$ appears across the gate and the two ends of the channel and that the maximum desired resistance can be achieved. The static characteristics of this circuit are shown in Fig. 5. The characteristics are very nearly those of a linear, variable resistance, and are of unequal lengths to satisfy the requirement that neither end of the channel becomes forward-biased or pinched-off. Fig. 5 shows the shape of the depletion region at seven different operating points. For a given gate bias, the movement caused

Fig. 2—Normalized drain characteristics of field-effect transistors.

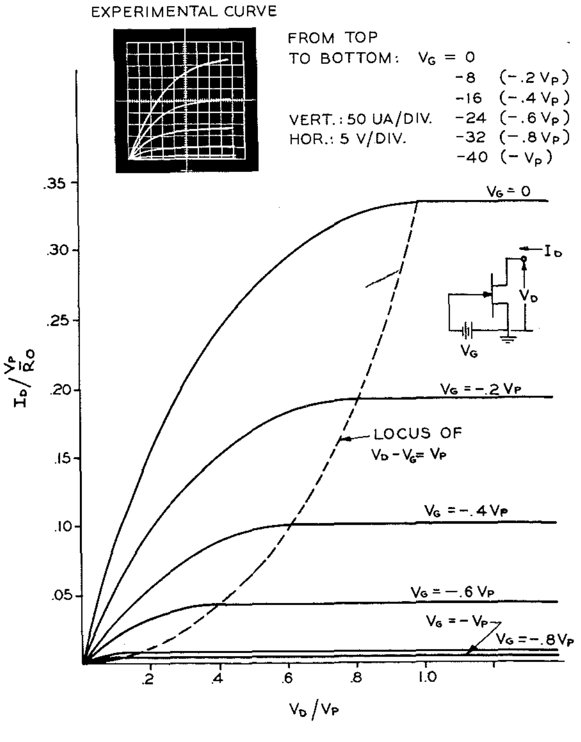
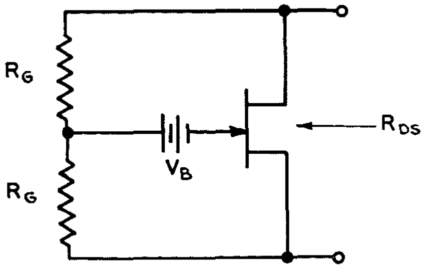


Fig. 4—Symmetrical gate-bias circuit.



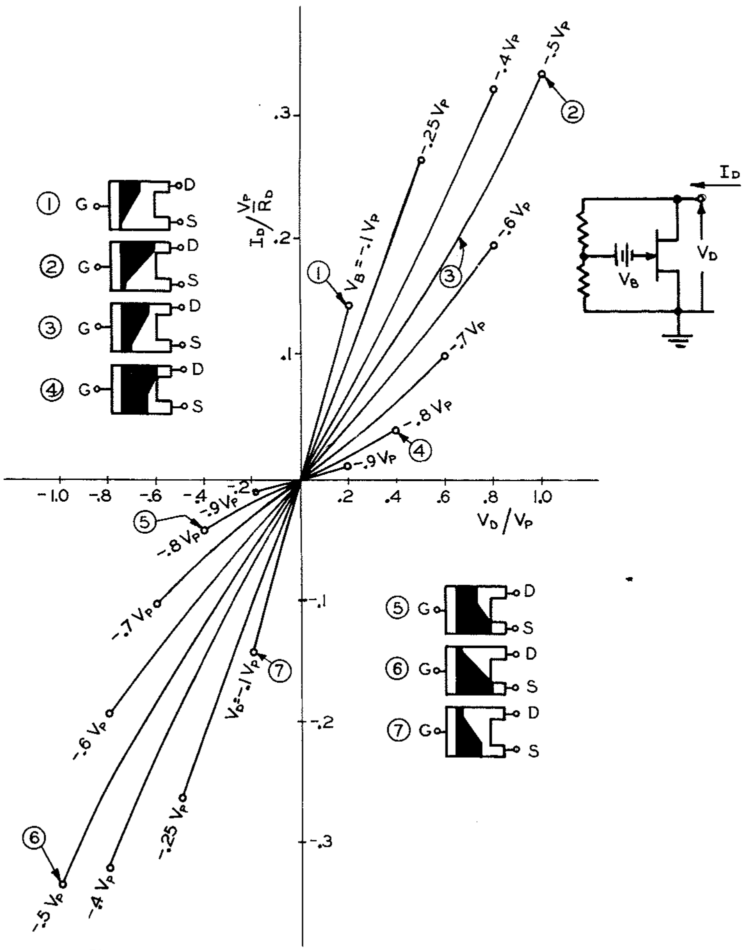


Fig. 5—Drain characteristics for center-tapped bias arrangement.

by an impressed drain voltage does not change the width of the depletion region at the center of the channel. However, the upper half of the region moves in a direction opposite the motion of the lower half. Thus, the channel resistance tends to remain nearly constant. For example, consider point 3 on Fig. 5. The movement of the top half of the depletion region has tended to increase the channel resistance while the movement of the lower half has tended to decrease the channel resistance. Points 2 and 6 of Fig. 5 represent the widest possible range in $V_D$, since pinch-off of one end of the channel is reached at the same time that zero bias occurs across the other end.

The cause of the slightly upward curvature of the center-tapped drain characteristic is the result of the square-root relationship between the width of the junction and the voltage across it. Fig. 6-a shows the normal drain characteristics of Unit 1792 for both first- and third-quadrant operation. Fig. 6-b shows the center-tapped drain characteristics of the same unit. The range of drain voltage for each fixed gate voltage was made equal to the largest possible value while still avoiding either the forward-biasing or pinching-off either end of the channel. Note that the maximum value of drain current is the same for both cases, as expected from Figs. 2 and 5. The agreement between the experimental curves and those predicted from the device equations is excellent.

Normally, the range of operation of the field-effect transistor as a variable resistance would not encompass the complete characteristics, but would be over a somewhat smaller excursion. Fig. 6-c shows the characteristics of Unit 1792 when varied between 33 kilohms and 2 megohms. Note that the characteristics are quite linear over the restricted range of ± 14 volts of drain voltage. The 2-megohm upper limit was determined by the value of $R_G$, which was set at 1 megohm, thereby presenting a total of 2 megohms in parallel with the channel resistance. Fig. 6-d shows the center-tapped characteristics of a higher-current, field-effect transistor. In this case the range of resistance was 1.1 kilohms to 2 megohms.

It is thus seen that the field-effect transistor, when used as an electronically variable resistance, provides the following outstanding features:

1) very small power necessary to vary the resistance
2) isolation between the resistor terminals and the signal that controls the value of the resistance
3) dynamic ranges of variation between 1000:1 and 10,000:1.
4) ability to handle signals of either polarity.
5) operation from d-c to several megacycles.

Some of the potential uses of an electronically variable resistance are:

1) agc circuitry
2) compression or expansion
3) mixing or analog multiplication
4) tunable phase shift oscillators
5) adaptive filters
6) variable RC time constant circuitry
7) chopper

### ANALOG MULTIPLIER

Perhaps the most interesting use of the field-effect transistor as an electronically variable resistance device is analog multiplication. The function of an analog multiplier, or linear mixer, is to electronically multiply two signals. When the frequencies of the inputs are $f_1$ and $f_2$, the output, or product, contains only $f_1 \pm f_2$. A circuit that can be used for multiplication is shown in

a) Normal Characteristic
Unit #1792
In Clockwise Direction
$V_G$ = 0, -8, -16, -24, -32
Sensitivity {V: .2ma/div. H: 10v/div.

b) Center-Tapped Characteristic
Unit #1792
In Clockwise Direction
$V_G$ = -2, -8, -16, -24, -30
Sensitivity {V: .2 ma/div. H: 10v/div.

c) Center-Tapped Characteristic
Unit #1792
In Clockwise Direction
$V_G$ = -8, -11, -16, -24, -44
Sensitivity {V: .1 ma/div. H: 5v/div.

d) Center-Tapped Characteristic
Unit #1793
In Clockwise Direction
$V_G$ = -5, -15, -29, -61
Sensitivity {V: 2 ma/div. H: 2 v/div.

Fig. 6—Variable resistance characteristics of field-effect transistors. Vertical: drain current. Horizontal: drain voltage. Running Parameter: gate voltage.

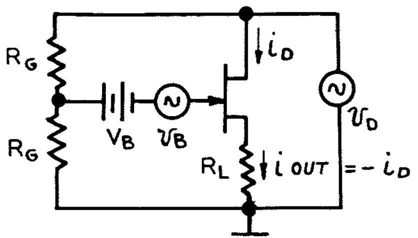


Fig. 7—Field-effect transistor analog multiplier.

Fig. 7; the two inputs are $v_D$ and $v_B$ and the current through $R_L$ is taken as the output. Over small ranges of $v_B$, the differential channel resistance is inversely proportional to $v_B$. The change in drain current caused by $v_B$ is therefore given as:

$$\Delta i_D = \frac{v_D}{\Delta R_{DS}} = k v_D v_B$$

where $k$ = a constant. The total output current contains a direct term of frequency, $f_D$, a term containing the product of the two input signals and error terms which result in harmonic distortion. The direct term may be neglected on the assumption that $f_D$ and $f_B$ are sufficiently apart, that $f_D$ may be removed by filtering, or that the presence of $f_D$ is unobjectionable.

Consider a unit with $R_o$ = 1000 ohms and $V_p$ = 50 volts. It is possible to determine the theoretical performance of this particular transistor from the device equations.[2] The effect of $v_B$, $v_D$, and $V_B$ upon the output current in the first quadrant as obtained with the device equations is presented in Fig. 8. This, together with the analytical results, show that 1) output signal is maximized by minimizing $R_o$ and $V_B$, and 2) distortion is minimized by maximizing $V_B$, and therefore $V_P$.

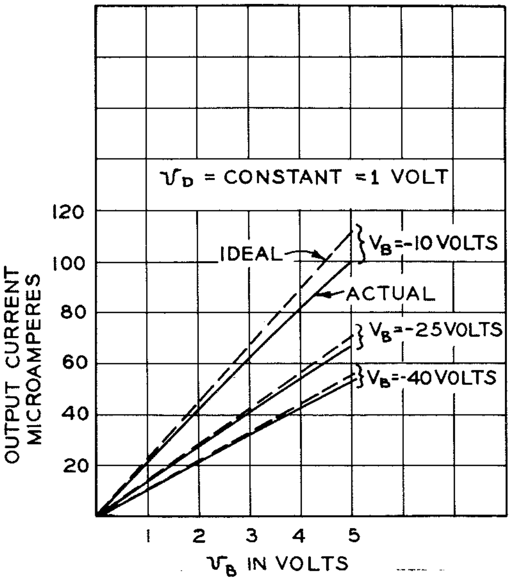


Fig. 8—Multiplier transfer characteristic.

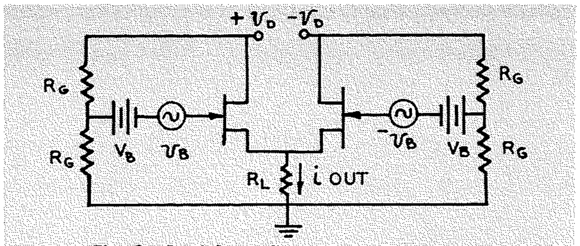


Fig. 9—Dual field-effect transistor multiplier.

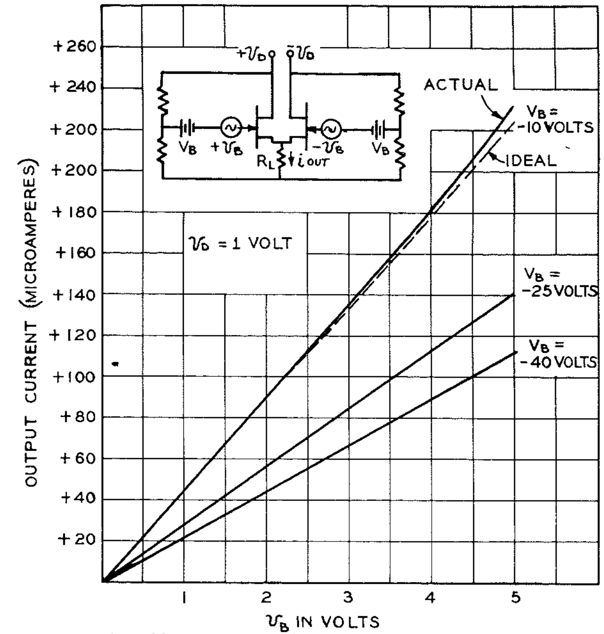


Fig. 10—Dual multiplier transfer characteristic.

It can therefore be deduced that a field-effect transistor with low $R_o$ and high $V_p$ would be optimum for the multiplier application.

The numerical results for the example chosen show that very low values of distortion can be achieved. For example, with $v_D$ and $v_B$ each 1 volt and $V_B$ equal to 40 volts, the output current is 10 microamperes and the distortion is only 0.70 percent.

### DUAL FIELD-EFFECT-TRANSISTOR MULTIPLIER

It is possible to eliminate the direct term and at the same time greatly reduce the distortion by using two matched field-effect transistors. In the circuit shown in Fig. 9, the direct term has been eliminated; the output current doubled, and the distortion greatly diminished. In any practical circuit, the realization of these advantages will depend on how well the two units are matched. The performance of the dual field-effect-transistor multiplier with the same device as used for the first example is shown in Fig. 10. One quadrant of the four is shown. Note the extremely low values of distortion that result from this arrangement. Considering again the case for which $v_D$ and $v_B$ are each 1 volt and $V_B$ is 40 volts, the output current is 22 microamperes and the distortion is only 0.01 percent.

### SUMMARY

The electronically variable resistance and multiplier are but two possible uses of the field-effect transistor. There are many other areas in which the field-effect transistor can supplement the conventional bipolar transistor. A more detailed description of the applications of the field-effect transistor is available in Reference 3.

### ACKNOWLEDGEMENT

The author would like to acknowledge that the use of the field-effect transistor as a variable resistance and analog multiplier was the outgrowth of joint work with S. M. Marcus and F. L. Putzrath.

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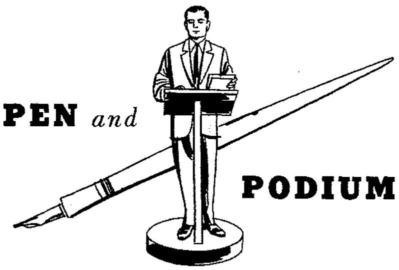

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**New X-Ray Equipment from RCA**
V. E. Buhrke: Symposium, Univ. of Illinois, Urbana, Ill., May 27, 1960.

**Data Communications**
E. D. Becken: Industrial Communications Assn. Conf., Miami, Florida, May 26, 1960.

**Standby Radio Facilities**
S. N. Fifield: Communications Conf., New York City, June 20-27, 1960.

**Interconnection of RCAC Semi-Automatic Telex**
W. Shaw: Communications Conf., New York City, June 20-27, 1960.

**Auto-Data—RCA's Automatic Store and Forward Message Switching System**
J. L. Owings: Communications Conf., New York City, June 20-27, 1960.

**An Automatic System for Handling International Telegrams**
E. D. Becken and R. K. Andres: Communications Conf., New York City, June 20-27, 1960.

**Frequency Division Multiplexing on Transoceanic Cables**
W. Lyons: Communications Conf., New York City, June 20-27, 1960.

**Digital Data Codes**
T. S. Maguire: Communications Conf., New York City, June 20-27, 1960.

**Alternate Use Voice Data Systems**
J. C. Hepburn: Communications Conf., New York City, June 20-27, 1960.

**50 Baud Conversion**
J. B. Moore: Communications Conf., New York City, June 20-27, 1960.

**Satellite Communications**
S. Metzger and D. S. Rau: Communications Conf., New York City, June 20-27, 1960.

**Auto-Data—RCA's Automatic Message Switching System**
H. P. Guerber, T. L. Genetta and A. S. Rettig: IRE, Professional Group on Military Electronics, Winter Convention, Los Angeles, Calif., Feb. 3-5, 1960.

**Remote Control of TV Microwave Equipment**
J. P. Bullock: IRE International Convention, New York City, March 21-24, 1960.

**TV Station Automation**
R. D. Houck: IRE, Professional Group on Broadcasting, Philadelphia Chapter, April 14, 1960.

**Tunnel Diode Computer Logic Circuits**
R. K. Lockhart: IRE, Professional Group on Electronic Computers, Washington, D. C., April 21, 1960.

**A Transistorized Portable Magnetic Recording Channel for Film**
C. E. Hittle, M. Rettinger and K. Singer: SMPTE Convention, Los Angeles, Calif., May 1-7, 1960.

## RCA STAFF

**Proposed Symbology for Digital Systems**
D. Drusdow and W. J. Smith (AFCRC): AIEE Northeastern Dist. Mtg., Providence, Rhode Island, May 2, 1960.

## DEFENSE ELECTRONIC PRODUCTS

**Korotkoff Sounds in Humans**
J. D. Wallace (D. H. Lewis, S. A. Khalil of Philadelphia General Hospital): 59th Mtg. Acoustical Society of America, Providence, R.I., June 9-11, 1960.

**Intracardiac Phonocardiography**
J. D. Wallace (D. H. Lewis, Philadelphia General Hospital and J. R. Brown, U.S. Naval Air Development Center): Third Internat'l Conf. of Medical Electronics, London, England, July 21-27, 1960.

**Systems Planning in Industry—The Link Between Military and Industrial Planning**
N. I. Korman: Defense Planning Seminar, Electronic Industries Assn., Washington, D. C., March 15, 1960.

**A Synchronous Detection System Utilizing a New Method of Frequency and Phase Control**
D. H. Sapp: M. S. Thesis, Univ. of Penn., June, 1960.

**Broadband UHF Amplifiers for High Power Transmitters**
A. H. Riccio: M. S. Thesis, Univ. of Penn., June, 1960.

**Submarine Interior Communications and Speech Processing**
M. L. Graham: Philadelphia Naval Reserve Company Mtg., April 25, 1960.

**Electronic Packaging Evaluation**
C. Dunaief: Electronic Circuit Packaging Symposium, Boulder, Colorado, August 18, 1960.

**Digital Evaluation Equipment—A Multipurpose Test System**
O. T. Carver: AIEE Aero-Space Transportation Conf., San Diego, California, Aug. 9-12, 1960.

**Pulse Radar for Trajectory Instrumentation**
D. K. Barton and S. M. Sherman: 6th Nat'l Flight Test Instrumentation Symposium, Instrument Sociey of America, San Diego, Calif., May 3, 1960.

**A Special Purpose Data Processor for Digital Data Communication**
E. Schlain: Assn. for Computing Machinery, Milwaukee, Wisconsin, Aug. 23-25, 1960.

**Customer Evaluation of Finished Products**
G. J. Conroy: Electrical Contacts Seminar, Univ. of Penn., June 6, 1960.

**Design and Operation of Four-Frequency Parametric Up-Converters**
J. A. Luksch, E. W. Matthews and G. A. VerWys: PGMTT Symposium, San Diego, Calif., May 9, 1960.

**Value Engineering**
S. Robinson: IRE Professional Group on Reliability and Quality Control, April 12, 1960.

**Reliability—An Engineering Problem for Management**
M. M. Tall: *Industrial Research,* July, 1960.

**Reliability Prediction—Its Validity and Application as a Design Tool**
T. C. Reeves: 1960 Design Engineering Conf., New York City, May 24, 1960.

**The Technical Writer and the Technical Proposal**
J. F. Biewener: Society of Technical Writing and Editors, Delaware Valley Chapter, Wilmington, Del., May 26, 1960.

**System Evaluation of Low Noise Radar Sensitivity**
S. Charton: 4th Nat'l Convention on Military Electronics, Washington, D. C., June 27, 1960.

**Circuit Applications of the Field Effect Transistor**
T. B. Martin: MSEE Thesis, Univ. of Penn, June, 1960.

**How to Write Operating Procedures**
D. P. Simonton: August, 1960 issue of *Product Engineering.*

**A Study of the Effects of Ground Reflections—A Monopulse Height Finding Radar**
W. H. Sheppard: RCA Microwave Symposium, June 8, 1960.

**Beacon Transmitters and Power Supply for an Inflatable Satellite**
J. C. McCubbin and H. B. Goldberg: AAS Sixth Annual Mtg., Jan., 1960.

**Optimizing the Design of a Solar Power Supply System**
S. Winkler and C. C. Osgood: AAS Sixth Annual Mtg., Jan., 1960.

**Weather Radar Observations from an Earth Satellite**
J. E. Keigler and J. Krawitz: AMS Weather Radar Conf., Feb., 1960.

**Electric Charge Transport through Organic Polymer Films**
J. J. Spokas: APS Princeton Univ. Conf., May, 1960.

**The Choice of Estimation Procedure and Loss Function for Radar System Synthesis**
J. E. Keigler: Symposium on Decision Theory, April, 1960.

**Structural Design of the Tiros Meteorological Satellite**
R. W. Northup: ARS Convention, April, 1960.

**Television Cameras for Space Exploration**
M. H. Mesner and J. R. Staniszewski: *Astronautics,* May, 1960.

**Operation and Control Equipment for the Tiros I Meteorological Satellite System**
V. D. Landon and E. A. Goldberg: *Astronautics,* June, 1960.

**Ground Data Processing for the Tiros I Meteorological Satellite System**
J. E. Keigler and C. B. Oakley: *Astronautics,* June, 1960.

**Self-Contained Ion Propelled Orbital Control Systems**
E. R. Walthall: USNRL Proceedings, May, 1960.

**Thermal Considerations for the Tiros I Meteorological Satellite**
M. Ritter: *Astronautics,* June, 1960.

**Electrostatic Tape Camera for Satellite Inspection**
S. W. Spaulding, B. P. Miller, and C. M. Burrill: ARS Convention, June, 1960.

**Spin-Stabilization of the Tiros I Satellite**
H. Perkel: *Astronautics,* June, 1960.

**The Tiros I Satellite Structure Design**
C. C. Osgood: *Astronautics,* June, 1960.

**The Tiros I Meteorological Satellite**
S. Sternberg and W. Stroud: *Astronautics,* June, 1960.

## RCA VICTOR CO., LTD.

***(Montreal, Canada)***

**Electromagnetic Properties of High Temperature Air**
M. P. Bachynski, T. W. Johnston and I. P. Shkarofsky: *Proceedings of the IRE*, March, 1960.

**The Noise Figure of Junction Transistors**
J. M. Stewart: Internat'l Convention on Transistors and Associated Semiconductor Devices, London, England and recently published by the Institution of Electrical Engineers.

**Silicon Junction Devices Detect Nuclear Particles**
R. W. Jackson and P. P. Webb: *Canadian Electronics Engineering*, April, 1960.

## SEMICONDUCTOR AND MATERIALS DIVISION

**Bias Stabilization of Semiconductor Circuits**
K. E. Loofbourrow: Semiconductor and Solid-State Circuits Course, St. Louis, Mo., March 1, 1960.

**Some Aspects of a Semiconductor Surface**
D. J. Donahue: Electrochemical Society Mtg., Dearborn, Michigan, March 17, 1960.

**Micromodule Components—A Review of the State of the Art**
R. A. Felmly: IRE Internat'l Convention, New York City, March 21-24, 1960.

**A New Concept in Transistor Converters**
L. Plus and R. A. Santilli: IRE Internat'l Convention, New York City, March 21-24, 1960.

**Maximum Stable Collector Voltage for Junction Transistors**
R. Schmeltzer: *Proceedings of the IRE*, March, 1960.

**Low-Frequency Operation of Transistors**
H. M. Kleinman: IRE Transistor Lecture Series, Chicago, Ill., April 6, 1960.

**High-Speed Logic Using Low-Cost Mesa Transistors**
D. Gipp, R. D. Lohman, R. R. Painter and B. Zuk: IRE Spring Technical Conf., Cincinnati Section, Cincinnati, Ohio, April 13, 1960.

**Forced Periodic Changes of Kinetic Energy of Gas Molecules as a Means of Vacuum Measurement**
H. J. Schwarz: *The Review of Scientific Instruments*, April, 1960.

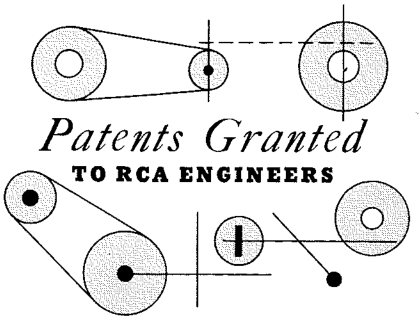
**Dielectric Properties of Zirconates**
B. Schwartz and H. Stetson: American Ceramic Society Mtg., Philadelphia, Pa., April 25, 1960.

**Hermetic Ceramic Package for Transistors Utilizing Ultrasonic Soldering**
H. D. G. Scheffer, W. Miller, W. H. Liederbach and A. J. Pikor: American Ceramic Society Mtg., Philadelphia, Pa., April 25, 1960.

**A Diffusion Mask for Germanium**
E. L. Jordan and D. J. Donahue: Semiconductor Symposium, Electrochemical Society, Chicago, Ill., May 2-5, 1960.

**Organiz Depolarized Dry Batteries**
G. S. Lozier: Fourteenth Annual Power Sources Conf., Atlantic City, N. J., May 17-19, 1960.

**Dynamic Temperature Coefficient of Micro-Element Inductors**
G. Hauser: IRE Internat'l Convention, New York City, March 21-24, 1960.



# Patents Granted
## TO RCA ENGINEERS

BASED ON SUMMARIES RECEIVED OVER A PERIOD OF ABOUT TWO MONTHS

## DEFENSE ELECTRONIC PRODUCTS

**Multiple Contact Connector**
2,942,229—June 21, 1960; R. H. Berger.

**Horn-Type Transducer**
2,942,071—June 21, 1960; A. L. Witchey.

**Radar Calibration System**
2,942,256—June 21, 1960; K. T. Strickland.

**Selective Detection of Radar Targets in the Presence of Noise Signals**
2,943,316—June 28, 1960; F. D. Covely III.

## INDUSTRIAL ELECTRONIC PRODUCTS

**Mechanical Movement**
2,939,332—June 7, 1960; R. H. Peterson.

**Sorting Apparatus**
2,935,732—May 3, 1960; H. P. Guerber.

**Color Television**
2,938,073—May 24, 1960; R. K. Lockhart.

**Data Input Control System**
2,939,631—June 7, 1960; A. Burstein.

**Controlled Sound Reproduction**
2,941,044—June 14, 1960; J. E. Volkmann.

**Picture and Sound Presentation Systems**
2,940,356—June 14, 1960; J. E. Volkmann.

**Printer Control System**
2,941,188—June 14, 1960; D. Flechtner and Carl C. Eckel.

**Electron Beam Controlling Apparatus**
2,943,218—June 28, 1960; B. R. Clay and L. R. Kirkwood.

**Electronic Extensor**
2,938,078—May 24, 1960; A. E. Canfora, H. J. Kishi, S. Sharin and A. Liguori.

## ELECTRON TUBE DIVISION

**Ruggedized Electron Tube**
2,937,305—May 17, 1960; L. P. A. De Backer.

**Method of Making Stem Assembly for Ultrahigh Frequency Electron Tubes**
2,941,279—June 21, 1960; N. S. Freedman.

**Apparatus for Use in the Manufacture of Mosaic Screens for Color-Kinescopes, Etc.**
2,941,457—June 21, 1960; M. R. Weingarten.

**Optical Projection System**
2,942,099—June 21, 1960; N. R. Goldstein.

**Electron Multiplier**
2,942,132—June 21, 1960; A. A. Rotow and E. A. Dymacek.

**Eyeletted Bulk Spacer Assembly**
2,943,224—June 28, 1960; M. D. Forte.

**Traveling Wave Type Tube and Method of Manufacture**
2,943,228—June 28, 1960; B. Kleinman.

## RCA VICTOR HOME INSTRUMENTS

**Plural Channel Wide Band Amplifier**
2,935,695—May 3, 1960; S. Wlasuk.

**Automatic Gain Control Systems**
2,937,235—May 17, 1960; C. W. Hoyt and L. P. Thomas, Jr.

**Interlock Assembly**
2,938,188—May 24, 1960; A. G. Lazzery.

**Frequency Converter Having Means to Prevent Self-Quenching**
2,939,000—May 31, 1960; L. M. Krugman.

**Remote Control System**
2,943,146—June 28, 1960; L. P. Thomas, Jr.

**Television Test Equipment**
2,943,144—June 28, 1960; S. Wlasuk.

## RCA STAFF

**Image Viewing Apparatus**
2,942,254—June 21, 1960; G. L. Beers.

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## ELECTRON TUBE DIVISION

**The Nuvistor, A New Design Concept for Electron Tubes**
R. N. Peterson: AIEE Philadelphia Section Mtg., Cherry Hill, N. J., March 9, 1960.

**Modern Traveling-Wave Tubes and Related Solid-State Microwave Amplifiers**
C. L. Cuccia: IRE Dallas-Fort Worth Section Mtg., Dallas, Texas, March 10, 1960.

**System Thinking**
W. K. Halstead: Northern New Jersey Chapter, American and Inventory Control Society, Newark, N. J., March 10, 1960.

**SSB Exciter Circuits Using a New Beam-Deflection Tube**
H. C. Vance, *QST*, March, 1960.

**Design Features and Fabrication of Traveling-Wave Tubes**
A. J. Biancull: IRE/AIEE Student Chapter Mtg., New York City, March 25, 1960.

**Receiving-Type Tubes**
M. B. Knight: *Electronics*, April 29, 1960.

**Non-Reciprocal Attenuation of Ferrites in Single-Ridge Waveguides**
T. S. Chen: *IRE Transactions*, Professional Group on Microwave Theory and Techniques, March, 1960.

**Theory and Practice of an EDP Application**
W. K. Halstead: EDP Seminar for Manufacturing and Production-Control Executives, Camden, N. J., May 2, 1960.

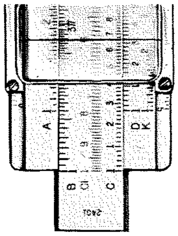
**Compact 20-Watt Hi-Fi Amplifier**
H. A. Wittlinger: *Electronics*, May, 1960.

**Super-Power Electronics**
M. V. Hoover: Rensselaer Polytechnic Institute, Troy, N. J., March 17, 1960.

**The Effectiveness of Ultrasonic Degreasing as Measured by Radiotracer Techniques**
E. L. Romero and H. A. Stern: IRE Internat'l Convention, New York City, March 21-24, 1960.

**Fiber Optics—New Tool in Electronics**
R. G. Neuhauser: SMPE Convention, Los Angeles, Calif., May 1-7, 1960.

# *Engineering* NEWS *and* HIGHLIGHTS



## 31 ENGINEERS EARN MS IN RCA GRADUATE STUDY PROGRAM

The engineers listed below have been awarded Master's Degrees through their participation in RCA's Graduate Study Program. Their achievement was honored at a recent dinner and by presentation of a distinctive certificate. This program, under the direction of **H. G. Matz,** Administrator, Graduate Study Program, College Relations, RCA Staff, Camden, was described in Vol. 5, No. 5 of the RCA ENGINEER in "The Engineer and the Corporation: RCA Graduate Study Program."

| | | | |
|---|---|---|---|
| **R. E. Atkins,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **J. Q. Butler,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **C. P. Clasen,** DEP, Moorestown | MS, Phys | University of Pennsylvania | 6/10/59 |
| **M. S. Crouthamel,** DEP, Camden | MSME | University of Pennsylvania | 6/10/59 |
| **D. L. Fried,** DEP, Princeton | MS, Phys | Rutgers the State University | 1/20/59 |
| **G. W. Hernan,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **G. J. Herskowitz,** DEP, Princeton | MSEE | Rutgers the State University | 6/3/59 |
| **J. E. Hoffman,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/10/59 |
| **V. D. Horvat,** DEP, Camden | MSEE | University of Pennsylvania | 6/10/59 |
| **J. A. Luksch,** DEP, Moorestown | MSEE | University of Pennsylvania | 2/13/60 |
| **E. P. McGrogan,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **T. B. Martin,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **G. Miller,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/15/60 |
| **R. K. Shreve,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/15/60 |
| **J. D. Sullivan,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/10/59 |
| **J. M. Anderson,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **J. M. Bailey,** IEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **W. T. Brennan,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **A. T. Crowley,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/15/60 |
| **D. Karlsons,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **R. F. Koontz,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/15/60 |
| **N. F. McAllister,** DEP, Moorestown | MSEE | University of Pennsylvania | 6/15/60 |
| **H. S. Miller,** RCA Laboratories | MSEE | University of Pennsylvania | 6/15/60 |
| **J. E. Palmer,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **E. J. Rapoza,** DEP, Camden | MSME | University of Pennsylvania | 6/15/60 |
| **J. D. Rittenhouse,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **G. L. Tate,** DEP, Moorestown | MSME | University of Pennsylvania | 6/15/60 |
| **E. L. Willette,** RCA Laboratories | MSEE | University of Pennsylvania | 6/15/60 |
| **P. E. Wright,** DEP, Camden | MSME | University of Pennsylvania | 6/15/60 |
| **K. K. Zeiger,** DEP, Camden | MSEE | University of Pennsylvania | 6/15/60 |
| **S. Skalski,** SC & M, Somerville | MS, Phys | Rutgers the State University | 6/8/60 |

### KALKMAN AND FRIEDMAN AWARDED SLOAN FELLOWSHIPS

Two RCA engineers have been chosen in national competition for 1960-61 Alfred P. Sloan Foundation on fellowships. They are: **Eugene C. Kalkman** of the Airborne Systems Division (DEP, Camden), and **Stan Friedman** of the West Coast Missile and Surface Radar Division (DEP, Van Nuys, Calif.). Mr. Kalkman will study at the Massachusetts Institute of Technology and Mr. Friedman will study at Stanford University. The Alfred P. Sloan fellowships are part of the Executive Development program of the School of Industrial Management at MIT. It is designed for a small number of exceptionally able young men whose employers nominate them because they show marked promise of growth.

### SOMMERS AWARDED GUGGENHEIM, FULBRIGHT FELLOWSHIPS

**Dr. Henry S. Sommers** of the RCA Laboratories technical staff and an author in this issue (see p. 4) was recently awarded both a Guggenheim Fellowship and a Fulbright Fellowship for study and lecture abroad for the academic year 1960-61. Dr. Sommers will begin a six-months' affiliation with Hebrew University, Jerusalem, in December, 1960, for research as well as lecture in solid-state physics and electronics. Prior to his residence in Jerusalem, he will spend several months in Zurich at Laboratories RCA, Ltd.

### RAMBERG AWARDED FULBRIGHT

**Dr. Edward G. Ramberg,** a Fellow of the RCA Laboratories technical staff, was recently awarded a Fulbright Fellowship for study and lecture abroad for the academic year 1960-61. He has been named Fulbright lecturer in thermo- and photoelectricity at the Technische Hochschule, Darmstadt, West Germany, for 1960-61.

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### ROYS ADVANCES TO NEW POST

**H. E. Roys** has been elevated to the post of Chief Engineer, RCA Record Division, by A. L. McClay, General Plant Manager at Indianapolis. Formerly Manager of Engineering, Mr. Roys has been active in RCA Engineering for many years and is also prominent author-inventor (see RCA ENGINEER Vol. 5—No. 2 "The Story of the Stereo Disc" for up-to-date biography.)

### NEW RCA RESEARCH LABORATORY IN JAPAN

RCA will open a new research laboratory in Tokyo, Japan in the near future to conduct fundamental studies in the physics and chemistry of solids. *Laboratories RCA, Inc., Tokyo* will be located in a building now under construction.

**M. E. Karns,** Director of License Operations, RCA International Division, announced the appointment of **Dr. Martin C. Steele** as Director of Research of the new laboratory. Dr. Steele, has been a physicist on the staff of RCA Laboratories, Princeton. The Tokyo laboratory will be staffed initially by scientists to be recruited from Japanese university graduates, with gradual expansion as required by the research program. The organization, devoted to fundamental research, will not be concerned with engineering development for the manufacture of electronic equipment.

The Tokyo laboratory is the second to be established abroad by RCA. The first was opened in 1955 at Zurich, Switzerland to conduct basic research in the European scientific environment.

### DR. GOLDSMITH KEYNOTES IEP ENGINEERING DINNER

Continuing its program of promoting communication within the professional staff, IEP on June 7, 1960, held its third Engineering Dinner (see photo) at the Ivystone Inn. At this meeting, **Dr. Alfred N. Goldsmith** (inset photo) discussed "The Engineer: His Scope and Opportunities in Modern Society." Dr. Goldsmith, a Founder and Charter Member of the IRE and widely known authority and inventor in the electronics industry, discussed the various phases of modern society and elaborated on professional occupation. He stressed the fact that ". . . the value of the professional man to society and to himself is his questing and questioning mind, his occasional nonconformity with accepted lines of thought, his particular brand of idealism, and his rare willingness to accept facts and eschew personal opinions or prejudices in his work."

Photo of IEP Engineering Dinner — Dr. A. N. Goldsmith inset.

## COMMITTEE APPOINTMENTS

In RCA Communications, Inc. **Walter Lyons** has been named an Alternate Member of the Administrative Committee, IRE Professional Group on Communications Systems. **C. G. Dietsch** has been named a Member of IRE Committee 15 on Radio Transmitters and IRE Subcommittee 15.2 on Radio Telegraph Transmitters.
—*D. S. Rau*

**L. J. Ortino,** Mgr., Design and Development, Radar Equipment Engineering, M and SR, DEP, Moorestown, has been appointed Chairman of a new EIA Committee on Solderless Wrapped Connections.

In the Electron Tube Division, Lancaster, **J. Forman,** Engineering Services, has been elected First Vice President of the Lincoln Chapter, Pennsylvania Society of Professional Engineers. **H. Walton,** Manufacturing Engineering, has been elected President, Lancaster Chapter, American Institute of Industrial Engineers. Also in the AIIE, **J. Spencer** of Manufacturing was elected Director of Publicity.

The Lancaster Engineering Education Committee has established a new subcommittee for professional activities and societies. Named to serve on this were **A. M. Morrell,** Chairman; and **R. G. Neuhauser, D. T. Copenhafer, P. W. Kaseman,** and **J. M. Forman.** —*H. S. Lovatt*

Elsewhere in the Electron Tube Division, **I. M. Rehm,** Color Development Shop, was named Chairman, JEDEC Subcommittee on Packaging of Transmitting and Microwave Tubes. **A. L. Smith,** Chem. and Phys. Lab., was chairman of a session at the Phosphor Symposium of the Spring Electrochemical Society Meeting in Chicago in May. **M. Berg,** Chem. and Phys. Lab., was Session Chairman Moderator of the Symposium on Ceramic-to-Metal Seals at the Spring Meeting of the Electrochemical Society. **P. D. Strubhar,** Chem. and Phys. Lab., was Chairman of a Tri-Chapter Meeting of the American Society of Metals held in May at Gettysburg, Pa.

From the engineering staff at Harrison, **T. M. Cunningham,** Engineering Administration, will be Business Chairman of the fall lecture series of the Northern New Jersey Section of the IRE. **R. E. Brown,** Applications Lab., will continue in his dual role as Chairman of their Student Activities Committee and Student Activities Editor of their *Newsletter.* **J. Hallahan,** Commercial Engineering, will continue as a Feature Editor. —*T. M. Cunningham*

In the Advanced Systems Group, Surface Communications Division, DEP, N.Y.: **Dr. F. Assadourian** is a member of the Planning Committee for RCA communication seminars. **P. G. Forsythe** is a member of the Papers Committee, IRE Professional Group for Communications Systems, and of the Papers and the Meeting Committee of the N. Y. Communications Division, AIEE. **Dr. C. Hammer** is Educational Chairman, District 46, Toastmasters International.

In SurfCom, Camden, **A. H. Coleman** is Chairman of the Executive Board for FAST (Fieldata Applications, Systems, and Techniques). **L. E. Mertons** is Vice Chairman, Meetings and Papers Committee, Philadelphia Section, IRE. **R. L. Rocamora** is serving on the AIEE's National Technical Committee (Data Communication) and is Chairman of their Student Activity Committee (Philadelphia Section). **D. Caplan** is a Member of the AIEE National Technical Committee (Automation, Data Processing). **J. L. Santore** is Treasurer of the Eleventh National Conference of IRE Professional Group on Voice Communications, to be held Dec. 1-2, 1960. **W. F. Meeker** is a member of the Speech Committee of the Acoustical Society of America, and a member of the American Standards Committee (Z-24, W-36). **P. Riley** is member of the National Administrative Committee of the IRE Professional Group on Industrial Electronics, and National Membership Chairman of the IRE Professional Group on Reliability and Quality Control. **G. V. Jacoby** is Chairman of the Science, Electronics, and Administration Technical Division of the AIEE. **L. Wolin** is a Member of the Steering Committee of the IRE-PGAC. **R. M. Carrell** is Chairman of the IRE Professional Group on Audio (Philadelphia Chapter). **P. S. Alday** is Chairman for Development, Engineers Club of Philadelphia. —*R. E. Patterson*

In IEP, F. J. Herrmann has been elected Vice President in Charge of Curricula for the Haddonfield Adult Evening School.

## DEGREES GRANTED

In Microwave Tube Operations, **P. Perrine** of Harrison received an MS in Industrial Engineering from Stevens Institute of Technology. **B. Halpern** of Harrison received an MS in Industrial Engineering from Stevens Institute of Technology. Polytechnic Institute of Brooklyn conferred the degree of Doctor in Electrical Engineering on **W. W. Siekanowicz** of Princeton.
—*H. J. Wolkstein*

Also at Harrison, **H. A. Stern,** Chemical and Physical Laboatory, received the MS in Chemistry from Franklin and Marshall College, Lancaster, Pa.

**R. V. Nikiel,** Engineering Leader, Receiving Tube Engineering, Harrison, received the MS in Electrical Engineering from Stevens Institute of Technology.
—*T. M. Cunningham*

**L. Brown** and **D. G. Hymes** of IEP received their MS in Electrical Engineering from the Moore School of Engineering, University of Pennsylvania. **Richard Barton** of the EDP Division received his MS in Electrical Engineering from the Graduate School of Arts and Sciences, University of Pennsylvania.

Drexel Institute of Technology has granted **F. J. Herrmann,** IEP, the MS in Physics.

## ENGINEERS IN NEW POSTS

**S. D. Heller,** Vice President, Government Services, RCA Service Company, has named **G. Denton Clark** Mgr., Missile Test Project, for RCA Service Company Operations at the Air Force Missile Test Center, Patrick AFB, Florida. Mr. Clark announced the appointment of **B. Sweatt, Jr.,** as Mgr., Range Service Engineering, who in turn has named engineering managers in his organization as follows: **R. A. Work,** Mgr., Instrumentation Engineering; **C. F. Feast,** Mgr., Field Service Engineering; and **A. R. Marcy,** Mgr., Engineering Services. Reporting to Mr. Work are **T. C. Knowles,** Mgr., Radar Engineering; **W. A. Price,** Mgr., Optics Engineering; and **R. T. Platt, Jr.,** Mgr., Telemetry Engineering. Reporting to Mr. Feast are **J. A. Meetze,** Mgr., Installation Engineering; **D. E. Wood,** Mgr., Field Installation; and **R. V. Zimmerman, Sr.,** Mgr., Field Engineering. Under the direction of Mr. Marcy are **W. H. Barrows,** Mgr., Range Support Engineer; **J. H. McDonough,** Mgr., Range Service Drafting; and **A. L. Cox,** Mgr., Range Service Control.

The organization of the SC&M Division, reporting to **A. M. Glover,** Vice President and General Manager, is announced as follows: **F. R. Buchanan,** Acting Mgr., Production; **N. H. Green,** Mgr., Consumer Semiconductor Products Dept.; **T. R. Hays,** Mgr., Marketing Dept.; **E. O. Johnson,** Chief Engineer; **R. E. Koehler,** Mgr., Microelectronics Dept.; **K. M. McLaughlin,** Mgr., Computer Products and Materials Dept.; and **A. M. Okun,** Administrator, Product Assurance. Under Mr. Hays, **B. V. Dale** has been named Administrator, Micro-electronics Planning.

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### MM-600 GETS USAF NOD

The RCA MM-600 microwave radio relay system has official approval as a standard USAF type-classification equipment. The MM-600 was developed as an RCA interdivisional project with corporate funds assistance. The 2-kmc version is being supplied to the Canadian National Telegraph in Canada, and the 6-kmc version to the Western Union Telegraph Company.

### AEP BECOMES DEP DIVISION

In an organizational shift effective July 1, 1960, the Astro-Electronics Division, Princeton, N. J. (formerly the Astro-Electronic Products Division), became the responsibility of the Communications and Aerospace organization of Defense Electronic Products. **Barton Kreuzer,** recently named Division Vice President and General Manager of AEP, will report to **Walter G. Bain,** Vice President and General Manager, Communications and Aerospace, DEP. **Sidney Sternberg** continues as Chief Engineer.

### SERVICE CO. ADDS TELETYPE

Teletypewriter service and maintenance has been added to the growing list of activities performed by the Technical Products Department of the RCA Service Company. The program was inaugurated in the Atlanta service area in Sept., 1959, and has progressed through the Boston and Pittsburgh service areas. By the end of 1960, this activity will be extended to Chicago, Dallas, and San Francisco service areas.

### HILLS RECEIVES AWARD

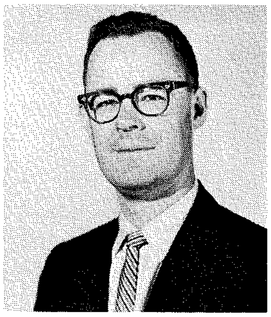
**Ernest J. Hills,** Methods and Processes Laboratory, Power Tube Engineering, Lancaster, recently received a $500 "Usership" award from *Steel* Magazine for his contest entry describing the possible use of a super-alloy spring wire.

## REGISTERED PROFESSIONAL ENGINEERS

| | | |
|---|---|---|
| **William Stonaker,** | Electron Tube Division | Prof. Eng., A-10092, N. J. |
| **R. M. Kurzrok,** | Surf. Comm. Systems Lab | Prof. Eng., 36940 P. E., N. Y. |
| **A. D. Zappacosta,** | Surf., Comm. Engineering | Prof. Eng., 11206, N. J. |

### GUNTER NAMED ED REP FOR ME & CD; DOBSON FOR SYSTEMS SUPPORT, ASD; ALSO IEP ED REPS REALIGNED

Two new RCA Engineer Editorial Representatives have been named in DEP locations. In the Missile Electronics and Controls Division, Burlington, Mass., **Dr. R. C. Gunter** has replaced **R. W. Jevon.** The Editors would like to extend their warmest thanks to Bob Jevon for his excellent work, and congratulate him on his new post as Mgr., Preliminary Design, in the Systems Engineering activities of ME & CE. Information on Dr. Gunter will be presented in the next issue of the RCA Engineer. In the Airborne Systems Division, Camden, the naming of D. B. Dobson for Systems Support Engineering marks the addition of a new Ed Rep spot for RCA Engineer activities. Both Gunter and Dobson will serve on **F. W. Whitmore's** DEP Editorial Board.



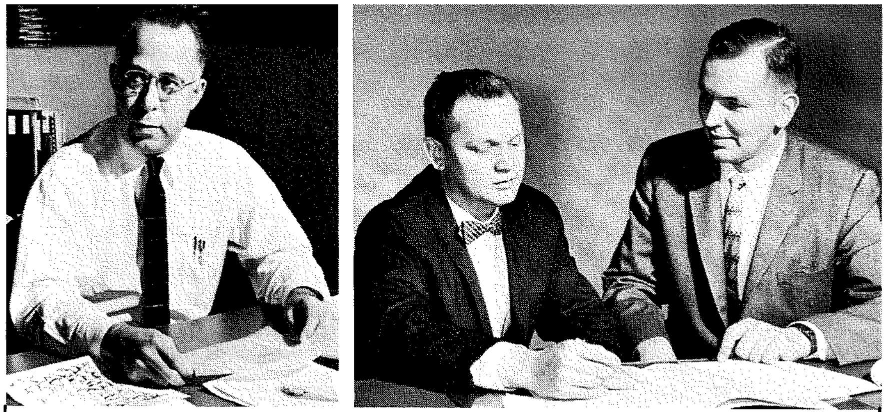
David B. Dobson

In IEP, **S. F. Dierk** has adjusted Ed Rep responsibilities to better match organizational structure. (See page opposite.)

**David B. Dobson** received his BEE from Rensselaer Polytechnic Institute, Troy, N. Y. He was employed by the Signal Corps Engineering Laboratories at Ft. Monmouth, N. J., working on high-powered audio amplifying systems. Subsequently, he was appointed the Electronics Member of the Army Psychological Warfare Board. Upon joining RCA in 1954 he was engaged in the production follow of the E-4, E-6, and MA-7 Fire Control Radars. In 1955 he was assigned as the Project Engineer for the Maintenance Engineering aspects of the MA-10 (AN/ASG-14). During 1957 he participated in the Maintenance planning for the ASTRA Weapon System and in 1958 was appointed an Engineering Leader with responsibility for the ASTRA Production and Depot Test Equipment Design and Development. With the formation of Systems Support Engineering in 1959, he became Projects Leader engaged in Application Engineering of RCA's Systems Support Products to new areas, and Engineering Support of the Marketing Activity. He is a Member of the IRE, the AIEE, the Society of American Military Engineers and has presented and published papers on all aspects of Systems Support.

### LYONS AN AUTHOR FOR SCIENCE ENCYCLOPEDIA

**Walter Lyons,** Mgr., Station Facilities (Equipment and System Design) for RCA Communications, Inc., has been chosen to prepare a 7500-word survey article on radio receivers and transmitters for the forthcoming McGraw-Hill *Encyclopedia of Science and Technology,* to be published this Fall in 15 volumes.



Left to right: Hobart Tipton, Frank Kay and Dave Carlson

### CREDIT TO ED-REP PLANNERS TIPTON, et al FOR THIS ISSUE

This excellent issue is the result of fine planning and coordination by Hobart Tipton, Ed Rep for SC & M, who had plenty of help in producing the tunnel-diode articles on theory and application from the following: Ed Dickey, RCA Laboratories; Charley Meyer and Eleanor McElwee, Electron Tube Division, Harrison; Dave Carlson, RCA Victor Home Instruments; and Sig Dierk of IEP, Camden. To round out the contents for this fine issue, Frank Whitmore and Wes Whittier in DEP, Lou Thomas in AEP, and Frank Kay in RCA Victor Home Instruments all helped out on papers of professional stature.

The Editors gratefully acknowledge this team work—typical of the kind of performance by Editorial Representatives that results in really fine issues.

## MEETINGS, COURSES AND SEMINARS

### ENGINEERING AS A CAREER

**Jules M. Forman,** Engineering Leader, Life Test and Data, Electron Tube Division, Lancaster, gave a talk on "Engineering as a Career" before parents and students at the Reynolds Junior High School, Lancaster, and to seniors at the York, Pa., William Penn Senior High School. —*H. Lovatt*

### COMMUNICATIONS CONFERENCE

A Communications Conference was held June 20-27 in New York City for European customers of communications networks, under the auspices of RCA Communications, Inc. The various presentations covered past, present, and future data communications systems and methods. —*S. Dierk*

### COMPUTER GROWTH

During the recent Western Joint Computer Conference (May 3-5), **J. Wesley Leas,** Chief Engineer, EDP Division, IEP, served as a panelist on "Computer Organization Trends." This session dealt with the trend in computer development and growth of that industry in the future. —*S. Dierk*

### DATA-HANDLING CONFERENCE

Nearly 100 key representatives of government and military agencies attended RCA's Data Handling and Display Conference held at Van Nuys, Calif. on July 12 and 13. The symposium featured talks by **D. B. Holmes,** Mgr., BMEWS, who discussed that program; **O. L. Patterson,** BMEWS Advanced Systems Mgr., who spoke on a new detection system for submarine-launched ballistic missiles, and **J. Lehmann,** of AEP who described the newest NASA weather observation satellite. Also on the agenda were the new RCA 601 EDP system, RCA micromodule programs, and the latest developments in visual display systems.

### ENGINEERING MEETINGS AND CONVENTIONS

**SEPTEMBER 19-22**
*National Symposium on Space Electronics & Telemetry (PGSET), Shoreham Hotel, Washington, D. C.*

**SEPTEMBER 21-22**
*Industrial Electronics Symposium (PGIE: AIEE), Sheraton-Cleveland, Cleveland, Ohio*

**OCTOBER 3-5**
*6th National Communications Symposium (PGCS)—Hotel Utica and Utica Municipal Auditorium, Utica, New York*

**OCTOBER 4-6**
*6th Conf. on Radio Interference Reduction (PGRFI: Armour Research Foundation)—Chicago, Ill.*

**OCTOBER 10-12**
*National Electronics Conference (IRE:AIEE:EIA:SMPTE), Hotel Sherman, Chicago, Ill.*

**OCTOBER 13-14**
*1960 National Symposium on Engineering Writing and Speech (PGEWS), Bismark Hotel, Chicago, Ill.*

**OCTOBER 17-21**
*88th Convention, Society of Motion Picture & Television Engineers, Sheraton-Park, Washington, D. C.*

**OCTOBER 19-21**
*Symposium on Space Navigation (PGSET), Deshler-Hilton Hotel and Civic Center, Columbus, Ohio*

**OCTOBER 20-22**
*Acoustical Society of America, San Francisco, Calif.*

**OCTOBER 24-26**
*7th East Coast Aero and Navigation Electronics Conference (PGANE, Baltimore Section), Lord Baltimore Hotel, Baltimore, Md.*

**OCTOBER 25-27**
*American Standards Association, 11th National Conference on Standards, Sheraton-McAlpin Hotel, New York*

**OCTOBER 27-29**
*1960 Electron Devices Meeting (PGED), Shoreham Hotel, Washington, D. C.*

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