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FIELD-EFFECT TRANSISTORS

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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 from 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 electron-valve 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_p, 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-drain circuits are useful as high input impedance amplifiers. The grounded-source circuit has the property that the transconductance, g_m, depends upon dc 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 volt-

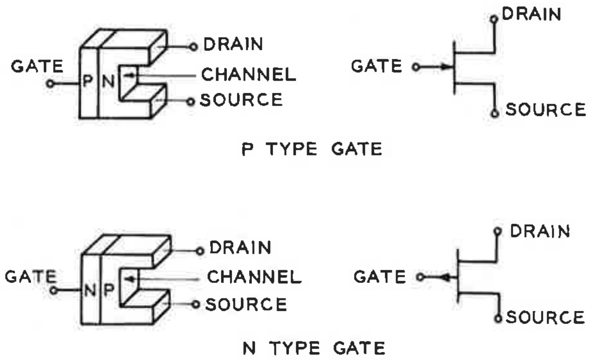

Fig. 1—Construction of field-effect transistors.

age. 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 valve, the operation is very similar to the cathode follower. An important

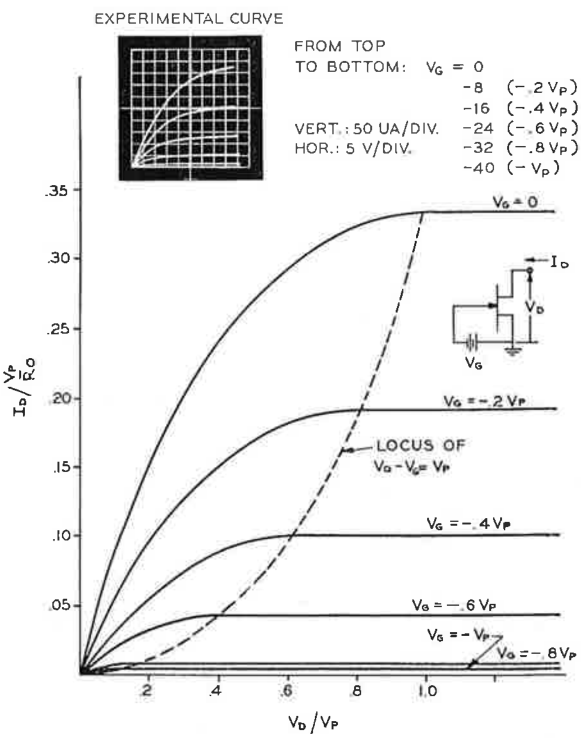

Fig. 2—Normalized drain characteristics of field-effect transistors.

property of this circuit is that the voltage gain is relatively independent of variations in g_m. The per cent-gain variations caused by per cent 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.

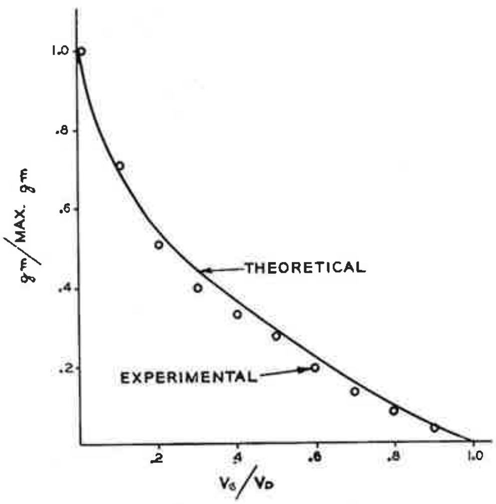

Fig. 3—Transconductance as a function of gate voltage.

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 non-linear 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

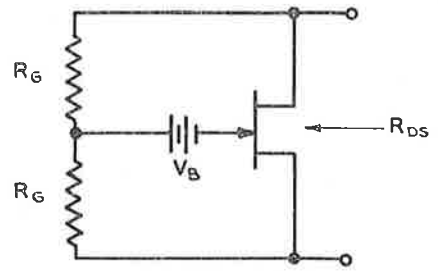

Fig. 4—Symmetrical gate-bias circuit.

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 by an impressed drain voltage does not change the width of the depletion region at the centre 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 centre-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 centre-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

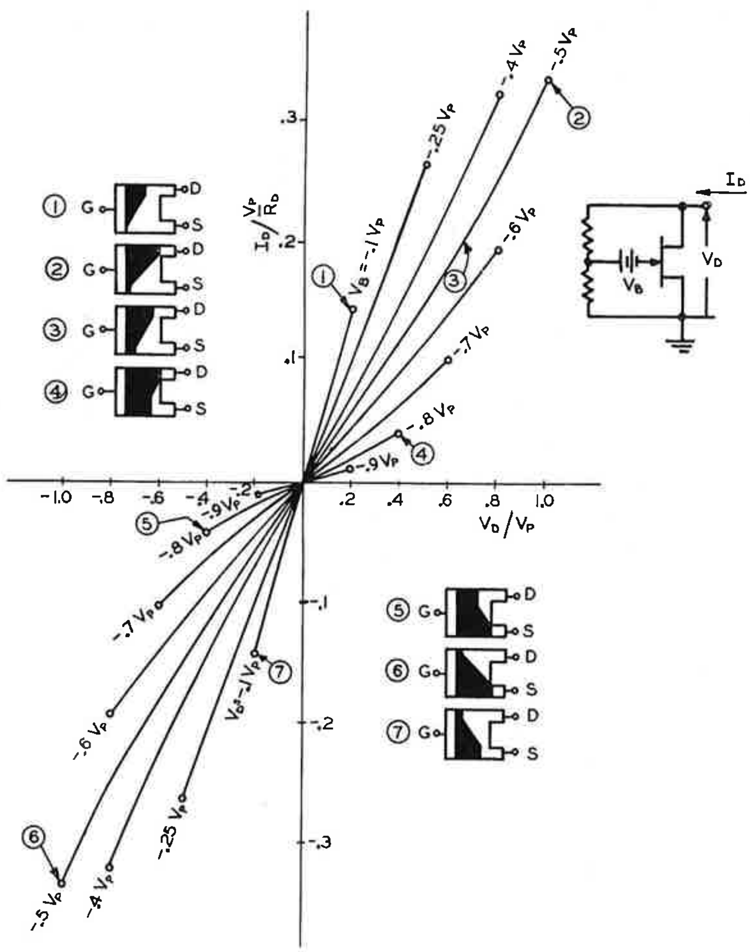

Fig. 5—Drain characteristics for centre-tapped bias arrangement.

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.

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 centre-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 dc to several megacycles

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

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

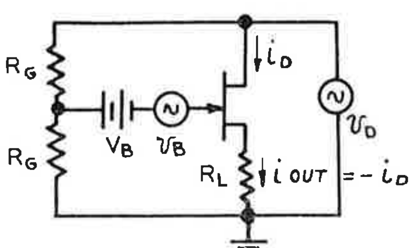

Fig. 7—Field-effect transistor analogue multiplier.

ANALOGUE MULTIPLIER

Perhaps the most interesting use of the field-effect transistor as an electronically variable resistance device is analogue multiplication. The function of an analogue 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 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 distor-

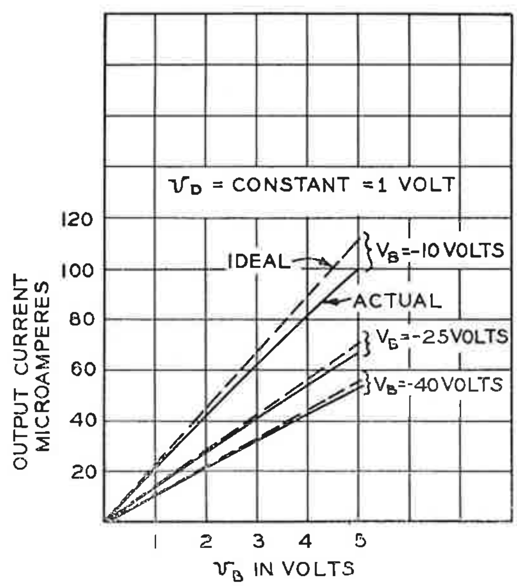

Fig. 8—Multiplier transfer characteristic.

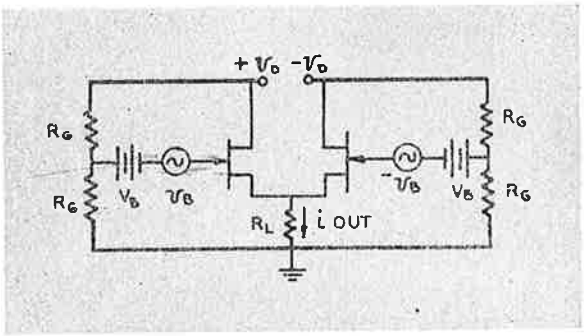

Fig. 9—Dual field-effect transistor multiplier.

tion. 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.

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.

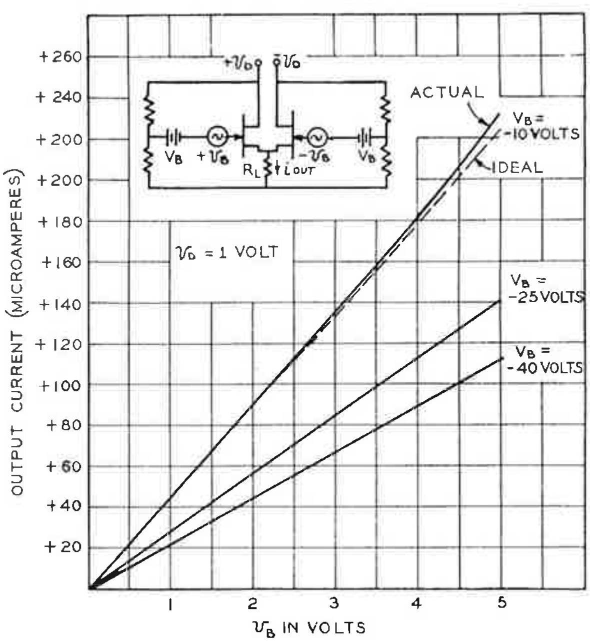

Fig. 10—Dual multiplier transfer characteristic.

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 per cent.

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 per cent.

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 analogue multiplier was the outgrowth of joint work with S. M. Marcus and F. L. Putzrath.

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(With acknowledgements to RCA)

A 10A TRANSISTOR POWER SUPPLY

by R. Walton, AWV Application Laboratory

This unit was made up for a specific purpose in the laboratory, and was not originally intended for publication. It was felt, however, that there may be some interest in the unit, as the basic idea could be used in many ways. The power supply was built to provide a dc output of up to 50 volts at 10 amperes, with a low ripple content.

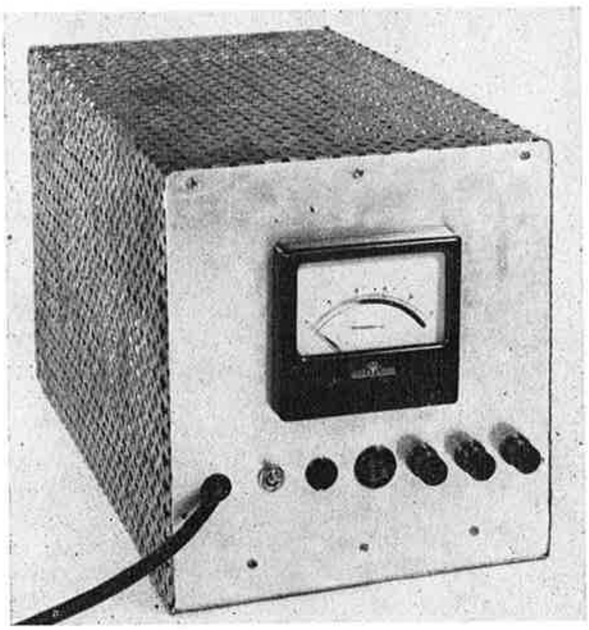
With the present trend of ever-increasing power capabilities in transistors, normal power sources in a laboratory or workshop are often strained to the limit of their capabilities. In some recent work on a high-power (200 watt) inverter, it was found that normal automobile batteries were inadequate as a primary power source. This led to a search for an alternative supply, and the unit described here was the result.

The inverter on which work was being done was designed for operation on 32-volt generating plants, in which voltage variations can be considerable. This fact must be allowed for in designing equipment intended to operate on such a system. An inverter of this power rating on a 32-volt supply can draw a current of the order of 8 amperes. To cover all expected needs, it was decided to make 50 volts at 10 amperes the goal for the power supply.

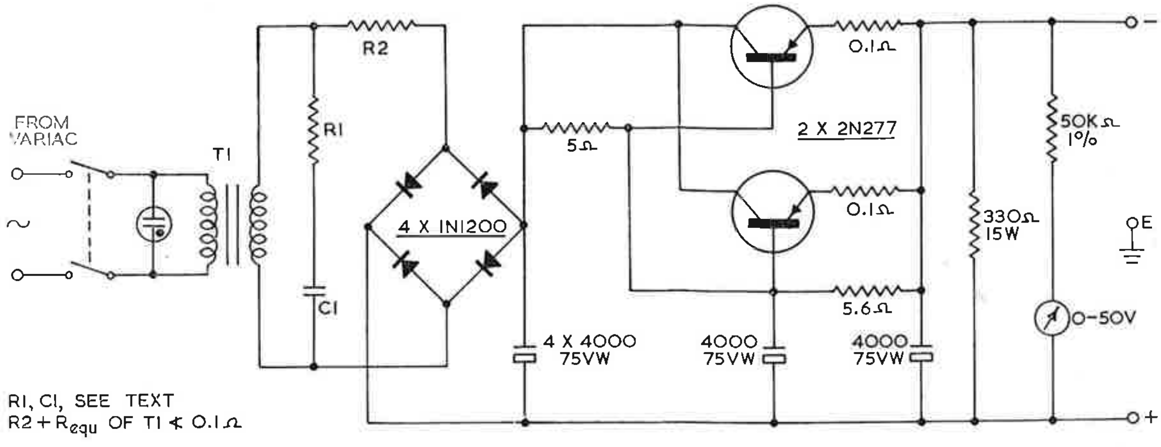
At this order of current, the easiest method of voltage variation in any power supply is to vary the ac input to the unit by means of a "Variac" or equivalent device. The high currents also necessitate special circuits for the rectifiers and filters.

A bridge circuit for the rectifier requires the lowest peak inverse voltage ratings for the rectifying elements, and also makes for the cheapest transformer. Calculations based on "Radiotron Designers' Handbook," Chapter 30 Sect. 2 indicate a transformer with a secondary rating of 52 volts at 12.5 amperes, and an equivalent resistance of the order of 0.1 ohm. This is within the rating for a type 1N1200 silicon diode, and four of these were therefore chosen as the rectifying elements.

The circuit shows four 4000 microfarad 75-volt working electrolytic capacitors as the input to the filter. The total capacitance of 16,000 microfarad will result in approximately 2.5 volts ripple at 50 volts, 10 amperes output. The next step in filtering calls for a "dynamic" filter, which has already been mentioned in the pages of this magazine. The arrangement shown gives an effective minimum capacitance of 140,000 micro-



Front view of the 10 A transistor power supply.



Circuit diagram of the 10 A power supply.

farads, that is, the product of the minimum current gain in the transistor and the capacitance in the base circuit.

The next problem was to dissipate the heat generated in the series transistors used as the "dynamic" filter elements. A pair of type 2N277 power transistors, each mounted on a modified Telecomponents type 7003 heat sink will dissipate about 50 watts up to 45°C ambient temperature. To ensure that the two power transistors share the current equally, an 0.1-ohm resistor has been connected in series with each emitter. The tolerance of these balancing resistors is not important, but they must be matched within 5%.

The "dynamic" filter, as indicated in previous articles, has a fairly high ac impedance above 5 Kc, which explains the use of the 4000 microfarad electrolytic capacitor across the output terminals. A bleed resistor is necessary to discharge the capacitor when a light load is used, and when the output voltage is reduced.

The power supply is provided with an 0-50 volt meter, consisting of a basic movement and a multiplier resistor. The final scale for the meter had not yet been fitted when the accompanying photographs were taken.

The series-connected resistor and capacitor across the secondary of the power transformer must be selected to suit the particular transformer used. Reference should be made to G.E.C. Application Report No. 32. Capacitor C = K × 2.2 microfarad, where

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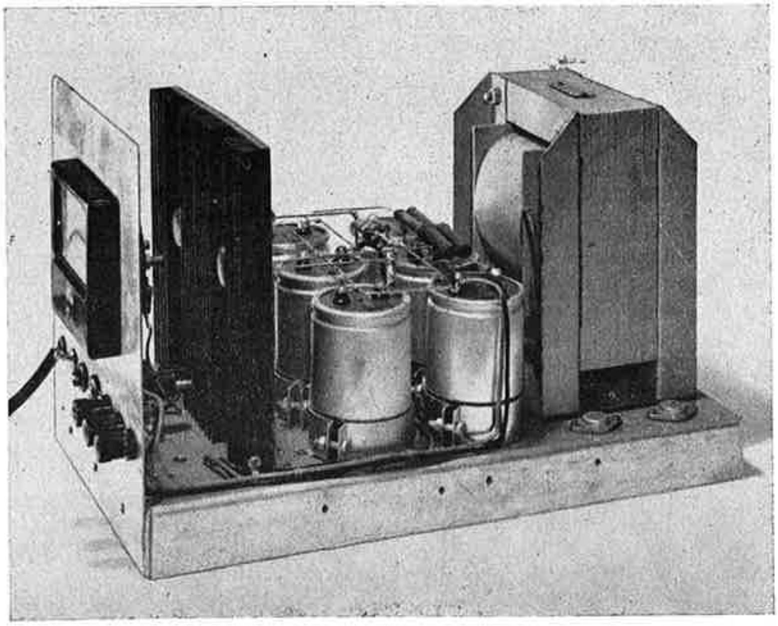
Interior arrangement of the power supply.



FLUORESCENT SCREENS

The screen materials used in all cathode ray tubes, picture tubes, kinescopes and similar devices consist of formulations of phosphors, which are designed to exhibit fluorescence and/or phosphorescence when the screen is activated by a beam of electrons. Different formulations are used according to the colour required, the persistence called for, and so on.

The type of phosphor used in any tube is, under the JEDEC (Joint Electron Device Engineering Council) System, incorporated into the type number of the tube. For example, a TV picture tube type 23CBP4 has a number 4 phosphor, indicated by the "P4" at the end of the type number. P4 is a phosphor exhibiting white fluorescence, and is therefore used for black-and-white TV picture tubes. Phosphor P11, on the other hand, has a high-intensity actinic blue fluorescence, and is therefore a good choice for photographic work.

The more common phosphors in use are described below, whilst the accompanying diagram charts the colour coordinates for them.

Phosphor P1 produces a brilliant spot having yellowish-green fluorescence and medium persistence. Types having this phosphor are particularly useful for general oscillographic applications in which recurrent wave phenomena are to be observed visually.

Phosphor P2 is a medium-short persistence screen which exhibits yellowish-green fluorescence and phosphorescence. The phosphorescence may have useful persistence for over a minute under conditions of adequate excitation and low-ambient illumination. Types utilizing this phosphor are particularly useful for observing either low- or medium-speed non-recurring phenomena.

Phosphor P4-Sulphide Type is a highly efficient screen having white fluorescence and medium-short persistence. Types having this phosphor are of particular interest for television picture tubes.

Phosphor P4-Silicate Type exhibits white fluorescence and has medium to medium-short persistence. Types having this phosphor are of particular interest for projection-type kinescopes.

Phosphor P4-Silicate-Sulphide Type exhibits white fluorescence and has medium to medium-short persistence. Types having this phosphor are of particular interest for projection-type kinescopes.

Phosphor P7 is a long-persistence, cascade (two-layer) screen. During excitation by the electron beam, this phosphor produces a white fluorescence. After excitation, the screen exhibits a yellowish-green phosphorescence which persists for several minutes. Types having this phosphor are particularly useful where either extremely low-speed recurrent phenomena or medium-speed non-recurrent phonomena are to be observed.

Phosphor P11 emits high intensity actinic blue fluorescence and has medium-short persistence to permit its use in all photographic applications except those in which film moves at high speed. P11 screens, because of their unusually high brightness characteristic, may also be used for visual observation of phenomena.

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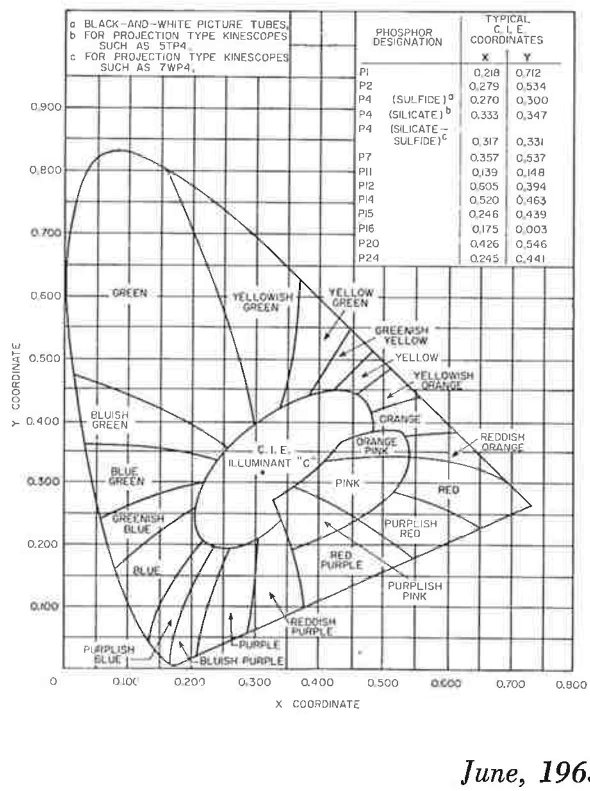

PHOSPHOR DESIGNATION		TYPICAL C. I. E. COORDINATES X	Y
P1		0.218	0.712
P2		0.279	0.534
P4	(SULFIDE)[a]	0.270	0.300
P4	(SILICATE)[b]	0.333	0.347
P4	(SILICATE—SULFIDE)[c]	0.317	0.331
P7		0.357	0.537
P11		0.139	0.148
P12		0.605	0.394
P14		0.520	0.463
P15		0.246	0.439
P16		0.175	0.003
P20		0.426	0.546
P24		0.245	0.441

CURRENT MICROWAVE RESEARCH AT RCA LABORATORIES

DR. S. BLOOM

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This paper describes a part of the work being done in the Microwave Research Laboratory, RCA Laboratories, Princeton. Several devices and fields of investigation are discussed as being representative of the main purpose of this Laboratory, to devise new means of generating greater power and of amplifying with less noise, at increasingly higher frequencies. The particular examples are drawn from three broad areas of research: solid state devices, electron beam tubes, and plasmas.

Because there is much overlap in the research to be described herein on solid-state devices, electron beam tubes, and plasmas, the division of this paper into these categories is somewhat arbitrary. For example, both solid-state devices and beam tubes are used as low-noise amplifiers; solids, tubes, and plasmas are all in the race to reach the millimetre wavelength region of the microwave spectrum; some of the plasma work deals with electron-hole plasmas in solids and some with electron beams shot into a plasma; some of the cathode work has application to low noise tubes, to plasma synthesis and to high power tubes—and so forth.

No mention will be made of masers or lasers. These quantum electronic devices go a long way toward answering the needs set forth in the abstract, above; however, they lie in another bailiwick and much information is available in the literature about them. Nor is there room here to discuss any of the other work in other laboratories at Princeton that relate to microwaves. For example, thin-film and tunnel-effect studies, photo-emission, ferroelectricity, superconductivity, acoustic waves and cyclotron resonance in solids, are all given short shrift even though it is such studies that lead to tomorrow's microwave devices.

SOLID-STATE MICROWAVE DEVICES

The field of solid-state microwave *active* devices is relatively new. Its impetus came from the discovery of low-noise microwave amplification via parametric diodes, and from the later discovery of intrinsic negative resistance in the tunnel diode. Much of the pioneering work on these two devices was done at RCA Laboratories. Today, the main emphasis is on solving certain difficult circuit problems, on extending the amplifiers to higher frequencies, and on realizing efficient solid-state millimetre-wave generators.

Although tunnel-diode amplifiers have been around now for several years and progressively lower noise factors have resulted from improvements in materials and fabrication, these devices have not yet taken their promised place in practical systems. The reason is that they suffer from a drawback indigenous to negative-resistance devices — *instability*. These are two-terminal affairs, and so variations in load and input impedances affect their gain. The noise factor also suffers from reflections from the load. Although a long-recognized solution to this problem has been to use nonreciprocal elements to isolate the amplifier from its terminations, little had been done in way of realization. To this end, studies are underway at this laboratory on a *travelling-wave tunnel-diode amplifier* using an L-band coaxial transmission line containing a distributed ferrite isolator.[1]

The amplifier (Fig. 1a) is in its early stages and does not yet incorporate the ideal situation wherein the nonreciprocal material is integrated with a distributed negative resistance. The four to six germanium tunnel diodes (obtained from RCA Semiconductor & Materials Division, Somerville) are part of the inner conductor, the separation between diodes being small compared to a wavelength. The ferrite cylinder provides 15-db reverse loss and only a few tenths of a db forward loss. Measurements, made at L-band for convenience, show the device to have a 15-db net gain. The bandwidth is 5 to 10 per cent, which is less than the bandwidth of the ferromagnetic resonance and so leaves room for further improvement. Both open- and short-circuit stability have been achieved. The noise factor, yet to be measured, is due to shot-noise current and so is proportional to the dc current and to the magnitude of the negative resistance. This speaks for getting better materials, perhaps gallium antimonide. Such diodes, in non-travelling-wave amplifiers, have given 2.5-to-3.0-db noise factors. In the travelling-wave version, the noise factors should be no worse and may well be better, since the ratio of negative resistance to input resistance, per unit length—which determines the noise factor—can be low, with the resulting small gain per unit length being made up by the use of a longer structure. With stability, which has been achieved, and with noise factors in the 2.5-to-3.0-db range (which appear likely) this dc "pumped" solid-state amplifier would be competitive with travelling-wave tubes in their common range of frequencies.

About two years ago, work in this laboratory resulted in a tunnel-diode down converter with unique properties.[2] This device was stable, had low noise factors, and exhibited conversion gain rather than loss. The gain resulted from the usual negative-resistance portion of the diode's *I-V* curve, the mixing resulted from the nonlinear portion of the *I-V* curve near the peak-current point, and the low noise factor was due to the exclusion of if noise by virtue of the conversion gain. Subsequent experiments showed that noise factors as low as 2 to 3 db could be achieved when the local oscillator signal was strong enough. This work was done at uhf; its success stimulated the attempts now in progress to realize a *millimetre-wave tunnel-diode down converter*.[3] The hope is that this converter, in contrast to present millimetre-wave detectors, will have a low noise factor.

This work is proceeding in two stages. First, tunnel diodes—in this case gallium antimonide with zinc dots—are being checked for their non-linearities by being operated as detectors in the positive-slope region of the *I-V* curve (Fig. 1b). The non-linearity in this region is greater than that obtainable from a conventional crystal diode; also, the losses are an order of magnitude less than in a conventional diode because the tunnel diode is heavily doped. A 1-kc modulation on a 55-Gc carrier was detected with a sensitivity 15 db greater than is obtainable at that frequency with a conventional 1N53 crystal diode. This is in agreement with calculation. The second stage of this study will be an investigation of the possibility of conversion with gain. In this case, the diode will still be biased in the positive slope part of the *I-V* curve, but the local-oscillator rf voltage will be strong enough to cause the total voltage to swing over into the negative-resistance part of the curve during part of the cycle.

During the course of this work, it was found that the tunnel diode at ultra-microwave frequencies exhibits nonlinear capacitor effects. In other words, at high enough frequencies the tunnel diode is mainly reactive and its parametric, or varactor, aspects became important. Furthermore, because the diode is heavily doped its series resistance is low. Thus, the tunnel diode, when acting as a varactor diode, offers promise of being useful for efficient harmonic generation of millimetre-wave power. Preliminary calculations indicate that generation of 1.2-mm waves as a fourth harmonic may be possible with as much as 10 per cent efficiency.

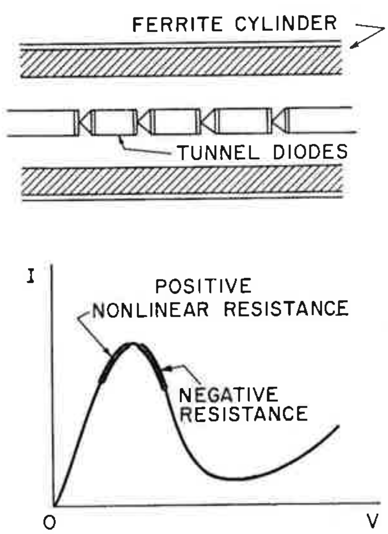

Fig. 1—(a) Travelling-wave tunnel diode amplifier uses diodes along the inner conductor of a coaxial transmission line, and uses a ferrite cylinder to achieve isolation. (b) The current-voltage characteristic of a tunnel diode has a region of positive nonlinear resistance useful for mixing, and a region of negative resistance for amplification.

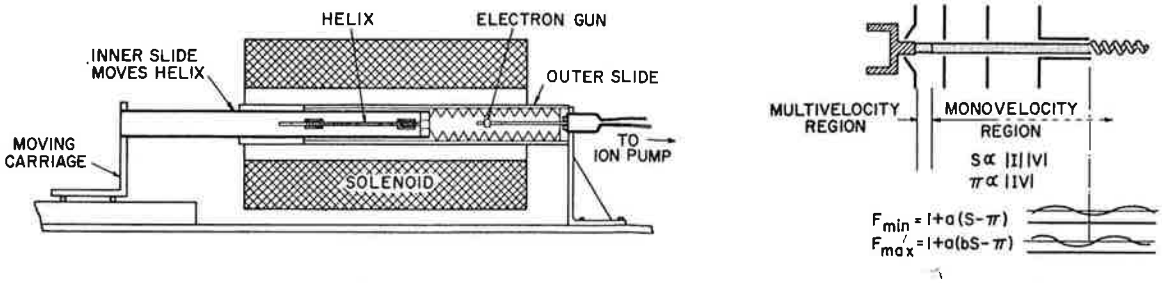

Fig. 2—The gun-to-helix separation in the "SπF apparatus" is adjustable by means of bellows. By varying this separation and measuring the resulting minimum and maximum noise factor of this travelling-wave tube, one can solve the theoretical equations to obtain the important noise parameters S and π; the constants a and b are fixed by the parameters of the tube and are known.

ELECTRON BEAM TUBES

Work on electron beam tubes represents a large fraction of the total effort in the Microwave Research Laboratory. This may seem somewhat contradictory in view of the fiat stated in the abstract which calls for "new" devices. The travelling-wave tube goes back to the late 1940's and is itself an off-shoot of the klystron, which predates it by another ten years.

Yet both these tubes, in one or more of their many forms, are still being researched. There are many reasons: Travelling-wave-tube amplifiers today offer a hard-to-beat combination of enormous bandwidth, high and stable gain, and ultra-low noise factors. There are cogent reasons for believing that further research will lead to further noise reduction. Also, the travelling-wave principle itself—i.e., the interaction mechanism resulting from the coupling between a beam and a circuit travelling-wave—has led to a bewildering array of beam-type devices whose capabilities have not yet been fully explored. Furthermore, the millimetre, and submillimetre-wave portion of the spectrum is a constant challenge; although beam tubes are certainly not easy to build at ultra-microwave frequencies, the fact remains that a beam tube—the backward-wave oscillator, a variant of the TWT—has so far made the most successful breach of the millimetre-wave barrier. There is still room for innovations, in terms of new types of circuits and new coupling schemes. Finally, beam tubes are interesting because electron beams are interesting. Beams exhibit many disconcerting phenomena when run under conditions that are closer to the real world than to the simplified models of the theorist. As a case in point, a new wave-damping effect occurring on beams that are "hotter" than usual has recently been observed and explained in this laboratory. This damping effect has important consequences, not only for low-noise tubes and power tubes, but also for plasma devices.

It was mentioned that today's travelling-wave tubes give excellent noise factors. This is true. However, it is not fully known why they are so good and whether or how they can be made still better. The noise factor of the TWT is determined by noise processes occurring in the beam very close to the cathode. An electron beam, like any noisy four-pole, contains two noise sources: a noise-current source due to shot noise, and a noise-voltage source due to thermal energy. In the jargon of the field, S measures the magnitude of these two sources, and π measures their mutual correlation. The tube noise factor F is proportional to the difference S-π; hence, one wants to design a gun which minimizes S and maximizes π. Also, one would like to know how S and π vary as one changes the beam geometry, current density distribution, and beam voltage near the cathode. Because the cathode region is difficult to probe directly, its properties must be inferred. To this purpose a research tool, known locally as the *SπF apparatus,* was conceived and developed[4] at this laboratory and is being used to measure S and π values for various low-noise TWT guns. The apparatus (Fig. 2) is essentially a TWT with adjustable cathode-to-helix separation. Values of S and π are inferred from measurements of the minimum and maximum noise factors that result when the helix position is varied. The technique differs from others in that it uses an electronically simple apparatus in which excellent vacuum conditions can be maintained. As a consequence, this is the first instrument of its type which has yielded S, π data on guns giving ultra-low noise factors. The measurements show that although the lowest noise factors occur when the gun-electrode voltages are such as to minimize S, the correlation π is essentially unchanged by this optimization. If follow-up measurements now in progress bear out the preliminary findings, serious doubts may be raised about the sacrosanct linearized theories of noise in the multivelocity region near the cathode.

It was mentioned that the noise factor of a travelling-wave tube—or, for that matter, of any beam-type amplifier—is proportional to S-π. The constant of proportionality contains the cathode temperature as a factor. To achieve a desired low temperature, work is in progress here on a novel *cesium hollow cathode*[5] (Fig. 3). Cesium vapour is introduced into a heated chamber made of tungsten or nickel. The atoms that adhere to the hot walls lower the metal's work function; this permits electron emission to occur at lower-than-usual temperatures. The electron emission per unit area is small, but the total emissive area is large. Furthermore, the few ions produced at the chamber walls effectively prevent the formation of a potential minimum within the chamber; this means that the electrons which are extracted all come from regions having essentially equal space potential. Hence, their low emission-temperature is not vitiated by an additional velocity spread. This is in contrast to previous hollow cathodes, which used no ions and, therefore, gave beam temperatures much higher than the wall temperature. The electrons are extracted through a small hole in the chamber. In this way beam currents adequate for TWT operation have been obtained at temperatures of about 700 to 750°K. This is in contrast to the 900°K beams of conventional TWT oxide cathodes. The next step will be to capitalize on this low temperature in an actual low noise tube.

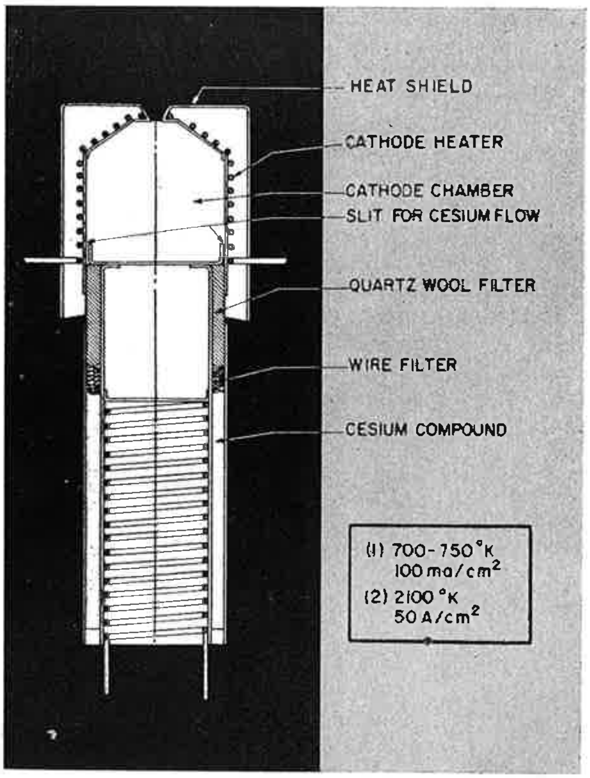

Fig. 3—Cesium vapour, evolved from the heated cesium chromate powder, is filtered before entering the cathode chamber through a fine slit. The hot chamber wall both emits electrons and ionizes the impinging cesium atoms.

The cesium hollow cathode has another important use. If the chamber walls are hotter than when used for low-noise purposes—say about 2000°K—no cesium adheres to the walls and the walls emit electrons by unassisted thermionic emission. If, furthermore, the cesium vapour pressure is relatively high, very many ions are produced at the hot walls. The resultant ion-electron plasma reduces the space charge in the chamber. This reduction alleviates the old problem of hollow cathodes—space-charge limitation—and so allows one to extract beams of very high current density through the aperture. Thus, this potent cathode has important applications in microwave superpower tubes. To date, 100-mil-diameter beams of 50 amps/cm^2 have been extracted. The efforts now are focused on raising this current density to 100 amps/cm^2 and to tame the attendant problems of cesium-vapour trapping and recirculation.

Another area of research, alluded to above, concerns the *spatial decay of space charge waves* along slowly drifting beams. Electrons in a neutralized cloud can oscillate, if disturbed, with a characteristic frequency called the plasma frequency. If the cloud moves, as in a directed electron beam, the Doppler effect produces two waves, one having a phase velocity slightly greater than the beam's drift velocity, the other slightly slower than the drift velocity. These are the "fast" and "slow" space charge waves. Their interference results in a standing wave along the beam. For example, if a beam is velocity-modulated by a klystron buncher cavity, then the rf current along the beam exhibits a standing sinusoidal pattern, the klystron catcher cavity then being put at the first current maximum. This is the "normal" situation, all of which is well known to designers of conventional beam tubes. However, recent experiments done here show that this simple sinusoidal pattern changes into a decaying sinusoidal when the beam velocity is sufficiently low. The experimental arrangement is sketched in Fig. 4, together with representative results. The wave decay is attributable to multi-velocity effects. When the electrons travel with a wide spread of velocities, the Doppler shift mentioned above is different for each velocity class; hence, instead of one fast space-charge wave, there are many. Each travels with a different phase velocity, and so they get out of phase with each other as they move downstream. Since the electron velocities are random, so are the phases; the upshot is a complete phase-mixing or decay of all waves except the single slow space charge wave. But this surviving slow wave has no fast-wave counterpart to beat against; hence, the power output detector sees a constant signal level. These effects occur at low beam velocities, since then the velocity spread becomes proportionately greater. In the experiments of Fig. 4, the velocity spread is due to electrostatic lens effects caused by the beam moving from, say, a 500-volt input

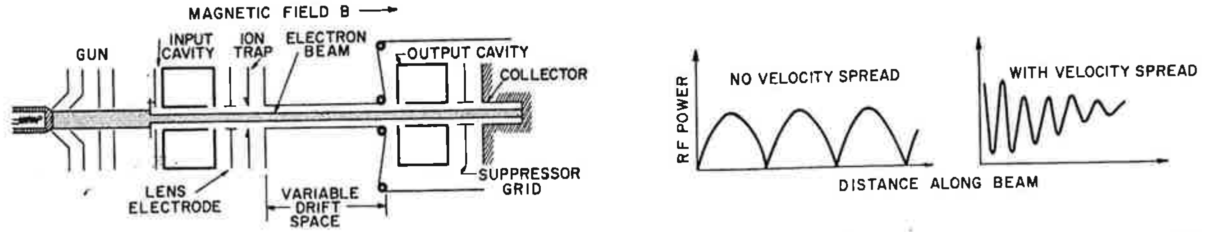

Fig. 4—The beam, after being velocity modulated by the input cavity, travels through a variable-length drift region. The standing wave pattern is detected with the output cavity. When the cavities and drift space are at equal voltages the pattern is the usual sinusoid. But when the drift space is at a low voltage so that lens effects produce a velocity spread, the pattern is a decaying sinusoid with finite minima.

cavity into a 50-volt drift space. Velocity spread can also arise from the space-charge depression of potential occurring in the core of very dense beams. In short, then, multivelocity effects can cause an undesirable decay of signal modulation and so must be more fully understood and controlled for proper design of beam-type power tubes.

Velocity spread also plays a role in low-noise tubes. In this case, the spread is due to the temperature of the cathode; the beam temperature is small (about 1000°K) but the drift voltage of the beam near the cathode is also small (about 1 volt). Hence the *relative* velocity spread is just as large as it is in the 50-to-100-volt beams of the preceding paragraph with their 10^5 °K equivalent "temperatures." The multivelocity effects near the cathode of a low-noise travelling-wave tube determine the S and π values mentioned earlier, and so determine the tube noise factor. Because the equations describing the noise waves in this cathode region are nasty, theorists for many years have had to settle for machine solutions—expensive and physically unrevealing. This situation has been remedied by a recent investigation at this laboratory. Closed-form, analytical solutions have been obtained for the multivelocity noise equations;[7] special cases of these solutions also describe the aforementioned wave-decay phenomena on signal-excited beams. It is hoped that these new results will teach us how to *design* the cathode region of a TWT for optimal noise reduction.

Although the low-noise-tube art is fairly well in hand at low microwave frequencies (say, 1 to 10 Gc)—with current efforts being devoted to obtaining greater consistency and fractions-of-a db improvement of noise factor—the situation at ultra-microwaves is less happy. This goes for *any* tube, low noise or not. Conventional tubes employ beams interacting with slow-wave circuits, and circuits become progressively more difficult to make as the frequency goes up. At millimetre wavelengths, the circuit (which for example may be a helix or ridged waveguide) must have minute transverse dimensions and must be virtually lossless. Also, since the fields decrease rapidly away from the circuit, the high-current-density beam must be kept close to the circuit. This imposes severe requirements on the beam-focusing system, lest the beam hit the surrounding circuit and produce an expensive puddle. These are formid-

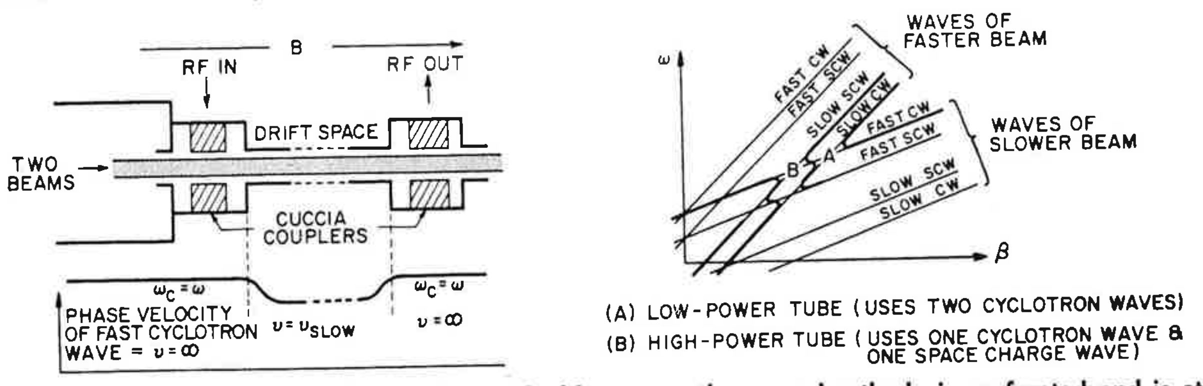

Fig. 5—Two cathodes are used to produce a double-stream; the second cathode is perforated and is at positive potential with respect to the first. The axial magnetic field in the coupler region is high so that the cyclotron frequency ω can equal the high signal frequency ω. The coupler excites the fast cyclotron wave of the slower beam. However, in the drift (or gain) region this fast wave must be slowed down to get in synchronism with the slow wave of the faster beam. This slowing down is accomplished by reducing the magnetic field.

able problems that tax the ingenuity of the engineer designing the tube and the dexterity of the watchmaker saddled with the job of putting it together. Nevertheless, a high-voltage, high-current beam is a potent source of power and the beam-to-wave mechanism should not be too quickly abandoned. The circuit, the main cause of difficulties, is fortunately not a crucial ingredient. One approach to the problem is to use two beams, one acting as the circuit for the other. In such a two-beam tube, amplification arises from the interaction between the fast wave of the slower beam and the slow wave of the faster beam. These waves may be[8] space-charge waves due to longitudinal oscillations, or they may be cyclotron waves due to the axial drift of transversely orbiting electrons in an axial magnetic field. Or, the waves may be hybrid—space-charge waves on one beam interacting with cyclotron waves on the other beam.[9] To test these ideas, a novel *two-stream cyclotron-wave tube* was conceived and is now being made in this laboratory (Fig. 5). Also shown is the frequency ω versus the wave number β of a two-beam system. Mode A uses two cyclotron waves and is a low-power mode of operation; hence, efforts will be directed toward its low-noise possibilities. Mode B is a hybrid mode and is more suitable for high-power amplification. Although this tube requires no slow-wave circuit to produce amplification, it does need circuit structures for coupling the signal into and out of the beams. However, because the tube uses cyclotron waves, one can use a very simple coupling circuit consisting essentially of two parallel plates (invented some years ago by C. L. Cuccia of The RCA Electron Tube Division for another purpose). The transverse rf field in this coupler changes the size of the orbits of the cyclotroning electrons and, thus, imparts the signal. Furthermore, if the signal frequency is very high, the cyclotron frequency—and, therefore, also the axial magnetic field—must also be high in the coupler regions; however, the magnetic field need not be high in the amplification region. This greatly reduces the need for super solenoids over the entire tube.

The tubes mentioned so far derive their amplification from the conversion of dc kinetic energy of the electrons into rf energy. Thus, these tubes have limited efficiencies under usual operating conditions because the beam slows down and gets out of synchronism with the circuit wave. In contrast to these "0-type" tubes, another variety exists, called "M-type," in which the rf gain comes from the potential energy of the beam electrons. In such tubes (shown in prototype in Fig. 6) mutually perpendicular dc magnetic and electric fields are applied normal to the direction of the beam's dc drift. Because the beam is not slowed down, these *cross-field tubes* have high efficiency. However, when the efficiency is high the gain is low. This is mainly due to two interrelated causes: First, the decrease in potential energy that supplies the gain arises from the beam moving closer to the slow-wave circuit, which means that the beam must start out far from the circuit; but at high frequencies the rf fields are very weak far from the circuit. Secondly, to get respectable amounts of power from these tubes requires the use of large beams. There are three waves on the beam, only one of which is the growing wave. But as the beam is made thicker, the input signal excites a smaller and smaller amount of this growing wave. These considerations evolved from a recent theoretical study[10] that has also led to some possible remedies. These include the use of an auxiliary slow-wave structure as part of the sole at its input end to excite the growing wave in the region where the rf fields are high. Also, by modifying the main slow-wave structure, one should be able to reshuffle the energy in the three waves in such a way as to allocate a larger fraction to the growing wave. A tube to test these proposals is being made.

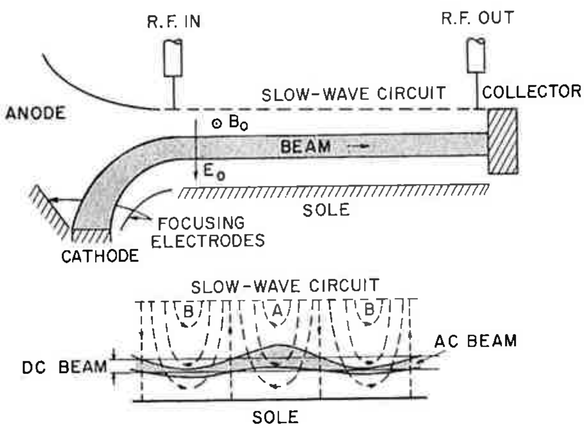

Fig. 6—In the decelerating rf electric field regions (A) the beam moves upward, the electrons doing work against the field and losing potential energy. In the accelerating regions (B) the opposite takes place . . . Also, the fields tend to bunch more electrons into A-regions than into B, thus warping the beam as shown. Because more electrons thus experience a decelerative field than accelerative, there is a net transfer of beam potential energy into field energy and the field grows.

PLASMAS

The reader will recall that during Scott Carpenter's orbital flight on May 24, 1962, there was a period during the capsule's re-entry when contact with the tracking stations was first lost. The cause of the expected blackout—*plasma*. The sheath of electrons and ions that covered the capsule grew so dense that the plasma frequency exceeded the frequency of Carpenter's radio and the sheath became impervious to signals. Furthermore, the plummeting capsule left behind a plasma trail that reflected the radar signals, thus confusing the trackers. This, coupled with the loss of voice communication, made it difficult for the radar trackers to decide whether they were watching Carpenter—or Carpenter as he used to be.

Controlled thermonuclear fusion, with its required fuel temperature in the hundreds of millions of degrees, also presents plasma problems—problems of confinement and of instabilities. The early enthusiasm of workers in the fusion field eventually gave way to a sober realization that these were not "just engineering problems" but, rather, were evidence of a need for a long-haul programme of basic plasma research.

In addition to space technology and controlled fusion, a third stimulus for the heavy, present-day efforts in plasma research comes from microwave electronics. Here, one looks for ways of *using* plasmas rather than of fighting them. For example, particular types of electromagnetic and hydromagnetic instabilities that plague the fusion workers can, when approached with equanimity, lead to useful electronic devices. Cases in point are described below, as examples of some of the plasma topics being investigated in the Microwave Research Laboratory.

Previously mentioned was the useful interaction that results from the passage of an electron beam along an electromagnetic wave. In the ubiquitous travelling-wave tube, this interaction, or "instability," arises from the use of a metallic circuit to support the wave. The double-stream tube was cited as a special case of the TWT in which a second beam was used to replace the circuit. As a special case of the double-stream tube, one can take the drift velocity of the slower beam to zero. Thus one beam can be replaced by a stationary electron cloud or plasma. This is the beam-plasma microwave tube.

Such a tube has one distinct advantage over the conventional double-stream tube, as far as very high frequencies are concerned. The electron densities obtainable in a plasma are much higher than those obtainable in an electron beam. Thus the plasma frequency ($f_p = 9000 \sqrt{N}$ Mc, where N is the density in cm^{-3}) can be higher for the plasma and the beam-plasma tube has a greater chance of operating in the millimetre wave region. An electron density of $10^{15}/cm^3$ is required if the plasma frequency is to be 300 Gc (1-mm wavelength). A plasma having nearly this density has recently been produced here by means of a *cesium discharge*.[11] The novel feature is not the high density but the *combination* of high density and high (20 per cent) degree of ionization. This means fewer scatterers are present to dampen the desired waves. The PIG (Philips ionization gauge) discharge which produces the plasma is shown as part of Fig. 7. Cesium vapour is fed into the metal cylinder, which is at 5 to 20 volts positive with respect to two annular cathodes. The positive voltage accelerates the cathode electrons sufficiently to cause a discharge. The cathode loading is small, since the emitted electrons are needed only to start the avalanche and to replenish whatever electrons are lost by recombination. Fig. 7 also shows a *beam-plasma tube*[12] made in this laboratory, which has oscillated at 24 Gc (12.5-mm wavelength). This tube uses helices for coupling the signal energy in and out of the beam. Eventually, for millimetre-wave work with the $10^{15}/cm^3$-density plasmas, couplers other than helices must be devised. Some novel schemes, using either direct coupling to a varying-density plasma or using optical "interferometer" methods, have been suggested; these will first be tested on the 24-Gc tube.

It was mentioned that plasmas exhibit hydromagnetic instabilities. One such instability occurs in a gaseous-discharge column when a sufficiently high magnetic field is applied in the direction of flow of a sufficiently high current. This instability arises somewhat as follows: If no H-field is present, any initial helically shaped disturbance of a current filament in the discharge soon dies out because of thermal diffusion of charges out of the dense filament. With a magnetic field, there is an additional force $j \times H$ which acts to increase

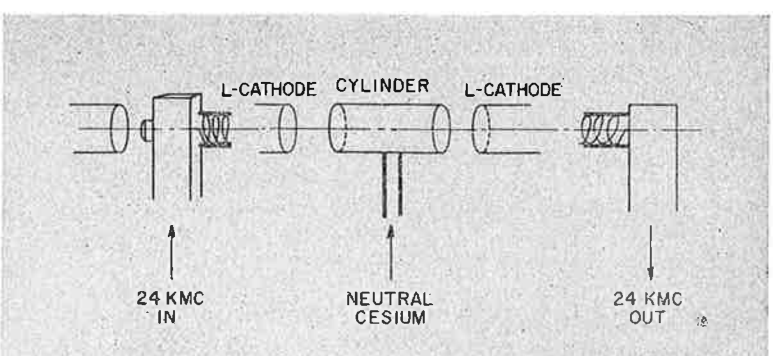

Fig. 7—A highly efficient discharge is initiated between the metal cylinder at positive potential and the electron-emissive L-cathodes. The signal to be amplified is put on the beam with a helix coupler, and removed with another helix coupler. The gain interaction takes place when the slow space-charge wave on the beam couples to the plasma oscillations of the plasma; then both the wave and the oscillations grow with distance.

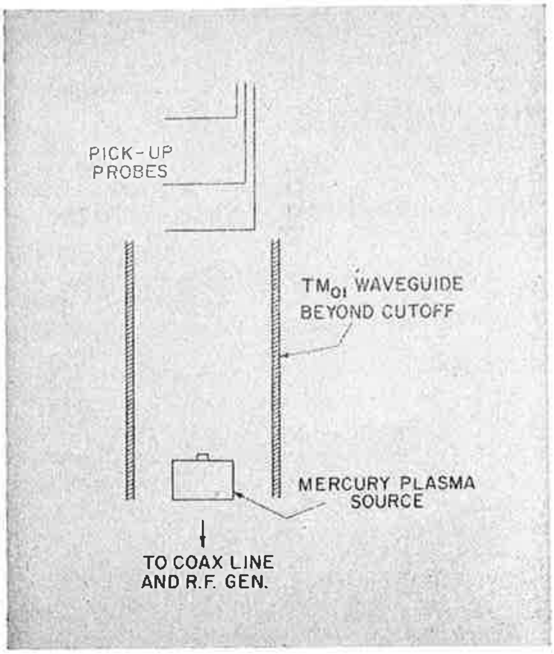

Fig. 8—A slug of plasma from the mercury plasma source is shot into the axially nonuniform rf field of the waveguide. This field accelerates the plasma and the imparted velocity is determined with the pick-up probes which measure the time of flight. The force on the plasma comes about in the same way that an ordinary dielectric is drawn into a condenser through the action of the nonuniform fringe fields.

the disturbance. When the current j and field H are high enough, this force overcomes the damping effect due to diffusion. The disturbance then grows, or "oscillates." These effects can also occur in the electron-hole plasmas existing in solids. The germanium *oscillator*[13], whose properties have been extensively investigated at the RCA Laboratories, has recently been shown[14] to operate on the basis of the same *hydromagnetic instability* as exists in the aforementioned discharge.

Because it behaves as a polarizable dielectric, a plasma experiences a body force when exposed to a nonuniform rf electric field. This fact has received much attention because of its possible use in confining thermonuclear plasmas and as a means of propulsion in space. A plasma has a dielectric constant κ which differs from unity by $\kappa = 1 - (\omega_p/\omega)^2$. If E is the rf field of frequency ω applied to the plasma blob, then the energy density within the plasma is $\epsilon_0 E^2/4$, where ϵ_0 is the free space permitivity. The potential energy of the polarized plasma is the difference between this internal energy and the energy density in the absence of plasma, $U = (\epsilon_0 E^2/4) \cdot (\kappa^{-1} - 1)$. A force, $F = dU/dx = (\frac{1}{2}\epsilon_0 E dE/dx)\ \omega_p^2/(\omega^2 - \omega_p^2)$, therefore exists if the field is nonuniform, and the direction of the force depends on whether ω is less than or greater than the plasma frequency ω_p. Experimental measurements of *plasma acceleration and deceleration* have been made in this laboratory, in collaboration with the RCA Astro-Electronics Division.[15] The setup (Fig. 8) consists of a mercury plasma source which shoots plasma blobs of short duration into the nonuniform field of a waveguide and thence past pick-up probes that measure the time of flight of the plasma. These measurements, performed at 140 and 350 Mc, and more recently at 2450 Mc, confirmed the theoretical dependence of the acceleration on ω and ω_p. This dependence is somewhat more complicated than the simple formula above predicts, because the dielectric constant depends also on the collision frequency.

Incidentally, going back to Carpenter's radio blackout, the expression for the dielectric constant $\kappa = 1 - (\omega_p/\omega)^2$ shows why the plasma becomes reflective when ω_p exceeds ω. Actually, this "dispersion" formula is altered not only by collisions but also by dc magnetic fields. Under certain conditions, a magnetic field can even cause the plasma to become transmissive. Now, some people are thinking of fitting space capsules with superconducting magnets for the purpose of shielding the astronaut from cosmic radiation. If present-day ground rules allow the astro people to propose this, then they also permit the microwave engineer to suggest that magnetic fields be used to keep the astronaut communicative, as well as healthy.

ACKNOWLEDGEMENTS

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(With acknowledgements to RCA)

10A POWER SUPPLY

(Continued from page 116)

$$K = \frac{\text{magnetizing current}}{\text{full load current}} \times 100\%$$

$$\text{OR } \frac{\text{leakage reactance referred to sec.}}{\text{secondary reactance}} \times 100\%$$

whichever is the greater. Resistor R = 60/Ims, where Ims is the magnetizing current referred to the secondary.

The output voltage of this power supply is 70 volts on no load, and 50 volts at 10 amperes load current. In the latter condition the ripple on the output is 20 millivolts rms. The mechanical arrangement of the original unit is shown in the accompanying photographs, but this could of course be changed to suit the user.

For a supply unit of this voltage and current rating, the use of an external "Variac" or similar device is not only a good solution to the problem of varying the output voltage, in that it considerably reduces the problem of power dissipation in the power unit itself, but with a unit of this size is also likely to be comparable in cost with other methods that could be used. The decision will ultimately rest on the intended application of the unit, the range of output voltages required, and similar considerations. It would then be necessary in each case to equate cost of voltage variation methods against required performance.

FLUORESCENT SCREENS

(Continued from page 117)

Phosphor P12 is a long-persistence phosphor which exhibits both orange fluorescence and phosphorescence. Types utilizing this phosphor are particularly useful for observing low- and medium-speed recurring phenomena.

Phosphor P14 is a medium-persistence, cascade (two-layer) screen. During excitation by the electron beam, this phosphor exhibits purplish-blue fluorescence. After excitation, it exhibits a yellowish-orange phosphorescence which persists for a little over a minute. Types utilizing this phosphor are particularly useful for observing either low- and medium-speed non-recurring phenomena or high-speed recurring phenomena.

Phosphor P15 emits radiation in the visible green region and in the invisible near-ultraviolet region. The ultraviolet radiation has very-short persistence which is appreciably shorter than that of the visible radiation. This phosphor finds application in flying-spot cathode-ray tubes.

Phosphor P16 has bluish-purple as well as near-ultraviolet fluorescence and phosphorescence with very-short persistence. This phosphor has a stable, exponential decay characteristic and is particularly useful for the high-speed scanning requirements of a flying-spot video-signal generator.

Phosphor P20 has high luminous efficiency, yellow-green fluorescence and medium to medium-short persistence. The screen may be used in applications requiring relatively short persistence and good visual efficiency.

Phosphor P24 is a short-persistence phosphor with green fluorescence and phosphorescence. Its spectral-energy emission characteristic has sufficient range to provide usable energy over the visible spectrum required for generating colour signals from colour transparencies.

Editor .... **Bernard J. Simpson**

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