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RADIOTRONICS

MARCH, 1961

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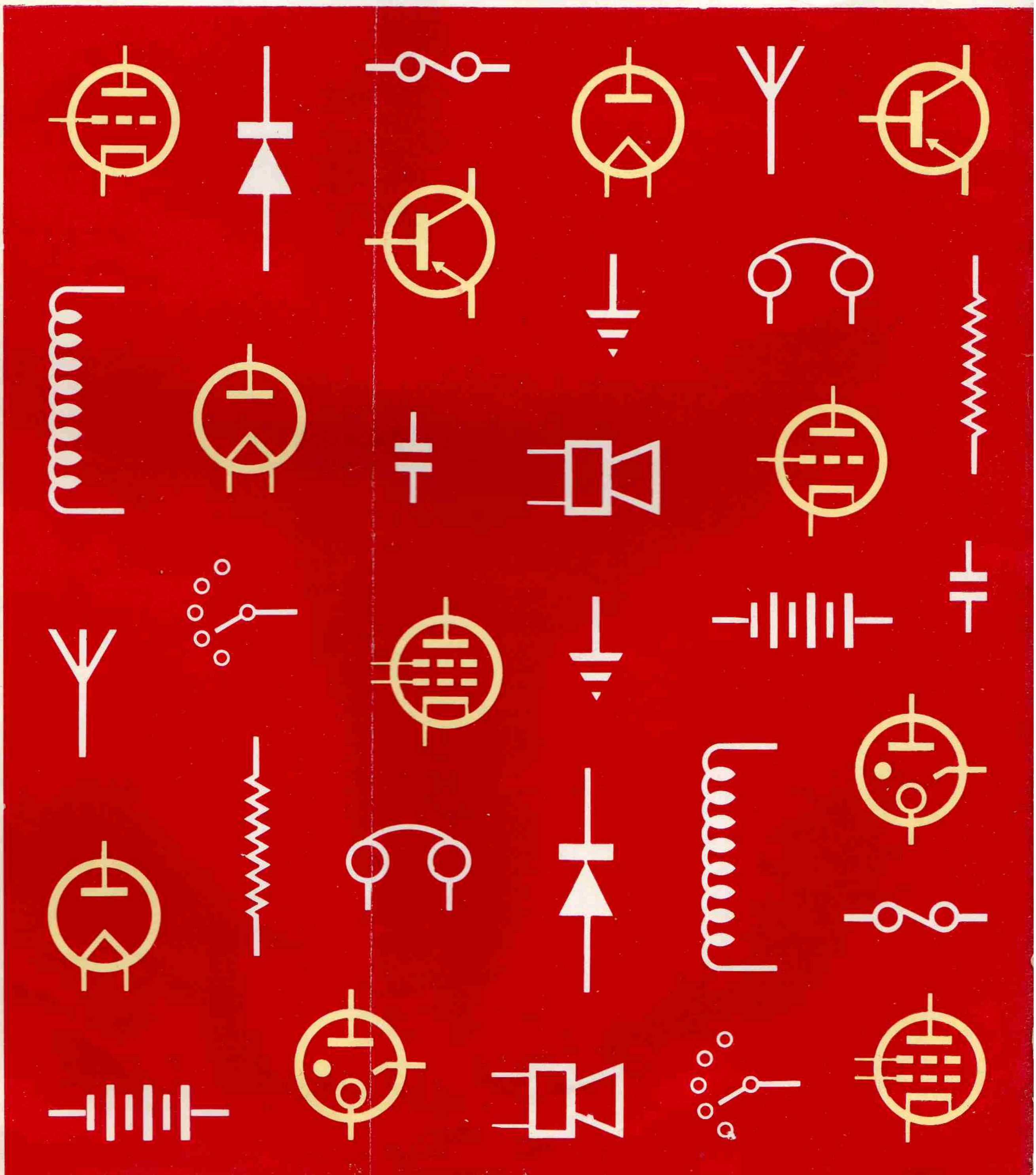

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Editor, Bernard J. Simpson

IN THIS ISSUE

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Photometry as Applied to Camera and Cathode Ray Tube Applications

This article defines the terms, symbols, and units used in photometry, and shows their relationship to corresponding radiometric terms. In addition, methods for the calculation of tube-face illumination for camera tubes and screen luminance for cathode-ray tubes are outlines.

Photometric Terms and Symbols

Photometry deals with the evaluation of radiant energy with respect to its ability to produce visual sensation. Because the degree of visual sensation depends on the wavelength of the radiant energy, there is a difference between radiant and luminous measurements. The actual energy which is transferred by electro-magnetic waves is called radiant energy, but only the radiant energy which stimulates visual sensation is called luminous energy.

A luminosity curve shows the relative effectiveness of various wavelengths of radiant energy to evoke visual sensation for a particular observer under particular conditions. The standard luminosity curve (photopic vision) shown in Fig. 1 has been adopted to make photometry as objective as possible and to provide a standard for the specification of photometric data. This curve represents an average of a number of observations; it is not necessarily the response of any one observer. The ordinate at any point on the relative luminosity curve is often called the "visibility factor" for the wavelength corresponding to that point.

The luminous efficiency of a source of light is the rate of emission of luminous energy per watt of input power. Luminous efficiency should not be confused with the luminous coefficient, which is a ratio of the luminous energy to the radiant energy emitted from a source.

The committee on colorimetry of the Optical Society of America has recommended the nomenclature in Table I, which gives the names, symbols, and basic MKS units used in photometry and radiometry.

Any radiometric unit is converted into the corresponding photometric unit by evaluation with respect to the standard luminosity function, $\bar{y}(\lambda)$. As an example, the standard luminosity curve of Fig. 1 shows that a source at 600 millimicrons will produce 430 lumens of luminous flux for every watt of radiant flux.

Luminous flux F is the rate of flow of luminous energy Q. One lumen corresponds to a luminous flux of one talbot per second. One lumen is also defined as the luminous flux F_c emitted through

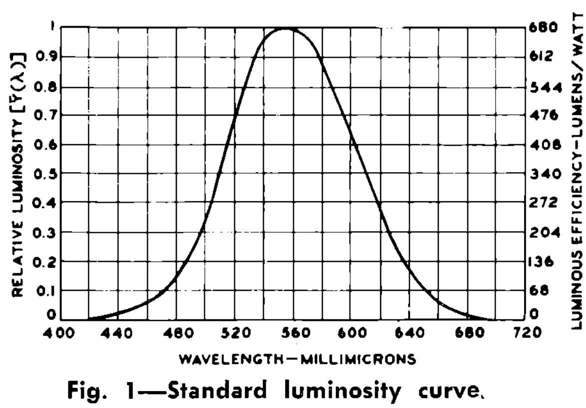


Fig. 1—Standard luminosity curve.

Radiometry Term	Symbol	MKS Units	Photometry Term	Symbol	MKS Units
Radiant Energy	U	joule	Luminous Energy	Q	talbot
Radiant Flux	P	watt	Luminous Flux	F	lumen
Radiant Emittance	W	watt/m^2	Luminous Emittance	L	lumen/m^2
Radiant Intensity	J	watt/ω	Luminous Intensity	I	lumen/ω (candle)
Radiance	N	watt/ω m^2	Luminance	B	lumen/ω m^2 (candle/m^2)
Irradiance	H	watt/m^2	Illuminance	E	lumen/m^2 (lux)
	m	meter		ω	steradian

Table I—Recommended names, symbols, and MKS units for radiometric and photometric terms.

a unit solid angle from a point source of one candle; i.e., a one-candlepower point source emits a total of 4π lumens.

Luminous emittance L is the luminous flux density of an emitting surface and is expressed as $L = F_e/A$ lumens per square metre, or lumens per square foot, where A is the area of the emitting surface.

Luminous intensity I is the ratio of luminous flux emitted by an element of source in a solid angle to the solid angle, or the luminous flux per steradian ω emitted by a source. It is expressed as $I = F_e/\omega$ lumens per steradian.

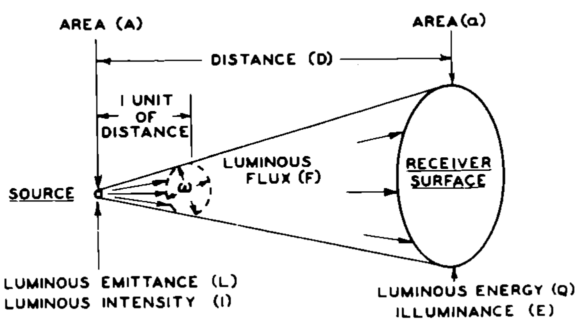


Fig. 2 — Graphic representation of important photometric terms.

Luminance B is the quantitative attribute of light that correlates with the sensation of brightness. The term "brightness" should be used only for nonquantitative statements, especially with reference to sensations and perceptions of light. Units of luminance fall into two classes. The first, "intensity-luminance", is given by the expression $B_i = I/A_{\cos}\theta$ lumens per steradian per square metre, or candles per square metre, where I is the luminous intensity of the area A, as viewed along a direction that makes an angle θ with the normal to the surface. The second unit of luminance applies to sources obeying Lambert's Law, which states that "the intensity from a surface element of a perfectly diffusing radiator is proportional to the cosine of the angle between the direction of emission and the normal to the surface". An element of a surface which obeys this law appears equally bright when observed from any direction. This unit, called lambert-luminance B_l is given by: $B_l = \pi B_i$ footlamberts, metre-lamberts, or the like.

Illuminance E, or illumination, is the density of luminous flux F_r on a surface which is uniformly illuminated. Expressed mathematically, $E = F_r/a$ lumens per square metre (luxes), lumens per square foot (footcandles), or lumens per square centimetre (phots), where a is the area of the illuminated surface.

For perfectly diffusing radiators, the luminous emittance L is equal to π times the "intensity-luminance" B_i. Therefore, a perfectly diffusing radiator having a luminance of B_i candles per square metre emits πB_i lumens per square metre. Fig. 2 graphically illustrates the important terms used in photometric calculations.

Calculation of the Tube-Face Illumination of Camera Tubes

A scene to be televised is optically focused on the face of a camera tube by means of a lens. For practical purposes, the illumination of the camera-tube face can be calculated from the following formula:

$$E = \frac{E_s R T}{4f^2(m + 1)^2}$$

where E is the tube-face illumination in footcandles,
E_s is the scene illumination in footcandles,
R is the reflectance of the scene,
T is the transmission of the lens,
m is the linear magnification from scene to tube face, and
f is the f/number of the lens.

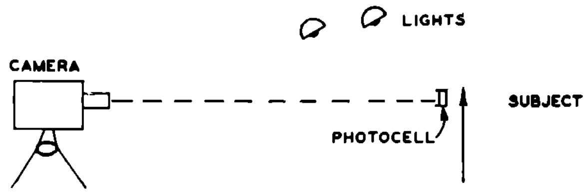


Fig. 3—Set-up for determining illumination of a scene with a photocel meter.

These factors are obtained as follows:

As shown in Fig. 3, the illumination of a scene E_s is usually obtained by means of a calibrated photocell placed near the principal subject in the scene, with the cell facing the camera. The illumination is indicated in footcandles on the photocell meter. For outdoor camera-tube use, the following table may be useful as a "rule-of-thumb" in determining the approximate scene illumination.

Lighting Conditions	Scene Illumination (Footcandles)
Full daylight*	1000-2000
Overcast day	100
Very dark day	10
Twilight	1
Deep twilight	0.1
Full moon	0.01
Quarter moon	0.001
Starlight	0.0001
Overcast starlight	0.00001

* Not direct sunlight.

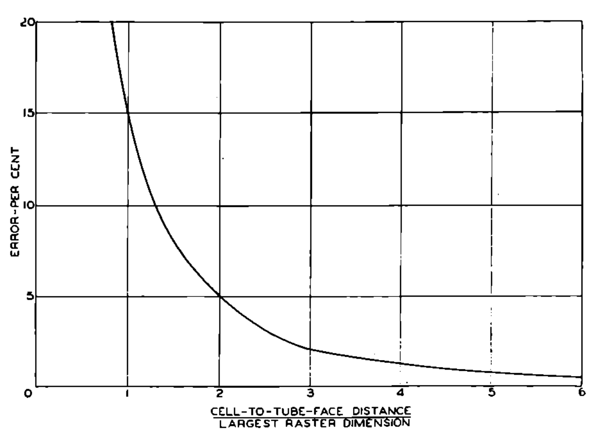


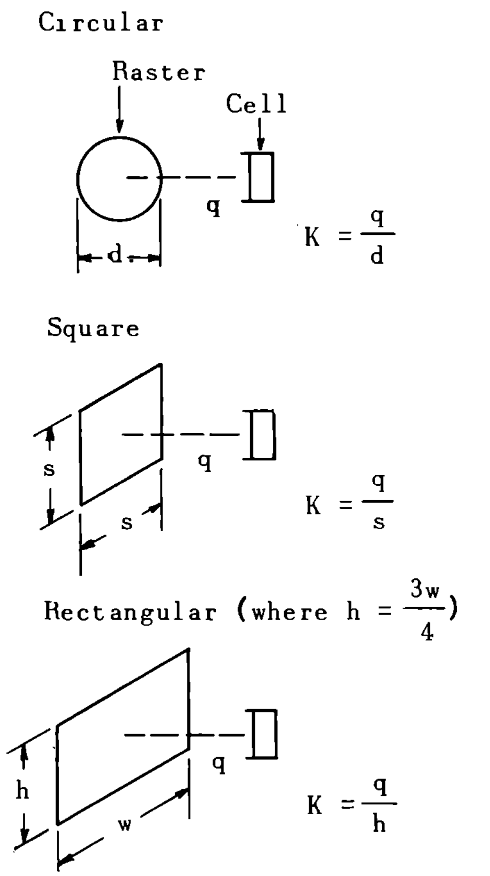
Fig. 4—Per-cent error in luminance calculation of Method II, when cell-to-tube-face distance is less than five times largest raster dimension. Luminance value is low by indicated per cent.

If a neutral density filter is used over the lens, the tube-face illumination is decreased by a factor equal to the filter transmission.

The value of reflectance R of the principal subject in the scene must usually be estimated. The value of reflectance for a test pattern is about 0.9; for live scenes, 0.5 is often used as a typical value.

The product E_sR may be replaced by the luminance of the principal subject, as obtained with a luminance meter. If E is to be in foot-candles, the luminance reading must be in foot-lamberts.

The transmission T of lenses varies. For standard Eastman Kodak Ektanon image-orthicon lenses, 0.9 is a typical value; for some lenses, however, the transmission may be 0.75 or less.



K	Circular	Square	Rectangular
0	1.0	1.0	1.0
1	5.0	4.4	3.4
2	17.0	13.8	10.5
3	37.0	29.6	22.6
4	65.0	51.3	38.8
5	101.0	79.8	60.4
6	145.0	114.7	86.2

Table II—Ratio of luminance B_i to illumination E for various values of K.

The symbol f is the actual f/number used. It should be remembered that the lower the f/number, the greater the illumination on the face of the camera tube.

The magnification m is the quotient of the image size divided by the object size. So long as the distance from the camera lens to the subject is large compared to the focal length of the lens, this term is negligible. For example, if the lens-to-subject distance is 17 times the focal length, the value of the term $(m + 1)^2$ is only 1.12; i.e., a 12-per-cent error is introduced if the magnification is neglected. (The focal length of a lens is usually stamped on the front of the lens mounting.)

Measurement and Calculation of the Luminance of Cathode-Ray-Tube Screen

In general, there are four methods used to measure the luminance of a cathode-ray-tube screen. Methods I and IV measure the average luminance of the area directly beneath the measuring device. Methods II and III measure average luminance over the entire raster area. In these measurements, there are many sources of error and accuracy within 10 per cent is considered satisfactory.

Method 1—Use of a luminance meter or "brightness" meter

When a "brightness" meter is used (several types are commercially available), luminance, usually expressed in footlamberts, can be read directly from the meter provided it is properly calibrated and is pointed at the cathode-ray-tube raster. Because these meters usually view a small area, they can be moved to other areas of the raster to measure luminance variations over the face of the tube. The value obtained from a luminance meter is not a function of distance from the tube face.

Method II—Use of an illumination meter when cell-to-tube-face distance is at least five times the largest dimension of the raster

The eye-corrected Weston Photronic cell is a popular unit for measuring illumination from a source in footcandles. The measured value is a function of distance from the tube face, which can be less than five times the largest raster dimension if errors due to geometry of more than one per cent can be tolerated. Fig. 4 shows percent error in luminance as a function of the ratio of cell-to-tube-face distance to largest raster dimension. Luminance B_i in footlamberts may be obtained from the following equation:

$$B_i = \pi E \frac{D^2}{A}$$

where E is the illumination read on the meter in footcandles, D is the distance from the cell to tube face in feet, and A is the area of the raster in square feet. The cell must be placed perpendicular to the centreline of the tube.

Method III—Use of an illumination meter when cell-to-tube-face distance is less than five times the largest raster dimension

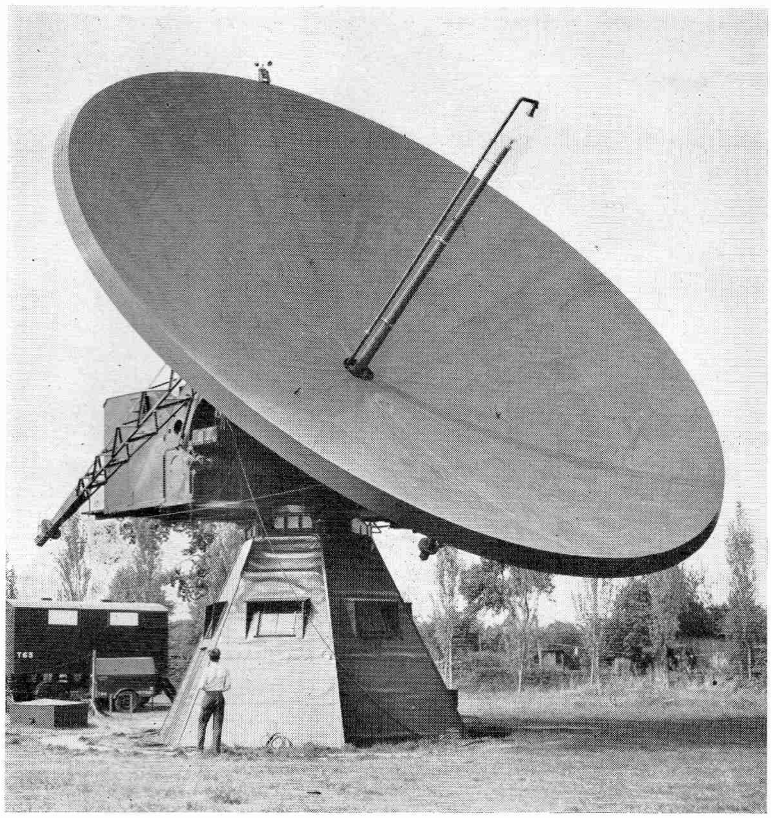
Calculations of luminance are more complicated in this case and must take into consideration the shape of the raster, as well as its area and distance from the cell. The results of such calculations for circular, square, and rectangular (3-by-4 aspect ratio) sources are tabulated in Table II as the ratio of luminance to illumination for various values of K, where K is the ratio of cell-to-raster distance divided by some raster dimension. Because the cell has a finite area an error exists in the calculations, but this error is small compared to the raster area. The B_i/E value given in the table multiplied by the meter reading in footcandles equals the luminance in footlamberts.

Method IV—Use of an illumination meter to read luminance directly

Table II indicates another method of measuring luminance. For values of K equal to zero, or when the cell is placed directly against the tube face, the reading in footcandles is numerically equal to luminance in footlamberts. This relationship is true only if the raster is larger than the cell.

(With acknowledgements to RCA)

MAGNETRON HELPS PROBE MOON'S SURFACE



A view of the 45 ft. radio telescope which was used during investigations by the Royal Radar Establishment at Malvern into the roughness of the moon's surface as a radar reflector. Future investigations may possibly include Venus, Mars and Jupiter. (Crown copyright reserved, reproduced by permission of the Controller H.M. Stationery Office.)

In a letter to "Nature" recently, Mr. V. A. Hughes of the Royal Radar Establishment at Malvern (U.K.) gave news of recent investigations carried out into the roughness of the moon's surface as a radar reflector. From the results of these experiments the assumption is made that the surface has vertical irregularities which are greater than a wavelength and a horizontal scale which is considerably larger. The angular scattering properties of the lunar surface were measured at a wave length of 10 centimetres by a 45ft. radio telescope, the radar transmitter being powered by an English Electric 2.5 Mw M543 magnetron.

This is one of the long-anode magnetrons currently being produced by EEV, capable of delivering a very high mean power. Under typical operating conditions the mean power is 3.75 Kw at 300 pps, with a pulse length of 5 μseconds.

In this experiment a pulse length of 5 μseconds, corresponding to a range resolution of 0.75 kilometre, was used, and since the area of the lunar surface illuminated by the pulse at any one time is constant, the angular scattering law may be obtained by converting the distance travelled by the pulse after the first contact with the moon to angle of incidence of the surface. The law of scattering derived by Mr. Hughes is consistent with scattering from a rough surface which has irregularities much greater than a wavelength and a horizontal scale about twenty times the vertical deviations.

(With acknowledgements to EEV)

PICTURE TUBE PERFORMANCE

GRID-DRIVE AND CATHODE DRIVE COMPARED

This article compares the performance of picture tubes operated under cathode-drive and grid-drive conditions, and points out the relative advantages of each type of service. In addition, the article presents curves of light output, and ultor current as a function of video drive from raster cutoff for both types of service. These comparisons permit the equipment designer to select the type of drive on the basis of individual requirements.

Cathode Drive versus Grid Drive

Many picture-tube types operate satisfactorily when a video signal having the correct polarity is applied to either the control grid (grid-No. 1) or the cathode of the electron gun. The comparative advantages of each operating mode for a given gun design capable of operation in either manner are summarized as follows:

Grid-Drive Service

Optimum beam focus is retained. Because the grid-No. 2-to-cathode voltage remains constant, variations of the electric fields within the gun structure are minimized.

Cathode-Drive Service

Increased beam current and light output are available. Because the grid-No. 2-to-cathode voltage is also video modulated, the modulation sensitivity of the electron gun is increased.

The solid-line curves of Fig. 1 compare the modulation (drive) sensitivity of an electron gun operated in grid-drive and cathode-drive service. These curves show the performance of a typical 17DSP4 picture tube, which employs one of the more widely used gun designs.

The dashed-line curves illustrate the influence which the voltage applied to the grid-No. 2 electrode has upon the drive performance of the electron gun. The curves show that a given gun structure requires less video signal to provide a

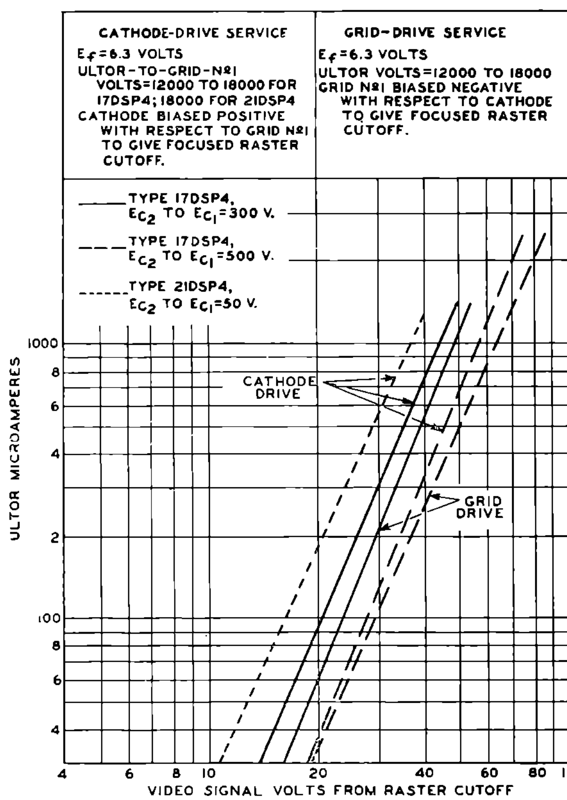


Fig. 1—Ultor current as a function of video signal volts from raster cutoff for both cathode-drive and grid-drive service.

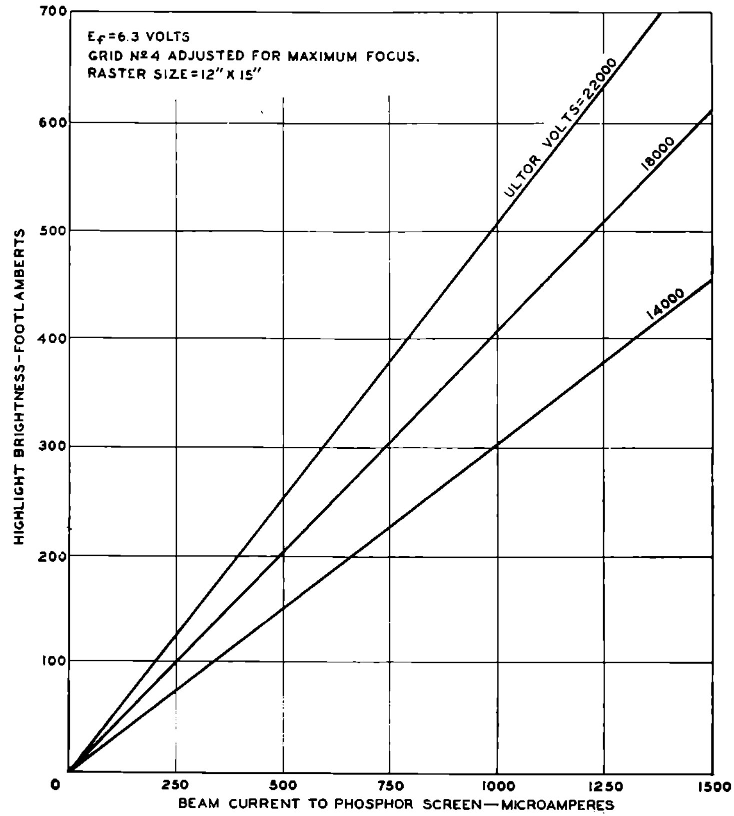


Fig. 2—Light output as a function of beam current to phosphor screen for various values of ultor voltage.

given beam current when a lower value of grid-No. 2 voltage is used. However, operation of an electron gun at a lower grid-No. 2 voltage normally results in some degradation of electron-beam focus because of the influence of the grid- No. 2 electrode potential on the electric fields in the k-g_1-g_2 (cathode-grid-No.1-grid-No.2) region of the gun. These field variations in turn, result in a loss of picture sharpness and clarity. Moreover, lower grid-No. 2 voltages cause some reduction in the maximum beam current which the gun can provide.

Recently, a family of electron guns intended for cathode-drive service only was introduced. These guns were designed to minimize the problem of maintaining optimum beam focus in cathode-drive service while retaining the benefits of the cathode-drive system. Although this problem was not completely solved, commercially acceptable performance was obtained. In addition to improved beam focus, the operating voltage required by the grid-No. 2 electrode was also significantly reduced. The dotted curve in Fig. 1 shows the drive characteristic of picture-tube type 21DSP4 which utilizes a typical gun from this family.

Beam Current versus Light Output

Fig. 2 shows typical curves of light output from a picture tube as a function of beam current to the phosphor screen. In the range from zero to 700 microamperes, beam current and ultor current are substantially equal; above 700 microamperes, beam current becomes increasingly less than ultor current. However, in the normal operating range, the curves of Fig. 2 provide a useful guide if the following approximations are made, when all other factors are constant: For small changes of ultor voltage of the order of ± 3 kilovolts, provided the ultor voltage rating is not exceeded, light output is proportional to ultor voltage (proportionality factor may be approximated from Fig. 2). For small changes in raster area, light output is inversely proportional to raster area (raster width times raster height—not the quoted area for the non-rectangular picture-tube screen).

Summary

In summary, Figs. 1 and 2 show that some increase in beam current and, consequently, light output can be realized from a given amount of video signal by use of cathode-drive service on a conventional gun. Further gain is possible through the use of a gun designed specifically for cathode-drive service. There is, in addition, a possible economic advantage in that the output requirements from the video-amplifier stage are reduced. However, some undesirable effects such as impaired beam focus and picture sharpness normally also result.

Because all these factors are subject to variation from one picture-tube type to another, and from one application to another, the equipment manufacturer should base his final decision upon electrical and visual engineering tests of the specific tube types and chassis designs under consideration.

(With acknowledgements to RCA)

EEV AT THE 1961 PHYSICAL SOCIETY EXHIBITION

The English Electric Valve Co. Ltd. demonstrated design and development advances in several fields including high gain image intensification, signal storage, low noise amplification and ceramic manufacturing techniques for high power rf amplifiers.

High Gain Image Intensifier

The equipment in this demonstration comprised an image intensifier tube together with its focussing solenoid, input and output optical systems and power supplies. A test chart was viewed at a low light level so that it was possible to see the tube's amplification and resolution capabilities. Tubes have been made giving a photon gain of 100,000 to 200,000, a gain of 7,000 being provided by five stages of secondary electron multiplication (thin-film secondary emitting dynodes) and the remainder arising from the bombardment of the output fluorescent screen.

The image to be intensified is focused on the photo-cathode at one end of the tube, and the photoelectrons are accelerated and focused on to the first thin film dynode. Secondary electrons, five or six to each primary, are emitted from the other side of the dynode and these are in turn accelerated and focused on to the second dynode. This process is repeated at each dynode, and the secondaries emitted from the final dynode are accelerated on to a fluorescent screen to produce an intensified replica of the original weak light image.

The usable diameter of the thin film dynodes is one inch. If desired, the picture may be reduced in size between the photo-cathode and the first dynode, so that the diameter of the input picture need not be limited to one inch. Similarly a larger picture may be formed on the screen if required.

Storage Tube Demonstration

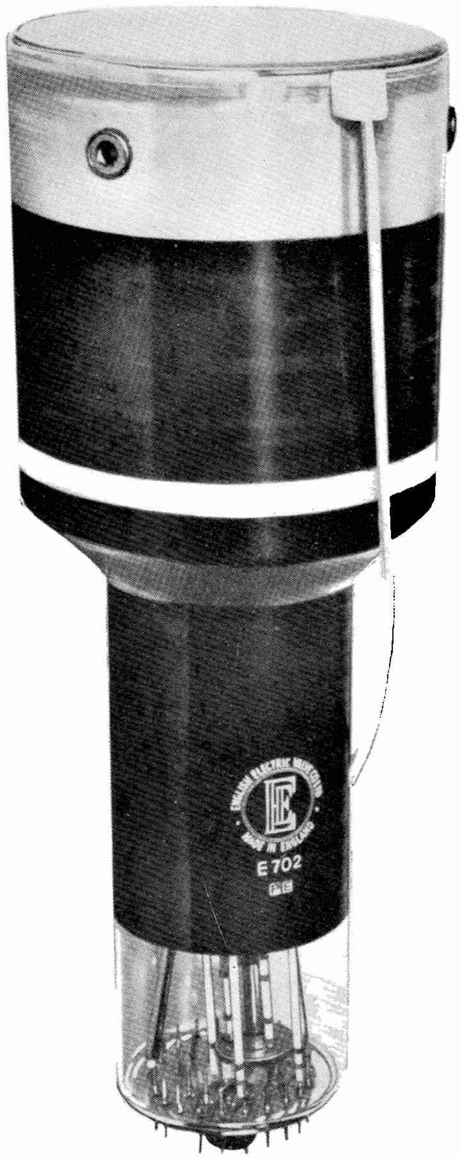
An E710 bi-stable storage tube displayed for analysis a decaying sine wave, generated by a free swinging pendulum. The tube was incorporated into an experimental oscilloscope of very simple design. The E710 will store information indefinitely and allow the continuous or intermittent display of the stored wave-form. Sequential signals can be displayed simultaneously and instantly erased when no longer needed.

Three other prototype tubes were on display. The E701 magnetically-deflected direct-view storage tube intended for bright radar displays; the E709, half-tone storage tube with electrical input and output; and the E711 half-tone tube, effectively the well known E702 in a new outline and with improved characteristics.

Low Noise Amplification

In this demonstration the noise figure of a 600 Mc parametric amplifier was measured by means of loads at room temperature and when cooled by liquid oxygen. The equipment comprised a parametric amplifier with focusing unit and power supply, all of which have been designed as compatable accessories for use in radar, radio astronomy and tropospheric scatter equipments where the 2 db expected noise figures and 20 db gain provide ideal input characteristics.

These tubes cut off under large input signal conditions and this affords complete protection to any following stages, a characteristic very



important in radar applications. Circuit paralysis or recovery time following the application of excessive input overload is extremely short.

The quadrupole parametric amplifier is a transverse field device in which an electron beam interacts with a radio frequency field transverse to the normal path of the beam. A beam produced by the electron gun is directed through an input coupler, a quadrupole or amplifying section, and an output coupler to a collector electrode. This beam is focused by an axial magnetic field, the field strength being such that the cyclotron frequency of the electrons and the input signal frequency are approximately equal.

Couplers of the Cuccia type are employed for injecting and removing the signal from the beam, and the input coupler performs the additional function of removing transverse noise generated by the cathode. The input coupler and the beam can be considered as two coupled transmission lines and, ideally, all transverse noise in the electron beam is transferred to the coupler whilst the input signal is impressed on the beam in the form of a fast wave.

In the input coupler section, the combined action of the transverse signal field and the axial magnetic field causes the beam electrons to describe helical paths with radii increasing linearly as they travel through the section. After leaving the input coupler, the electrons in the modulated beam continue to travel in helical paths of constant radii until they reach the quadrupole section, where the actual amplification takes place.

The quadrupole structure consists of four plates, equispaced round the beam and suitably interconnected, with four coils for tuning to the pump frequency. A pumping signal of twice the cyclotron frequency is applied to this structure and an alternating field is produced. This field may be resolved into two counter-rotating fields in the manner usually employed with four pole electrical machines, each field rotating with a frequency equal to the cyclotron frequency. One field component rotates synchronously with electrons of the correct phase to produce an exponential increase in the radii of their paths and hence to give linear amplification. Electrons in phase opposition to the rotating field are decelerated and the radii of their paths decreases exponentially. The average kinetic energy of the electrons, derived over a number of electrons entering the quadrupole section with every possible phase angle, is increased, resulting in an overall gain.

In the output coupler section, the spiralling electrons transfer their rotational energy to the output circuit and the radii of their paths is reduced to near zero. Unlike the travelling wave tube, the input and output stages of the parametric amplifier are coupled by the beam only; the device is unilateral and is therefore unconditionally stable. In addition, excessive input signals cause the beam to spiral out and strike the tube electrodes, thus "cutting-off" the tube. Since the beam power is very low, no damage is incurred by the tube electrodes under these conditions. The "cut-off" characteristic provides complete protection to any following stages in an equipment and is of particular importance in radar applications.

(With acknowledgements to EEV)

SILICON POWER TRANSISTORS

AN APPLICATION GUIDE

(CONTINUED)

THERMAL RATINGS AND STABILITY CONDITIONS

Thermal ratings of transistors are usually based on the maximum permissible temperature for the collector junction [T_j(max)]. This temperature is based upon the manufacturers' experience in design of transistors, and on life-test and reliability studies. In actual circuit applications the limiting factor may not be Tj(max) but thermal runaway. Therefore, to establish comprehensive thermal ratings for transistors, particularly power transistors, it is necessary to limit maximum junction temperature and prevent thermal runaway. Each of these considerations is discussed in detail below.

Maximum Dissipation as a Function of T_j(max)

For a given value of Tj(max) the maximum instantaneous dissipation is a function of transistor thermal impedance, contact impedance, heat-sink impedance, and the temperature of the absolute reference or heat sink (usually free air).

Heat transfer in power transistors has been analyzed in "Transistor Dissipation Ratings for Pulse and Switching Service", "Radiotronics", Vol. 24, No. 11, November, 1959, which gives a graphical solution for the diffusion equation of heat flow for pulse response. This graphical solution is shown below in Fig. 3. From a knowledge of two transistor thermal constants [steady-state thermal resistance junction-to-case (R), and thermal time constant junction-to-case (τ_1)], and the transistor-case temperature (at the transistor mounting flange), the maximum transistor dissipation for any transient, pulse, or steady-state condition can be found from Fig. 3. Fig. 3 can also be used for applications involving the use of heat sinks, provided the thermal time constant of the chassis is much larger than the thermal time constant τ_1 of the transistor, which is usually the case. The steady-state thermal resistance R_T is the total thermal resistance between junction and free air or ultimate heat sink; that is, $R_T = R + R_C + R_S$, where

R is the junction-to-case thermal resistance for the transistor
R_C is the contact thermal resistance
R_S is the heat-sink thermal resistance.

The dissipation nomogram is represented in Fig. 3 by a family of repetitive-pulse-power curves with pulse width (t_O) normalized by τ_1 for various duty-cycle values of the ratio

$$\frac{\triangle Tj(t)}{P(max) R_T}$$

(shown as the ordinate), where

P(max) is the peak value of the pulse power dissipated in the transistor.
$\triangle$Tj(t) is the temperature rise between junction and reference (case, heat sink, free air, etc.)

Values of τ_1, Tj(max) and R for the three families of RCA silicon power transistors are shown below:

Family	τ_I (msec)	Tj(max) (°C)	R (°C/W)
Medium-Power Types	10	175	37·5
Intermediate-Power Types	10	175	10
High-Power Types	12	175	2·5

Some typical values of contact thermal resistance (R_C) for RCA silicon power transistors are :

No insulator	0·4 ohm
0·125″ anodized-aluminium insulator	0·5 ohm
0·002″ mica insulator	0·5 ohm
0·003″ glass-cloth insulator ..	1·25 ohms

Equation (6) gives the approximate value of heat-sink thermal resistance (R_S)

$$R_S = \frac{100}{A_S} \qquad (6)$$

where A_S is the exposed heat-sink area in square inches.

For applications involving a single pulse of power dissipation, the 0·00001 duty-cycle curve should be used. For steady-state applications use the ratio

$$\frac{T_j(\max)}{P(\max)R_T} = 1$$

(For steady-state applications $t_0 \rightarrow \infty$)

When the actual value of $T_j(\max) / P(\max)R_T$ has been determined from Fig. 3, determine P(max) for a given maximum rise in temperature $T_j(\max)$. This rise in temperature is the difference between the maximum permissible junction temperature and the temperature of the ultimate heat sink (usually free air). The same procedure may be used to determine rise in junction temperature $T_j(t)$ and junction temperature T_j, for a given P(max), pulse conditions, and ultimate heat-sink temperature.

To illustrate the application of this thermal-rating system, determine P(max) for a type 2N1490 in a power-switching application, if the dissipation is a single pulse of 6 milliseconds duration. The ambient temperature is 125°C and the transistor is mounted by means of a 0·002″ mica insulating washer on a 50 sq. in. (exposed) heat sink. Also calculate the maximum steady-state dissipation for the same thermal conditions.

For the 2N1490 :

$$\tau_1 = 12 \text{ msec}, R = 2{\cdot}5°C/W, T_j(\max) = 175°C$$

For 0·002″ mica :

$$R_C = 0{\cdot}5 \ °C/W$$

From equation (6) :

$$R_S = \frac{100}{50} = 2{\cdot}0 \ °C/W$$

$$R_T = R + R_C + R_S = 2{\cdot}5 + 0{\cdot}5 + 2{\cdot}0 = 5{\cdot}0 \ °C/W$$

$$\triangle T_j(\max) = T_j(\max) - T_{ambient} = 175 - 125 = 50°C$$

From Fig. 3 for the pulse case :

$$t_0/\tau_1 = 0{\cdot}5, \text{ duty cycle} = 0{\cdot}00001$$

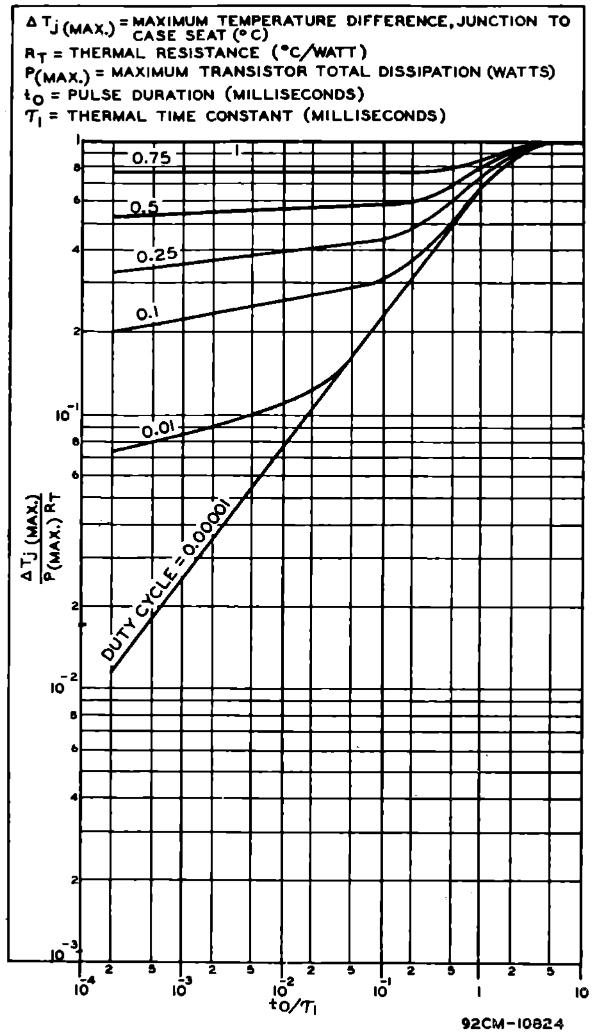


Fig. 3—Chart for Determining Maximum Permissible Dissipation of RCA Silicon Power Transistors for Pulse Service.

$$\frac{\triangle T_j(\max)}{P(\max)\,R_T} = 0{\cdot}48, \ P(\max) = \frac{\triangle T_j(\max)}{0{\cdot}48 \ R_T}$$

$$P(\max) = \frac{50}{(0{\cdot}48)\ (5{\cdot}0)} = 20{\cdot}7 \text{ watts}$$

For the steady-state case :

$$\frac{\triangle T_j(\max)}{P(\max)R_T} = 1{\cdot}0, \ P(\max) = \frac{\triangle T_j(\max)}{1{\cdot}0 \ R_T}$$

$$P(\max) = \frac{50}{(1{\cdot}0)\ (5{\cdot}0)} = 10 \text{ watts}$$

This example demonstrates the significant increase in allowable transistor dissipation when the transistor is used in a pulse-type application.

Temperature-Stability Considerations and Stability Factor

The operating point of a transistor changes with temperature because collector current (I_C) is dependent on I_{CBO}, which is a function of temperature.

The rate of change of I_C with respect to I_{CBO} is called the " stability factor " (S). Because S is partly determined by circuit components other than the transistor, the circuit designer is able to exercise some control over its value. A very low value of S will provide very stable operation at the expense of reduced efficiency and, in some circuit configurations (particularly resistance-capacitance coupled amplifiers), at the expense of reduced power gain. A high value of S, on the other hand, will result in less stability of the operating point, but will minimize loss of efficiency and power gain.

Under conditions of extreme instability, it is possible for a transistor to overheat and destroy itself by a process known as " thermal runaway ". As the junction temperature of a transistor increases due to an increase in ambient temperature, power dissipation, or both, the I_{CBO} of the transistor increases. The increase of I_{CBO} and consequent increase in I_C causes a further increase in junction temperature and, hence, in I_{CBO}. Thus, an infinite series of increases in junction temperature occurs. If the series does not converge, thermal runaway occurs, the junction temperature increases without limit, and the transistor is destroyed by heat. It can be shown that the criterion for the convergence of the series is :

$$S(max) < \frac{1}{R_T V_{CE} \frac{dI_{CBO}}{dT_j}} \qquad (7)$$

A major advantage of RCA silicon power transistors as compared with germanium power transistors is that the derivative dI_{CBO}/dT_j for the silicon types is much smaller over the useful operating-temperature range. This feature permits the silicon types to operate stably with higher values of S, and minimizes loss of efficiency and power gain.

The value of dI_{CBO}/dT_j can be found by drawing a line tangent to the curve of I_{CBO} versus junction temperature at the highest junction temperature at which the transistor is to be operated, and using the following relationship :

$$\frac{dI_{CBO}}{dT_j} = \left(2{\cdot}3\, I_{CBO}(max)\right) \times \left(\frac{\log_{10} I_{CBO1} - \log_{10} I_{CBO2}}{T_{j1} - T_{j2}}\right) \qquad (8)$$

where $I_{CBO}(max) = I_{CBO}$ at the highest junction temperature, T_{j1}, I_{CBO1} and T_{j2}, I_{CBO2} are points on the tangent line.

The designer must prevent thermal runaway of the *limit* transistor. Because the curve referred to is for a *typical* transistor, the value of $I_{CBO}(max)$ read on the curve should be multiplied by the ratio of the maximum value of I_{CBO} at 175°C to the typical value of I_{CBO} at 175°C, and the result used as $I_{CBO}(max)$ in equation (8).

Note that in equation (8) dI_{CBO}/dT_j is found by multiplying $d \log_e I_{CBO}/dT_j$ by the magnitude of I_{CBO}. The reason it is possible for dI_{CBO}/dT_j to be considerably smaller for silicon than for germanium is that for silicon the magnitude of I_{CBO} is much smaller over the useful operating-temperature range.

After dI_{CBO}/dT_j has been determined, the maximum permissible value of S which will assure freedom from thermal runaway may be calculated from equation (7). If the calculated value of S is greater than $h_{FE} + 1$, thermal runaway cannot occur in any single-transistor circuit having a bias network composed of linear circuit elements, because in such a circuit S can only assume values from 1 to $h_{FE} + 1$. This rule, however, may not apply for circuits using direct-current coupling between two or more transistors.

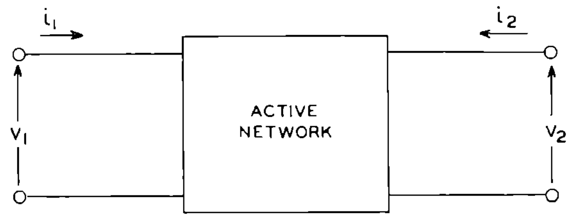


Fig. 4—Basic Four-Terminal Linear Network.

In many circuits using RCA silicon transistors, prevention of thermal runaway is not difficult. However, the maximum permissible value of S may be dictated not by the necessity for preventing thermal runaway, but by the necessity to stabilize the operating point to the degree necessary to keep distortion of the signal at a tolerable level. This consideration may make it necessary to use a value of S lower than that necessary to prevent thermal runaway. The value of S necessary to minimize distortion may be determined from :

$$S = \frac{DI_C}{100 \triangle I_{CBO}}$$

where D is the permissible per cent. change in I_C, and $\triangle I_{CBO}$ is the change in I_{CBO} over the operating temperature range (this value may be found from the curve of I_{CBO} vs. T_j).

As an example, a 2N1480 is to be used in an application where $V_{CE} = 25$ volts, $I_E = 40$ ma. The transistor must operate over an ambient temperature range of 25°C to 100°C. A clamp exerts pressure on the transistor so that its case seat makes firm contact with a piece of 0·003″ glass cloth which separates the transistor case seat from a heat sink of 9 sq. in. exposed to free air. Requirements : (1) choose a value of S_{max} which will prevent thermal runaway : (2) determine a value of S_{max} which limits distortion to a degree such that $D<5\%$.

First, determine R_T :

$$R_T = R + R_C + R_S$$
$$= 37{\cdot}5°C/Watt + 1{\cdot}25°C/Watt + \frac{100}{9}°C/Watt$$
$$= 49{\cdot}85°C/Watt$$

Now, calculate $\triangle T_j$:

$$\triangle T_j = R_T\ V_{CE}\ I_E$$
$$= 49{\cdot}85°C/Watt \times 25 \text{ volts} \times 0{\cdot}04 \text{ ampere}$$
$$= 49{\cdot}86°C \text{ or } 50°C \text{ (approx.)}$$

If it can be assumed that I_E and V_{CE} remain constant, T_j varies from 75°C to 150°C.

Next, draw a line tangent to the curve at 150°C and calculate dI_{CBO}/dT_j from equation (8) :

$$\frac{dI_{CBO}}{dT_j} = \left(2{\cdot}3\,I_{CBO}(max)\right)\left(\frac{\log_{10}I_{CBO1} - \log_{10}I_{CBO2}}{T_{j1} - T_{j2}}\right)$$

$$\frac{dI_{CBO}}{dT_j} = \left(2{\cdot}3 \times 1{\cdot}5 \times 10^{-4} \text{ amperes}\right) \times \left(\frac{\log_{10}3 - \log_{10}0{\cdot}15}{125°C - 75°C}\right)$$

The value of $1{\cdot}5 \times 10^{-4}$ ampere for $I_{CBO}(max)$ was found by multiplying the value read on the curve ($0{\cdot}12 \times 10^{-4}$ ampere) by the ratio 750 μa/60 μa.

$$\frac{dI_{CBO}}{dT_j} = 8{\cdot}97 \times 10^{-6} \text{ ampere/°C}$$

From equation (7) the maximum permissible value of S is :

$$S_{max} < \frac{1}{R_T\ V_{CE}\ \frac{dI_{CBO}}{dT_j}}$$

$$S_{max} < \frac{1}{49{\cdot}86°C/W \text{ x } 25 \text{ volts x } 8{\cdot}97 \text{ x } 10^{-6} \text{ amp/°C}}$$

$$S_{max} < 89{\cdot}3$$

Thus, if I_E and V_{CE} remain constant, any value of S less than 89·3 will prevent thermal runaway. Of course, I_E does not remain constant. If there is sufficient resistance in the emitter or collector circuit, the rise in I_E may be compensated for by a fall in V_{CE}. The worst possible case occurs when V_{CE} remains approximately constant as I_E increases, which is usually the case in a transformer-coupled circuit. To assure that runaway will not occur in such a case, a value of $\triangle I_C = S\triangle I_{CBO} = \triangle I_E$ may be calculated using S = 89·3 ; then a new value S_1 calculated using the new value of I_E, $(I_E + \triangle I_E)$ to determine T_j and dI_{CBO}/dT_j, and the process continued until a final maximum value of S_n which will prevent thermal runaway is found.

To choose the value of S based on the distortion requirement :

$$S = \frac{DI_C}{100\triangle I_{CBO}}$$
$$= \frac{5 \times 40 \text{ ma}}{100 \times (150 \times 10^{-3} - 20 \times 10^{-3}) \text{ ma}}$$
$$= 15{\cdot}35$$

PARAMETERS AND EQUIVALENT CIRCUITS

Although RCA silicon power transistors are used mainly in large-signal-amplifier and switching applications they find some use in medium-to-small signal service. To provide design assistance in this area equivalent circuits and measured parameters for the RCA silicon power types at selected dc operating points are presented below.

Consider the generalized four-terminal linear network shown in Fig. 4. If the equations relating all the currents and voltages are known, the network is completely defined. Since the transistor is an active network, four independent parameters are needed to determine the network.

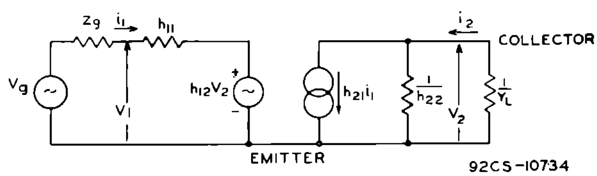


Fig. 5—"h" Parameter Equivalent Circuit of an RCA Silicon Transistor.

The three commonly used sets of parameters for transistors are the "z" parameters, "y" or "hybrid-π" parameters, and "h" parameters. Because in most circuits a transistor has an output impedance much greater than the load impedance, it is usually most convenient to use an equivalent circuit with an output-current generator in parallel with an output impedance. With this type of equivalent circuit the output impedance can be ignored whenever it is much greater than the load impedance. Although both the "y"-parameter and "h"-parameter equivalent circuits contain an output-current generator, the "h" parameters are usually preferred at low frequencies because they are more easily measured at these frequencies. At the higher frequencies the "y" parameters are preferred because they more closely represent the device.

"h" Parameter Equivalent Circuit

The "h"-parameter equivalent circuit with the generator and load is shown in Fig. 5.

TABLE II

Typical "h" and "hybrid-pi" Parameters for RCA Silicon Power Transistors

Type	Common-Emitter Circuit, Base Input Ambient Temperature = 25° C												
	Conditions		"h"-Type				"hybrid-pi" Type						
	DC Collector to-Emitter Volts V_{CE}	DC Collector Milli-amperes I_C	$h_{11} = h_{ie}$ (ohms)	$h_{12} = h_{re}$	$h_{21} = h_{fe}$	$h_{22} = h_{oe}$ (μmhos)	$r_{bb'}$ (ohms)	$r_{b'e}$ (ohms)	$C_{b'e}$ (μf)	$r_{b'c}$ (megohms)	$C_{b'c}$ ($\mu\mu$f)	g_m (mhos)	$r_{c'e}$ (ohms)
Medium Power													
2N1479 to 2N1482	4	5	300	144×10^{-6}	48	54	20.8	279	0.0316	1.59	333	0.172	27600
	4	10	188	128×10^{-6}	52	99	19	169	0.058	0.96	386	0.308	19800
Intermediate-Power													
2N1483 to 2N1486	4	10	186	120×10^{-6}	56	80	6	180	0.0417	1.16	490	0.312	19800
	4	25	98	110×10^{-6}	62	175	4.4	93.6	0.0995	0.625	605	0.66	10600
	4	50	63	116×10^{-6}	64	320	3.96	59	0.19	0.378	730	1.086	4300
High-Power													
2N1487 to 2N1490	4	50	66	248×10^{-6}	78	750	2.4	63.6	0.073	0.216	650	1.22	2160
	4	100	38.5	207×10^{-6}	78	1270	2.3	36.2	0.148	0.144	820	2.15	1270
2N1511 to 2N1514	4	50	66	248×10^{-6}	78	750	2.4	63.6	0.073	0.216	650	1.22	2160
	4	100	38.5	207×10^{-6}	78	1270	2.3	36.2	0.148	0.144	820	2.15	1270

Definitions of the "h" parameters for the common-emitter connection are:

Short-circuit input impedance (v_c constant):

$$h_{11} = h_{ie} = \frac{V_{be}}{i_b}$$

Open-circuit reverse voltage gain (i_b constant):

$$h_{12} = h_{re} = \frac{V_{be}}{V_c}$$

Short-circuit forward current gain (v_c constant):

$$h_{21} = h_{fe} = \frac{i_c}{i_b}$$

Open-circuit output admittance (i_b constant):

$$h_{22} = h_{oe} = \frac{i_c}{V_c}.$$

The low-frequency "h"-parameter design-centre values for the three families of RCA silicon power transistors are shown in Table II. The corresponding mesh equations for the circuit in Fig. 5 are:

$$V_g = i_1 (z_g + h_{11}) + h_{12} V_2 \tag{9}$$

$$h_{21} i_1 = -V_2 (h_{22} + Y_L) \tag{10}$$

Defining the circuit performance:

(a) Input impedance:

$$z_{in} = h_{11} - \frac{h_{12} h_{21}}{h_{22} + Y_L} \tag{11}$$

(b) Output impedance:

$$z_{out} = \frac{z_g + h_{11}}{z_g h_{22} + h_{11} h_{22} - h_{12} h_{21}} \tag{12}$$

(c) Current gain:

$$A_i = \frac{h_{21} Y_L}{h_{22} + Y_L} \tag{13}$$

(d) Power gain:

$$G = \left(\frac{h_{21}^2}{h_{11}h_{22}}\right) \left(\frac{1}{1 + h_{22}/Y_L}\right) \left(\frac{1}{1 - \frac{h_{12}h_{21}}{h_{11}h_{22}} + \frac{Y_L}{h_{22}}}\right) \tag{14}$$

(e) Voltage gain:

$$A_V = \frac{-h_{21}/h_{11}}{Y_L + h_{22}\left(1 - \frac{h_{12}\ h_{21}}{h_{11}\ h_{22}}\right)} \tag{15}$$

For RCA silicon power transistors using the values shown in Table II, the "h"-parameter equivalent circuit is useful at frequencies up to 10 Kc. At higher frequencies, the "hybrid-pi" equivalent circuit discussed below should be used.

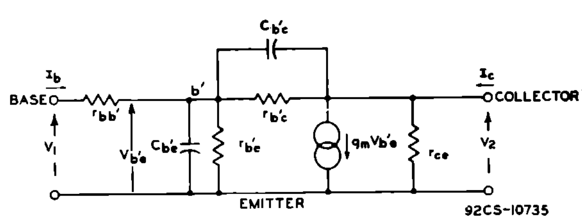


Fig. 6—"hybrid-pi" Equivalent Circuit of an RCA Silicon Power Transistor.

" Hybrid-Pi " Equivalent Circuit

The " hybrid-pi " equivalent circuit not only represents the transistor fairly well over its useful frequency range, but in addition, its parameters are frequency-independent and easily related to the physics of the device. The " hybrid-pi " common-emitter equivalent circuit is shown in Fig. 6.

Definitions of the " hybrid-pi " parameters are :

$r_{b'b}$ = lead resistance plus base-spreading resistance
$r_{b'e}$ = base-to-emitter resistance
$C_{b'e}$ = diffusion capacitance plus depletion capacitance
$r_{b'c}$ = feedback resistance
$C_{b'c}$ = feedback capacitance
g_m = intrinsic transconductance
r_{ce} = collector-to-emitter resistance

The resistance terms in this circuit are related to the parameters of the low-frequency common-emitter " h " type circuit in the following manner :

$$r_{b'e} = h_{ie} - r_{bb'}$$

$$r_{b'c} = \frac{h_{ie} - r_{bb'}}{h_{re}}$$

$$g_m = \frac{h_{fe}}{h_{ie} - r_{bb'}} \qquad (16)$$

$$\frac{1}{r_{ce}} = h_{oe} - \frac{(1 + h_{fe})\, h_{re}}{h_{ie} - r_{bb'}}$$

The design-centre measured values of all the " hybrid-pi " parameters for the three families of RCA silicon power transistors are shown in Table II. These parameters are measured for two different dc collector currents. For other values of dc collector current, the parameters are modified by the following expressions :

$$r_{b'e} = \frac{K_1\,(h_{FE} + 1)}{I_C}$$

$$g_m = \frac{K_2\, I_C\, (h_{FE})}{h_{FE} + 1} \qquad (17)$$

$$C_{b'e} = K_3\, I_C$$

where K_1, K_2, K_3 are constants that can be determined from the two sets of parameters in Table II.

(TO BE CONTINUED)

BILLIONS OF ATOMS

The germanium utilized in the fabrication of transistors and diodes must be brought to an exceptionally high degree of purity. By repeatedly heating the germanium crystal by regions up to the fusion temperature, the foreign bodies contained in the germanium crystal are expelled, so that, at the end, the ratio of impurity to germanium is only 1 part to 10 thousand millions.

However, in order to render the germanium effective and to produce the drift of electrons necessary for the generation of current, a definite quantity of certain foreign bodies must be added to the pure metal. Depending on the type of transistor desired, the quantity added will amount to 50 to 5000 billion impurity atoms to each cubic centimetre. Even this enormous quantity gives such a low degree of impurity, however, that it cannot be detected by chemical or spectroscopic analysis. Only the electrical measurement of the resistance, which reacts with high sensitivity to impurities, makes it possible for manufacturers to determine with considerable accuracy the degree of purity of the metal.

The following comparison will give an idea of the extremely small size of a germanium atom with which researchers work as with a tangible object. For instance, a cube with a volume of one cubic centimetre is a conceivable object. Such a cube consists of 45 x 10^{21} atoms, twenty billion times the entire population of the earth.

The smallness of transistors is one of their advantages in modern circuits. For instance, one typical sub-miniature transistor weighs only 1/2 gramme, and standard types have a weight of about 1 gramme.

The quantity of germanium contained in these transistors is even much smaller. It is only a wafer of about 6 square millimetres and approximately 1/10th millimetre thick. The higher the operating frequency of the transistor, the thinner the germanium wafer must be. In the high-frequency transistors used to-day, which are capable of performing at vhf frequencies, the germanium wafer is reduced to a thickness of 40μ, or four one hundredths of a millimetre. The transit time of electrons between both poles of the transistor (emitter and collector) is only 2.5 x 10^{-9} = 2.1/2 millionths of a second.

(With acknowledgements to Telefunken)

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