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DIAMOND JUBILEE OF TRANSATLANTIC RADIO

Sixty years ago—on the 12th December, 1901—Guglielmo Marconi became the first to send a wireless signal across the Atlantic. This remarkable achievement with such primitive equipment marked the birth of world-wide communication.

During the spring of 1900, Marconi had succeeded in sending reliable signals from St. Catherines in the Isle of Wight to The Lizard in Cornwall, a distance of 186 miles. This encouraged his belief that by using larger aerials and far more powerful transmitters he would be able to achieve transatlantic distances. Scientists were highly sceptical, many said it was impossible because of the curvature of the earth.

Marconi determined to make the attempt. A transmitting station nearly one hundred times more powerful than any previously constructed was built at Poldhu near Mullion in Cornwall. Enormous aerials were erected at Poldhu and at Cape Cod in Massachusetts but both were wrecked in severe gales. Another, less ambitious in design, was put up at Poldhu while Marconi and his two assistants sailed to Newfoundland where, from the top of Signal Hill, a receiving aerial was hoisted, at the third attempt, by means of a kite.

At 12.30 p.m. (Newfoundland time) on December 12th, 1901, Marconi and his assistant G. S. Kemp, using one of the primitive receivers of the period with a telephone earpiece heard a faint succession of S's in Morse code. Signals from Poldhu, 2,200 miles away, had crossed the Atlantic.

SPANNING THE ATLANTIC

During the early 1890's, many of the leading physicists were closely interested in the properties of 'Hertzian waves' but none expressed a thought that these waves would be of the slightest value for the purpose of communication.

In 1895, Guglielmo Marconi, working at his parents' residence at Pontecchio in Italy, discovered the great increase in range which could be obtained by the use of an elevated aerial. It was this discovery which paved the way for the use of Hertzian waves in a practicable system of wireless telegraphy.

Early in the following year—1896—Marconi arrived in England and applied for the world's first patent for wireless telegraphy. He had chosen to come to England partly because this country was then the most powerful maritime nation in the world and it seemed likely that wireless telegraphy would be of value to shipping, and partly because of a national affinity, his mother being Irish.

In 1897 he founded the Wireless Telegraph & Signal Company (which in 1900 became MARCONI'S WIRELESS TELEGRAPH COMPANY LIMITED), and this provided him with the money and technical resources necessary for his future developments.

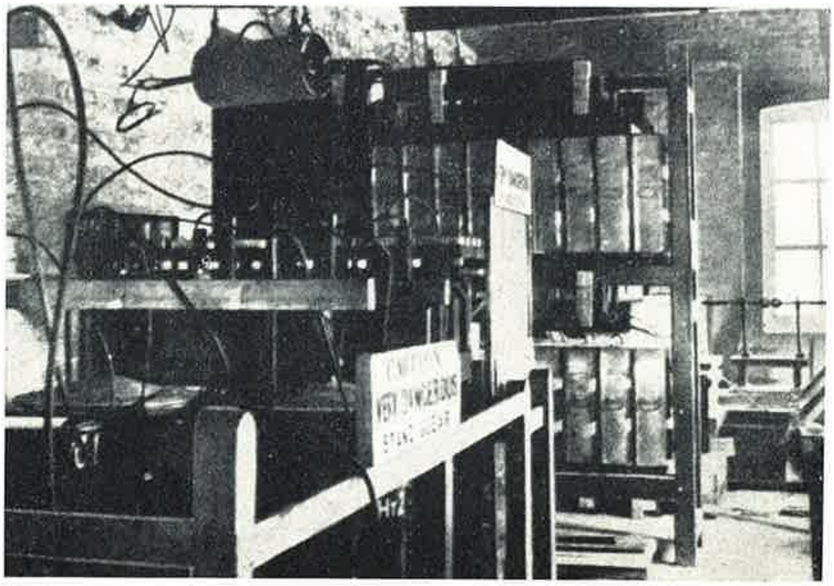
Marconi spent the next four years in an almost continual round of experiment, development and demonstration, his object being continually to improve the reliability and range of his apparatus. At first only covering a mile or so on Salisbury Plain, he was soon communicating regularly from Alum Bay near the Needles in the Isle of Wight to Bournemouth and then to Sandbanks at the entrance to Poole Harbour, a distance of 18 miles. In March 1899 he spanned the English Channel and early in the following year he set up reliable communication from the Isle of Wight to The Lizard in Cornwall, a distance of 186 miles.

The shipping companies had shown mild interest but very little enthusiasm to install wireless equipment on their ships. It seemed, in fact, that far greater ranges and a chain of land stations would be required before wireless telegraphy would have a wide appeal. The scientists of the day, however, were almost united in believing that wireless waves, like light waves, would not follow the curvature of the earth. Therefore, they said, really long ranges were impossible.

Marconi thought otherwise. Experiments had led him to believe that the key to longer ranges lay in the employment of larger aerials and higher transmitter powers. He therefore determined to build two super-power transmitting stations, one on each side of the Atlantic, and to attempt two-way communication. Accordingly, a site was selected at Poldhu in Cornwall and the other at Cape Cod in Massachusetts.

It is difficult to visualise the stupendous problems which confronted him. The aerial system, at both Poldhu and Cape Cod, was of a size and complexity which had never been attempted before, for it consisted of twenty 200 ft. masts in a circle with an inverted cone of about 400 wires leading down to the transmitter. As to the transmitter itself, it was to be 100 times more powerful than any hitherto built, and no precedents whatever existed for the design. Marconi delegated the responsibility for this to his scientific adviser, Professor J. A. (later, Sir Ambrose) Fleming, and Fleming carried it out brilliantly.

Some details of the transmitter may be of interest. The prime mover for the generation of power was a Hornsby-Ackroyd oil engine which drove a Mather and Platt 2000v 50 cps alternator. This was capable of delivering 25 Kw, although from a paper read by Fleming to the Royal Society of Arts in December 1921 it would appear that the plant was considerably under-run at the time of the transatlantic tests.



Part of the transmitting apparatus at Poldhu with which the first signals were made across the Atlantic.

The transmitter proper, which embodied a form of the new syntonic tuning with all its advantages, employed two 20 Kw Berry transformers parallel-connected to step up the input voltage to 20,000 volts. This was fed through rf chokes to a closed oscillatory circuit in which a capacitor discharged across a spark gap via the primary of a "jigger" or rf transformer. The secondary of this transformer connected to a second spark gap and capacitor and the primary of a second rf transformer, the secondary winding of this transformer being in series with the aerial. Keying was effected by the short circuiting of the chokes in the alternator output.

The capacitors were made of 20 glass plates each 16″ square, backed on one side with one square foot of tinfoil. The plates were immersed in linseed oil contained in stoneware boxes; each box had a capacity of approximately 0.05 mfd.

Both the Poldhu and Cape Cod stations were all but ready when a double catastrophe struck; severe gales wrecked the aerial arrays and masts at both stations almost simultaneously.

With £50,000 already spent on the project, Marconi elected not to wait until both stations were repaired. Instead, a new aerial system was erected at Poldhu, consisting of 54 copper wires arranged in a fan-shape and upheld by a triatic slung between two 150 ft. masts. The current into the bottom of this aerial is stated by Fleming to have been 17 amperes and the radiated frequency is thought to have been between 100-150 Kc. No one knows for certain, however, as no reliable means of measurement existed at the time and individual estimates made by those on the spot differ considerably.

With the encouraging news the Poldhu's signals were being strongly received at Crookhaven in Ireland, 225 miles away, Marconi, with two assistants — Kemp and Paget — took passage to St John's, Newfoundland, the nearest landfall in the New World, taking with them large canvas kites and several small balloons with which Marconi proposed to raise the aerial. This latter course of action was decided upon for two reasons; to avoid the public speculation that the erection of tall masts would bring, and to save time.

At St John's all possible assistance was given them by the Governor of Newfoundland, Sir Cavendish Boyle, and the Prime Minister, Sir Robert Bond. Six hundred feet up on the clifftop of Signal Hill, overlooking St John's harbour, was the disused Barracks Hospital; a ground-floor room in this building was placed at Marconi's disposal and here he set up his instruments.

On December 9th a cable was sent to Poldhu instructing the engineers to begin transmissions on the 11th, between 3.0 pm and 7.0 pm GMT. The signals were to consist of repetitions of three dots (the Morse letter 'S'). This letter was chosen because—to quote Marconi himself—'the switching arrangements at Poldhu were not constructed at the time to withstand long periods of operation—especially if letters containing dashes were sent—without considerable wear and tear, and if S's were sent an automatic sender could be employed'.

Across the Atlantic, Marconi and associates erecting a kite-supported aerial to receive transmissions from the high-power station in England.

Heavy gales were sweeping Newfoundland, however, and the next two days were spent in unsuccessful attempts to keep an aerial aloft. A balloon and a kite were lost in these endeavours.

On December 12th a full gale was still blowing, but despite this a kite was flown carrying an aerial to a height of 400 ft. Marconi began a listening watch, using his latest syntonic receiver, but could receive no signals because the erratic movements of the kite were continually altering the angle of the aerial to earth, and therefore its capacity. He decided, therefore, to revert to the older, untuned receiver, using a telephone earpiece in series with the coherer.

Various types of coherer were tried, one of which was the so-called "Italian Navy" device. This is of particular technical interest in that it is described as consisting as a glass tube with a plug of iron at one end and another of carbon at the other, with a globule of mercury between them. The device was self-restoring and had to be used in conjunction with a telephone earpiece. It would seem, therefore, that what is described as a coherer was in fact a true semi-conductor rectifier with either the dissimilar plugs, or oxide film on the mercury, or possibly other surface impurities, performing the rectification process.

At 12.30 pm, Newfoundland time, on December 12th 1901 Marconi heard, faintly but distinctly, the groups of three dots which could only have been emanating from Poldhu, 2200 miles away. He passed the earpiece to Kemp, who confirmed that he had not been mistaken. Paget, to his lifelong regret, was ill on that day and was not present.

The feat was all the more remarkable when it is remembered that the onus was almost entirely on the transmitter, for no amplification was possible at the receiver, and so the received signal itself had to be strong enough to operate the earpiece.

The use of a telephone in place of a recording tape and the absence of any unbiased witness had unfortunate consequences for, immediately the news was made public, a stormy controversy arose as to whether Marconi and Kemp had been deceived into misinterpreting the noise of static as Morse signals. In this matter events conspired against Marconi in that the Anglo-American Telegraph Company, which had a message-carrying monopoly covering Newfoundland, threatened legal action if further experiments were carried out, and so there was no opportunity of giving a public demonstration. But two months later tests were carried out between Poldhu and the liner *Philadelphia* en route from Southampton to New York in which S's were received on the ship at a distance of 2099 miles and these were amply verified by witnesses. Ten months later—in December 1902—two-way communication was effected between Poldhu and a new high-power transmitting station at Glace Bay, Canada—a circumstance made possible by the generous action of the Canadian Government in donating £16,000 towards the cost of the station.

There remained the problem of reconciling the theories of the scientists with the practical results achieved by Marconi. At that time no one knew of the existence in the upper atmosphere of an ionised layer which serves to reflect radio waves and so to make long-distance communication possible. In 1902 Heaviside in England and Kennelly in America independently postulated the existence of such a belt to account for Marconi's achievement but its actual physical existence remained a matter for controversy until the 1920's.

There were, in fact, many unknowns at the time. Until the tests between Poldhu and the liner *Philadelphia* in February 1902 it had not been realised that much longer ranges were obtainable at night. Indeed, it was only then that it was realised that for the transatlantic experiment a listening watch had been kept at the worst possible time of the day! Again, the very success of the operation led to a universal acceptance of the rule 'the lower the frequency the greater the range' and it was not until 1924 that the value of the short waves for long distance communication was realised, largely as a result of the pioneering work of amateurs. The inauguration in 1924 of the Marconi-Franklin short-wave beam-radio service ushered in a completely new era in international radio communication. Incidentally, it was at the Poldhu site that much of the experimental work in connection with short-wave beam transmission took place.

In the same way as Marconi by the introduction of the aerial/earth system had taken wireless waves out of the laboratory into the realm of practical communications, so by the 1901 transatlantic experiment did he introduce the concept of high-power radio engineering and world coverage. And although the spark telegraphy of that day was not electronically generated, it did lead directly to the invention of the thermionic valve and through this to the dawn of the electronics age.

(Courtesy Science Museum, London)

THERMAL IMPEDANCE OF SILICON RECTIFIERS

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PART 2

The thermal resistance R_T from junction to case is equal to the difference between the change in temperature of the junction and of the case divided by the corresponding change in power dissipated at the junction, as follows:

$$R_T = \frac{(T_{J2} - T_{C2}) - (T_{J1} - T_{C1})}{P_2 - P_1} \qquad 7$$

In this equation, the subscript 1 indicates the power level and the temperature corresponding to the steady measuring current I_m (no heating current). The subscript 2 indicates the power level and junction temperature occurring during the I_h-I_m cycle. Equation (7) may be rewritten as follows:

$$R_T = \frac{(T_{J2} - T_{J1}) - (T_{C2} - T_{C1})}{P_2 - P_1} \qquad 8$$

or

$$R_T = \frac{\Delta T_J - \Delta T_C}{\Delta P} \qquad 9$$

In this equation, ΔP is the difference in power dissipation between the steady I_m current and the I_h-I_m cycle. For either level, the power is given by

$$P = \frac{\Sigma VIdt}{\Sigma dt}$$

where V and I are, respectively, the instantaneous voltage and current at time t.

For the current levels shown, P_1 = approximately (0.5 volt) (0.010 ampere) = approximately 0.005 watt, and P_2 = approximately (1.0 volt) (20 amperes) = approximately 20 watts; thus P_1 is negligible compared to P_2.

The quantity T_J is the difference in junction temperature between the steady I_m current and the I_h-I_m cycle. As shown previously by equation (6), ΔT_J is related to the change in forward voltage V_m at current I_m as follows:

$$\Delta T_J = m(\Delta V_m) \qquad 6$$

or

$$\Delta T_J = m(V_{m1} - V_{m2})$$

where m is the slope of the curve of forward voltage drop as a function of temperature (m = 2 mv/°C at low current for the silicon rectifiers measured). Because the quantity needed to obtain ΔT_J is ΔV_m rather than the actual values of V_{m1} and V_{m2} the measurement may be made by use of an oscilloscope. The voltage V_{m1} then establishes a base line or reference voltage level on the oscilloscope, and the shift in this line when the I_h-I_m cycle is applied defines the quantity $V_{m1} - V_{m2}$ or $\triangle V_m$. Although the complete trace of forward voltage drop as a function of time appears as shown in Fig. 12, the scope sensitivity and voltage base level are set so that only the forward voltage drop at current I_m appears on the scope, as shown in Fig. 13.

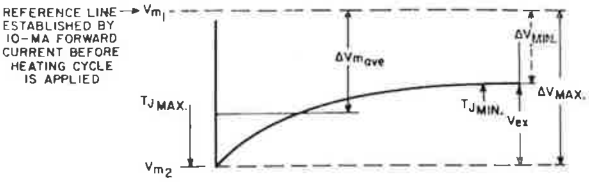

Fig. 13—Forward-voltage characteristic of Fig. 12 as displayed on an oscilloscope using a differential preamplifier.

The quantity $\triangle T_C$ in equation (9) is the change in case temperature between the steady current I_m and the I_h-I_m cycle. This change can be measured by means of a thermocouple attached to the case. Fig. 14 shows temperature curves for a 20-ampere stud-mounted silicon rectifier mounted on a 3″ x 3″ x $\frac{1}{16}$″ copper heat sink. The junction temperature was measured by application of an 8.2-millisecond square wave of 40 amperes peak and 20 amperes average current, and measurement of the change in forward voltage drop at a forward current of 10 milliamperes. Case temperature was measured with a copper-constantan thermocouple peened into a hole drilled in the hexagon flange of the device in the manner prescribed by JEDEC.*

The curves of Fig. 14 illustrate the time required for the heat-sink temperature to stabilize and the relatively constant difference between junction and case temperatures as the case temperature rises. The magnitude of $\triangle T_C$ varies greatly depending on the heat sink on which the device is mounted. Measurements of thermal resistance can be made most accurately when $\triangle T_C$ is kept small. The ideal way would be to use an infinite heat sink on which $\triangle T_C$ would be zero. This method would also permit measurements to be made in the shortest possible time. The time required for device temperatures to stabilize depends on the device size, but is generally less than one minute. The time required for the heat-sink temperature to stabilize may reach several minutes, however, depending on how high $\triangle T_C$ rises and the quantity of heat required to raise the heat sink to this stable condition.

For example, if the temperature rise of the heat sink is negligible at a point x, and if point y stabilizes at some temperature above point x, enough heat must be supplied to raise the temperature of the heat sink between x and y. Because point x is effectively at point y on an infinite heat sink, only enough heat is required to raise the device temperature, and stability is reached much more quickly. Various types of heat sinks have been tried in attempts to approximate an infinite heat sink. Although a copper fin attached to the device is generally unsatisfactory, its effectiveness can be increased by forced-air cooling. Placing the device on a heat sink in a circulating oil bath provides adequate cooling, but is inconvenient. However, a water-cooled heat sink has been found to approximate an infinite heat sink and be fairly easy to use. Such a heat sink was made by drilling passages for water in a copper block, as shown in Fig. 15. With this heat sink, the case-temperature rise is small enough to be neglected and units can be measured fairly accurately without use of a case thermocouple.

Because the type of heat sink used can also affect the temperature distribution over the device, the same type of heat sink and same point of attachment of the case thermocouple should be used when devices are being compared. Tests of two groups of typical 20-ampere stud-mounted silicon rectifiers indicated little correlation of thermal-impedance measurements when different methods were used.

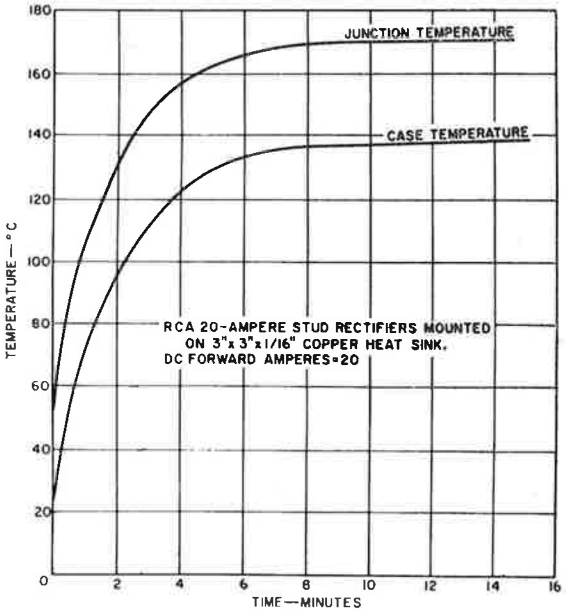

Fig. 14—Curves of junction temperature and case temperature for 20-ampere silicon rectifier mounted on copper fin.

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*Joint Electron Device Engineering Council.

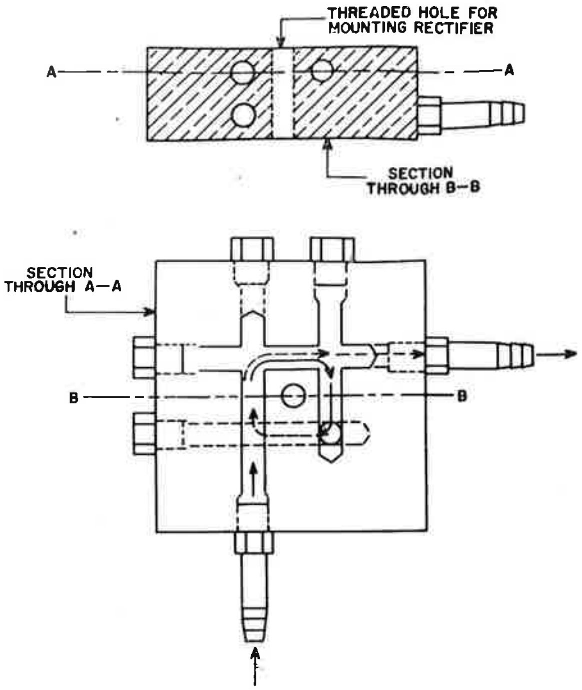

Fig. 15—Water-cooled heat sink used for thermal measurements.

When a heat sink is used in conjunction with an oscilloscope, the procedure for measurement of R_T is as follows:

1. Mount the device on the heat sink.
2. Apply a steady current of 10 milliamperes to the device to establish a reference voltage line on the oscilloscope.
3. Measure case temperature T_{C1}.
4. Begin the I_h-I_m cycle and allow the temperature to stabilize. (Stability is indicated by lack of further change in the forward-voltage-drop curve.)
5. Measure V_h and average forward current.
6. Measure case temperature T_{C2}.
7. Measure $\triangle V_m$ by the change of position of the scope trace.

The thermal characteristics of a typical 20-ampere stud-mounted silicon rectifier might also be measured by use of a cycle consisting of 40-ampere heating-current pulses 8 milliseconds long and 10-milliampere measuring pulses 8-milliseconds long, as shown in Fig. 16. To date, forward-current cycles in which I_h and I_m are equally long have produced the most useful measurement of thermal characteristics of silicon rectifiers. The circuit used for these measurements is shown in Fig. 17. This circuit, together with the use of a water-cooled heat sink, made possible repeatable measurements of much greater accuracy than those obtained with a shorter measuring current.

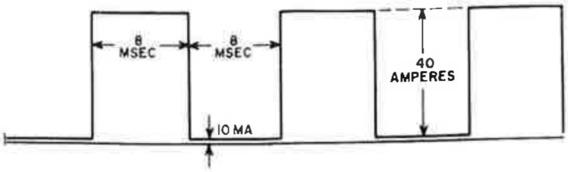

Fig. 16—8-millisecond square pulses used to measure thermal characteristics of 20-ampere rectifiers.

With the cycle shown in Fig. 16, the peak current is 40 amperes and the average current 20 amperes. Consequently, the junction temperature rises as shown in Fig. 10 and finally stabilizes as shown in Fig. 18, with the junction temperature continuing to fluctuate up and down around some average value. Because the heating and cooling portions of the curve are conjugates, the average junction temperature is just halfway between the maximum and minimum temperatures occurring each cycle.

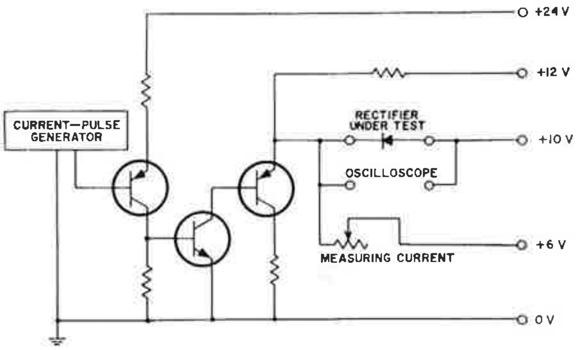

Fig. 17 — Switching circuit used for thermal-resistance measurements.

Because the average junction temperature is equal to the junction temperature resulting from a steady current which produces the same power dissipation in the device as the I_h-I_m cycle, the true thermal resistance must be found from the average power input. The following expression for thermal resistance was given previously in equation (8):

$$R_T = \frac{(T_{J2} - T_{J1}) - (T_{C2} - T_{C1})}{P_2 - P_1} \tag{8}$$

In the earlier discussion, T_{J2} and P_2 were, respectively, the maximum junction temperature and maximum power input. This equation also applies, however, if T_{J2} and P_2 are, respectively, the average junction temperature and average power during the I_h-I_m cycle. Equation (8) still gives the true thermal resistance, the same value that would be determined by a cycle consisting of a long heating current I_h and a short measuring current I_m.

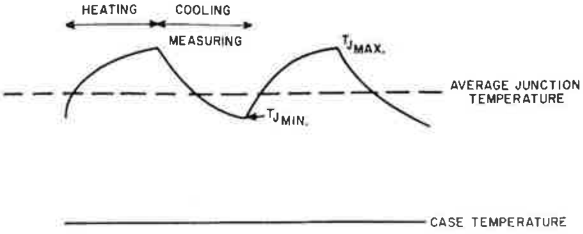

Fig. 18—Temperature-rise curve produced by pulse of Fig. 16.

When the I_h and I_m portions of the cycle are equally long, the curve of forward voltage drop as a function of time appears as shown in Fig. 19. Changes in V_m may be observed by presentation of the measuring portion of this curve on an oscilloscope in conjunction with a high-gain preamplifier, as shown previously in Fig. 13. ΔV_{max}, the greatest change in V_m, is given by $V_{m1} - V_{m2}$. The average value of ΔV_m is given by $1/2(\Delta V_{max} - \Delta V_{min})$, where ΔV_{min} is measured at the end of the I_m cycle.

Thermal resistance R_T is given by

$$R_T = \frac{\Delta T_{J\,ave}}{\Delta P_{ave}} = \frac{m(\Delta V_{m\,ave})}{\Delta P_{ave}} \tag{12}$$

An approximate value for thermal resistance when this cycle is used is given by

$$R_T = \frac{\Delta T_{J\,max}}{\Delta P_{ave}} = \frac{m(\Delta V_{max})}{P_{ave}}\,. \tag{13}$$

Although the value of R_T given by equation (13) is always slightly higher than the actual thermal resistance given by equation (12), it is a useful quantity for comparing one device with another or for evaluating changes in the thermal characteristics of a device. Any such comparisons, of course, require that the same measuring cycle be used each time and for each device.

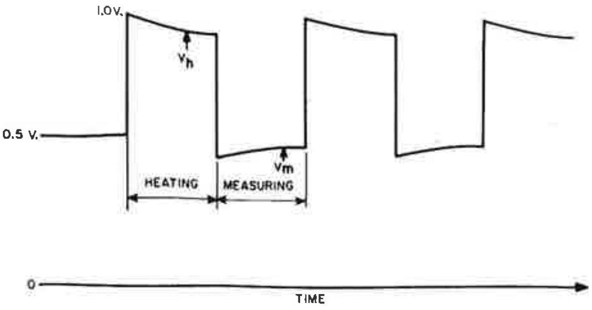

Fig. 19—Curve of forward voltage as a function of time when I_h and I_m are equally long.

DISCUSSION OF TEMPERATURE MEASUREMENTS

Another quantity which may be obtained when the heating current and measuring current are equally long, and which is useful in comparing one unit with another, is the temperature excursion T_{ex} which occurs during each cycle. This quantity is given by

$$\begin{aligned} T_{ex} &= T_{J\,max} - T_{J\,min} \\ &= mV_{ex} = m(\Delta V_{max} - \Delta V_{min}) \end{aligned}$$

If the "on" and "off" times of the squarewave pulse are each 8.3 milliseconds, the resulting data are extremely useful for evaluating thermal cycling because these times correspond to the duty cycle in conventional (USA) 60-cps systems.

In any of these methods, the importance of rapid measurement must be emphasized. The extremely small thermal capacitance and correspondingly small time constant of the semiconductor pellet cause a very rapid temperature drop in the pellet when the heating current stops. Time constants of 85 microseconds have been reported in small-area pellets. Consequently, the true maximum junction temperature is obtained only when the heating current is cut off very sharply and the temperature measurement is made within a few microseconds of this cutoff time.

If there is a delay of the order of several hundred microseconds between the end of the heating pulse and the application of the measuring pulse, the temperature of the junction drops to that of the material on which it is mounted and the measured temperature is not the true peak junction temperature occurring during the heating pulse.

The following examples illustrate the calculation of thermal resistances and capacitances for a typical 20-ampere silicon rectifier:

1. Given a disc of silicon:

$$\begin{aligned}
&\text{radius } r = 0.229 \text{ cm}\\
&\text{thickness } h = 0.0203 \text{ cm}\\
&\text{area } A = \pi r^2 = 0.165 \text{ cm}^2\\
&\text{volume } V = \pi r^2 h = 0.00334 \text{ cm}^3\\
&\text{mass } M = PV = 2.33\ V = 0.00778 \text{ gm}
\end{aligned}$$

Thermal capacitance C_T = (specific heat) (mass)

$$\begin{aligned}
&= (0.162 \text{ cal/g}^\circ\text{C})(0.00778 \text{ g})\\
&= (0.00126 \text{ cal/}^\circ\text{C}) \text{ x}\\
&\quad (4.18 \text{ watt-sec/cal})\\
&= 0.00527 \text{ watt-sec/}^\circ\text{C}
\end{aligned}$$

Thermal resistance R_T = (thickness)/(area) (thermal conductivity)

$$\begin{aligned}
&= h/AK_t = 0.0203 \text{ cm}/(0.165 \text{ cm}^2)\\
&\quad (0.837 \text{ watt-cm/cm}^{2\circ}\text{C})\\
&= 0.148\ ^\circ\text{C/watt-mil}\\
&= 1.18\ ^\circ\text{C/watt}
\end{aligned}$$

2. Given a disc of solder (99% Pb):

$$\begin{aligned}
&\text{radius } r = 0.229 \text{ cm}\\
&\text{thickness } h = 0.00254 \text{ cm}\\
&\text{area } A = 0.165 \text{ cm}^2\\
&\text{volume } V = 0.000419 \text{ cm}^3\\
&\text{mass } M = PV = 0.00475 \text{ g}
\end{aligned}$$

Thermal capacitance C_T =

$$\begin{aligned}
&(0.031 \text{ cal/g}^\circ\text{C})(0.00475 \text{ g}) \text{ x}\\
&(4.18 \text{ watt-sec/cal})\\
&= 0.00616 \text{ watt-sec/}^\circ\text{C}
\end{aligned}$$

Thermal resistance R_T =

$$\begin{aligned}
&(0.00254 \text{ cm})/(0.165 \text{ cm}^2) \text{ x}\\
&(0.347 \text{ watt-cm/cm}^{2\circ}\text{C})\\
&= 0.0447\ ^\circ\text{C/watt-mil}
\end{aligned}$$

A thermal time constant for a silicon pellet such as that shown in Fig. 20 may be determined as follows: It is assumed that the entire pellet is at a uniform temperature T above the base temperature, and that heat can be transferred out only through the base. Because the heat is distributed throughout the pellet, it may be assumed that the heat must pass through one-half the thermal resistance of the pellet and through the entire thermal resistance of the solder. As in the electrical analog circuit described earlier, the time constant T is given by

$$\begin{aligned}
T &= R_T\ C_T\\
&= (1/2\ R_{Tp} + R_{Ts})(C_{Tp} + 1/2C_{Ts})\\
&= (0.0735 + 0.0447)^\circ\text{C/watt x}\\
&\quad (0.00527 + 0.00031) \text{ watt-sec/}^\circ\text{C}\\
&= (0.1182)(0.00558) \text{ second}\\
&= 0.00066 \text{ second}\\
&= 660 \text{ microseconds}
\end{aligned}$$

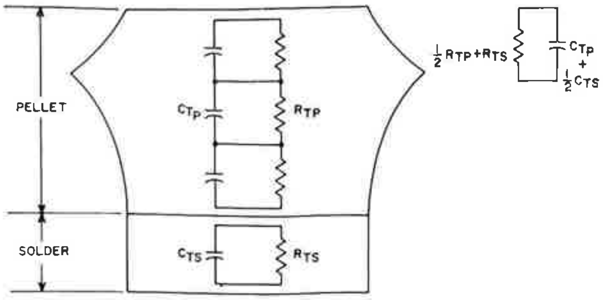

Fig. 20—Representation of silicon pellet showing quantities used in electrical analogy.

In an actual device, the copper disc on top of the pellet increases this time constant if the thermal resistance between disc and pellet is low because the disc then increases the thermal capacitance of the pellet.

Considerable difficulty has been encountered in attempts to get a sharp enough current cutoff. In a circuit which used power transistors to switch a current of 40 amperes on and off with an 8-millisecond-on, 8-millisecond-off square-wave pulse, the current was found to decay as shown in Fig. 21. The current decays exponentially, decreasing by half its value every 100 microseconds. Consequently, there is a delay of approximately 1.8 milliseconds between the instant the 40-ampere current begins to decay and the instant it becomes smaller than the 10-milliampere measuring pulse.

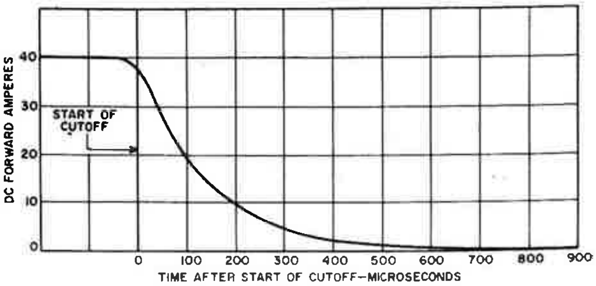

Fig. 21 — Current-decay curve for 40-ampere pulser.

In this transistorized pulser, the switching equipment sets a minimum limit on the time delay which must occur between the heating pulse and the measuring pulse. However, a minimum time delay is also imposed by the lifetime of the carriers in the device to be measured. When the heating current is cut off, a large number of free carriers remain in the junction region. These carriers decrease the forward voltage drop at the measuring level until they either recombine or are pulled out of the crystal. For example, if the pulse is switched from a 40-ampere level to a 10-milliampere level in a device in which the minority-carrier lifetime is 10 microseconds, approximately 110 to 120 microseconds must elapse before the number of carriers left by the 40-ampere current decays to a level which is negligible compared to the 10-milliampere level. For material having a longer or shorter lifetime, the delay time is correspondingly longer or shorter.

Excess carriers might be removed in a shorter time if the 10-milliampere pulse is applied immediately after the heating pulse to sweep them out in the forward direction, or if a short inverse voltage pulse is applied immediately after the heating pulse to sweep them out in the reverse direction.

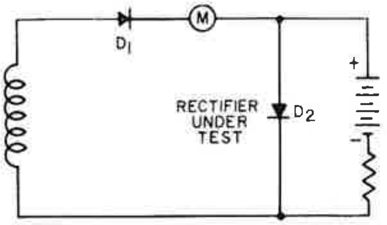

Fig. 22—Thermal-resistance test circuit utilizing forward characteristics of silicon rectifier.

Although the minimum temperature-measurement delay time is limited by both the time required for the switching transistors to turn off and the time required for carriers to be removed from the base region of the device being measured, these two quantities have opposite effects on the forward voltage drop. The delay in turn-off time causes the forward voltage drop to be higher than if it were governed by temperature alone because it prevents the current from dropping immediately to level I_m from level I_h. The carriers remaining in the junction region of the device, however, cause the forward voltage drop to be lower than if it were governed by temperature alone because these excess carriers are available to keep forward current flowing with less than the normal voltage drop.

Fairly accurate measurements of peak junction temperature can be obtained in spite of the carrier-lifetime and switching transients by extrapolation from the voltage curve after these transients have decayed. If the measuring pulse is long enough, for example, V_m may be measured at 200 microseconds, 400 microseconds, and 1 millisecond after the heating current begins to decay; the measured values at these points may then be extrapolated to the minimum value of V_m. Such a method is necessarily limited in its accuracy because the temperature drop is not a true exponential decay and because of heat generated during the current decay.

A pulser using controlled rectifiers has recently been found to have a much shorter switching time. With this pulser, the switching time to date seems to be limited only by the lifetime of the carriers in the rectifier under test.

DISCUSSION OF THERMAL IMPEDANCE

Many different methods have been used to measure thermal resistance. In the method shown in Fig. 22, a half-wave rectifier D_1 and a battery are connected to the rectifier under test D_2 in the forward direction. The rectifier under test receives half-wave heating current at the rated forward current of D_1; the battery supplies the small level of measuring current. The difference between the initial forward voltage drop with no high current through the device and the forward drop with rated forward current gives the change in temperature. The equation for calculating the thermal resistance R_T is given below:

$$R_T = \frac{\Delta T}{P} \qquad (15)$$

$$P = (V_o - M\Delta T) \times I_f + (I_{rms})^2 R_p$$

where V_o is the dc forward voltage drop (volts), m is the slope of the forward-voltage-vs.-temperature curve (mv/°C), and R_p is the dynamic resistance of the slope of the forward-voltage-vs.-current curve (ohms).

Calculation of power by this method is laborious. Values obtained are in error to a varying degree according to the thermal time constants of the device. The temperature calculated from the voltage at the low current-injection level does not correspond to the peak temperature in the junction because peak temperature occurs close to the peak of the rated forward-current cycle and the thermal time constant of silicon rectifiers is much shorter than the half-cycle of rated current.

Another method for measuring thermal resistance or thermal impedance utilizes the reverse characteristics of the rectifier. The rectifier is connected in the test circuit as shown in Fig. 23. The rectifier under test, D_2, is first operated at rated forward current, rated case temperature, and rated peak inverse voltage (PIV). The average reverse current I_R and case temperature T_{C1} are recorded. The forward current is then cut off, and with the rectifier operated at rated PIV the case temperature is

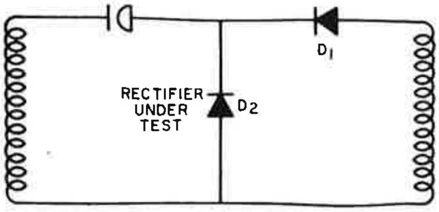

Fig. 23 — Test circuit utilizing reverse characteristics of rectifier.

increased until I_R reaches the previously recorded value. The case temperature T_{C2} at this point is also recorded. The resulting difference in case temperature is assumed to be equal to the difference between junction and case temperature when the unit is operating with rated forward current.

Thermal resistance is then expressed as follows:

$$R_T = \frac{\Delta T_C}{P}$$

where ΔT_C is the difference in case temperatures ($T_{C1} - T_{C2}$), and P is the forward power dissipation as given by equation (15).

This method has several disadvantages and is not as accurate as methods based on forward characteristics. The leakage current depends upon junction temperature, among other things, and if the thermal resistance or thermal impedance were excessively high, then the leakage current would be excessively high. The disadvantages are apparent. First, considerable time is required for the rectifier to reach equilibrium and to return to the previous leakage current. Second, if a high leakage is recorded, it is more probably due to a poor surface condition than to temperature alone. Third, as discussed previously, the peak temperature occurs at the peak of the forward-current half-cycle rather than the peak of the inverse half-cycle, where the measurements are taken. Consequently, what is measured is the junction temperature which exists during the "off" cycle, not the actual peak junction temperature.

The "null" method for testing thermal resistance uses the circuit shown in Fig. 24. The equation for thermal resistance of a transistor or rectifier under test is given by

$$R_T = (T_J - T_x)/P$$

where T_J is the junction temperature, T_x is the reference temperature (case, heat sink, or ambient), and P is the power input.

The forward current through the test device is supplied at room temperature by a constant-current generator adjusted to produce a voltage drop of 0.5 volt. The device is then heated in a high-temperature oil bath so that the forward-voltage drop can be measured at an elevated junction temperature. (The junction temperature is the same as the bath temperature because the power dissipation in the device is very low.) The device is then removed from the oil bath, and a steady value of forward current is applied to produce the same forward-voltage drop obtained in the oil bath. This forward current is interrupted for 100 microseconds at a rate of 200 times per second.

THERMAL FATIGUE

The results obtained from thermal-resistance tests are extremely important in classifying and applying semiconductor devices. In most industrial applications, rectifiers are required to operate for a minimum of 10,000 hours. In constant operation from a 60-cycle power line, therefore, a rectifier is subjected to 2,160,000,000 cycles. One of the most serious problems encountered with medium and high-current rectifiers is thermal fatigue, caused by the different coefficients of expansion of the metals and materials used in the construction of the devices. The lower the temperature excursions in the rectifier, therefore, the higher the number of thermal cycles before failure is caused by thermal stresses.

It is extremely important that the user of silicon rectifiers recognize the symptoms of thermal fatigue. Too often the cause of failure of a semiconductor device is attributed to a "transient". The initial electrical measurements which are usually made on silicon rectifiers (i.e., forward voltage drop and inverse electrical characteristics) have no bearing on thermal fatigue. The following

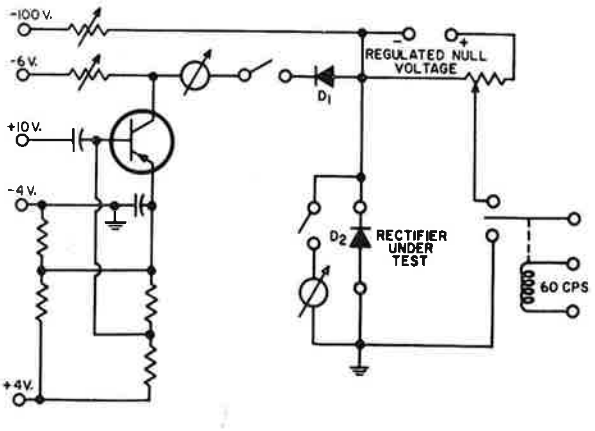

Fig. 24 — Null method for testing thermal resistance.

measurements should be made to investigate thermal fatigue:

1. Instantaneous forward voltage drop should be measured at a current 5 to 6 times the average current rating of the device.
2. Thermal impedance should be measured with a reliable system.
3. Temperature excursion should be measured with a given on - off pulse time.

These measurements should be made both initially and after the unit has been operated in service or subjected to a number of temperature cycles. An increase in forward voltage drop indicates that the rectifier is dissipating more heat. Consequently, the initial cooling method may no longer be adequate and the rectifier may be damaged. An increase in thermal impedance indicates that the thermal connection between the silicon junction and the base of the unit is becoming fatigued, provided the inverse characteristics of the device remain unchanged. If the reverse characteristics also change, the reason for failure may or may not be fatigue.

CONCLUSIONS

The thermal impedance in semiconductor devices is analogous to the impedance in equivalent electrical circuits. Consequently, the effects of equivalent resistance and capacitance are extremely important. Temperature excursions in a device can be minimized by the addition of a maximum amount of thermal mass (which is equivalent to capacitance) in very close proximity to the junction, with a minimum increase of thermal and electrical resistance.

Tests are still being conducted to determine the optimum values and most satisfactory circuit for measuring thermal resistance. The major problem has been to switch rapidly enough from the high current level used for heating to the low level of current used for measurement. The effect of left-over minority carriers also poses a problem. Changes in forward voltage drop must be measured within a few microseconds after the heating pulse has stopped. Differences in change of the forward voltage drop must be measured accurately in millivolts. The case temperature must also be measured, and allowance made for case temperature rise.

ACKNOWLEDGMENT

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(With acknowledgements to RCA)

VALVES FOR HIFI

PART II — POWER-OUTPUT TYPES

By M. Y. Epstein

RCA Electron Tube Division, Harrison N.J.

The major requirements for an audio power valve are high efficiency, good power sensitivity, low distortion, and low internal valve impedance. The valve impedance shunts the plate load, and a low impedance is effective in damping out undesirable output peaks caused by the large resonant response of the loudspeaker at certain frequencies. This feature suggests the use of a triode rather than a pentode as an output valve. However, the plate potential of a triode is at the lowest value of its voltage swing just when the greatest cathode current is required, as shown in Fig. 1-a. A pentode, on the other hand, always has a large, fixed value of screen-grid voltage available to draw current from the cathode regardless of how the plate voltage varies, as shown in Fig. 1-b. Consequently, the cathode size (and hence the heater power) of a triode must be significantly greater than that of a pentode supplying equivalent current.

Because the cost of a power valve is almost directly proportional to its cathode size, the more efficient pentode or beam power valve is almost invariably used in modern-day audio systems in preference to the triode. The advantage of the triode's low internal impedance is thus sacrificed, but loudspeaker damping for the high-

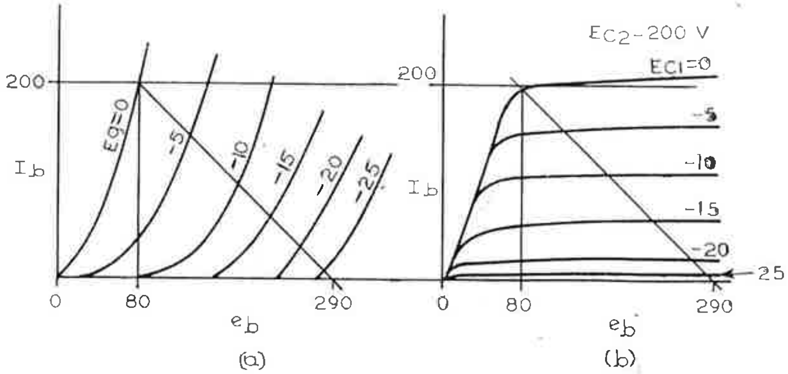
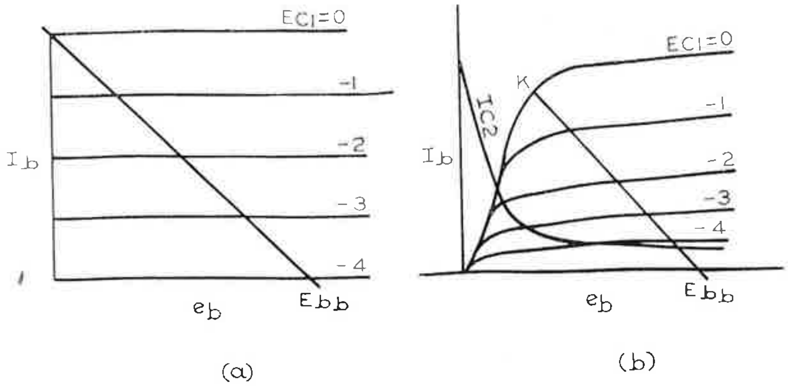


Fig. 1—Comparison between triode and pentode power output valves.



(a)

(b)

Fig. 2—Comparison between a perfect pentode and a practical pentode.

internal-impedance pentodes is easily obtained by feedback networks and by external resistance-capacitance shunting of the loudspeaker load.

Power Output and Efficiency

Fig. 2 shows plate characteristics for a theoretically perfect pentode and a practical pentode. Power output is readily calculated from these curves as follows: $P.O. = (E_{max}-E_{min})(I_{max}-I_{min})/8$. For increased power output and efficiency of the practical valve, it would be necessary both to lower the knee voltage and to reduce the current that is lost to the screen grid. The knee voltage has a finite value because the negative charge of the electrons passing through the screen grid produces a potential dip in front of the plate. At low plate potentials, the attractive force of the plate is not strong enough to pull all the electrons through this space-charge barrier, and many return to be picked up by the screen grid. At plate potentials higher than the knee voltage (the voltage at which the plate overcomes the space-charge dip), all electrons passing through the screen-grid wires reach the plate. Screen-grid current exists even at high plate potentials, however, due to the cathode current that flows directly into the screen-grid wires.

Beam power valves operate at higher efficiencies than pentodes because the alignment of the control grid and screen grid greatly reduces both the direct interception of current by the screen grid and the tendency of the electrons to return to the screen grid at low plate potentials[1], as shown in Fig. 3. The electron lens formed by the aligned grids in a beam power valve is designed so that the electrons emitted from the cathode are formed into beams that come to a relatively sharp focal point in the plane of the screen grid between the screen-grid wires. Furthermore, the shape of the electron beams causes the electrons to pass into the region beyond the screen grid along well-directed paths so that there is little tendency for the electrons to turn around and return to the screen grid. This feature increases the effectiveness of the plate potential in assisting the electrons in their forward direction, and thus lowers the knee voltage. The many electron beams combine uniformly in front of the plate to form a space-charge-potential minimum that prevents the flow of secondary-emission electrons from the plate to the screen grid.

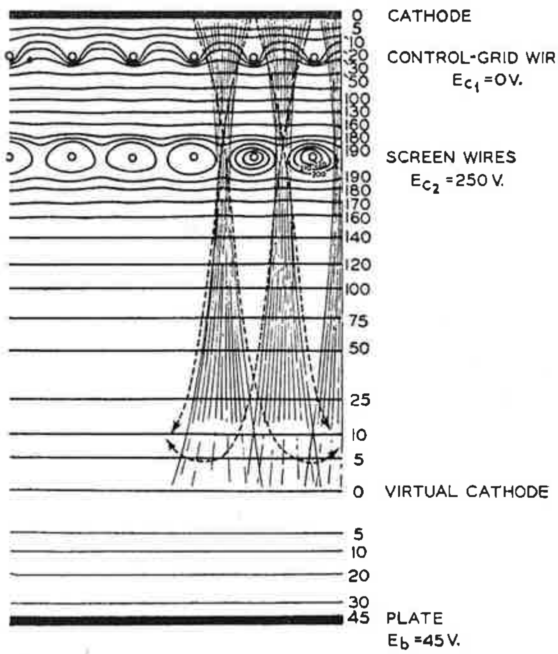


Fig. 3—Alignment of grid lateral wires and focusing of electron stream in a beam power valve.

In conventional pentodes, little effort is made to direct individual electrons and, consequently, many electrons flow directly into the screen-grid wires or follow curved paths that cause the electrons to return to the screen grid even after they have passed through it. The randomly directed electrons combine nonuniformly in front of the plate so that a special suppressor grid is required to prevent the flow of secondary electrons from the plate to the screen grid.

Many studies have shown a strong inverse relationship between valve life and dependability and valve temperature. The higher the valve temperature, the greater is the danger of valve failure from grid emission, insulation breakdown of the glass between the electrode lead-in wires, and cathode poisoning from gases evolved from the hot electrodes.

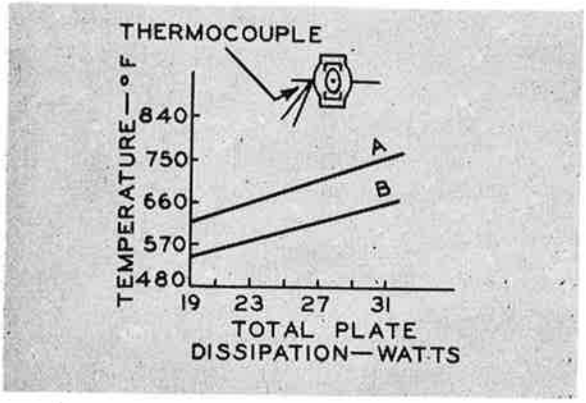


Fig. 4—Temperature curves for plates using carburized-nickel-clad steel (A) and copper-cored steel (B). All parts seven thousandths of an inch thick.

Within the lower power-output ranges, heat removal from the valves is generally no problem, and small power pentodes enjoy a cost advantage over beam power valves because the alignment of the grids in beam power valves requires extra care and effort. Within the higher power ranges, however, heat removal is a problem, and the cost of providing means for such removal becomes a major factor in over-all valve cost. Consequently, in spite of the relatively costly grid-alignment process, the efficient beam power valves are less expensive to fabricate than are the less efficient pentodes.

The 6L6-GC and 7027-A beam valves are specially designed to produce high power output with long life and dependability at low cost. High voltage ratings are achieved by the use of special high-resistance glass and careful spacing and insulation of the electrodes. Large quantities of heat are drained from the valves by radiation from large fins and from the black-body plate, and by conduction through low-thermal-resistance electrodes and connecting leads.

Both the 6L6-GC and the 7027-A plates use 11-mil, carburized nickel-clad steel instead of the 6- or 7-mil material normally used in RCA receiving power valves. Fig. 4 shows a temperature comparison between plates using 7-mil ca burized-nickel-clad steel and 7-mil copper-core steel.[2] The plate material used in the 6L6-G and 7027-A gives results comparable to those obtained with copper-cored steel, but is less expensive. Because the electron stream impinges on the plate only over a small portion of its area, hot spots tend to form on the plate that give off cathode-poisoning gases. The thicker material used in the 6L6-GC and 7027-A rapidly dissipates the heat and prevents hot-spot formation. Use of 11-mil plate material on the 6L6-GC increased its possible plate dissipation by 40 percent as compared to that of the 6L6-GB.

The control-grid lateral wires of both the 6L6-GC and the 7027-A are made of high-work-function and low-thermal-resistance gold-plated molybdenum to lower the temperature of the grid and to prevent grid emission.

Power Sensitivity

For a given power output, the necessary grid-driving voltage and the required voltage amplification of the input signal are largely determined by the power sensitivity ($P.S. = P.O./E_g{}^2$) of the output valve. In low-cost systems where each stage of the complete system is pushed to its maximum output, the available driving voltage is often limited and high-power-sensitivity output valves are essential. In higher-priced systems, however, power sensitivity usually plays a secondary role to valve efficiency and distortion.

Part I of this article discussed the *"inselbildung"* effect in reference to microphonics. Actually, this phenomenon plays a major role in many aspects of valve design, and is especially pertinent to a discussion on power sensitivity.

When the ratio of control-grid-to-cathode spacing (D_{g1k}) to control-grid-turns pitch ($1/TPI$) becomes less than unity, the electrostatic fields at the surface of the cathode lose their uniformity and "islands" of potential are formed, as shown in Fig. 5. Consequently, electrons emitted from different sections of the cathode are exposed to different electrostatic fields and, in effect, are controlled to varying degrees by the instantaneous potentials on the control grid.

Measured valve characteristics, therefore, are actually the summation of many tiny valve sections in parallel.

The family of curves in Fig. 6 shows the variation of amplification factor (μ) over the length of the cathode. These curves are normalized with respect to the composite calculated amplification factor. As would be expected, the cathode sections furthest from the grid turns are the least controlled by the grid, i.e., have the lowest values of amplification factor (μ). As the control grid becomes more negative, therefore, the electron emission of the high-μ sections of the cathode is cut off quickly, and the valve characteristics are established by the sections which are more difficult to control. Thus, for close-spaced valves, μ is hardly a constant, but varies in some inverse manner with grid bias. The effects of this variation on power sensitivity can be explained as follows:

Plate current and grid voltage are related primarily by the following expression:

$$I_b = f \frac{\left(-E_g + \dfrac{E_{c2}}{\mu} \right)^{3/2}}{(D_{g1k})^2}$$

Where, I_b is the plate current, E_g is the control-grid voltage, E_{c2} is the screen-grid voltage, μ is the triode amplification factor, and D_{g1k} is the control-grid-to-cathode spacing.

At maximum power output, the amount of power output depends on plate-current swing, or maximum current at zero bias. (It is assumed that the valve is cut off at the other end of its grid-voltage swing.) Power sensitivity is determined both by the power output and by the grid-signal voltage. Thus, to raise power sensitivity (with a given cathode and heater power) it is necessary to raise the power output or to lower the signal voltage required to drive the valve from zero bias to cut-off (for maximum output). Thus, E_g may be considered as the voltage required to cut off the valve; from the above equation, $E_g = E_{c2}/\mu$.

Raising the cutoff value of μ to lower E_g unfortunately also tends to lower I_b and P.O. A decrease in D_{g1k} would increase I_b and P.O., but because of the "inselbildung" effects noted above, would tend to lower the cut-off μ and thus require an increase in E_g again. There are two different μ values of concern here. One is the value of μ at cutoff, which affects E_g; the other is the value of μ at zero bias, which controls I_b and *P.O.* The two are related by means of the geometry of the valve (Fig. 2).

Thus, a decrease in D_{g1k} tends to lower the cutoff μ and requires an increase of the grid turns per inch (TPI) to maintain the a/b ratio parameter of Fig. 6. Higher TPI, of course, improves the shielding effects of the grid and undesirably raises the zero-bias μ. This effect may then be countered by the use of fine grid lateral wire. The net result is that, in order to raise power sensitivity, it is necessary to reduce the electrode spacings and to use high TPI and fine-wire grids. Limitations, then, on possible power sensitivities occur mostly from the mechanical problems of making the various parts and of combining them into a structurally sound valve.

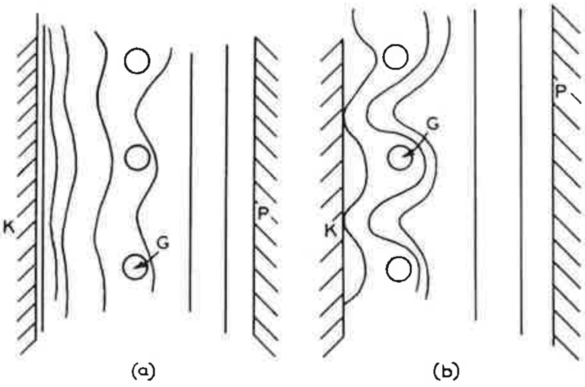


Fig. 5—Illustration of the "inselbildung" effect.

Beam power valves are far more sensitive to these restrictions than pentodes because of their need for proper alignment of the lateral wires of the control and screen grids. The limitations imposed by present grid-making and valve-assembly techniques are so severe that the TPI of beam-valve grids can be only about half that of grids available for pentodes.

With present techniques, the valve designer has just about reached the mechanical limits in his attempts to improve power sensitivity. New techniques are required and are being developed. The most spectacular new development, of course, is the nuvistor. As a power valve, the nuvistor will be a beam power tetrode or pentode whose unique construction will permit the use of automatically aligned grids with relatively small electrode spacings, small grid pitch, and fine-wire grids. For conventional structures, new grid forms, such as the frame grid, are being developed which will also permit automatic alignment with higher-TPI grids, and higher power sensitivity.

Distortion

The most significant aspect of the high-fidelity audio system and hence the high-fidelity valve is, of course, its distortion level. Distortion is generally described in terms of percentage of total output, i.e., total distorted output divided by total output. However, total distortion consists of many component parts, some more objectionable than others. For example, the normal push-pull type of output circuit tends to cancel even-order harmonic distortion present in the valve outputs. Furthermore, high levels of even-order harmonic distortion are not especially objectionable, whereas even small amounts of odd order harmonic distortion are objectionable.

Vacuum valves are inherently non linear devices because the valve currents vary with the electrode voltages to powers other than one. However, the valve designer can control the power to which plate current varies with control-grid voltage to some extent by proper grid-geometry design. In addition, under dynamic conditions, the choice of load line will determine the amplitude relations between the control-grid voltage and the plate voltage. Together, these two factors may be used to shape the dynamic transfer characteristic of the valve and hence to determine the nature of distortion that the valve will generate.

Again, the "inselbildung effect" is significant. Severe changes in curvature of the tail of the valve transfer characteristic are caused by the variation in amplification factor as the high-μ sections of the cathode are suddenly and completely cut off with increasing negative grid voltage, as described above. Because operation of the valve in this region of the transfer curve produces extra distortion of high harmonic order, the valve designer must compromise between high power output at low low distortion (obtained by high a/b ratios and cutoff μ's) and high-power-sensitivity valves with high-μ, high-TPI grids operating in the "inselbildung" regions.

The 6L6 beam power valve was originally designed to produce a square-law dynamic characteristic when operated at its maximum power-output point (i.e., with its load line intersecting the knee of the e_b-i_b family of curves), and its harmonic content is composed almost exclusively of the second harmonic. The 6L6-GC, 7027-A, and 6973 are similar in this respect to the original 6L6 and, therefore, provide excellent low-distortion power output when operated in push-pull. At maximum levels of output, these valves can be operated with less than 2-percent total distortion.

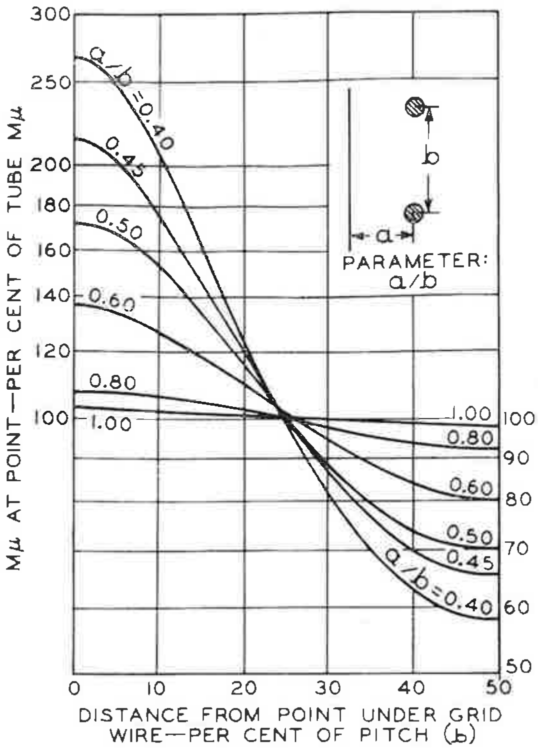


Fig. 6—Variation of amplification factor along the length of the cathode.

In pentodes, unfortunately, the grid geometries necessary to provide high power sensitivity tend to produce odd-order harmonic distortion when the load lines are chosen to produce maximum power output. When sufficient negative feedback voltage to reduce distortion is used, however, high-sensitivity pentodes provide satisfactory performance in low-cost, good-quality audio systems.

The difference between pentode and beam-power-valve performance is shown by a comparison between the 6973 and the 6BQ5. The 6973 beam power valve is designed for high-efficiency, high-quality, medium-power systems. Although the 6BQ5 pentode, which operates in the same power range, has four times the power sensitivity of the 6973, the 6973 has lower distortion, higher valve efficiency, and lower heater power than the 6BQ5. At the relatively low power levels at which these valves operate, however, valve cost and over-all system cost are largely determined by the power sensitivity, with valve efficiency playing somewhat of a secondary role.

Acknowledgment

The author wishes to thank T. Boyer for his helpful discussions and assistance.

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(With acknowledgements to RCA)

TRANSISTOR MARKINGS

Transistors of the encapsulated flying lead type, such as the 2N217, 2N408 and so on, have in the past been marked with a red dot adjacent to the collector electrode to assist in identifying the leads and their connections. Transistors will now be marked EITHER with the red dot as heretofore, OR with the letter "C" adjacent to the collector lead.

READING BOOKS

From time to time we review in these pages new books on electronics produced by various publishing houses. These books are examined by engineers with many years experience in the industry, and in many cases the matter in the books is something that they first met many years before, and have seen many dozens of times since then. This could be the case for example where the book deals with fundamental aspects of a subject.

This of course does not imply that the books are without merit. On the contrary, the standards of knowledge available to us are constantly improving, and new books are called for that contain the new data, relate it to former information, and use the ever-improving presentation techniques we see today. To illustrate this, we have only to look at a high school physics book of pre-war vintage and compare it with the modern equivalent.

How ever many times we read the same facts in different books, we can almost always deepen our understanding of the subject, very often because a new book presents the matter in a different viewpoint. In the same way, re-reading a book always shows a dividend of increased understanding. This was expounded very neatly in a bookmark received with a recent submission for review; the text is reproduced here. in its entirety, with acknowledgements to the McGraw-Hill Book Coy.:

HOW TO LEARN

"This book was written so that you might learn. In order to learn from it, however, you must know how to use it properly. The book will teach you little unless you put organized effort into reading it, for active, directed work is necessary if you wish to understand and remember what you read. Careful, intelligent reading of this book will mean that you understand better what you learn in class. What is equally important, if you have studied this book properly in the first place, it will serve as a convenient and quick refresher for future reference. Nearly everyone knows that we easily forget what we learn when we do not use it. What many students do not realize, however, is that we can relearn what we have once learned, providing we have learned it well in the first place. Thus, this book provides a convenient auxiliary memory that can serve you all of your life.

In order to understand and remember the contents of this book it is essential that you do more than read. It means that you must actively recite, question, and review the material you have read. See the reverse of this card for suggestions that will help you to study this book. By following these suggestions, you will find the book will be more valuable to you both in the course in which it is assigned and as a part of your permanent library.

HOW TO GET THE MOST OUT OF A BOOK

1. Skim through the assigned reading so that you will know what it is you are to study.
2. Read the text carefully. Do not forget that many important ideas are presented in graphs, diagrams or maps.
3. As you read, stop now and then and recite to yourself, in your own words, the important ideas in what you have just read.
4. Make brief notes in the margin. These will serve as cues for subsequent self-recitation.

5. Mark important or key passages for later review.
6. Review the material at least once between the first time you study the assignment and study for exams. Make use of your marginal notes as cues for self-recitation.
7. Remember that a little relearning is necessary each time you wish to use what you have learned for an examination, a related course, or for independent study. If you use the author's headings, marked passages, and brief notes for cues it will help you relearn easily.
8. Coordinate what you read with what you learn in the classroom. Keep well-organized lecture notes. Lecture notes that are legible and accurate will, like your text-book, serve in the years to come as quick and inexpensive keys to the knowledge that you are acquiring".

Now that was a fine bookmark to give away with the book, because it forms a key to the book itself, or to any other book for that matter.

In conclusion, let us go to one of the most famous of all authors in the English language; Sam Johnson, who had something to say on most things, said this about books:

"Books have always had a secret influence on the understanding; we cannot at pleasure obliterate ideas: he that reads books of science, though without any desire fixed of improvement, will grow more knowing; he that entertains himself with moral or religious treatises, will imperceptibly advance in goodness; the ideas which are often offered to the mind, will at last find a lucky moment when it is disposed to receive them".

The section on "How to Get the Most out of a Book" was written by James A. Deese, Associate Professor of Psychology in the John Hopkins University, and author of "The Psychology of Learning".

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