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DAMPING FACTOR

One of the features of audio amplifiers around which argument has centred over the years has been the question of damping factor, and how large the factor should be. At the same time, the types of speakers commonly used with high fidelity systems have changed, putting perhaps a different slant on the discussion. But this is running ahead somewhat; let us go back first of all to see what is meant by damping factor, and what effect it has in practice.

Definition

The damping factor of an amplifier is defined as the ratio between the load (loudspeaker) impedance and the source impedance (output impedance of the amplifier). But such a simple definition is very far from telling anything like the complete story.

In the case of an amplifier driving a loudspeaker, the obvious objective is that movement of the speaker should follow exactly the electrical impulses fed into it from the amplifier. It will be convenient for the purpose of this discussion, although not completely accurate, to separate the electrical impulses we are thinking about into (a) simple variations of frequency in a steady-tone input, and (b) transients.

In the case (a), failure of the speaker to response is attributable to a frequency response that is less than perfect; this feature of the speaker is of no interest to us in this context. In the case (b), we would like a speaker subjected to a transient pulse to follow the pulse accurately. We will find cases where the coil and the cone (or part of the cone where flexure occurs) cannot be moved from rest quickly enough. Conversely, when the transient is over, the moving assembly cannot be brought to rest quickly enough. The result of all this will naturally be considerable distortion of the transient waveform, and when we consider that a great deal of music is transient in character, the importance of this matter will be obvious.

The over-shoot at the end of the transient will occur at the natural frequency of the speaker or resonant element, and not at the frequency of the applied signal. In a loudspeaker, the worst offender is the resonant element formed by the mass of the cone and coil, and the stiffness of the cone suspension. This gives a pronounced resonance at a frequency determined by the expression:

$$f = \frac{1}{2\pi}\sqrt{\frac{\text{stiffness of suspension in dynes/cm}}{\text{mass of cone and coil in grammes}}}$$

This figure is more familiar to us as the natural free-air resonant frequency of the speaker.

At this resonant frequency, the amplitude of movement of the cone may be 20 times or more greater for constant input current than it is at other frequencies removed from the resonant point. It is interesting to note in passing that this effect can make the speaker appear to have a better bass response, judged solely from a subjective point of view, than will be found in normal steady-tone frequency response tests. The experienced ear, however, will readily detect the spurious effect from the "soft and flabby" sound, as opposed to the firm and precise sound that should be heard.

It is well-known that over-shoot of the type mentioned can be ameliorated by the judicious use of damping, which acts in much the same way as the shock-absorbers in an automobile. Damping on a loudspeaker arises in three ways, all of which are present at the same time, but in different proportions in each case. The damping is partly acoustical, partly frictional and partly electromagnetic.

It is the last-mentioned in which we are mainly interested in this note, although the other two will be mentioned again later. Electromagnetic damping can be improved by feeding the speaker from an amplifier which has a low output impedance. Whilst this will generally give an improvement at the onset of a transient, it is after the transient is over that the most important improvement will result. If the end of a transient is considered, and we have the case where the speaker cone does not instantly come to rest, a decaying oscillation at the speaker's natural resonance will follow, very much like a tennis ball dropped onto a hard surface. Whilst this oscillation is going on and there is no driving signal from the amplifier, the speaker cone is generating a voltage which is being fed back into the output of the amplifier. The current which will flow is determined by the speaker constants and by the output impedance of the amplifier. The higher the current that is allowed to flow, the quicker will be the decay of the damped wave, because energy will be absorbed from it quicker.

Whilst we cannot do much about the dc resistance and the inductance of the speaker itself, we can make an improvement in reducing the output impedance of the amplifier, and this is where damping factor becomes important.

How Much Damping?

The main point about damping factor on which complete agreement still seems to elude us is in deciding just what the damping factor should be. Different authorities have differing views on the magnitude of the damping factor in terms of the speaker characteristics, and others have quoted desirable values for the damping factor ranging from 5 to 100 or more. Why so much disagreement?

In order to untangle the skein of thought, it will be useful to consider some the pertinent aspects, and then see whether they can be used to reconcile some of the points of difference.

We know, for example, that each speaker possesses dc resistance, so that whatever we do outside the speaker, the damped wave output resulting from overshoot will be limited by that resistance. This would appear to render futile some of the efforts that have been made over the years to improve the situation by using amplifiers with zero output impedance. (This is generally done by a combination of negative and positive feedback.)

Langford - Smith, in "Radiotron Designer's Handbook," gives us an expression for the output impedance of an amplifier in terms of the reduction of the Q of the resonant element of the speaker, where the constants of the speaker are known or can be measured. He quotes a value of 0.5 for critical damping, with two authorities to support him, but also admits that some engineers take a value of 1 for critical damping. However, he also states that a reduction of the output impedance of the amplifier below one-fifth of the resistance of the voice coil, transformer secondary and connecting leads has little effect on the Q of the circuit, and goes on to say that there is no advantage in using greater than critical damping. Although it is only stated in this way by inference, there is in fact a danger that if we go beyond critical damping, the situation will worsen. It is important to note that this authority calls for a determination of damping factor in terms of a specific speaker.

Another well-known authority, Moir, in "High-Quality Sound Reproduction," states that critical damping is generally secured when the output impedance of the amplifier is one-fifth the dc resistance of the speaker voice coil. We have now found two highly-respected authorities that support this view. It is now clear that if the speaker itself is to determine the output impedance of the amplifier, this could explain many of the differing values of damping factor loosely mentioned as being desirable. It is also clear that it may be better to measure damping factor in terms of the dc resistance of the voice coil rather than in terms of the nominal voice coil impedance, but there are obvious practical difficulties here.

There are one or two other factors that have appeared in more recent years which could alter the desirable magnitude of the damping factor, if not the basic theory behind the derivation of the correct value. Some years ago, the tendency was to use speakers on flat baffles, then enclosures of increasing degrees of complexity appeared, and now we have high-quality low-efficiency speakers mounted in infinite baffles or similar enclosures that do not add to the bass response and afford better acoustical and frictional damping. Coupled with this progress, higher flux densities have been generally introduced, and this, by increasing the voltage fed back to the amplifier during overshoot, results in a more rapid decay and tends to call for a lower value of damping factor. All this means that today, we are likely to require in general lower values of damping factor than were necessary in the past.

Whilst in the past the tendency was to make an amplifier with the highest possible damping factor and hope for the best, this is certainly not

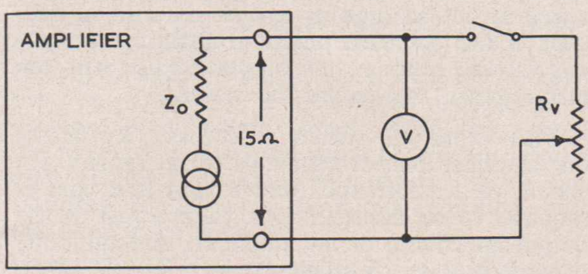

Fig. 1

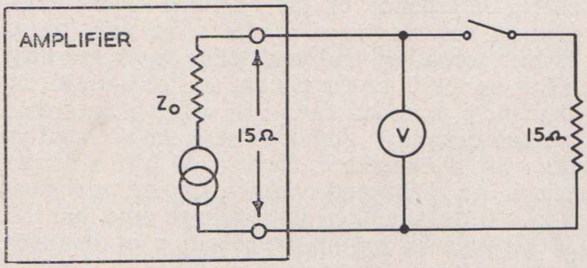

Fig. 2

the case today, and the magnitude of the factor becomes less and less meaningful as a measure of quality, unless related to a specific combination of amplifier and speaker. This is yet another reason for the complete system concept so often mentioned in these pages. Too-high a damping factor will have one serious effect that will be noticeable, in reducing the bass response of the system, and must have accounted for many cases where a speaker that sounded good on one amplifier was not so good when used with a different amplifier.

Is it possible that one day speaker makers may specify an optimum damping factor or amplifier output impedance for use with their units? It is realised that this could be difficult, partly because the final results are evaluated subjectively by the user. Just as some like different settings of tone controls, so some could like different values of damping factor, or at least the different response curves, particularly in the bass region, that would result from their use.

Measuring Damping Factor

Having, one may hope, aroused or renewed interest in the subject of damping factor, the next logical step is to see how this property of an amplifier can be measured. It is, in fact, one of the easiest characteristics to measure, and we have a choice of two satisfactory methods. They are known as the "variable resistance method" and the "voltage regulation method."

Variable Resistance Method. The basic scheme of this method is shown in Fig. 1. The normal load for the amplifier is removed, and a signal is fed into the amplifier at any convenient frequency. Care must be taken to see that the input signal is not so high as to run the risk of arcover in the unloaded output transformer or output valves. Measure the output voltage with the meter shown and record the reading.

Then close the switch to connect Rv across the amplifier output terminals. The resistor Rv should be a substantial wire-wound unit, and is required to be adjusted to a value equal to the output impedance of the amplifier. A total value of perhaps 8 to 15 ohms should suit most cases. The resistor Rv is now adjusted until the voltmeter reading falls to one-half of the unloaded reading already recorded. We now know from basic theory that the adjusted value of Rv is equal to the output impedance of the amplifier.

Resistor Rv is now disconnected and accurately measured. Let us assume in this case that the measured value is 1 ohm. The damping factor of the amplifier is then the nominal load impedance (15 ohms) divided by the value of Rv, giving a value of 15 for the factor.

Just in case an amplifier is encountered that has a negative output impedance, in this case the output voltage will rise when the switch is closed and Rv is connected across the output terminals. Adjust Rv until the output voltage rises to twice the original value; the measured value of Rv will then be equal to the NEGATIVE output impedance of the amplifier. In this case, somewhat higher values may be required for Rv than were indicated above.

Difficulty may be experienced in using this method in certain cases. For example, if the amplifier has a high damping factor and/or the nominal load impedance is low, it may be difficult to adjust Rv to the required value. Further, with very low values of resistance connected across the output transformer secondary, the primary impedance of the transformer falls and excessive plate current may flow in the output stage.

Voltage Regulation Method. This method will yield the same answers as the previous method, without the disadvantages mentioned. It is the method which has been standardised in most laboratories, including our own, and is the method called for in testing procedures laid down by the Audio Group of the British Radio and Electrical Manufacturers' Association and by the U.S.A. Institute of High Fidelity Manufacturers.

The general arrangement of this test is shown in Fig. 2. With the switch open, a signal is applied to the amplifier, taking care not to make the signal level too high, as mentioned before, and the output voltage is measured. Call this voltage E1. Then the switch is closed, and a resistance equal to the nominal load impedance for the amplifier is connected across the terminals.

The amplifier is now correctly loaded, and a second voltage reading is taken, which may be called E2. The damping factor of the amplifier is equal to the loaded voltage divided by the difference between the no-load and loaded voltages, that is, damping factor equals E2(E1—E2).

The test just described is, as the name tells us, a test of amplifier regulation. The lower the output impedance, the better the regulation and the higher the damping factor.

Page 168, please.

TWO WARNING LIGHTS

Introduction

We have received a number of enquiries of late for what may perhaps be described as a transistorized light flasher. Several such circuits have appeared in the literature, so it was decided that a slightly different approach might fill the need, and also make the item a little more interesting. We therefore came up with the idea of a small unit fitted with a red lamp and a long pair of leads fitted with battery clips. The prime thought was that this could be used as a safety warning lamp, operated from a car battery, where the car was immobilised at night.

Such a situation might arise, for example, where it was necessary to change a wheel and there was no space to pull off the carriageway to do so. Oncoming vehicles at night would need as much warning of the obstruction as possible. We ultimately came up with two units, one of them using one light only, and the other using two lights which flash alternately.

Whilst these units are suitable for the purpose described, the basic circuit is capable of being adapted for several purposes where automatic periodic switching is required. The use of a relay with either unit would extend its usefulness still further. This note, however, describes the units as originally conceived, and leaves it to readers to adapt the circuit as required.

UNIT NUMBER 1

This is the simpler of the two units, and consists of a lamp housing with red lens, mounted on a small steel box, with two terminals. The circuitry of the unit is accommodated inside the box. If it is required or desirable to make the unit completely weather-tight, one could cut a gasket for the lid of the box. The accompanying photographs show the general arrangement. Whilst it was thought that the warning light could be stood on the roadside alongside the disabled vehicle, with the red lens either at the top or facing towards the rear, the unit could just as easily be placed on the rear parcel shelf (in sedans), hung on the rear bumper bar, or if provided with a felt pad glued to the underside, could be placed on top of the roof. An adequate length of lead should be provided.

Circuit

The circuit of this model is shown in Fig. 4, and consists of an n-p-n small-signal transistor type 2N647 (or 2N649) and a power transistor type 2N301, in a simple direct-coupled amplifier circuit. The lamp that is to be switched forms the collector load for the power transistor. In this circuit, the signal at the base of the n-p-n transistor is in phase with the signal at the collector of the power transistor, so that positive feedback will take place through the RC combination connected between these two points.

To study the operation of the circuit, let us imagine that the lamp is off. The potentiometer is advanced sufficiently towards the positive supply line to allow a small current to flow through the n-p-n transistor. This will in turn bring the power transistor into conduction. Positive feedback will now cause both transistors to move rapidly into the saturated condition, and the lamp will light.

With the lamp on, the feedback capacitor is being charged, and the current supplied to the base of the n-p-n transistor through the RC combination is falling. A point is reached where the current through the capacitor falls too low to hold the n-p-n transistor in saturation, resulting in a decrease in collector current. With the aid of the feedback circuit, the small change of current now rapidly transfers the amplifier from the saturated condition to the cutoff condition, and the lamp goes out.

The circuit is now held in the cutoff condition by the charge on the capacitor, placed there

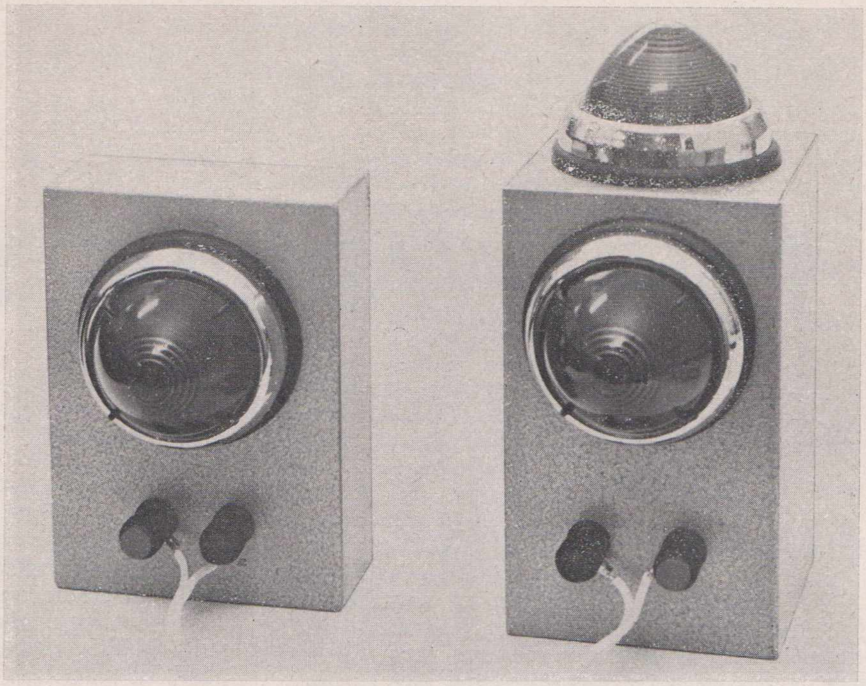

Fig. 1—This photograph shows the two warning light units described in this note.

whilst the lamp was lit. The circuit will remain in the off condition until the charge on the capacitor leaks away sufficiently to allow a small current flow in the n-p-n transistor. The circuit then re-cycles.

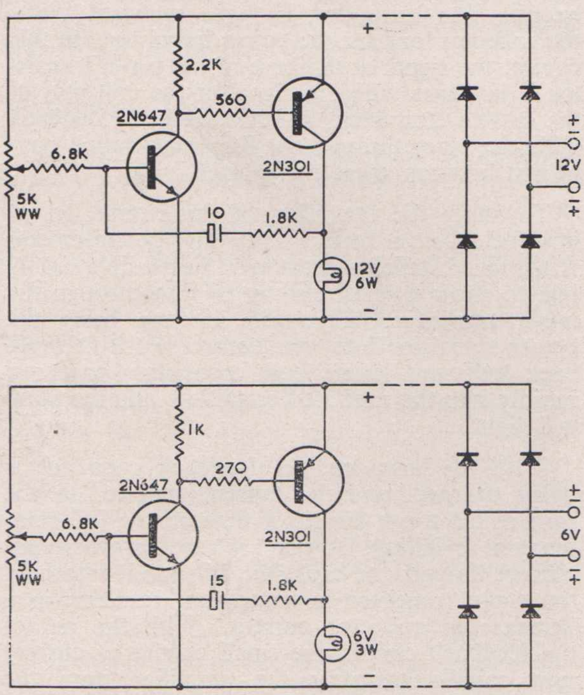

Fig. 2—Circuit diagram of the single-lamp unit with optional polarity corrector. The upper circuit shows the 12-volt version and the lower circuit the 6-volt version.

The initial adjustment of the circuit to make it operate is carried out at the potentiometer, which should be advanced towards the positive line until the lamp starts to blink. There is then an area over which adjustment of the control gives a small variation of the on/off times of the lamp, but if the control is taken too far in the positive direction, the lamp will remain on. The values shown in the circuit are safe for all settings of the control.

The values shown in the circuit give a flashing rate of about once per second, which seemed a reasonable rate based on a subjective test. The rate may be changed by altering the time constant of the feedback elements, an increase in the time constant giving a longer period. In order to protect the n-p-n transistor from excessive current, the resistor used in the feedback circuit should not be reduced below 250 ohms in the 6-volt circuit and 500 ohms in the 12-volt circuit. Making the resistor too large may stop the unit completely.

The circuit shows our unit in both 6-volt and 12-volt versions. The current drawn in the on condition is usually a little less than half an ampere, the wattage rating of the bulbs being of course only nominal. The emitter current of the n-p-n transistor in the on condition is of the order of 3.5 ma in the 6-volt version and 8.5 ma in the 12-volt version.

There is only one other small point in relation to the circuit, and that is the four diodes shown at the input. These diodes are optional. Inspection of the arrangement will show that the purpose of the four diodes is to make the unit un-

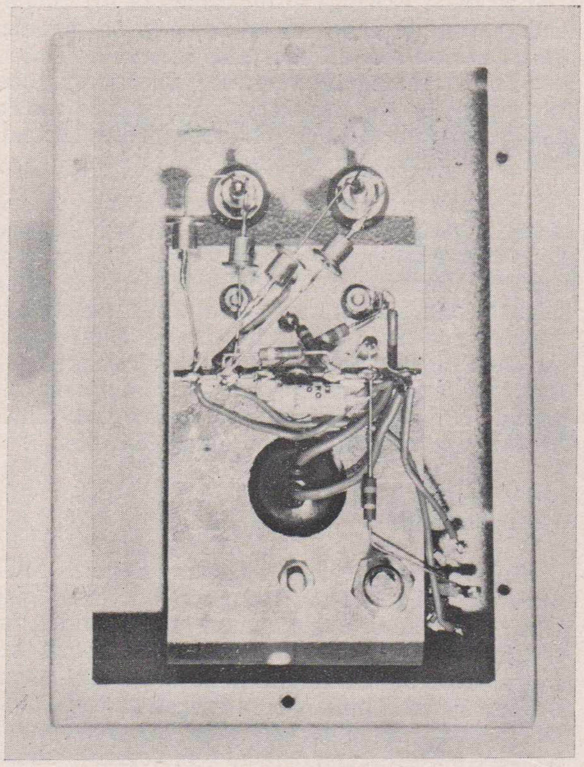

Fig. 3—Interior view of the single-lamp unit, showing the general arrangement of the components.

critical as to the polarity of the input voltage. This is a little device frequently used by us as a protective measure on equipment under development, where frequent connection and disconnection of the dc power supply could lead to accidental damage as a result of connecting the equipment with the wrong polarity. In this case, it was thought that some people might appreciate the idea, as the lamp unit could be quickly connected up in the dark without regard to polarity. Any 500 ma silicon diodes could be used here, the forward voltage drop being negligible.

Construction

We constructed the unit in a Heating Systems steel box type MC42, with a Campbell CC25/D/R lamp housing. This lamp housing has a double-contact holder, so that the electrical circuit can be kept isolated from the box in which the unit is constructed. Two heavy terminals were provided for the input lead, although a captive lead run into the box would do as well.

The two bolts holding the lamp housing to the box were made long so that they could also support a small metal plate inside. The potentiometer and the power transistor are mounted on this small plate, together with a tag strip to support the small components. The dissipation in the power transistor is small, and a larger heat sink is not required.

If desired, it would be possible to make up this unit with its own dry battery as a self-contained warning lantern, in which case a heavy-duty battery of the lantern type should be used. Doubtless other variations and applications will occur to readers.

UNIT NUMBER 2

This unit is a little more ambitious than the previous one, in that it uses two lamps, which are lit alternately at intervals of about one second. It follows, of course, that this unit is a little more expensive to make, but could be more useful in certain circumstances. In this case, a larger steel box is used for the case, and we mounted one lamp on the side of the box and one on the top. This, however, is a matter of discretion for the user.

Circuit

The circuit diagram of the double-lamp unit is shown in Fig. 5, and consists of two n-p-n transistors in a symmetrical multivibrator circuit, which in turn control two power transistors. The basic operation of a multivibrator is so well known today that there is little point in explaining the operation in detail.

As usual, the timing of the circuit is controlled by the values of the base and collector resistors of the n-p-n transistors, and the values of the coupling capacitors. A variation of these values

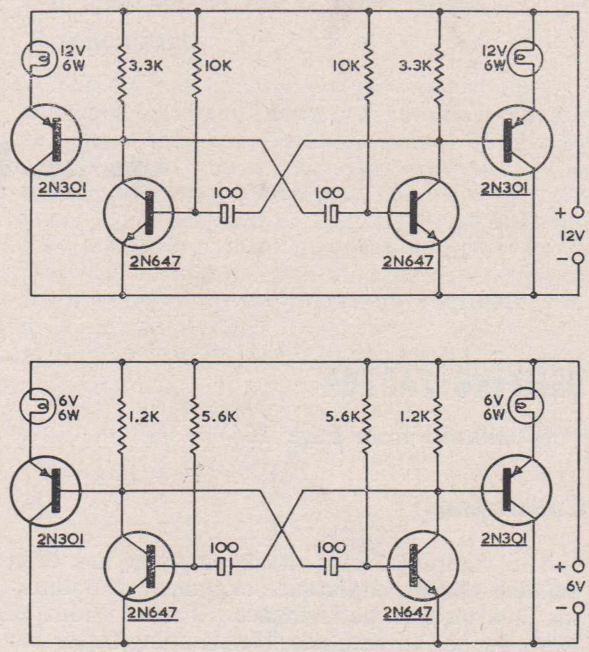

Fig. 4—Circuit diagram of the double-lamp unit. The 12-volt version is shown at the top and the 6-volt version at the bottom.

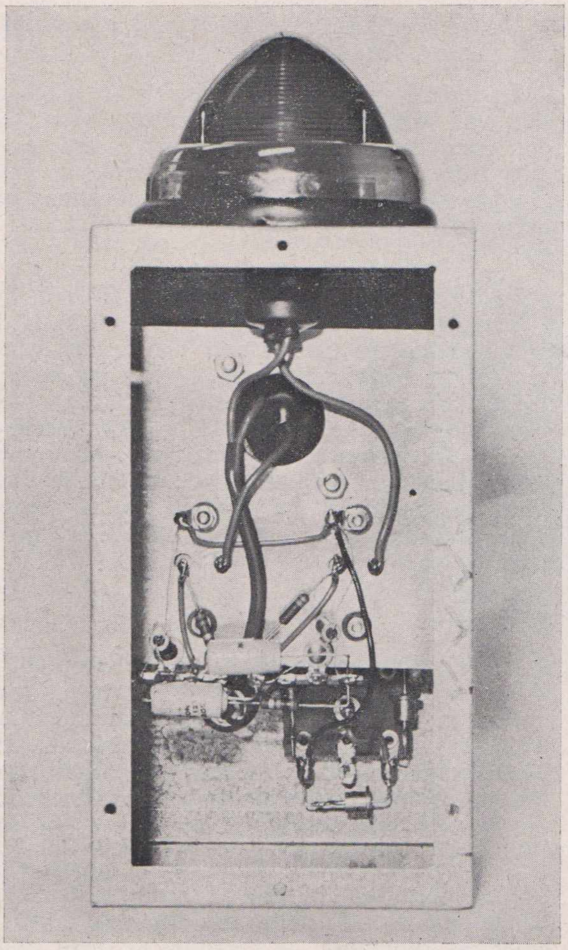

Fig. 5—Interior view of the double-lamp unit.

will produce a change in the timing of the circuit, and the circuit can also be adjusted so that the lamps are on for different time intervals. This is done by arranging different time constants for the two multivibrator transistors. The values shown allow each lamp to light for about half a second, and this gives a good subjective result.

The polarity corrector shown in the circuit diagram of the single-lamp unit can also be used with the two-lamp unit if desired. The current drain of this unit in the condition with one lamp alight is of the order of 700 ma, whilst the emitter currents for the n-p-n transistors is about 2 ma, the same figures applying to both versions of the unit. If experimenting with the timing of this circuit, it is advisable not to reduce the values of the base resistors below about 500 ohms.

Construction

This unit is built into a Heating Systems MC4 metal case, two of the red lamp housings being used. As before, two terminals are provided for the input lead. The two power transistors are mounted on an aluminium plate of 18 swg, about 4 in. x 3½ in, which is so arranged as to be held by the two mounting bolts for the front lamp housing. The accompanying photograph shows the arrangement.

Most of the remarks made about the single-lamp unit will apply also to this one, both as regards modifications and applications. Note, however, the higher operating current for the two lamp unit if attempting to operate it on dry batteries.

DAMPING FACTOR

(Continued from Page 164)

Conclusion

The damping factor of an amplifier has been defined, and its importance explained in optimising loudspeaker performance. The old idea of going for the highest possible damping factor has been discounted in favour of adjusting the amplifier output impedance, and therefore the damping factor, to suit the loudspeaker being used. Two eminent authorities have been mentioned for those who wish to read further into the subject, and each of those in turn quotes a large number of sources of further reading. Two simple methods of measuring damping factor have been explained, the latter being the preferred method.

CURRENT LIMITING FOR TRANSISTOR SERIES VOLTAGE REGULATORS

By H. T. Breece III

(RCA Semiconductor and Materials Division)

Introduction

The increased reliability, reduced size, and low cost of transistors makes them increasingly desirable for series voltage regulator applications. One of the problems encountered in the design of series transistor voltage regulators is protection for the series control element from excess dissipation resulting from current overloads, short circuits, and capacitive loads. This note describes several methods of current limiting for series transistor voltage regulator circuits.

General Information

Overloading most series voltage regulator circuits results in permanent damage to the series control transistor. As an example (Figure 1) when the output terminals are shorted the full input voltage and current capability are applied to the series control transistor. This usually is many times greater than the dissipation ratings of the series transistor.

A series fuse is sometimes used in an attempt to protect the series transistor from this excessive dissipation; however, a series fuse is usually not capable of providing the necessary protection under all overload conditions. This is because the thermal time constant of the fuse is normally much greater than that of the transistor.

Protection for all overload conditions may be accomplished through the use of a circuit which limits the current to a safe value, determined by the dissipation rating of the series regulator transistor.

To be effective, current limiting circuits must respond fast enough to protect the series transistor and yet be capable of returning to normal regulator operation as soon as the overload condition is removed. It is desirable to achieve current overload protection with minimum degradation of regulator performance.

One method of achieving current limiting is to use a resistor in series with the regulator transistor. The large resistance normally required, however, wastes a great deal of power and degrades the regulator performance.

Applications

The current limiting section (dashed lines) of the regulator circuit shown in Figure 2a is designed to appear as a large series resistance during current overload and a negligible resistance during normal operating conditions. The value of resistance, R_5, is designed so that during normal regulator operation, transistor, Q_4, operates in the saturated condition. For the overload condition, R_4 is adjusted so the maximum allowable overload current through it produces a voltage drop high enough to cause D_1 to conduct. Conduction

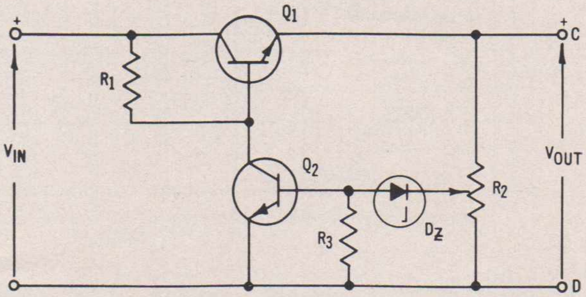

Fig. 1

of D_1 reduces the bias to Q_4, making the transistor appear as an increasing series resistance in the regulator circuit.

Under short-circuit conditions, the entire value of input voltage (V_{in}) appears across Q_4 simultaneously with the limit value of current. Transistor Q_4 must be capable of withstanding the resulting dissipation. When the limit is reached, the junction temperature of transistor Q_4 rises to a value considerably above the ambient temperature. This increase in junction temperature causes the value of short circuit current to drift slightly upward due to the inherent variation of V_{be} with temperature in transistors. This effect is minimized by mounting the silicon rectifier D_1 and transistor Q_4 on a common heat sink so their respective junction temperatures may reach the same value, since the values of their respective V_{be} and forward voltage drop temperature coefficients are comparable.

Performance characteristics for the transistor series voltage regulator of Figure 2a are shown in Figure 2b.

Although the circuit of Figure 2a provides adjustable current limiting with simple circuitry and minimum power loss during normal operation, it does have the disadvantage of requiring a second series transistor capable of withstanding short circuit output current and total input voltage simultaneously.

In many high-current and high-voltage regulator circuits, it is necessary to use parallel or series connections to obtain the required voltage and current ratings for the series control transistor. The method of Figure 2a may not be practical in this application due to the additional series transistors required.

The circuit shown in Figure 3a eliminates the need for an additional series transistor by using the series regulator transistor as the current limiting element. This method is very effective when a Darlington connection is used for the series control transistor. A desirable feature of this circuit in high-current regulators is that it functions well, even with R_4 adjusted to zero.

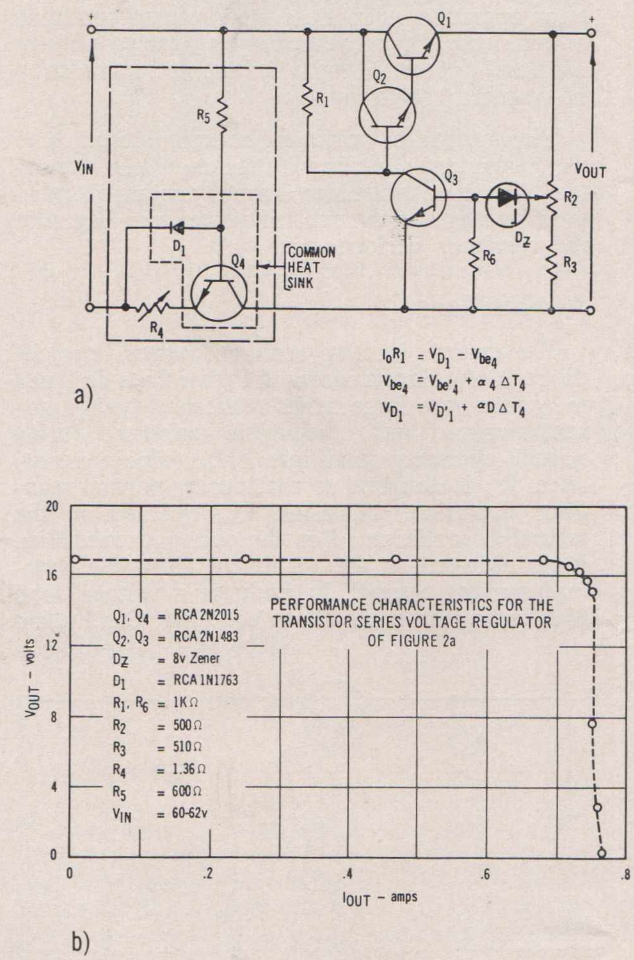

Fig. 2

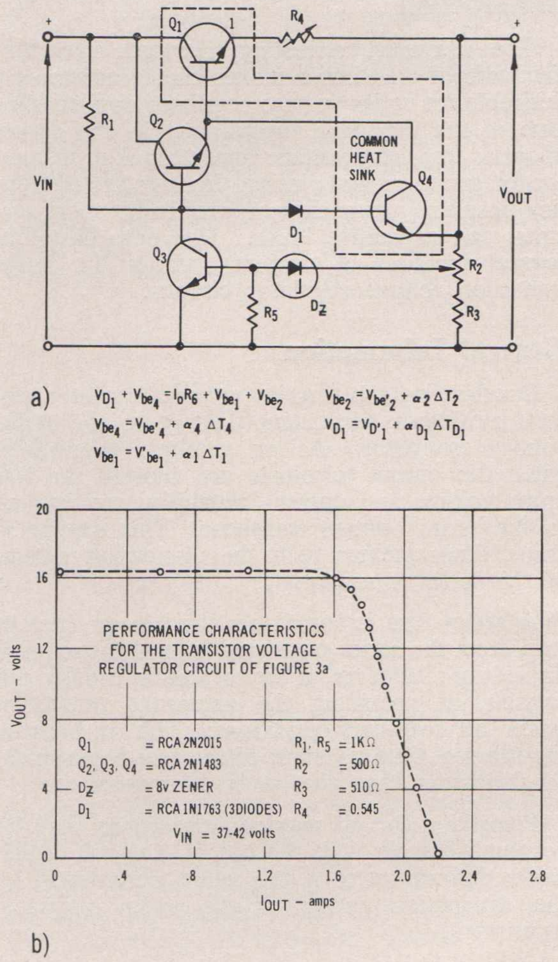

Fig. 3

Current limiting is achieved by the combined action of the components shown in the dashed lines. The voltage developed across R_4 and the base-to-emitter voltages of Q_1 and Q_2 are proportional to the circuit output current. During current overload, these voltages add up to a value great enough to cause D_1 and Q_4 to conduct. As D_1 and Q_4 begin to conduct, Q_4 shunts a portion of the bias available to the series regulator transistor. This, in turn, increases the series resistance of Q_1.

The value of current in the circuit of Figure 3a is adjusted by varying the value of resistance, R_4.

Higher current ranges may be obtained by increasing the number of rectifiers represented by D_1. The temperature drift problem is minimized by mounting transistors Q_1 and Q_4 on a common heat sink. Performance characteristics for this circuit are shown in Figure 3b.

Figure 4a shows a circuit similar to that of Figure 3a. Current limiting is adjusted by varying R_4 and by changing the number of silicon rectifiers represented by D_1. Temperature drift is minimized by mounting the series control transistor Q_1 and rectifier D_1 on a common heat sink.

Performance characteristics for this circuit are shown in Figure 4b. The circuits of both Figures 3a and 4a are applicable to high-current and high-voltage regulators since additional series power transistors are not required.

Figure 5a shows a current limiting circuit consisting of transistors Q_1, Q_2, Q_4, D_1 and R_4. Again, the regulator series control transistor is used as the current limiting element.

The series element must be capable of withstanding input voltage and short-circuit current simultaneously. The value of short-circuit current is selected by adjusting the value of resistor R_4. Performance characteristics of this circuit are shown in Figure 5b. The circuit functions equally well with resistor, R_4, located in the positive output lead.

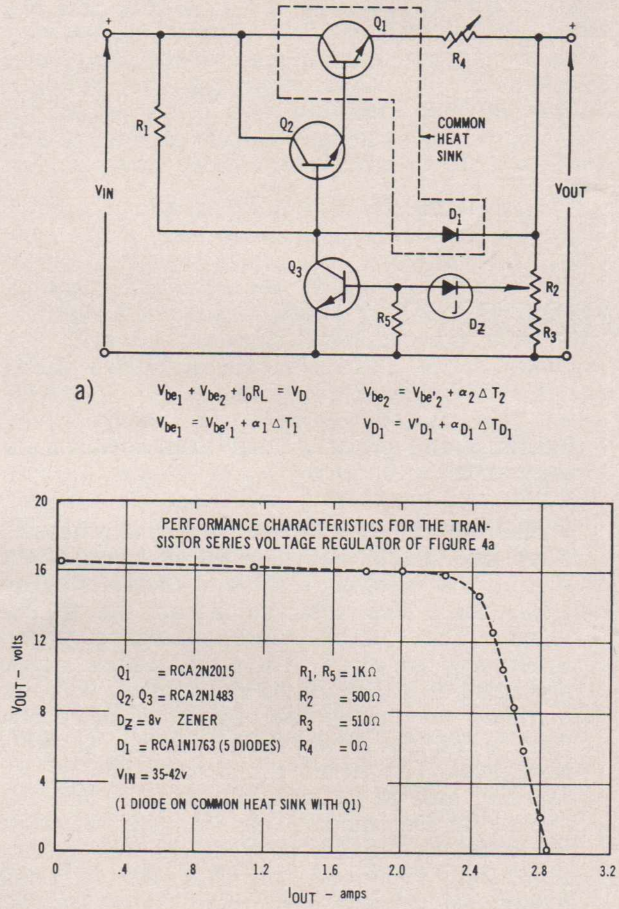

Fig. 4

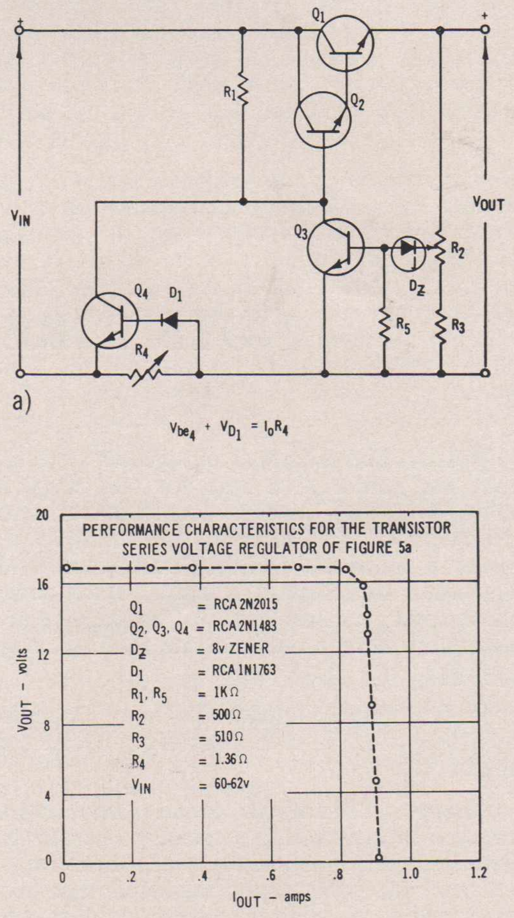

Fig. 5

Conclusions

The circuits described in this Note provide adjustable current limiting with minimum degradation of the normal regulator operation and automatic return to normal regulator operation upon removal of the current overload condition.

The circuit of Figure 2a provides the most desirable performance, as is shown in Figure 2b, but requires an additional series transistor.

The circuit of Figure 5a has good performance characteristics, but requires a somewhat larger value of series resistance than the circuits of Figure 3a and 4a, which will operate with zero series resistance.

The choice of a current limiting circuit depends upon the particular application, as to desired performance characteristics, available power, and available number of components.

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New address stencils now coming into use will carry information regarding the duration of the subscription, allowing readers to check for themselves when their subscriptions are due to terminate. These stencils will be used progressively as renewals are made. The stencil carries a code consisting of two numerals and a letter. The two numerals are the last two figures of the expiry year, whilst the letter shows the last monthly issue due. The letters commence with "a" for January, and go through, omitting "i" and "l", to "n" for December. Thus, for example, code 64j indicates the last copy due to be that for September, 1964. At this date, only a small number of readers have these coded stencils, but it is expected that all will have them shortly.

4: Multiplier Phototubes

Part I

Construction and Principles of Operation

Although **photoelectric emission** is a relatively efficient process on a per-quantum basis, the **primary photocurrent** for low light levels is so small that special amplification techniques are required for most applications. The multiplier phototube, which uses **secondary electron emission** to provide current amplification in excess of 10^6, is a very useful detector for low light levels.

In a multiplier phototube, the photoelectrons emitted by the photocathode are, in general, electrostatically directed to a secondary emitting surface called a **dynode.** When normal operating voltages are applied to the dynode, 3 to 6 secondary electrons are emitted per primary electron. These secondaries are focused to a second dynode, where the process is repeated. In addition to 6 to 14 dynodes, a multiplier phototube may contain other electrodes for focusing the electron stream, reducing space-charge effects, or accelerating the electrons to reduce transit-time effects. The last dynode is followed by an anode which collects the electrons and serves as the signal-output electrode in most applications.

Dynode Properties

Secondary emission[1–3] in many respects is similar to photoelectric emission. The impact of primary electrons rather than incident photons causes the emission of electrons. One primary electron excites several low-energy secondary electrons near the surface of the emitter; some reach the surface and overcome the work function (in the case of metals) or the electron affinity (for semiconductors) and escape into the vacuum. In general, the number of secondaries created increases as the primary electron energy is increased. However, the depth from which the electrons must escape also increase as the primary energy increases because of the greater primary penetration; this factor tends to reduce the number of secondaries at higher dynode voltages.

Fig. 30[4] shows the number of emitted secondary electrons σ per primary electron (secondary emission coefficient) as a function of the energy of the primary electrons for a number of practical dynode materials. The same general pattern is observed for metals, but the yield is insufficient for use in multiplier phototubes.

An important property of the secondary electrons is the **energy distribution** of the emitted secondaries. A typical distribution curve is shown in Fig. 31. The peak at the extreme right corresponds to the energy of the primary electrons, and probably represents elastically scattered primary electrons. The true secondaries are represented by the peak at the left. Although the spread of secondary-electron velocities of good secondary emitters is generally much less than that shown in Fig. 31, it is nevertheless large in comparison with that of the photoelectron velocities. This velocity spread dictates to some extent the type of electron optics needed for efficient utilization of the secondary electrons.

The materials **silver-oxygen-cesium** (Ag-O-Cs) and **cesium-antimony** (Cs_3Sb) used in the photocathodes of phototubes having spectral responses of S-1 and S-4 are also useful as secondary emitters. They are practical from a manufacturing standpoint because the activation process is nearly identical to that used in the production of the corresponding photocathode. However, it is very difficult to produce both the cesium-antimony and the silver-oxygen-cesium emitters in the same envelope. Furthermore, because of the generally high dark emission and instability of the Ag-O-Cs surface, it is no longer used commercially.

The Cs_3Sb emitter has the highest secondary emission of the materials in the practical working range near 100 volts. This material, however, has

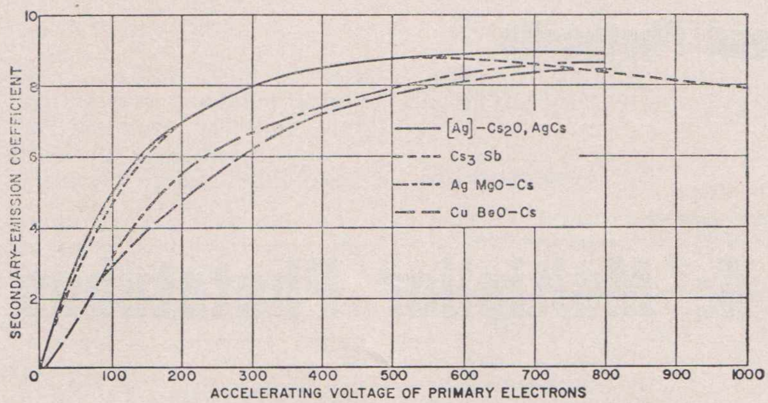

Fig. 30—Secondary emission coefficient for a number of dynode materials.

certain limitations: it cannot tolerate exposure to air, it is damaged by temperatures in excess of 75 degrees centigrade, and it does not have stable characteristics when subjected to current densities in excess of approximately 100 micro-amperes per square centimeter. During manufacture, the cesium-antimony dynode requires a slightly different technique to achieve optimum secondary emission and stability than does the cesium-antimony photocathode. Although this difference is not of major consequence, the result is that on the average the photocathode sensitivity in tubes having cesium-antimony dynodes is slightly less than that achieved with certain other combinations.

A very practical secondary emitter can be made from an oxidized **silver-magnesium** alloy containing approximately 2 per cent of magnesium. Oxidation by means of low-pressure water vapor or carbon dioxide produces a concentration of MgO on the surface which does not occur when the alloy is heated in oxygen directly (probably because the large H_2O or CO_2 molecules do not diffuse as far into the surface, and therefore Mg migrates to the surface before oxidation). When cesium vapor is present during the processing of multiplier photocathodes, it has the further benefit of increasing the secondary-emission ratio.[5-7] Although Ag-Mg-O dynodes do not have as high a secondary-emission ratio as Cs_3Sb dynodes, the material is easily processed and is more stable at relatively high currents. In addition, it can tolerate higher temperatures, and can be outgassed at higher temperatures during exhaust. This surface has a low thermionic background emission, which is important in applications requiring detection of low-level light. Without the cesium activation, the oxygen-activated silver-magnesium layer is used in demountable systems for detecting ions and other particles.

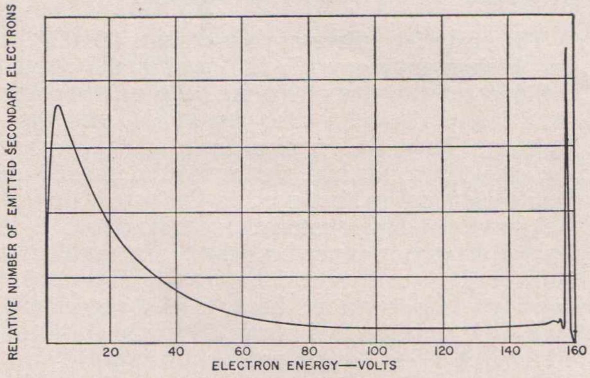

Fig. 31—Typical secondary-electron energy distribution[4] for a silver target: primary electron energy is approximately 150 volts.

A material with characteristics similar to those of Ag-Mg-O can be formed from an oxidised layer of copper-beryllium alloy[8] in which the beryllium component is about 2 per cent of the alloy. Oxidation of the beryllium is accomplished in a manner similar to that used to oxidize the magnesium in the Ag-Mg emitter; secondary emission is enhanced by the bake-out in cesium vapor. Secondary emission and stability are similar to those of the Ag-Mg-O dynode, although copper-beryllium has some advantages in ease of handling and dynode manufacture.

Dynode Configurations

One of the primary problems of design in a multiplier phototube is the shaping and positioning of the dynodes (usually in a recurrent geometrical pattern) so that all stages are properly utilized and no electrons are lost to support structures in the tube or deflected in other ways. Although it is not necessary that the electrons come to a sharp focus on each succeeding stage, the shape of the fields should be such that electrons tend to return to a centre location on the next dynode, even though the emission point is not at the optimum location of the preceding dynode. If this requirement is not met, the electrons increasingly diverge from the centre of the dynode in each successive dynode stage. This effect in turn leads to skipping of stages and loss of gain. Magnetic fields may be combined with

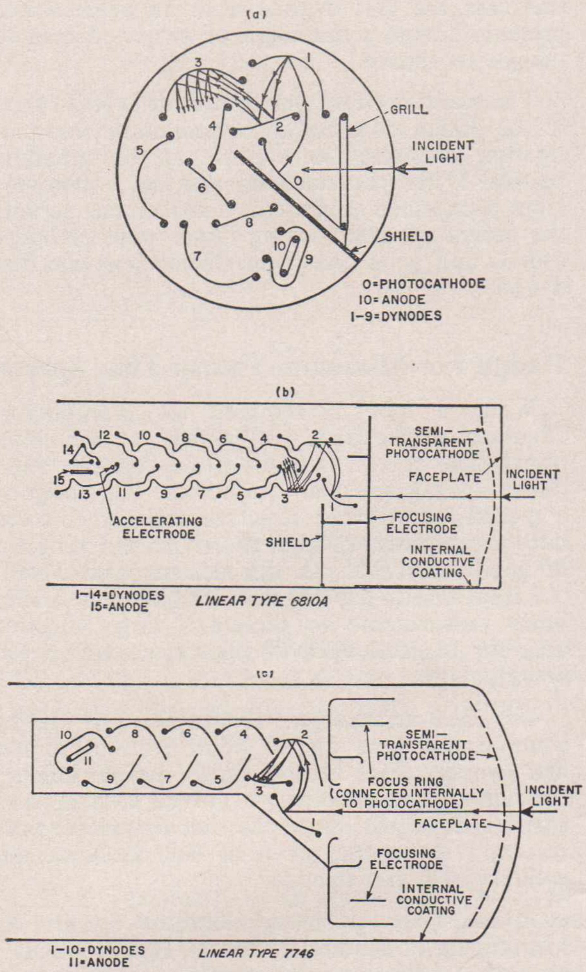

Fig. 32 — Various dynode configurations in general use: (a) circular-cage type; (b) and (c) linear types.

electrostatic fields to provide the required electron optics, although today most multiplier phototubes are electrostatically focused.

A number of different dynode configurations (see Figs. 32 and 33) are used in multiplier phototubes. The circular arrangement of dynodes of the 931A and similar types permits a compact layout, but allows little flexibility in adding dynodes beyond the circle; however, fewer than the full circle of nine dynodes can be used. The collection between stages and the transit-time dispersion is remarkably good for the **circular cage** which was one of the earliest systems developed.[9]

The **Rajchman linear-dynode** structure[10] (as in type 6810A) provides a good recurrent-field system, although the dynode shapes are rather complex. This design is further complicated by a curvature not only as shown in the drawing, but also at right angles to the plane of the drawing. This curvature provides a focusing field which maintains the electron stream near the middle of the dynode structure and prevents bombardment of the supporting spacers at the edges of the dynodes. For this reason, a larger number of dynodes can be used successfully without problems caused by lateral spreading of the electron stream. In general, linear-style dynode systems have good transit-time characteristics because of the focusing properties and the good withdrawal fields at the dynode surface.

Although focused-dynode arrays have a minimum of stray electrons between stages, the acceptance area of the first stage is generally small. If the first stage is used as a cathode, as in the 931A, the cathode area is too small for many applications.

Box-type dynodes provide very efficient collection of electrons between boxes, except for losses to the grid wire. However, because of the lack of specific focusing properties and the wide variation in withdrawal fields, the dynodes do not provide a good transit-time dispersion characteristic.

The **venetian-blind** type of dynode (as in type 8053) can be coupled simply and in a relatively small space. More dynodes can easily be added to the chain if the system is opaque to feedback either by light or by ions. A disadvantage of this type of focusing is the rather low value of electric field at the emitting surfaces, which results in relatively large transit-time dispersion. Some electrons are lost from one dynode to the next because of the low withdrawal field, and some electrons are lost to the wire grid used to prevent field interaction between dynodes.

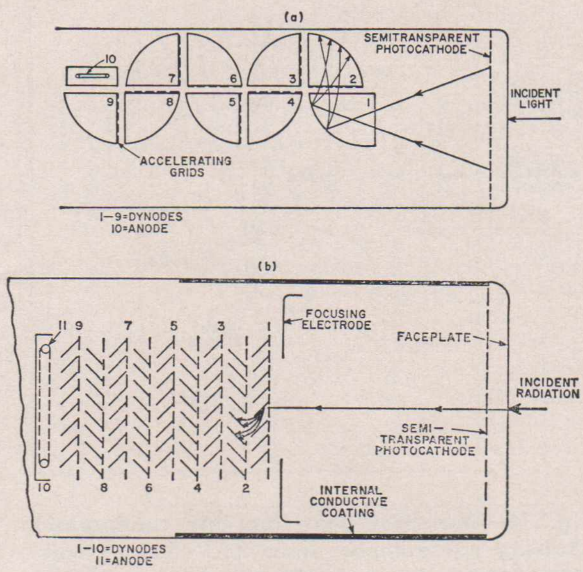

Fig. 33 — Various dynode configurations in general use: (a) box type; (b) venetian-blind type.

At the present time, **the transmission** type of dynode is only used experimentally in multiplier phototubes. The dynode is a thin membrane on which primary electrons impinge from one side and secondary electrons are emitted from the other. Dynodes are mounted in closely spaced parallel planes. For this reason, the transit time and transit-time spread can be made very small. Transmission secondary-emission dynodes are not practical for ordinary applications because of the difficulty and expense of construction and because present dynodes have very poor life at ordinary current levels.

Coupling Dynode System To Cathode

One of the design problems of a dynode system is the coupling of the recurrent dynode chain to the photocathode. In the **circular-cage** type, the first stage of the circle serves as cathode (type 931A); however, for many applications this limited-area cathode is too small. In scintillation counting, it is desirable to have relatively large flat photocathodes at the end of the tube for efficient coupling of the tube to the scintillation crystal. When the first stage of the circular cage is used as a dynode, as it is in the 6342A, the effective size of the first dynode is critically small for collecting all the electrons from the photocathode.

In the **linear-cage type** (type 6810A), the problem is increased by the requirement of housing the whole assembly in an axial configuration. In this case the first dynode is at an angle which presents almost a minimum of projected area to the photocathode.

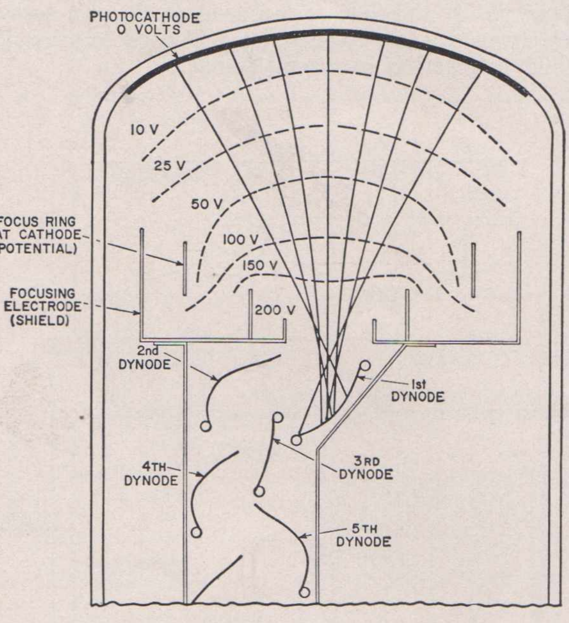

Fig. 34—Matheson-type front-end configuration showing equipotential lines and electron trajectories feeding into a modified linear-type dynode cage which exposes more of the first dynode area to the photoelectron stream.

The venetian-blind type of dynode is well suited to the design of a multiplier phototube for scintillation counting because it can be mounted parallel to the photocathode and has a relatively large acceptance area. This arrangement permits the design of tubes having larger photocathodes (8054) and good collection efficiency at the first dynode.

Design For Minimum Transit-Time Spread

When a tube is required for scintillation-counting applications in which a minimum transit-time spread is desired, the venetian blind dynode is not suitable. Matheson[11] has designed a special focused cage structure designed to solve both the problem of high speed and the problem of good collection. In this design (type 7746), the front end of the cage deviates from a strictly linear construction to present a large effective area for the collection of photoelectrons on the first dynode.

The ideal arrangement should also provide for equal transit time for all photoelectrons to the first dynode. Fig. 34 also shows the Matheson[12] solution to this problem: a curved cathode and annular rings just above the first dynode to correct and shape the potential field between the cathode and first dynode.

A multiplier phototube devised by G. A. Morton, R. M. Matheson, and M. H. Greenblatt[11] provides minimum interdynode transit-time spread; accelerator electrodes placed between dynodes, as shown in Fig. 35, are connected to a highly positive potential. The proximity of the high voltage provides a large withdrawal field for the electrons, and although they are slowed down after passing the accelerator electrodes, the transit time and transit-time spread are very short.

The output sections of some multiplier phototubes have special terminals for very high-speed pulse counting and analysis. The construction may be specially designed to provide a maximum pulse current before space-charge limits the response of the tube.

Properties of Multiplier Phototubes

Gain Characteristics

When several secondary-emission stages are coupled together, so that the secondary electrons from one become the primary electrons of the next, the total gain μ of the multiplier phototubes is given by

$$\mu = \delta^n \qquad \text{.. .. ..} \qquad (26)$$

where δ is the secondary emission per stage

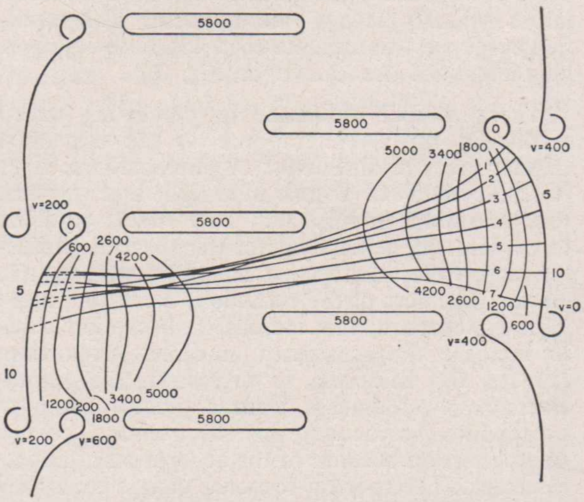

Fig. 35—Interdynode accelerator-electrode system designed by G. A. Morton to provide minimum transit-time spread.

(assumed to be equal for each stage) and **n** is the number of stages. It is also assumed in this expression that all the secondary electrons are collected at the next stage.

In practice, some of the electrons may skip stages, or become lost to the amplification process by impinging upon nonproductive secondary-emission areas.

It is customary to describe the gain of the multiplier phototube as a function of the applied voltage. Fig. 36 shows two such curves on a semilog scale. These curves illustrate the wide range of amplification in a multiplier phototube. They also indicate the necessity of providing a well regulated voltage supply for the dynode stages.

It is possible to operate a multiplier phototube so that each stage is at the voltage required for maximum secondary emission, as shown in Fig. 31. In such cases, the gain could be made practically independent of voltage over a small range. However, such a condition would require approximately 500 volts per stage; thus the total voltage required would be very high for the amount of gain achieved.

In the design or operation of a multiplier phototube having a fixed supply voltage, the number of stages can be chosen so that the gain of the tube is maximum. For this purpose, the optimum voltage per stage is that value at which a line through the origin (unity gain on the log-gain scale) is tangent to the curve, as shown in Fig. 36. This point is identified on the graph as the point of **maximum gain per volt.** (Note that this argument neglects the voltage used between the last dynode and the anode and any discrepancy resulting from nonuniform distribution of voltage per stage.) In most applications of multiplier phototubes, the tubes are operated above the point of maximum gain per volt. It is customary to present tube data with both the gain and the voltage on a logarithmic scale; over the normal range of operation the resultant curve is then closely approximated by a straight line.

Spectral Response

Photocathodes developed for diode phototubes are also used in multiplier phototubes. Photoelectrons emitted from the photocathode are directed to the first dynode of the tube instead of to the anode as in a photodiode. However, several special photocathodes which are rarely used in photodiodes have been used in multiplier phototubes.

Cathodes of the **transmission** type are often used in multiplier phototubes, in contrast with the **opaque** type used in most photodiodes. A transmission-type photocathode is one in which a semi-transparent layer is applied to the inside surface of the envelope window. Light impinges on the outer (glass) side of the photocathode, and electrons are emitted on the inner or vacuum side. The vacuum-evaporation and processing of a transmission-type cathode require careful control to achieve uniformity and high sensitivity. The most common transmission-type photocathode is made of antimony and cesium, the same elements used for the opaque photocathode in tubes having an S-4 spectral response. During the processing of the tube, antimony is vacuum-evaporated onto the inner surface of the window. A metallic substrate is often put on the

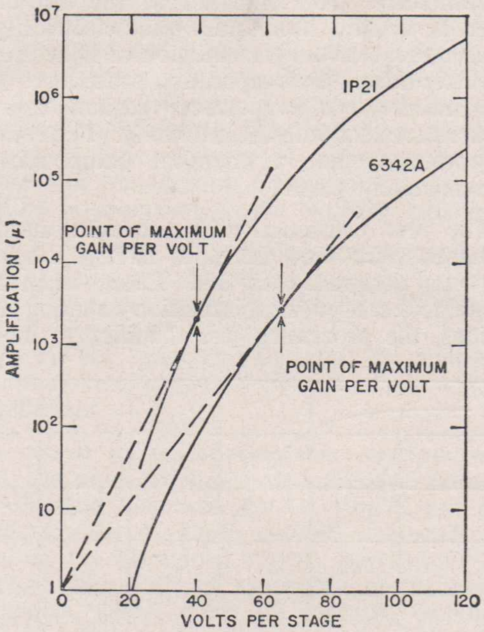

Fig. 36—Log of gain as a function of volts per stage for a tube (1P21) with Cs-Sb dynodes and for a tube (6342A) with Cu-Be dynodes.

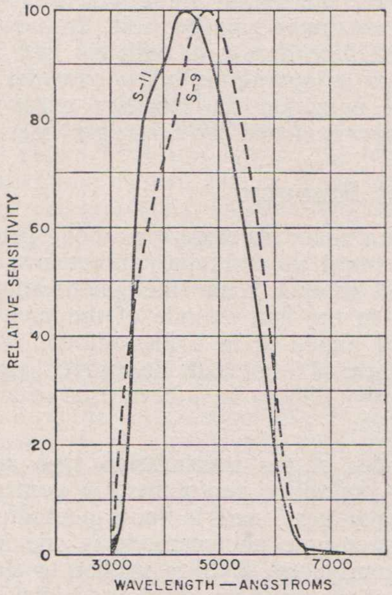

Fig. 37—Comparison of the S-9 and S-11 spectral response characteristics. Both curves are for transmission-type cesium-antimony photocathodes. The S-11 response was evolved to provide maximum blue response for scintillation counting.

window first to improve the conductivity of the photocathode or to facilitate the activation process. The antimony is usually evaporated from a heated filament pre-beaded with antimony; the beads are positioned to provide a uniform layer of antimony. The thickness of the evaporated layer is usually monitored photoelectrically by measurement of the transmission of light through the layer during the evaporation procedure. After the antimony is evaporated, cesium vapor is allowed to react with the antimony. The resultant photocathode has a chemical composition of approximately Cs_3Sb.

The first transmission-type cesium antimony cathodes were used in tubes having a spectral response designated as S-9. Later, when tubes began to be used in scintillation-counting applications, the processing was modified to increase

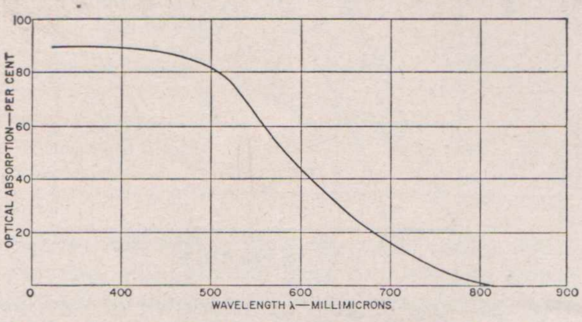

Fig. 38 — Optical absorption of Cs_3Sb layer having a typical S-11 spectral response.

the blue sensitivity of the cathode because scintillators typically have a blue emission. This modified response was designated S-11; both response characteristics are shown in Fig. 37.

The principal difference in processing which results in the S-11 response is the use of a thinner layer of antimony. A photocathode layer of Cs_3Sb tends to absorb blue light and transmit red, as shown in Fig. 38. Two effects combine to explain the dependence of the spectral characteristic on the thickness of the photocathode. (1) As the thickness of the cathode is increased, more light is absorbed; this increased absorption tends to increase the sensitivity in direct proportion. (2) As the thickness is increased, the photoelectrons must emerge from a greater depth to escape into the vacuum; this effect tends to reduce photoemission because of the absorption of photoelectrons. The spectral-response characteristic S-9 was evolved for use with a typical tungsten light source. The resulting rather thick surface provided better response at the red end of the spectrum. A compromise to a thinner cathode improved the blue response (S-11) with the loss of some red response.

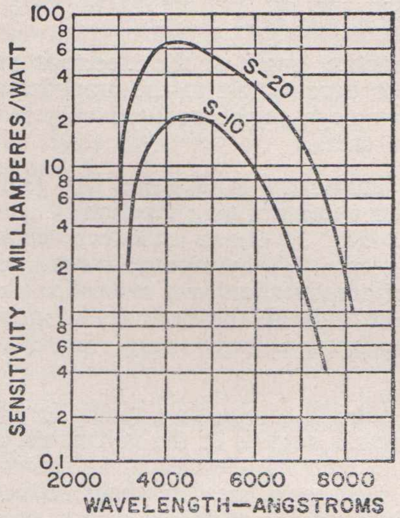

Fig. 39—Spectral response characteristics for S-10 (Bi-Ag-O-Cs photocathode) and S-20 [(Cs) Na_2 KSb photocathode].

Two transmission-type photocathodes of importance, particularly in the red region of the spectrum, are the **bismuth-silver-oxygen-cesium** (Bi-Ag-O-Cs) cathode used in tubes having S-10 spectral response, and the **multialkali** ([Cs]Na_2KSb) cathode, used in tubes having S-20 spectral response. These spectral-response characteristics are shown in Fig. 39.

The semitransparent Bi-Ag-O-Cs cathode is prepared by first evaporating a thin layer of bismuth and then a thin layer of silver. The silver is then oxidized, and cesium vapor is allowed to react with the layer.

The multialkali photocathode[13] is very difficult to process to uniformly high sensitivity. The process is complicated and involves alternate treatment with evaporated antimony and alkali vapors. The resultant photocathode is the most sensitive known for the region from the ultra-violet to the red end of the spectrum. However, compared with the cesium-antimony photocathode, the multi-alkali photocathode has only a slight advantage in the blue region.

The **bialkali photocathode** (Na_2KSb) also deserves mention although at the present time its use is only experimental. The Na_2KSb cathode has a spectral response (tentatively identified as S-24) similar to the S-11 (see Fig. 40), but has the advantage of a lower thermionic emission at room temperature of the order of 10^{-16} amperes per square centimeter. In fact, some data have indicated values as low as 10^{-19} amperes per square centimeter. Another advantage of the Na_2KSb cathode is its ability to withstand somewhat higher temperatures than other cathodes; it can be used to about 100 degrees centigrade. The bialkali cathode promises to be useful as a cathode for low-energy scintillation counting because of its good blue sensitivity and very low dark current.

Another two-alkali antimony cathode has recently been announced by A. H. Sommer, $(K,Cs)_3Sb$. Although its properties have not been thoroughly explored, it is apparent that it has

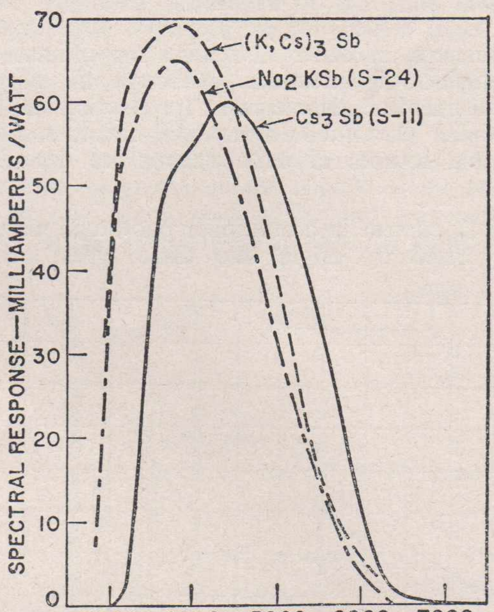

Fig. 40—Comparison of the absolute spectral sensitivities of three antimony alkali photocathodes: Cs_3Sb, Na_2KSb, and $(K,Cs)_3Sb$. The S-11 curve shown is representative of the cathode on type 8053: the S-24 and $(K, Cs)_3Sb$ curves are tentative.

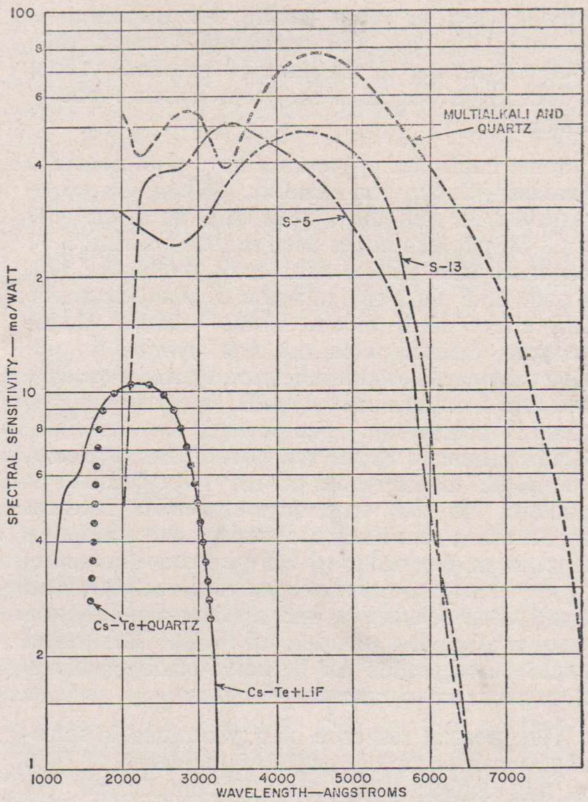

Fig. 41 — Spectral response characteristics of various photocathode-window combinations useful for the ultraviolet: S-5 (Cs-Sb with 9741 glass); S-13 (transmission type Cs-Sb with quartz); Cs-Te with quartz; Cs-Te with LiF window; multialkali with quartz.

high blue sensitivity and low dark current. Its tentative spectral response is also given in Fig. 40. At the present time, it seems that the most promising application is in scintillation counting, perhaps for low energy work, although the dark current and temperature characteristics may not be quite as promising as the Na_2KSb cathode.

As in vacuum photodiodes, the spectral response of multiplier phototubes is generally limited in the ultraviolet region by the type of window used in the device, as shown in Fig. 19. Fig. 41 compares the spectral responses of multiplier phototubes having several photo-cathode-window combinations.

In recent years, considerable developmental effort has been expended on cathodes made of the following materials: Cs_2Te, Rb_2Te, and CsI and CuI. All of these cathode materials are useful in the ultraviolet region, particularly because of their lack of sensitivity in the longer wavelength region. They are generally combined with a window made of a material such as LiF or sapphire.

The only photocathode useful in the infrared region is the **silver-oxygen-cesium** (Ag-O-Cs) cathode used in tubes having S-1 response, as shown in Fig. 18. For wavelengths longer than 8000 angstroms to its limit of response, about 11000 angstroms, it is the most sensitive photocathode.

Some multiplier phototubes have photosensitive dynodes—Cs-Sb, for example. When the transmission-type cathode is quite thin, as in the case of Cs-Sb photocathodes used in tubes having S-11 response, transmitted light may strike the first dynode and cause the emission of photoelectrons. This effect is of second order because photoelectrons emitted from the first dynode do not have the benefit of multiplication by the secondary emission from the first dynode; nevertheless, the effect is observable. The increase in sensitivity occurs primarily at the red end of the spectrum, where the transmission of the photocathode is greatest. In the blue region, which is most important in scintillation counting, the maximum effect is of the order of 2.5 per cent (assuming 10 per cent transmission, identical cathode and first-dynode sensitivity, and a secondary-emission ratio at the first dynode of 4:1). The effect would be negligible for dynodes made of silver-magnesium or copper-beryllium.

The spectral response of a phototube is somewhat sensitive to the **angle of incidence** of light on the cathode. The effect is complicated, but to a first approximation it is possible to explain the increased red sensitivity with angle of incidence in tubes having S-11 response (shown in Fig. 42) by the increased absorption of red light with greater angle of incidence.

The effect of temperature on spectral response is minor and occurs primarily in the red region of the spectrum,[14] as shown in Fig. 43. At higher temperatures, the distribution of electrons in the semi-conductor photocathode is shifted to higher energy levels and, consequently, more electrons are emitted by the near-threshold energy radiation.

Dark Current

Even though a multiplier phototube is in complete darkness, electron currents may be observed to flow from dynode to dynode. These currents increase in magnitude toward the anode end of the tube since any electron component of the dark current is amplified in the same way as the photo-originated electrons. Usually, only the dark current in the anode circuit of the tube is of importance to the observer. This **anode dark current** is described in this section as the "dark current" of the multiplier phototube.

The dark current and the resulting **noise** of the multiplier phototube is of particular concern because it is usually the critical factor in limiting the lower level of light detection. It is important to understand the variation of dark current in the multiplier phototube as a function of various parameters, in order to realize the ultimate in low-light-level detection. Although all of the pecularities of dark-current production in the multiplier phototube are not well understood, the possible sources of dark current are described below:

Dark current in a multiplier phototube may be categorized by origin into three types: ohmic

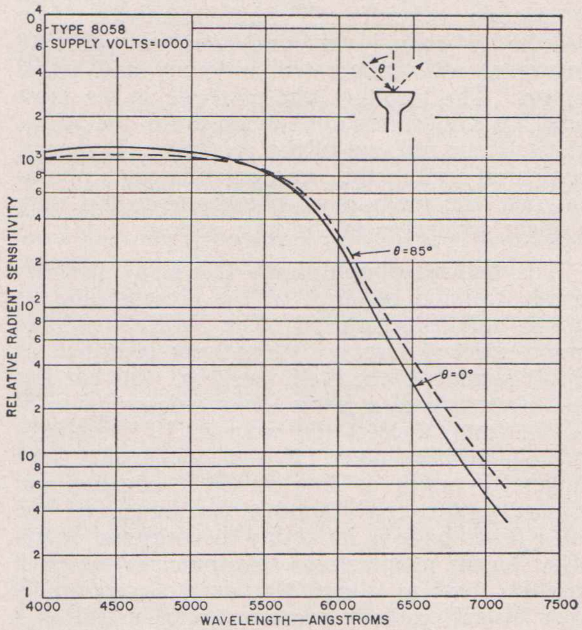

Fig. 42—Spectral-response characteristics for a Cs-Sb transmission-type photocathode (S-11) for two different incident angles of radiation. The increased red response for the large angle of incidence may be explained in part by the increase in absorption due to the longer path; compare the absorption characteristic shown in Fig. 38.

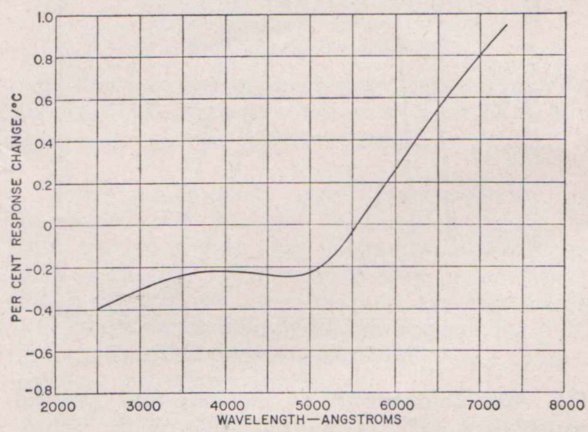

Fig. 43 — Temperature coefficient of Cs-Sb cathodes as a function of wavelength; data taken at 20 degrees Centigrade. Note the large positive effect near the threshold.

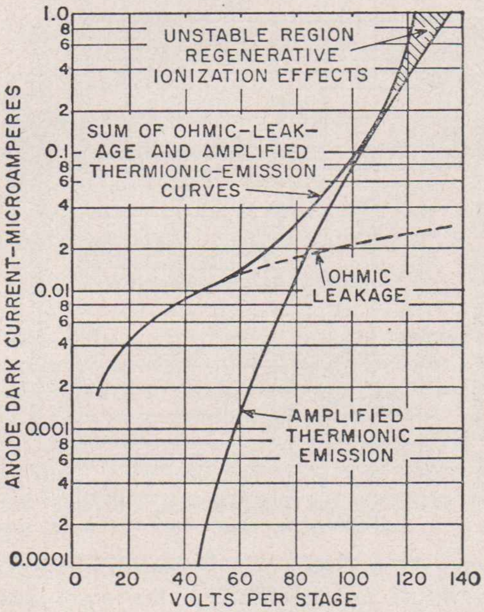

Fig. 44—Typical variation of dark current with voltage for a multiplier phototube.

leakage, dark or "thermionic" emission of electrons from the cathode and other elements of the tube, and regenerative effects.

Ohmic leakage, which results from the imperfect insulating properties of the glass stem, the supporting members, or the plastic base, is always present. This type of leakage is usually negligible, but in some tubes it may become excessive because of the presence of residual metals used in the processing of the photocathode or the dynodes. Condensation of water vapor, dirt, or grease on the outside of the tube may increase ohmic leakage beyond reasonable limits. Simple precautions are usually sufficient to eliminate this sort of leakage. In unfavorable environmental conditions, however, it may be necessary to coat the base of the tube with moisture-resisting materials, which may also prevent external arc-overs resulting from high voltage.

Ohmic leakage is the predominant source of dark current at low-voltage operating condition. It can be identified by its proportionality with applied voltage. At higher voltages, ohmic leakage is obscured by other sources of dark current.

Fig. 44 shows the typical variation of dark current of a multiplier phototube as a function of applied voltage. Note that in the mid-range of voltage, the dark current follows the gain characteristic of the tube. The source of the gain-proportional dark current is the dark or **thermionic emission** of electrons from the photocathode and the first-dynode stage. Because each electron emitted from the photocathode is multiplied by the secondary-emission gain of the tube, the result is a unipotential output pulse having a magnitude equal to the charge of one electron multiplied by the gain of the tube. (There are statistical amplitude variations which will be discussed later.) Because the emission of thermionic electrons is random in time, the output dark current consists of random unidirectional pulses. The time average of these pulses, which may be measured on a dc meter, is usually the principal dc component of the dark current at normal operating voltages. The limitation to the measurement of very low light levels is the variable character of the thermionic dark-current component. It is not possible to balance out this wide-band noise component of the multiplier phototube, as it might be to balance out a steady ohmic-leakage current. Nevertheless, it is usually advantageous to operate the multiplier phototube in the range where the thermionic component is dominant. In this range, the relationship between

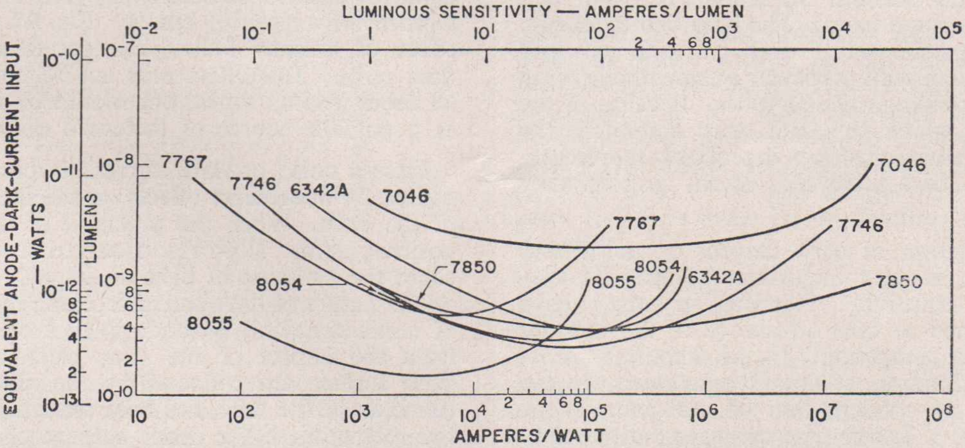

Fig. 45—"Equivalent anode dark current input" as a function of the luminous sensitivity for various multiplier phototubes. The Equivalent Anode Dark Current Input represents the light flux which would result in an output current change just equal to the dark current. Optimum operating range is usually where this function is near a minimum.

sensitivity and noise is fairly constant as the voltage is increased because both the photoelectric emission and the thermionic emission are amplified by the same amount.

At higher dynode voltages, a regenerative type of dark current develops, as shown in Fig. 44. The dark current becomes very erratic, and may at times increase to the practical limitations of the circuit. Continued flow of large dark currents may cause damage to the sensitized surfaces. Some possible causes of the regenerative behaviour will be discussed in more detail later. All multiplier phototubes eventually become unstable as the gain is increased.

The best operating range can generally be predicted from a consideration of the ratio of the dark current to the output sensitivity of the tube. This ratio, known as the **equivalent anode-dark-current input** (EADCI), is shown as a function of luminous sensitivity in Fig. 45. The EADCI is equivalent to the light flux on the photocathode which would result in an output-current change equal to the dark current observed. If thermionic emission were the sole source of dark current, the EADCI would be a horizontal line on the graph. In some tubes a regenerative condition sets in before the really flat operating region has been attained.

As may be expected, the thermionic component of the dark current is very much a function of the temperature. Fig. 46 shows the temperature variation of the equivalent of the dark current at the photocathode (anode dark current divided by gain) per unit cathode area. The data for this figure were obtained from a number of tubes having well defined (flat) minima of the EADCI curves. In some cases, the photocathode equivalent of the output dark current was derived from noise measurements because of the difficulty of separating the leakage from the electronic component of dark current at low temperatures. (The relation between thermionic emission and noise is discussed later.) The data shown in Fig. 46 represent the actual dark emission per unit area associated with particular photocathode types rather than with particular tubes. Because of the manner in which the data were compiled, the thermionic emission shown is probably representative of the best achievable at the present state of the art.

The variation of dark current or noise with temperature is most important for ultimate low-light-level sensitivity. Various cryostats have been designed to take advantage of the reduced noise at low temperature.[15] An important practical consideration at low temperatures is the prevention of condensation of moisture on the window. In a Dewar-type arrangement, condensation is not a problem; in simpler set-ups moisture condensation may be prevented by a controlled low-humidity atmosphere at the external window.

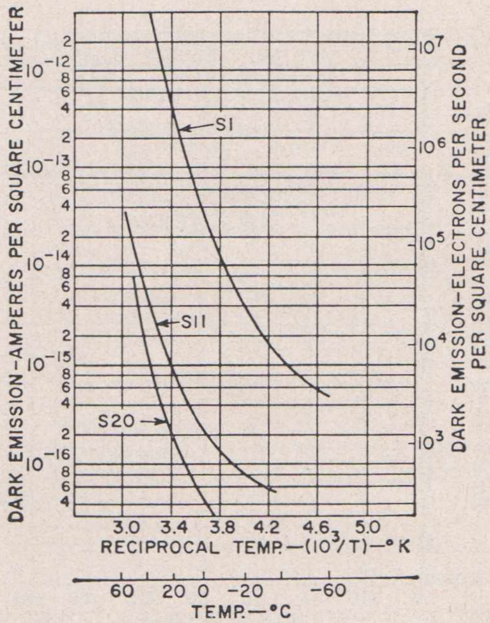

Fig. 46—Temperature variation of dark current for multiplier phototubes in the manner of a Richardson plot.

In most multiplier phototubes, the electrostatic potential of the walls surrounding the photocathode and dynode-cage region is important particularly with respect to the onset of the regenerative dark-current component. The bulb wall can be maintained near photocathode potential if the bulb is wrapped or painted with a metallic coating maintained at cathode potential; the connection of the metallic coating to the cathode is usually made through a high impedance to avoid shock hazard. A positive potential on the bulb wall can cause noisy operation. Even though the bulb is not connected to a positive potential, the proximity of a shield or container at positive potential may lead to the development of a positive charge. Fig. 47 shows the effect of various bulb-shield potentials on the dark noise. This effect may not be observed for all tubes and all types, but should be recognized as a possible source of increased noise.

Excess noise or dark current is often accompanied by **fluorescent effects** on the inner surface of the bulb. When the potential of the bulb is positive, stray electrons attracted to the bulb cause the emission of light on impact, depending on the nature of the glass surface and the presence of contamination. Secondary electrons resulting from the impact of the stray electrons on the glass surface are collected by the most positive elements in the tube and help maintain the positive potential of the inner surface of the glass. Under these circumstances, it is possible to observe the formation of glowing spots on the inside of the glass bulb, provided the eye is dark-adapted and the applied voltage is sufficiently

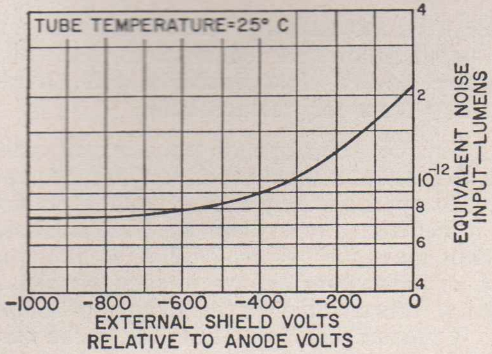

Fig. 47—Effect of external-shield potential on the noise of a 1P21 multiplier phototube. Note the desirability of maintaining a negative bulb potential.

high. Some of this emitted light may reflect back to the cathode and cause regenerative dark current.

Other surfaces within the multiplier phototube are also important in the control of regenerative dark current. In the 6342A, for example, the focus-shield potential shows a point of distinct noise minimum,[16] as shown in Fig. 48. The mechanism of this behaviour is not understood at present.

Another phenomenon which deserves mention in connection with dark current is the effect of previous exposure to light. If a photocathode is exposed to strong light, with or without an applied voltage, a measurement made immediately afterward shows a higher-than-normal dark current which decreases rapidly with time. The effect

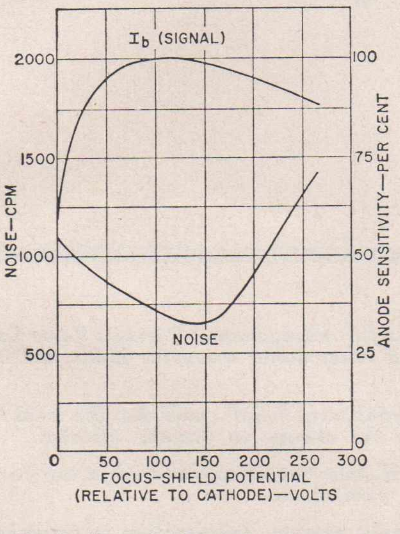

Fig. 48—Typical variation of noise output of a 6342A multiplier phototube as a function of focus-shield potential. Also indicated is the output signal current showing how only a minimum loss in signal sensitivity results from a choice of shield potential which minimizes noise.

is more marked when the exposure is richer in shortwavelength radiation. Very long periods are required to attain the low dark current associated with equilibrium, as shown in Fig. 49.

An analysis of the rate of decay of the dark emission suggests the following reciprocal relationship:

$$i_d = \frac{a}{1 + bt} + i_o \quad \ldots \quad (27)$$

where i_d is dark emission as a function of time, a and b are constants, t is time, and i_o is the constant level of dark emission. However, the significance of this relationship is not apparent at present. It is possible that the heavy initial exposure to light alters the Fermi level of the photocathode as a result of the introduction of excited states and that the photocathode then decays to its initial state. For best operation and low noise, therefore, it is recommended that the photocathode be kept dark at all times, or at least for many hours before making low-level measurements.

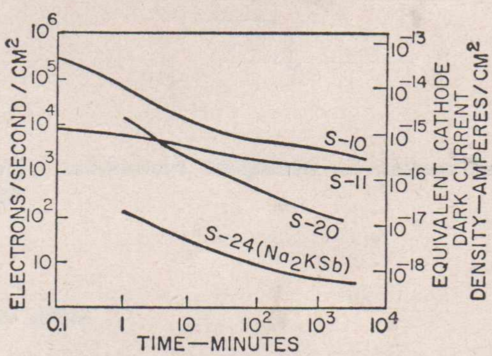

Fig. 49—Variation of dark current following exposure of cathode to cool-white fluorescent-lamp radiation. The various cathodes are identified by their spectral-response symbols.

Origins of Dark Current. In a metal, the electrons which escape as thermionic emission are generally from the top of the conduction band (see Fig. 1). Thus, the **work function** for photoemission and thermionic emission is the same. Thermionic emission as a function of work function ø in volts and temperature T in degrees Kelvin is given by the familiar Richardson equation:

$$j = \frac{4\pi emk^2T^2}{h^3} e^{-\phi e/kT} \quad \ldots \quad (28)$$

where j is the thermionic current density; e, the electron charge; m, the electron mass; k, Boltzman's constant; and h, Planck's constant. In MKS units, the equation becomes

$$j = 1{\cdot}2 \times 10^6 T^2 e^{-\phi e/kT} \quad \ldots \quad (29)$$

For semiconductor photocathodes, the work functions of photoemission and thermionic emission may be quite different. The work function for photoemission (see Fig. 5) is the potential height from the top of the valence band to the vacuum level, or E_a (the electron affinity) plus E_g (the forbidden gap, i.e., the separation of valence and conduction bands). For an intrinsic semiconductor, thermionic emission originates from the valence band, as does photoemission but the "work function" is not the same as for photoemission. In the case of an intrinsic semiconductor, thermionic-emission density can be expressed as

$$j + \frac{4\pi\ emk^2T^2}{h^3}\ e^{-\frac{\left(E_a + \frac{E_g}{2}\right)e}{kT}} \quad .. \quad (30)$$

For an excess impurity semiconductor where thermionic emission originates from the impurity centres, the equation for thermionic emission is written as follows:

$$j = \frac{4\pi emk^2T^2}{h^3}\ n_o^{1/2}\ \frac{h^{3/2}}{(2\pi mkT)^{3/4}}\ e^{-\frac{(E_a + R/2)\,e}{kT}} \quad .. \quad .. \quad (31)$$

where n_o is the impurity concentration and R is the depth below the conduction band to the impurity level.[17] Photoemission still originates primarily from the valence band.

In Fig. 46, the slope of the curve of cathode dark current as a function of reciprocal temperature has approximately the magnitude predicted, but it is difficult to explain the change in slope. Perhaps there are "islands" of different impurity level or concentration, or it may be that the impurity concentration is a function of temperature. Exposure to temperatures above the normal operating range sometimes results in permanent reduction in dark current.

Another source of dark emission from the photocathode results from radioactive elements in the tube or surroundings which cause scintillations in the glass envelope of the tube. For example, it is very difficult to obtain glass that does not contain potassium, which has a natural component of radioactive K^{40}. Some glasses, such as fused silica, have comparatively low background radiation.

This section on Multiplier Phototubes is too long for inclusion in one issue. The second half will appear next month.

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

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