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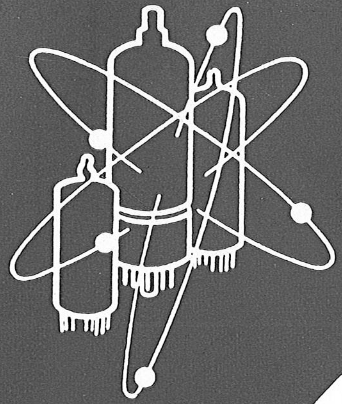

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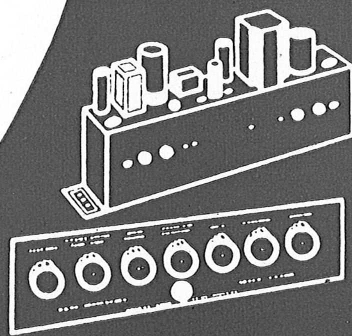

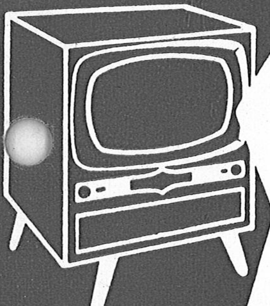

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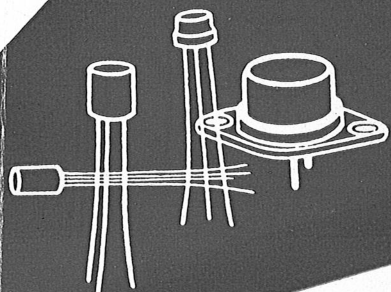

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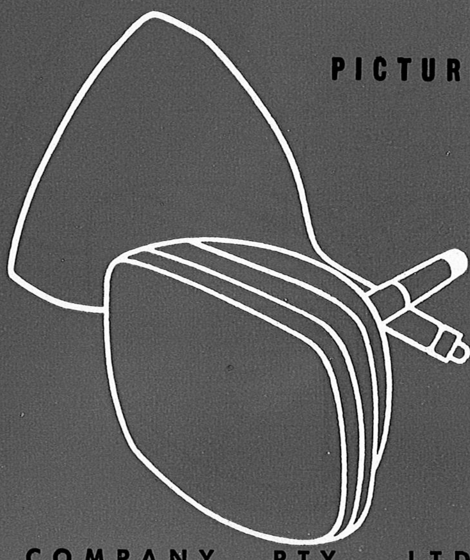

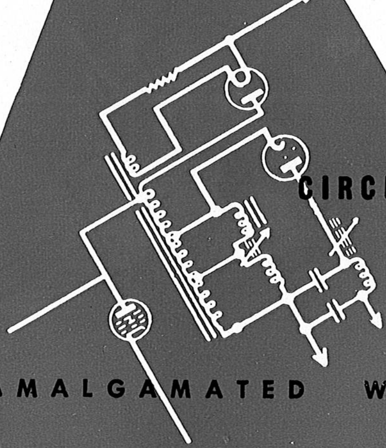

CIRCUITRY

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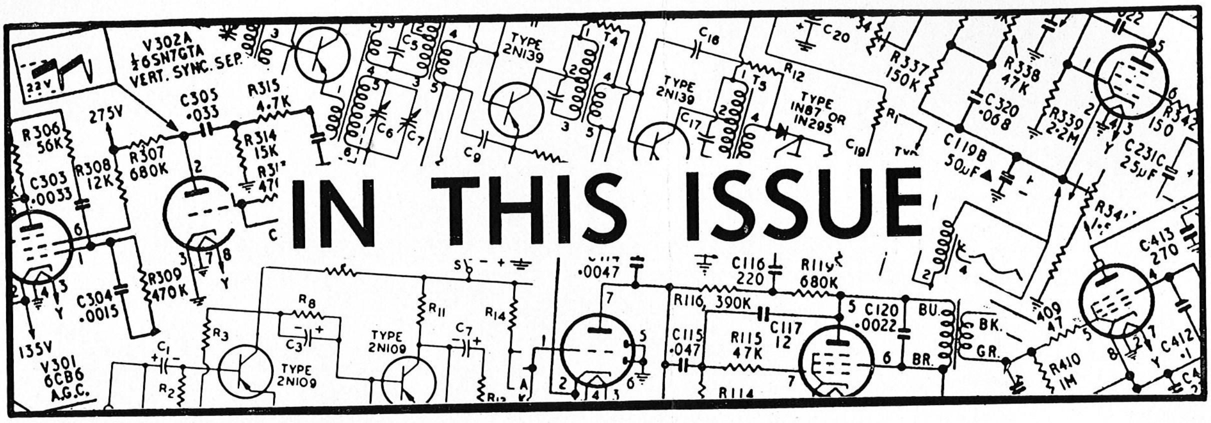

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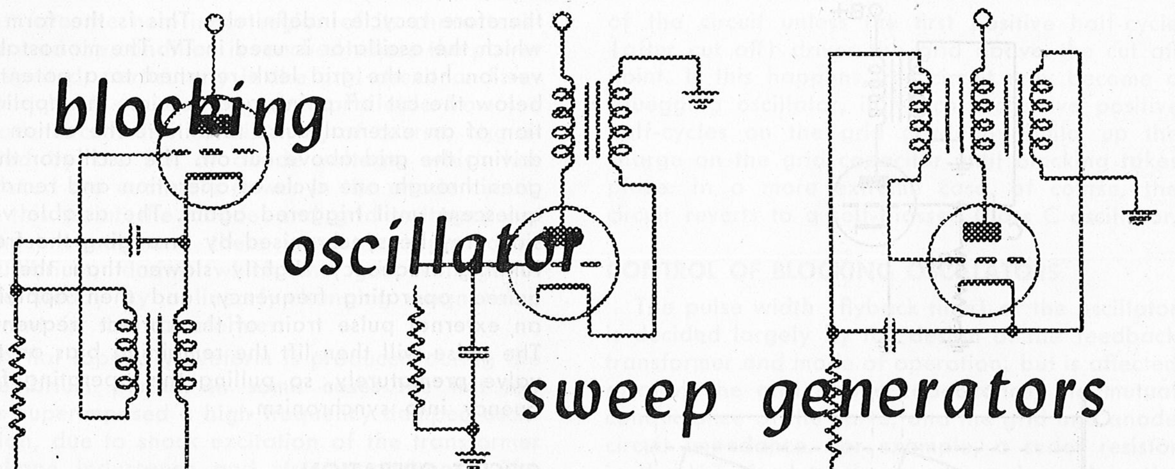

blocking oscillator sweep generators

By B. J. SIMPSON, M.I.R.E. (Aust.)

INTRODUCTION

The blocking oscillator is essentially a transformer — coupled feedback oscillator which is so arranged that anode current flows for one half-cycle, after which a bias voltage developed in the grid circuit prevents further oscillation. The circuit is a development of the quenching or squegging oscillator, and may be considered as a limiting case of that circuit.

Grid-leak biassed oscillators may be caused to squegger by arranging the circuit values so that after a short burst of oscillation, grid current charges the grid capacitor to the point where the valve is cut off. Oscillation then ceases until the bias relaxes sufficiently for anode current to flow again. The output of the oscillator therefore consists of short bursts of oscillation separated by quiescent periods. The frequency of oscillation is determined by the inductance/capacitance in the circuit, the operating and quiescent periods mainly by the grid leak/grid capacitor values. Characteristics of the squegging oscillator are heavy positive feedback provided by high mutual inductance in the feedback transformer, and vigorous oscillation during the short operating period. Where the oscillator is used as a time base generator, the sawtooth sweep voltage is derived across the grid capacitor. The trace is provided by the exponential discharge of the capacitor, and flyback by the rapid charging during the oscillating period. The oscillator may also be used as the discharge switch in a capacitive time base; this aspect is discussed later.

Turning now to consideration of the blocking oscillator, this circuit is distinguished from the generalised squegging oscillator by the fact that the blocking or squegging action takes place in one half-cycle of operation. The half-cycle does not necessarily have any direct relationship with the natural resonant frequency of any part of the circuit. The circuit magnification factor of the tuned circuit is reduced, whilst at the same time a high L/C ratio is used with heavy feedback to retain strong oscillation during the operating period. The capacitance of the tuned circuit is usually provided by the distributed capacitance of the transformer plus strays.

In a feedback oscillator producing a sinusoidal output, part of the circuit is resonant at the required operating frequency. In the blocking oscillator however, the pulse repetition frequency is the quality in which we are mainly interested. This is determined, not by the resonance of the circuit, but mainly by the time constants of the self-biassing arrangements. The resonant frequency has an effect on the pulse rise and fall times, and its period is always very short compared with the required pulse duration. A low-impedance feedback path and short time lag in the transformer permit the generation of pulses with very rapid rise and fall times, whilst pulses of very short duration are possible, down to the order of 0.1 microseconds. The blocking oscillator is further characterised by low power consumption and rapid recovery time compared with other sweep and pulse generators.

APPLICATION

Blocking oscillators find many applications as sweep generators and pulse generators for oscillography, TV, radar ranging circuits, counting circuits, frequency division and the like. At the present time however, we are considering the use of the device as a sweep generator, mainly as applied to TV.

The waveforms available from the oscillator fall into three classifications. A voltage pulse may be taken from the anode or from a tertiary winding on the feedback transformer. A typical output impedance at the anode is 1,000 ohms, whilst a tertiary winding may be designed for

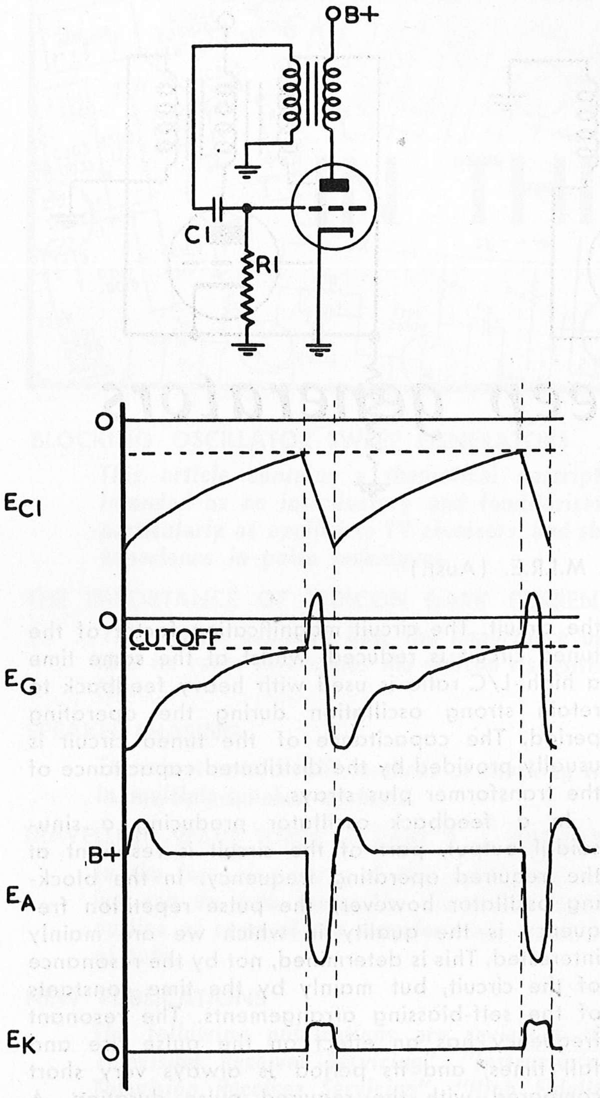

FIG. 1

the desired load impedance and output level. A current pulse may be derived across a resistor placed in the anode, cathode or grid circuits; the output impedance in these cases is very low. The third method of using the oscillator is to take the sawtooth voltage waveform of the self-bias voltage across the grid capacitor. The output impedance in this case is low during the charging time (flyback) due to the large grid/cathode current, and is determined by the circuit values during the discharge time. This third application is now rare, and when used as a sweep generator, the oscillator is today invariably used as a discharge switch in a capacitive time base, except where a specialised circuit is employed.

The blocking oscillator is used in two basic forms, the astable or free-running, and the monostable or triggered version. In the astable version, the grid leak is returned to a potential above the valve cut-off voltage, and the circuit will therefore recycle indefinitely. This is the form in which the oscillator is used in TV. The monostable version has the grid leak returned to a potential below the cut off point, and requires the application of an external pulse to initiate the action by driving the grid above cut off. The oscillator then goes through one cycle of operation and remains quiescent until triggered again. The astable version may be synchronised by arranging the free-running frequency slightly slower than the required operating frequency, and then applying an external pulse train of the correct frequency. The pulse will then lift the remaining bias on the valve prematurely, so pulling the operating frequency into synchronism.

CIRCUIT OPERATION

The blocking oscillator operates in four recognised modes, the distinguishing features of which are the shape and duration of the grid and anode pulses. The modes are exponential, exponential/sine, sine, and leakage-inductance controlled. The first three named modes produce pulses of decreasing duration, and other values being equal, are determined by decreasing values of anode circuit impedance; they are so called from the shape of the grid waveform produced. The leakage—inductance controlled mode is so called because it produces a pulse of very fast rise and fall time, in which the rate of change of anode voltage and grid current is limited only by the leakage inductance of the feedback transformer. Viewing the circuit as a whole, the mode of operation is determined by circuit parameters, and particularly the design of the feedback transformer. The mechanism of operation is basically the same for each mode, and the following description may be applied to all, although it is related specifically to the exponential mode.

It should be noted that the R/C grid circuit may be replaced with an L/C circuit or even with a delay line for particular applications, but the function of the circuit is always the same, to hold the valve cut off pending discharge or retriggering.

The basic blocking oscillator circuit is shown in Fig. 1. For the purpose of explanation let it be assumed that the decay of a negative charge on the grid capacitor has just allowed the grid to rise above cut off. The valve commences to conduct. The change in anode current induces a positive voltage at the grid, which increases anode current still further. When the gain around the feedback loop becomes greater than unity, oscillation occurs. The anode current and grid voltage then rise increasingly quickly, attaining a rate limited by the leakage inductance of the transformer. As the grid of the valve is driven positive, grid current commences, and the reduced grid impedance imposes a heavy resistive loading on the transformer. The valve is driven well into the non-linear region of its characteristics, and the condition is reached where the power dissipated, particularly in the grid circuit and external load, equals the power available from the anode circuit. The condition is, of course, aggravated by the fall of anode voltage, which has reduced the available power in the anode circuit. At this point the anode and grid voltages are almost equal, and there is a short period of equilibrium, during which the grid voltage is held steady by a linearly-changing magnetising current in the transformer.

A flat topped waveform is produced during the equilibrium period. In some cases the flat top has superimposed a high frequency damped oscillation, due to shock excitation of the transformer leakage inductance and stray capacitance. The short equilibrium period is terminated by the collapse of the magnetic field in the transformer and the discharge of the distributed transformer capacitance. The induced grid voltage undergoes a polarity reversal. The decreases of anode current is delayed by the heavy loading of the transformer secondary by grid current. Furthermore, as the anode/cathode potential increases, the increasing gain in the stage also contributes a delay. The comparatively slow decrease of anode current persists until the grid voltage falls below zero and grid current ceases. With the heavy loading now removed from the transformer secondary, the collapse of the magnetic field is considerably accelerated. This is followed by a heavy oscillatory overswing which drives the grid negative to an amount determined by circuit losses, and which adds to the negative charge on the grid capacitor.

The negative voltage across the grid capacitor rose sharply when heavy grid current flowed during the rise time of the pulse. During the fall time the effective grid voltage is that supplied by the transformer, less the charge on the capacitor. The failure of the transformer to hold the grid positive may occur at any time during the fall time of the pulse, and so initiate the regenerative shut off. In some cases the voltage across the grid capacitor built up during the initial rise in anode current is of such magnitude as to cut the valve off in the absence of the large positive pulse from the transformer. In this case the valve is cut off as the rate of change of anode current falls to zero and the positive pulse is removed from the grid. The regenerative shut off does not occur; the oscillation is heavily damped and rapidly reduced to a negligible amplitude.

The waveform produced by the cut off of the oscillator depends on the degree of damping present. Where the damping is heavy, as is usually the case, the overswing will be limited to approximately one half cycle, as shown in the waveforms of Fig. 1. Where the degree of damping is less, a train of damped oscillations will be produced. This will not seriously affect the operation of the circuit unless the first positive half-cycle (after cut off) drives the grid above the cut off point. If this happens, the circuit may become a squegging oscillator, in which successive positive half-cycles on the grid gradually build up the charge on the grid capacitor until blocking takes place. In a more extreme case, of course, the circuit reverts to a self-biassed Class C oscillator.

CONTROL OF BLOCKING OSCILLATORS

The pulse width (flyback time) of the oscillator is decided largely by the design of the feedback transformer and mode of operation, but is affected also by the grid/anode capacitance, the mutual conductance of the valve, and the grid and anode circuit impedance. For example, a series resistor in the H.T. feed to the anode as is used in the simple capacitive time base of Fig. 6, will in general produce a longer pulse width with increasing values of resistor. Increasing the grid/anode capacitance will produce a similar effect. In some sweep generators the flyback time is not critical, and no problem exists, but in TV applications flyback time can become important.

The pulse repetition frequency of the oscillator is determined by the expression

$$T + RC\log_e (V_1 + V) (V_1 + V_2),$$

where T is the pulse width, V the peak negative voltage attained by the grid capacitor, V_1 the potential to which the grid leak resistor is returned, and V_2 the cut off voltage of the valve. R and C are, of course, the grid circuit values. Control of repetition frequency can therefore be obtained by a variation of one or more of R, C, or any of the three voltages. In practice, the variation of C, V, or V_2 does not provide an easy

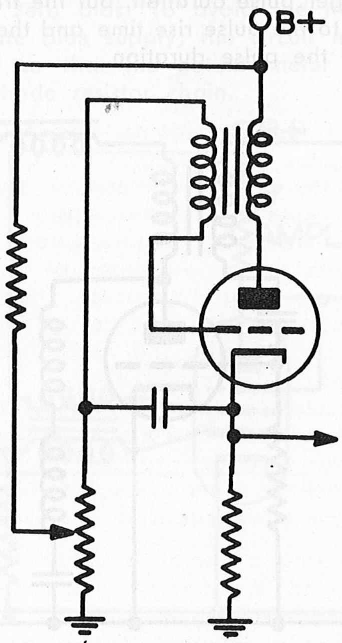

FIG.2

answer to the problem because the variation of each raises other problems. The range of control available by variation of R is approximately 25:1, this is ample for TV applications and is the method generally used. Where a wider range of control is required, a simultaneous variation of R and V_1 is used, as shown in Fig. 2. Other more elaborate means of controlling the repetition frequency for special applications are described in the references.

It is interesting to note that by using a diode or other means to control the voltage to which the grid capacitor is allowed to charge, the repetition rate of the oscillator can be made virtually independent of valve characteristics and mains voltage variations. When the oscillator is synchronised however, as in TV applications, the design of the oscillator for a very stable free-running frequency is not required.

Free-running blocking oscillators may be readily synchronised by the application of a series of pulses which lift the remaining bias prematurely and cause anode current to flow. Negative-going pulses may be applied to the cathode, or to the anode where they are transformer-coupled to the grid. Positive pulses may be applied to the grid directly if they have a high source impedance. If the source impedance is low, they are usually applied across a low-value resistor in series with the earthy side of the grid capacitor. A tertiary transformer winding could also be used. The oscillator is reasonably tolerant of the synchronising pulse shape, but ideally the shape should be as sharp as possible. The synchronising pulse should be of shorter duration than the oscillator pulse width. The pulse width is then largely independent of the trigger pulse duration, but the trigger may contribute to the pulse rise time and therefore indirectly to the pulse duration.

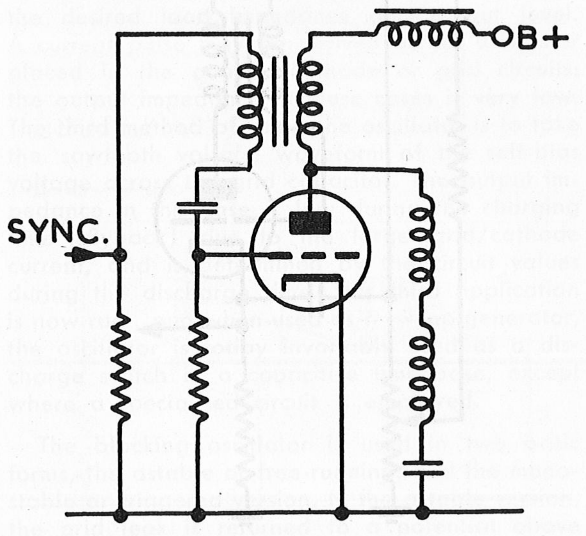

FIG. 3

Methods discussed here of linearising the sweep output of the blocking oscillator are largely of interest only, because in most TV applications the sweep is not supplied directly by the oscillator. The problem of producing a linear sweep cannot be isolated at the blocking oscillator, because a sweep output stage and coupling transformer are necessary, but must be dealt with on the basis of the complete deflection circuit. The generator output waveform must be shaped to give overall linearity. Except in simple cases, therefore, linearisation is outside the scope of this article, but an indication of the methods used is given later when two typical modern sweep circuits are discussed.

Where the oscillator is to be used directly as a sawtooth generator for electrostatic deflection, appreciable linearisation of the output can be achieved by returning the grid-leak resistor to a high positive potential, such as the B+ line. This increases the ratio between the total voltage across the grid capacitor and the incremental (charge) voltage, so that the discharge of the capacitor follows a substantially linear portion of the exponential discharge characteristic. This method can be used to provide a satisfactory time base for oscillography.

Many circuits have been evolved for the linearisation of blocking oscillator electromagnetic scanning generators, in which the linearisation is achieved by using a saturable reactor. This component has a variable value of inductance according to the DC current flowing in it and the consequent degree of saturation. The current/inductance relationship is arranged to be linear over the workng range. This device may be used where the sweep generator drives the deflection coils directly, but is obsolete as far as TV is concerned. Fig. 3 shows a representative circuit, several variations of which are possible. When the synchronising pulse arrives, anode current commences and the grid is driven positive. The reactor now controls the rate of change of current in the feedback transformer primary, and therefore the rate of rise of grid voltage. The rate of change of anode current is arranged by circuit design and the characteristics of the reactor so that a substantially linear change of current through the deflection coils results. The effect of the reactor is therefore to convert the sharp rise of anode current into a linear rise, thereby producing a linear change of potential across the deflection coils and DC blocking capacitor. When the inductance of the deflection coils is low, this will produce a substantially linear variation of current through them.

CIRCUIT VARIATIONS

There are many variations of the circuit shown in Fig. 1, and only the main variations can be mentioned here. A few variations are shown in Fig. 4. The grid/cathode coupled circuit shown is popular for TV applications, one advantage being its greater noise immunity. This results from the fact that the grid circuit is isolated from the output during the cut off period; the second or anode/cathode coupled circuit also has this advantage. A further advantage of the grid/cathode coupling is the absence of high DC voltages across the transformer between windings and between

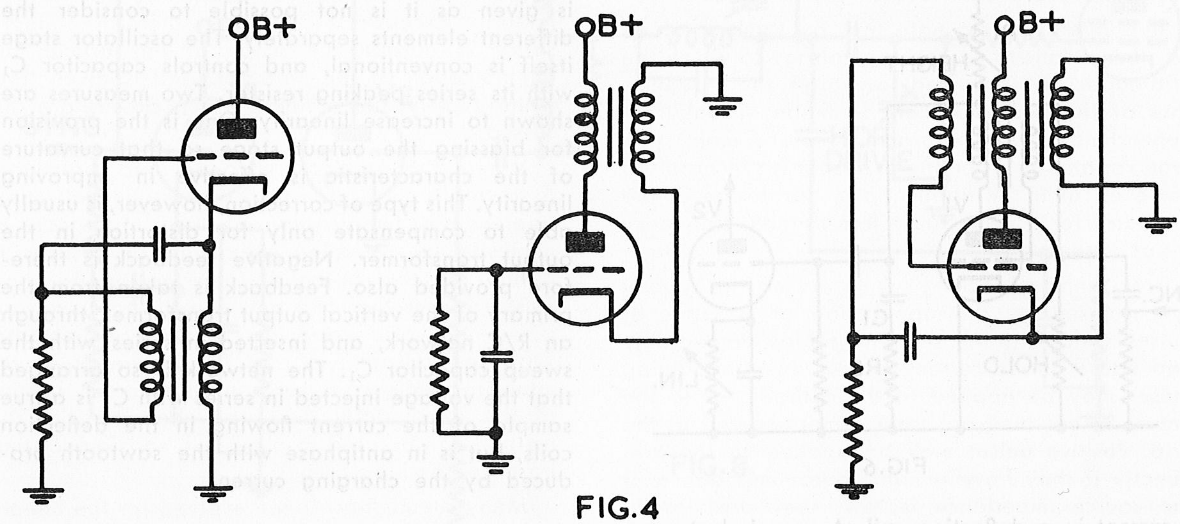

FIG.4

windings and core. Degeneration due to the cathode winding must be overcome by using a step-up ratio in the feedback transformer. If the circuit is so arranged that an element of negative feedback is retained, this will reduce the effect of changes of valve characteristics on the oscillator. The low value of stray capacitance at the grid renders this circuit particularly suitable for synchronisation from low-power sources.

On the anode-grid coupled circuit of Fig. 1, the degenerative effect of the transformer winding in the cathode circuit is not present, but the circuit has the disadvantage that noise arriving during cut-off is coupled to the output circuit through the inter-winding capacitance of the transformer. When used in TV applications, noise superimposed on the output can cause picture jitter. Other variations shown in Fig. 4 are the anode/cathode coupled circuit and the anode/grid/cathode coupled circuit. Both of these variations are rarely met, although the noise immunity of the anode/cathode coupled circuit should be comparable with that of the cathode/grid coupled version. The anode/cathode coupled circuit appears to be less susceptible to changes of repetition frequency resulting from power supply voltage variations than do most versions of the basic circuit. In some cases the grid (blocking) capacitor is placed in the cathode circuit. The basic manner of operation is the same as previously described except that in this case the cut off is accelerated by the fact that both grid and cathode current contribute to the charging of the capacitor, with a consequent decrease in flyback time.

An example is given in Fig. 5 of a pentode used as a blocking oscillator/discharge valve. The cathode grid and screen form the oscillator itself, whilst the anode controls the discharge of the capacitor. A variation of this circuit uses the anode alone to control the capacitor with a fixed charging potential. Amplitude control of output is provided by returning the suppressor grid to a potentiometer across a negative bias supply. Precautions must be taken to ensure that the potential of the suppressor grid does not approach too close to zero bias. To avoid the necessity for a separate bias supply, the circuit may be rearranged so that the potentiometer forms part of a cathode resistor chain.

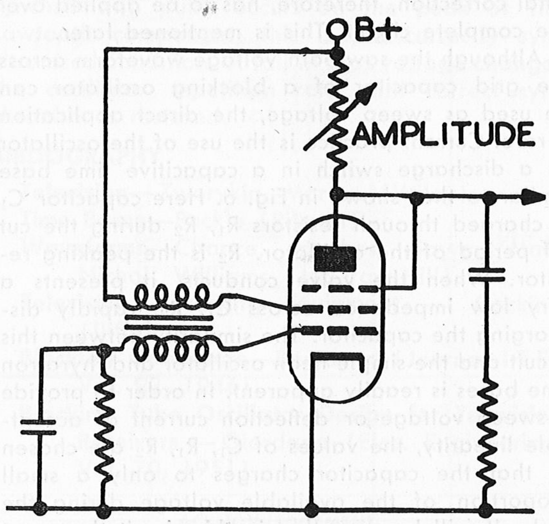

FIG.5

TV APPLICATIONS

Due to the diverse circuit arrangements used, it is possible to give only few examples of the use of blocking oscillators in TV receivers. It will be remembered that a peaked waveform consisting of combined rectangular and sawtooth components is required to produce a linear rise

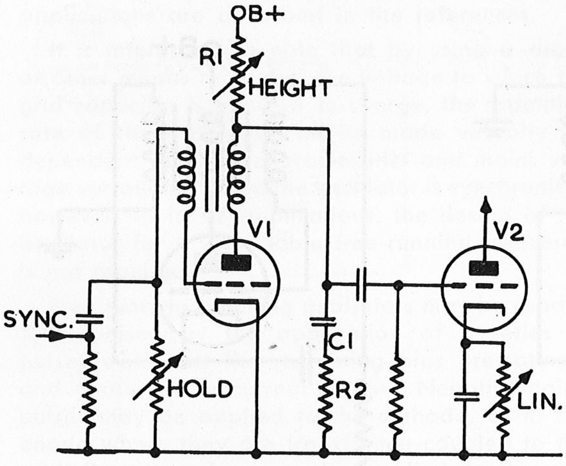

FIG.6

of current in a deflection coil. A pure inductance requires a rectangular waveform to produce the required deflection, whilst the inevitable resistance of the coil requires the sawtooth component. The proportions are a function of the particular coil. When the sawtooth is produced in a capacitive time base, a series or peaking resistor is used in series with the capacitor to produce the required output waveform. In practical applications this does not, of course, complete the formation of the waveform, the final shaping of which takes place in the deflection output stage and couplings. Final correction, therefore, has to be applied over the complete circuit. This is mentioned later.

Although the sawtooth voltage waveform across the grid capacitor of a blocking oscillator can be used as sweep voltage, the direct application is rare. Current practice is the use of the oscillator as a discharge switch in a capacitive time base similar to that shown in Fig. 6. Here capacitor C_1 is charged through resistors R_1, R_2 during the cut off period of the oscillator. R_2 is the peaking resistor. When the valve conducts, it presents a very low impedance across C_1, R_2, rapidly discharging the capacitor. The similarity between this circuit and the simple neon oscillator and thyratron time bases is readily apparent. In order to provide a sweep voltage or deflection current of acceptable linearity, the values of C_1, R_1, R_2 are chosen so that the capacitor charges to only a small proportion of the available voltage during the cycle. It will be seen that in this circuit the onset of oscillation in V_1 is affected by the rising anode voltage as well as by the relaxation of the self bias. The pulse duration of the oscillator (flyback time) must be sufficient to allow C_1 fully to discharge. The peaked sawtooth waveform at C_1 is R/C coupled to the ouput stage V_2, which is biassed by the linearity control into a curved portion of the characteristic where non-linear operation provides correction to the waveform.

A vertical deflection circuit using a blocking oscillator is shown in Fig. 7. This circuit is typical of modern practice. The complete basic circuit is given as it is not possible to consider the different elements separately. The oscillator stage itself is conventional, and controls capacitor C_1 with its series peaking resistor. Two measures are shown to increase linearity. One is the provision for biassing the output stage so that curvature of the characteristic is effective in improving linearity. This type of correction, however, is usually able to compensate only for distortion in the output transformer. Negative feedback is therefore provided also. Feedback is taken from the primary of the vertical output transformer, through an R/C network, and inserted in series with the sweep capacitor C_1. The network is so arranged that the voltage injected in series with C_1 is a true sample of the current flowing in the deflection coils, but is in antiphase with the sawtooth produced by the charging current.

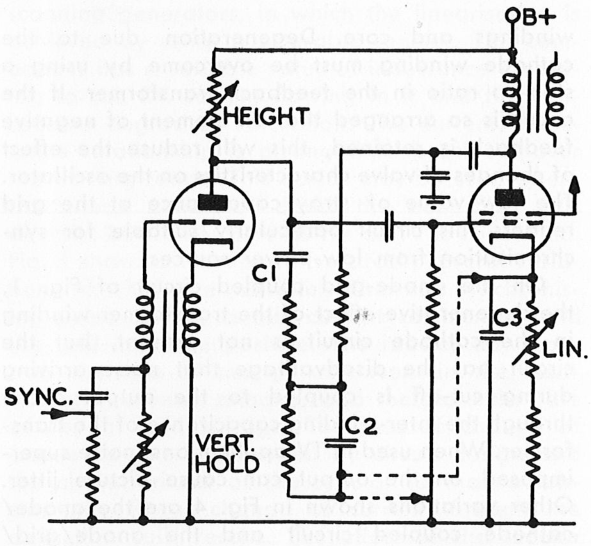

FIG.7

Alternative connections for the lower end of the charging circuit are shown, as both are used. Where the ground connection is used, a very large value of capacitor is required to by-pass the cathode resistor of the output stage to avoid the introduction of further distortion. This necessity is avoided with the cathode connection. Furthermore, whereas with the ground connection the grid/cathode voltage of the output stage is that on $C_1 + C_2 - C_3$, in the case of the cathode

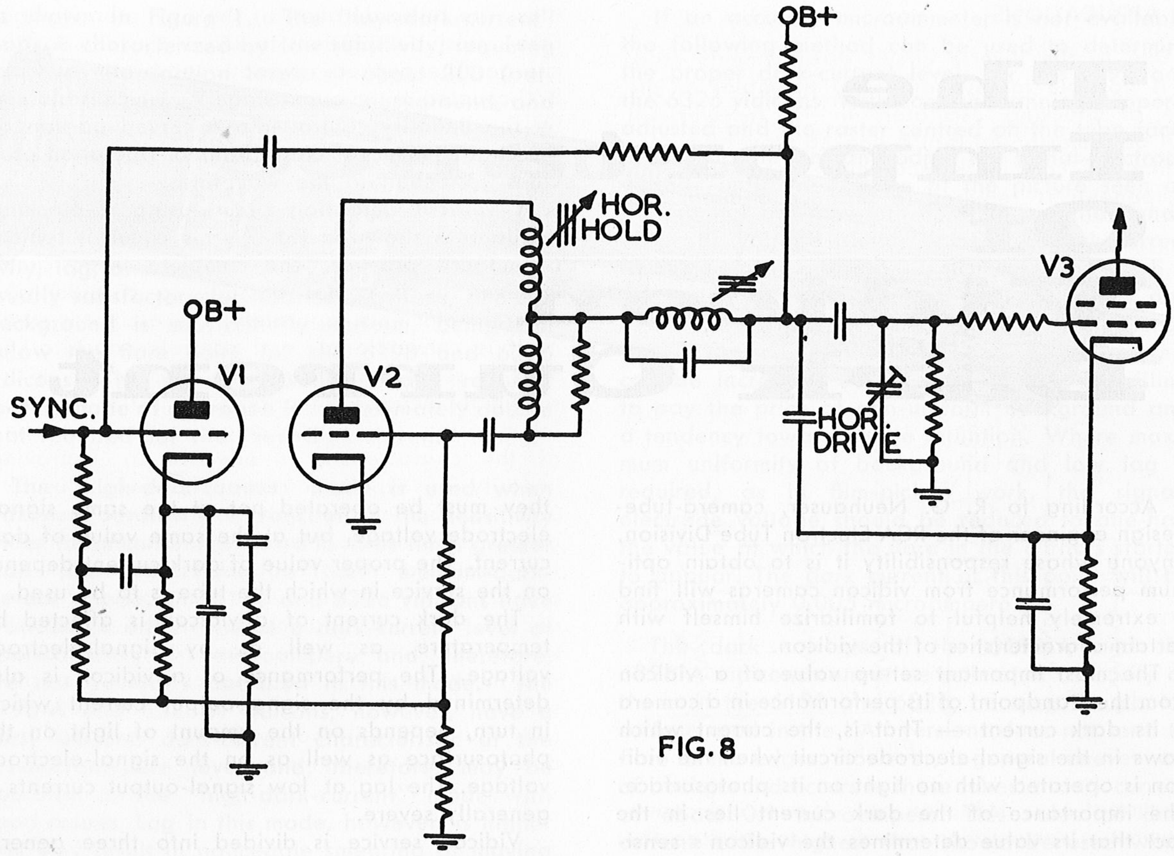

FIG. 8

connection it is that on $C_1 + C_2$ only, and the voltage on C_3 is added to the H.T. (charging) voltage. The bucking of the H.T. voltage by that of C_3 produces a more nearly exponential sawtooth waveform, which is more readily corrected than the S-shaped waveform produced by the ground connection.

Fig. 8 shows a blocking oscillator of interesting design used in a horizontal deflection system incorporating pulse-time automatic frequency control. The oscillator uses an auto-transformer with H.T. feed at the tapping point. In this oscillator, where a bias voltage is used to control the repetition rate, the cut off is accompanied by a damped oscillation train. The damped oscillations are filtered out by the parallel-connected inductor and capacitor in the H.T. feed to the auto-transformer. This parallel circuit is tuned approximately to the horizontal sweep frequency, and is shock-excited by the operation of the oscillator. These oscillations appear superimposed on the oscillator grid, and cause the grid to rise more sharply as it approaches the cut off point of the valve at the end of the cut off period. This makes the starting time of the oscillator pulse more certain, stabilises the oscillator, and prevents jitter in the horizontal sweep.

The AFC system depends for its operation on the simultaneous application of two voltages, one the synchronising pulse and the other a sample of the oscillator output to the grid of V_1, the oscillator control stage. Integration in this stage produces a cathode current, the peak value of which is limited by grid current and the magnitude of which depends on the area of the combined waveform input above cut off. It will now be seen that the cathode current of V_1 is at a maximum when the synchronising pulse occurs on the peak of the sawtooth component; this is the correctly synchronised condition. A fall in bias voltage changes the oscillator repetition rate until correct synchronism is restored.

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The Importance Of Vidicon Dark Current

According to R. G. Neuhauser, camera-tube-design engineer of the RCA Electron Tube Division, anyone whose responsibility it is to obtain optimum performance from vidicon cameras will find it extremely helpful to familiarize himself with certain characteristics of the vidicon.

The most important set-up value of a vidicon from the standpoint of its performance in a camera is its dark current — That is, the current which flows in the signal-electrode circuit when the vidicon is operated with no light on its photosurface. The importance of the dark current lies in the fact that its value determines the vidicon's sensitivity, uniformity of background, and lag (tendency of moving objects to smear). The sensitivity increases as the dark current is increased, at the expense of increased lag and decreased uniformity of background.

When two or more vidicons of the same type are set up so that they have the same sensitivity, uniformity of background, and lag characteristics, their signal-electrode voltages may differ by as much as three to one, but their dark currents will be identical. Similarly, if these vidicons are set up so their dark currents are equal, they will have remarkably similar sensitivities, background uniformity, and lag characteristics, although their signal-electrode voltages may differ by as much as three to one. This variation in the amount of signal-electrode voltage required to produce equal dark currents in vidicons of the same type is due to normal differences in photosurface thickness from tube to tube. Although any vidicon will operate over a three-to-one range of signal voltage, its dark current and, therefore, its performance in a given application will vary widely over this range. To obtain optimum performance from a vidicon in a given application, the signal-electrode voltage must be very carefully adjusted to provide the proper value of dark current for the application. Similarly, when two or more vidicons must perform identically in a given application — that is, have the same sensitivity, uniformity of background, and lag characteristics — they must be operated not at the same signal-electrode voltage, but at the same value of dark current. The proper value of dark current depends on the service in which the tube is to be used.

The dark current of a vidicon is affected by temperature, as well as by signal-electrode voltage. The performance of a vidicon is also determined by the signal-output current which, in turn, depends on the amount of light on the photosurface as well as on the signal-electrode voltage. The lag at low signal-output currents is generally severe.

Vidicon service is divided into three general types or modes, called the "low-", "medium-", and "high-dark-current" modes, corresponding, respectively, to operation with dark currents of approximately 0.005, 0.02, and 0.1 microampere

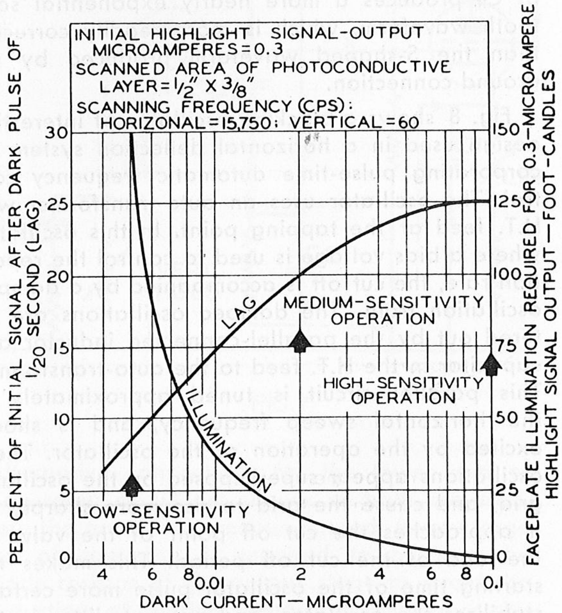


Fig. 1. Faceplate illumination requirements and lag characteristics of vidicons, as functions of dark current.

as shown in Figure 1. The "low-dark-current" mode is characterized by low sensitivity, requiring faceplate-illumination levels of about 200 foot-candles, by perfect uniformity of background, and by negligible lag. It is used for film-pickup work or outdoor pickup in bright sunlight, and requires the lowest signal-electrode voltage. The "medium-dark-current" mode is the one most frequently used and provides a good balance between sensitivity, lag, and background uniformity. The lag is usually satisfactory for most applications, and the background is substantially uniform, being just below the flare point for the 6198 and 6326 vidicons. The signal-electrode voltage required for this mode of operation is approximately double that required for the "low-dark-current" mode.

The "high-dark-current" mode is used when maximum sensitivity is required. The sensitivity under these conditions is two to three times greater than that obtainable in the "medium-dark-current" mode. The 6198 and 6326 vidicons have pronounced edge flare at a dark-current level of approximately 0.1 microampere and, therefore, are not generally operated in this mode. The 6198-A and 6326-A vidicons, however, have a fairly uniform dark-current characteristic at the 0.1-microampere level and, therefore, may be operated in the "high-dark-current" mode with good results. Lag in this mode, however, is higher and may result in noticeable smearing of moving objects and in a tendency towards long-lasting "image-burn". The signal-electrode voltage required for the "high-dark-current" mode of operation is about 50% higher than that required for the "medium-dark-current" mode, or three times that required for the "low-dark-current" mode.

Because dark current is the principal consideration in setup and operating adjustments for vidicons, facilities for the measurement of this current should be provided in all vidicon camera installations. The most suitable instrument for this purpose is a multirange microammeter having a full-scale sensitivity of 0.01 microampere or less. This instrument should be connected in the lead to the signal electrode, as shown in Figure 2.

To measure the dark current, set up the vidicon to produce a picture under normal conditions of light and lens opening for the mode of operation employed — i.e. adjust the scanning width and height to the proper values, and adjust the vidicon beam so that it just discharges the picture highlights. Then cap the lens and read the signal-electrode current. Next, cut off the vidicon beam and read the signal-electrode (circuit-leakage) current. The difference between the two readings is the true dark current of the vidicon. Adjust the signal-electrode voltage to obtain the desired value of vidicon dark current.

If an accurate microammeter is not available, the following method can be used to determine the proper dark-current levels for the 6198 and the 6326 vidicons. With camera scanning properly adjusted and the raster centred on the tube faceplate, cap the lens and adjust the signal-electrode voltage until the corners of the picture seen on the monitor begin to turn light. Record the signal-electrode voltage at this point. The dark current at this point is about 0.02 microampere, which is the optimum value for most industrial uses and live-scene pickups. If the highest possible sensitivity is desired, the signal-electrode voltage safely can be increased 50%, provided one is willing to pay the price of non-uniform background and a tendency toward image retention. Where maximum uniformity of background and low lag is required, as in film-pickup work, the signal-electrode voltage should be reduced to one-half the value at which the flare in the corners started to appear. The dark current at this point will be approximately 0.005 microampere.

The dark currents of the 6198-A and the 6326-A vidicons cannot be estimated as easily as those of the 6198 and 6326, because the photosurface used in the "A" versions does not tend to flare at the edges. However, a general unevenness of background occurs in these tubes at dark current of about 0.1 microampere. This value of dark current is the highest that should be used when the 6198-A and 6326-A are adjusted for maximum sensitivity.

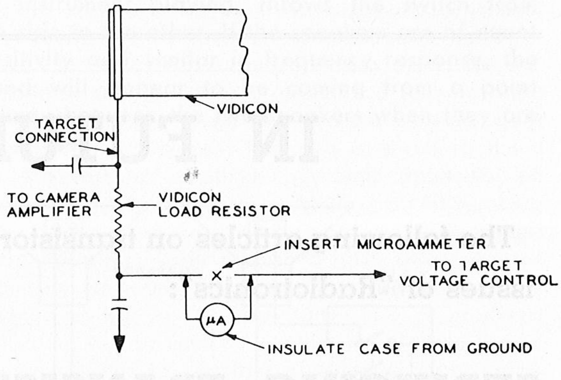


Fig. 2. Circuit used to measure the signal-electrode current of a vidicon camera tube.

In some vidicon cameras in which metering facilities are not provided, a calibration signal is made available which may be substituted for the vidicon signal. This signal usually corresponds to a vidicon signal-output current of about 0.3 microampere. Once the level of this calibration signal has been established on the waveform cathode-ray oscilloscope, it can be used to measure the signal-output current and relative dark current by comparing the amplitude of both the dark current and the signal current above the zero-current level.

Non-uniformities in wave shape of the dark current signal may make accurate measurement difficult, but adequate determination of approximate level is readily made.

Because the dark current of a vidicon increases with temperature, adjustments made when the camera is first turned on may have to be changed as the temperature rises to its stable operating value. The effects of dark-current drift with increasing temperature will be noticeable either as flare or as an increase in pedestal height or setup in the signal. Readjustment of the signal-electrode voltage will restore pedestal and flare to original conditions.

Another major consideration affecting the performance of a vidicon is the value of signal-output current. The signal-output current is the difference between the signal-electrode current developed in the dark and the total current developed during pickup. Excessive signal-output currents result in loss of resolution while low values result in unnecessary lag or smearing of moving objects. For best performance, the peak value of the signal-output current should be 0.3 to 0.4 microampere.

The average signal-output current can be measured by the same method used to measure the dark current. Peak signal-output is somewhat more difficult to measure. A good approximation can be obtained by the use of the current meter and video-waveform monitor oscilloscope. Focus the camera on a plain white card or wall, or on the open film gate of the projector, and measure the resulting signal-electrode current, which is approximately equal to the peak signal-output current. Also record the height of the resulting trace on the oscilloscope screen above the position of the trace with the lens capped. The oscilloscope can then be used to compare the peak signal-electrode current developed during any scene with the equivalent-peak value produced by the uniformly bright scene. After these measurements have been made, the video gain of the camera should be kept constant and light level controlled to maintain the peak signal value constant at the level observed on the oscilloscope.

The fact that the lag depends upon the signal-output current as well as on the dark current may introduce problems in certain applications. For example, if the light level on the tube is fixed, it is sometimes possible to reduce the lag by increasing the signal-electrode voltage so as to obtain both higher signal-output and dark currents. In these instances, there will be an optimum signal-electrode voltage where the lag will be at a minimum. Note that the curves of Figure 1 are given for constant signal-output level.

A common fault in vidicon camera setup is the practice of increasing the signal-output current just to a point where a satisfactory signal-to-noise ratio is attained. This practice usually results in a signal output that is too low to assure best performance from the standpoint of lag.

This article is reprinted with acknowledgements to RCA.

IN FUTURE ISSUES

The following articles on transistorized radios will be presented in forthcoming issues of "Radiotronics":

SERVICING TRANSISTORIZED RADIOS —see P.204

TRANSISTOR RADIO CIRCUITRY —P.183

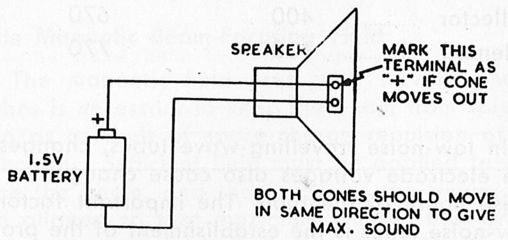
TRANSISTOR RADIO SERVICING TECHNIQUES —See Jan. 59, P.3.

Speaker Phasing

(With acknowledgements to RCA)

Sound is produced by the movement of a column of air impinging upon the eardrums of the listener. In the case of a loud speaker, the cone moves in and out, thus creating pressure and rarification waves, or movement of the air column in unison with the speaker cone movement.

Two speakers are often used to give a more uniform audio response. The sound intensity is maximum when the two speakers are "in phase", or when the cones of each speaker move in unison with each other. However, should the speakers be so connected that they are "out of phase", the cone of one speaker will move out when the other moves in. This will cause a cancellation effect and the intensity of the air movement reaching the ear will be reduced. Of even greater importance is the fact that the "out-of-phase" relationship of the two sounds will often result in distortion as well as cancellation.

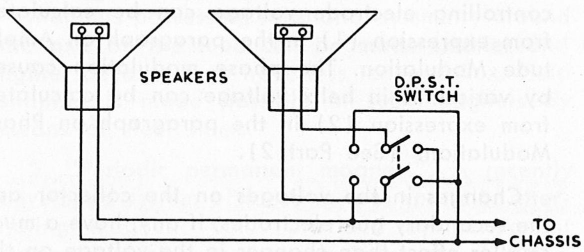


This phasing is of greatest importance where the speakers are positioned less than a few wavelengths apart. At 2000 cycles, one wavelength is between 5 and 6 inches, at 200 cycles this is between 50 and 60 inches. In the woofer/tweeter system found in some high-fidelity instruments, it seldom makes much difference whether or not the high-frequency speakers are in phase with the speaker reproducing the low and middle range frequencies. This is due to the fact that the tweeters have little effect below 5000 cycles.

The phasing is of greatest importance where two low frequency speakers are connected together in one cabinet. It is also important where an external accessory speaker is located near the internal speaker of the instrument.

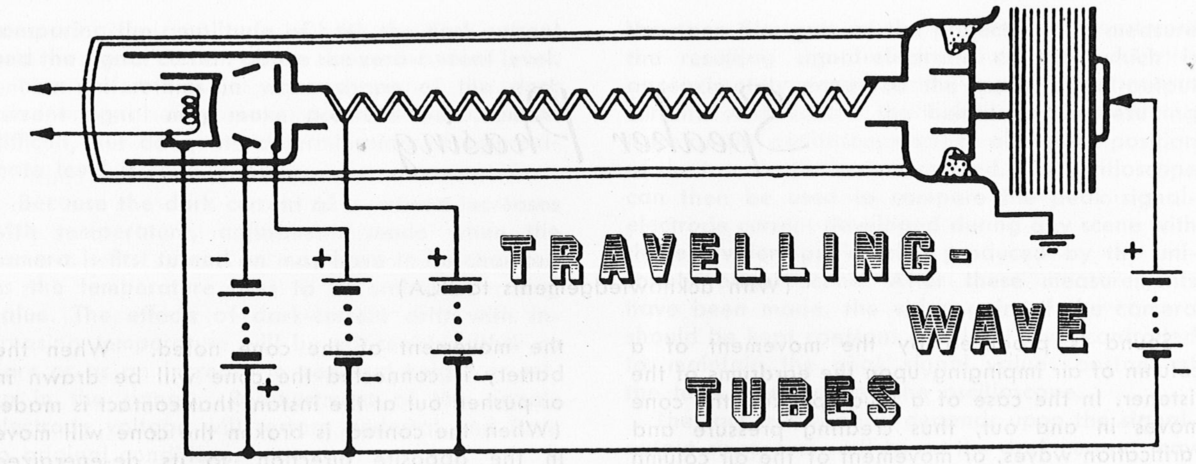
There are several methods that may be used to determine the polarity of speakers and if they are "in phase". One method employs a single cell ($1\frac{1}{2}$ volt) flashlight battery. The battery is connected across the voice coil of a speaker and the movement of the cone noted. When the battery is connected the cone will be drawn in or pushed out at the instant that contact is made. (When the contact is broken the cone will move in the opposite direction, to its de-energized position.) Considering that the cone has been PUSHED OUT, when contact was made, mark as (+) that terminal of the speaker to which the positive (centre) terminal of the battery was connected. The same test is then applied to the other speakers and their terminals similarly marked. These marked terminals are then the ones that must be connected together for proper phasing. As previously noted, phasing of tweeters has little effect on sound quality.

Another method of checking the phasing of speakers employs a double-pole double-throw switch and a listening test. The D.P.D.T. switch is prepared with long leads so that it may be connected in the speaker circuit in the instrument as shown. After connecting in the switch the listener goes to the front and as far from the instrument as the switch leads will permit, and with the instrument playing, throws the switch from one side to the other. If the speakers are of equal sensitivity and similar in frequency response, the sound will appear to be coming from a point midway between the two speakers when they are "in phase".



It must be kept in mind that when checking speakers with different frequency characteristics, such as a woofer and an intermediate, the frequencies that are common to both speakers are the ones that will be affected and are the ones that must be listened for in the audible test.

When performing these tests it will be found that the effect will be more readily noticeable if a constant tone signal is used rather than music or voice.



TRAVELLING-WAVE TUBES

PART 3
APPLICATION CONSIDERATIONS

Stability of Voltages

In a travelling-wave tube, the interaction between the electron beam and the circuit is essentially linear, and the broadband circuit has an essentially constant phase angle. The interaction circuit of the travelling-wave tube is electrically very long, usually ten wavelengths or more. Because the electron beam slightly alters the phase velocity of the wave travelling along the interaction circuit, slight changes in operating parameters can alter the total phase shift throughout the tube. The amount of change tolerable is, of course, determined by the particular application in which the tube is used. In microwave relay systems or certain radar systems in which phase-modulated or frequency-modulated information is amplified, even a very small amount of phase distortion is extremely disturbing. The phase modulation due to variations in beam-current-controlling electrode voltage can be calculated from expression (1) in the paragraph on Amplitude Modulation. The phase modulation caused by variations in helix voltage can be calculated from expression (2) in the paragraph on Phase Modulation. (See Part 2).

Changes in the voltages on the collector and the secondary gun electrodes, if any, have a much smaller effect than changes in the voltage on the beam-current-controlling electrode because they produce smaller changes in beam shape. Experimental tests have shown that the variation in phase from these causes is approximately one order of magnitude less than variations caused by the same percentage change in the current-controlling-electrode voltage.

Calculations for a typical low-noise travelling-wave tube show that for a maximum phase change of $+ 0.1°$, the following voltage variations can be tolerated:

Element	DC Volts (Approx.)	Maximum Ripple (RMS Millivolts)
G_1	0	10
G_2	20	10
G_3	40	50
G_4	200	350
Helix	375	2.4
Collector	400	670
Solenoid	100	770

In low-noise travelling-wave tubes, changes in the electrode voltages also cause changes in the noise figure of the tube. The important factor in low-noise tubes is the establishment of the proper voltage gradients between the cathode surface and the start of the interaction circuit. Under correct conditions, the initial velocity and current variations present at the cathode surface are suppressed to a large degree, resulting in what is termed a "de-amplification" of noise in the electron beam. The voltage variations permissible for minimum noise figure are much larger than those which cause significant phase distortion. Typical noise-figure curves for a 3000-megacycle low-noise travelling-wave tube are shown in Fig. 17. The variation in helix voltage affects both the noise figure and the gain, as was shown previously in Fig. 5. Changes in the heater power of a low-noise tube affect the cathode temperature and the emission, and thus affect the noise figure, as shown in Fig. 17. The effect of changes in grid No. 3 and grid No. 4 voltage is also shown in Fig. 17.

The gain of a travelling-wave tube varies directly as the third root of the current. A change in the voltage on the current-controlling electrode, therefore, changes the gain of the tube and may introduce amplitude modulation. Again, the amount of amplitude change permissible is determined by the application.

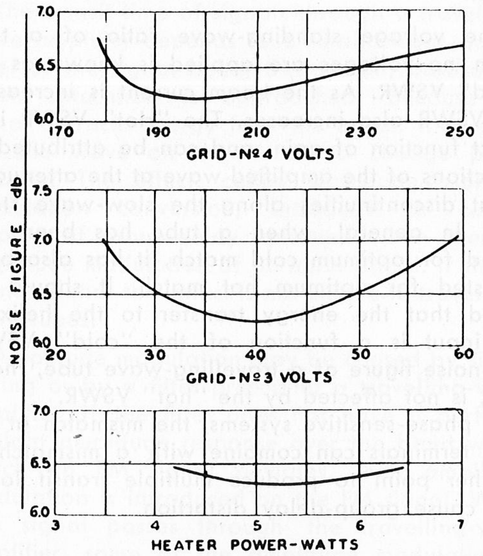


Fig. 17.—Effect of Variations in Heater power and grid No. 3 and grid No. 4 Voltages on Noise Figure of a Travelling Wave Tube.

The Magnetic Beam-Focusing Field

The magnetic field used with travelling-wave tubes is necessary to keep the beam from spreading as a result of space-charge repulsion of the electrons. Sufficient field must be applied to confine the beam, and the field and the beam must be aligned so that maximum transmission of the beam through the helix is obtained. A maximum helix interception current of the order of 3% to 5% of the beam current is usually tolerable.

In low-noise travelling-wave tubes, the problem of beam alignment in the magnetic field is more complex because helix interception current, especially at the start of the helix, causes partition noise and increases the noise figure of the amplifier. The alignment of a low-noise tube in the magnetic field must be such that the helix interception current is less than one per cent of the total beam current. In addition, the magnetic field must be considerably stronger than the value just barely necessary to obtain the minimum interception. The stronger magnetic field overcomes some of the beam-velocity variations and reduces the noise figure.

Nonuniformities in the focusing field may cause changes in the beam diameter, and produce an effect known as "scalloping". This effect changes the interaction of the beam and helix because different portions of the beam are closer or further from the rf field on the helix. External stray fields transverse to the beam axis can also cause this effect, and, if such fields are strong enough, complete beam interception by the helix may occur.

AC ripple in the magnetic field can cause a small ac phase variation by changing the beam diameter. However, because relatively large field changes are necessary to change the over-all phase angle, ripple is a minor problem. As a general rule, 1% ripple is tolerable for most applications.

Solenoids for travelling-wave tubes can be designed in a variety of ways depending on the ultimate application. For airborne applications, where weight reduction is important, it is possible to use aluminum-conductor solenoids and higher current densities at the expense of increased dissipation. This method is practical, however, only when the weight saved exceeds the additional weight necessary to supply the extra power and cooling. Solenoids wound of aluminum foil are particularly suitable for this application.

When greater reliability is desired at higher current densities, high-temperature wire insulation may be used to eliminate the need for an external blower. Permanent magnets show great promise for travelling-wave tubes because of several obvious advantages. Some information on the use of permanent magnets is given below:

1. "Uniform-Field" permanent magnets. Some permanent-magnet designs use a U-shaped mag-Because such a magnet has large leakage flux, it must be rather large and heavy to produce the desired magnetic field. For example, a U-shaped magnet used on a developmental 2000 Mc/s, 500-volt tube weighs approximately 20 pounds. As frequency and tube length increase, magnet weight also increases. A U-type magnet capable of producing 600 gausses across a 10-inch air gap weighs between 100 and 150 pounds.

Magnets possessing rotational symmetry around the axis of the tube can be made smaller and lighter because they have less leakage flux. Field shaping is made possible in such magnets by means of steel shims.

2. "Periodic permanent magnets". A recently developed system of focusing uses spatially alternating magnetic fields provided by permanent magnets which have considerably less size and weight. Although the weight of a uniform magnet is proportional to the third power of the length of the magnetic field, the weight of a periodic magnet is directly proportional to the field length. The periodic structure, therefore, has a weight advantage equal in the ideal case to the square of the magnetic-field length.

The basic concept of the operation of the periodic magnet is that the electron beam remains equally well focused when the vector of the magnetic field is in the same or the opposite direction as that of the electron flow. The magnetic-field distribution shown in Fig. 18 (a) can be realized by the physical structure shown in Fig. 18 (b). A cross section of a developmental medium-power travelling-wave tube using periodic permanent magnets was shown in Fig. 2.

A comparison of approximate magnet weights is given below for a developmental 3000-megacycle, medium-power travelling-wave tube.

Copper-wire solenoid (400 watt)	50 pounds
Uniform permanent magnet	100 pounds
Periodic permanent magnet	7 pounds

Recent advances indicate the possibility of further weight reduction. The choice of the type of magnet depends, however, not only on minimum-weight considerations, but on many other factors, such as tube structure, operating characteristics and application.

The use of electrostatic-focusing techniques would make possible further reductions in the size and weight of the travelling-wavetube "package". Research work is currently under way to determine the practicability of using such techniques.

Stray Magnetic Fields

The operation of the travelling-wave tube requires the use of an axial magnetic field for aligning the electron beam and keeping it collimated. Any transverse field tends to push the beam off centre and increase the helix interception current. Transverse fields arise from two major sources: (1) nonuniformities present in the focusing magnet, and (2) transverse components of stray fields present at the beam.

The transverse fields in the focusing structure can be minimized and practically eliminated by careful design and fabrication. Stray fields present a more serious problem, especially when the travelling-wave tube is mounted in close proximity to high-power magnetrons or klystrons which employ large magnetic structures. In such cases, it may become necessary to employ mumetal shielding around the solenoid. Mumetal shields are extremely effective. In one particular case, a 4J50 magnetron could be brought to within three inches of a shielded low-noise travelling-wave tube before the helix interception was affected appreciably. When the shield was removed, the travelling-wave tube was affected as soon as the magnetron was brought within three feet of the solenoid.

Alignment Procedures

The usual method for aligning a travelling-wave tube in the magnetic field is to apply all voltages except the beam-current control voltage. This voltage is then increased until a small amount of beam current flows, and the tube is aligned for minimum helix current. This process is repeated until rated beam current is obtained. Caution must be used in increasing the voltage, particularly with power tubes, because excessive current to the helix may cause a localized burnout.

Match

The voltage standing-wave ratio of a tube when no voltages are applied is known as the "cold" VSWR. As the beam current is increased, the VSWR also increases. The "Hot" VSWR is a direct function of gain, and can be attributed to reflections of the amplified wave at the attenuator or at discontinuities along the slow-wave structure. In general, when a tube has been adjusted for optimum cold match, it has also been adjusted for optimum hot match. It should be noted that the energy transfer to the helix at the input is a function of the "cold" VSWR. The noise figure of a travelling-wave tube, therefore, is not affected by the "hot" VSWR.

In phase-sensitive systems, the mismatch at the tube terminals can combine with a mismatch at another point to produce multiple transit loops and cause group-delay distortion.

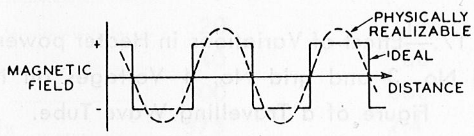


Fig. 18 (a).—Ideal Magnetic Field Distribution for Periodic Permanent Magnets.

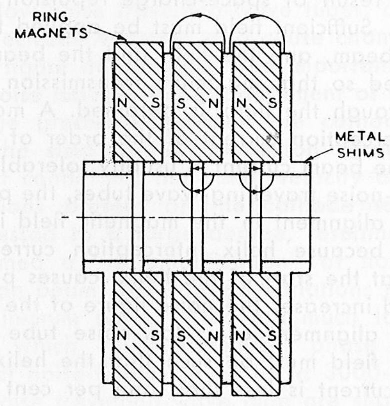


Fig. 18 (b).—Physical Structure used to Approximate Ideal Distribution.

Cooling Requirements

Sufficient cooling must be available at the collector to remove the heat dissipated by the tube and maintain the temperature of the seals below 175°C. Solenoids may also require cooling if the magnet wire insulation is enamel, so that the maximum winding temperature is not exceeded.

The solenoids may be designed for convection cooling as an alternative, especially if high-temperature insulation is used.

AM — PM Conversion

The transit time of signals through a travelling-wave amplifier depends to a small extent on the level of the signal being transmitted, particularly when saturation is approached, because the signal extracts energy from the beam, thus causing the wave on the circuit to change phase velocity. The effect, called amplitude-to-phase-modulation conversion, introduces a certain amount of group-delay variation when there is amplitude modulation of an input signal. It should be noted that this effect is not unique to travelling-wave tubes.

Amplitude modulation may be caused by signal fading or by a filter preceding a travelling-wave amplifier. If the filter does not have a perfectly straight amplitude response over the band within which an FM signal deviates, then amplitude modulation is introduced on the FM signal. When this signal passes through the travelling-wave amplifier, some of the amplitude modulation is converted to phase modulation because of the delay variation through the tube caused by the signal level. The resulting group-delay variation within the band of deviation may cause distortion in a frequency-modulated signal passing through the amplifier. This effect is extremely sensitive to signal level, and exists in measurable quantities only when the tube is operated at or near saturation. The effect can be overcome by operating the tube well below the saturation level.

The amount of AM — PM conversion is usually so small that it becomes important only in applications involving great sensitivity to phase. In microwave FM relaying of colour-television signals, for example, the cumulative distortion caused by passing the signal through ten or more relay stations may no longer be negligible.

AM — PM conversion may also affect the noise figure of travelling-wave tubes used in an FM or other phase-sensitive system. At low carrier-to-noise ratios, the input signal is effectively a carrier plus a band of noise having a total power at some given level with respect to the carrier. The output signal is effectively a carrier containing both FM and AM. In a conventional FM receiver, the AM is removed by limiters, and the FM component remains to provide a certain signal-to-noise ratio on demodulation. If, however, the travelling-wave tube converts some of the AM noise to PM, this PM noise cannot be removed by limiters, but adds to the noise already present in the receiver and degrades the output signal-to-noise ratio. The maximum amount of degradation usually is in the order of 1 db, and is negligible provided the output of the travelling-wave tube is at least 3 db below saturation.

Interaction Effects between Two Signals

When a strong signal and a weak signal are simultaneously fed into the same travelling-wave tube, the degree of amplification experienced by the weak signal is essentially determined by the operating point on the output-vs-input-power curve (shown in Fig. 3) established by the strong signal. The gain of the weak signal, therefore, varies with the level of the strong signal, particularly near the saturation point of the tube, and inter-modulation distortion may be produced in the weak signal.

In a similar manner, the action of a strong signal can produce phase modulation of a weak signal by AM-to-PM conversion. Particularly in the operating region near saturation, the phase velocity of the weak signal is affected as the level of the strong signal is varied. This phase modulation produces a spurious sideband on the other side of the strong signal, as shown in Fig. 19 for a 3000-megacycle low-noise travelling-wave tube.

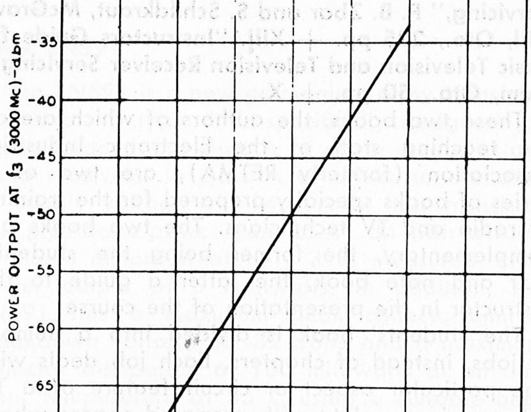


Fig. 19.—Spurious Response Characteristic produced by Interaction Effects between two signals.

Associated Microwave Circuitry

A low-noise travelling-wave tube located before a local-oscillator-mixer combination may cause an apparent 3 db degradation in noise figure unless an output filter is used to remove noise generated in the travelling-wave tube at the image frequency.

When the tube feeds into a filter, frequencies in the "reject" band are reflected back into the tube. As the wave travels back through the tube, it suffers little attenuation until it is absorbed by the attenuator. If there is appreciable reflection from the attenuator or from other discontinuities inside the travelling-wave tube, oscillations may occur depending on the gain from the attenuator or discontinuity to the end of the tube. Travelling-wave tubes are often designed to be short-circuit stable, i.e. the reflections from the attenuator are smaller than the gain of the tube from attenuator to output. Although this condition may exist at the rated value of beam current, the gain increases as the current is increased beyond the rated value until a value of gain is reached at which the system will oscillate.

In the use of a high-gain microwave amplifier tube such as the travelling-wave tube, special care must be taken to prevent distortion or oscillations due to feedback through external circuitry. Coaxial or waveguide switches are sometimes at fault. In addition, some types of filters may show satisfactory attenuation characteristics in and near the frequency band of interest. However, because of the extremely large bandwidth of most travelling-wave tubes, oscillations can still occur as a result of holes in the filter characteristic at frequencies far removed from the band of interest. Attenuation of filters should be checked over wide bands, therefore, and the holes, if any, should be filled in by supplementary simple filter designs.

New Publications

"Basic Television and Television Receiver Servicing," P. B. Zbar and S. Schildkraut, McGraw-Hill, Qto., 295 pp. + XIII. "Instructors Guide for Basic Television and Television Receiver Servicing", idem, Qto., 50 pp. + X.

These two books, the authors of which are on the teaching staff of the Electronic Industries Association (formerly RETMA), are two of a series of books specially prepared for the training of radio and TV technicians. The two books are complementary, the former being the students' text and note book, the latter a guide to the instructor in the presentation of the course.

The students' book is divided into a number of jobs, instead of chapters. Each job deals with one particular aspect or circuit feature of a TV set, and is complete with prepared spaces where the student can fill in measurements and observations on the job. Whilst it is not to be claimed that this form of presentation is original, it is ideally suited to the purpose in hand. A total of 55 jobs tackled under expert guidance could not fail to enhance the technical standing of the student, especially when they contain such a wealth of practical "on-the-bench" data and wrinkles.

Although the students' volume is primarily designed for organised instruction, this does not preclude its use for private study, or as a reference book.

Summing up, this book would be difficult to improve for class work, and at the same time will, it is felt, fill admirably the need of the serviceman or technician who wishes to improve his knowledge and experience of TV.

The Instructors' Guide, as its name suggests, is for the information and guidance of the class instructor. It deals with preparation of jobs, test gear, bench arrangement, and all the other multitudinous things that engage the attention of the instructor before the class presents itself. The data and notes it contains are such as the average instructor normally gathers over quite a period of time, and are certain to ease considerably the instructor's task. With a specialised book of this kind it is difficult to review in general terms. Let it be said however, that both the Instructors Guide and the students' volume reflect the dynamic and practical approach of skilled instructors to the problems of trade training.

"High Fidelity Sound Reproduction", Gen. Ed. E. Molloy, George Newnes Ltd., 8to, 200 pp. + XII.

Within the compass of 200 pages this handy volume deals with all aspects of "Hi-Fi" reproduction, with each chapter written by a specialist in the particular subject. A nice balance is struck between over-simplification and weighty treatment of the subject, and this has produced an informative and pleasant-to-read volume which has an appeal for everyone from the beginner to the old hand in the fascinating world of "Hi-Fi". That the book necessarily leans towards U.K. practice is no criticism as this is to be expected.

Noted with pleasure were the inclusion of stereo reproduction and a complete chapter on the controversial electrostatic speaker, whilst the chapter on speaker enclosures by E. Jordan of Goodman's Industries contains constructional details of a number of enclosures. The text contains, in addition to the usual explanatory diagrams, many commercial circuits (with values) of amplifiers, preamplifiers, tape recorders and tuners to illustrate current practice and design trends. This volume carries a very interesting introduction by Mr. H. J. Leak, of "Point One" fame, whose concluding remarks I adopt as my own — "I hope you will enjoy it as much as I have."

RADIOTRON 7212

The Radiotron 7212 is a small beam power valve electrically interchangeable with the 6146, but designed specifically for applications where dependable performance under severe shock and vibration is essential. It is intended for use as an rf power amplifier and oscillator as well as an af power amplifier and modulator.

The 7212 has a maximum plate dissipation of 25 watts under ICAS conditions in modulator and CW service. In the latter service, it can be operated with full input to 60 Mc and with reduced input to 175 Mc. Because of its high power gain and high efficiency, the 7212 can be operated with relatively low plate voltage to give large power output with small driving power.

RADIOTRON 6J6-WA

The 6J6-WA is a premium medium-mu twin triode of the 7-pin miniature type. It is constructed and processed to meet the exacting requirements of military specifications. The 6J6-WA is intended primarily for use as a class A amplifier and control valve in mobile and airborne equipment, where uniformity of characteristics and dependable performance under conditions of shock and vibration are primary considerations.

RADIOTRON 5670

The 5670 is a "premium" medium-mu twin triode of the 9-pin miniature type. It is constructed to give dependable performance under conditions of shock and vibration, and is similar in characteristics to the 2C51. The 5670 is intended for use in r-f or i-f amplifiers, mixer and multivibrator circuits, and in oscillator circuits in industrial and communication equipment.

AWV 2N591

The 2N591 is a new germanium alloy transistor of the p-n-p type. It is specifically designed for use in large-signal class A af driver stages of car radio receivers. In class A af driver service at a dc supply voltage of -14.4 volts, this transistor can provide a power gain of 41 db with a total harmonic distortion of only 3%, measured at a power output of 5 milliwatts. In addition, the 2N591 has a maximum peak collector-emitter breakdown voltage of -32 volts and a maximum collector dissipation of 100 milliwatts at an ambient temperature of 55°C using a suitable heat sink. The 2N591 has the Jetec TO-1 outline with flying leads.

180.

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