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70-WATT HIGH-FIDELITY STEREO AMPLIFIER

In February of last year we published some information on an experimental high-fidelity system, and this system was demonstrated to several interested audiences. One of the main objectives of that exercise was to show that transistors have a place in high-fidelity. We feel that the demonstration was completely convincing and, moreover, material published elsewhere since then, and some of the commercial units that have appeared on the market, particularly overseas, has supported these views.

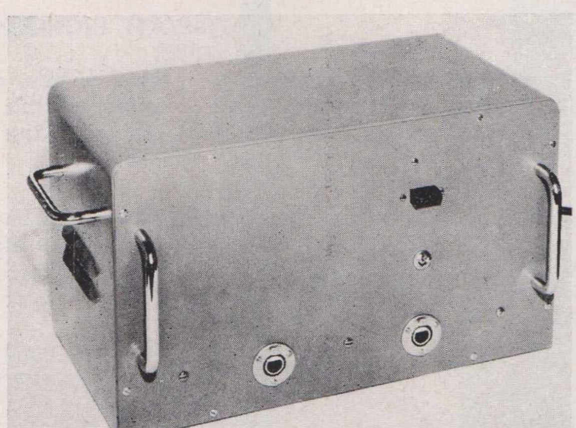
Even further than this again, several indications have appeared in the literature of the art that there is some grounds for thinking that "transistor sound" is superior in some respects to that produced by more conventional equipment. We do not propose to enter this phase of the argument at this stage, not because the arguments do not appear to have merit, but because some of the investigations supporting the theories have not yet been completed.

All of this is assuming that the transistorized unit on which any argument is based is one of good quality. Unfortunately, it cannot be denied that whilst many fine equipments have been made available, particularly overseas, units have also appeared which leave something to be desired. This is a natural sequence of events at any time, and we know the tendency for a new technique to be branded with the shortcomings of the worst examples that are available.

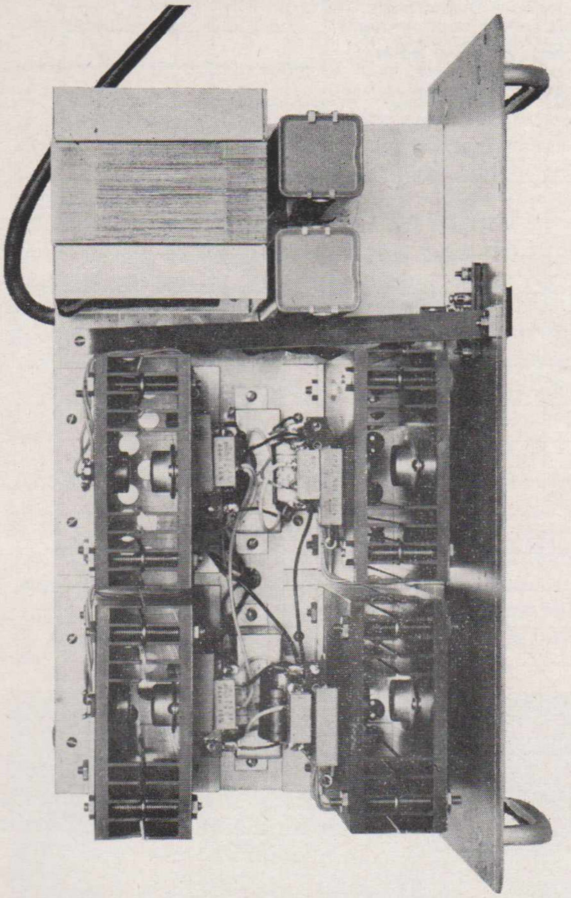
Looking back over the events of the last two years in the high-fidelity field, it cannot now be denied that transistorized high-fidelity is here, and is here to stay. Further, it has been amply demonstrated that most of the drawbacks claimed for transistorized equipment either do not exist, or are far from being the stumbling blocks that they were originally claimed to be. Typical is the question of cost.

There was a great tendency in the earlier days to assess the cost of the semiconductors in a design and immediately claim that the unit was too costly. It is true that with transistorized amplifiers, the proportion of the total cost represented by the purchase of the active devices will often be higher than in the case of roughly comparable units using electron valves. A careful survey of the position, however, will generally show that the cost of other components is lower and that, in many cases, although transistorized high-fidelity is comparatively young, costs will favourably compare even at this time.

Still on the subject of cost, it is usually possible to discard the costly output transformer, and the power supply can be a simpler design with a transistor amplifier. It is interesting to note that some of the better transistor amplifiers are so good that it would be very difficult, if not impossible, to make an output transformer to suit



Photograph of the 70-watt stereo main amplifier unit.



View of the interior of the amplifier, from the top, showing the arrangement of the heat sink sub-assemblies, mains transformer and main filter capacitors.

the performance of the amplifier. This is the case with the amplifier which forms the subject of this article. To this extent, it could perhaps be claimed with some truth that the transistorized amplifier is moving into an area where the conventional electron valve amplifier cannot follow.

It is realised that there are some specialized amplifiers using electron valves that do not use an output transformer. This is generally done by designing for a low value of load impedance, and then coupling a speaker directly to the amplifier. The speaker will generally be one with a much higher impedance than normal, 45 and 80 ohms being typical values. Whilst these amplifiers perform very well, it is a fact that the unusually high load impedance represented by the special speakers used reduces the flexibility and application of the amplifier. What is needed for general use is still an amplifier capable of driving standard 15-ohm speakers.

Some of the foregoing may seem to be in the nature of a digression, but it is useful in setting the stage on which the amplifier to be described is to perform, and establishing its place in the art of high-fidelity as it exists at present.

Past History

We took a course in the original release of data on the experimental amplifier which is contrary to our usual policy. Normally, we would not release such information through these pages until all the work we wished to do on the design had been completed and evaluated. The preliminary release of experimental data was partly engendered by the speed with which developments in transistorized amplifiers were appearing, and partly by the enormous and gratifying number of enquiries for information that followed the earlier demonstrations.

Since that time, a great deal of further work has been done on the amplifier, some of an investigatory nature, some to rationalise component values and ratings, and some with a view to making a stereo unit. It will be remembered that when the amplifier appeared in its original form, the stereo system consisted of two complete main amplifiers with their own power supplies, together with a preamplifier powered from one of the main amplifiers.

The position at the moment is that a suitable mains transformer has been obtained and a stereo main amplifier is now available. We had thought at one time that it would be a good thing to integrate a preamplifier with this final stereo version, but this was decided to the contrary. For one thing, the type and nature of the preamplifier used would depend on the type of use that the builder had for the amplifier. Secondly, we appear at the moment to be moving into a further generation of preamplifiers, which, if present promises are fulfilled, will result in much better circuits within the reasonably near future than we have available for publication at this time.

This is one of the problems of the art at the moment, that developments both in the devices themselves, and the techniques of using them, are coming along very fast. They can sometimes overtake the normal delay of three to six months in getting developments into print. At least two units over the last two years have been so badly overtaken in this way that plans to publish them were abandoned.

But against this we have the stimulation of things happening, the excitement of promises to come, and the certain knowledge that out of it all we are making substantial progress. From a comparatively static field, high-fidelity has become a forward-reaching branch of the electronic art.

Design Considerations

One of the limitations of power transistors until recently has been the relatively low cut-off frequency, with the attendant high phase shift and restrictions in the use of feedback. The development of the power drift type with a gain-bandwidth product $f_T = 4$ Mc has greatly reduced this problem, and made it possible to design high power amplifiers with flat response extending to 100 Kc.

The causes of differences other than amplitude between the output and the input signals, normally referred to as distortion, can be grouped under the following three main headings:—

(i) non-linear transfer characteristics for each half of the push-pull stage,
(ii) asymmetry between halves of the push-pull stage,
(iii) cross-over transients caused by hole storage effects.

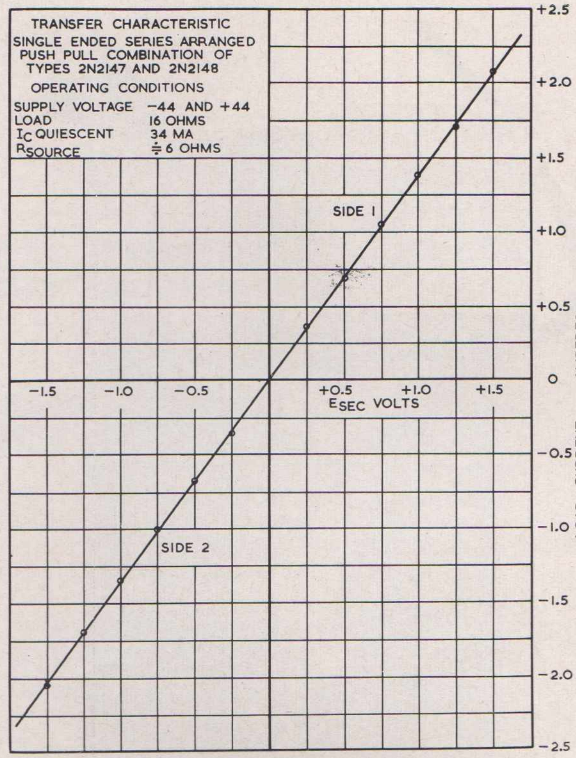
Of these, the first item (i) produces mainly odd-harmonic distortion and has a number of causes. The major factors are the non-linear input impedance of the transistor and its non-linear transfer characteristic. Each of these is a result of the variations of a number of parameters, e.g., Z_{in} in the common emitter configuration is given approximately by $Z_{in} = r_b + (h_{fe} + 1)r_e$, where Z_{in} is the input impedance, r_b is the base resistance of the "T" equivalent circuit, r_e is the emitter resistance of the "T" equivalent circuit, and h_{fe} is the common emitter current gain. Since r_b, h_{fe} and r_e all change to a varying degree with emitter current, the variation of Z_{in} can be very complex. The method of driving the base then must be very important in reducing this form of distortion.

Until recently, for common emitter circuits the use of a constant voltage source (low impedance) has been preferred to the constant current or high impedance source, since the transfer characteristic of collector current versus base-to-emitter voltage has been more linear. One writer[1] has drawn attention to the very linear transfer characteristics, particularly in the cross-over region, which can be obtained using a high source impedance drive. However, the amplifier described here uses a relatively low impedance base drive.

The transfer characteristic for a set of AWV 2N2147 and 2N2148 transistors, operating under the conditions of the amplifier being described here, is shown in an accompanying diagram. It will be seen that a line drawn through the plotted values is a straight line.

The reduction of distortion due to asymmetry between halves of the push-pull stage, mostly even harmonics, is complicated, because although the parameters r_b, h_{fe} f_{hfe} can be matched at one or a few conditions, they each vary with I_E and should be matched over the full range of I_E experienced in the design.

The smaller the variation, the easier it is to match over the whole range of emitter current,



Transfer characteristic of a set of AWV 2N2147 and 2N2148 power drift transistors, operating under the conditions obtaining in the 70-watt main amplifier.

and here the 2N2147, with its linear relationship between these characteristics and the emitter current, is of great assistance.

The third, and perhaps unexpected cause of disturbance in a transistor high-power wide-band audio frequency amplifier is the hole storage effect. It is found that as the speed of switching a transistor on and off increases, a point is found where the base and emitter currents will not exactly follow the switching signal. A current will flow in the base circuit after the base-to-emitter voltage has been reversed to switch the transistor "off." This current is in the reverse direction to the normal forward bias current, and is due to the minority carriers (holes in p-n-p type), which are in the base region and which must be removed before the base impedance can become high in the switched-off condition.

The higher the frequency, the more pronounced are these effects, since the magnitude of the hole storage effect is a measure of the switching time of the transistor. The rapid change in base and emitter current at the step, and the increased sharpness of the switch-off as a result of the hole-storage current, can induce ringing in any transformer in the associated circuit. The use of transistors with a high cut-off frequency and therefore lower hole-storage effects, and the re-

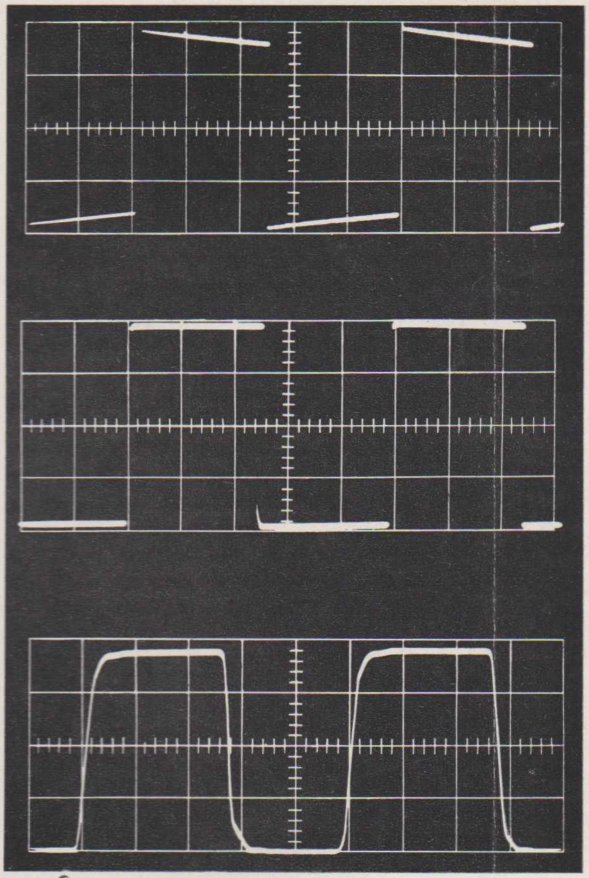
[1] "High Impedance Drive for the Elimination of Cross-over Distortion," I.R.E. Trans. on Audio, July/Aug. 62, 99. J. J. Faran and R. G Fulks.

output transformers, is required to minimise the disturbance.

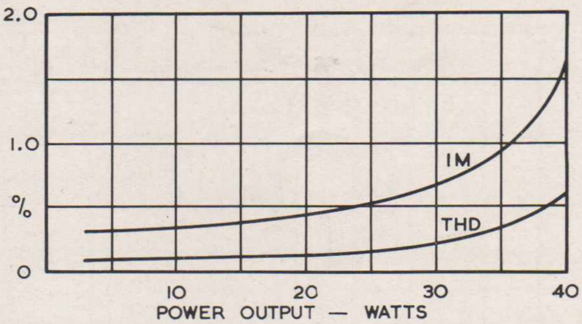
It is worth mentioning at this point that when transistors are used in series to increase voltage or power ratings, their switching speeds are most important. As is the case with semi-conductor rectifier diodes in series, they can switch with different time delays, and produce a very unequal division of the peak voltage across each. One or more of the following precautions must be taken: (a) ensure similar switching times by selecting transistors and/or the use of an appropriate circuit technique as discussed later, (b) suppress the resultant "spike" or (c) use devices with a breakdown voltage rating adequate for the worst combination of conditions.

Main Amplifier Circuit

It has already been mentioned that this amplifier has been designed around the AWV 2N2147 and 2N2148 power drift transistors, and this design is typical of those referred to as being able to stand favourable comparison with similar units



Oscillograms of the square-wave response of the main amplifier at frequencies (reading from the top) of 100 cps and 1 Kc, for a power output of 25 watts, and at 20 Kc for a power output of 10 watts.



Total harmonic distortion and intermodulation distortion of the main amplifier, versus power output.

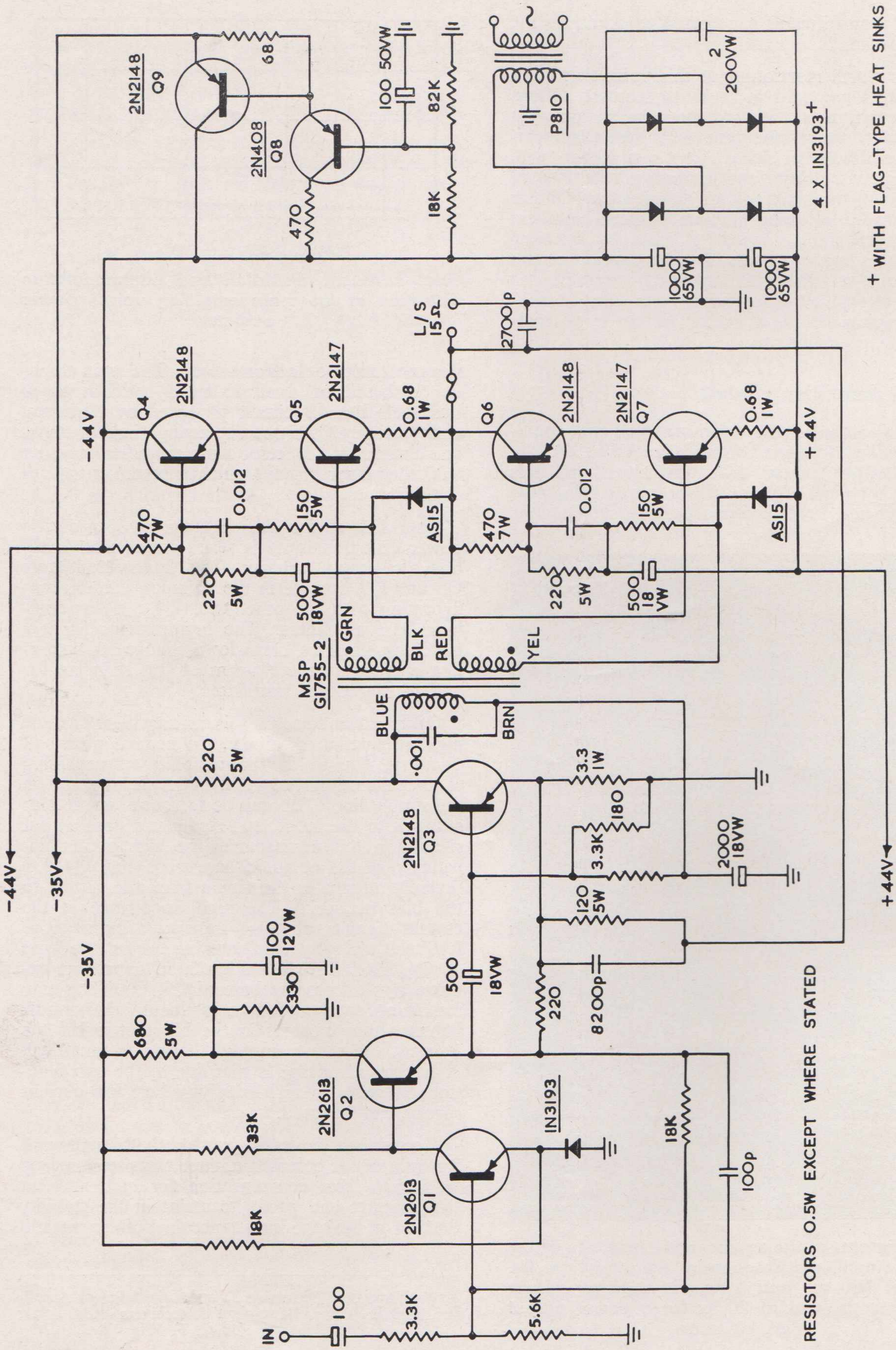
designed around electron valves. The main amplifier is based on a single-ended series-arranged push-pull class B circuit proposed by Kleinman and Wheatley.[2] An accompanying diagram shows one channel of the stereo main amplifier, together with the power supply, which is common to both channels.

Each half of the class B stage consists of two power drift transistors in series across a relatively high dc supply voltage. The driven transistors Q5 and Q7 operate in the common emitter configuration with the other two (Q4, Q6) as common base amplifiers. The requirements for h_{FE} matching, linearity, breakdown voltage and saturation current are therefore much less stringent for half of the output transistors.

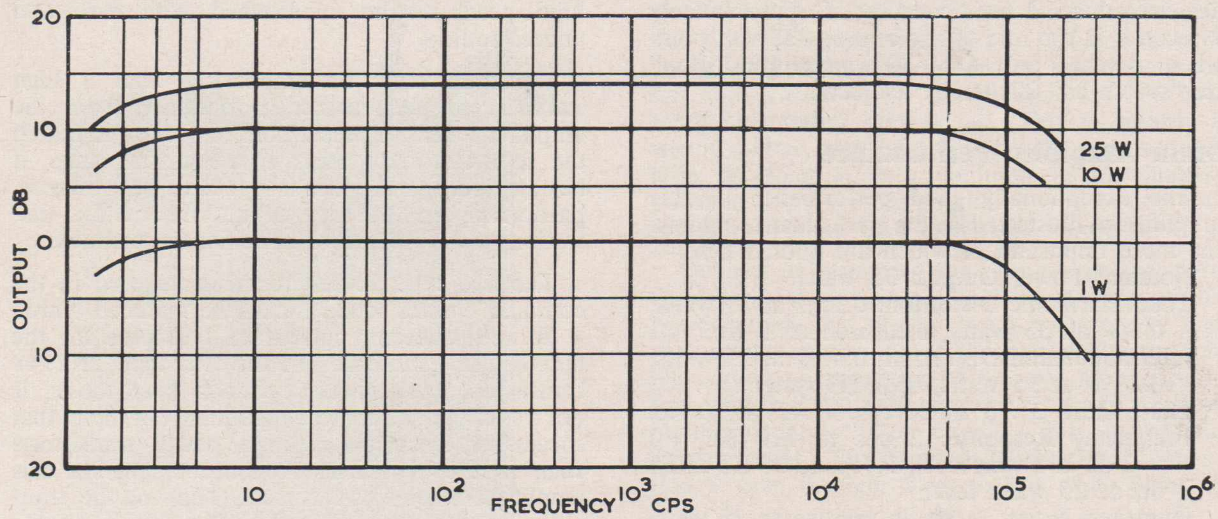
Other advantages of this arrangement include high attenuation of ripple. In normal push-pull circuitry, due to the impossibility of providing exact balance, an added hum ripple appears on the supply line. This can be fed back via the bias network to the input, and result in a hum output signal. In this arrangement the high output impedance of the common base transistors Q4, Q6 severely attenuates the ripple from the supply to the bias network. The high impedance in the emitter circuits of these transistors by negative feedback virtually eliminates ripple on the base of Q4, Q6. The 500 microfarad capacitors remove ripple from the base of Q5, Q7. The only remaining source of hum is due to the out-of-balance hum currents in the bias networks. As a result the power supply filtering is uncritical. A further advantage is that due to the two junctions in series, a high-supply-voltage low-current supply can be used.

Germanium diodes are used in the bias network for each of the common emitter transistors, partly to provide bias compensation for high ambient temperature, and partly to maintain the stability duction of leakage inductance in both input and

[2] "An Ultra Low Distortion Transistorized Power Amplifier," I.R.E. Trans. BC and TV Receivers, July, 1961, H. M. Kleinman and C. F. Wheatley.



Circuit diagram of one channel, with the power supply which is common to both channels.



Frequency response of the main amplifier, taken at power levels of 1, 10 and 25 watts.

of the operating point with signal level and thus produce a better transient response.

In amplifiers of this power rating it is possible under certain circumstances for high peak voltages to be developed across the output transistor. The conditions causing the development of these high voltage "spikes" are the application to the amplifier of high level signals of low or high frequency, either momentarily or continuously. The momentary application can be caused by a noisy gain control, the inadvertent connection of the pickup to the preamplifier with the gain control at the maximum gain position, or the connection of a measuring instrument to one of the preamplifier stages. In each case the signal level can be sufficient to drive the output stage to produce up to a 70-watt output which will be of square wave form. This will be true for signals of any frequency within the range of the system, and will produce a square wave output with a rise time of approximately 4 microseconds. Under these conditions the relative switching times of the two series-connected transistors in each half of the amplifier are important.

In this amplifier the development of high peak voltages across any one of the output transistors is prevented by the connection of 12,000 picofarad capacitors across each of the base resistors of the common base transistors. These capacitors serve to "speed up" the application of the cutoff signal to the common base stages. A variation in the value of the "speed-up" capacitors will allow the time at which the common base transistors (Q4, Q6) cutoff to be adjusted to occur before or after their respective common emitter drivers. If either of the capacitors is omitted, Q4 or Q6 will switch later than Q5 or Q7 respectively, with the result that a very high voltage "spike" can appear between the collector and emitter of Q4 and Q6. The condition giving rise to this situation were mentioned previously.

A power drift type is used as the driver, since power with low distortion at high frequencies is required. The coupling transformer is RC coupled to the driver, and uses grain-oriented silicon steel. The secondaries are bifilar wound between the halves of the primary. Negative dc feedback to the base of the driver provides very good stability for the operating point.

Two direct-coupled stages are used in the predriver, with the high-performance low-level audio-frequency transistor 2N2613. Q2 operates as an emitter follower and provides a low impedance source for the relatively low input impedance of the driver. The silicon diode in the emitter circuit of Q1 provides very good stability for the operating bias of the two predrivers, and thus maintains low distortion.

Negative feedback totalling approximately 39 db is used around the driver and output stage. At 1000 cps, the voltage feedback between the load and the driver emitter provides approximately 20 db, while the driver emitter resistor accounts for approximately 18 db. At 20 cps, the combined feedback is 27 db, while at 100 Kc it is 15 db. The loop around the predriver from the emitter of the emitter follower to the base of Q1 provides at 1000 cps approximately 21 db of negative feedback.

The power supply for the amplifier is a very simple one. Two full-wave circuits are used to produce separate negative and positive 44-volt supply voltages, which are each filtered by a single 1000 microfarad capacitor. These two supplies are used for the output stage, whilst dc power for the driver and predriver stages is obtained from the —44 volt point through a "dynamic" filter using a 2N2148 and a 2N408. This arrangement provides a —35 volt supply with a ripple of about 4 millivolts, which represents about 60 db attenuation from the —44 volt point. The arrangement also introduces a time constant of approximately 2 seconds in the rate of rise of the —35 volt point when switched on. A 50-70 watt output "plop" on "switch-on" is thereby avoided.

Main Amplifier Performance

The exceptionally good performance of this amplifier is illustrated in the performance figures, the more important of which are quoted here:—

Nominal Power Output: 35 watts.

Total Harmonic Distortion: 0.15% at 25 watts, 0.3% at 35 watts, measured at 1 Kc.

Intermodulation Distortion: 0.45% at 25 watts, 0.87% at 35 watts, SMPTE system.

Phase Shift: 3°.15′ at 20 cps, 4°.42′ at 20 Kc.

Frequency Response: 2 cps to 110 Kc +0 —3 db at 1 watt level, 10 cps to 35 Kc ±0.5 db at 25 watts level.

Hum and Noise: —80 db relative to 35 watts output.

Sensitivity: 120 mv rms at 1 Kc for 25 watts, 140 mv rms for 35 watts.

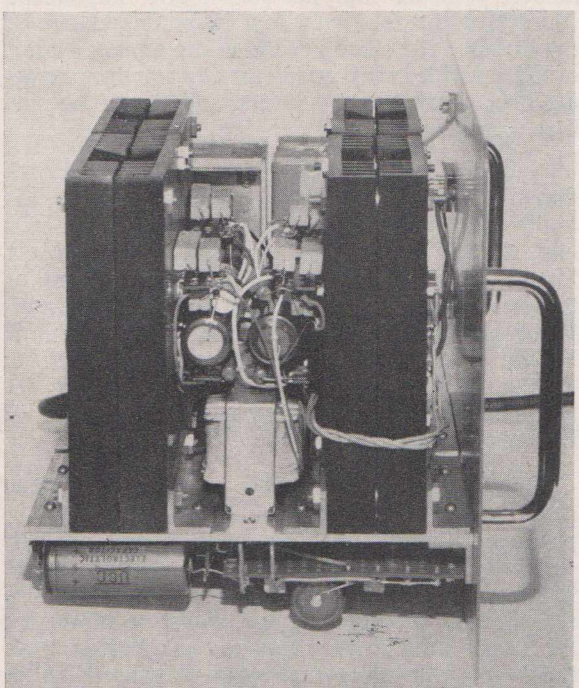
Output Impedance: 1.1 ohm.

Damping Factor: 14.5.

Load Impedance: 15 ohms.

Construction

It must be pointed out that this amplifier is a comparatively advanced design. It is therefore not recommended to beginners, not because the design as published is not safe, but because the building of a high-power amplifier of this kind needs to be approached with care and understanding.



Side view of the amplifier with the case removed, showing further details of the assembly.

Whilst it must be emphasised that a high standard of performance is obtained from the amplifier when no precautions are taken to match the semiconductors used in the output stage, it has also been found that to optimise performance, particularly in the area of intermodulation distortion, some degree of matching is desirable.

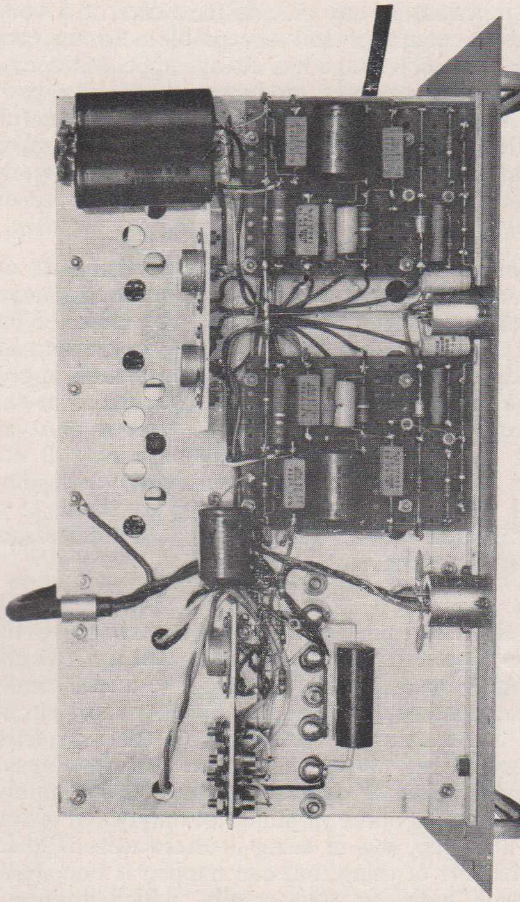
The two type 2N2147 transistors used in the common emitter mode should be matched within 4 db at a collector current of 1 ampere for the best results. Matching of the two type 2N2148 transistors, used in the common base mode, is not required. It must be pointed out here that successful operation of high power transistors, such as those used in the output stage of this amplifier, relies upon correct fitting of the transistor on the heat sink, and silicone grease should always be used to reduce the thermal resistance between the transistor and the heat sink.

The two AWV compensating diodes type AS15 have a bearing on the intermodulation distortion and the question of thermal stability. For the best results, these diodes should be matched in pairs for use in this amplifier, and should be so ordered. Further, flag-type heat sinks should also be ordered and used with them, and the diode heat sinks should be bolted to one of the heat sinks of the output stage; this latter requirement ensures the correct functioning of the diode as it holds the diode at essentially the same temperature as the output transistors.

The general layout adopted in the construction of our models is shown in the photographs. In basic terms, each side of each output stage (involving two transistors and one diode) is assembled as a sub-assembly around the two heat sinks. Four of these sub-assemblies then make up the two output stages. The driver transformers are located between pairs of the sub-assemblies.

The power drift transistors used for the driver stages do not require heat sinks as such, and are mounted on a bracket under the 12-gauge aluminium bedplate on which the amplifier is constructed. The pre-driver stages for each amplifier are constructed on small sections of matrix board mounted under the bedplate. One of these assemblies is the mirror-image of the other in order to bring the input leads and grounds into the most favourable positions. A second bracket under the bedplate carries the rectifiers and filter circuit.

It must be noted that at the time these photographs were taken, we were using larger stud-mounted diodes in the power supply, instead of those called for in the circuit diagram. However, it was decided later that the cheaper diodes now called for would be entirely satisfactory, and these are the ones that should be used. The case



Underside view of the amplifier, with case removed, showing assembly of the driver and pre-driver stages, main rectifiers and dynamic filter.

used for our amplifier is an Imhof type 1054A, approximately 15″ wide, 8″ deep and 9″ high, excluding the handles. Cannon Series XL plugs and sockets were used as we intended the units for demonstration purposes, involving more rigorous use than might otherwise be the case. However, the choice of this type of component is at the discretion of the builder.

It will be seen in the photographs that the portion of the aluminium bedplate covered by the heat sink assemblies is pierced with a number of ventilating holes, although the class B operation of the unit does not give rise to a lot of heat. It would be neater and better in a finished design to cut out a complete rectangle under the finned areas of the heat sinks. At the same time, provision must be made in the underside, and either the top or the top of the back, for ventilation through the heat sinks. The heat sinks are Telecomponents type 7002, 4″ x 6″; smaller heat sinks, or those with a smaller equivalent area, must not be used.

Turning to the question of wiring, and assuming that a layout similar to that we used is employed for the pre-driver stages, giving fairly short leads, the only point to mention is the question of grounds. This is most important, and experience has shown that unless the grounds are run correctly, instability may occur. This is entirely a wiring matter, and not a design fault. In fact, this type of thing is very common in the higher-powered transistorized amplifiers, and can be corrected by proper lead dress and sequence.

Dealing in turn with the grounds in the main amplifier and power supply, and starting at the power supply end of the circuit diagram, we first pick up the centre-tap of the mains transformer, the common connection of the two filter capacitors, and the 100 mfd capacitor and 82K resistor in the dynamic filter circuit. This forms the starting point. The common loudspeaker connection is then brought back to this same point.

We then have four more grounds in each amplifier in the driver and pre-driver stages. These are: (a) the junction of the 3.3-ohm and 180-ohm resistors in the driver, (b) the earthy end of the 2000 mfd capacitor in the driver circuit, (c) the 330-ohm resistor plus the 100 mfd capacitor decoupling the pre-driver stages, (d) the 5.6K resistor and diode in the input stage. The method of dealing with these will now be described.

The two connections (a) are commoned in the two amplifiers by a bridge, shown in the photographs. To the centre of this bridge are then brought the other connections (b) through (d), plus the ground already mentioned in connection with the power supply. Finally, a ground is taken from the common connection of all these leads and connected to the chassis at the input socket. The ground brought in by the 3-core mains lead may be connected to the chassis at any convenient point.

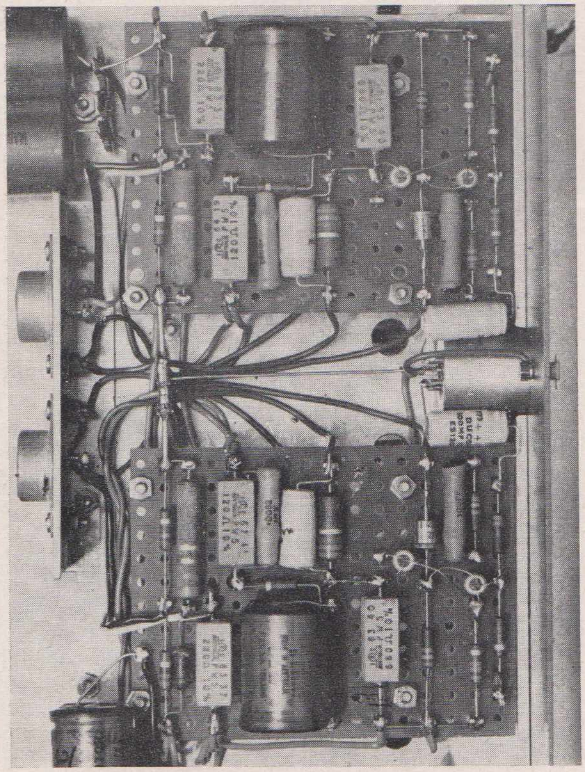
Testing

The fact that it is very easy to cause catastrophic failure in semiconductors by incorrect application of operating voltages makes it important in this type of amplifier to carry out a very careful circuit check before applying power to the circuit. We have ourselves evolved a basic routine for this type of thing, based on the fact that as there are no heaters to worry about in semiconductor units, it is easy to vary the operating potentials by control of the mains input to a unit.

Our starting-up routine therefore consists of a very careful circuit check, preferably carried out by someone other than the person who wired the unit. Then, when power is applied first, a small voltage only is applied, using a tapped auto-transformer or a "Variac." The initial application of voltage could be, say, 20% of the working voltage. Do not apply any signal at this stage, but a resistive dummy load must be applied.

The procedure now is to make a quick check on the voltages around the circuit, with a view to ascertaining that they are approximately what would be expected with the lower input voltage, that is, in about the same proportion. This has often revealed a wiring or circuit fault in our laboratory without causing damage to the equipment. In the matter of this amplifier specifically, at this stage one should check (a) that the voltage division across the two halves of the output stages is equal or almost so (zero or near-zero dc voltage across the dummy load), and (b) that the division of voltage between the two junctions in each half of the output stages is about equal.

By way of example, the dc voltage under quiescent conditions between the live output terminal (with resistive load connected) and ground, and with full operating voltage applied, should be less than 1 volt, polarity unimportant, and will generally be considerably less, about a quarter of a volt or less being common. With lower applied voltages, similar or lower figures will be found. Turning now to the two halves of the output stages, we have Q4 and Q5, and Q6 and Q7. Each pair of junctions takes half the total applied voltage, and half this voltage should in turn appear across each junction. However, the equal sharing of the voltage is not so important in this case as in the two halves of each output stage. A difference in the voltages appearing across the two junctions in any pair of the order of 5 volts or so is common and acceptable. So far, the slightly lower voltage has always appeared across the common emitter mode transistor of a pair. Here again, voltages are quoted for the full applied voltage, although it has in fact been found that most of the difference will be already noticeable when the amplifier is being run up on reduced voltage.



Close-up of driver and pre-driver stages, showing layout and common ground arrangements.

Assuming that everything appears normal on the initial reduced voltage test, it is then usual to raise the voltage in steps of, say, 20% of the full voltage, rechecking as one proceeds that all indications are correct. The reason for not applying signal during these checks, apart from the obvious one of avoiding peak voltages until all tests well, is that the pre-driver stages will not pass the required signal level without clipping until the applied voltages are up to about 75% of the rated values. This could cause unnecessary worry unless it were known.

It must be pointed out most firmly that this amplifier, in common with others that use the same basic circuit configuration, will not tolerate a short-circuit across the output terminals under driven conditions. Even a momentary short-circuit can cause catastrophic failure of the four transistors in the output stage, and may also result in consequential damage to the power supply. This must be borne in mind not only when testing the amplifier, but in using it afterwards. Typical of the sort of thing that can happen is one stereo balance indicator sold locally, which applies a momentary short-circuit as a switch on the indicator is moved from a "setting" to a "use" position. Extreme care must therefore be used in regard to this type of amplifier, not only when working on the amplifier, but also in connecting up to other equipment.

Once the initial checks just described have shown that there is no wiring or circuit error, then the rest of the testing programme can be carried out in the normal way. There will be some who will say that this is an unnecessary warning, but in view of the extended bandwidth of this amplifier, it is essential that the response of the meters and other indicators used in testing the amplifier be linear over the range of frequencies concerned, or spurious results will be found.

Analyses

It has not been our practice in the past to present voltage or current analyses, because of the well-known fact that figures measured on a specific unit may vary considerably from those quoted, whilst the apparatus is working perfectly. Any such figures that are quoted must therefore be regarded as a guide only, and interpreted as such; they do not represent precise figures which must be found to ensure correct operation.

Two analyses have been prepared for the assistance of those who would like to build this amplifier. The first analysis consists of a listing of voltages measured at strategic points in the circuit, with full mains voltage and resistive load applied, but without signal. Figures were taken over two stereo units, i.e., over four main amplifiers; the four sets of figures are given, together with an average of the four.

The second analysis that has been prepared is concerned with signal levels. These measurements were taken on one channel at a time, with the input signal level adjusted for an output of 25 watts rms at 1 Kc. As in the previous case, an averaged figure has been added, although here it is only a point of interest.

Preamplifier

There is a reason for the fact that, so far, no mention has been made of a preamplifier for this unit. The type of preamplifier, and the facilities required, will vary with different users. Readers will remember, however, that a lot of information was recently published in these pages on the subject of preamplifiers (October, 1964), and readers are referred to the previous article for suitable circuits.

Measured at	Amplifier 1		Amplifier 2		Average
	LH	RH	LH	RH	
Q1 Collector	—3·3	—3·4	—3·4	—3·6	—3·42
Q1 Base	—0·75	—0·75	—0·73	—0·71	—0·74
Q1 Emitter	—0·62	—0·62	—0·6	—0·59	—0·61
Q2 Collector	—10·3	—10·1	—9·8	—9·8	—10·0
Q2 Emitter	—3·1	—3·2	—3·1	—3·4	—3·2
Q3 Collector	—13·5	—14·0	—13·6	—14·9	—14·0
Q3 Base	—0·7	—0·62	—0·69	—0·65	—0·67
Q3 Emitter	—0·45	—0·46	—0·45	—0·42	—0·45
Q4 Collector	—48·0	—48·0	—48·5	—48·5	—48·25
Q4 Base	—21·0	—21·0	—21·8	—22·0	—21·45
Q5 Collector	—21·5	—21·0	—21·0	—21·8	—21·65
Q5 Base	—0·1	—0·13	—0·23	—0·08	—0·135
Q5 Emitter	0·78	0·58	Zero	—0·16	0·3
Q6 Collector	0·82	0·68	0·08	0·17	0·44
Q6 Base	27·5	26·0	27·8	28·0	27·3
Q7 Collector	28·0	26·0	28·0	28·2	27·55
Q7 Base	49·0	49·0	49·0	49·0	49·0
Q7 Emitter	49·5	49·5	49·5	49·5	49·5
Supply "—35 V"	+49·5, —48·0 —38·5		+49·5, —48·5 —38·5		— —38·5

Voltage Analysis
Measurements made with a 20,000 ohms/volt meter on appropriate ranges.

Measured at	Amplifier 1		Amplifier 2		Average
	LH	RH	LH	RH	
Input	115	92	110	120	109
Q1 Base	5·8	5·8	6·0	6·1	5·9
Q1 Collector	585	580	555	598	579
Q2 Emitter	570	560	540	576	562
Q3 Base	570	560	540	576	562
Q3 Collector	2,820	2,700	2,750	2,750	2,755
Q3 Emitter	560	550	530	570	553

Signal Analysis
Signal levels in millivolts for 25 watts output.

We ourselves have used these amplifiers for demonstration purposes in reproducing records. We therefore used the previously-published module with RIAA replay characteristic, followed by gain and balance controls, and then the same module again with feedback tone control. This gives us a system which provides a gain of about 40 times (32 db) at 1 Kc, allowing for a 6 db margin in the setting of the gain control. Thus a 5 mv input signal gives us about 200 mv output, plus an available 6 db margin. This is more than adequate to drive the main amplifier being described here.

There is just one small point to note in relation to the connection of a preamplifier, and that concerns the 100 mfd electrolytic input capacitor. As shown in the circuit diagram, the polarity is correct for the connection of a circuit which applies a dc voltage between about —0.5 volt and a positive voltage depending on the rating of the capacitor. Where the external circuit applies a higher negative voltage, as will be the case if the capacitor is connected directly to the collector of a preceding p-n-p stage, or to the emitter of a common collector p-n-p stage, then the polarity of the 100 mfd capacitor will have to be reversed. This must then be remembered when testing the unit individually.

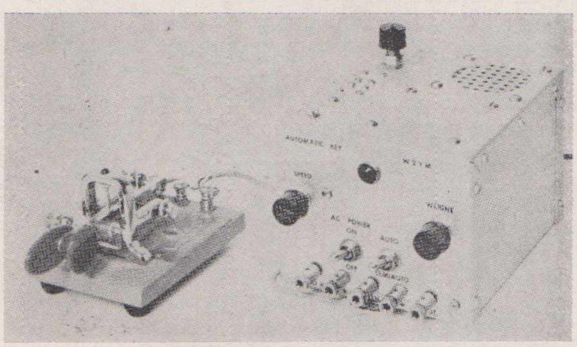
A TRANSISTORIZED KEYER

By George D. Hanchett, W2YM

RCA Electronic Components and Devices

The subject of semi- and fully-automatic keying has long held considerable fascination for the author, who has traced its development since introduction of the electronic semi-automatic key, or "bug," in the middle thirties. Various developments since then have resulted in numerous and sundry systems, all employing either relays or combinations of electron tubes and relays to perform the keying function. To the author, it seemed that the next logical step in the evolution of keying systems should be the design and construction of a practical transistorized electronic keyer. The compact unit described in this article represents his efforts in that direction.

The transistorized keyer may be operated either as a semi-automatic key (automatic dots) or as a fully automatic key (automatic dots and dashes). The keying function is performed by a high-speed relay, which is located at the output of the unit so that the keyer will be electrically isolated from the circuits being keyed. A double-pole relay is employed so that one set of relay contacts may be used to mute the receiver during the key-down condition. Keying speed is controlled by a voltage source in order to permit the use of a single potentiometer instead of the ganged dual potentiometer usually required for this purpose. Another feature of the keyer is the built-in tone oscillator which allows the operator to monitor his keying at all times.

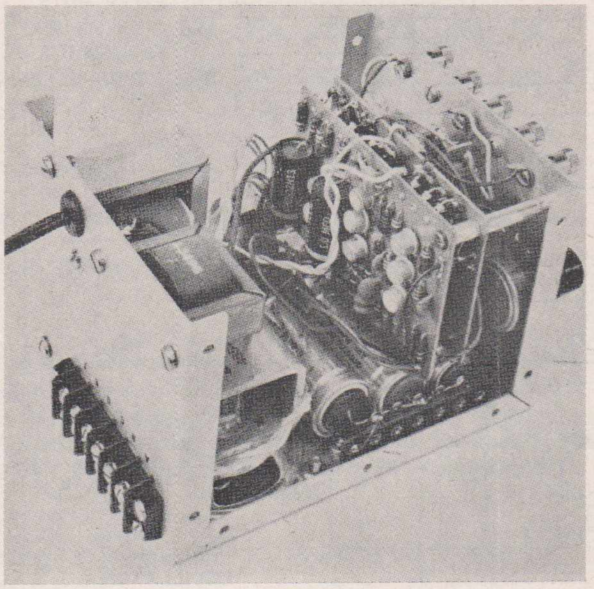


W2YM's transistorized keyer can be operated either as a semi-automatic "bug" key or fully automatic key. Instrument is shown here with a Vibroplex "Vibro-Keyer" connected to it. Standard hand key can be connected to two binding posts at left.

Circuit Details

The schematic and parts list of the transistorized keyer appear in Figure 1. The actual keying circuits consist of a free-running multivibrator, a flip-flop multivibrator, an OR gate, and a transistor-controlled relay circuit. A half-wave rectifier provides the DC voltage to control the keying speed, and a voltage double from the 6.3-volt winding of the power transformer provides the DC supply voltage for the unit. A tone oscillator provides an audible indication of the keying.

The dot multivibrator, as its name implies, controls the formation of the dots, and the repetition rate of this multivibrator determines the rate at which the dots are produced and hence the speed of the keying. When the "Vibro-Keyer," S_1, is in the open position, the multivibrator is held inoperative (transistor Q_2 is not conducting) by the biasing action of clamp-transistor Q_3. When the paddle of S_1 is moved to the dot position, the clamp-transistor becomes inoperative, and the dot multivibrator becomes a free-running circuit. The square-wave signal developed at the emitter of multivibrator transistor Q_2 is then applied to the base of transistor Q_7 in the OR gate. During the positive alternation of this signal, the OR gate will permit current to flow through the relay-control transistor, Q_9, and through the keying relay, K_1, in series with this transistor.



Bottom view of keyer shows location of speaker, transformers, and filter capacitors. Terminals "A" through "F" (left) are connected to the double-pole, double-throw relay contacts.

Once a dot is initiated by moving the paddle of S_1 to the dot position, the action will continue—regardless of the position of the paddle—until both the dot and the space that follows it are formed. This feature is provided by the feedback circuit from the base of clamp transistor Q_3 to the collector of multivibrator transistor Q_1, which assures that clamp transistor Q_3 will be held inoperative, and that the operation of the multivibrator—once begun—will continue until a full cycle is completed.

The ratio of the "on time" to the "off time" of the dot multivibrator is controlled by the setting of potentiometer R_5. This ratio is usually referred to as "weight." Thus, R_5 is called the "weight control." In most cases, an operator will want this weight control in the central, or neutral, position, but occasionally it may be desirable to change the ratio of the dot time to space (i.e., under conditions of slow sending, when the change can be readily made).

The rate at which the dots are produced is controlled by the voltage applied to the combination of C_1 and R_4 and combination of C_3 and R_6. The more negative this voltage on the movable arm of R_5, the faster the timing capacitors will charge to the conducting potential of the multivibrator transistor not conducting at that instant. In other words, the greater the multivibrator-repetition rate, the more rapid the keying speed. In the author's model, the maximum charging potential was set at 60 volts. This corresponds to a keying speed of about 40 words per minute.

A keyer of higher speed can be obtained by reducing the value of R_{41} to produce a corresponding rise in the speed-control voltage and keying speed.

The value of R_{41} should not be reduced below 1,000 ohms because the voltage across filter capacitor C_{13} would then exceed its working-voltage rating. If desired, the minimum speed of the keyer (approximately five words per minute) can be decreased by increasing the values of the timing capacitors, C_1 and C_3. To insure good stability, it is important that these timing capacitors be of the paper or plastic type. Electrolytic capacitors are not stable enough for this application.

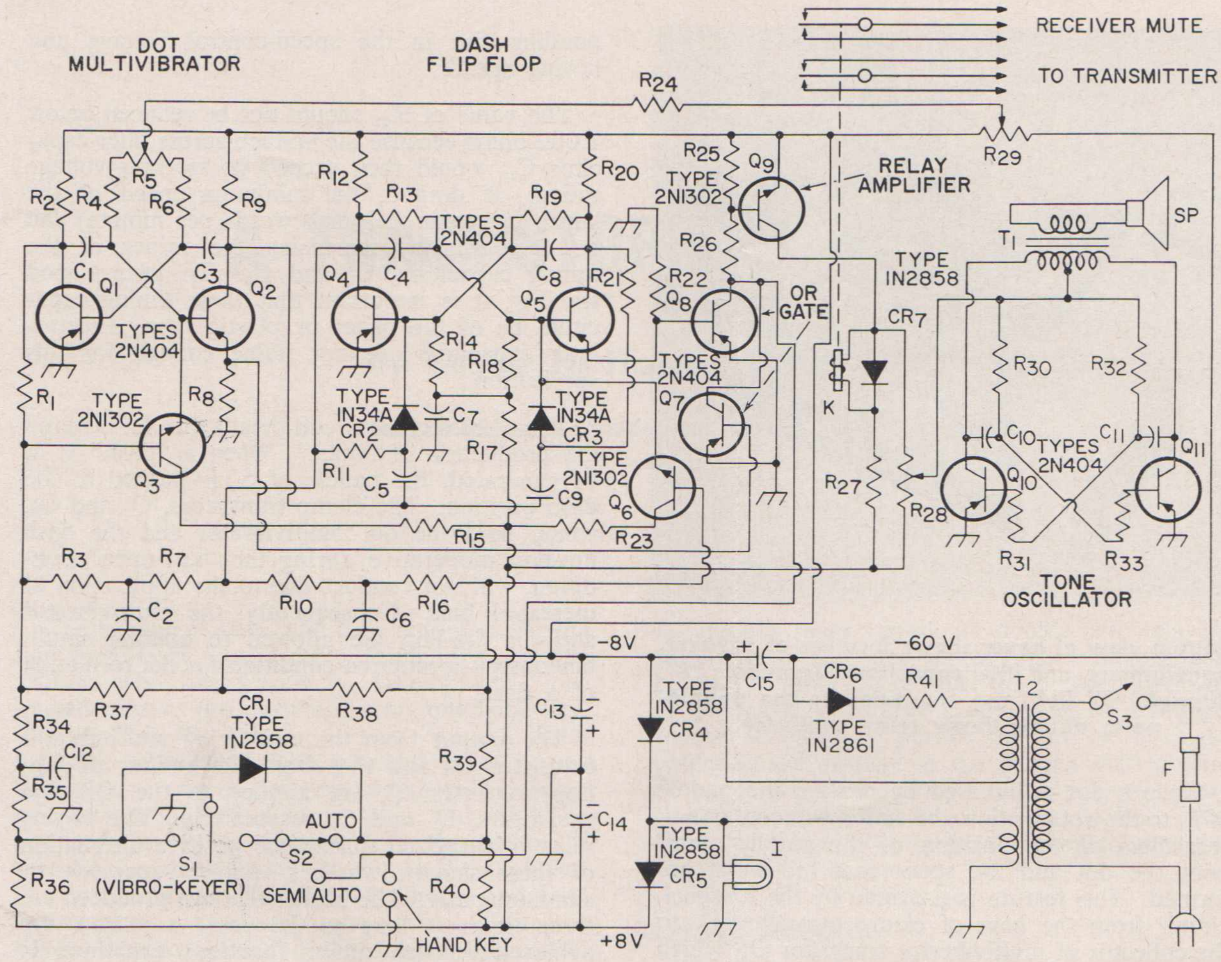
The discussion of circuit details thus far covered the production of "dots." When a "dash" is to be generated, the paddle of S_1 is pushed to the dash position. The clamp transistors, Q_3 and Q_6, which hold the dot multivibrator and the dash flip-flop inoperative during the "key-open" condition, will not conduct due to the application of increased bias. Consequently, the multivibrator and the flip-flop are allowed to operate simultaneously—a required condition for the formation of a dash.

The output from the emitter of multivibrator transistor Q_2 and that from the emitter of flip-flop transistor Q_5 are applied to the OR-gate transistors, Q_7 and Q_8 respectively. The keying relay is energized during the positive alternation of these signals, whether applied separately or simultaneously. The dashes that are produced are three times as long as the dots—a relationship achieved through timing functions explained in the paragraphs that immediately follow.

Probably the best way to explain the keying function is through a graphical presentation such as Figure 2. Assume, for the purpose of this discussion, that there is no voltage drop across the transistors of the multivibrator and of the flip-flop when they are conducting. Assume also that the switching time for these transistors is zero (i.e., the transistors can be switched from "off" to "on" or from "on" to "off" instantaneously). As mentioned previously, the keying relay will be energized whenever the positive alternation of the signal from either multivibrator transistor Q_2 or flip-flop transistor Q_5—or both—is applied to the OR gate.

When a dot is being produced, only the dot multivibrator supplies the keying signal to the OR gate. For this condition, OR-gate transistor, Q_7, controls the operation of the relay circuit. The relationship between the current through this transistor and that through the relay are shown by the dot-formation wave-forms in Figure 2.

When the paddle of S_1 is positioned to connect the dash contact to ground, the dot contact is also connected to ground through steering diode CR_1—resulting in simultaneous operation of the dot multivibrator and the dash flip-flop. Signals

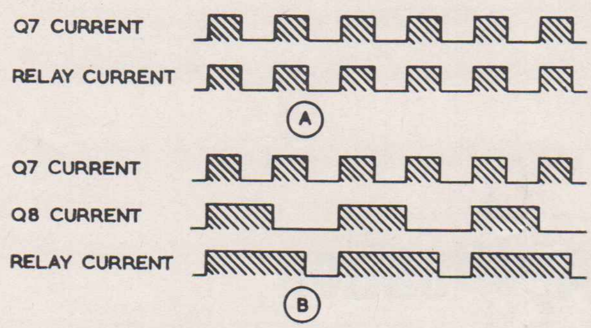


C_1, C_3—1.0 μf paper (or plastic), 200 volts
C_2—0.47 μf, ceramic, 25 volts
C_4, C_8—560 pf, ceramic, 600 volts
C_5, C_9—330 pf, ceramic, 600 volts.
C_6, C_7—0.01 μf, ceramic, 50 volts
C_{10}, C_{11}—0.02 μf, ceramic, 50 volts
C_{12}—0.1 μf, ceramic, 50 volts
C_{13}, C_{14}—2,000 μf, electrolytic, 15 volts
C_{15}—16 μf, electrolytic, 150 volts
F—Fuse, 1 ampere
I—Indicator lamp No. 47
K—DC relay; coil resistance—2,500 ohms; operating current—4 ma.
R_1—39,000 ohms, 0.5 watt
R_2, R_9, R_{12}, R_{20}—3,900 ohms, 0.5 watt
R_3, R_{16}—18,000 ohms, 0.5 watt
R_4, R_6—51,000 ohms, 0.5 watt
R_5, R_{29}—Potentiometer, 10,000 ohms
R_7, R_{10}—22,000 ohms, 0.5 watt
R_8, R_{22}, R_{25}—68 ohms, 0.5 watt
R_{11}, R_{21}—15,000 ohms, 0.5 watt
R_{13}, R_{19}—33,000 ohms, 0.5 watt
R_{14}, R_{18}, R_{30}, R_{32}—27,000 ohms, 0.5 watt
R_{15}, R_{23}—270 ohms, 0.5 watt
R_{17}—68,000 ohms, 0.5 watt
R_{24}—100,000 ohms, 0.5 watt
R_{26}—560 ohms, 0.5 watt
R_{27}—1,200 ohms, 0.5 watt
R_{28}—Volume-control potentiometer, 50,000 ohms.
R_{31}, R_{33}—10,000 ohms, 0.5 watt
R_{34}—6,800 ohms, 0.5 watt
R_{35}—8,200 ohms, 0.5 watt
R_{36}, R_{39}, R_{40}—15,000 ohms, 0.5 watt
R_{37}, R_{38}—47,000 ohms, 0.5 watt
R_{41}—10,000 ohms, 1 watt
S_1—Vibroplex keyer or equiv.
S_2—Toggle switch; double-pole, double-throw
S_3—Toggle switch; single-pole, single-throw
SP—Replacement speaker; 3½-inch, 3.2-ohm voice coil.
T_1—Push-pull output transformer (14,000 ohms to V.C.).
T_2—Power transformer.
Secondary One—125 v at 15 ma or more;
Secondary Two—6.3 v at 0.6 ampere or more

Schematic diagram and parts list of W2YM's transistorized keyer.

will now be applied to both OR-gate transistors, and the relay will be energized for an interval three times as long as that used to produce a dot. The dash-formation waveforms in Figure 2 illustrate this relationship.

The voltage drop across the keying relay and resistor and R_{27} is the DC supply voltage for transistors Q_{10} and Q_{11} in the tone oscillator. For current to flow through these components, the relay-amplifier transistor, Q_9, must receive a keying signal from the OR gate. The tone oscillator, therefore, operates only when dots or dashes are being produced. Its output is then applied to the speaker to provide an audible indication of the



Graphical representation of keying function showing waveforms for Q7, Q8, and relay currents. (A) Dot formation, Paddle connecting dot contact of Vibro-Keyer to ground, (B) Dash formation, Paddle connecting dash contact of Vibro-Keyer to ground.

keying. Potentiometer R_{28} controls the volume of this output.

The transistorized keyer may also be operated as a semi-automatic or manual key. In the semi-automatic mode, switch S_2 (see Figure 1) is placed in the SEMI-AUTO position. Although the dots are still produced automatically, the automatic-keying circuits are by-passed when the paddle of S_1 is moved to the dash position, and the dashes must then be produced manually. If desired, a hand key may be connected to the unit (across the terminals marked HAND KEY in Figure 1) so that the automatic-keying circuits can be by-passed when producing both dots and dashes.

Construction

The complete keyer is housed in a miniature aluminium case which is only 4 inches high, 5 inches wide, and 6 inches deep. Internal mounting details for the unit are shown in the photographs

The major circuitry is mounted on two phenolic boards. One board contains the multivibrator and its clamp transistor, the flip-flop and its clamp transistor, and the OR gate. The other board contains the tone oscillator and voltage bridge for the dot and dash clamp circuits. [Author's Note: The voltage bridge and clamp circuits are similar to those used by James C. MacFarlane, W30PO, in the December, 1962, issue of "QST" magazine.]

The power supply, relay, speaker, output transformer, switches, and potentiometers are all mounted directly to the case. The cone of the speaker is protected from damage by covering it with a small piece of perforated aluminium. In the model constructed by the author, binding posts were used to connect the Vibro-Keyer, S_1, to the automatic keyer, but any method of connection that suits the fancy of the builder would be equally satisfactory.

All relay contacts are brought out to the rear of the keyer and the connections to them are made through a six-terminal Jones strip. This arrangement permits two circuits to be keyed simultaneously. It also allows the relay to provide either normally closed action or normally open action, whichever is preferred. In the unit described here, the second set of contacts are used to mute the receiver during the key-down condition. This feature required that the relay be normally closed. [Author's Note: Some relays do not have non-metallic strikers on either the pole pieces or the armature and, consequently, may be sluggish. This condition can be corrected by drilling and tapping the armature and installing a No. 2-56 brass screw. The screw should be adjusted so that about 2 to 6 mils protrudes from the armature. A lock nut should be used to prevent any shift in the position of the screw.]

The transistorized keyer constructed by the author has been in constant use for more than a year. Once an operator becomes familiar with the instrument, he will find it extremely easy to send perfect copy.

(With acknowledgements to R.C.A.)

6: General Application Considerations

Many criteria must be considered before choosing a photosensitive device for a particular application. Some application requirements can be filled by only one type of device; however, for many applications, any one of several possible devices may be suitable. This section provides a general guide to the selection of photosensitive devices for most typical applications.

Level of Light or Radiation

All the devices described in this manual can be used over wide ranges of light or radiational level. Of course optical, environmental, or circuit adjustment may be necessary to accommodate the particular device. Fig. 88 shows the approximate useful range of each class of photosensitive device. Multiplier phototubes are indicated for the lowest levels of radiation because of their inherent advantage of secondary-emission gain. Higher levels of radiation should be avoided in the case of multiplier phototubes because of fatigue, saturation, and life problems. However, the range can be increased upward for a multiplier phototube by decreasing the applied voltage.

Because the gain of a gas-filled phototube is much less than that of a multiplier phototube, it cannot be used at as low a level of light; however, such tubes can operate satisfactorily at higher ranges. Vacuum phototubes can be used at higher levels of radiation than either gas-filled phototubes or multiplier phototubes. Cadmium sulfide photocells are usually most useful in the higher ranges of light intensity. At the lower ranges, they are somewhat limited by response-time problems. Solar cells (photovoltaic cells, energy-conversion typed) are of use only at the upper range of radiation levels. At low levels, the power developed is too small to be of value for most applications.

The far-infrared detectors are generally useful only for the very low levels of radiation. If a large amount of infrared radiation is available, the complication of special cooling can be avoided by other simpler devices such as the bolometer or thermocouple.

Spectral Energy Distribution

Although many of the different detectors of radiation have a broad spectral sensitivity range, there are times when the spectral energy distribution of the source is of special importance in selecting the proper device. For the far-infrared region, only the solid-state photoconductors are

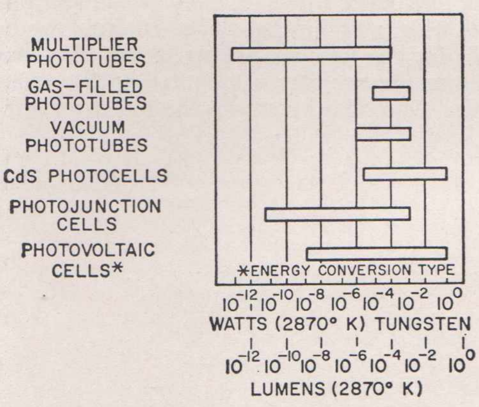


Fig. 88—The range of power or luminous flux for which the various detector types are useful. The indicated ranges are only for guidance, and may at times be exceeded. The overlapping scales of luminous and power fluxes are scaled in reference to a tungsten lamp operating at 2870 degrees Kelvin color temperature. For a different reference source, the luminous equivalent of the power flux would be different.

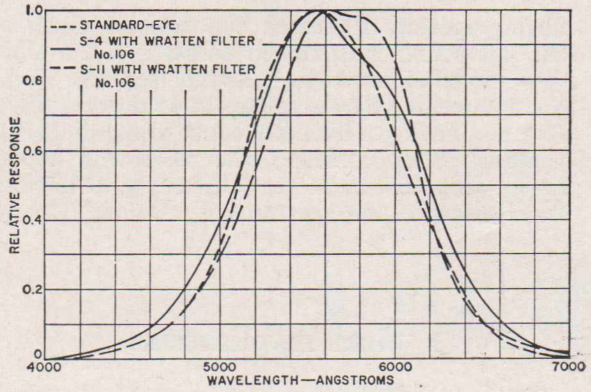


Fig. 89—Spectral responses S-4 and S-11 modified by the Wratten filter 106 closely match spectral sensitivity characteristic of the eye.

suitable. Similarly, in the ultraviolet region, only photoemissive devices are practical; however, in the visible range, many devices can be used. The spectral response characteristics of the various types of devices are included in the pertinent sections of data books.

Some applications require a device having a specific spectral sensitivity to match a particular response. However, for many applications, no such detector exists. For example, measurement of light levels requires that the receiver has the same spectral sensitivity as the eye. Although the devices described in this manual have spectral response covering the range of visible radiation, none of the responses exactly matches that of the eye. In this case, the spectral response may be modified by means of special color filters to provide a combination detector and filter which does match the eye response. The filter characteristic required may be obtained by dividing the spectral sensitivity characteristic of the eye by the spectral response of the device to be corrected. Fig. 89 shows the close approximations which can be obtained by using commercial color filters.

Fig. 90 shows the transmission characteristics of a number of other useful filters and radiation transmitting substances. Color filters are frequently used to isolate various regions of the spectrum to improve detection or matching characteristics of photodetectors. The transmission of various glasses and the atmosphere are also provided because these factors have an important influence on practical optical systems.

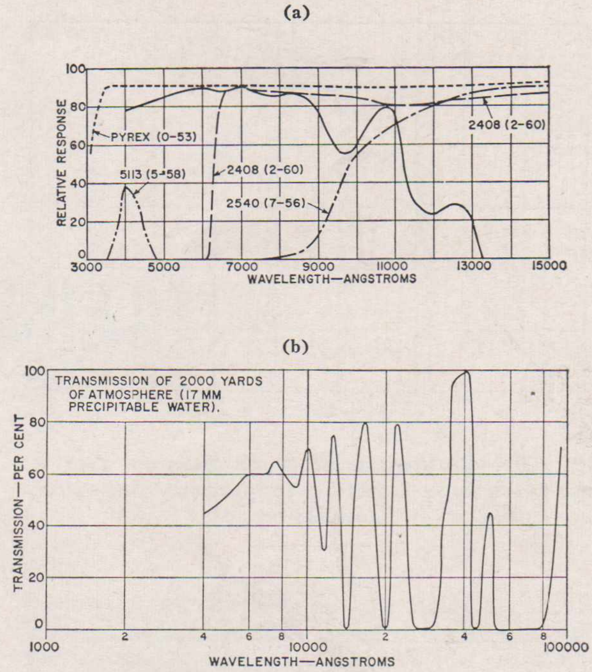


Fig. 90—Transmission characteristics of various useful color filters, glasses, and the atmosphere. ((A) 2540, 5113, 2408, Pyrex (B) 2000 yds. of air.)

Frequency Response

The approximate useful range of frequency response for the various types of photosensitive devices is shown in Fig. 91. All the detectors are useful from dc operation to a specific cutoff frequency. However, for photocells of the germanium type, in which the dark current is quite substantial, the detection of unmodulated signals is not practical. A modulated signal can be automatically discriminated from the dark-current component. A comparison of ranges in Fig. 91 shows that the multiplier phototube is especially useful in detecting very-high-frequency modulations.

In some cases the light to be detected is modulated by the nature of the application, or is in the form of light pulses. For cases in which the light is unmodulated, it is often desirable to provide discrimination between signal current and dark current by one of several means, such as a light-chopper wheel.

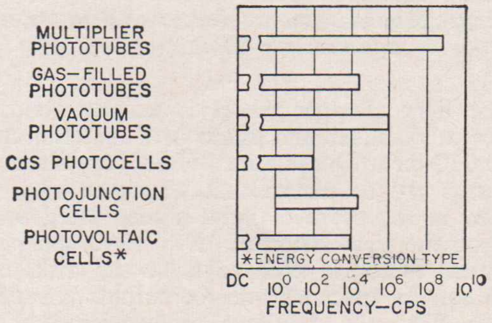


Fig. 91—Range of frequency response of various photodetectors. Recommendation of operation for dc (unmodulated light) sources is indicated by the symbol dc at the left end of the frequency range.

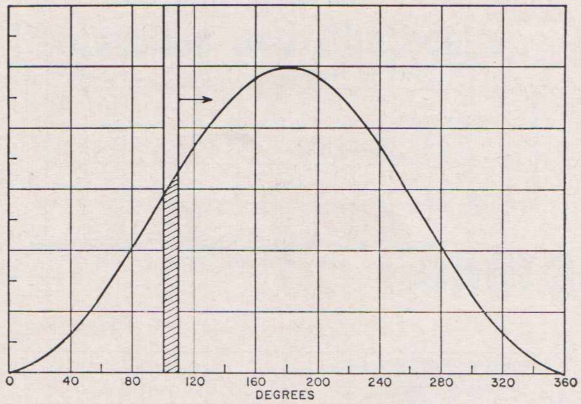


Fig. 92a—Illustration of a slit passing over a sine-wave type aperture to produce sine-wave modulation.

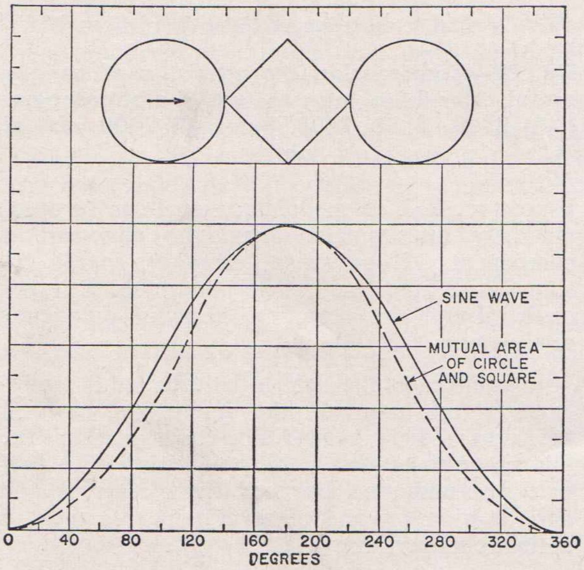


Fig. 92b—Production of sine-wave approximation by use of circular holes and a square aperture of 45 degrees.

The light chopper wheel is usually about 8 inches in diameter and made of a material such as 0.030-inch Duralumin. Holes or alternate segments are so placed that when the wheel is rotated in the beam of light, a modulated beam falls on the light detector. If a 1800 rpm synchronous motor is used to drive the disk, and three equally spaced 60-degree cutouts are made in the disk, the resultant modulation (as the cutouts pass over a radial slit) approximates a 90-cycle square wave, which is a convenient frequency because it is not a multiple or submultiple of the 60-cycle current. Almost any type of modulation can be accomplished with properly shaped apertures. For example, if the moving aperture is shaped like the area under a sine curve and it is moved across a slit, a sine wave modulation can be achieved (see Fig. 92a). A circle passing over a square at 45 degrees (Fig. 92b) and having a diameter equal to the diagonal produces an approximate sine wave with about 10-per-cent first-harmonic distortion (Uniform illumination of the apertures is assumed.)

Typical Applications

As discussed in the previous sections, photosensitive devices are extremely versatile devices which are being used in an ever-increasing variety of applications. Shown below is a brief listing of several typical applications for each category of photosensitive devices. Choice of a particular type for a specific application should be based on their ratings and characteristics.

Vacuum Phototubes

Photometry
Spectrophotometry
Industrial controls
Facsimile
Colorimetry

Gas Phototubes

Industrial controls
Sound reproduction

Multiplier Phototubes

Scintillation counting
Photometry
Spectrophotometry
Flying-spot generator
Star tracking
Cerenkov radiation measurement
Laser detection
Industrial controls
Colorimetry
Timing measurement

Photoconductive Cells

Industrial controls
Camera iris control
Street light control

Photojunction Cells

Sound reproduction
Data processing

Photovoltaic Cells

Solar power conversions
Industrial controls
Photometry

A VLF OSCILLATOR

This note describes a very low frequency oscillator which uses transistors as the active elements, and which is suitable for vibrato applications in electronic musical instruments. But before going on to describe the oscillator, it will be necessary to clear up a popular misconception surrounding the terms "vibrato" and "tremolo."

Vibrato consists of a rhythmic variation of the frequency of a musical tone. This is generally achieved by the performer himself, because it would be a very difficult thing to do electronically. For example, in one of the many types of stringed instruments, where the note is determined by manually stopping the vibrating string at some point along its length, vibrato can be introduced by the performer merely by vibrating the finger that is holding the string down.

Tremolo, on the other hand, consists of a variation in the amplitude of the note. This is a much easier thing to do electronically, because all that is needed is a low frequency oscillator which is used to amplitude-modulate the signal waveform. This is the same process as that used to impress audio signals onto an rf carrier.

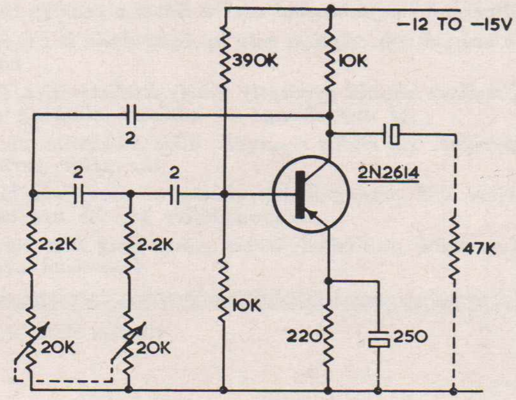
There is a demand in electronic instruments for a "vibrato" facility, but what is in fact provided is a tremolo facility. We all know what the idiosyncrasies of the human ear are, and we can use these peculiarities to deceive the ear. What is done is to use tremolo; the ear then detects these changes in amplitude, but due to its own characteristics, finds it easy to translate the changes in level into apparent changes in pitch. See the Fletcher-Munson hearing contours reproduced in the September, 1964, issue of this magazine.

VLF Oscillator

The tremolo facility in an electronic instrument consists of some means of placing a very-low-frequency modulation on the musical signal. Various ways have been found of doing this, and particularly in supplying the very low frequency required, generally in the range 2 to 10 cps. The very low frequency waveform should be reasonably free of distortion to avoid possible harshness, and to assist in producing a modulated music waveform which is reasonably symmetrical. This latter point also has a bearing on the modulation system used, although that is outside the scope of this note.

Care must be taken in the amplifier system to prevent the very low frequencies themselves getting into the amplifier, which results in a phenomenon usually described by those interested as "pumping." This is most objectionable, and can also lead to damage to the equipment. In this regard the production of a symmetrical modulated waveform, that is, one in which the centre point has not been shifted, is most important. If this is done, then the normal high-pass filters inherent in the amplifier will take care of the very low frequency.

The object of this note is to describe a very simple transistor oscillator which may be used as the very low frequency source, and to indicate



some of the ways in which it could be adapted to suit various requirements.

Circuit

The circuit, shown in the accompanying diagram, is very simple. The oscillator is an example of the current-derived phase-shift type, using an AWV 2N2614 transistor. In fact, almost any small-signal audio transistor could be used, provided its current gain was high enough. As in all phase-shift oscillators, the operation depends on the fact that a phase shift of 180° takes place in the RC network between collector and base, at the operating frequency. The required phase shift occurs only at one frequency because at other frequencies the capacitive reactance of the network increases or decreases with respect to that needed for the 180° phase shift.

The operating frequency depends, therefore, on the constants of the network. If the frequency of the oscillator has to be varied, then one or more of the elements of the RC network may be varied. As a rule, the resistive elements are varied, because this is the easiest approach. In this case, two of the resistive elements are varied simultaneously to get the frequency coverage required, which is 2 to 10 cps.

A collector supply at 12 to 15 volts will be satisfactory. The current consumption is of the order of 0.5 ma at —12 volts supply and 0.75 ma at —15 volts supply. The output level at the collector of the transistor is about 2.7 volts rms at 10 cps and about 2.2 volts rms at 2 cps, when using the —12 volt supply. Operation with a supply lower than —12 volts is not recommended.

When a —15 volt collector supply is used, the output levels are higher. The levels then become about 4.25 volts rms at 10 cps and about 3.8 volts rms at 2 cps. The output load impedance placed on the collector of the transistor needs to be kept high to avoid too-high a reduction in signal level, leading eventually to the oscillator stopping. A load of 47,000 ohms is just beginning to show a reduction in the level at the collector.

Because the available output level from the oscillator is far in excess of what will usually be required, the comparatively high load value will in general present no difficulty. One simple solution would be the use of a potentiometer with a fixed resistor in series, as the load on the oscillator. The values would be so chosen that the potentiometer can provide the range of output voltages required, whilst the potentiometer, external load and series resistor present a high load to the oscillator at all times.

Construction

There is really nothing to say about the construction. The layout is obviously unimportant, and the only point on which difficulty may be experienced is in leakage in the capacitors forming the phase-shifting network. In the model, paper capacitors of the 200 VW type were used in preference to electrolytic capacitors for that reason. It was found that leakage of the order of 250,000 ohms across one of the capacitors could stop oscillation. As for the rest, it is a matter of how the oscillator can best be fitted into the overall scheme that the builder has in mind.

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

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