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Continued

SINGLE-SIDEBAND

By B. J. Simpson

In the last two instalments of this article, we discussed just what is meant by SSB, the advantages of this system of transmission, and then went on to take a look at methods of carrier suppression. The next step that must be taken is the suppression of one of the sidebands. In the measures described last month to suppress the carrier, we were left with an output consisting of two sidebands but no carrier. This is double sideband (DSB), and it is now necessary to remove or suppress one of these sidebands.

As previously explained, both of the two sidebands carry the same intelligence, and the transmission of only one of them is required, so that the full advantage of the system may be realised. There are two basic methods of removing the unwanted sideband, one depending on frequency discrimination and the other on phase discrimination. The former uses a filtering technique and the latter a system of phase shift.

Filters

Filters come in many forms, so much so that many of them are not readily recognised as filters by those less well versed in the art. They all have one thing in common, that they offer a higher impedance to some frequencies than to others. High impedance (and attenuation) is offered to the frequency or frequencies that are not wanted, whilst low attenuation is offered to those that are wanted.

For general purposes, filters may be divided into three basic forms, the low-pass filter, the high-pass filter and the band-pass filter. The operation of these configurations is largely explained by the names, but it may be well to emphasise a couple of points. In the case of the low-pass filter, the filter is designed to pass all frequencies below the designed cutoff frequency, with minimum attenuation, and to offer high attenuation to all frequencies above the design figure. The high-pass filter is the converse of this arrangement, whilst the band-pass filter is intended to pass only frequencies within a specified band. The station selecting arrangements of your radio or TV set are in fact band-pass filters.*

In the design of practical filters, it is not possible to offer infinite attenuation to unwanted frequencies, any more than it is to offer zero attenuation to the wanted frequencies. The wanted frequencies will always suffer some attenuation in the filter, and this is called the insertion loss. A little reflection will also show that a sudden sharp cutoff cannot be achieved, although something approaching it can be obtained. The cutoff frequency then is commonly regarded as that at which the response is down by 3 db. It may however vary with different manufacturers.

Another important characteristic of filters is the sharpness with which the attenuation of the unwanted frequencies increases beyond the cutoff frequency. This is a property which, within the limits already implied, can be controlled by the designer according to the application. The rate of increase of attenuation with change of frequency outside the cutoff frequency is called skirt selectivity.

* The filters described are series-type units. Shunt-type units can also be used, their operation being the converse of that already described.

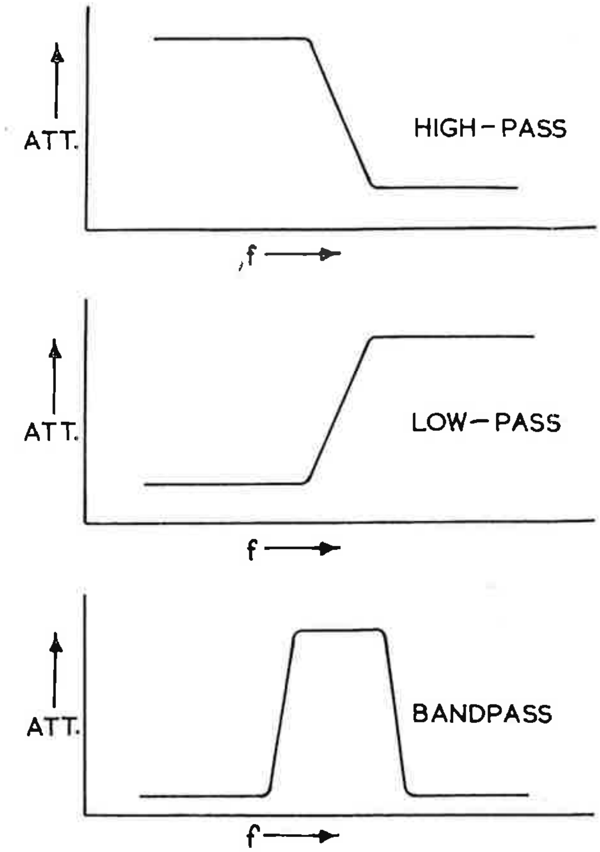

Fig. 17

When dealing with SSB, we are particularly concerned with band-pass filters, and with their bandwidth. The bandwidth, as the name implies, is the difference between the highest and the lowest frequencies that the filter will pass at the specified attenuation, often at 3 db down. This figure may vary between different makers, and the attenuation should therefore always be quoted in relation to the pass-band or bandwidth.

Typical characteristic curves for the three types of filter mentioned are shown in Fig. 17. There are other filters besides these, used for specialised applications, but these special types are in general combinations of, or variations of, these three basic types.

Most of the filters we encounter in the electronics industry are made up of combinations of lumped inductances and capacitances. In fact, there are cases where capacitors and resistors only are used, as for example in audio cross-over networks of the simpler type. But cases like this, where only a slow rate of attenuation beyond cutoff is required, are rather rare, and are the exception to the rule.

In recent years, two other types of filter have appeared. These are the crystal filter, in which combinations of crystals form the frequency selecting elements, and the mechanical filter, which depends for its operation on a series of plates which are so dimensioned as to be resonant at specific frequencies.

Crystal Filters

For many years now, crystal filters have been in use in communication receivers to provide very fine selectivity in the intermediate frequency section of the receiver, particularly when ICW transmissions were being received. In such cases, it has been possible to achieve a pass-band of only a few hundred cycles or less, using only one crystal.

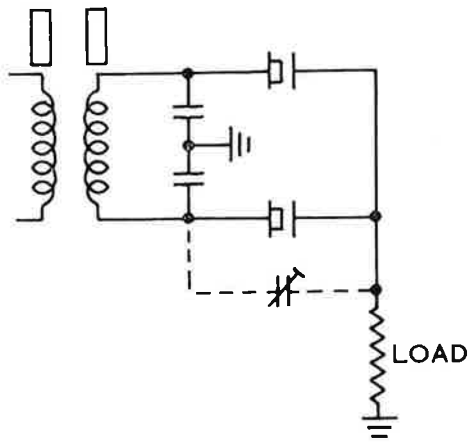

Fig. 18

For SSB use, the simplest type of crystal filter is the half-lattice shown in Fig. 18. Those familiar with communication receivers will recognise it as similar to the narrow-band filter already mentioned, but with the phasing capacitor replaced by a second crystal. In this arrangement, the shunt capacitances of the two crystals cancel each other, and the bandwidth of the filter depends on the resonant frequencies of the two crystals. This simple arrangement normally has rather poor skirt selectivity, especially where a wider bandwidth is required.

The skirt selectivity of the simple half-lattice crystal filter can be improved by the addition of the small preset capacitor shown connected into Fig. 18 by dotted lines. Correct adjustment of this capacitor will sharpen the skirt selectivity, but there comes a point at which side responses appear; this limits the improvement obtainable. There is a further solution to the problem, in the use of multiple sections of filter, one after the other, as shown in Fig. 19.

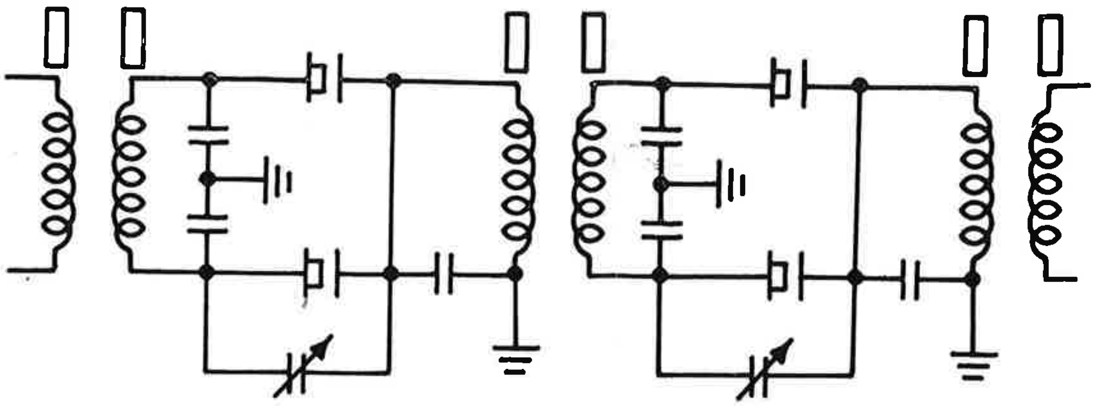

Fig. 19

The full lattice type of crystal filter is shown in Fig. 20, where the diagram shows two consecutive sections. The bandwidth of this system is determined largely as before, by the difference in the frequencies for which the crystals are ground. The crystals in each section are in two pairs, denoted A and B in the diagram; the two crystals in each pair should have the same resonant frequency within a few cycles, or the characteristics of the filter will suffer. The separation between A and B frequencies then fixes the bandwidth.

Mechanical Filters

Mechanical filters are a comparatively recent arrival on the scene. They depend for their operation on the fact that a metal disc, or any other body for that matter, will have a natural resonant frequency. By a careful adjustment of the diameter and thickness of a disc, we can select the resonant frequency; we then have in this disc a resonant circuit similar to the tuned radio circuit with which we are familiar. The disc will resonate mechanically when excited mechanically at the resonant frequency, just as our circuit will be excited electrically when a signal of the right frequency is fed into it.

The practical mechanical filter consists of a number of such discs, depending on the characteristics required of the filter. The discs will be coupled together by coupling rods, as shown in Fig. 21. Basically then, the filter centre frequency is determined by the individual discs, the skirt selectivity by the number of discs used, and the bandwidth by the total area of the coupling rods.

Coupling of the electrical signal into and out of the mechanical filter is done by means of electromechanical transducers. Each of these transducers consists of a coil wound around a core of magnetostrictive material (e.g., nickel), with the end of the core in contact with the end disc of the filter. Each transducer has a polarising magnet.

An input signal to one of the transducers produces contraction and expansion of the core material, which in turn exerts a pull or a push to the stack of discs. The electrical signal is thus translated into mechanical vibration applied to the end of the filter. If this vibration is at or very close to the natural frequency of the discs, they will be resonated, and will transfer the vibration from one to the other through the stack, and thence to the second transducer at the far end.

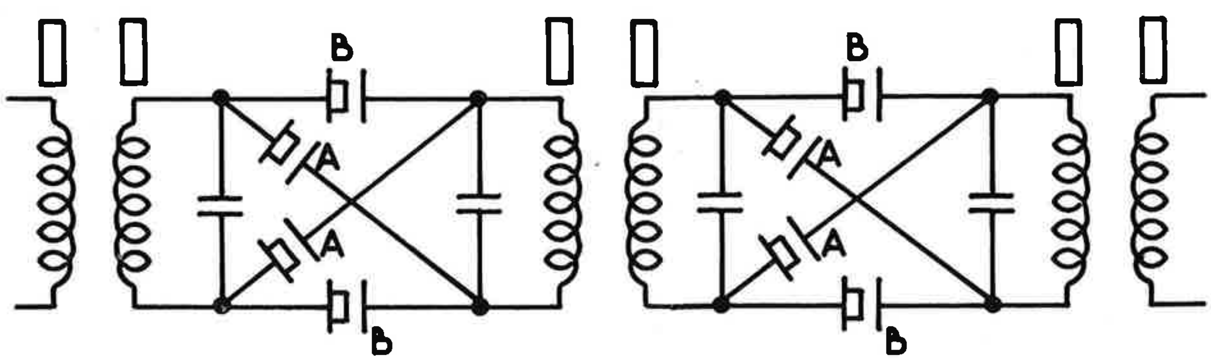

Fig. 20

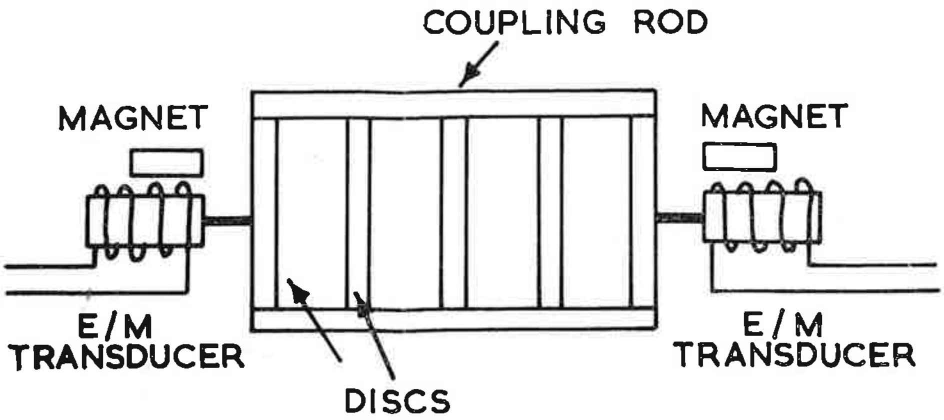

Fig. 21

Here the mechanical impulses are re-translated back to electrical signals and thence to the following equipment.

For those who are not familiar with the term "magneto-striction," it refers to the property of some partially-magnetic materials, of which nickel is perhaps the best known, by which mechanical contraction and expansion of the material takes place when subjected to a magnetic field. This property is widely used in transducers for all sorts of purposes. A typical case is in echo-sounding transducers for ships, where electrical impulses are converted to sound waves in the water, and vice versa.

Sideband Suppression

We have so far said quite a lot about filters, but little about sideband suppression, but filters form the core of this part of the problem, and an understanding of them is essential to an understanding of the complete system. As most readers will already have realised, the unwanted sideband is removed quite simply by filtering it out with a suitable filter.

The characteristics of the filter used, and in particular its bandwidth, will depend on the width of the sideband that we wish to transmit. It will be obvious that the wider the sidebands generated in the balanced modulator, the more difficult it may be to provide a suitable filter. Aside from the consideration of bandwidth as such, there is the related matter of what bandwidth is necessary to transmit the intelligence. Because SSB transmission is used mainly for communication work, a maximum sideband of about 3.5 Kc would probably satisfy the requirement.

In fact, it is common practice to restrict the speech frequency range used in commercial transmissions to something of the order of 300 to 3,000 cps. This restriction, whilst affecting intelligibility to a negligible extent, considerably reduces the amount of power required, and greatly assists in the suppression of the unwanted sideband when SSB transmission is used.

Typical of the use of a band-pass filter to remove an unwanted sideband is the system shown in Fig. 22. Here we have a filter 3 Kc wide at the —3 db points, and with sharp skirt selectivity. The filter characteristic is now disposed on either side of the carrier frequency, depending on which sideband is to be removed, so that the carrier appears at the —20 db point. The two alternative carrier positions in the diagram are designated C1 and C2.

Assuming that we wish to suppress the upper sideband, then the system would be arranged with the carrier on the C2 position. It will then be seen that the lower sideband will pass through the filter with only the insertion loss of the filter, but the upper sideband will be considerably attenu-

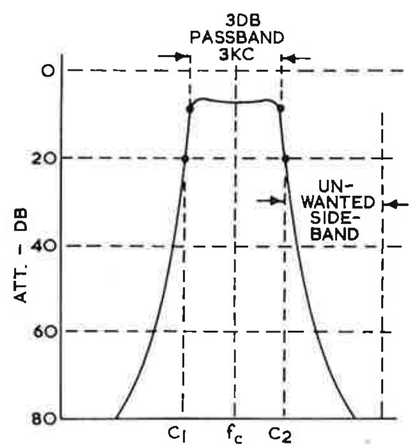

Fig. 22

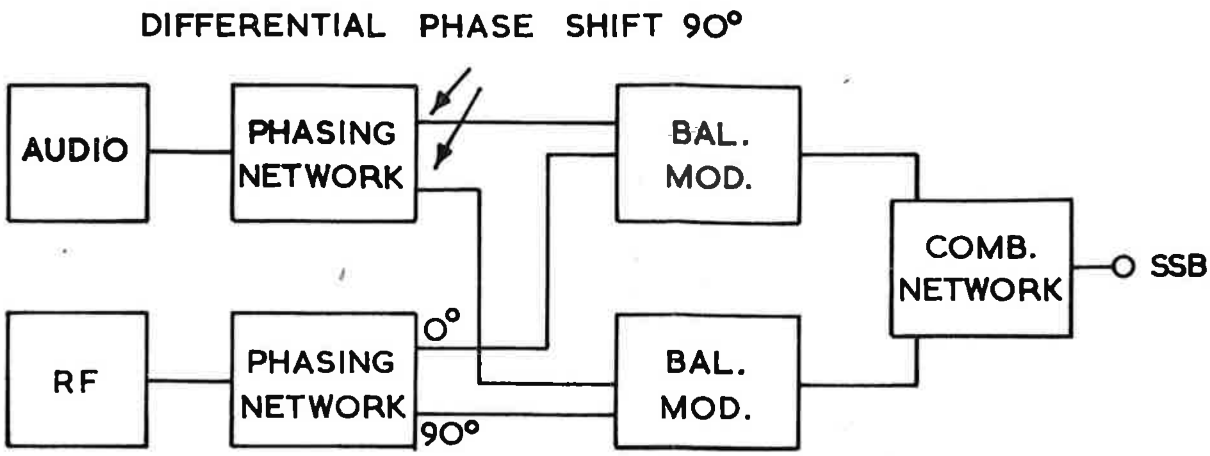

Fig. 23

ated. The reason for placing the carrier at the —20 db point is partly to attenuate the very low frequencies (below about 300 cps) in the wanted sideband, and partly to increase the suppression of the carrier frequency which was started in the balanced modulator previously discussed.

Frequency Changing

It has for some years been common practice to carry out the processes of sideband generation and suppression at comparatively low frequencies, two typical frequencies being 50 Kc and 465 Kc. Where the latter frequency is used, standard intermediate frequency transformers can be used for many of the stages. Another reason for some standardisation of frequencies is the availability of suitable filters. It is obviously more economic to be able to use a standard range of such components than to have different units made for every application.

All this means that once the sidebands have been produced, and the carrier and unwanted sideband removed, the sideband that is left is fed to a frequency changer where it is mixed with another frequency to produce sum and difference frequencies in the usual way. One of these sets of frequencies, which of course carries the intelligence, is then arranged to be at the required transmission frequency, whilst the other is suppressed in the tuned circuits at the output of the frequency changer. This possibility was mentioned earlier.

By way of example, let us assume that we have derived a single sideband centred on 50 Kc, with a bandwidth of 3 Kc. We will also assume that this is the lower sideband. The requirement now is to transmit this information on a carrier frequency of 3.6 Mc. If the sideband is mixed with a 3.55 Mc carrier, then the sum frequency sideband will lie between 3.6 and 3.597 Mc. At the same time, the difference frequency sideband will lie between 3.5 and 3.503 Mc. Note that by a suitable choice of the mixing frequency, a sideband that starts off as an upper sideband at the initial frequency (say 50 Kc) can end up as a lower sideband with respect to the final transmitted (carrier) frequency, and vice versa.

Phase Shift Selection

The phasing system of single-sideband production was developed by R. Hartley in the 1920's, and is widely used at the present time. The system has a number of advantages over the previously-described system using filters, the most obvious one being that the filters are not required. These can be quite costly components. A further advantage is that the required sideband can be generated at the operating frequency, so that the frequency-changing techniques mentioned in the foregoing section are not required.

The phase-shift system is not, however, without its disadvantages. Phase shift and amplitudes in the circuit must be held to very close tolerances, and the system requires two balanced modulators instead of only one. The phase-shift system can therefore be as complex, if not more so, than the filter method.

The phasing system of sideband selection makes use of the fact that if two balanced modulators are fed with a carrier of the same frequency, but 90° apart in phase, and with the same audio signal, but also 90° apart in phase, the combined outputs of the two modulators will contain only one sideband. The relative phasing between the two sideband outputs of the modulators result in the cancellation of one of the sidebands and the reinforcement of the other. Selection of the alternative sideband can be achieved by reversing either the rf or the audio phase relationship between the two modulators.

A block diagram of the phase-shift selection system is shown in Fig. 23. This diagram is largely self-explanatory, and shows the basic system already described. The phasing network for the rf input is a simple arrangement, operating at one frequency only. When we come to the audio side, however, where a band of frequencies has to be transmitted, the use of a simple phase shift circuit is not possible. This is because the phase shift through a simple circuit will vary with frequency. Instead, it is necessary to employ a differential phase shift network, with a differential phase shift of 90°, between the audio source and the balanced modulators. In this way a difference in phase of 90° can be maintained within the limits of 2 or 3 degrees over a wide range of audio frequencies, and so meet the requirement.

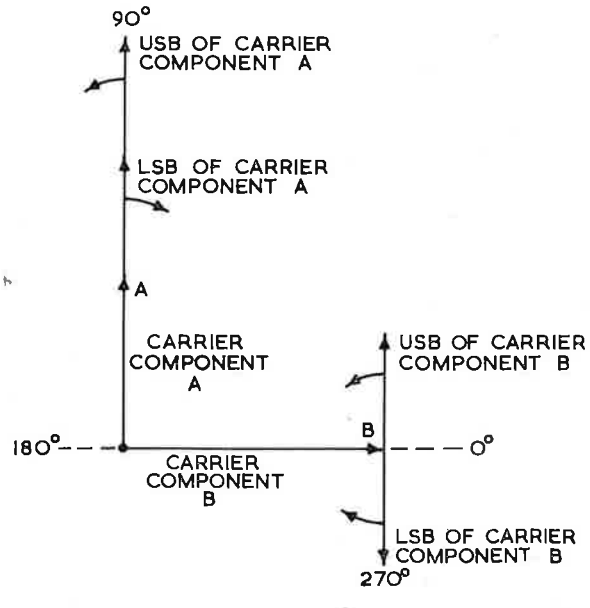

Fig. 24

A vector diagram showing the operation of the phase-shift sideband selector is shown in Fig. 24. Here the two carrier components, labelled A and B, are shown with A leading B by 90°. As in the previous diagrams, the two vectors are assumed to be rotating in an anticlockwise direction about their origin.

At the instant shown in the diagram, the two sidebands of carrier A are adding, and the two sidebands of carrier B appear to be cancelling. The lower sideband of carrier A is generated at the same frequency as that of carrier B, but because there is a phase difference of 90° in both the audio and the rf signals, the two lower sidebands appear 180° out of phase. The two lower sidebands rotate in the same direction, and at the same rate, because they are both at the same frequency. This means that the antiphase relationship will always be maintained, and their combined net amplitude, in the common tuned output circuit, will be zero.

The same consideration does not apply in the case of the upper sidebands. Here, the frequency is the same, and therefore the speed of rotation, but the phase difference is zero, and will remain so. They therefore add, and the amplitude of the combined outputs will be twice that of one of them. It will be seen therefore, that provided the correct phase relationships are maintained, the output will consist of the upper sideband only. As previously indicated, if the required output is the lower sideband, the phase of either the audio or the rf inputs to the system can be reversed to produce this result. The construction of a vector diagram similar to that shown in Fig. 24 will then demonstrate the operation of the new set of conditions, which produce the lower sideband output.

Summary

With this portion of the article we complete the description of the basic considerations involved in the production of a single-sideband signal. It is not within the scope of this article to go more fully into practical circuit applications, which are adequately covered elsewhere. In the concluding article which follows, the processes involved in the reception of the SSB signal, and the recovery of the transmitted intelligence, will be dealt with.

COLD CATHODE TRIGGER TUBES

The characteristic curves of cold cathode tubes are entirely different from those of other well-known gas discharge tubes; it is therefore difficult to summarise their properties without special explanations. Moreover, in actual fact it is only the combination of the "ignition curve" with the "transfer curve" which provides adequate information. In this description, an attempt is therefore made to explain the significance of this representation and to summarise at the same time the special characteristics of cold cathode trigger tubes.

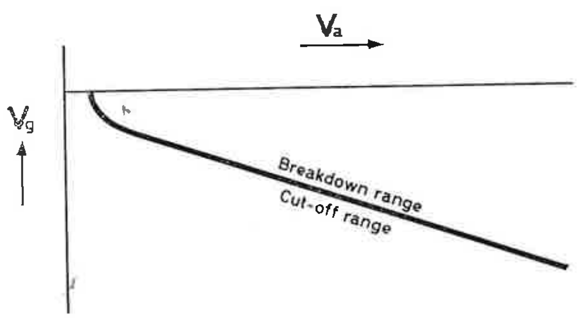

Fig. 1: Breakdown characteristic of a thyratron. This represents the boundary between cutoff and breakdown range.

The normal breakdown characteristics for thyratrons (Fig. 1) are well-known. Fig. 2 shows the corresponding characteristics for cold cathode trigger tubes, which are obtained for breakdown between the anode-cathode section (main section). If the associated positive trigger voltage $V_{st}=0$, the main section anode will only strike on reaching the anode voltage V_{az}. An increase of the trigger voltage in the positive direction increases the anode breakdown voltage. If the trigger voltage reaches the value V_{sz}, i.e., the trigger breakdown

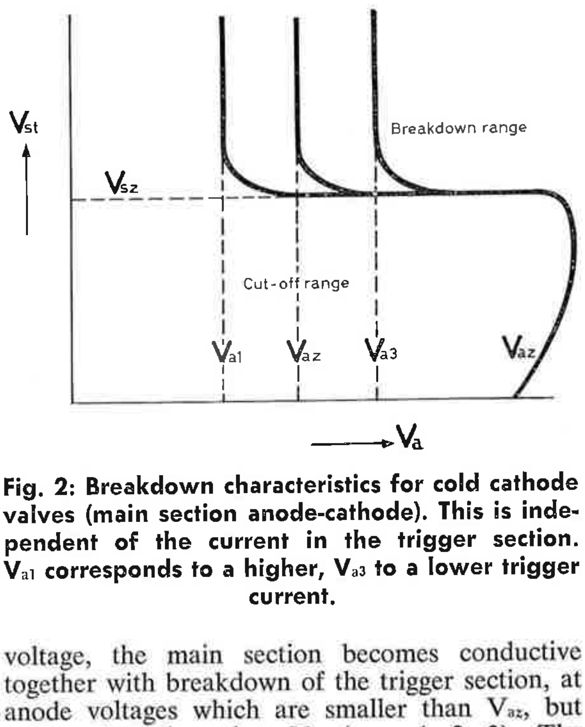

Fig. 2: Breakdown characteristics for cold cathode valves (main section anode-cathode). This is independent of the current in the trigger section. V_{a1} corresponds to a higher, V_{a3} to a lower trigger current.

voltage, the main section becomes conductive together with breakdown of the trigger section, at anode voltages which are smaller than V_{az}, but greater than the values V_{an} ($n = 1, 2, 3$). The value of this V_{an} voltage is determined by the current I_{st} in the trigger circuit; ($n = 1$ corresponds to the maximum trigger current, i.e., the smallest resistance; $n = 3$ corresponds to the smallest trigger current and therefore the highest resistance in the trigger circuit).

It is a well-known fact, that the breakdown characteristics of thyratrons can also be varied by altering the resistance in the control grid line. Of course, this only becomes noticeable with very high resistances exceeding 1 megohm. In cold Cathode tubes resistances of the order of 0.1 megohm lead to limitation of the trigger current. Since this effect is extremely important in the design of circuits, the relationship of the values of V_{an} to the associated trigger currents I_{st} are quoted as the so-called transfer characteristics (Fig. 3). They refer to the point at which the discharge in the trigger cathode space is "transferred" to the anode.

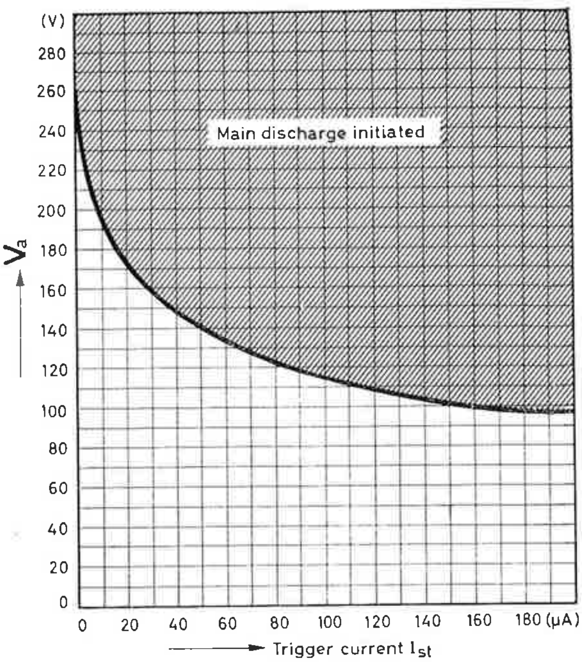

Fig. 3: Transfer characteristic. The V_a values correspond to the $V_{a\ 1,\ 2,\ 3,\ \ldots}$ values of Fig. 2 in relation to the trigger current.

The transfer values can be improved by adopting an ingenious method. The trigger time characteristics are arranged, so that breakdown always occurs first in the trigger cathode section, conduction in the main section being started when the trigger current is sufficiently high. As breakdown occurs in the trigger section, the trigger voltage falls by about 20 V. By connecting a capacitance in parallel to the trigger section as shown in Fig. 4, this capacitance will provide a corresponding capacitative discharge in the form of a current surge into the trigger when the critical trigger voltage is reached, thus temporarily reinforcing the trigger current. The effect of this is that the trigger characteristic is apparently lowered, thus providing the transfer characteristics of Fig. 5, depending on the capacitance value. The capacitance should not exceed 5000 pf, otherwise the phase shift in ac operation becomes too high on the one hand, while on the other hand the cathode may be damaged by excessive discharge current surges.

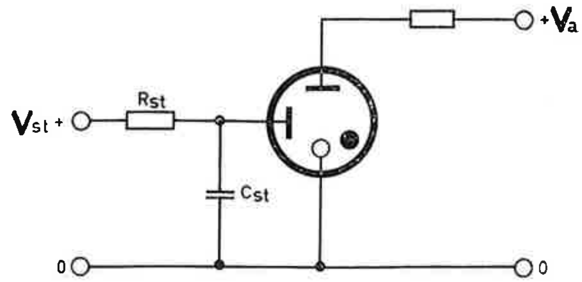

Fig. 4: Capacitance in parallel to the trigger-cathode section to improve the transfer characteristics.

It is a well-known fact that the voltages in the reverse phase or the negative grid bias of gas-filled tubes cannot be increased to any desired level without inviting self-ignition with all its harmful effects on the circuit and tube. This also applies to cold cathode trigger tubes. Since the limits in this case are somewhat closer and therefore more easily reached, it has become common practice to make due allowance for these values in the breakdown voltage graph. The same graph (see Fig. 6) therefore includes the useful "forward breakdown values" as well as the harmful "reverse breakdown values." In this graph, no reference is made to the particular discharge at any given time—a total of six different discharges possible between the three electrodes with two directions each—but simply

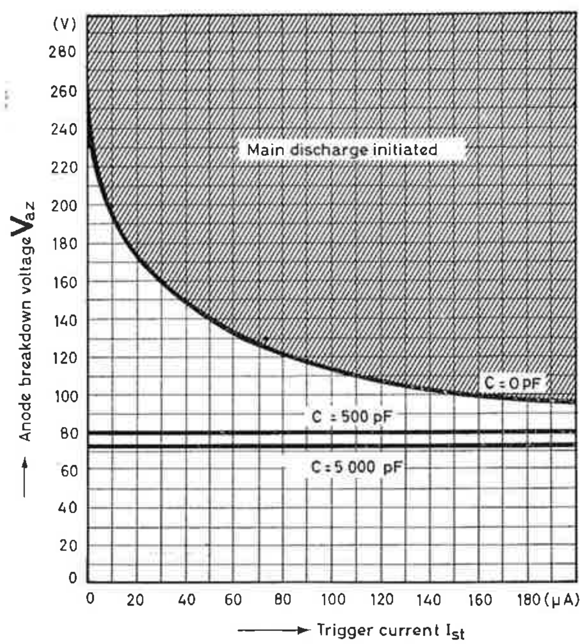

Fig. 5: Transfer characteristics as in Fig. 3, but together with those obtained in parallel with the capacitance shown in Fig. 4.

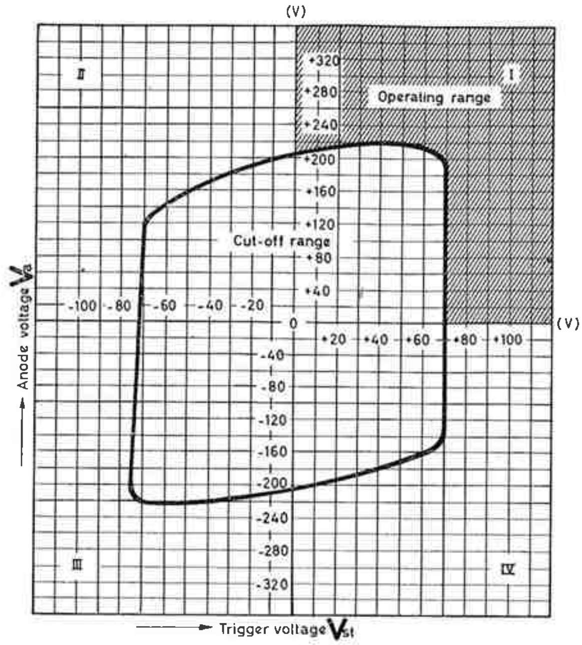

Fig. 6: Breakdown graph, showing the working range.

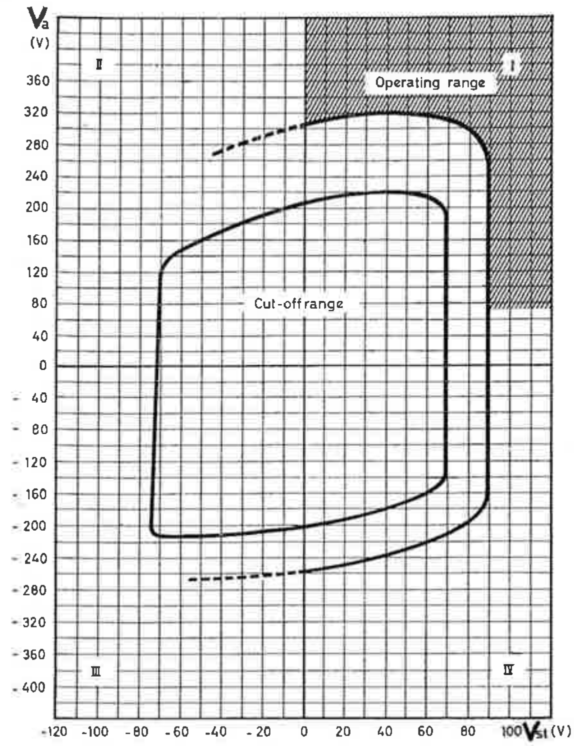

Fig. 7: Breakdown graph with tolerance values.

the limitation of the cutoff range, within which definitely no current flows in the tube represented (see Fig. 6).

The line representing ignition range is therefore a warning limit which must not be exceeded in any of the ranges in which a discharge in the wrong direction is possible, in accordance with the description above, concerning reverse or self-ignition.

This applies to all three quadrants with the exception of the upper right-hand quadrant, which generally refers to the normal operating range. This operating range contains the ignition characteristics of Fig. 1 (shown in a different position) but in this representation it is possible to dispense with the curvature towards higher pre-ignition voltages, since with V_{sz}, the trigger cathode section is at any rate always ignited and the breakdown graph merely states whether or not a discharge is maintained.

Normally, the breakdown graph for the operating range is specified with tolerances (Fig. 7), since it is important to know the extent to which the trigger values may vary between tube specimens, and in the course of the useful working life of the tube. The internal trigger characteristics refer to the minimum values for the warning ranges, which must not be exceeded; failure to observe this precaution may lead to self-ignition in certain tube specimens.

To summarise, the following may be said:

1. The trigger characteristics in the operating range (Quadrant I) indicate the voltages which must be supplied by the circuits to ensure reliable ignition. The same graph also specifies the voltage combinations in the other ranges (Quadrants II, III and IV) which must not be exceeded if damage to the tube is to be avoided.
2. The transfer characteristic refers to the currents required to flow towards the trigger electrode to ensure ignition of the main anode-cathode section. Its value may be the smaller, the greater the capacitance in parallel to the starter section.

(With acknowledgements to Telefunken)

WIDE-BAND AUDIO AMPLIFIERS

By B. J. Simpson

Introduction

One of the frequent arguments we hear on the subject of audio, and there are many of them to choose from, concerns the merits or demerits of having audio amplifiers with a bandwidth extending far beyond the limits of human hearing. On the face of it, this may seem a completely unnecessary thing to do, and a waste of time and money.

It is a fact, however, that some of the best amplifiers produced have had a wide bandwidth, extending at least two or three octaves beyond the limits of human hearing. In terms of frequency, taking the limits of hearing to be typically 15 cps to 15 Kc, this means a bandwidth, over which the response of the amplifier is substantially level, ranging from about 3 cps to 60 Kc.

It has also been shown subjectively that, where good quality components are used in the rest of the audio system, the amplifier with the extended bandwidth sounds better. In other words, if we take two amplifiers, both of which have substantially the same response characteristic within the limits of hearing, but one of them also has an extended response of perhaps two octaves beyond the top and bottom limits, then the amplifier with the extended bandwidth will sound better.

Music Reproduction

It has been argued, and with considerable justification, that whilst we carry out most of our tests on audio amplifiers with sine wave inputs, a programme input bears little resemblance to a sine wave. If it did, there would be no point in having lots of different instruments in the orchestra, for they would all sound the same. Musical instruments produce a complex wave sound structure, which is peculiar to each instrument, and is the property that enables us to recognise the instruments by ear when they are played.

To carry the matter even further, it would be true to say that each individual instrument has its own character, which allows us to differentiate between good instruments and bad. In poor amplifiers, with restricted frequency response and high distortion levels, it is possible for the reproduction of an instrument to bear so little resemblance to the original as to make it difficult to decide what the instrument is.

Whilst the waveform produced by a musical instrument is far from being a square wave, there is some justification, in view of its complex structure, for the view that a successful amplifier will be one that performs well with inputs more nearly approximating a square wave than with sine wave inputs. As we all know, the successful reproduction of square wave inputs requires a level frequency response and also a wide bandwidth.

Amplifier Distortion

A measure of the performance of an audio amplifier is the harmonic content in the output, when the amplifier has a pure tone input. But it is generally recognised today that a closer measure of performance, and one that can more easily be related to subjective tests, is the degree of intermodulation distortion produced in the unit. Any departure from linearity in the amplifier will increase both figures mentioned.

If we had to consider only fundamental frequencies and harmonics lying within the generally-accepted audio spectrum, then a response that was level over that range would be entirely satisfactory. But this is far from being the case. It has been shown that harmonics and other unwanted products lying well outside the audio range contribute distortion, particularly intermodulation distortion, where the beat from two supersonic overtones can and often does fall within the audio range. Lack of linearity in the supersonic range could therefore be capable of increasing distortion.

If we accept, even if only to a limited extent, the proposition that the general characteristics of our amplifier should be directed to the amplification of square waves rather than pure tones, then, apart from questions of intermodulation distortion, the requirement for a wide bandwidth is clear, as mentioned in the foregoing section.

These facts have been put forward for some years now by Williamson, Martin and others, who have also shown how to get the required results. It may be useful here to review very briefly the most important steps that have to be taken to get the sort of performance that seems to be indicated, together with a little philosophy.

Design Problems

There is little trouble experienced today in the production of an amplifier with a flat response over the audio spectrum and low values of distortion when measured with sine wave inputs. Amplifiers with distortion levels well below 1% and with a response within 1 db from 20 cps to 20 Kc are everyday matters, and practically anyone can build one.

With the restricted frequency range mentioned (in relation to the figures previously quoted), a common formula is to get a moderately good amplifier and then lean heavily upon negative feedback to reduce distortion. The feedback will also operate to flatten the response of the amplifier, but in general, with the small frequency range mentioned, little feedback is needed for that purpose alone.

In the better amplifiers, and especially when we are going for a wide bandwidth, not only is a different approach essential, but other considerations intrude. It is not proposed to discuss these in detail, but to mention the more important ones in order to draw the general picture.

A problem that always looms very large is the output transformer, assuming that we are not using an OTL* design. The transformer presents problems both at the very low frequencies and at the very high frequencies. In general, therefore, it must be the best component available.

In the rest of the amplifier, frequency range is limited at the low end mainly by the impedance of the interstage couplings, which increase sharply at the lower frequencies, and present severe attenuation. Increasing the values of the coupling capacitors is an expedient that has practical limitations, as we find ourselves calling for impossibly high values of capacitor. Increasing the input and output impedances of the individual stages helps, but usually at the expense of the higher frequencies, where shunt capacitances will offer a comparatively lower impedance. Direct coupling is a good solution, but is often beset with considerable difficulty.

At the higher frequencies, the two biggest problems are shunt capacitances, already mentioned, and Miller effect, which itself can offer comparatively low shunt impedances and high frequency losses. Neutralized balanced drivers can assist greatly here, although they are usually more costly than the more conventional systems.

With our feedback we can do two things. We can extend the frequency range, and we can lower the distortion. It is like having a fixed sum of money to distribute between two heirs; one of them can only gain at the expense of the other. Where our amplifier is concerned, the most effective solution is to put one of the factors into the position where he does not need much money; this allows the other to get most of the benefit.

Reduced to practical terms, we make the amplifier to have as low a distortion content as possible BEFORE we apply the feedback, so that the maximum benefit can be taken in regard to frequency response. In other words, we are not relying 100% on the feedback to give us low distortion figures.

All this calls for a more careful approach to the problem. When we consider what feedback we can apply, we find that the figure available is limited by phase shift in the amplifier, and low phase shift again implies a good amplifier without feedback. Direct-coupled drivers and internal phase-correcting feedback loops can be very beneficial here. Large amounts of negative feedback can then be used around the driver and output stages to extend frequency response and reduce distortion still further at the most troublesome point. Note that this heavy feedback loop will include what is probably the most trouble-

*Output transformer-less.

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A 120-WATT 50-MC TRANSMITTER

By George D. Hanchett, W2YM

RCA Semiconductor and Materials Division

Although an active ham for more than three decades, the author had never tried transmission on six metres—until this year. Preparatory steps involved the usual search around the shack for parts suitable for a low-powered rig. W2YM decided that only the rf unit and antenna had to be built. Power supplies and the modulator were borrowed from a two-metre transmitter. The available power supplies limited the rig to a power level of about 100 watts. Because the author's location is in a Channel-2 area, the 120-watt 50-megacycle transmitter—well shielded throughout—is of proved straightforward design.

Front view of W2YM's 120-watt 50-megacycle transmitter. [Note the air-intake holes on the side of the blower (or bottom) chassis. They each measure $\frac{3}{8}$ inch in diameter.]

Circuit Considerations

Initial step in planning the 120-watt 50-megacycle transmitter was to lay out a three-stage rf section having a VFO-driven multiplier or crystal oscillator as the first stage and an 829B as the final stage (see Figure 1).

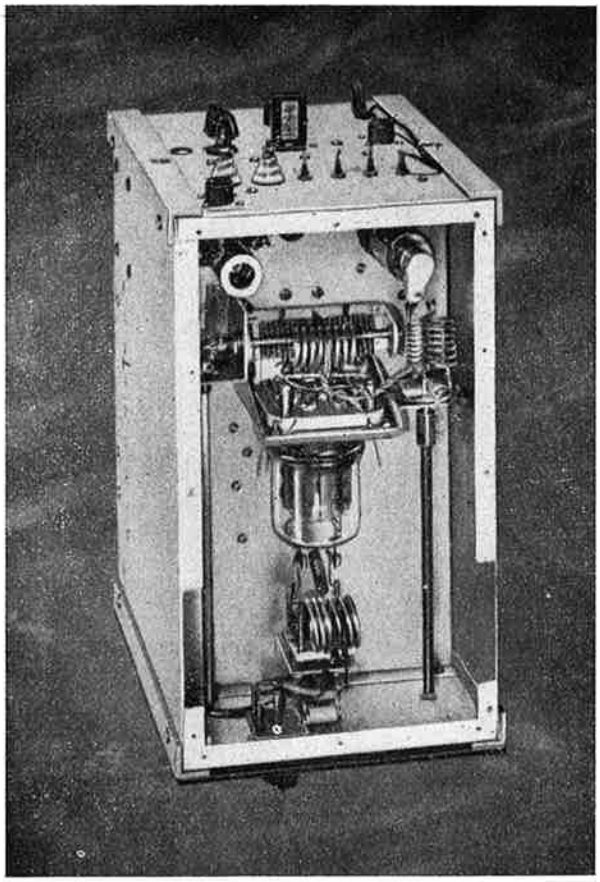
A 12BY7A oscillator-tripler was arranged so that it operates as a grid-plate oscillator in the crystal-control position. This oscillator uses 8-megacycle crystals and its output is tuned to the third harmonic. In the VFO position, the 12BY7A stage can be either an amplifier or a multiplier, and can be driven by a VFO with 8-, 12-, or 25-Mc output.

The oscillator-multiplier is capacitively coupled to a 2E26 doubler which has a 50-Mc output. The 2E26 is link-coupled to the grids of the 829B. Link coupling was used because it facilitates coupling of a single-ended stage to push-pull grids.

In addition, the use of the double-tuned circuit provides extra selectivity in the grid circuit of the 829B amplifier and, thereby, reduces the possibility of harmonic interference to FM and TV reception.

The 829B final-amplifier plate circuit is tuned by a butterfly capacitor. The rotor section of this capacitor is ungrounded to improve balance. The antenna is link-coupled to the final tank circuit and is equipped with a 50-micromicrofarad capacitor which tunes out the inductance of the link winding.

Metering of different circuits is accomplished by use of a 0-1 milliampere meter. Suitable meter shunts are used in order to meter oscillator plate current (30 milliamperes full scale), doubler-grid current (2 ma full scale), doubler plate current (100 ma full scale), final grid current (30 ma full scale), final screen-grid current (100 ma full scale), and final plate current (300 ma full scale). A tuning switch is incorporated because it not only aids in the tune-up procedure, but also saves valves and prevents possible damage to other components.



Top view of inside of utility box shows details of grid and plate circuit for the 829B. Also note that a portion of the utility box flanges have been removed to allow for insertion of subchassis.

Construction

Completely contained in a 12- by 7- by 6-inch aluminium utility box, the 50-megacycle transmitter is fitted with an aluminium subchassis. This subchassis has small, ½-inch lips which are bent on the long sides of the chassis to provide stiffness.

Half-inch tabs—bent on the front and rear of the chassis—serve as mounting brackets. One set of these tabs is bent up, the other down. Without this feature it would be impossible to insert the subchassis into the utility box.

For the same reason, two slots are cut in the top flanges of the utility box. These slots are visible in the photograph, which shows a top view of the transmitter.

A rectangular cutout at the rear of the subchassis fits around the power-lead filters which are mounted on the rear wall of the utility box. All leads entering or leaving the utility box are brought out through low-pass filters.

When complete, the utility box is mounted on a 12- by 7- by 3-inch aluminium chassis that serves as a bottom cover as well as a housing for the cooling fan and filament transformer. (The schematic for the bottom chassis is shown in Figure 2.)

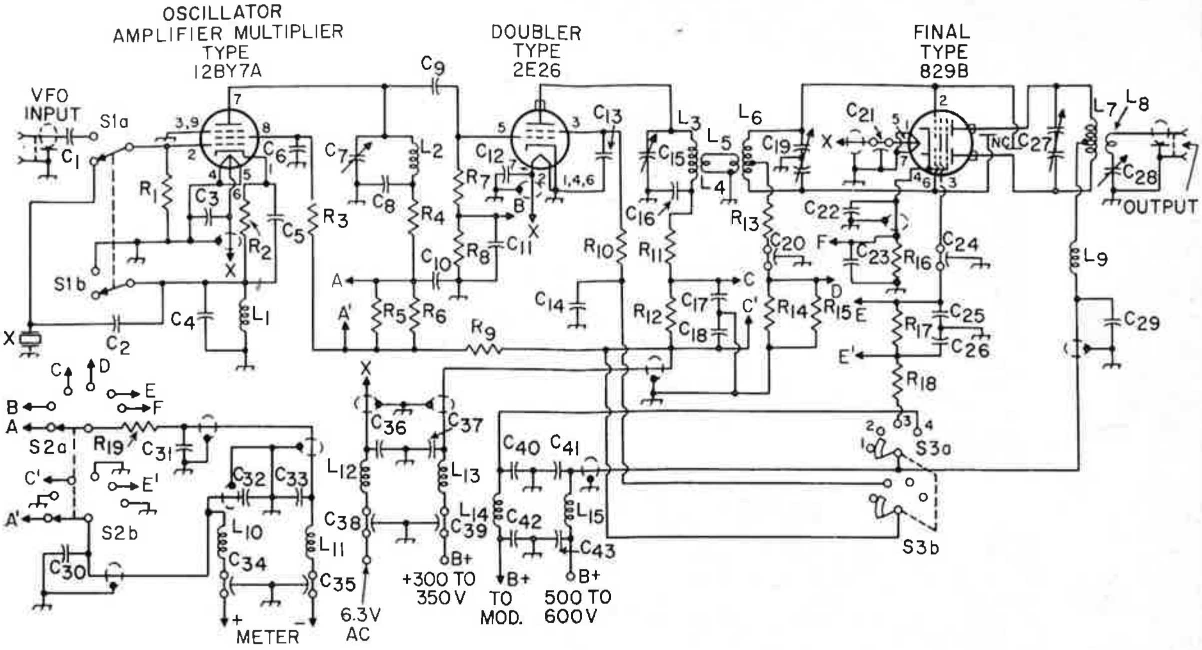
A right-angle drive for the final-amplifier grid capacitor was made from two brass-bevelled gears. Each gear is secured on its shaft with two Allen-set screws spaced 90 degrees apart.

As shown in the sketch of the grid assembly (see Figure 3), the socket, grid coil, and neutralizing capacitors for the 829B are mounted on an aluminium bracket. The two top-mounting screws of the socket, together with a polyethylene strip, are used as feed-through connections.

The holes in the brackets must be enlarged so that the neutralizing capacitors do not short to ground. In the Figure 3 sketch, these holes are enlarged to ⅜ inch.

Note that, during construction, the grid leads of the 829B are criss-crossed. The neutralizing capacitors are small pieces of No. 12 wire which are close to the plate region of the 829B. Neutralization is accomplished by adjustment of the length of these wires, as described below.

The heater, screen-grid, and control-grid by-pass capacitors (as well as the output capacitor of the low-pass filters for the meter leads, heaters, and 350-volt B+) are 0.001 μf feed-throughs. The high-voltage and modulator power-lead feed-throughs are ceramic units which are externally by-passed.



C_1—220 pf, mica, 500 volts.

C_2—10 pf, mica, 500 volts.

C_3, C_5, C_6, C_8, C_{10}, C_{11}, C_{12}, C_{13}, C_{14}, C_{16}, C_{17}, C_{18}, C_{22}, C_{23}, C_{25}, C_{26}, C_{30}, C_{31}, C_{32}, C_{33}, C_{36}, C_{37}—1,000 pf, disc ceramic, 1,000 volts.

C_4—100 pf, mica, 500 volts.

C_7, C_{28}—3.7-52 pf, variable, air gap 0.015 inch.

C_9—47 pf, mica, 500 volts.

C_{13}—5.2-30 pf, variable, air gap 0.045 inch.

C_{19}—5.0-28.5 pf double-section variable, air gap 0.045 inch.

C_{20}, C_{21}, C_{24}, C_{34}, C_{35}, C_{38}, C_{39}—1,000 pf, feed-through, ceramic, 500 volts.

C_{27}—4.8-27.3 pf, butterfly, variable, air gap 0.030 inch.

C_{29}, C_{40}, C_{41}, C_{42}, C_{43} — 1,000 pf, disc ceramic, 3,000 volts.

L_1—RF choke, 1 mh.

L_2—10 turns of No. 20 tinned on $\frac{1}{2}$-inch diameter form, winding length $\frac{3}{4}$ inch.

L_3—5$\frac{1}{2}$ turns of No. 10 solid on $\frac{5}{8}$-inch diameter, winding length 1 inch.

L_4, L_5—2 turns of No. 20 plastic covered on $\frac{1}{2}$-inch diameter.

L_6—8 turns of No. 10 solid on $\frac{5}{8}$-inch diameter, winding length 1$\frac{1}{2}$ inches.

L_7—6 turns of No. 10 solid on $\frac{7}{8}$-inch diameter, winding length 1 inch.

L_8—2 turns on No. 14 Enam. covered with insulation tubing on $\frac{5}{8}$-inch diameter.

L_9, L_{10}, L_{11}, L_{13}, L_{14}, L_{15}—RF choke, 7 μh, 1,000 ma.

L_{12}—RF choke, 25 turns of No. 16 Enam. close-wound on $\frac{1}{4}$-inch diameter.

NC — Neutralizing capacitors: No. 12, tinned wire ($\frac{1}{2}$-inch length placed in proximity of 829B plates); see text.

R_1—100,000 ohms, 0.5 watt.

R_2—120 ohms, 0.5 watt.

R_3—33,000 ohms, 0.5 watt.

R_4, R_8, R_{11}, R_{19}—1,000 ohms, 0.5 watt.

R_5, R_{14}—47 ohms, 0.5 watt.

R_6, R_{15}—130 ohms, 0.5 watt.

R_7—47,000 ohms, 1 watt.

R_9—3,300 ohms, 1 watt.

R_{10}—10,000 ohms, 2 watts.

R_{12}—10 ohms, 0.5 watt.

R_{13}—56,000 ohms, 2 watts.

R_{16}—3.3 ohms, 0.5 watt.

R_{17}—33 ohms, 0.5 watt.

R_{18}—15,000 ohms, 10 watts, wire wound.

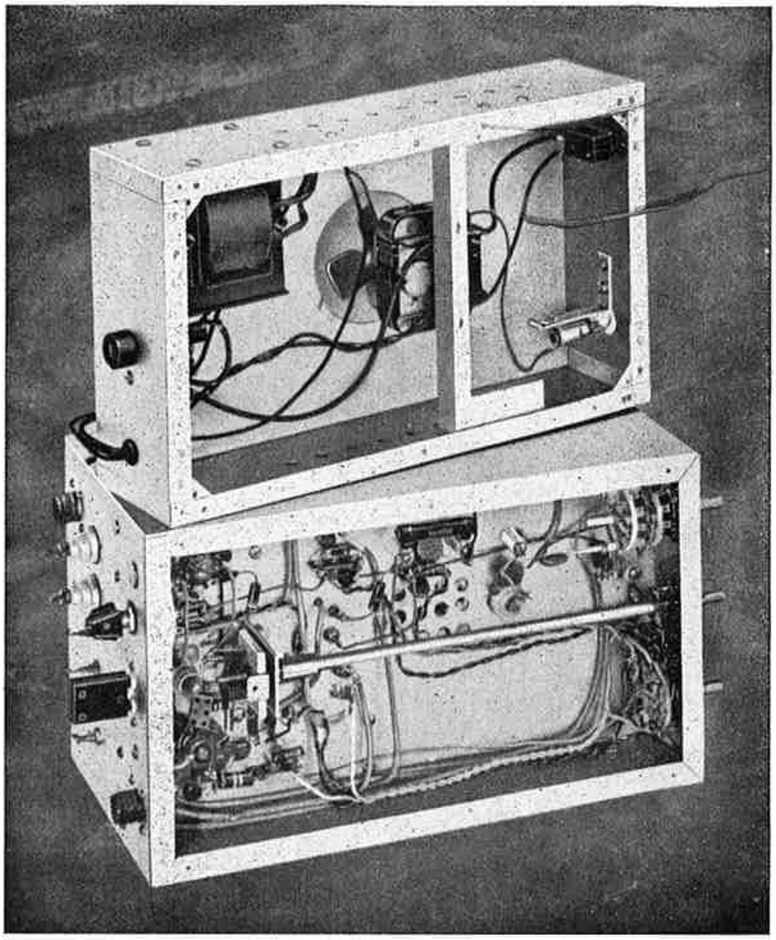
S_1—Crystal-VFO Switch; two-pole, two-position, wafer, non-shorting, rotary.

S_2—Meter Switch; two-pole, six-position, wafer, non-shorting, rotary.

S_3—Tuning Switch; 60-degree indexing; two progressively shorting 30-degree wafers, using every second contact.

Miscellaneous—One crystal socket; one 829B socket; one octal socket; one 9-pin min. socket and shield; one 6- x 7- x 12-inch utility box; one 7- x 12- x 3-inch chassis; two ceramic feed-through insulators; two bevelled gears, 90 degrees; one pilot light (green); one fuse; one cord set; one insulated plate cap; two coaxial connectors. **Note: All resistors have 10% tolerance.**

Figure 1: Schematic diagram and parts list of the rf section of W2YM's 50-megacycle transmitter.

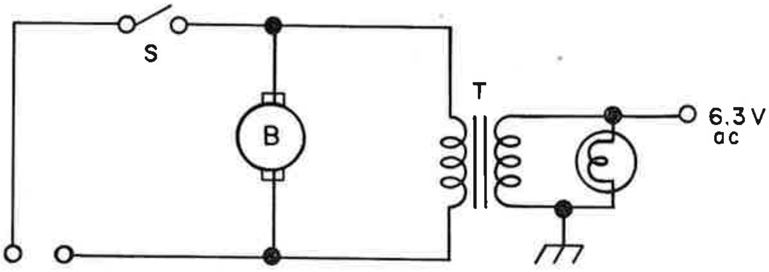


Bottom views of both utility box and blower chassis before assembly as combined unit. Note blower fan in the blower chassis. Also note that the 12 holes in the subchassis underneath the 829B allow for free flow of air around the valve.

The tuning switch is constructed from a 60-degree detent assembly and two progressively shorting wafers. In the first position, the screen-grid voltage is removed from both the 2E26 and the 829B, as well as the plate voltage to the external modulator. Advancing the switch in a clockwise manner activates each stage in turn until, in the fourth position, the complete transmitter is in operating condition.

As previously mentioned, the filament transformer and cooling fan are mounted in the blower (or bottom) chassis (Figure 2). The cooling-fan blade is positioned in a $2\frac{3}{4}$-inch hole in the blower chassis. Holes drilled in the sides of this chassis provide an air inlet for the fan. As a result, air is freely circulated around the 829B.

Although the bottom cover is one of the original utility box covers, the top cover is a piece of perforated aluminium. The utility box and the blower chassis are held together by four $\frac{3}{4}$-inch angles which can be made from "do-it-yourself aluminium." The front panel is cut from $\frac{1}{8}$-inch aluminium stock and fastened to the front



B—Cooling fan and motor.
S—Toggle switch (SPST).
T—6.3-volt filament transformer (6 amperes).

Figure 2: Schematic diagram and parts list of 50-megacycle transmitter's bottom chassis.

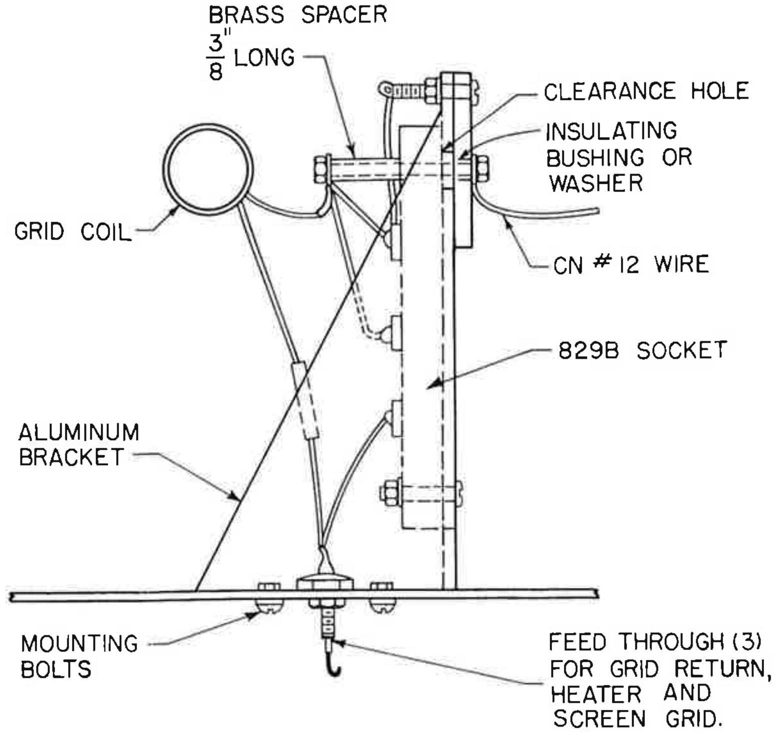
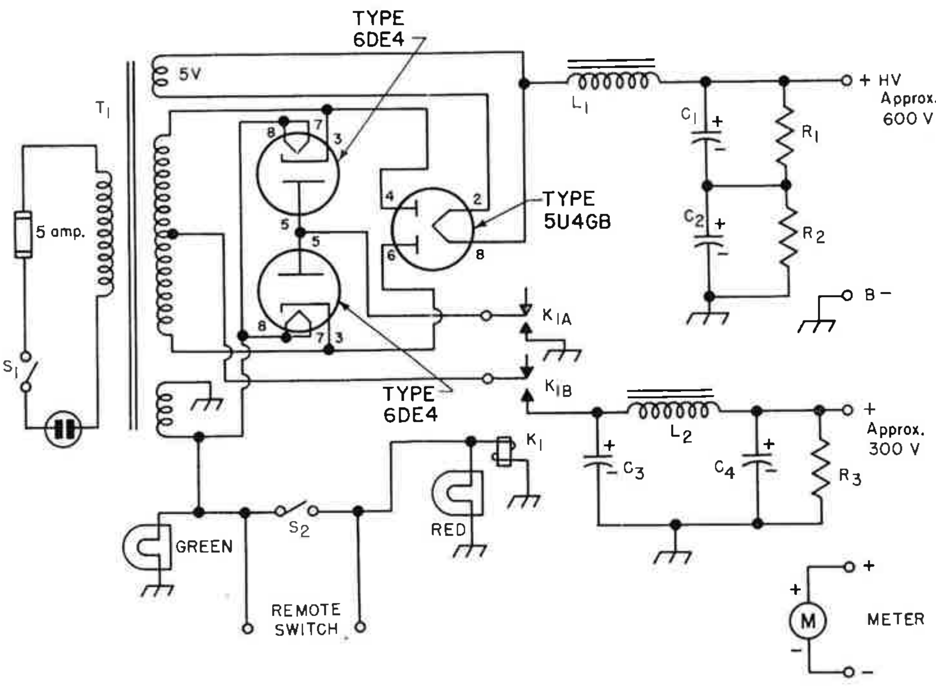
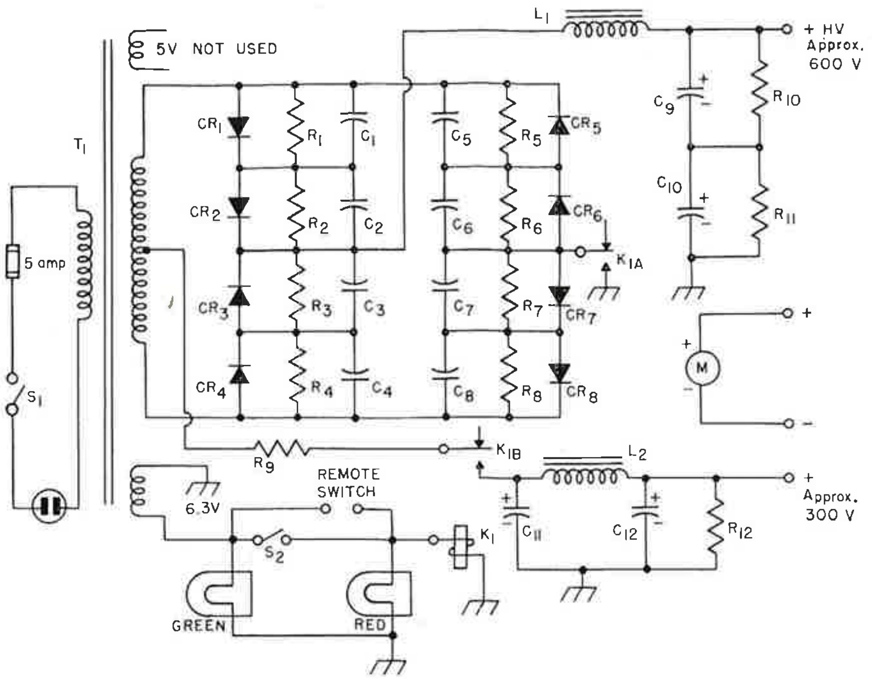


Figure 3: Sketch of grid assembly shows details of grid coil mounting and neutralizing capacitors.



C_1, C_2, C_3, C_4—40 μf, 450 volts, electrolytic.
K_1—Relay.
L_1—2.8 henry, 300 ma.
L_2—4 henry, 175 ma.
M—Milliammeter 0-1 ma dc.
R_1, R_2—15,000 ohms, 10 watts.
R_3—47,000 ohms, 2 watts.
S_1—Toggle switch (SPST).
T_1—Power transformer, 360-0-360 V, 250 ma.

Figure 4: Schematic diagram and parts list of suggested power supply circuit using vacuum rectifier valves.



C_1, C_2, C_3, C_4, C_5, C_6, C_7, C_8—.001 μf, 1,000 volts, disc ceramic.
C_9, C_{10}—40 μf, 450 volts.
CR_1, CR_2, CR_3, CR_4, CR_5, CR_6, CR_7, CR_8—1N2864.
K_1—Relay.
L_1—2.8 henry, 300 ma.
L_2—4.0 henry, 175 ma.
M—Milliammeter 0-1 ma meter, dc.
R_1, R_2, R_3, R_4, R_5, R_6, R_7, R_8—100,000 ohms, 0.5 watt.
R_9—47 ohms, 1 watt.
R_{10}, R_{11}—15,000 ohms, 10 watts.
R_{12}—47,000 ohms, 2 watts.
S_1, S_2—Toggle switches (SPST).
T_1—Power transformer, 340-0-340 V, 300 ma.

Figure 5: Schematic diagram and parts list of suggested power supply circuit using silicon rectifiers.

aluminium angles of the transmitter. The panel may then be painted and lettered.

In the transmitter, the 0-1 milliampere meter is not included because it is an integral part of the power supply unit used. Figures 4 and 5 show the power supplies that have been successfully used with this transmitter.

Transmitter Adjustments

With the top and bottom covers removed and the utility box detached from the blower chassis, make temporary connections to the heater circuit and ground the utility box to the blower chassis. Then, after checking all wiring, turn on the ac power to the fan and heater-filament transformer.

See that all valves are properly lit. With the tuning switch in position No. 1, temporarily connect the 300-volt B+ to its proper terminal. Turn on the power and adjust C_7 for the maximum grid current of the 2E26 (approximately 1.0 to 1.2 milliampere). Turn the meter switch so that the oscillator plate current can be read. (This value should be between 12 and 18 milliamperes.) With a wave meter or grid-dip meter in the diode position, check that the plate circuit of the 12BY7A is three times the crystal frequency.

Shut off the plate supply and advance the tuning switch to position No. 2. Reapply the 300-volt B+ and quickly adjust C_{15} and C_{19} for maximum 829B grid current. Adjust the link coupling so that a grid current of approximately 10 milliamperes is flowing to the 829B grid. In making adjustments of the link coupling, be sure you turn off the B+ voltage because 300 volts is exposed at the 2E26 plate coil. At this point, by using a wave meter or grid-dip meter, check again that these circuits are on 50 megacycles.

If a dip in the grid-current occurs with excitation while C_{27} is tuned, the 829B neutralization must be adjusted by cutting $\frac{1}{8}$-inch lengths from the neutralizing wires until there is no noticeable change in the grid current as C_{27} is tuned through resonance.

Next, connect a load to the antenna connector of the transmitter. (If you do not have a non-inductive 50- or 75-ohm load, a 100-watt light bulb can be substituted.)

Temporarily attach the 600-volt high-voltage lead to its proper terminal. With the tuning switch in position No. 3 and the meter switch in the 829B cathode-current position, turn on the plate supplies. Quickly rotate the plate tank capacitor C_{27} for minimum dip. Now load the amplifier by adjusting L_8 and C_{28} until the total cathode current is approximately 200 milliamperes.

Return the meter to the 829B grid position and adjust the link coupling between the 2E26 and the 829B until the grid current is 15 milliamperes. Again return the meter switch to the cathode-current position and adjust the loading until the cathode current is approximately 240 milliamperes.

The 829B screen-grid current should now be between 25 and 30 milliamperes. If the screen-grid current is above 30 milliamperes, the screen-grid dropping resistor must be increased; if below 25 milliamperes, it must be decreased. It is unlikely, however, that you will require any adjustment of this dropping resistor.

The utility box now can be assembled to the blower chassis. The bottom cover, as well as the top perforated cover, also can be permanently attached. The transmitter and modulator leads can be connected permanently, too. What else? Connect a six-meter antenna and you are ready to go on the air.

A few words of precaution: Never attempt to modulate this transmitter unless the 829B is completely loaded. If the transmitter is not loaded, flash-over will occur, since the butterfly capacitor voltage breakdown ratings will be exceeded. An output in the order of 75 to 85 watts should be attained.

Valve Selection

Potential builders of this 120-watt 50-megacycle transmitter may be interested in knowing why particular valve types were selected. Here are the more important reasons:

- This writer was interested in attaining high reliability consistent with a moderate cost. The valves chosen have been in production for many years and have demonstrated excellent life characteristics.

- Instead of the 829B, two 6146's could have been used for the final stage. However, the spacing and construction would have been more difficult. In addition, while the 6146 would have cost less, the difference in price between this valve and the one used was not great enough to justify the extra mechanical problems that would have been brought about as a result.

- By replacement of the 2E26 with its 12-volt version (the 6893), and by rewiring of the 12BY7A and 829B, the transmitter can be used —with 12-volt heaters—in mobile installations.

All in all, the valve line-up selected appears to provide the best balance of power, cost and reliability.

(With acknowledgements to RCA)

WIDEBAND AMPLIFIERS

CONTINUED FROM PAGE 240

some single component in the amplifier, the output transformer.

This, then, in very general terms, is the formula for an amplifier that is just that little extra. It raises the interesting and provocative thought that a good guide to the inherent quality of an amplifier is the performance obtained with the main feedback loop disconnected.

Summary

Some of the popular expressions of amplifier performance tell a story that is an expurgated version, and many amplifiers, in spite of their figures, have a less than adequate performance. That little bit extra cannot of course be had without spending a little more money; what is sometimes surprising is just how small that extra contribution can be. If the opportunity offers, we hope next year to demonstrate some of the things mentioned here, and to show how theory is converted to practice.

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

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