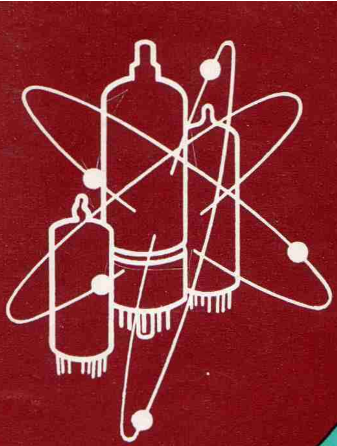

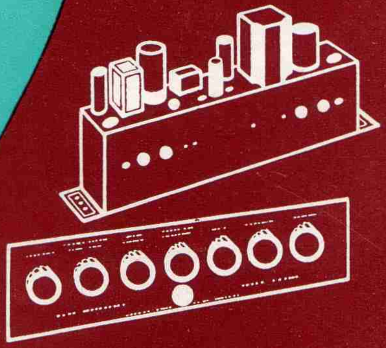

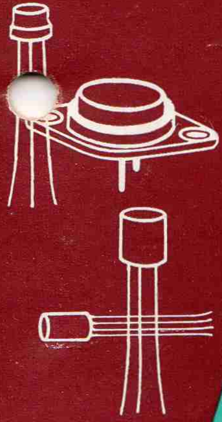

RADIOTRONICS

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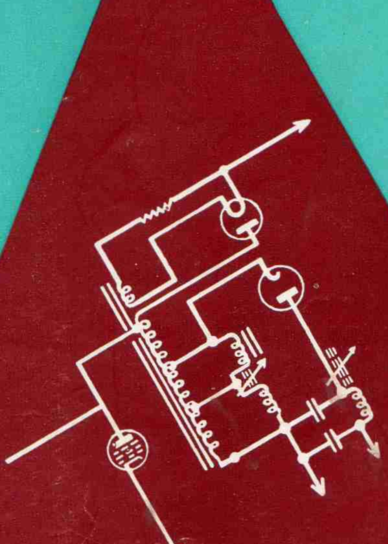

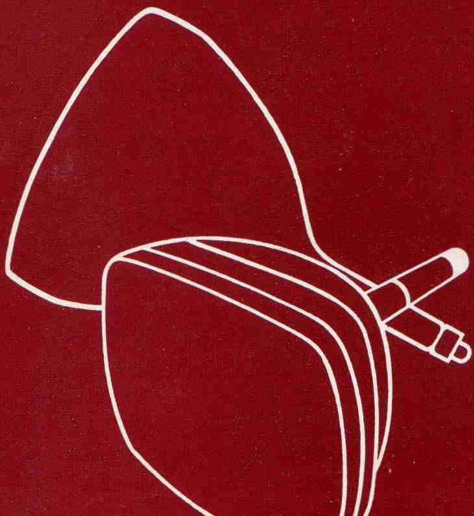

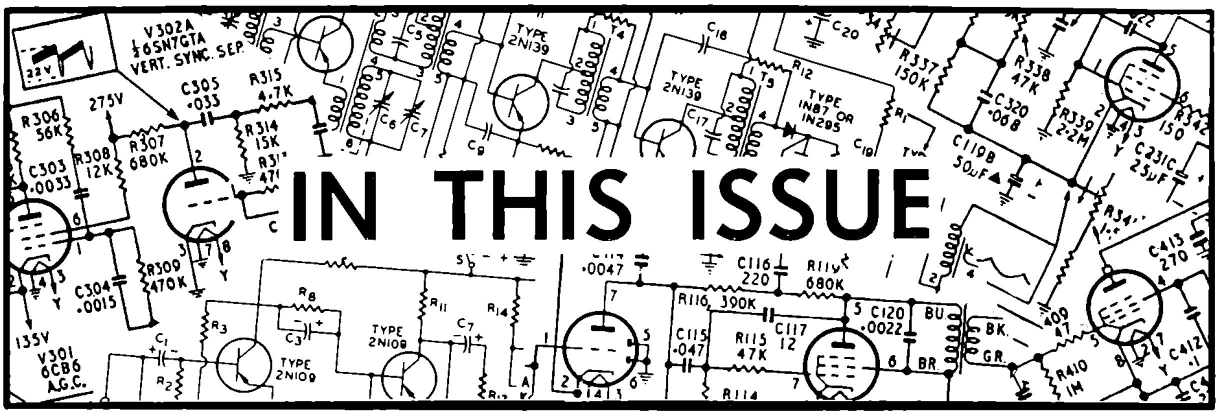

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Part two of this interesting and instructive article by J. van der Goot (AWV Application Laboratory) deals with horizontal oscillators and afc systems, and follows logically the discussion of vertical oscillators which was given in the first part last month.

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The conclusion of this excellent article by Harry F. Olson completes the presentation of our first article on stereo. Mr. Olson, of the RCA Princeton Laboratories, is a world authority on acoustical engineering, and his article carries the mark of that authority.

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The 21" 110° picture tube is described, with full technical data.

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Three new publications are reviewed this month:
"Servicing Transistorized Radios"
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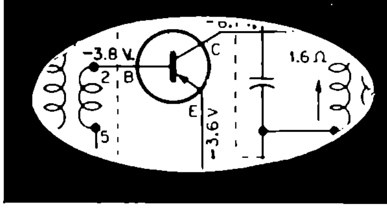
DESIGN-MAXIMUM SYSTEM FOR RATING VALVES 102

An introductory note on the design-maximum system is appropriate at this time and foreshadows its great use in the electronics industry consequent upon the advantages it offers over the two other systems. Further information will be presented soon in these pages.

The swing over to the Design-Maximum System of rating valves announced by RCA and other manufacturers will take effect with new entertainment types issued. Later, ratings for the more popular entertainment types already published under the Design-Centre System will be reissued in the Design-Maximum System. The merits of the new system are described briefly in the note printed in this issue on page 102.

TRANSISTOR APPLICATIONS



CHAPTER ONE

TRANSISTOR AMPLIFIERS

Before undertaking the study of transistor amplifiers it should be pointed out that the manner of amplification in a transistor differs from that of a thermionic valve. It is well known that a valve is a **voltage** operated device. That is, an ac voltage is applied to the grid of the valve to control the current flow between the cathode and plate. A transistor, however, is a **current** operated device. That is, the **current** flowing in the emitter-base circuit controls the current flowing in the collector circuits. The symbols α (alpha) and/or β (beta) are used to indicate the current gain in a transistor as compared to the symbol μ (mu) to indicate voltage gain in a valve. Another important factor is the transistor input and output resistances. It is the current flowing through this resistance that determines the voltage or power gain of a transistor amplifier.

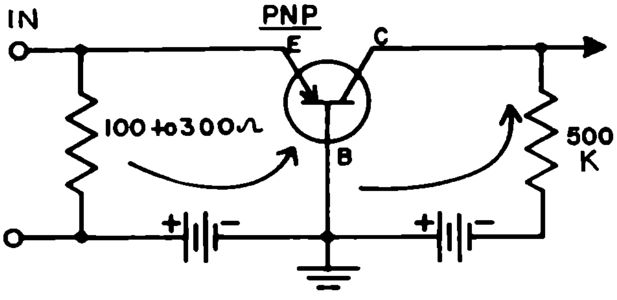
There are basically three types of transistor amplifiers; the common or grounded base, the common or grounded emitter and the common or grounded collector. The term **grounded** will be used throughout the lesson since it has greater usage in the field than the word common.

Grounded-Base Amplifier

The grounded-base amplifier circuit is similar to the grounded-grid valve amplifier which has extensive use as an rf amplifier in television tuners. A comparison of these two basic circuits is shown in Figure 1. As can be seen, the base of the transistor and the grid of the valve are grounded. The emitter is biased in the direction of greatest electron flow and the collector is biased in the direction of least electron flow. With this bias arrangement the input of the transistor has a low resistance in the order of 20 to 50 ohms and the output has a high resistance which is approximately 1 to 2 megohms. The current gain (or alpha) is always less than unity in this type of circuit and is usually in the order of 0.98 to 0.99. The resistance gain between base and collector is very high. The voltage gain in this type of circuit may be in the order of 1500.

The characteristic curve of a grounded-base circuit is shown in Figure 2. It can be seen from the characteristic curve that the collector current never exceeds the emitter current.

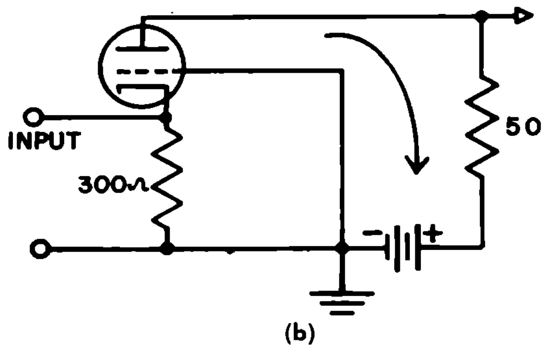
It is very important to maintain the proper collector voltage and emitter current so as not to exceed the maximum dissipation of the transistor. The grounded-base circuit is used when a circuit calls for a very low input impedance and very high output impedance.



(a)

Arrows indicate direction of electron flow.

Emitter arrow indicates direction of current flow.



(b)

FIGURE 1

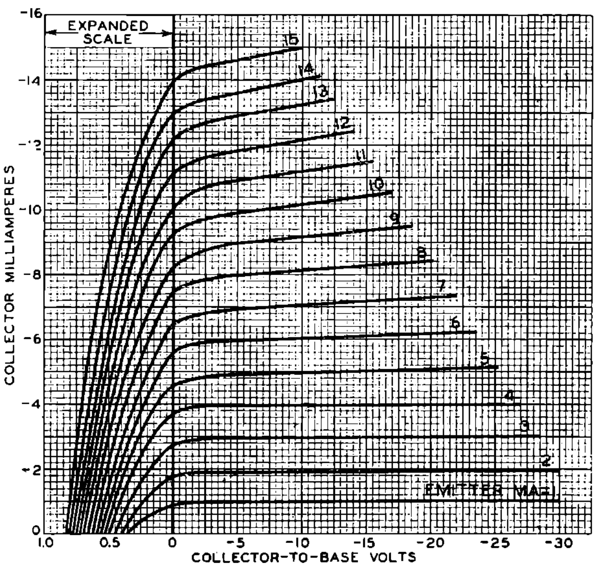


FIGURE 2

The biasing arrangement of the grounded-base circuit can be simplified as shown in Figure 3. Figure 1 shows two bias supplies, one for the emitter and the other for the collector. The insertion of a resistor in the base circuit biases the emitter positive with respect to the base. Thus, by selecting the proper resistance the proper emitter current can be obtained. Signal degeneration will be introduced by this resistor, therefore, it is by-passed with a capacitor to put the base effectively at ac ground.

Grounded-Emitter Amplifier

The grounded-emitter amplifier is similar to the conventional, grounded-cathode, valve amplifier. The signal is applied to the base of the transistor, whereas in the valve the grid is the driven element. A comparison between the transistor and valve circuits is shown in Figure 4.

As in the case of the grounded-base amplifier, the emitter is biased in the direction of greatest current flow and the collector is biased in the direction of least current flow. The resistance of the input circuit is normally in the range of 1,000 to 2,000 ohms, however, it may be as low as 100 ohms or as high as 10,000 ohms. The output resistance is normally about 50,000 ohms, however, it may be as low as 5,000 ohms and as high as 500,000 ohms.

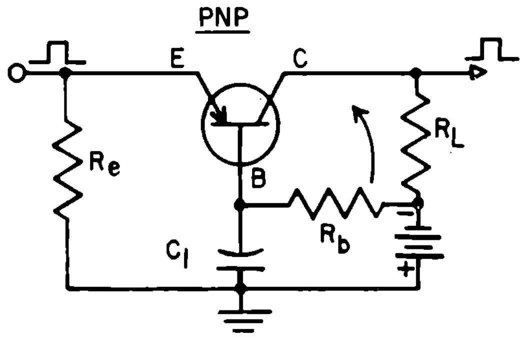


FIGURE 3

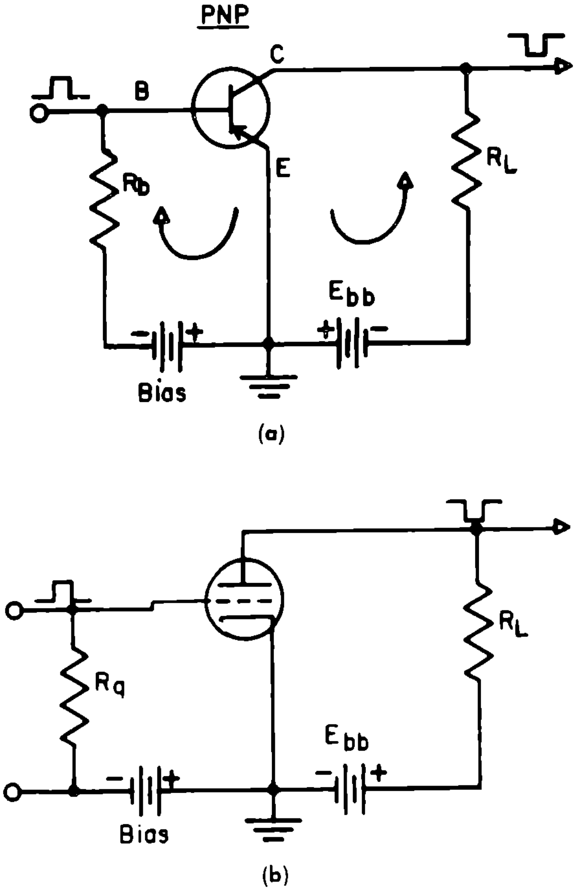
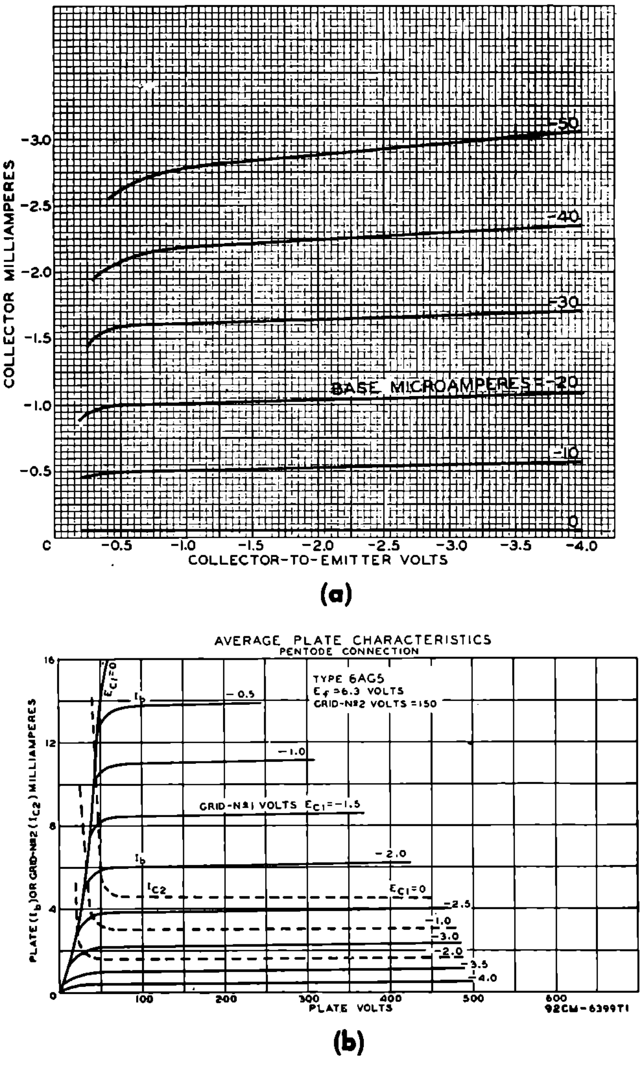


FIGURE 4

The characteristic curve for a grounded emitter circuit is shown in Figure 5. This curve is for the AWV Type 2N105 p-n-p junction transistor and is compared with the I_p-E_p curve of a 6AG5 valve. The major difference lies in the fact that in a transistor the collector current is controlled by the emitter current, whereas, the valve plate current is controlled by the grid to cathode bias voltage. Aside from this, the curves are used in a similar manner. From the characteristic curve shown in Figure 5 (a), it can be seen that for a very small change in base current, a relatively large change in collector current is possible. For example, as shown in Figure 5 (a), with a collector voltage of minus 4 volts, a change of base current from .20 to 30 microamperes produces a change of collector current from 1.1 to 1.7 milliamperes. We can say then, that a 10 microampere change in base current produces a 600 microampere change in collector current; thus, a current gain is realised under these operating conditions for the transistor. This current gain is called the **beta** (β) gain between the base and collector as compared to the alpha (α) gain between the emitter and collector in the grounded-base amplifier.

Power gains of 42 db or approximately 10,000 times can be realized with this circuit arrangement. The voltage gain of this circuit arrangement is the same as that of the grounded-base connection, but the current gain is considerably higher. Due to this increase of gain over the grounded-base connection, this circuit is popular for many circuit designs.

As with the grounded cathode valve circuit, a voltage reversal takes place between base and collector. A positive signal at the base opposes the bias voltage causing a smaller base current, thus decreasing the collector current. The decrease in collector current causes the collector to become more negative. Therefore, a positive signal at the base develops a negative signal at the collector.



(a)

(b)

FIGURE 5

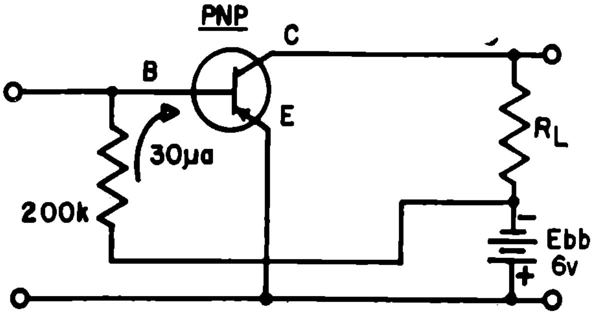


FIGURE 6

Various methods have been established for biasing the base in the grounded-emitter circuit. In Figure 4, two separate bias supplies are used. It can be noted, however, that both the base and collector are negative with respect to the emitter. Therefore, the bias arrangement can be simplified as shown in Figure 6. This method can be called **fixed bias.**

The base current can be established by choosing the proper value of base resistance R_b. For example, if a base current of 30 microamperes is desired:

$$R_b = \frac{E_{bb}}{I_b}$$

$$= \frac{6 \text{ volts}}{30 \times 10^{-6} \text{ amps}}$$

$$= 200\text{K ohms}$$

This value of base resistance includes the internal emitter to base resistance, however, this represents only a few hundred ohms and the internal resistance can normally be neglected.

Fixed bias is not the most satisfactory method of biasing the base. Due to variations between transistor units and their sensitivity to temperature changes, it is difficult to maintain a critical base current. One method of partially overcoming

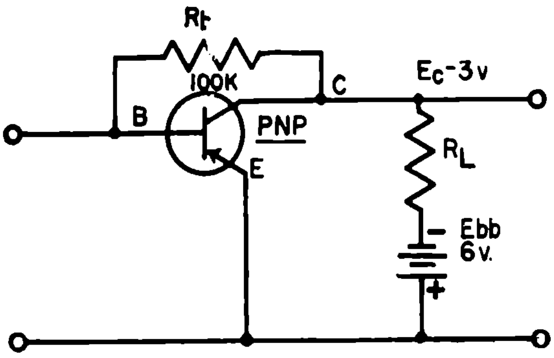


FIGURE 7

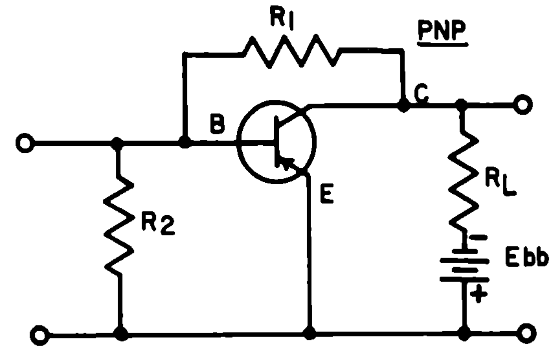


FIGURE 8

this problem is to tie the base resistor directly to the collector as shown in Figure 7. This arrangement provides degeneration in a form of **automatic control** of the base bias and can be called **self bias.** To determine the value of R_b, the supply voltage (E_{bb}) is replaced by the collector voltage (E_c) in the previous formula:

$$R_b = \frac{E_c}{I_b}$$

$$= \frac{3 \text{ volts}}{30 \times 10^{-6} \text{ amps}}$$

$$= 100\text{K ohms}$$

This method of self bias causes ac negative feedback which, although it overcomes many of the disadvantages of fixed bias, reduces the effective gain of the amplifier.

Both fixed and self bias can be used to provide even better circuit stability. This method is illustrated in Figure 8. Here a voltage divider composed of R_1 and R_2 biases the base negative with respect to the emitter. Bleeder current through the voltage divider fixes a bias at the base. However, any change in collector voltage due to a change in emitter current will automatically change the base bias. This circuit is commonly used due to its inherent stability.

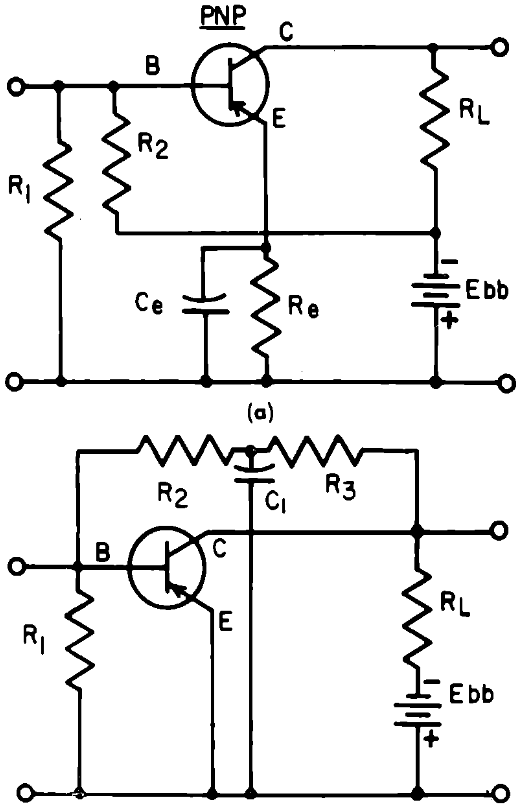
To minimize loss of gain either of the two circuits shown in Figure 9 may be used. In Figure 9 (a) a resistor is added to the emitter circuit and R_2 is returned to the negative terminal of the battery instead of the collector. The emitter resistor (R_e) provides additional stability and is usually $\frac{1}{5}$ to $\frac{1}{10}$ the value of R_1. To prevent emitter degeneration, capacitor C_c is added. The value of this capacitor is usually about 50 μf; however, the value may be much higher depending, among other things, upon the lowest frequency to be amplified. The emitter resistor in this case is similar to the cathode resistor in a valve circuit.

Another method is shown in Figure 9 (b). Here the voltage divider is split and all ac variations are bypassed by capacitor C_1. The value of R_3 is usually 5 to 10 times the value of R_2. The total resistance of R_2 and R_3 should equal the resistance of R_1 shown in Figure 8. In some circuit applications a combination of Figure 9 (a) and 9 (b) may be used. In other cases a bypassed emitter resistor may be added to the circuit of Figure 8. Voltage gain and circuit stability are usually the determining factors when selecting the proper circuit.

Grounded-Collector Amplifier

The grounded-collector amplifier circuit is similar to the valve cathode follower circuit. Both circuits are illustrated in Figure 10. The input impedance (resistance and reactance) of the transistor circuit is high and the output impedance is low, being similar to the valve circuit. The output impedance of the transistor circuit is dependant on the input impedance; however, this is not the case in valve circuits. The voltage gain is less than unity and the power gain of the stage is usually lower than either the grounded emitter or grounded base stages. The circuit is mainly used as an impedance matching device.

As in the case of the grounded base amplifier there is no phase reversal of the signal between the input and output. The same is true in the cathode-follower and the grounded grid valve circuits.



(a)

(b)

FIGURE 9

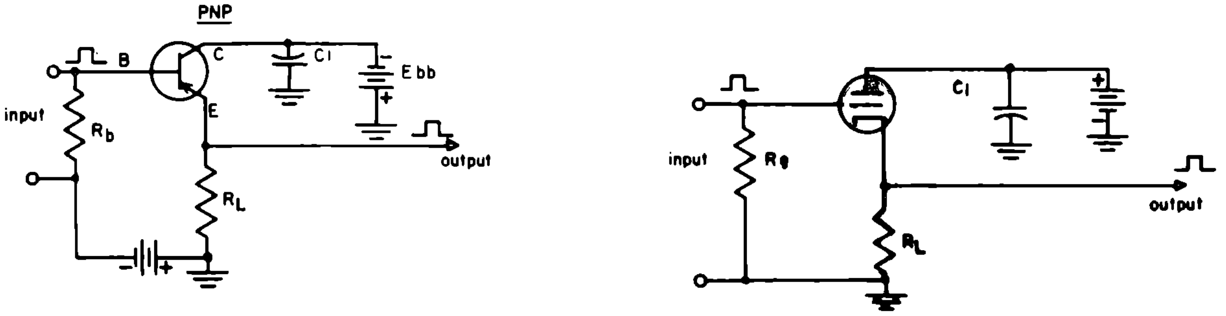


FIGURE 10

METHODS OF COUPLING

The basic methods of coupling transistor stages are similar to those used in valve circuits. The major difference lies in the fact that the input and output resistance of transistors as compared to valves vary widely. These resistances depend on the type of transistor used and the operating conditions. Also a change in input or output load resistance reflects into the input or output, whichever the case may be. For example, as the load resistance increases, the input resistance decreases. This is not generally true with valves since changes in the plate load do not normally reflect into the grid circuit. In some cases, however, plate to grid capacitance becomes a factor and plate loading will affect the input impedance of a valve circuit.

The coupling requirements for transistors can be met by various methods, such as, transformer, resistance - capacitance, impedance and direct coupling. Each of these methods will be discussed.

Transformer Coupling

The method of transformer coupling transistor stages is illustrated in Figure 11. As can be seen, this grounded emitter circuit employs fixed and self bias and an emitter resistor (R_e) for stabilization. The biggest advantage of this circuit is that the input and output impedance of the transistor can be matched for maximum power gain. A step down transformer, T_1, is used from the collector of the preceding stage to the base of the following grounded emitter stage as shown in Figure 11. Due to this step down it would seem that a voltage loss appears across the secondary of T_1 and would defeat our purpose. However, it must be remembered that a transistor is a **current operated** device, not a voltage operated device such as the valve. This step down provides best power transfer and the change in base current, due to the presence of the signal, activates transistor action and a power gain can be measured across the primary of T_2. This step down can be compared to the output stage of an audio amplifier, where a step down transformer is required to drive a loudspeaker, which is a current operated device. The purpose of the transformer is usually for maximum power transfer.

The circuit components include a voltage divider (R_1 and R_2) which provides the proper bias. The voltage divider is bypassed by C_1 to avoid signal attenuation. The emitter resistor (R_e) is the stabilizing resistor which allows variations of the transistor and circuit elements to be absorbed automatically without adverse effects. This resistor (R_e) is bypassed by C_2 prevent loss of gain due to degeneration. The supply voltage E_{bb} is also bypassed to prevent feedback due to possible ac signal voltages being developed across the power supply. Capacitors C_1 and C_2 may be replaced by a single capacitor connected between the emitter and the bottom of the secondary of T_1.

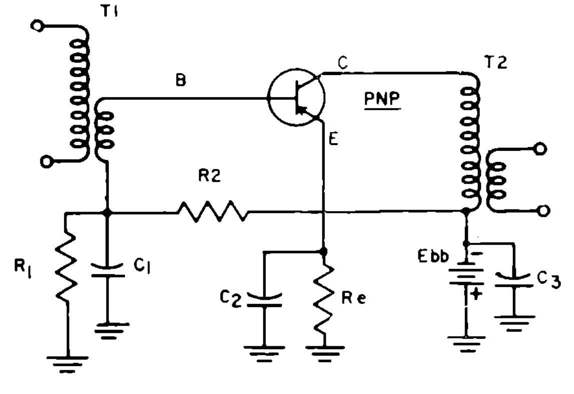


FIGURE 11

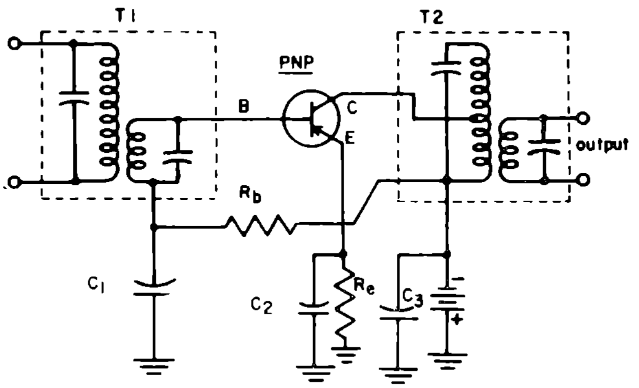


FIGURE 12

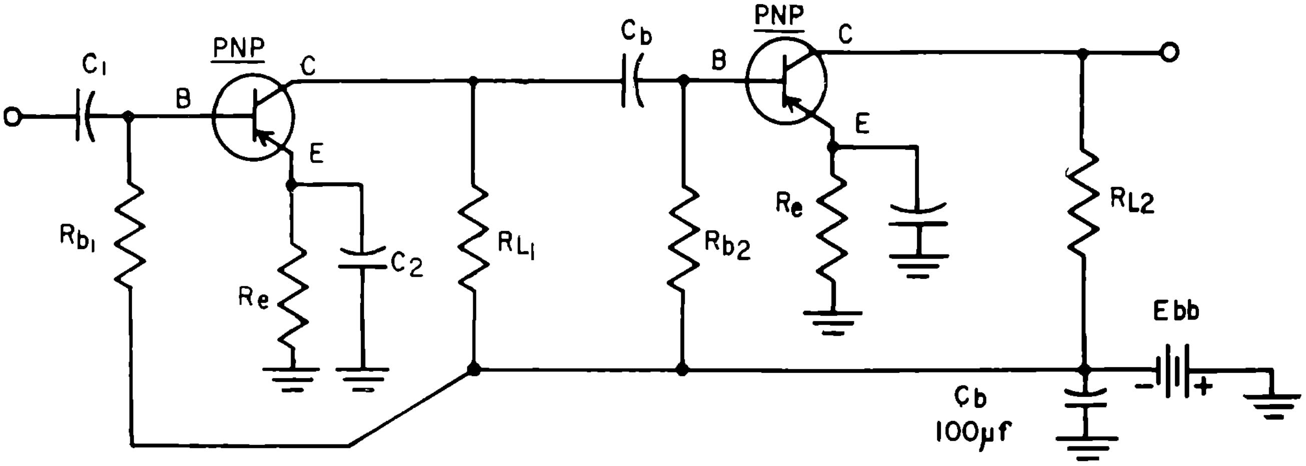


FIGURE 13

Transformer coupling in if stages can present problems in design. Basically, the if transformer is used to select the desired signal frequency and has a band width capable of passing the sidebands of the desired signal frequency but rejects adjacent signals. This means the Q of the if transformer must be carefully specified and maintained. In the case of valve circuits the output resistance of an amplifier is sufficiently high that adjusting the transformer to have the proper Q is no problem. If the output resistance is low (100K or under) the Q of the transformer is lowered, possibly causing poor selectivity. This is the case when using transistor amplifiers. A typical output resistance of a transistor radio if stage (455Kc) is approximately 25K ohms. If the collector is shunted across the entire primary of transformer T_2, as shown in Figure 11, the selectivity will be seriously affected due to the output resistance of the transistor reducing the Q of T_2. To overcome this problem, the circuit is connected as shown in Figure 12. By selecting the proper tap on T_2 for the collector connection, the Q of the circuit can be satisfied as well as properly matching the output resistance of the transistor.

An exact match is not always desirable since feedback is also a major problem in transformer design. In some cases neutralization is incorporated as was done in old triode if amplifiers. In other circuits the primary is not tapped for the collector connection. By selecting the proper L/C ratio the correct Q can be maintained, however, neutralization may be necessary. Also some circuits use a centre tapped secondary in order to balance the feedback currents. In this case no neutralization is necessary.

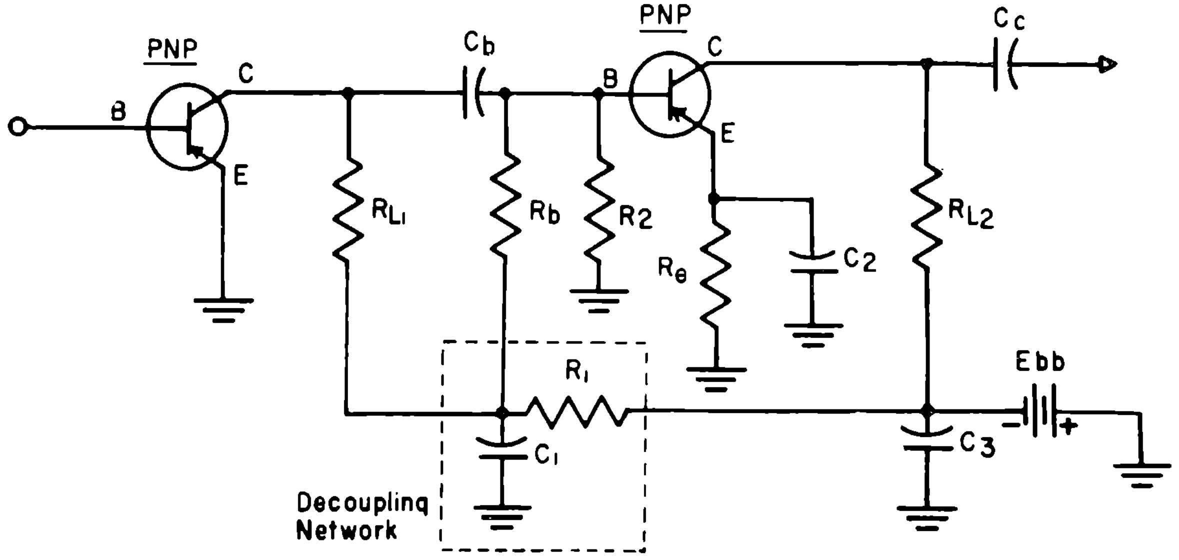


FIGURE 14

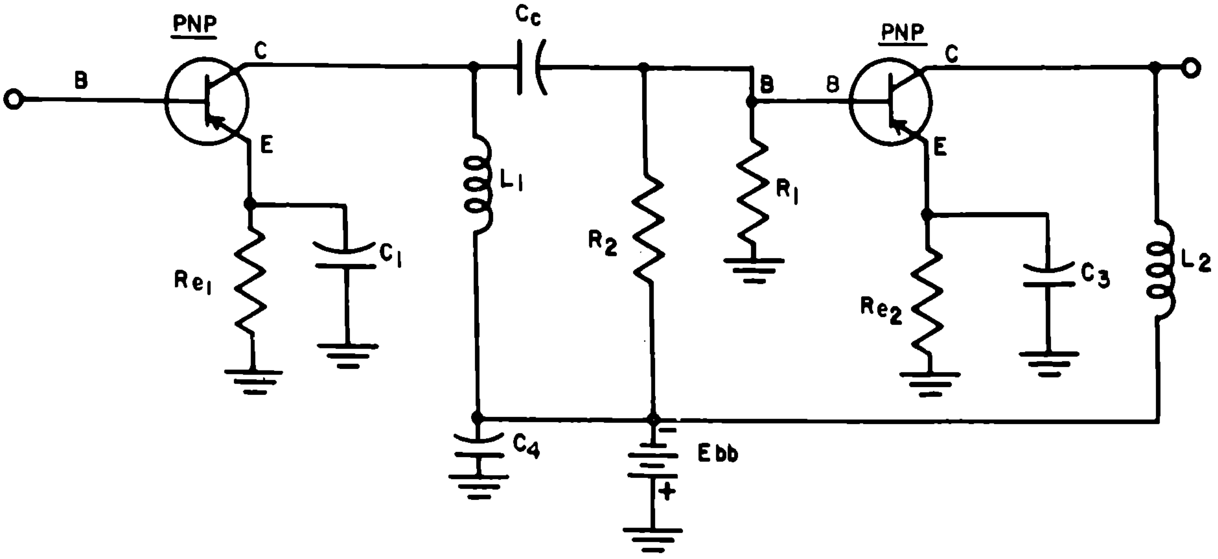


FIGURE 15

Resistance-Capacitance Coupling

Transformer coupling can be used to advantage in if strips or any rf application where selectivity is required. They are also used in large signal audio circuits such as drivers and output stages. However, R/C coupling is desirable where low level audio signals are involved since transformers are more susceptible to hum pick-up and also take up space. Figure 13 illustrates a two stage R/C coupled circuit. The method of bias is similar to that used in transformer coupling. The major additions are R_L (collector load) and C_c (coupling capacitor). The coupling capacitor (C_c) must be made very large (2 to 10 μf) due to the small output and input resistances involved.

It should be noted that electrolytic capacitors are used for coupling whereas they are not so used in valve circuits. Therefore, polarity should be observed or damage to the capacitors and possibly to the transistor may occur. Leakage current is not as critical in transistor circuits as in valve circuits.

When cascading R/C coupled amplifiers as shown in Figure 13, it is necessary to decouple one or more stages in order to prevent feedback. One method of de-coupling is illustrated in Figure 14. This is accomplished by inserting resistor R_1 in series with the base resistor and bypassing R_1 by means of capacitor C_1. The R_1C_1 time constant should be adjusted to insure that the lowest frequency to be amplified is adequately

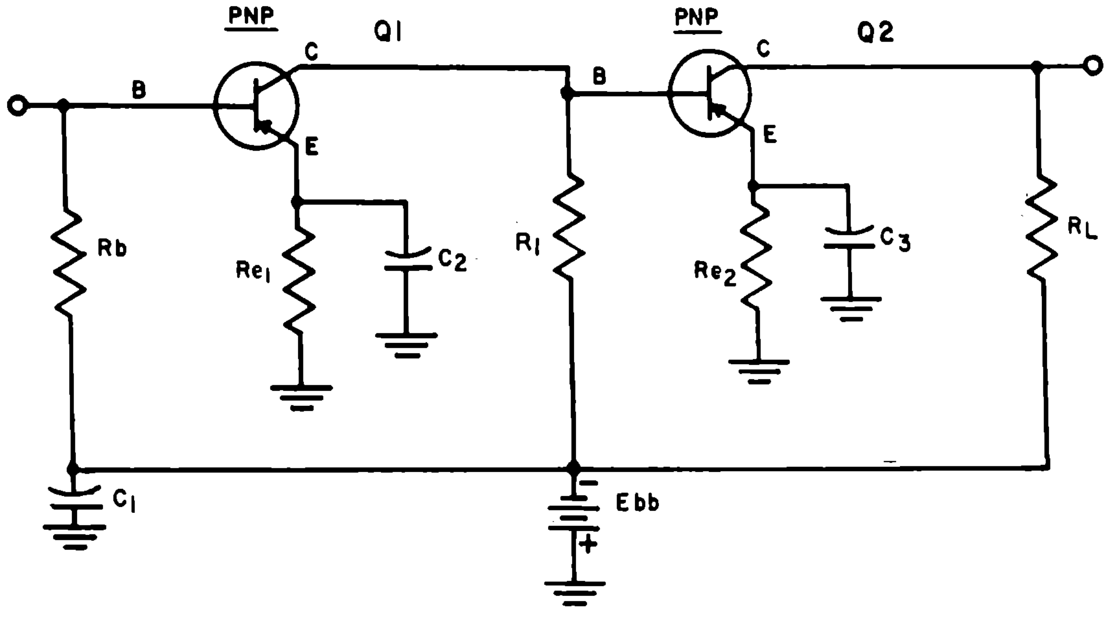


FIGURE 16

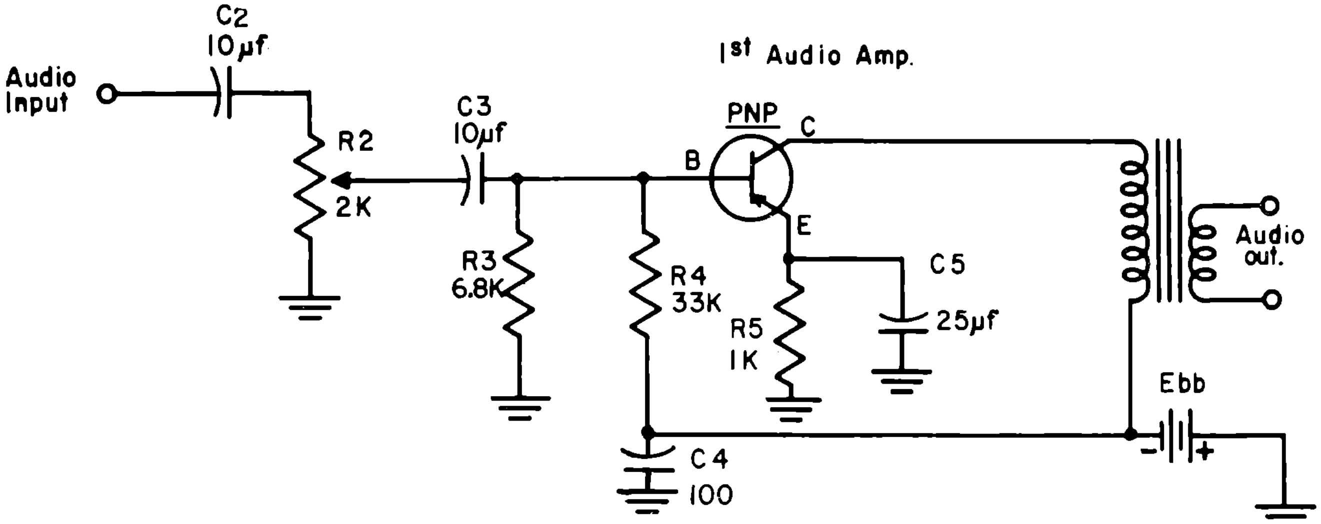


FIGURE 17

bypassed. Generally, the value of R_1 must be kept small so that the supply voltage is not drastically reduced to the previous stages. Therefore, the value of C_1 must be kept very large, usually 100 μf or larger.

Impedance Coupling

Impedance coupling is similar to R/C coupling. The major difference lies in the fact that inductances are used to replace the load resistors in Figure 5. This type of coupling may be used for circuit applications above audio frequencies. Resistors R_1 and R_2 are still necessary in order to provide the proper emitter-base bias.

Both series and shunt peaking can be accomplished in this type of coupling. Peaking coils similar to those associated with valve circuits in television receivers may be used in transistor video amplifiers.

Direct Coupling

Direct coupling is used generally where cost is a factor. However, it is also used where partial dc restoration is required in television receivers and where the dc component must be amplified. A direct coupled amplifier is shown in Figure 16. In cases where the dc component must be retained in whole the capacitors must be eliminated. As can be seen in Figure 16, resistor R_1 serves as both the collector load of Q_1 and the bias resistor of Q_2.

GAIN CONTROLS

As in valve circuits, volume controls must be used in transistor audio amplifiers to provide means for suitable audio level adjustment by the listener. Although these controls are generally associated with R/C coupled amplifiers, they are

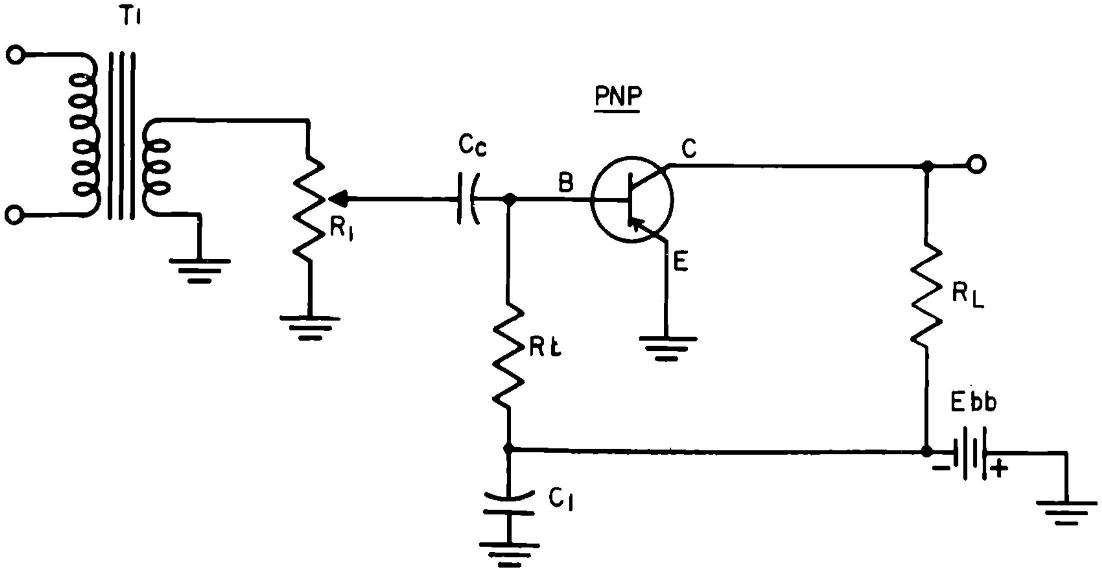


FIGURE 18

also used where transformer or direct coupled circuits are involved.

A volume control circuit is illustrated in Figure 17. The circuit includes the volume control and the 1st audio amplifier. The capacitor C_2 prevents the volume control R_2 from changing the dc operating point of the previous stage. Capacitor C_3 prevents the base current from flowing through the volume control. Resistors R_3 and R_4 provide the base bias. The base must be negative with respect to the emitter for p-n-p transistor operation. The coupling capacitors C_2 and C_3 must be large in value due to the low circuit impedances involved.

Where transformer coupling is used, a circuit as shown in Figure 18 may be used. The resistance of the volume control R_1 should be about 2 to 3 times the impedance of the secondary of the coupling transformer to prevent excessive loading.

Continued next month.

Some Aspects of Synchronization in TV Receivers

By J. VAN DER GOOT, M.I.R.E. (Aust.)

PART 2

HORIZONTAL OSCILLATORS AND AFC. SYSTEMS

Horizontal scanning oscillators are not usually directly synchronized. Directly synchronized horizontal oscillators are too vulnerable to noise. Firstly the duration of the noise pulse is often comparable with the horizontal pulse duration. Secondly the frequency at which the noise pulses occur is often high compared with the line — or horizontal frequency. Noise pulses that occur a relatively short time before the horizontal sync pulse would tend to trigger a directly synchronized oscillator as may be seen from Fig. 2. This is the reason why the frequency of the horizontal oscillator is controlled by some form of automatic frequency control (afc) system. The action of the afc system is often called "fly-wheel" action because of its analogy with a fly wheel.

Strictly the expression "automatic frequency control" is incorrect or at least incomplete, since, apart from frequency, the phase of the oscillator must also be controlled. The systems of frequency and phase control that will be described are of the phase comparator type. The phase relation between pulses derived from the oscillator output and the sync pulses is detected and the frequency of the oscillator is adjusted so that the phase relation between those pulses remains nearly constant. The two most commonly used horizontal oscillator systems are:

(1) The blocking oscillator with a type of phase comparator which is known as the "synchro-guide".

(2) The multivibrator with the phase discriminator.

The first part of this article appeared last month, and deals with Vertical Oscillators. The third and concluding part of the article will be presented next month, and deals with processing of the sync pulse.

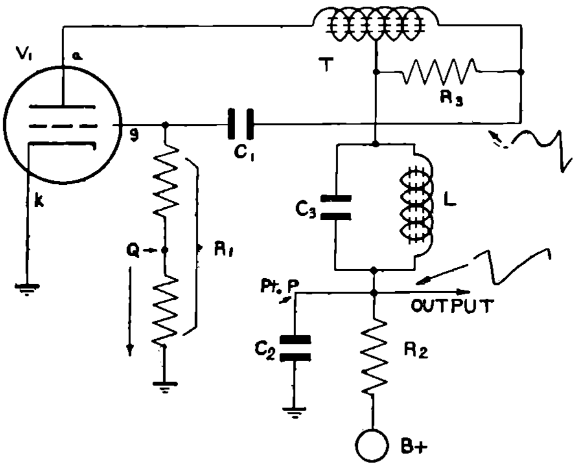


Figure 10.—Circuit diagram of a horizontal blocking oscillator.

The Horizontal Blocking Oscillator

Fig. 10 shows the circuit diagram of a horizontal blocking oscillator. The principle of the oscillator is similar to that of the vertical blocking oscillator. Feedback is accomplished by means of plate coupling. The oscillator period is partly determined by the time constant of R1 and C1. R3 is the resistor that damps the natural oscillation of T. Because the oscillator is indirectly synchronized the requirements for stability are very stringent. For this purpose a so-called sine wave coil is incorporated. L is called the "sine wave coil" because it superimposes a sine wave on the oscillator plate — and grid voltage waveforms. Fig. 11 (a) shows the oscillator grid voltage waveform. Fig. 11 (b) shows the general shape; for simplicity's sake the damped natural oscillations of the oscillator have been deleted. The short duration plate current pulse (Fig. 11 (c)) shocks LC3 into oscillation. LC3 is tuned to approximately the horizontal line frequency. Fig. 11 (d) shows the sine wave voltage which appears across L and C3. The plate current pulse discharges C2. Between plate current pulses C2 is charged again through R2 giving a positive going sawtooth waveform. The voltage at the tap of T is the resultant of the waveform of Fig. 11 (d) superimposed on that of Fig. 11 (e). The waveform is drawn in Fig. 11 (f). The waveform at the anode of V1 will have the same general shape as in Fig. 11 (f) but actually the voltage resulting from the natural oscillations of T is superimposed on it. The sinusoidal voltage of Fig. 11 (d) will also appear across R1 and will be superimposed on the blocking oscillator grid voltage waveform of Fig. 11 (b). Fig. 11 (g) shows the resulting general waveform. Actually the voltage resulting from the natural oscillations of T is superimposed on Fig. 11 (g) again. Fig. 12 shows an oscillo-

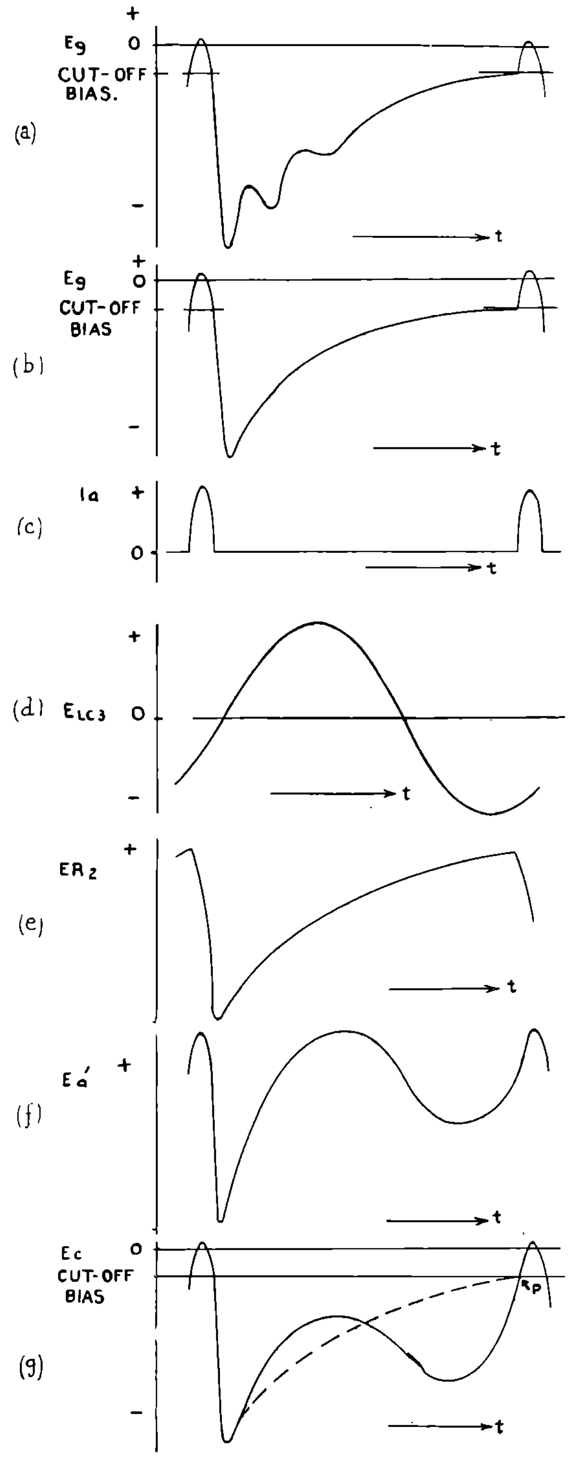


Figure 11.—Waveforms at the horizontal blocking oscillator.

- (a) and (b) Voltage at the grid (without sine wave).
- (c) plate current.
- (d) voltage across the sine wave coil.
- (e) voltage across C2 = output voltage.
- (f) voltage at tap of the transformer.
- (g) voltage at the grid (with sine wave).

gram of the grid voltage waveform. The advantage of this type of grid voltage waveform is that the approach to the grid cutoff potential is much faster than without the addition of the sine wave. The effect of variations in plate current cutoff grid potential, e.g., due to mains voltage variations, are thus very much reduced. Fig. 13 illustrates this clearly. The intersection of the grid voltage and the cutoff potential which is the point at which the oscillator starts another cycle (point P in Fig. 11 (g)) is drawn on a larger scale. The effect of a shift of cutoff bias on the oscillator period is $\delta t1$ without sine wave and $\delta t2$ with sine wave.

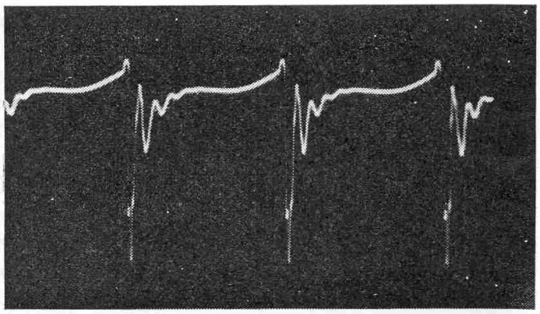


Figure 12.—Oscillogram of blocking oscillator grid voltage. Sweep: 20 μsec/cm.

The frequency of the oscillator is kept constant by controlling the current in the lower part of R1 which in turn controls the grid potential of V1. When the grid potential is made more negative, V1 will remain cut off for a longer time and the frequency of the oscillator is decreased. On the other hand when the grid potential is made less negative the oscillator frequency is increased. The manner in which this grid potential control can be obtained by the synchro-guide circuit is described next.

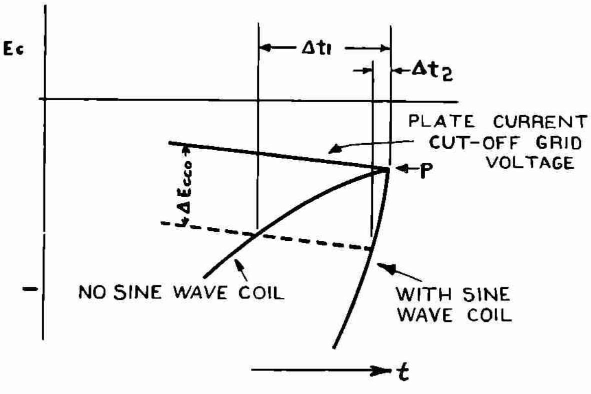


Figure 13.—Drawing illustrating the frequency stabilizing effect of the sine wave coil. Enlargement of intersection at P in Fig. 11 (g).

The Synchro-guide

Fig. 14 shows the circuit diagram of the synchro-guide. A pulse derived from the oscillator output, Fig. 11 (e), is applied to the grid of V2. The grid bias is such that the positive peak of the pulse is slightly above plate current cutoff grid potential, Fig. 15 (a). At the same time horizontal sync pulses are also applied to the grid, Fig. 15 (b). Now assume that the horizontal sync pulses occur at the same time as the steep trailing edges of the feedback pulses. Then the grid voltage waveform of V2 will be as shown in Fig. 15 (c). During the hatched portions plate and cathode current will flow in V2. Capacitor C1 and subsequently C2 are charged. Between pulses C1 is discharged through R4 and R1', and C2 is discharged through R5, R4 and R1'. The charging time constant of C1 is comparable with the pulse duration. The discharge time constants of C1 and C2 are of the order of several line durations. The result is that the cathode, k, is maintained at a positive potential which depends mainly on the height of the hatched pulses in Fig. 15 (c). The shape of these pulses will also affect the cathode voltage but not to the same extent as the height. Fig. 15 (d) shows the voltage waveform at the cathode of V2. Resistor R1' is common to the hori-

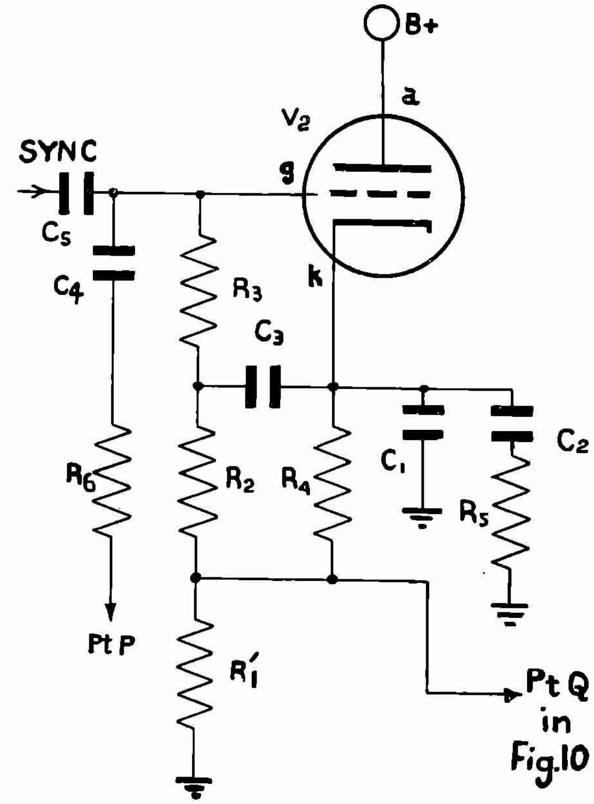


Figure 14.—Circuit diagram of the synchro-guide.

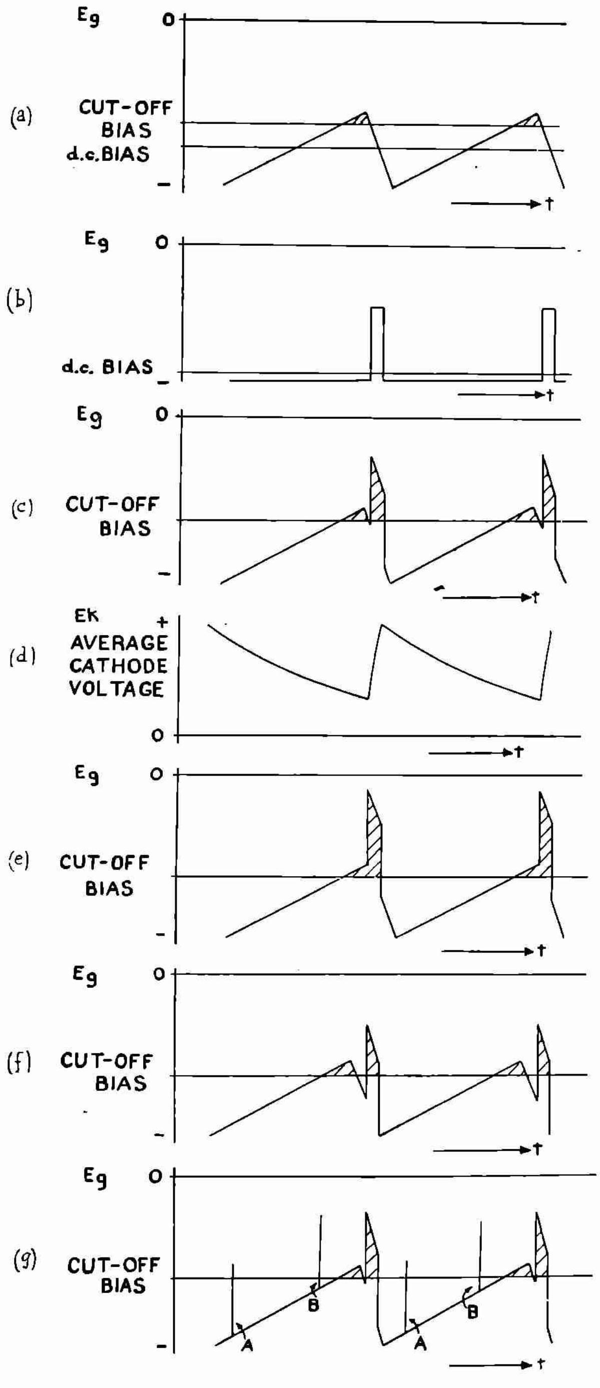


Figure 15.—Voltage waveforms of the synchro-guide.
(a) Grid; feed back pulses;
(b) grid; sync pulses;
(c) grid with (a) superimposed on (b) showing the phase relation between sync and feed back pulses in the centre of the frequency control range;
(d) cathode voltage waveform;
(e) as (c) but showing phase relation at lower end of the frequency control range;
(f) as (c) but showing phase relation at higher end of the frequency control range;
(g) as (c) with noise pulses.

zontal blocking oscillator grid leak, the lower part of R1 in Fig. 10.

If the frequency of the oscillator tends to decrease, the relative phase of sync and oscillator pulses will shift and the height of the hatched pulses will increase as shown in Fig. 15 (e), therefore the cathode voltage will increase. The current through R1' in Fig. 14 will also increase. As was explained in the previous section an increase in current in the common resistor (the lower part of R1 in Fig. 10) in the indicated direction increases the frequency of the oscillator. On the other hand if the frequency of the oscillator tends to increase the height of the hatched pulses will decrease, Fig. 15 (f). The oscillator grid potential will become more negative and the frequency of the oscillator will decrease. This system of feedback therefore tends to keep the oscillator feedback pulses and the horizontal sync pulses in a constant phase relation. Fig. 16 shows an actual oscillogram of the grid voltage waveform. Differentiated sync pulses are superimposed on the feedback pulses.

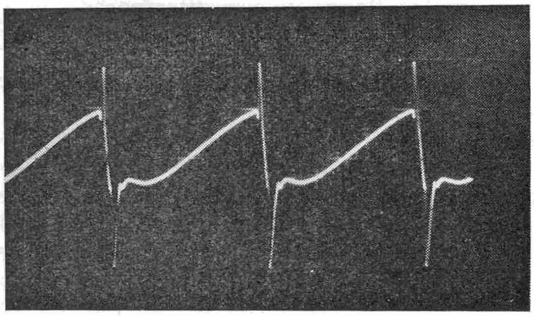


Figure 16.—Oscillogram of control valve grid voltage with differentiated sync pulses (see Fig. 15 (c)).
Sweep: 20 μsec/cm.

Fig. 15 (g) shows the effect of noise pulses. Here the assumption is made that the amplitude of the noise pulses is never higher than the amplitude of the sync pulses. This can be achieved by sync clipping and limiting, as will be described in later sections. Noise pulses of type A do not contribute to the synchro-guide cathode voltage whereas pulses of type B do. However, the overall effect of a large number of noise pulses occurring at random is drastically reduced. It is of course desirable that the height and shape of the horizontal sync pulses remain constant during the entire scanning field if the oscillator frequency is to remain constant. This requirement presents problems which will be dealt with in later sections.

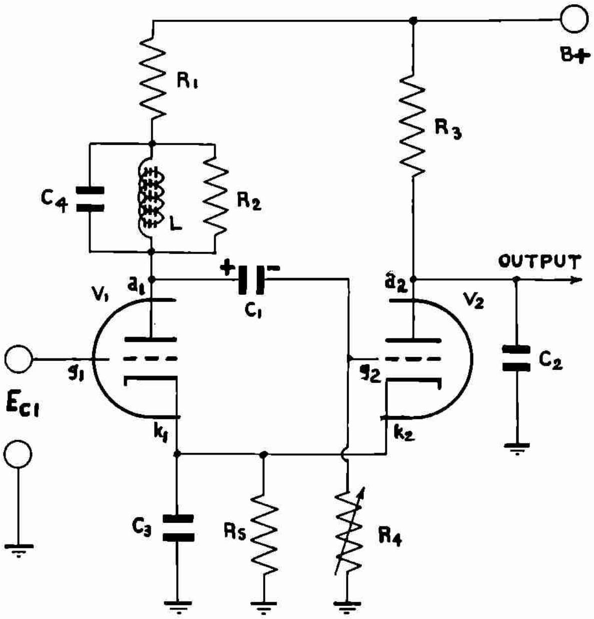


Figure 17.—Circuit diagram of a cathode coupled multivibrator.

A fraction of the oscillator grid voltage waveform, Fig. 12, will appear across R1'. R2 and C3 serve as a smoothing filter to prevent this waveform appearing at the grid of V2.

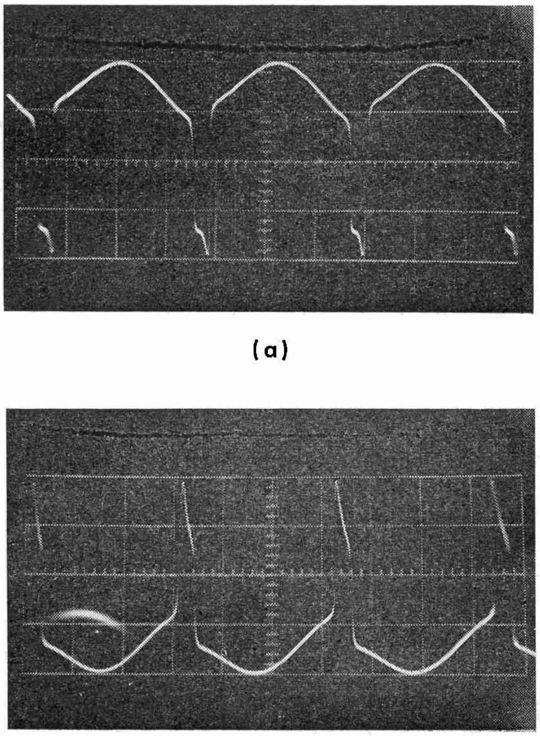
The Horizontal Multivibrator

The most common type of horizontal multivibrator is the cathode coupled type. Fig. 17 shows the circuit diagram of such an oscillator. The principle action of the cathode coupled multivibrator is that of a double switch. The two switches consist of valves. One valve switches off the second one. The second valve remains cutoff for a period which is determined by circuit constants. After this period has elapsed the second valve switches off the first one, which again switches off the second valve. And so the whole cycle repeats itself. Usually the two valves are combined in the one envelope. Often double tri odes are used.

The following is a more detailed description of the circuit of Fig. 17. Assume that C1 is charged with a polarity as indicated. V1 is conducting and V2 is cut off. C1 discharges slowly through R4, R5, and V1. The discharge time constant is mainly determined by the values of C1 and R4. When the grid potential of V2 reaches plate current cut-off, plate and cathode currents start to flow in V2. The voltage at k2 and k1 therefore increases. The plate current through V1 decreases and the voltage at a1 rises. This causes the voltage at g2 to rise. The result of this positive feedback is that V2 is rapidly driven from cutoff to grid conduction. At the same time the voltage at a1 has also risen rapidly. V1 will be nearly cut off. As soon as grid current flows in V2, C1 charges very quickly through R1, R2, V2 and R5. The charging time constant is mainly determined by the values of R1 and C1 (R1 is much larger than R5).

While V1 is conducting and V2 is cut off, C2 is charged through R3. While V2 conducts, C2 is rapidly discharged through V2 and R5. As C2 loses its charge the discharging current decreases and therefore the potential at k2 and k1 decreases. The plate current in V1 increases and the voltage at a1 decreases. This causes the potential at g2 to drop and causes a further decrease of cathode current in V2. The result of this positive feedback is that the potential at a1 drops sharply and g2 is rapidly driven negative. V2 remains cut off until C1 is discharged to the point where plate current starts to flow in V2. Then the whole sequence is repeated again.

L and C4 are tuned to approximately the oscillator frequency. When V1 is driven into conduction L and C4 are shocked into oscillation. The result is that a sinusoidal voltage is superimposed on the a1 and g2 voltage waveforms. The advantage of the stabilizing coil, as L is generally called, is the same as the sine wave coil in the blocking oscillator. It causes the g2 voltage to



(a)

(b)

Figure 18.—Oscillograms of voltage waveforms for the multivibrator shown in Fig. 17. (a) At a point a1; (b) at point g2.
Sweep: 20 μsec/cm.

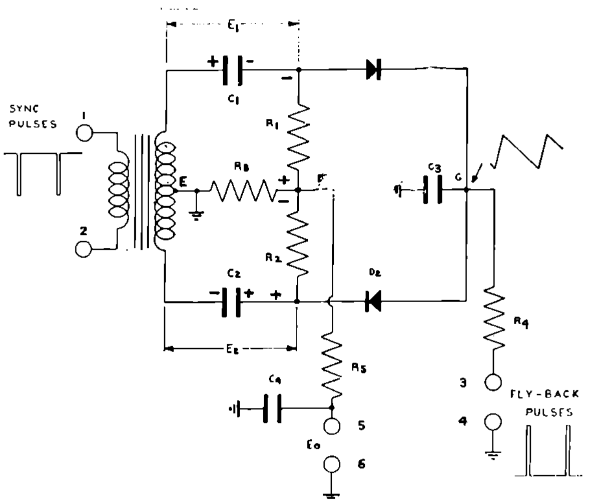


Figure 19.—Circuit diagram of horizontal phase discriminator.

approach the plate current cut off grid potential more rapidly. Fig. 18 (a) shows an oscillogram of the waveform at a1, Fig. 18 (b) shows that at g2. The output waveform is similar to that of Fig. 11 (e).

As the time during which V2 is cut off is greatly influenced by the value of R4, the latter is often made variable and serves then as a horizontal hold control. The frequency can also be controlled by the voltage of g1 (Ec1 in Fig. 17). When V2 conducts, V1 is driven towards cut off. The amplitude of the plate voltage pulse of V1 (Fig. 18 (a)) is therefore determined by the standing plate current of V1. The larger the V1 plate current the larger the plate voltage swing and the more g2 will be driven negative. The more g2 is driven negative the longer it will take for C1 to discharge and the longer the oscillator period will be. On the other hand the smaller the plate current of V1 the shorter the period of the oscillator will be. Making the grid of V1 more positive will therefore result in a decrease of oscillator frequency and making g1 more negative will result in an increase of oscillator frequency. This is how automatic phase and frequency control of the cathode coupled multivibrator can be obtained. The most common type of control circuit used in conjunction with the cathode coupled multivibrator is the phase discriminator. One type of horizontal phase discriminator and its principle of operation will be discussed in the next section.

The Phase Discriminator

Fig. 19 shows the circuit diagram of one type of phase discriminator. The sync pulses are applied to the primary of the transformer. The secondary is centre tapped. First assume that G is at ground potential. The pulses appear across the two halves of the secondary with opposite polarity.

During the pulses C1 and C2 will be charged with a polarity as indicated. Between pulses C1 and C2 will discharge slowly through R1 and R2. The potential of F will be the same as that of G, i.e., ground potential. Suppose the potential of G to ground is made positive, C1 will then be charged

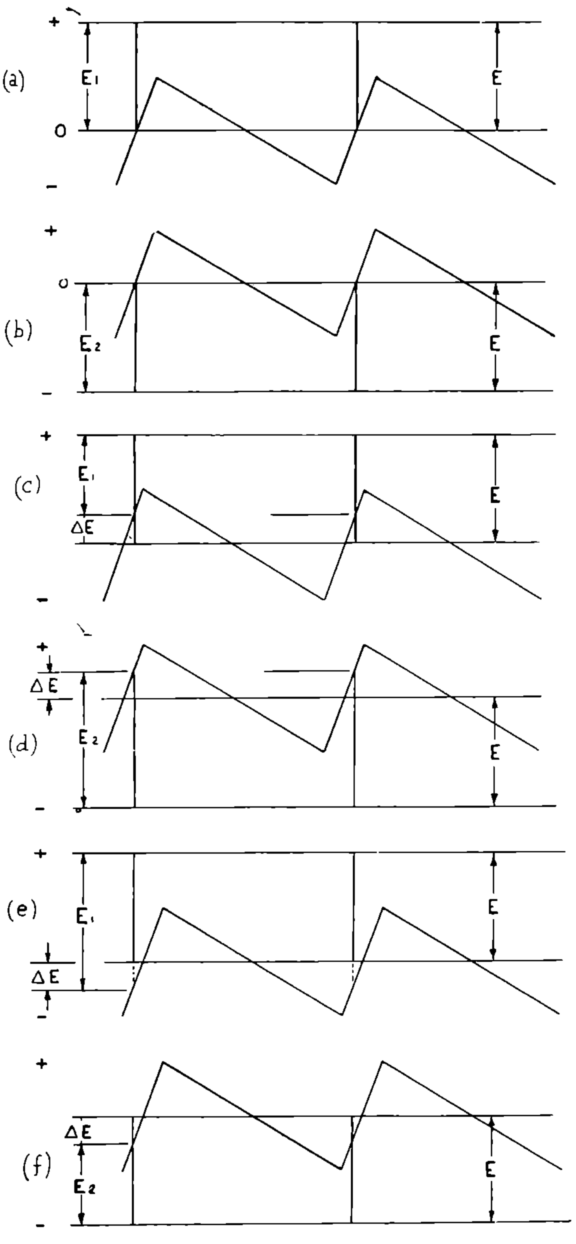


Figure 20.—Voltage waveforms for the phase discriminator (Fig. 19).

(a), (c) and (e) at D1.
(b), (d) and (f) at D2.
(a) and (b) free running period = sync pulse period.
(c) and (d) free running period < sync pulse period.
(e) and (f) free running period > sync pulse period.

to a lower potential and C2 to a higher potential. Point F will assume a positive polarity with respect to ground.

Now the polarity of G is made to vary in a sawtooth fashion. The sawtooth can be derived from the horizontal fly-back pulses. The time constant of R4 and C3 is long compared with the duration of a fly-back pulse. During the pulse C3 is charged to an amplitude which is only a fraction of the amplitude of the fly-back pulses; thus the voltage across C3, which is the potential of point G with respect to ground, rises nearly linearly.

If the polarity to ground of G is zero at the time a horizontal sync pulse occurs, Fig. 20 (a) and (b), then $E1 = E2 = E$, where E is the pulse amplitude. The potential of F to ground is then zero. If the potential of G to ground is positive when a horizontal sync pulse occurs, Fig. 20 (c) and (d), then $E1 = E - \delta E$ and $E2 = E + \delta E$. The potential of F with respect to ground is then $+ 2\delta E$. If the potential of G to ground is negative when a horizontal sync pulse occurs, Figs. 20 (d) and (e), then $E1 = E + \delta E$ and $E2 = E - \delta E$. The potential of F to ground is then $- 2\delta E$. The output of the phase discriminator is connected to the control grid of the first valve of the multivibrator, Fig. 17. If the frequency and phase of the multivibrator are such as shown in Fig. 20 (a) and (b), g1 is at zero potential. Now if the frequency of the multivibrator tends to increase a phase relation will exist as shown in Fig. 20 (c) and (d). The potential of F will become positive. The positive potential of g1 will decrease the frequency of the multivibrator as described in the previous section. In the other case where the frequency tends to decrease, Figs. 20 (e) and (f), the potential of g1 will become negative and the frequency will increase. This negative feed-back maintains an equilibrium which keeps the phase relation between the sync pulses and feedback pulses nearly constant. Since between pulses C1 and C2 discharge a small amount, E1 and E2 are not pure dc voltages. Neither will any voltage appearing at F be pure dc and so R5 and C4 are used as a smoothing filter.

The effect of noise is similar to that described in the discussion of the synchro-guide circuit.

Instead of the transformer in Fig. 19, a phase splitter in the form of a valve can be used. The valve can then also be used as a pulse clamp and limiter. Such a "sync splitter" is described later.

Although both the blocking oscillator-synchro guide and the multivibrator phase discriminator have a fly-wheel action which removes the necessity for individual sync pulses to initiate all the line traces, the absence of sync pulses during several line scans will affect the output voltage of the synchro-guide or phase discriminator. This in turn will affect the phase of the blocking oscillator or multivibrator. Similarly a change in pulse height and/or pulse shape during several line durations will affect the phase of the oscillator and consequently affect the picture. The next section deals with these considerations in more detail.

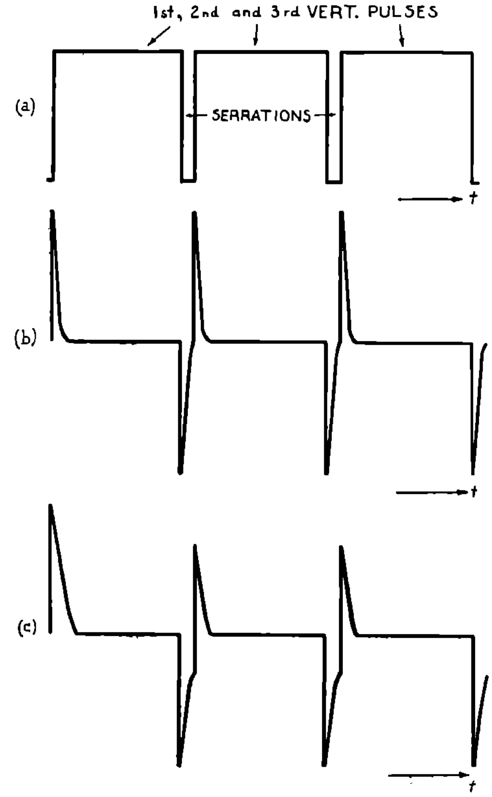


Figure 21.—Voltage waveform of three vertical sync pulses.

(a) Input;
(b) output; short time constant;
(c) output; time constant too long.

Necessity for Differentiation

From the previous sections it will be clear that the free running frequency of the horizontal blocking oscillator with synchro-guide afc control will be lower than the horizontal or line frequency of the received composite video signal. The synchro-guide adjusts the grid bias of the oscillator in a positive direction so as to increase the oscillator frequency until it keeps in step with the sync pulses. The free running frequency of the horizontal multivibrator with phase discriminator afc control can be either higher or lower than the frequency of the sync pulses. The phase discriminator adjusts the bias of the first valve of the multivibrator either in a positive or a negative direction and decreases or increases the frequency of the multivibrator until it keeps in step with the sync pulses.

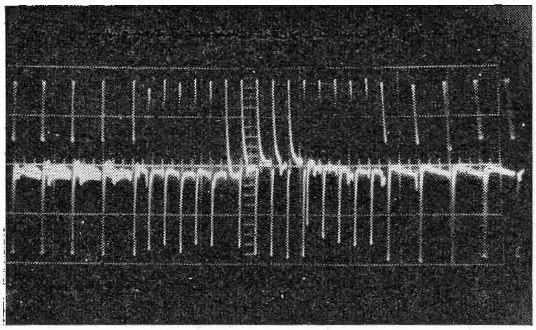


Figure 22.—Oscillogram of differentiated sync pulses (time constant: 1 μsec), showing end of odd field, pre-equalizing pulse train, vertical sync block, post-equalizing pulse train and start of even field.
Sweep: 100 μsec/cm.

If the energy content of the horizontal sync pulses is reduced for several line durations the frequency and phase of the horizontal oscillator will tend to change towards its free running frequency. The rate of change will depend on the time constants of the afc system. One of the most frequent causes of changing energy content of the horizontal sync pulses stems from the sync clipper as will be explained later when dealing with the sync clipper. Another cause is insufficient differentiation of the sync information. The sync information as supplied by the transmitter is such that the leading edges of all the pulses are separated by exactly one line duration (64 μsec) during most of the field and exactly half a line duration (32 μsec) during the pre-equalizing pulse train, the vertical sync block and the post-equalizing pulse train. If pulses of equal energy content are to be obtained from the sync information it is necessary for the input voltage level of the synchro-guide or phase discriminator to be the same at the start of each pulse. This calls for a differentiating time constant of not more than 1 μsec (see appendix). This may be obtained from a simple CR differentiating network. In Fig. 21 (a) the first three vertical pulses of a vertical sync block have been drawn. When this signal is applied to a differentiating network with a time constant of 1 μsec, the resultant output voltage will be as shown in Fig. 21 (b). If the time constant is longer than 1 μsec, positive going pulses derived from the second vertical pulse on, will be reduced in energy content as shown in

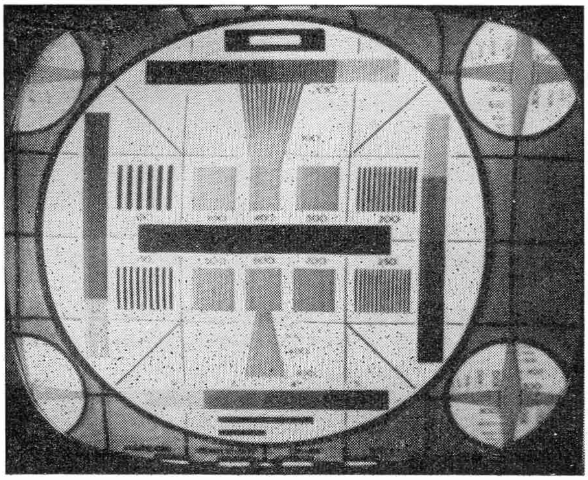
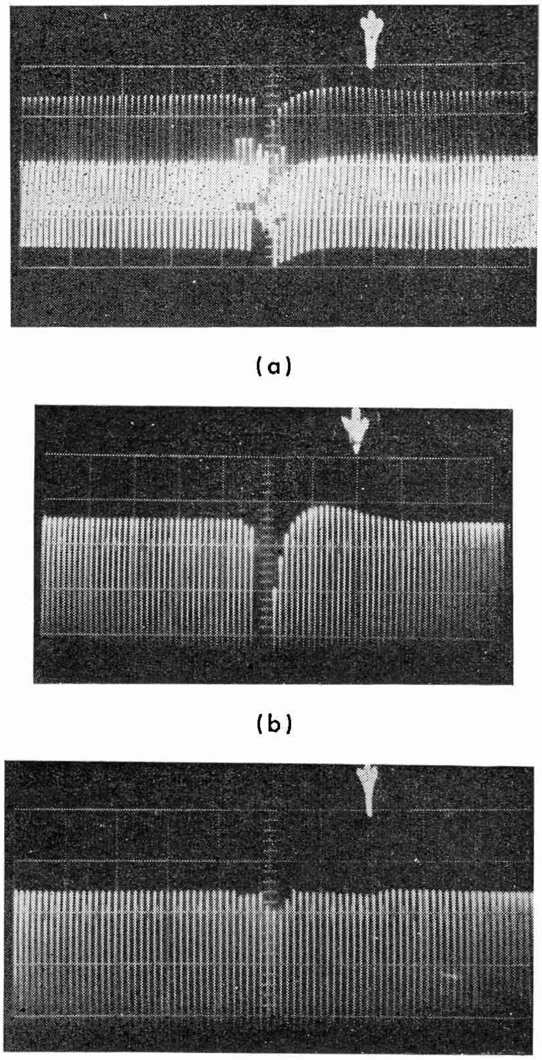


Figure 23.—Photograph of picture tube showing "bend" towards the left of the vertical lines at the top of the picture.



(a)

(b)

(c)

Figure 24.—Oscillograms of the waveform at the grid of the synchro-guide control valve.
(a) Sync pulse differentiating time constant $\cong$ 25 μsec.
(b) As (a) but showing the top of the waveform on a larger scale.
(c) As (b) but with a time constant $\cong$ 1 μsec. Arrows indicate the end of the field blanking.
Sweep: 500 μsec/cm.

Fig. 21 (c). Fig. 22 is an actual oscillogram of part of a field of pulses derived from a differentiating network with a time constant of approximately 1 µsec.

If the energy content of the differentiated pulses (positive going) is reduced during the vertical sync block the time taken for the oscillator to recover its original phase and frequency, depends entirely on the time constants of the afc system. If that time extends past the end of the field blanking period the result will be distortion at the top of the picture as shown in Fig. 23. This photograph was taken using a network with a time constant of approximately 25 µsec which provides insufficient differentiation. Fig. 24 shows oscillograms of waveforms at the grid of the synchro-guide control valve. Fig. 24 (a) was taken with a differentiation time constant of approximately 25 µsec. Consequently the sync pulses derived from the vertical sync block are very much reduced in amplitude and energy content. Fig. 24 (b) shows the same waveform as in Fig. 24 (a) but it shows the sync pulse tips only, on a larger scale. When the time constant is reduced to 1 µsec the sync pulse amplitude variation becomes very small, Fig. 24 (c).

To obtain a differentiation time constant of 1 µsec and sufficient output amplitude with a simple circuit is rather difficult in practice. The positive peak amplitude of the sync pulses required at the input of the synchro-guide or phase discriminator is of the order of 20 volts. In Fig. 25 an equivalent circuit is drawn of the source of the synchronization waveform, g1, with its equivalent impedance (R1 and C1), the differentiating coupling capacitor, C2, the equivalent input impedance of the synchro-guide (R2 and C3) and the source of the sawtooth feedback pulses, G2 with its equivalent series resistance R3. In practice C1 and C3 are of the same order of magnitude, 10 — 15 µµf. C1 is the output capacitance of the sync amplifier and C3 is the input capacitance of the control valve, each with stray circuit capacitances. The maximum amplitude of the sync pulses across R2 is obtained when R2 is much larger than R1 and when C2 is much larger than C3. On the other hand the time constant of C2, C3 and R2 must be approximately 1 µsec. The amplitude of horizontal feedback pulses in the vertical circuits must be kept down to a minimum to obtain good interlace. In most cases R1 is common to the vertical sync source, therefore R1 must be smaller than R3 so that a minimum amplitude of feedback pulses will appear across R1. All these conflicting requirements are difficult to meet in a circuit as represented in Fig. 25.

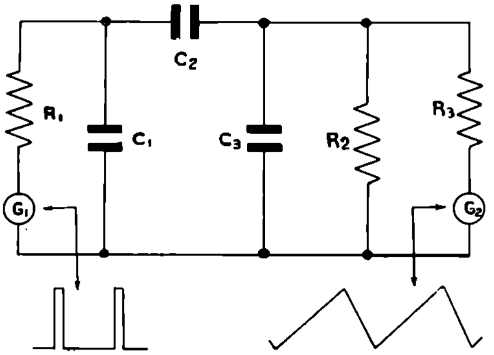


Figure 25.—Equivalent circuit diagram of a differentiating network with a valve as sync source and the synchro-guide control valve as a load.

One solution to the problem is to differentiate the sync information at the output of the sync clipper (see later discussion on the sync clipper) then amplify the differentiated pulses before applying them to the horizontal control valve. A circuit which embodies this principle is described at the end of the article. But first sync amplification and clipping will be discussed.

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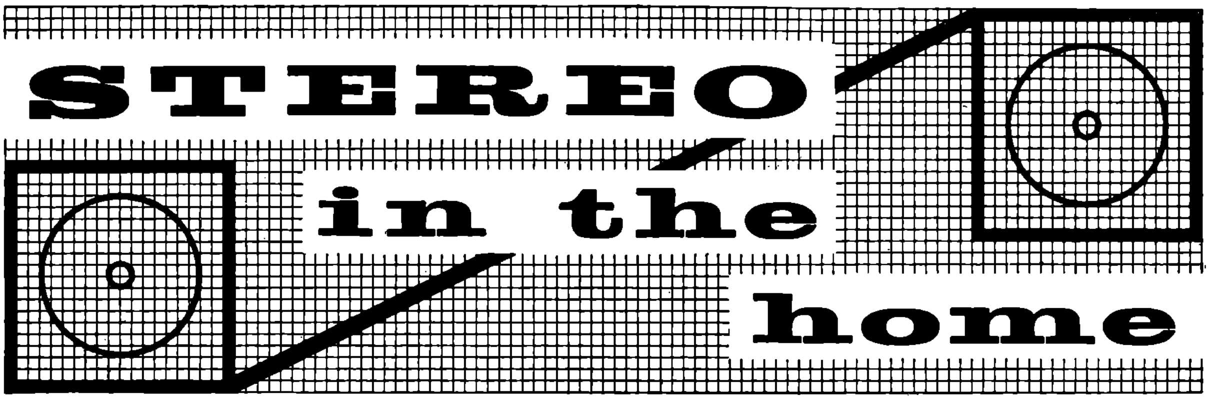

STEREO in the home

By HARRY F. OLSON

Part 2

LOCALIZATION TESTS[7]

From the standpoint of the listener, in any true stereophonic sound reproducing system there should be a subjective effect of a distribution of the reproduced sound sources in lateral directions as well as in depth. It is the purpose of this section to describe a series of experiments that demonstrate stereophonic effects in sound reproduced by means of a two-channel system.

A. Lateral Localization

There are two factors that determine angular or lateral localization of a source of sound, namely, phase and intensity. An experiment which demonstrates angular or lateral localization with regard to phase and intensity is shown in Fig. 7. The same signal is reproduced from loudspeakers 1 and 2. The signal from loudspeaker 1 can be delayed by means of the delay system. For each value of delay, the ratio of the voltage input to the two loudspeakers is varied until it is impossible to distinguish which loudspeaker appears to be the source. The results of this test are shown by the graph of Fig. 7. This experiment shows that there can be considerable unbalance in amplitude before the sound ceases to appear to come from the undelayed source, and therefore, that both phase and intensity play a part in angular localization.

The experiment of Fig. 7 was carried further to determine the exact subjective location of the sound.

In performing the lateral localization experiments, some precautions must be taken to avoid errors in observation. For example, if the observer can see the loudspeakers he tends to point to either one or the other as a source. This is a natural state of affairs because there are two visible sources of sound, and, obviously, the sound must come from one or the other. In order to avoid this subjective bias, the loudspeakers were located behind a light - opaque, sound - translucent screen. The arrangement[8] is shown in Fig. 8.

When the intensity and the phase of the sound emanating from the two loudspeakers is the same, the sound appears to originate from a point midway between the two loudspeakers, designated as S1'. If the intensity of the sound emanating from both loudspeakers is the same but the phase of loudspeaker A' of Fig. 8 is delayed 2 msec with respect to loudspeaker B', the sound appears to come from the point S4' of Fig. 8. If the delay is increased to 5 msec, the sound appears to come from point S3' of Fig. 8. If the intensity of A' is made 5 db higher than B', and A' is delayed 5 msec with respect to B', the sound appears to come from point S4'. If the intensity of B' is made 5 db higher than A' but the phase of the two is the same, the sound appears to come from point S3'. If the intensity of B' is made 2 db higher than A', but the phase is the same, the sound appears to come from point S4'. If the intensity of B' is 5 db higher than A' and the sound from loudspeaker A' is delayed about 5 msec, the sound appears to come from point S2'. This experiment shows that the apparent position of the reproduced sound can be shifted over wide limits by varying the relative phases and/or relative intensities of the sound sources.

As an extension of the above experiments, the stereophonic arrangement of Fig. 9 was employed. The configuration of the microphones and loudspeakers is shown in Fig. 9. A person speaking was located at the different positions S1, S2, S3, S4, and S5 in the free-field room. The corresponding

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[7] The localization tests were performed over a period of time extending from 1949 to 1955.

[8] In the experiments depicted in Figs. 8, 9, 10, and 11, the listener was asked to locate the apparent lateral location of the reproduced source of sound at the surface of the curtain. In other words, the apparent location of the source of sound with respect to depth was not determined. In the section on depth localization, experiments relating to the apparent location of the source of sound with respect to depth will be described.

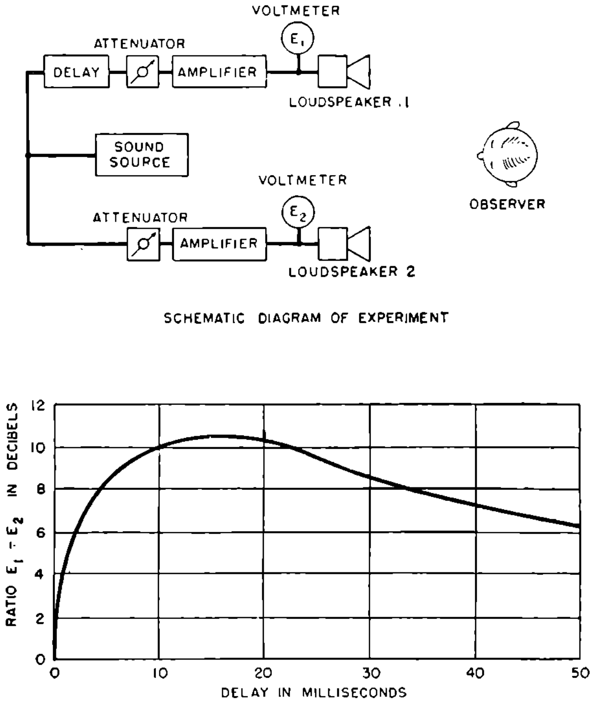

Fig. 7.—Schematic arrangement of the apparatus which illustrates the effect of phase delay and amplitude upon the localization of a sound source. The graph shows the ratio of the voltages E1 and E2 as a function of the delay for the condition when it is impossible to distinguish which loudspeaker is the source.

apparent locations of the reproduced sound in the listening room are shown as S1', S2', S3', S4', and S5'. The relative estimated sound levels in db are also shown, with positive S1' arbitrarily designated as the 0 db reference level. This experiment shows that the apparent position of the reproduced sound follows the actual position of the sound in the studio with small deviations. The deviations are as follows: The reproduced sound tends to be spread out in middle positions. The overall spread of the sound sources is less than the distance between the loudspeakers. The reproduced sound level appears to be lower in the middle positions.

In the next experiment an attempt was made to reduce the spreading out and reduction of level of the reproduced sources in the middle positions. The position of the sources in the studio which was found to give practically equal spacing of the apparent sound sources in the listening room, and the orientation of the microphones required, are shown in Fig. 10. The experiment of Fig. 10 shows that it is possible to obtain an equal spread of the sounds in reproduction. Furthermore, there is only a slight reduction in level for the middle positions.

In the preceding experiments the listener has been located on a line perpendicular to the midpoint of the line joining the two loudspeakers. With the studio sound source positions the same as in Fig. 10, the listener was next located $2\frac{1}{4}$ feet from the centre line, as shown in Fig. 11. The location of the apparent sources then appeared as shown. The apparent sources are concentrated toward the loudspeaker nearest the observer. As would be expected the apparent intensity falls off with the distance from the loudspeaker nearest the observer. In spite of the fact that the apparent sources are not equally spaced or of constant intensity, from a practical standpoint, the stereophonic aspects are preserved.

B. Depth Localization

In any true stereophonic sound reproducing system, there should be a sense of relative depth as well as a sense of lateral distribution. However, in this connection, it is generally recognized that the subjective effect of lateral distribution is more important. Many factors contribute to the subjective sense of depth in stereophonic reproduction of sound. These include differences in: time of arrival of the sound from two sources spaced in depth, intensity of the sources, response frequency characteristics of the sources, and reproduced reverberation of the sources.

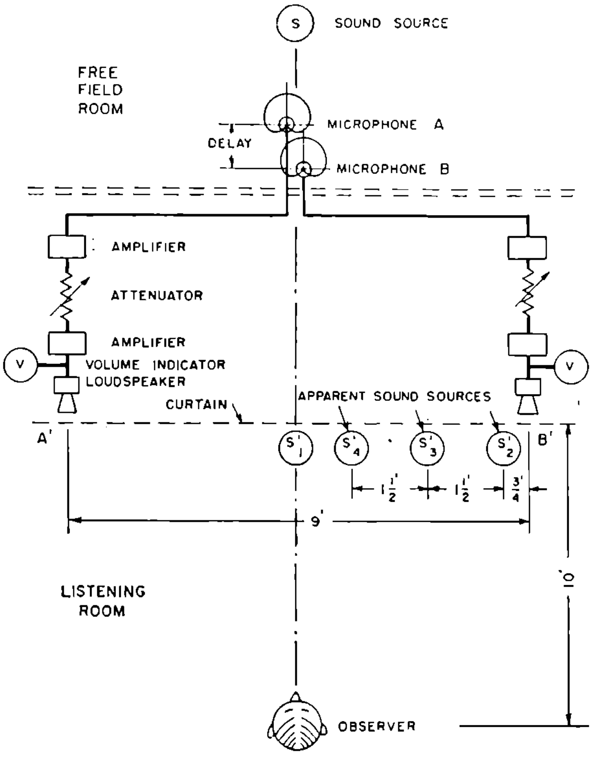

Fig. 8.—Schematic arrangement of the apparatus which illustrates the effect of the relative delay and amplitude of two spaced sound sources in determining the apparent lateral location of the reproduced sound source.

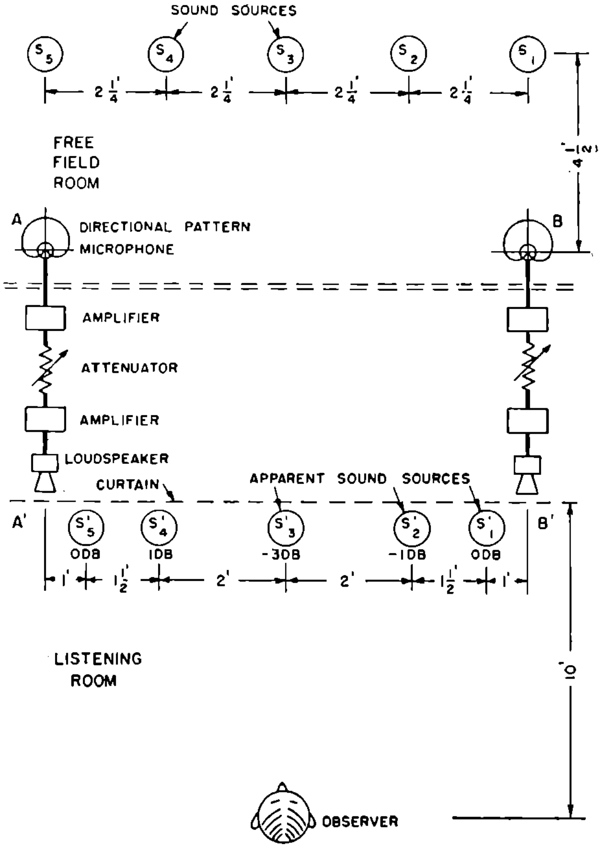

Fig. 9.—Schematic arrangement of the apparatus of a two-channel stereophonic system showing the locations of the sound sources in the free-field room and the apparent sound sources as reproduced in the listening room.

To determine the subjective effect of depth, an experiment was carried out as shown in Fig. 12. A person speaking was located at the positions S1, S2, S3, S4, S5, and S6 in the free-field room. The corresponding apparent estimated locations of the reproduced sound in the listening room are shown as S1', S2', S3', S4', S5', and S6'. The relative estimated sound levels in decibels are also shown with positions S1' designated arbitrarily as the O db reference level. It appears that in this experiment the apparent locations in depth are determined by intensity.

In another experiment, the person speaking was located in the free-field room at position S1 of Fig. 12. The apparent location of the reproduced sound could be moved from S1' in the listening room when a uniform overall response characteristic was used, to a position toward the listener when the response was accentuated in the frequency range from 1,000 to 4,000 cps, or to a position away from the listener when the response was reduced in the same frequency range. The frequency range from 1,000 to 4,000 cps plays an important role in determining the presence[9] of the reproduced sound. When the response in this region is increased, the presence is increased, and the effect is to move the source closer to the listener.

In another experiment, the person speaking was located in a room of 5,500 cubic feet and reverberation[10] time of 0.7 second at position S1, S2, S3, S4, S5, and S6 of Fig. 13. In this experiment the level of the reproduced sound was adjusted so that the output was the same for all locations of the reproduced sound. The corresponding relative locations of the reproduced sound in the listening room was S1', S2', S3', S4', S5', and S6'. As the distance between the source of sound and the microphone is increased, the effective reverberation of the source of sound is increased. As the effective reverberation of the reproduced sound is increased, the presence is decreased, and the apparent location of the source of reproduced sound will move away from the listener.

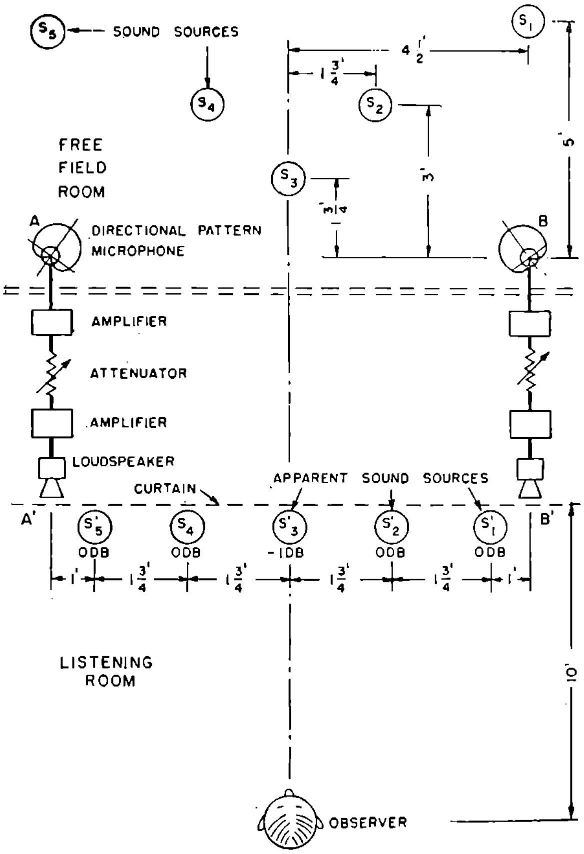

Fig. 10.—Schematic arrangement of the apparatus of a two-channel stereophonic system showing the arrangement of the sound sources in the free-field room which will give a uniform spacing of the sound sources at the curtain in the listening room.

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[9] H. F. Olson, "Musical Engineering", (McGraw-Hill Book Company, New York, N. Y., 1952).

[10] The reverberation time of the studio as a function of the frequency corresponds to that recommended in Olson, "Acoustical Engineering", (D. Van Nostrand Company, Princeton, N. J., 1957).

The systems and experiments depicted in Figs. 8, 9, 10, 11, 12, and 13 demonstrate that practical stereophonic reproduction of sound with the subjective effects of both lateral and depth distribution of the reproduced sound sources can be achieved by the use of a two-channel system.

EXPEDIENTS USED IN STEREOPHONIC SOUND RECORDING

It has been demonstrated in the preceding sections that the apparent location of a source of sound in stereophonic reproduction of sound is determined by the relative intensities and phases of the sounds emanating from the two loudspeakers. These parameters may be employed to convert a single-channel sound pick-up to sterophonic sound reproduction. The arrangement of the apparatus for this is shown in Fig. 14.

In certain musical aggregations it may be desirable to accentuate certain performers. This objective can be realized by the use of additional microphones as shown in Fig. 15. In the system of Fig. 15 the sound source S1 is to be emphasized. This could be accomplished by placing close to the microphone. This may be in The same result can be obtained by th second microphone in channel A place source S1. In this way the relative le source S1 and the other sources can b by means of the gain controls follow phones Ma1 and Ma2. The sound sourc other example of a source to be e However, in this case the output is char above expedients and arrangements c tended to obtain the desired distribut apparent sources in the reproduced sc

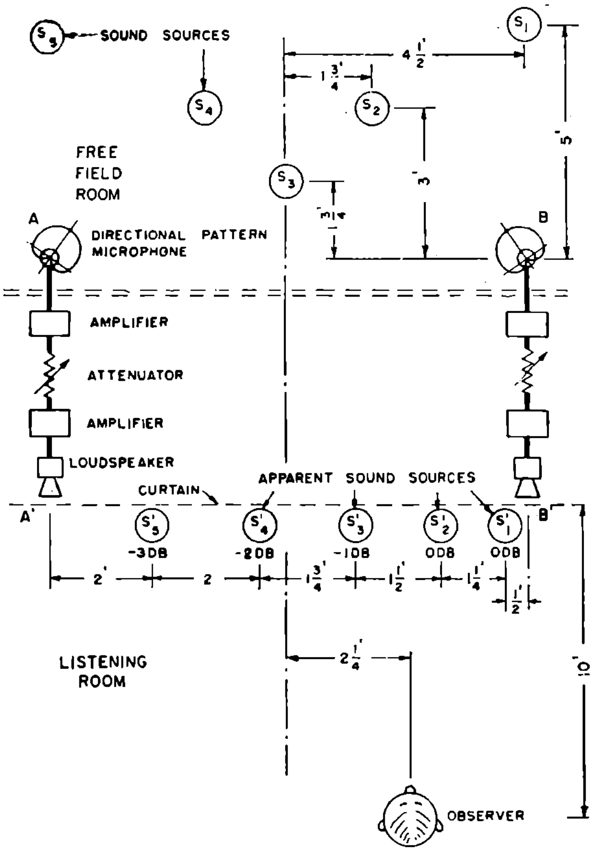

Fig. 11.—Schematic arrangement of the apparatus of a two-channel stereophonic system using the arrangement of the sound sources in the free-field room of Fig. 10 and showing the location of the apparent sound sources at the curtain in the listening room with the observer located 2¼ feet from the centre line.

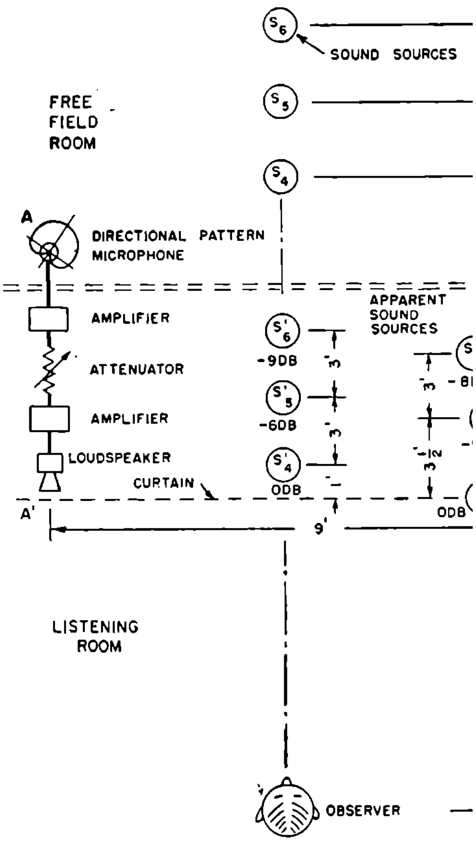

Fig. 12.—Schematic arrangement of the of a two-channel stereophonic systen the arrangement of the sound sources i field room and the apparent location a sound level of the sound sources in th room.

REVERBERATION TESTS

Room acoustics, as exemplified by t reverberation of the reproduced sounc important role in the establishment The effective reverberation[11] is primar tion of the ratio of the direct to refle at the observation point. In the case of

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[11] H. F. Olson, "Acoustical Engineering", (D. Company, Princeton, N. J., 1957).

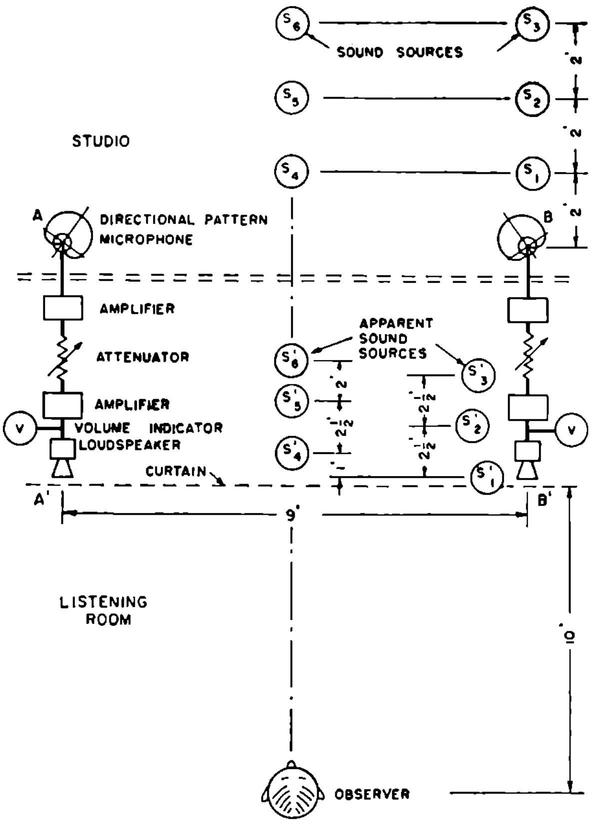

Fig. 13.—Schematic arrangement of the apparatus of a two-channel stereophonic sound system showing the arrangement of the sound sources in the studio and the apparent location of the sound sources in the listening room. The reproduced level of the sound was adjusted so that the output was the same for all locations of the reproduced sound.

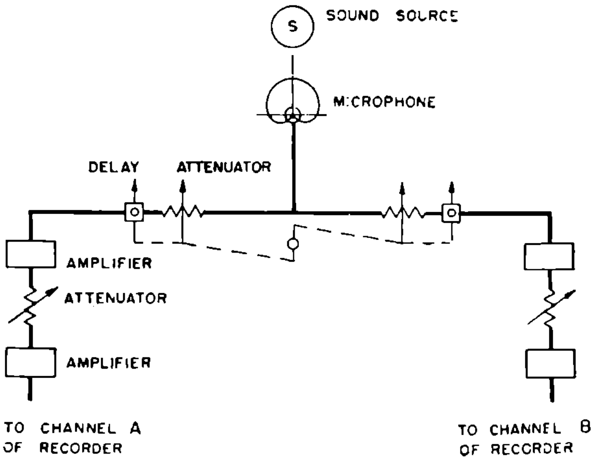

Fig. 14.—Schematic arrangement of the apparatus for converting a single-channel sound pickup into stereophonic sound reproduction.

sound the effective reverberation of the sound picked up in the studio is the ratio of direct to reflected sound signal output of the microphone. The effective reverberation of the signal output of the microphone is primarily a function of the reverberation time of the studio, the distance between sound source and the microphone, and the directional characteristics of the microphone. The effective reverberation of the reproducing system in the listening room is primarily a function of the reverberation time of the listening room, the distance between the loudspeaker and the listener and the directional characteristics of the loudspeaker. The effective reverberation of the overall system is the effective reverberation of the complete chain from the sound source in the studio to the listener in the listening room.

In listening to live sound the acoustics of a single room are involved. However, in the case of the reproduced sound the acoustics of two rooms are involved, namely, the room in which the sound is picked up and the room in which the sound is reproduced. If the room in which the sound is picked up is a free-field room as was the case of the "Stereophonic Frequency Preference Test", then the only acoustics involved are those of the listening room. In effect, the sound sources are transferred to the listening room by means of a sound reproducing system. If the acoustics of the average living room in the home were suitable for all types of music, then the logical procedure would be to record all of the material in a free-field sound room. However, for most recorded material additional reverberation is required. The optimum reverberation of the reproduced sound is

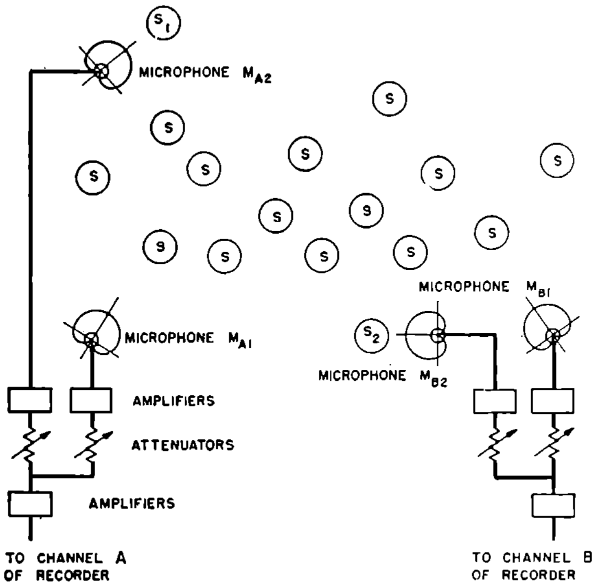

Fig. 15.—Schematic arrangement of the apparatus of a two-channel stereophonic system in which certain sound sources in a large group of sound sources are accentuated in the stereophonic sound reproduction of the aggregation.

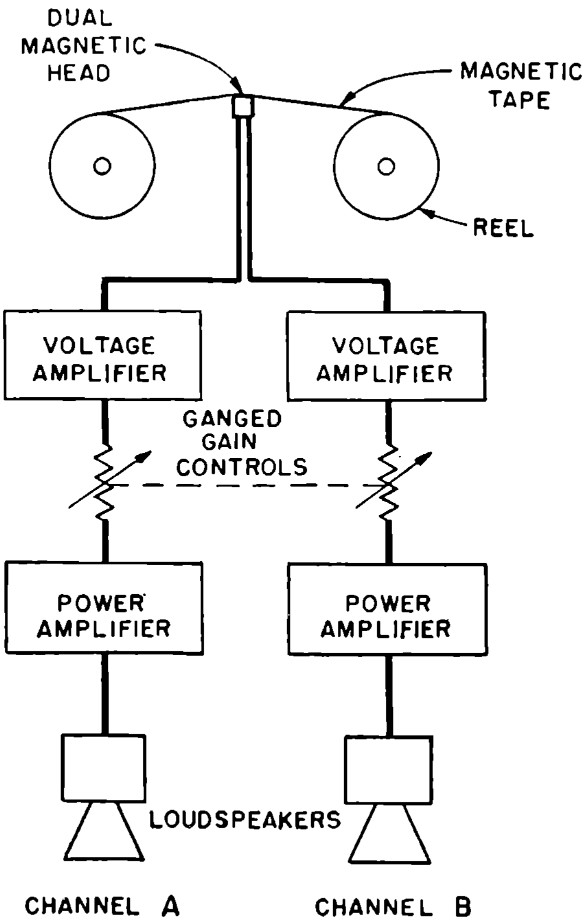

Fig. 16.—Schematic diagram of a two-channel stereophonic magnetic tape sound reproducer.

the sum of the effective reverberation of the studio and listening room. In arriving at the correct value of reverberation of the reproduced sound in practice, it is assumed that the listening room will exhibit the acoustics of the average living room. Then the acoustics of the studio and sound pickup system are selected so that the optimum value of effective reverberation will be produced when the records are reproduced in the average living room.

In this connection, it should be pointed out for example, that true concert hall acoustics cannot be achieved in the sound reproduced in the living room of a home. Nevertheless, a reasonable facsimile of the acoustics of the concert hall can be obtained in the living room by the proper selection of the overall effective reverberation.

RECAPITULATION

As stated in the introduction, in order to achieve realism in a sound reproducing system the following conditions must be satisfied, namely: first, the frequency range must be such as to include all the audible components of the various sounds to be reproduced; second, the volume range must be such as to permit noiseless and distortionless reproduction of the entire range of intensity associated with the sounds; third, the spatial and reverberation characteristics of the original sound must be preserved.

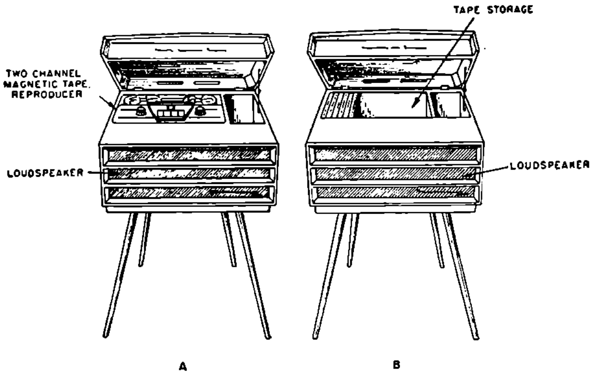

Fig. 17.—Perspective view of a two-channel magnetic tape stereophonic sound reproducer.

As a result of the experiments described in this paper, it appears at the present time that a two-channel stereophonic magnetic tape sound reproducing system will satisfy the above conditions for

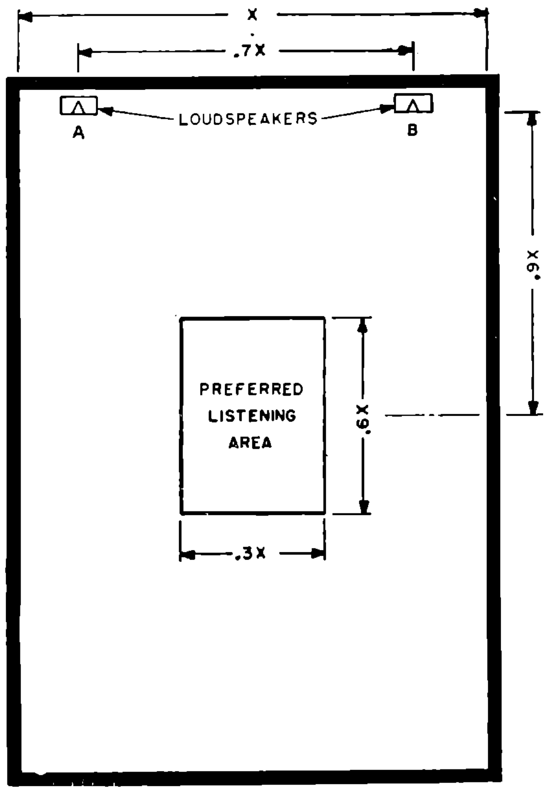

Fig. 18.—The arrangement of a two-channel stereophonic tape reproducing system in a room showing the optimum relative dimensions and the preferred listening area.

obtaining a high order of realism in the home-type sound reproducing systems.

Accordingly, home-type magnetic tape, two-channel, stereophonic-type reproducers have been commercialized; and prerecorded, two-channel, stereophonic tapes are now available for reproducing machines.

SYSTEM OF REPRODUCTION OF STEREOPHONIC MAGNETIC TAPES

The elements of a two-channel system for reproducing two-channel, stereophonic magnetic tape are shown schematically in Fig. 16. The system consists of a two-channel magnetic head, voltage amplifiers for each channel, volume controls for each channel ganged together, power amplifiers for each channel and loudspeakers for each channel.

A perspective view of the commercial system for reproducing prerecorded magnetic tape is shown in Fig. 17. Cabinet A houses the two-channel magnetic tape reproducer, operating controls, amplifiers and the loudspeaker for channel A. Cabinet B houses the loudspeaker for channel B and provides storage space for tape.

PRERECORDED STEREOPHONIC MAGNETIC TAPES

A catalogue of music selections recorded on magnetic tape for stereophonic reproduction of sound in the home is now available in the U.S.A. These tapes have been recorded for reproduction in living room environments employing the principles outlined in the preceding sections. Any standard two-channel stereophonic sound reproducer can be used to reproduce the prerecorded magnetic tapes.

REPRODUCTION OF PRERECORDED STEREOPHONIC MAGNETIC TAPES IN THE HOME

A plan view of a typical living room setting for the reproduction of two-channel stereophonic magnetic tapes is shown in Fig. 18. The arrangement and relative dimensions shown in Fig. 18 apply to practically any room in the average home. The ratios shown in Fig. 18 preserve the dimensional ratios for the most important subjective effect, namely, the angular distribution of the sound sources.

ACKNOWLEDGMENTS

The author wishes to express his thanks to Mr. John Preston for developing the acoustic filter used in the acoustic frequency preference test, to Mr. E. G. May for developing the equipment used in the stereophonic frequency preference test, to Mr. A. R. Morgan for developing the equipment used in the nonlinear distortion tests, to Mr. Preston for assistance in conducting the localization tests, and to Dr. J. G. Woodward for conducting the orchestra and arranging the music for the various tests, and to the members of the orchestra, Messrs. G. A. Arleth, C. A. Hurford, E. G. May, D. G. Murray, J. Preston, C. S. Windham, and J. G. Woodward.

The means and procedures for recording of the commercial stereophonic magnetic tapes were carried out under the direction of Messrs. A. A. Pulley and W. H. Miltenberg. The development of the duplication process in the production of prerecorded magnetic tapes was carried out under the direction of Messrs. H. I. Reiskind and H. E. Roys. The development of the commercial stereophonic reproducer was carried out under the direction of Mr. J. L. Franke. A very brief description of the commercial tapes and systems used for the reproduction of stereophonic sound in the home is given in this paper. A complete description of the commercial procedures and equipment developed for the reproduction of stereophonic sound in the home will be presented in papers by those responsible for this work.

ACKNOWLEDGMENT

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RADIOTRON 17BZP4 PICTURE TUBE

The Radiotron 17BZP4 has a 16-9/16" envelope diagonal and an overall length of only 12-9/16". It features a new electron gun of the "straight" type, designed to minimise deflection distortion. This gun permits a short neck — only 5-7/16" long, and eliminates the need for an ion-trap magnet.

The Radiotron 17BZP4 utilises low-voltage electrostatic focus and employs a 110° deflection angle. It has a spherical filterglass faceplate, an aluminized screen 14-3/4" x 11-11/16" with slightly curved sides and rounded corners, and a minimum projected screen area of 155 square inches. In addition, the 17BZP4 has an external conductive bulb coating which provides a capacitance value ranging between 800 and 1500 μμf.

GENERAL

Heater Voltage	6.3 volts
Heater Current	0.6 amp.
Direct interelectrode Capacitance:	
Grid No. 1 to all other electrodes	6 μμf
Cathode to all other electrodes	5 μμf
External conductive coating to ultor	1,500 max. μμf 800 min. μμf
Faceplate, Spherical	Filterglass
Light Transmission (Approx.)	78%
Phosphor	P4-Sulphide Type aluminized
Fluorescence	White
Phosphorescence	White
Persistance	Short
Focusing Method	Electrostatic
Deflection Method	Magnetic
Deflection Angles (Approx.)	
Diagonal	110°
Horizontal	105°
Vertical	87°

Electron Gun:
Requires no external Ion-trap Magnet.

Tube Dimensions:		
Overall Length	12-9/16"	± 1/4"
Greatest Width	15-5/8"	± 1/8"
Greatest Height	12-3/4"	± 1/8"
Diagonal	16-9/16"	± 1/8"
Neck Length	5-7/16"	± 1/8"

Screen Dimensions (Minimum):	
Greatest Width	14-3/4"
Greatest Height	11-11/16"
Diagonal	15-3/4"
Projected area	155 sq. in.
Weight (Approx.)	10 lbs.
Mounting position	Any
Cap	Recessed small cavity (JETEC No. J1-21)
Bulb	J132-1/2
Base	Small Button Eightar 7-pin Style B (JETEC No. B7-183)

SOCKET CONNECTIONS

As for Radiotron 21CEP4, see page 98

GRID-DRIVE SERVICE

Grid drive is the operating condition in which the video signal varies the Grid-No. 1 potential with respect to cathode.

(Unless otherwise specified, voltage values are positive with respect to Grid No. 1)

MAXIMUM RATINGS, DESIGN-CENTRE VALUES:

ULTOR VOLTAGE	16,000 volts (max.) 12,000 volts (min.)
GRID No. 4 VOLTAGE	
Positive Value	1,000 volts
Negative Value	500 volts
GRID No. 2 VOLTAGE	500 volts
GRID No. 1 VOLTAGE	
Negative peak value	200 volts
Negative bias value	140 volts
Positive bias value	0 volts
Positive peak value	2 volts
PEAK HEATER-CATHODE VOLTAGE	
Heater negative with respect to cathode	180 volts
Heater positive with respect to cathode	180 volts

EQUIPMENT DESIGN RANGES:

With any ultor voltage (E_{c5k}) between 12,000 and 16,000 volts
and Grid No. 2 voltage (E_{c2k}) between 200 and 500 volts

Grid No. 4 Voltage for Focus*	0 to 400	volts
Grid No. 1 Voltage for Visual Extinction of Focused Raster	— 9.3% to — 24% of E_{c2k}	volts
Grid No. 1 Video Drive from Raster Cutoff (Black Level):		
White-Level Value (Peak Positive)	9.3% to 24% of E_{c2k}	volts
Grid No. 4 Current	— 25 to + 25	µamp
Grid No. 2 Current	— 15 to + 15	µamp
Field Strength of Adjustable Centring Magnet	0 to 8	gausses

EXAMPLES OF USE OF DESIGN RANGES:

With Ultor Voltage of	14000	16000	volts
And Grid No. 2 Voltage of	300	300	volts
Grid No. 4 Voltage for focus	0 to +400	0 to +400	volts
Grid No. 1 Voltage for Visual Extinction of Focused Raster	— 28 to — 72	— 36 to — 94	volts
Grid No. 1 Video Drive from Raster Cutoff (Black Level):			
White-Level Value (Peak Positive)	28 to 72	36 to 94	volts

MAXIMUM CIRCUIT VALUE:

Grid No. 1 Circuit Resistance	1.5	megohms

CATHODE-DRIVE SERVICE

Cathode drive is the operating condition in which the video signal varies the cathode potential with respect to Grid No. 1 and the other electrodes.

(Unless otherwise specified, voltage values are positive with respect to Grid No. 1)

MAXIMUM RATINGS, Design-Centre Values:

ULTOR TO GRID No. 1 VOLTAGE	16,000 volts (max.) 12,000 volts (min.)
GRID No. 4 TO GRID No. 1 VOLTAGE	
Positive Value	1,000 volts
Negative Value	500 volts
GRID No. 2 TO GRID No. 1 VOLTAGE	640 volts
GRID No. 2 TO CATHODE VOLTAGE	500 volts
CATHODE TO GRID No. 1 VOLTAGE	
Positive Peak Value	200 volts
Positive Bias Value	140 volts
Negative Bias Value	0 volts
Negative Peak Value	2 volts
PEAK HEATER-CATHODE VOLTAGE	
Heater Negative with respect to Cathode	180 volts
Heater Positive with respect to Cathode	180 volts

EQUIPMENT DESIGN RANGES:

With any ultor to Grid No. 1 Voltage (E_{c5g1}) between 12,000 and 16,000 volts
and Grid No. 2 to Grid No. 1 Voltage (E_{c2g1}) between 225 and 640 volts

Grid No. 4 to Grid No. 1 Voltage for focus*	0 to 400	volts
Cathode to Grid No. 1 Voltage for Visual Extinction of focused Raster	8.5% to 19.4% of E_{c2g1}	volts
Cathode to Grid No. 1 Video Drive from Raster Cutoff (Black Level):		
White-Level Value (Peak Negative)	— 8.5% to — 19.4% of E_{c2g1}	volts
Grid No. 4 Current	— 25 to + 25	µamp
Grid No. 2 Current	— 15 to + 15	µamp
Field Strength of Adjustable Centring Magnet	0 to 8	gausses

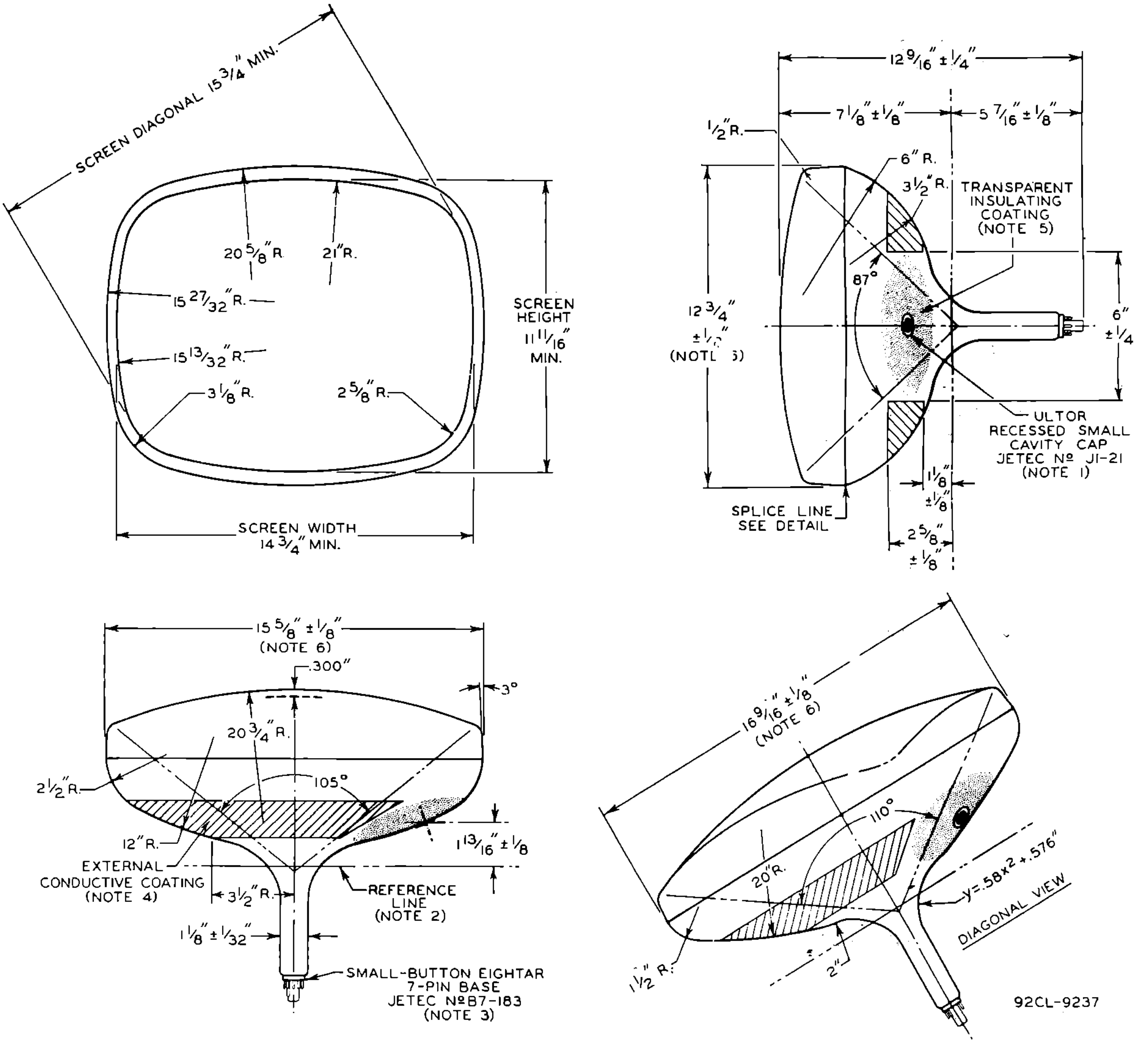
EXAMPLES OF USE OF DESIGN RANGES:

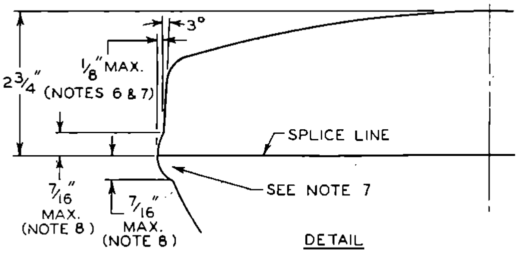
With Ultor to Grid No. 1 Voltage of	14000	16000	volts
And Grid No. 2 to Grid No. 1 Voltage of	300	400	volts
Grid No. 4 to Grid No. 1 Voltage for focus	0 to 400	0 to 400	volts
Cathode to Grid No. 1 Voltage for Visual Extinction of Focused Raster	28 to 60	36 to 78	volts
Cathode to Grid No. 1 Video Drive from Raster Cutoff (Black Level):			
White-Level Value	— 28 to — 60	— 36 to — 78	volts

MAXIMUM CIRCUIT VALUE:

Grid No. 1 Circuit Resistance	1.5 megohms

* The Grid No. 4 Voltage or Grid No. 4 to Grid No. 1 Voltage required for focus of any individual tube is independent of Ultor current and will remain essentially constant for values of Ultor Voltage (or Ultor to Grid No. 1 Voltage) or Grid No. 2 Voltage (or Grid No. 2 to Grid No. 1 Voltage) within design ranges shown for these items.





NOTES:
For Notes, see page 101.

RADIOTRON 21CEP4 PICTURE TUBE

The Radiotron 21CEP4 has a 21-3/8" envelope diagonal and an overall length of only 14-7/16". It features a new electron gun of the "straight" type designed to minimize deflection distortion. This gun permits a short neck — only 5-7/16" long, and eliminates the need for an ion-trap magnet.

The 21CEP4 utilizes low-voltage electrostatic focus, employs a 110° deflection angle, and has a maximum ultor-voltage rating of 18 Kv (design centre). It has spherical Filterglass face-plate, an aluminized screen 19-1/16" x 15-1/16" with slightly curved sides and rounded corners, and a minimum projected screen area of 262 square inches. In addition, the 21CEP4 has an external conductive bulb coating which provides a capacitance value ranging between 2,000 and 2,500 $\mu\mu$f.

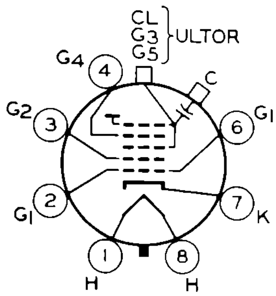
GENERAL

Heater Voltage	6.3 volts
Heater Current	0.6 amp
Direct Interelectrode Capacitances:	
Grid No. 1 to all other electrodes	6 $\mu\mu$f
Cathode to all other electrodes	5 $\mu\mu$f
External conductive coating to ultor	2,500 max. $\mu\mu$f 2,000 min. $\mu\mu$f
Faceplate, Spherical	Filterglass
Light transmission (approx.)	73%
Phosphor, metal-backed	P4 Sulphide Type
Fluorescence	White
Phosphorescence	White
Persistance	Short
Focusing Method	Electrostatic
Deflection Method	Magnetic
Deflection Angles (approx.):	
Diagonal	110°
Horizontal	106°
Vertical	85°
Electron Gun:	
Requires No Ion-Trap Magnet	
Tube Dimensions:	
Overall Length	14-7/16" ± 5/16"
Greatest Width	20-1/4" ± 1/8"
Greatest Height	16-3/8" ± 1/8"
Diagonal	21-3/8" ± 1/8"
Neck Length	5-7/16" ± 1/8"

Screen Dimensions (minimum):	
Greatest width	19-1/16"
Greatest Height	15-1/16"
Diagonal	20-1/4"
Projected area	262 sq. in.
Cap	Recessed small cavity (JETEC No. J1-21)
Bulb	J171
Base	Small Button Eightar 7-pin (JETEC No. B7-183)
Weight (approx.)	23 lbs.
Mounting Position	Any

SOCKET CONNECTIONS

(bottom view)



Pin 1 — Heater
Pin 2 — Grid No. 1
Pin 3 — Grid No. 2
Pin 4 — Grid No. 4
Pin 6 — Grid No. 1
Pin 7 — Cathode
Pin 8 — Heater
Cap — Ultor (Grid No. 3, Grid No. 5, Collector)
C — External Conductive Coating.

GRID-DRIVE SERVICE

Grid drive is the operating condition in which the video signal varies the Grid-No. 1 potential with respect to cathode.

(Unless otherwise specified, voltage values are positive with respect to cathode.)

MAXIMUM RATINGS, Design-Centre Values:

ULTOR VOLTAGE	18,000 volts
GRID No. 4 VOLTAGE:	
Positive value	1,000 volts
Negative value	500 volts
GRID No. 2 VOLTAGE	500 volts

GRID No. 1 VOLTAGE	
Negative peak value	200 volts
Negative bias value	140 volts
Positive bias value	0 volts
Positive peak value	2 volts
PEAK HEATER-CATHODE VOLTAGE:	
Heater negative with respect to cathode	180 volts
Heater positive with respect to cathode	180 volts

EQUIPMENT DESIGN RANGES:

With any Ultor Voltage (E_{c5k}) between 12,000 and 18,000 volts
and Grid No. 2 Voltage (E_{c2k}) between 200 and 500 volts

Grid No. 4 Voltage for Focus*	0 to 400	volts
Grid No. 1 Voltage for Visual Extinction of Focused Raster	— 9.3% to — 24% of E_{c2k}	volts
Grid No. 1 Video Drive from Raster Cutoff (Black Level):		
White Level Drive (Peak Positive)	9.3% to 24% of E_{c2k}	volts
Grid No. 4 Current	— 25 to + 25	μamp
Grid No. 2 Current	— 15 to + 15	μamp
Field Strength of Adjustable Centring Magnet	0 to 8	gausses

EXAMPLES OF DESIGN RANGES:

With Ultor Voltage of	14000	16000	volts
And Grid No. 2 Voltage of	300	400	volts
Grid No. 4 Voltage for Focus	0 to 400	0 to 400	volts
Grid No. 1 Voltage for Visual Extinction of Focused Raster	— 28 to — 72	— 36 to — 94	Volts
Grid No. 1 Video Drive from Raster Cutoff (Black Level):			
White-Level Drive (Peak Positive)	28 to 72	36 to 94	volts

MAXIMUM CIRCUIT VALUE:

Grid No. 1 Circuit Resistance	1.5 megohms

CATHODE-DRIVE-SERVICE

Cathode drive is the operating condition in which the video signal varies the cathode potential with respect to Grid No. 1 and the other electrodes.

(Unless otherwise specified, voltage values are positive with respect to Grid No. 1)

MAXIMUM RATINGS, Design-Centre Values:

ULTOR TO GRID No. 1 VOLTAGE	18,000 volts
GRID No. 4 TO GRID No. 1 VOLTAGE:	
Positive value	1,000 volts
Negative value	500 volts
GRID No. 2 TO GRID No. 1 VOLTAGE	640 volts
GRID No. 2 TO CATHODE VOLTAGE	500 volts
CATHODE TO GRID No. 1 VOLTAGE:	
Positive peak value	200 volts
Positive bias value	140 volts
Negative bias value	0 volts
Negative peak value	2 volts
PEAK HEATER-CATHODE VOLTAGE:	
Heater negative with respect to cathode	180 volts
Heater positive with respect to cathode	180 volts

EQUIPMENT DESIGN RANGES:

With any Ultor to Grid No. 1 Voltage (E_{c5g1}) between 12,000 and 18,000 volts
and Grid No. 2 to Grid No. 1 Voltage (E_{c2g1}) between 225 and 640 volts

Grid No. 4 to Grid No. 1 Voltage for Focus*	0 to 400	volts
Cathode to Grid No. 1 Voltage for Visual Extinction of Focused Raster	8.5% to 19.4% of E_{c2g1}	volts
Cathode to Grid No. 1 Video Drive from Raster Cutoff (Black Level):		
White-Level Value (Peak Negative)	— 8.5% to — 19.4% of E_{c2g1}	volts
Grid No. 4 Current	— 25 to + 25	μamp
Grid No. 2 Current	— 15 to + 15	μamp
Field Strength of Adjustable Centring Magnet	0 to 8	gausses

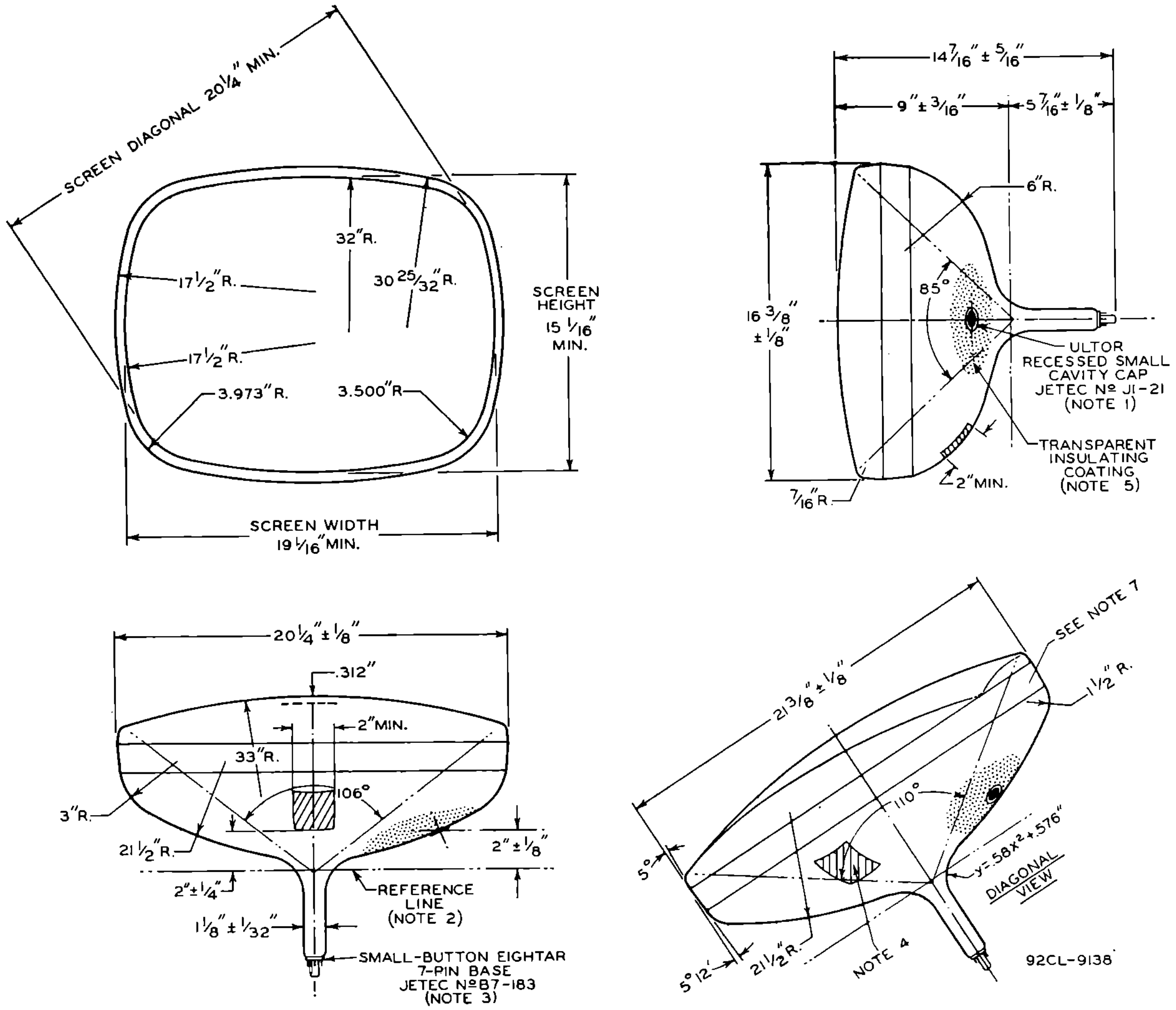
EXAMPLES OF USE OF DESIGN RANGES:

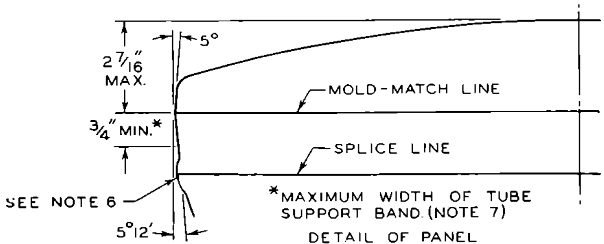
With Ultor to Grid No. 1 Voltage of	14000	16000	volts
And Grid No. 2 to Grid No. 1 Voltage of	300	400	volts
Grid No. 4 to Grid No. 1 Voltage for Focus	0 to 400	0 to 400	volts
Cathode to Grid No. 1 Voltage for Visual Extinction of Focused Raster	28 to 60	36 to 78	volts
Cathode to Grid No. 1 Video Drive from Raster Cutoff (Black Level): White-Level Value (Peak Negative)	—28 to — 60	— 36 to — 78	volts

MAXIMUM CIRCUIT VALUE:

Grid No. 1 Circuit Resistance	1.5 megohms

* The Grid No. 4 Voltage or Grid No. 4 to Grid No. 1 Voltage required for focus of any individual tube is independent of Ultor current and will remain essentially constant for values of Ultor Voltage (or Ultor to Grid No. 1 Voltage) or Grid No. 2 Voltage (or Grid No. 2 to Grid No. 1 Voltage) within design ranges shown for these items.





DETAIL OF PANEL

Note 1:
The plane through the tube axis and pin No. 4 may vary from the plane through the tube axis and bulb terminal by angular tolerance (measured about the tube axis) of ± 30°. Ultor terminal is on the same side as Pin No. 4.

Note 2:
With tube neck inserted through flared end of reference-line gauge (JETEC No. 126) and with tube seated in gauge, the reference line is determined by the intersection of the plane CC' of the gauge with the glass funnel.

Note 3:
Socket for this base should not be rigidly mounted; it should have flexible leads and be allowed to move freely. Bottom circumference of base wafer will fall within a circle concentric with bulb axis and having a diameter of 1-3/4".

Note 4:
The drawing shows minimum size and location of the contact area of the external conductive coating. The actual area of this coating will be greater than the contact area so as to provide the required capacitance. External conductive coating must be grounded.

Note 5:
To clean this area, wipe only with soft dry lintless cloth.

Note 6:
Bulge at splice-line seal may increase indicated maximum for envelope width, diagonal and height by not more than 1/4" (17BZP4) or 1/8" (21CEP4), but at any point around the seal, the bulge will not protrude more than 1/8" (17BZP4) or 1/16" (21CEP4) beyond the envelope surface at the mold-match line.

Note 7:
(17BZP4) measured 2-9/32" ± 1/32" from the plane tangent to the surface of the faceplate at the tube axis.

(21CEP4) undisturbed area between mold-match line and splice line is 3/4" minimum. This should be the maximum width of the tube support band.

Note 8 (17BZP4 only):
The tube should be supported on both sides of the bulge. The mechanism used should provide clearance for the maximum dimensions of the bulge.

New Publications

"Servicing Transistorized Radios", Amalgamated Wireless Valve Co. Pty. Ltd., Qto., 24 pp.

This Radiotron publication has been produced in response to a demand for a series of articles on the servicing of transistorized radios which appeared in "Radiotronics". The articles deal with transistor circuitry, servicing considerations, and servicing practice, and are here assembled into a convenient booklet. No serviceman can afford to be without a copy of this book, and it is recommended to all service organisations and workshops.

"Phototubes", Amalgamated Wireless Valve Co. Pty. Ltd., Qto., 50 pp.

This book is designed to present a range of over fifty phototubes, photomultipliers and photoconductive cells, and gives technical data on each unit. It goes further however, by incorporating a series of articles on the selection, application and operation of phototubes, and on the principles governing their operation. For those who use or intend to use phototubes, or even for those who would like to know more about how they work, this book is an essential item on the library shelf. It is yet another Radiotron publication, with the accent on straight-forward, easy-to-read presentation.

"Basic Electronics", 2nd Edition, P. B. Zbar and S. Schildkraut, McGraw-Hill, Qto., 148 pp.

This book is another of the series of excellent volumes prepared by the teaching staff of the Electronic Industries Association. It will be remembered that two books in this series dealing with TV were reviewed in "Radiotronics", Vol. 23, No. 10, October 1958. This particular volume sets about providing the reader with a groundwork in electronics, by carrying out a series of 59 "jobs" or experiments. Nine of these, incidentally, deal with transistors. The book is intended for those entering the electronic industry, and in your reviewer's opinion is excellent for the purpose. In fact, for anyone desiring an introduction to electronics, this volume can be strongly recommended.

The objectives of each "job" or experiment are clearly stated, together with the materials required. Introductory information to each job includes a down-to-earth discussion of the relevant theory and practice, including basic principles, instrument techniques, circuit parameters, and the effect of varying parameters on circuit performance. Each job concludes with a questionnaire on the work done, with provision for logging results obtained. This is undoubtedly one of the best books for beginners that has appeared in a long time, and would be difficult to better.

DESIGN MAXIMUM SYSTEM FOR RATING VALVES

Ratings are established on valve types to guide and assist equipment designers in utilizing to best advantage the performance and service capabilities of each type. Rating values are provided for those characteristics for which careful study and experience indicate limiting values are required to ensure satisfactory performance. In order that the numerical values of a rating system have significance, the system used must be accurately defined and properly applied.

Valve Rating Systems

Three rating systems are in use by the industry. The oldest is known as the Absolute-Maximum System, the next as the Design-Centre System, and the latest and newest is the Design-Maximum System. Definitions of these systems formulated by the Joint Electron Tube Engineering Council (JETEC) and standardized by NEMA and EIA will be explained in a future series of articles.

The numerical values of the various ratings shown under the three systems are not directly comparable because of certain allowances (factors of safety) made for equipment, component, and adjustment tolerances, supply-voltage variations, environmental conditions, and valve variations as prescribed by the definition of the system used.

The significant differences between the three Rating systems can be summarized as follows:

Absolute-Maximum System

In this system the ratings are the maximum capabilities of any valve of the type rated.

Design-Centre System

Here the ratings are the maximum capabilities of any valve of the type rated, less an allowance for valve variations, and less an allowance for component and supply variations.

Design-Maximum System

In this system the ratings are the maximum capabilities of any valve of the type rated, less an allowance for valve variations.

These expressions show clearly why a valve of any given type might be rated with entirely different numerical values, depending on the system used, although in effect it is rated at the same capability level in each system. When the equipment designer uses valve rating values, he must be careful to note the system used and must take into consideration the appropriate allowances.

The definitions specify whether bogey or limit valves of a type should be used when the equipment designer wants to determine whether his design uses that valve type within its ratings. For the Design-Centre and Design-Maximum Systems, normal, average, or bogey valves should be used; for the Absolute-Maximum System, limit valves are necessary.

Under the Design-Maximum System, a bogey valve should be put into the socket under test while measurements of voltages, currents, temperatures, or other characteristics as designated by the data sheet are made under the worst probable or limit values of line voltage, environmental conditions, equipment components, or equipment adjustment.

Note that this procedure differs from that used in checking under the Design-Centre System where the same bogey valve would be used but the measurements would be made under the normal, or bogey, conditions for line voltage, environment, equipment adjustment, and component tolerances.

If the check were to be made under the Absolute-Maximum System, it would be necessary to make the measurements not only at the worst probable conditions for the equipment variables, but also with a limit valve of the type under consideration instead of the bogey one which is used in the other two systems.

Benefits of Design-Maximum System

The use of the Design-Maximum System of ratings for entertainment-type receiving valves is expected to benefit both equipment manufacturers and valve manufacturers. One of the benefits of this system is that it provides for the safety factors to be determined by the equipment or valve design groups best able to evaluate them. The equipment designer is better able to allow for the variations in use and design of his equipment; the valve manufacturer is better able to judge the allowance in ratings needed due to variations of his product.

Because better judgment can be used in determining these allowances, improved valve performance can be expected through:

1. Better life. Valve manufacturers can cover ratings more realistically on their life tests.
2. Fuller benefit of valve ratings. Equipment designed with close tolerances can safely use valves at the published numerical ratings.
3. More reliable valve usage. The appropriate allowance for equipment is determined by the equipment designer who is better able to make this evaluation than the valve manufacturer.

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