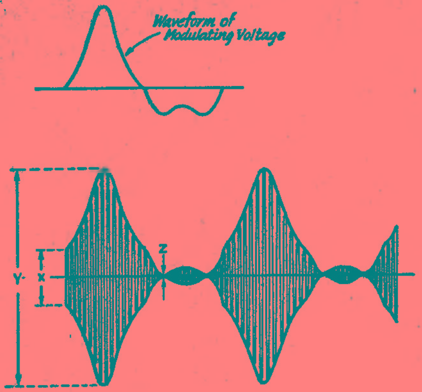


Fig. 12-3 — Modulation by an unsymmetrical wave form. This drawing shows 100-percent downward modulation along with 300-percent upward modulation. There is no distortion, since the modulation envelope is an accurate reproduction of the wave form of the modulating voltage.

modulation automatically fulfills the requirement that the *instantaneous* power at the modulation uppeak be four times the carrier power. Consequently, systems in which the additional power is supplied from outside the modulated rf stage (e.g., plate modulation) usually are designed on a sine-wave basis as a matter of convenience. Modulation systems in which the additional power is secured from the modulated rf amplifier (e.g., grid modulation) usually are more conveniently designed on the basis of peak envelope power rather than average power.

The extra power that is contained in a modulated signal goes entirely into the sidebands, half in the upper sideband and half into the lower. As a numerical example, full modulation of a 100-watt carrier by a sine wave will add 50 watts of sideband power, 25 in the lower and 25 in the upper sideband. With lower modulation percentages, the sideband power is proportional to the *square* of the modulation percentage, i.e., 50-percent modulation will add 12.5 watts of sideband power, 6.25 watts in each sideband. Supplying this additional power for the sidebands is the object of all of the various systems devised for amplitude modulation.

No such simple relationship exists with complex wave forms. Complex wave forms such as speech do not, as a rule, contain as much average power as a sine wave. Ordinary speech wave forms have about half as much average power as a sine wave, for the same peak amplitude in both wave forms. Thus for the same modulation percentage, the sideband power with ordinary speech will average only about half the power with sine-wave modulation, since it is the peak envelope amplitude, not the average power, that determines the percentage of modulation.

Unsymmetrical Modulation

In an ordinary electric circuit it is possible to increase the amplitude of current flow indefinitely, up to the limit of the power-handling capability of the components, but it cannot very well be decreased to less than zero. The same thing is true of the amplitude of an rf signal; it can be modulated *upward* to any desired extent, but it cannot be modulated *downward* more than 100 percent.

When the modulating wave form is unsymmetrical it is possible for the upward and downward modulation percentages to be different. A simple case is shown in Fig. 12-3. The positive peak of the modulating signal is about 3 times the amplitude of the negative peak. If, as shown in the drawing, the modulating amplitude is adjusted so that the peak downward modulation is just 100 percent ($Z = 0$) the peak upward modulation is 300 percent ($Y = 4X$). The carrier amplitude is represented by X, as in Fig. 12-2. The modulation envelope reproduces the wave form of the modulating signal accurately, hence there is no distortion. In such a modulated signal the increase in power output with modulation is considerably greater than it is when the modulation is symmetrical. In Fig. 12-3 the peak envelope amplitude, Y, is four times the carrier amplitude, X, so the peak-envelope power (PEP) is 16 times the carrier power. When the upward modulation is more than 100 percent the power capacity of the modulating system obviously must be increased sufficiently to take care of the much larger peak amplitudes. Such a system of modulation, often called "supermodulation," was popular among amateurs in the early 1950s. (See bibliography at the end of this chapter.)

Overmodulation

If the amplitude of the modulation on the downward swing becomes too great, there will be a period of time during which the rf output is entirely cut off. This is shown in Fig. 12-4. The shape of the downward half of the modulating wave is no longer accurately reproduced by the modulation envelope, consequently the modulation is distorted. Operation of this type is called **overmodulation.**

The distortion of the modulation envelope causes new frequencies (harmonics of the modulating frequency) to be generated. These combine

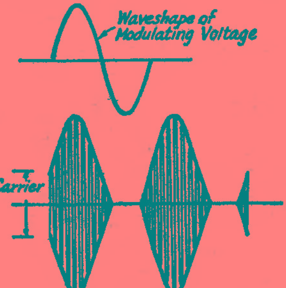


Fig. 12-4 — An overmodulated signal. The modulation envelope is not an accurate reproduction of the wave form of the moduleting voltage. This, or any type of distortion occurring during the modulation process, generates spurious sidebands or "splatter."

with the carrier to form new side frequencies that widen the channel occupied by the modulated signal, as shown in Fig. 12-1C. These spurious frequencies are commonly called "splatter."

It is important to realize that the channel occupied by an amplitude-modulated signal is dependent *on the shape of the modulation envelope.* If this wave shape is complex and can be resolved into a wide band of audio frequencies, then the channel occupied will be correspondingly large. An overmodulated signal splatters and occupies a much wider channel than is necessary because the "clipping" of the modulating wave that occurs at the zero axis changes the envelope wave shape to one that contains high-order harmonics of the original modulating frequency. These harmonics appear as side frequencies separated by, in some cases, many kilohertz from the carrier frequency.

Because of this clipping action at the zero axis, it is important that care be taken to prevent applying too large a modulating signal in the downward direction. Overmodulation downward results in more splatter than is caused by most other types of distortion in a phone transmitter.

AMPLITUDE MODULATION METHODS

MODULATION SYSTEMS

As explained in the preceding section, amplitude modulation of a carrier is accompanied by an increase in power output, the additional power being the "useful" or "talk power" in the sidebands. This additional power may be supplied from an external source in the form of audio-frequency power. It is then added to the unmodulated power input to the amplifier to be modulated, after which the combined power is converted to rf. This is the method used in plate or collector modulation. It has the advantage that the rf power is generated at the high-efficiency characteristic of Class C amplifiers – of the order of 65 to 75 percent – but has the accompanying disadvantage that generating the audio-frequency power is rather expensive.

An alternative that does not require relatively large amounts of audio-frequency power makes use of the fact that the power output of an amplifier can be controlled by varying the potential of a tube or transistor element – such as a control or screen grid or a transistor base – that does not, in itself, consume appreciable power. In this case the additional power during modulation is secured by sacrificing carrier power; in other words, a tube is capable of delivering only so much total power within its ratings, and if more must be delivered at full modulation, then less is available for the unmodulated carrier. Systems of this type must of necessity work at rather low efficiency at the unmodulated carrier level. As a practical working rule, the efficiency of the modulated rf amplifier is of the order of 30 to 35 percent, and the unmodulated carrier power output obtainable with such a system is only about one-fourth to one-third that obtainable from the same amplifier with plate modulation.

PLATE OR COLLECTOR MODULATION

Fig. 12-5 shows a system of plate modulation, in this case with a triode rf tube. A balanced (push-pull Class A, Class AB or Class B) modulator is transformer coupled to the plate circuit of the modulated rf amplifier. The audio-frequency power generated by the modulator is combined with the dc power in the modulated-amplifier plate circuit by transfer through the coupling transformer, T. For 100-percent modulation the audio-frequency power output of the modulator and the turns ratio of the coupling transformer must be such that the voltage at the plate of the of the modulated amplifier varies between zero and twice and dc operating plate voltage, thus causing corresponding variations in the amplitude of the rf output. The tubes of Fig. 12-5 may be replaced with transistors, either bipolar or FET, for collector or drain modulation.

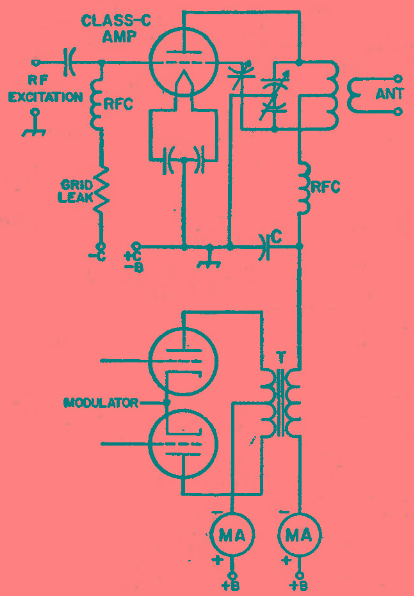


Fig. 12-5 – Plate modulation of a Class C rf amplifier. The rf plate bypass capacitor, C, in the amplifier stage should have reasonably high reactance at audio frequencies. A value of the order of .001 μF to .005 μF is satisfactory in practically all cases for vacuum-tube circuits. A considerably higher value will be required if the vacuum tubes are replaced by transistors – in the order of a few microfarads.

Audio Power

As stated earlier, the average power output of the modulated stage must increase during modulation. The modulator must be capable of supplying to the modulated rf stage sine-wave audio power equal to 50 percent of the dc input power. For example, if the dc input power to the rf stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Although the total power input (dc plus audio-frequency ac) increases with modulation, the dc plate or collector current of a modulated amplifier should not change when the stage is modulated. This is because each increase in voltage and current is balanced by an equivalent decrease in voltage and current on the next half cycle of the modulating wave. Dc instruments cannot follow the af variations, and since the average dc plate or collector current and voltage of a properly operated amplifier do not change, neither do the meter readings. A change in current with modulation indicates nonlinearity. On the other hand, a thermocouple rf ammeter connected in the antenna, or transmission line, will show an increase in rf current with modulation, because instruments of this type respond to power rather than to current or voltage.

Modulating Impedance; Linearity

The modulating impedance, or load resistance presented to the modulator by the modulated rf amplifier, is equal to

$$Z_m = \frac{E_b}{I_p} \times 1000 \text{ ohms}$$

where E_b = Dc plate or collector voltage
I_p = Dc plate or collector current (mA)
E_b and I_p are measured without modulation.

The power output of the rf amplifier must vary as the square of the instantaneous plate or collector voltage (the rf output voltage must be proportional to the plate or collector voltage) for the modulation to be linear. This will be the case when the amplifier operates under Class C conditions. The linearity depends upon having sufficient grid or base excitation and proper bias, and upon the adjustment or circuit constants to the proper values.

Screen-Grid RF Amplifiers

Screen-grid tubes of the pentode or beam-tetrode type can be used in Class C plate-modulated amplifiers by applying the modulation to both the plate and screen grid. The usual method of feeding the screen grid with the necessary dc and modulation voltages is shown in Fig. 12-6. The dropping resistor, R, should be of the proper value to apply normal dc voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

The modulating impedance is found by dividing the dc plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents gives the power input to be used as the basis for determining the audio power required from the modulator.

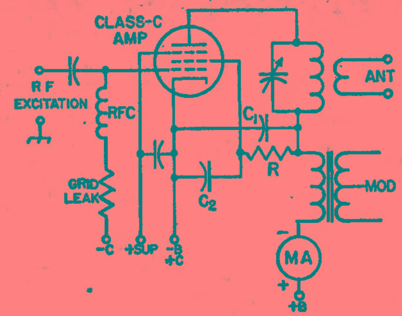


Fig. 12-6 — Plate end screen modulation of a Class C rf amplifier using a screen-grid tube. The plate rf bypass capacitor, C1, should have reasonably high reactance at all audio frequencies; a value of .001 to .005 μF is generally satisfactory. The screen bypass, C2, should not exceed .002 μF in the usual case.

Modulation of the screen along with the plate is necessary because the screen voltage has a much greater effect on the plate current than the plate voltage does. The modulation characteristic is nonlinear if the plate alone is modulated.

Choke-Coupled or Heising Modulation

One of the oldest types of plate modulating systems is the choke-coupled Class A or Heising modulator shown in Fig. 12-7. Because of the relatively low power output and plate efficiency of a Class A amplifier, the method is rarely used now except for a few special applications.

The audio power output of the modulator is combined with the dc power in the plate circuit through the modulation choke, L1, which has a high impedance at audio frequencies. This technique of modulating the rf signal is similar to the case of the transformer-coupled modulator but there is considerably less freedom in adjustment since no transformer is available for matching impedances. The dc input power to the rf stage must not exceed twice the rated af power output of the modulator, and for 100-percent modulation the plate voltage on the modulator must be higher than the plate voltage on the rf amplifier. This is because the af voltage developed by the modulator cannot swing to zero without a great deal of distortion. R1 provides the necessary dc voltage drop between the modulator and the rf amplifier. The voltage drop across this resistor must equal the minimum instantaneous plate voltage on the modulator tube under normal operating conditions. C1, an audio-frequency bypass across R1, should have a capacitance such that its reactance at 100 Hz is not more than about one-tenth the resistance of R1. Without R1-C1 the percentage of modulation is limited to 70 to 80 percent in the average case.

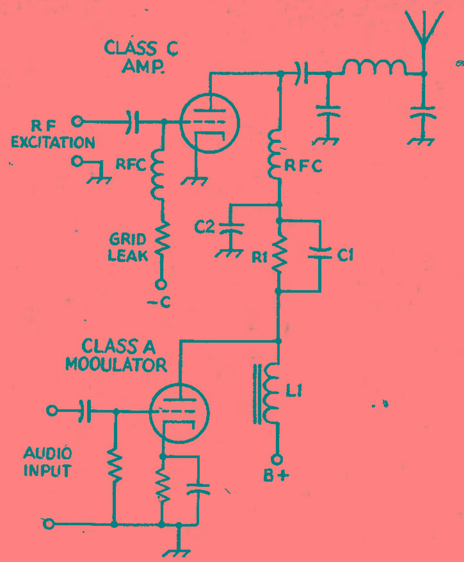


Fig. 12-7 — Choke-coupled Class A modulator. The modulation choke, L1, should have a value of 5 H or more. A value of .001 to .005 μF is satisfactory for C2. See text for discussion of C1 and R1.

GRID MODULATION

The principal disadvantage of plate modulation is that a considerable amount of audio power is necessary. This requirement can be avoided by applying the modulation to a grid element in the modulated amplifier. However, serious disadvantages of grid modulation are the reduction in the carrier power output obtainable from a given rf amplifier tube and the more rigorous operating requirements and more complicated adjustment.

The term "grid modulation", as used here applies to all types – control grid, screen, or suppressor – since the operating principles are exactly the same no matter which grid is actually modulated. (Screen-grid modulation is the most commonly used technique of the three types listed here.) With grid modulation the plate voltage is constant, and the increase in power output with modulation is obtained by making both the plate current and plate efficiency vary with the modulating signal. The efficiency obtainable at the envelope peak depends on how carefully the modulated amplifier is adjusted, and sometimes can be as high as 80 percent. It is generally less when the amplifier is adjusted for good linearity, and under average conditions a round figure of 2/3, or 66 percent, is representative. The efficiency without modulation is only half the peak efficiency, or about 33 percent. This low average efficiency reduces the permissible carrier output to about one-fourth the power obtainable from the same tube in cw operation, and to about one-third the carrier output obtainable from the tube with plate modulation.

The modulator is required to furnish only the audio power dissipated in the modulated grid under the operating conditions chosen. A speech amplifier capable of delivering 3 to 10 watts is usually sufficient.

Grid modulation does not give quite as linear a modulation characteristic as plate modulation, even under optimum operating conditions. When misadjusted the nonlinearity may be severe, resulting in considerable distortion and splatter.

Screen Grid Modulation

Screen modulation is probably the simplest and most popular form of grid modulation, and the least critical of adjustment. The most satisfactory way to apply the modulating voltage to the screen is through a transformer.

With practical tubes it is necessary to drive the screen somewhat negative with respect to the cathode to get complete cutoff of rf output. For this reason the peak modulating voltage required for 100-percent modulation is usually 10 percent or so greater than the dc screen voltage. The latter, in turn, is approximately half the rated screen voltage recommended by the manufacturer under maximum ratings for radiotelegraph operation. The audio power required for 100-percent modulation is approximately one-fourth the dc power input to the screen in cw operation, but varies somewhat with the operating conditions.

Controlled Carrier

As explained earlier, a limit is placed on the output obtainable from a grid-modulation system by the low rf-amplifier plate efficiency (approximately 33 percent) under unmodulated carrier

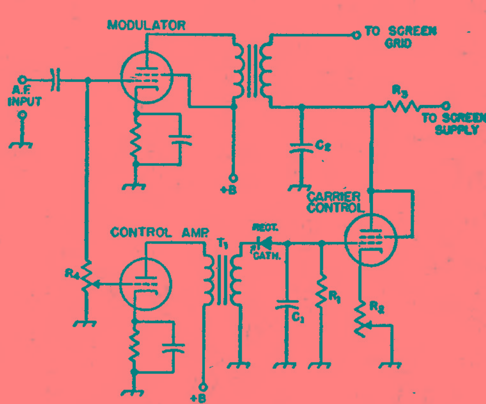


Fig. 12-8 — Circuit for carrier control with screen modulation. A small triode such as the 6C4 can be used as the control amplifier and a 6Y6G is suitable as a carrier-control tube. T1 is an interstage audio transformer having a 1-to-1 or larger turns ratio. R4 is a 0.5-megohm volume control and also serves as the grid resistor for the modulator. A germanium diode may be used as the rectifier. R3 may be the normal screen dropping resistor. C1-R1 and C2-R3 should have a time constant of about 0.1 second.

conditions. The plate efficiency increases with modulation, since the output increases while the dc input remains constant, and reaches a maximum in the neighborhood of 50 percent with 100-percent sine-wave modulation. If the power input to the amplifier can be reduced during periods when there is little or no modulation, thus reducing the plate loss, advantage can be taken of the higher efficiency at full modulation to obtain higher effective output. This can be done by varying the dc power input to the modulated stage in accordance with *average* variations in voice intensity, in such a way as to maintain just sufficient carrier power to keep the modulation high, but not exceeding 100 percent, under all conditions. Thus the carrier amplitude is controlled by the average voice intensity. Properly utilized, controlled carrier permits increasing the carrier output at maximum level to a value about equal to the rated plate dissipation of the tube, twice the output obtainable with constant carrier.

It is desirable to control the power input just enough so that the plate loss, without modulation, is safely below the tube rating. Excessive control is disadvantageous because the distant receiver's avc system must continually follow the variations in average signal level. The circuit of Fig. 12-8 permits adjustment of both the maximum and minimum power input, and separates the functions of modulation and carrier control. A portion of the audio voltage at the modulator grid is applied to a Class A "control amplifier," which drives a rectifier circuit to produce a dc voltage negative with respect to ground. C1 filters out the audio variations, leaving a dc voltage proportional to the average voice level. This voltage is applied to the grid of a "clamp" tube to control the dc screen voltage and thus the rf carrier level.

DOUBLE-SIDEBAND GENERATORS

The a-m carrier can be suppressed or nearly eliminated by using a balanced modulator. The basic principle in any balanced modulator is to introduce the carrier in such a way that it does not appear in the output but so that the sidebands will. This requirement is satisfied by introducing the audio in push-pull and the rf drive in parallel, and connecting the output in push-pull. Balanced modulators can also be connected with the rf drive and audio inputs in push-pull and the output in parallel with equal effectiveness.

Vacuum-tube balanced modulators can be operated at high power levels and the double-sideband output can be used directly into the antenna. Past issues of *QST* have given construction details on such transmitters (see, for example, Rush, "180-Watt D.S.B. Transmitter," *QST* July, 1966). A dsb signal can be copied by the same methods that are used for single-sideband signals, provided the receiver has sufficient selectivity to reject one of the sidebands. In any balanced-modulator circuit, no rf output will exist with no audio signal. When audio is applied, the balance of the modulator is upset so that sum and difference frequencies (sidebands) appear at the output. Further information on balanced modulators is presented in Chapter 13.

CHECKING A-M PHONE OPERATION

USING THE OSCILLOSCOPE

Proper adjustment of a phone transmitter is aided immeasurably by the oscilloscope. The scope will give more information, more accurately, than almost any collection of other instruments that might be named. Furthermore, an oscilloscope that is entirely satisfactory for the purpose is not necessarily an expensive instrument; the cathode-ray tube and its power supply are about all that are needed. Amplifiers and linear sweep circuits are by no means necessary.

In the simplest scope circuit, radio-frequency voltage from the modulated amplifier is applied to the vertical deflection plates of the tube, usually through blocking capacitors, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the instantaneous amplitude of the audio signal varies, the rf output of the transmitter likewise varies, and this produces a wedge-shaped pattern or trapezoid on the screen. If the oscilloscope has a built-in horizontal sweep, the rf voltage can be applied to the vertical plates as before, and the sweep will produce a pattern that follows the modulation envelope of the transmitter output, provided the sweep frequency is lower than the modulation frequency. This produces a wave-envelope modulation pattern.

The Wave-Envelope Pattern

The connections for the wave-envelope pattern are shown in Fig. 12-9A. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a low-impedance (coax, twisted pair, etc.) line and pickup coil. As shown in the alternative drawing, a resonant circuit tuned to the operating frequency may be connected to the vertical plates, using link coupling between it and the transmitter. This will eliminate rf harmonics, and the tuning control provides a means for adjustment of the pattern height.

If it is inconvenient to couple to the final tank coil, as may be the case if the transmitter is tightly shielded, the pickup loop may be coupled to the tuned tank of a matching circuit or antenna coupler. Any method (even a short antenna coupled to the tuned circuit shown in the "alternate input connections" of Fig. 12-9A) that will pick up enough rf to give a suitable pattern height may be used.

The position of the pickup coil should be varied until an unmodulated carrier pattern, Fig. 12-10A, of suitable height is obtained. The horizontal sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. When voice modulation is applied, a rapidly changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated 100 percent. This is illustrated by Fig. 12-10C.

If the height is greater than twice the unmodulated carrier amplitude, as illustrated in Fig. 12-10D, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 percent.

The Trapezoidal Pattern

Connections for the trapezoid or wedge pattern as used for checking a-m are shown in Fig. 12-9B. The vertical plates of the CR tube are coupled to the transmitter tank through a pickup loop, preferably using a tuned circuit, as shown in the upper drawing, adjustable to the operating frequency. Audio voltage from the modulator is applied to the horizontal plates through a voltage divider, R1-R2. This voltage should be adjustable so a suitable pattern width can be obtained; a 0.25-megohm volume control can be used at R2 for this purpose.

The resistance required at R1 will depend on the dc voltage on the modulated element. The total resistance of R1 and R2 in series should be about 0.25 megohm for each 100 volts. For example, if a plate-modulated amplifier operates at 1500 volts, the total resistance should be 3.75 megohms, 0.25 megohm at R2 and the remainder, 3.5 megohms, in R1. R1 should be composed of individual resistors not larger than 0.5 megohm each, in which case 1-watt resistors will be satisfactory.

For adequate coupling at 100 Hz, the capacitance in microfarads of the blocking capacitor, C, should be at least .05/R, where R is the total resistance (R1 + R2) in megohms. In the example above, where R is 3.75 megohms, the capacitance should be .05/3.75 = .013 μF or more. The voltage rating of the capacitor should be at least twice the dc voltage applied to the modulated element.

Trapeziodal patterns for various conditions of modulation are shown in Fig. 12-10, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line should be adjusted, by means of the pickup-coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher

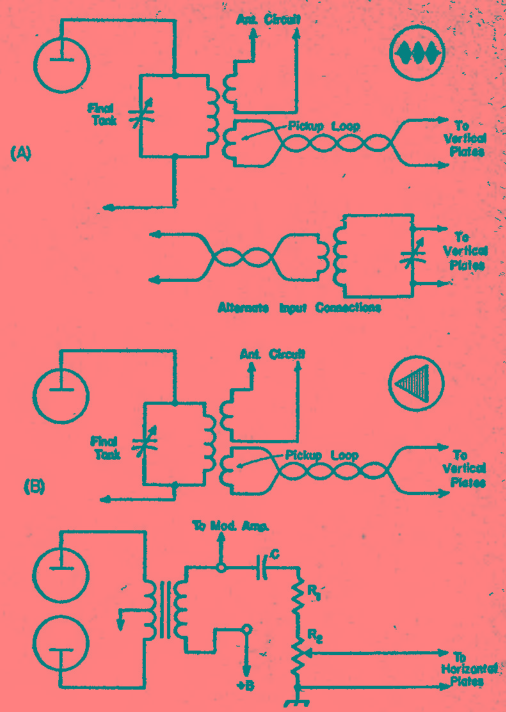


Fig. 12-9 — Methods of connecting the oscilloscope for modulation checking. A — connections for wave-envelope pattern with any modulation method; B — connections for trapezoidal pattern with plate or screen modulation.

the modulation percentage, the wider and more pointed the wedge becomes. At 100-percent modulation it just makes a point at one end of the horizontal axis, and the height at the other end is equal to twice carrier height. Overmodulation in the upward direction is indicated by increased height, at one end, and downward by an extension along the horizontal axis at the pointed end.

CHECKING A-M TRANSMITTER PERFORMANCE

The trapezoidal pattern is generally more useful than the wave-envelope pattern for checking the operation of the phone transmitter. However, both types of patterns have their special virtues, and the best test setup is one that makes both available. The trapezoidal pattern is better adapted to showing the performance of a modulated amplifier from the standpoint of inherent linearity, without regard to the wave form of the audio modulating signal, than is the wave-envelope pattern. Distortion in the audio signal also can be detected in the trapezoidal pattern, although experience in analyzing scope patterns is required to recognize it.

If the wave-envelope pattern is used with a sine-wave audio modulating signal, distortion in the

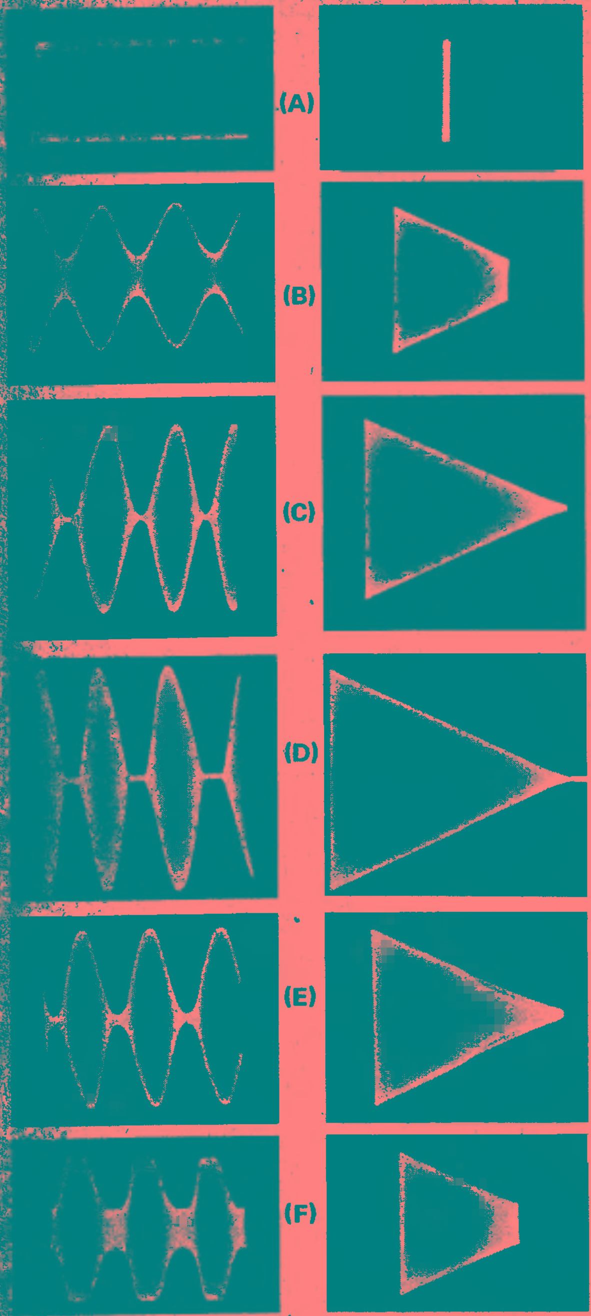


Fig. 12-10 — Oscilloscope patterns showing various forms of modulation of an rf amplifier. At left, wave-envelope patterns; at right, corresponding trapezoidal patterns. The wave-envelope patterns were obtained with e linear oscilloscope sweep having a frequency one-third that of the sine-wave audio modulating frequency, so that three cycles of tha modulation envelopa may be seen. Shown at A is an unmodulated carrier, at B approximately 50-percent moduletion, and at C, 100-percent modulation. The photos at D show modulation in excess of 100 percent. E and F show the results of improper operation or circuit design. See text.

modulation envelope is easily recognizable; however, it is difficult to determine whether the distortion is caused by lack of linearity of the rf stage or by af distortion in the modulator. If the trapezoidal pattern shows good linearity in such a case, the trouble obviously is in the audio system. It is possible, of course, for both defects to be present simultaneously. If they are, the rf amplifier should be made linear first; then any distortion in the modulation envelope will be the result of improper operation in the speech amplifier or modulator, or in coupling the modulator to the modulated rf stage.

Rf Linearity

The trapezoidal pattern is a graph of the modulation characteristic of the modulated amplifier. The sloping sides of the wedge show the rf amplitude for every value of instantaneous modulating voltage. If these sides are perfectly straight lines, the modulation characteristic is linear. If the sides show curvature, the characteristic is nonlinear to an extent shown by the degree to which the sides depart from perfect straightness. This is true regardless of the modulating wave form. If these edges tend to bend over toward the horizontal at the maximum height of the wedge, the amplifier is "flattening" on the modulation uppeaks. This is usually caused by attempting to get too large a carrier output, and can be corrected by tighter coupling to the antenna or by a decrease in the dc screen voltage. The slight "tailing off" at the modulation downpeak (point of the wedge) can be minimized by careful adjustment of excitation and plate loading.

Several types of improper operation are shown in Fig. 12-10. The patterns at E show the effect of a too long time constant in the screen circuit, in an amplifier getting its screen voltage through a dropping resistor, both plate and screen being modulated. The "double-edged" pattern is the result of audio phase shift in the screen circuit combined with varying screen-to-cathode resistance during modulation. This effect can be reduced by reducing the screen bypass capacitance, and also by connecting resistance (to be determined experimentally, but of the same order as the screen dropping resistance) between screen and cathode.

The pictures at the bottom, F, show the effect of insufficient audio power. Although the trapezoidal pattern shows good linearity in the rf amplifier, the wave-envelope pattern shows flattened peaks (both positive and negative) in the modulation envelope even though the audio signal applied to the amplifier was a sine wave. More speech-amplifier gain merely increases the flattening without increasing the modulation percentage in such a case. The remedy is to use a larger modulator or less input to the modulated rf stage. In some cases the trouble may be caused by an incorrect modulation-transformer turns ratio, causing the modulator to be overloaded before its maximum power output capabilities are reached.

GENERAL-PURPOSE AMPLITUDE MODULATORS

The two modulator circuits shown in Figs. 12-11 and 12-12 can be employed to deliver from 3 to 70 watts of audio power. The basic designs are taken from RCA's *Audio Design Phase 2.* The complementary-symmetry circuit, Fig. 12-11, is characterized by a Class A driver and a complementary pair (npn/pnp) of output transistors. The primary advantages of this circuit are simplicity and economy. Common conduction is minimized because the transistor which is "off" during half of the audio cycle is reverse biased. The output transistors are operated at zero bias, providing excellent dc stability. Elaborate regulated power supplies are not required. The complementary-symmetry amplifier is limited to about 20 watts output because of the high level of heat that the driver stage must dissipate. Component values and transistor types are given in Table 12-I for 3-, 5-, 12-, and 20-watt designs.

For higher power levels, the quasi-complementary circuit (Fig. 12-12) is usually chosen. Here a Class A predriver feeds a Class B npn/pnp driver

TABLE 12-I

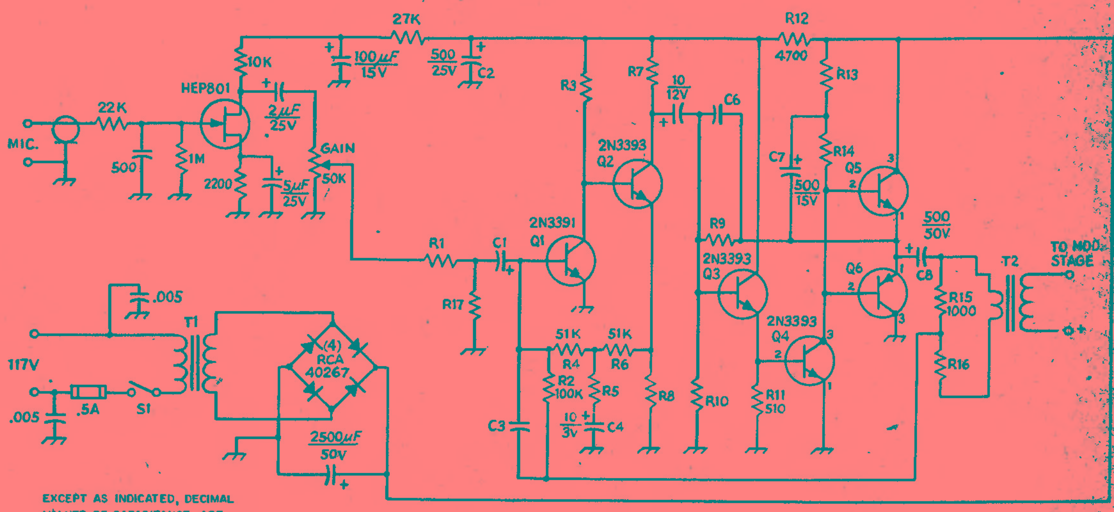
PARTS VALUES FOR COMPLEMENTARY – SYMMETRY CIRCUIT

Power (Watts)	R1	R3	R5	R7	R8	R9	R10	R13	R14	R16	R17	C1 µF	C3 pF	C6 pF	Q4	Q5	Q6	T1	
3	91k	68k	2.7k	3.9k	620	33k	5.6k	120	150	22	22k	0.1/6V	10	100	40611	40610	40609	15V, 1A	(Stancor TP-4)
5	51k	68k	3.3k	3.9k	620	27k	3.6k	110 (1W)	110 (1W)	27	22k	0.25/6V	5	150	40616	40615	40614	17V, 1A	(Stancor TP-4)
12	16k	91k	7.5k	2.7k	390	18k	1.8k	91 (2W)	91 (2W)	56	–	1/6V	10	220	40389	40622	40050	25V, 1A	(Stancor TP-4)
20	8.2k	91k	8.2k	2.2k	360	22k	1.3k	100 (2W)	100 (2W)	100	–	2/6V	10	270	40628	40627	40626	32V, 1A	(C. P. Elec. 10596)

TABLE 12-II

PARTS VALUES FOR QUASI-COMPLEMENTARY-SYMMETRY CIRCUIT

Power (Watts)	R3	R7	R8	R10	R11	R22 R23	Q4	Q5	Q6 Q7	T1		
25	12k	680	1800	2200	270	0.43(5W)	2N3568	2N3638	40632	37V	1.5A	(C. P. Elec. 10596)
40	15k	560	2200	2700	390	0.39(5W)	40635	40634	40633	46V	2A	(C. P. Elec. 10596)
70	18k	470	2700	3300	470	0.33(5W)	40594	40595	40636	60V	2.5A	(C. P. Elec. 10598)



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS; k = 1000, M = 1000 000

Fig. 12-11 — General-purpose amplitude modulator for 3 to 20 watts of audio power. Capacitors with polerity indiceted are electrolytic. See Table 12-I for parts not listed below.
S1 — Spst toggle.
T2 — See text.

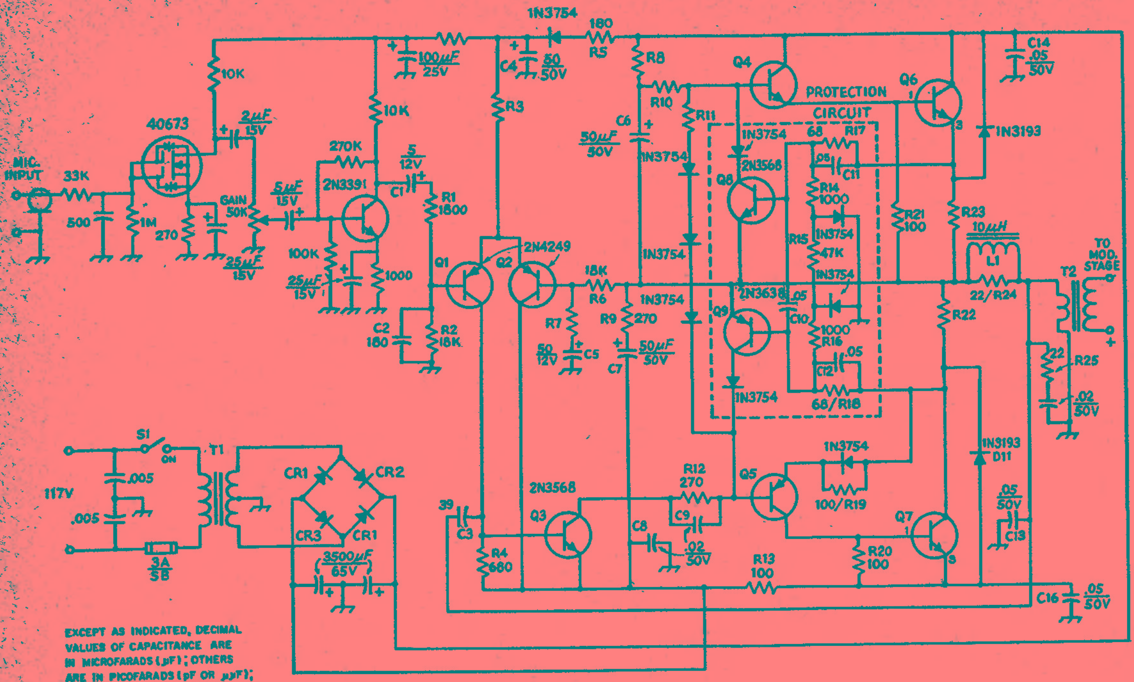


Fig. 12-12 – General-purpose amplitude modulator for 25 to 70 watts of audio power. Capacitors with polarity indicated are electrolytic. See Table 12-II for parts not listed below.

L1 – J. W. Miller 4622 or equiv.
S1 – Spst toggle.
T2 – See text.

pair, which, in turn, activates the npn output transistors. The danger of damage to the output stage from a short circuit is high, so protection is included. Table 12-II includesparts information for three power levels: 25, 40, and 70 watts.

All amplifiers are designed for an 8-ohm output, so T2 can be a standard audio output transformer in "reverse." The secondary impedance will depend on the impedance of the stage to be modulated.

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Single-Sideband Transmission

GENERATING THE SSB SIGNAL

A fully modulated a-m signal has two thirds of its power in the carrier and only one third in the sidebands. The sidebands carry the intelligence to be transmitted; the carrier "goes along for the ride" and serves only to demodulate the signal at the receiver. By eliminating the carrier and transmitting only the sidebands, or just one sideband, the available transmitter power is used to greater advantage. To recover the intelligence being transmitted, the carrier must be reinserted at the receiver, but this is no great problem with a proper detector circuit.

Assuming that the same final-amplifier tube or tubes are used either for normal a-m or for single sideband, carrier suppressed, it can be shown that the use of ssb can give an effective gain of up to 9 dB over a-m – equivalent to increasing the transmitter power 8 times. Eliminating the carrier also eliminates the heterodyne interference that so often spoils communication in congested phone bands.

Filter Method

Two basic systems for generating ssb signals are shown in Fig. 13-2. One involves the use of a bandpass filter having sufficient selectivity to pass one sideband and reject the other. Mechanical filters are available for frequencies below 1 MHz. From 0.2 to 10 MHz, good sideband rejection can be obtained with filters using four or more quartz crystals. Oscillator output at the filter frequency is combined with the audio signal in a balanced modulator, and only the upper and lower sidebands appear in the output. One of the sidebands is passed by the filter and the other rejected, so that an ssb signal is fed to the mixer. The signal is mixed with the output of a high-frequency rf oscillator to produce the desired output frequency. For additional amplification a linear rf amplifier must be used. When the ssb signal is generated around 500 kHz it may be necessary to convert twice to reach the operating frequency, since this simplifies the problem of rejecting the "image" frequencies resulting from the heterodyne process. The problem of image frequencies in the frequency conversions of ssb signals differs from the problem in receivers because the beating-oscillator frequency becomes important. Either balanced mixers or sufficient selectivity must be used to attenuate these frequencies in the output and hence minimize the possibility of unwanted radiations. (Examples of filter-type exciters can be found in various issues of *QST* and in *Single Sideband for the Radio Amateur*.)

Phasing Method

The second system is based on the phase relationships between the carrier and sidebands in a modulated signal. As shown in the diagram, the audio signal is split into two components that are identical except for a phase difference of 90 degrees. The output of the rf oscillator (which may be at the operating frequency, if desired) is likewise split into two separate components having a 90-degree phase difference. One rf and one audio component are combined in each of two separate balanced modulators. The carrier is suppressed in the modulators, and the relative phases of the sidebands are such that one sideband is balanced out and the other is augmented in the combined output. If the output from the balanced modulators is high enough, such an ssb exciter can work directly into the antenna, or the power level can be increased in a following amplifier.

Generally, the filter-type exciter is easier to adjust than is the phasing exciter. Most home built ssb equipment uses commercially made filters these days. The alignment is done at the factory, thus relieving the amateur of the sometimes tedious task of adjusting the filter for suitable bandpass characteristics. Filter-type exciters are more popular than phasing units and offer better carrier suppression and alignment stability. It is still practical for the builder to fabricate his own crystal-lattice filter by utilizing low-cost surplus crystals. This possibility should not be overlooked if the builder is interested in keeping the overall cost of the home-built exciter at a minimum.

BALANCED MODULATORS

The carrier can be suppressed or nearly eliminated by using a balanced modulator or an extremely sharp filter. In ssb transmitters it is common practice to use both devices. The basic principle of any balanced modulator is to

Fig. 13-1 — Single sideband is the most popular of all the modes for amateur hf communication.

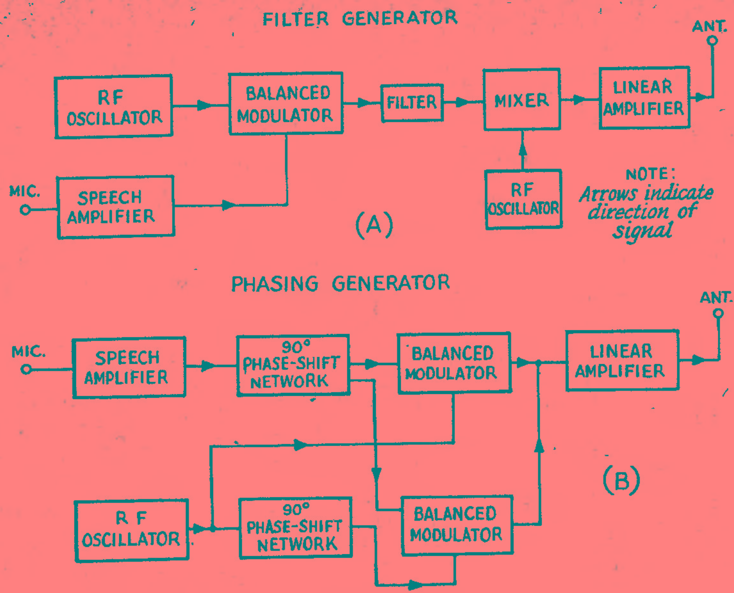


Fig. 13-2 — Two basic systems for generating single-sideband suppressed carrier signals.

introduce the carrier in such a way that it does not appear in the output, but so that the sidebands will. The type of balanced-modulator circuit chosen by the builder will depend upon the constructional considerations, cost, and whether diodes or transistors are to be employed.

In any balanced-modulator circuit there will be no output with no audio signal. When audio is applied, the balance is upset, and one branch will conduct more than the other. Since any modulation process is the same as "mixing" in receivers, sum and difference frequencies (sidebands) will be generated. The modulator is not balanced for the sidebands, and they will appear in the output.

In the rectifier-type balanced modulators shown in Fig. 13-3, at A and B, the diode rectifiers are connected in such a manner that, if they have equal forward resistances, no rf can pass from the carrier source to the output circuit via either of the two possible paths. The net effect is that no rf energy appears in the output. When audio is applied, it unbalances the circuit by biasing the diode (or diodes) in one path, depending upon the instantaneous polarity of the audio, and hence some rf will appear in the output. The rf in the output will appear as a double-sideband suppressed-carrier signal.

In any diode modulator, the rf voltage should be at least 6 to 8 times the peak audio voltage for minimum distortion. The usual operation involves a fraction of a volt of audio and several volts of rf. Desirable diode characteristics for balanced modulator and mixer service include: low noise, low forward resistance, high reverse resistance, good temperature stability, and fast switching time (for high-frequency operation). Fig. 13-4 lists the different classes of diodes, giving the ratio of forward-to-reverse resistance of each. This ratio is an important criterion in the selection of diodes. Also, the individual diodes used should have closely matched forward and reverse resistances; an ohmmeter can be used to select matched pairs or quads.

One of the simplest diode balanced modulators in use is that of Fig. 13-3A. Its use is usually limited to low-cost portable equipment in which a high degree of carrier suppression is not vital. A ring balanced modulator, shown in Fig. 13-3B, offers good carrier suppression at low cost. Diodes CR1 through CR4 should be well matched and can be 1N270s or similar. C1 is adjusted for best rf phase balance as evidenced by maximum carrier null. R1 is also adjusted for the best carrier null obtainable. It may be necessary to adjust each control several times to secure optimum suppression.

Varactor diodes are part of the unusual circuit shown in Fig. 13-3C. This arrangement allows single-ended input of near-equal levels of audio and carrier oscillator. Excellent carrier suppression, 50 dB or more, and a simple method of unbalancing the modulator for cw operation are features of this design. CR1 and CR2 should be rated at 20 pF for a bias of -4 V. R1 can be adjusted to cancel any mismatch in the diode characteristics, so it isn't necessary that the varactors be well matched. T1 is wound on a small-diameter toroid core. The tap on the primary winding of this transformer is at the center of the winding.

A bipolar-transistor balanced modulator is shown in 13-3D. This circuit is similar to one used by Galaxy Electronics and uses closely matched transistors at Q1 and Q2. A phase splitter (inverter) Q3, is used to feed audio to the balanced modulator in push-pull. The carrier is supplied to the circuit in parallel and the output is taken in push-pull. CR1 is a Zener diode and is used to stabilize the dc voltage. Controls R1 and R2 are adjusted for best carrier suppression.

The circuit at E offers superior carrier suppression and uses a 7360 beam-deflection tube as a balanced modulator. This tube is capable of

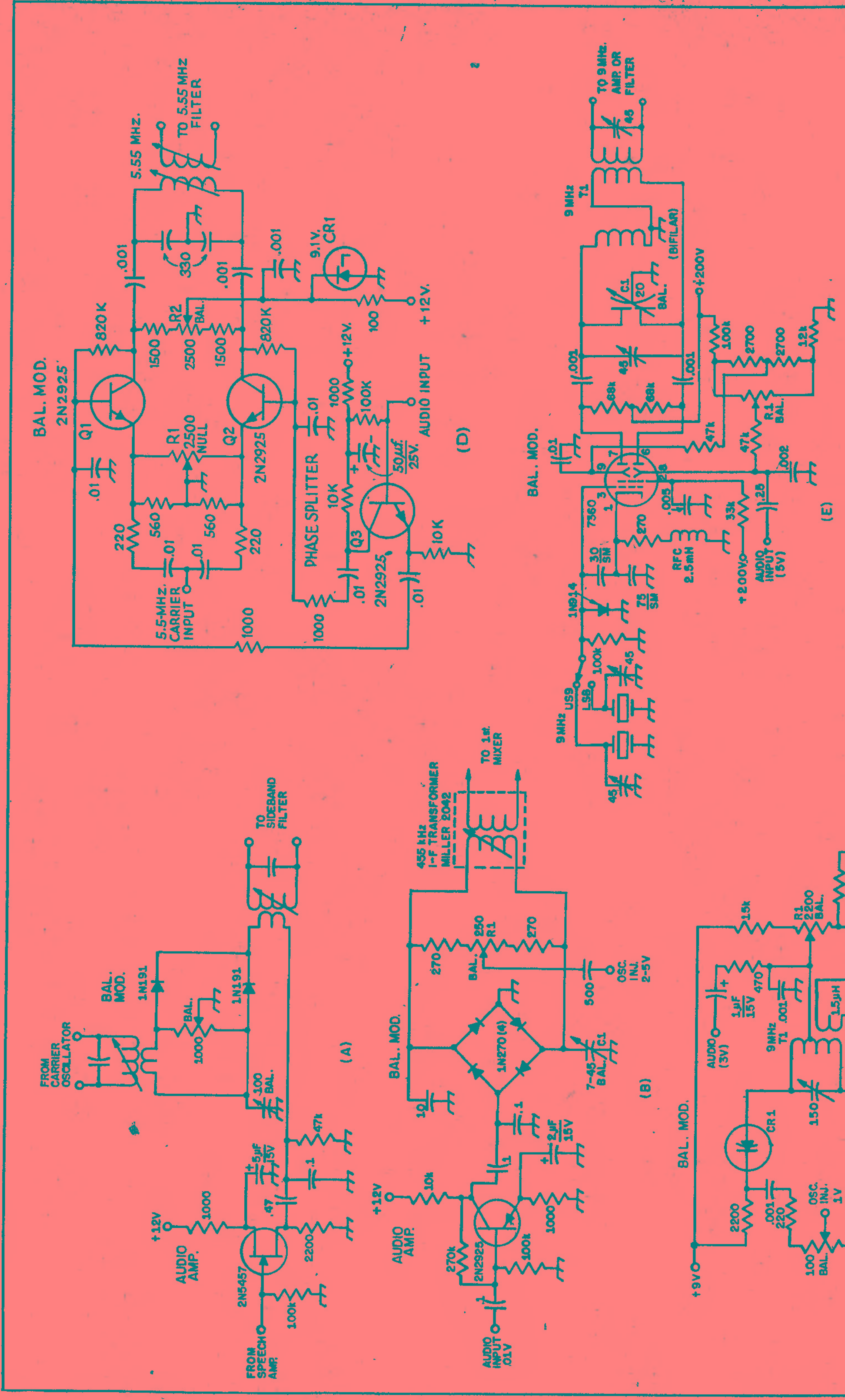


Fig. 13-3 — Typical circuits of balanced modulators. Representative parts values are given and should serve as a basis for designing one's own equipment.

Diode Type	Ratio (M = 1,000,000)
Point-contact germanium (1N98)	500
Small-junction germanium (1N270)	0.1M
Low-conductance silicon (1N457)	48M
High-conductance silicon (1N645)	480M
Hot-carrier (HPA-2800)	2000M

Fig. 13-4 — Table showing the forward-to-reverse resistance ratio for the different classes of solid-state diodes.

providing as much as 60 dB of carrier suppression. When used with mechanical or crystal-lattice filters the total carrier suppression can be as great as 80 dB. Most well-designed balanced modulators can provide between 30 and 50 dB of carrier suppression; hence the 7360 circuit is highly desirable for optimum results. The primary of transformer T1 should be bifilar wound for best results.

IC Balanced Modulators

Integrated circuits (ICs) are presently available for use in balanced-modulator and mixer circuits. A diode array such as the RCA CA3039 is ideally suited for use in circuits such as that of Fig. 13-5A. Since all diodes are formed on a common silicon chip, their characteristics are extremely well matched. This fact makes the IC ideal in a circuit where good balance is required. The hot-carrier diode also has closely matched characteristics and excellent temperature stability. Using broad-band toroidal-wound transformers, it is possible to construct a circuit similar to that of Fig. 13-6 which will have 40 dB of carrier suppression without the need for balance controls. T1 and T2 consist of trifilar windings, 12 turns of No. 32 enam. wire wound on a 1/2-inch toroid core. Another device with good inherent balance is the special IC made for modulator/mixer service, such as the Motorola MC1496G or Signetics S5596. A sample circuit using the MC1496 can be seen in Fig. 13-5B. R1 is adjusted for best carrier balance. The amount of energy delivered from the carrier generator effects the level of carrier suppression; 100 mV of injection is about optimum, producing up to 55 dB of carrier suppression. Additional information on balanced modulators and other ssb-generator circuits is given in the texts referenced at the end of this chapter.

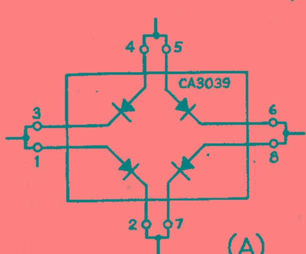
FILTERS

A home-built crystal lattice filter suitable for use in an ssb generator is shown in Fig. 13-7. This unit is composed of three half-lattice sections, with 2 crystals in each section, made with surplus hf crystals. The 330-ohm resistor between sections two and three reduces interaction and smooths the passband response. The leakage reactance between the two halves of L2 and L3 is tuned out by the capacitors connected in series with the center taps of these coils. L1 and L4, the input and output coils, resonate with the calculated value of terminating capacitance at 5060 kHz and reflect the needed inductance across the crystals. The 2000-ohm resistors complete the termination.

All the crystals were purchased as 5500-kHz FT-243s and etched to the desired frequencies with hydrofluoric acid. It is best to wash each crystal with soap and water and measure its frequency before etching. The crystals in each set of three should be as close to each other in frequency as possible, and the separation between the two groups should be about 1500 Hz.

Tuning the filter is quite simple since all four adjustements can be peaked for maximum output at a fixed alignment frequency. This frequency should be on the high side of the pass band and can be the carrier frequency used for lower-sideband transmission (5505.5 kHz in the case of the filter described). Using the carrier frequency it is only necessary to unbalance the balanced modulator to obtain a cw alignment signal. Of course, a signal generator and rf-probe-equipped VTVM can also be used. C1, C2, L1 and L4 are adjusted for maximum output.

A slightly better shape factor can be had by detuning the carrier oscillator to a lower alignment frequency corresponding to about the 4-dB-down point on the high-frequency side of the pass band. Fig. 13-8 shows the measured performance of the filter when aligned at 5505.2 kHz. The 6-dB bandwidth is 2750 Hz.



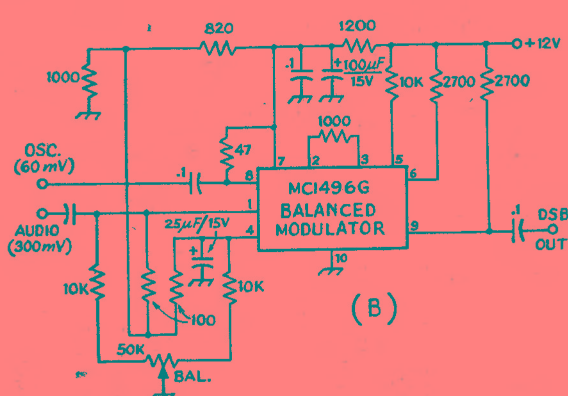


Fig. 13-5 — Additional balanced-modulator circuits in which integrated circuits are used.

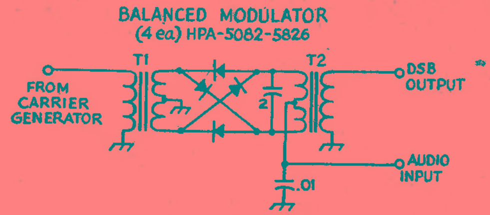


Fig. 13-6 — Balanced modulator design using hot-carrier diodes.

The (suppressed) carrier frequency must be adjusted so that it falls properly on the slope of the filter characteristic. If it is too close to the filter mid frequency the sideband rejection will be poor; if it is too far away there will be a lack of "lows" in the signal.

Ordinarily, the carrier is placed on one side of the curve, depending upon which sideband is desired, which is approximately 20 dB down from the peak. It is sometimes helpful to make provisions for "rubbering" the crystal of the carrier oscillator so that the most natural voice quality can be realized when making initial adjustments.

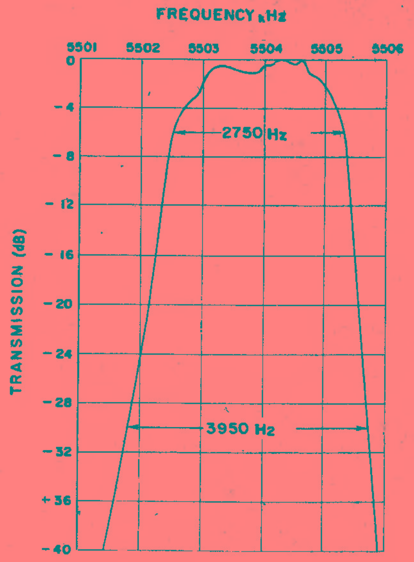


Fig. 13-8 — Measured selectivity characteristic of the filter when aligned at 5505.2 kHz. The 6-dB bandwidth is 2750 Hz and the 30-dB/6-dB shape factor is 1.44.

Using Commercial Crystal Filters

Some builders may not have adequate testing facilities for building and aligning their own filters. In such instances it is possible to purchase ready-made units which are prealigned and come equipped with crystals for upper- and lower-sideband use. Spectrum International[1] has two types for use at 9 MHz. Another manufacturer, McCoy Electronics Co.,[2] sells 9-MHz models for amateur use, and other filters are available surplus.[3]

Mechanical Filters

Mechanical filters contain elements that vibrate and establish resonance mechanically. In crystal filters the coupling between filter sections is achieved by electrical means. In mechanical filters, mechanical couplers are used to transfer the vibrations from one resonant section to the next. At the input and output ends of the filter are transducers which provide for electrical coupling to and from the filter. Most mechanical filters are designed for use from 200 to 600 kHz, the range near 455 kHz being the most popular for amateur use. Mechanical filters suitable for amateur radio circuits are manufactured by the Collins Radio Co. and can be purchased from some dealers in amateur radio equipment.

[1] McCoy Electronics Company, Mt. Holly Springs, PA.

[2] Spectrum International, Topsfield, MA.

[3] E. S. Electronic Labs, 31 Augustus, Excelsior Springs, MO.

FILTER APPLICATIONS

Methods for using typical sideband filters are shown schematically in Fig. 13-9. In the circuit of

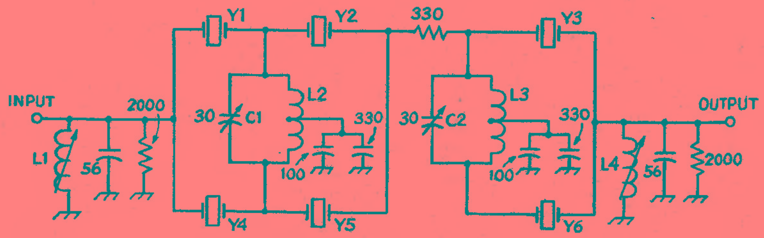


Fig. 13-7 — Circuit diagram of a filter. Resistances are in ohms, and resistors are 1/2-watt composition; capacitors are disk ceremic except as noted.

C1, C2 — Mica trimmer.
L1, L4 — 50 turns No. 38 enamel, close-wound on 17/64-inch dia ceramic slug-tuned form (CTC LS-6, National XR-81 or similar).
L2, L3 — 60 turns No. 38 enamel, close wound on 17/64-inch ceramic form (CTC LS-6, National XR-81 or similar with powered-iron core removed), center tapped.
Y1, Y2, Y3 — All same frequency (near 5500 kHz).
Y4, Y5, Y6 — All same frequency and 1500 to 1700 Hz different from Y1, Y2, Y3.

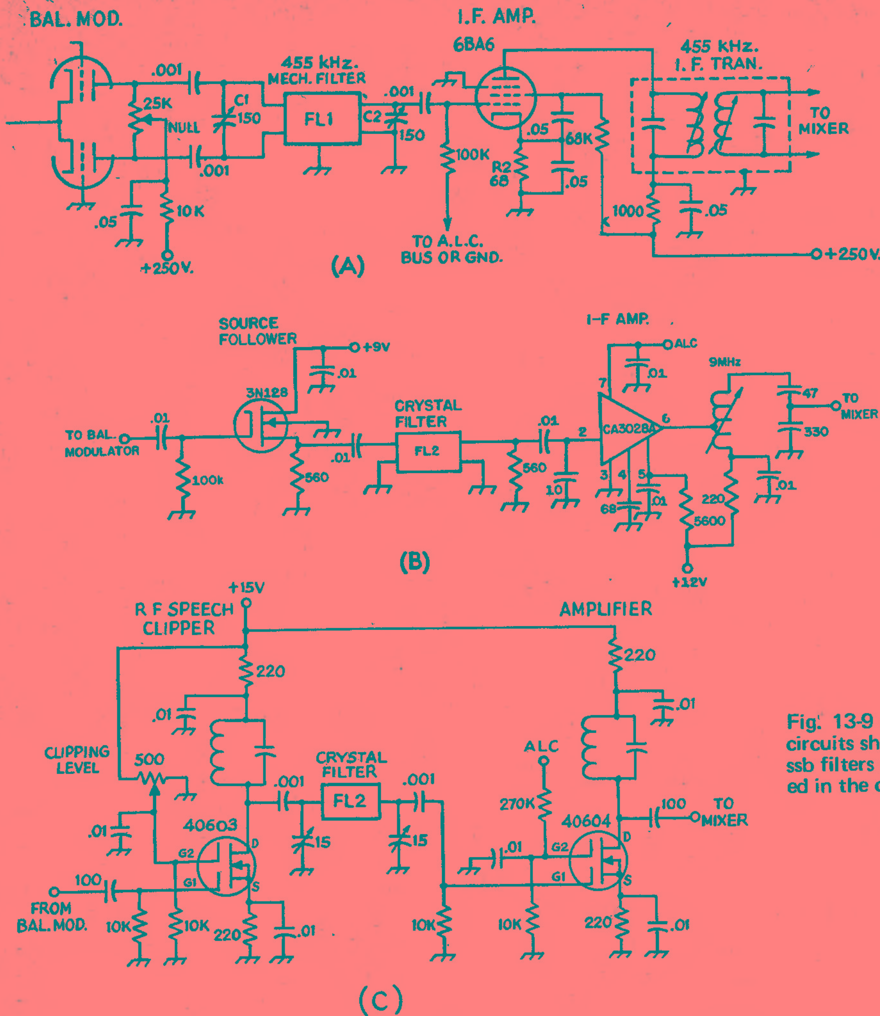


Fig. 13-9 — Typical circuits showing how ssb filters are connected in the circuit.

Fig. 13-9A a 455-kHz mechanical filter is coupled to the balanced modulator by means of two dc isolating capacitors. C1 is used to tune the input of FL1 to resonance (if a Collins type 455-FB-21 is used). Frequently, a fixed-value 120-pF capacitor will suffice at each end of the filter. C2 tunes the output of the filter. A stage of i-f amplification usually follows the filter, as shown, to compensate for the insertion loss of the filter and to provide a stage to which agc can be applied for alc (automatic level control) purposes. In the circuit shown the operator can ground R1 if alc is not used. R2 can be lifted from ground and a 5000-ohm control can be placed between it and ground to provide a means of manual gain control for providing the desired signal level to the mixer.

The circuit of Fig. 13-9B uses a 9-MHz crystal filter, followed by an IC i-f amplifier. Either the McCoy or Spectrum International filters are suitable. Most commercial ssb filters are supplied with a data sheet which shows recommended input and output circuits for matching the impedance of the filter. All are adaptable to use with tubes or transistors.

Another circuit which uses an hf crystal filter, preceded by a dual-gate MOSFET operating as an rf speech clipper, is shown in Fig. 13-9C. The advantages of rf clipping are explained later in this chapter. A second MOSFET amplifies the signal from the filter and provides a variable level of output which is controlled by the alc line.

CARRIER OSCILLATOR

The ssb-generation process starts with a crystal-controlled oscillator, as shown in Fig. 13-2. In a filter-type generator, the oscillator frequency is set on the low-frequency side of the filter bandpass to produce upper sideband and on the upper side when lower-sideband operation is desired. Suitable oscillator circuits are shown in Chapter 6.

MIXER

A single-sideband signal, unlike fm or cw, cannot be frequency multiplied. One or more mixer stages are employed in an ssb exciter to

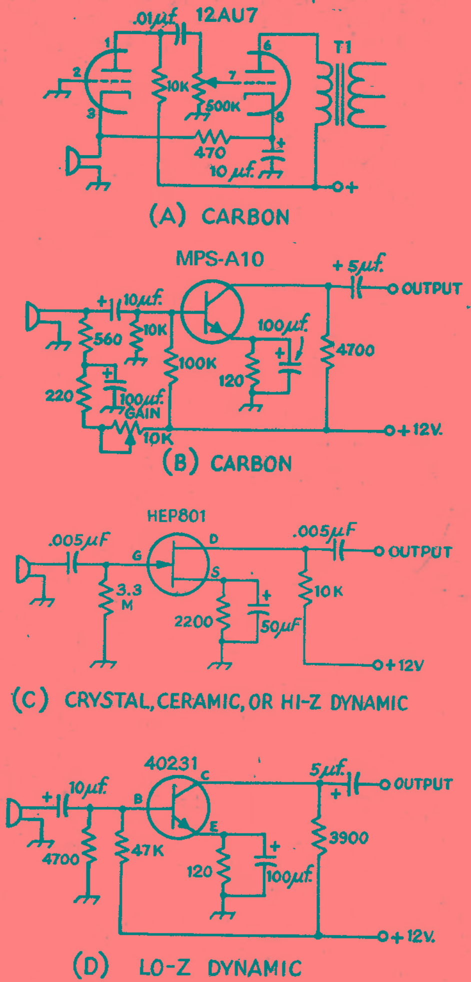


Fig. 13-10 — Speech circuits for use with standard-type microphones. Typical parts values are given.

heterodyne the output of a fixed-frequency ssb generator to the desired operating frequency. See Chapter 8 for details of mixer design and sample mixer circuits.

THE SPEECH AMPLIFIER

The purpose of a speech amplifier is to raise the level of audio output from a microphone to that required by the modulator of a transmitter. In ssb and fm transmitters the modulation process takes place at low levels, so only a few volts of audio are necessary. One or two simple voltage-amplifier stages will suffice. A-m transmitters often employ high-level plate modulation requiring considerable audio power, as described in Chapter 12. The microphone-input and audio voltage-amplifier circuits are similar in all three types of phone transmitters, however.

When designing speech equipment it is necessary to know (1) the amount of audio power the modulation system must furnish and (2) the output voltage developed by the microphone when it is spoken into from normal distance (a few inches) with ordinary loudness. It then becomes possible to choose the number and type of amplifier stages needed to generate the required audio power without overloading or undue distortion anywhere in the system.

MICROPHONES

The level of a microphone is its electrical output for a given sound intensity. The level varies somewhat with the type. It depends to a large extent on the distance from the sound source and the intensity of the speaker's voice. Most commercial transmitters are designed for the median level. If a high-level mike is used, care should be taken not to overload the input amplifier stage. Conversely, a microphone of too low a level must be boosted by a preamplifier.

The frequency response (fidelity) of a microphone is its ability to convert sound uniformly into alternating current. For high articulation it is desirable to reproduce a frequency range of 200-3500 Hz. When all frequencies are reproduced equally, the microphone is considered "flat." Flat response is highly desirable as peaks (sharp rises in the reproduction curve) limit the swing or modulation to the maximum drive voltage, whereas the usable energy is contained in the flat part of the curve.

Microphones are generally omnidirectional, and respond to sound from all directions, or unidirectional, picking up sound from one direction. If a microphone is to be used close to the operator's

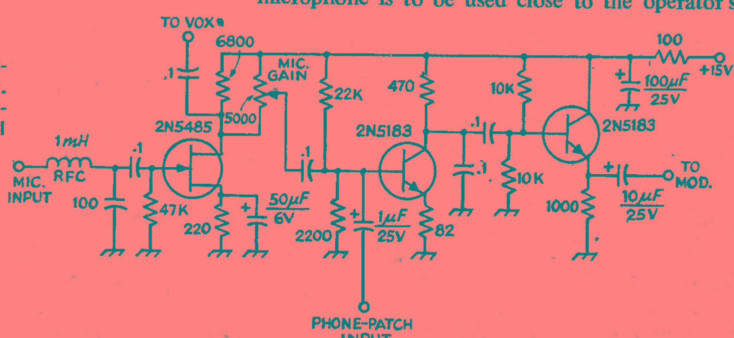


Fig. 13-11 — A resistance-coupled speech amplifier. Component values are representative of a typical circuit.

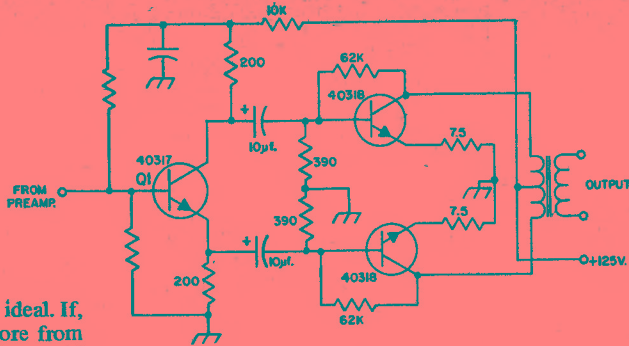


Fig. 13-12 — Typical phase-inverter circuits for transistor amplifier applications.

mouth, an omnidirectional microphone is ideal. If, however, speech is generated a foot or more from the microphone, a unidirectional microphone will reduce reverberation by a factor of 1.7:1. Some types of unidirectional microphones have proximity effect in that low frequencies are accentuated when the microphone is too close to the mouth.

Carbon Microphones

The carbon microphone consists of a metal diaphragm placed against a cup of loosely packed carbon granules. As the diaphragm is actuated by the sound pressure, it alternately compresses and decompresses the granules. When current is flowing through the button, a variable dc will correspond to the movement of the diaphragm. This fluctuating dc can be used to provide grid-cathode voltage corresponding to the sound pressure.

The output of a carbon microphone is extremely high, but nonlinear distortion and instability has reduced its use. The circuit shown in Fig. 13-10 will deliver 20-30 volts at the transformer secondary.

Piezoelectric Microphones

Piezoelectric microphones make use of the phenomena by which certain materials produce a voltage by mechanical stress or distortion of the material. A diaphragm is coupled to a small bar of material such as Rochelle salt or ceramic made of barium titanate or lead zirconium titanate. The diaphragm motion is thus translated into electrical energy. Rochelle-salt crystals are susceptible to high temperatures, excessive moisture, or extreme dryness. Although the output level is higher, their use is declining because of their fragility.

Ceramic microphones are impervious to temperature and humidity. The output level is adequate for most modern amplifiers. They are capacitive devices and the output impedance is high. The load impedance will affect the low frequencies. To provide attenuation, it is desirable to reduce the load to 0.25 megohm or even lower, to maximize performance when operating ssb, thus eliminating much of the unwanted low-frequency response.

Dynamic Microphones

The dynamic microphone somewhat resembles a dynamic loudspeaker. A lightweight coil, usually made of aluminum wire, is attached to a diaphragm. This coil is suspended in a magnetic circuit. When sound impinges on the diaphragm, it moves the coil through the magnetic field generating an alternating voltage.

Electret Microphones

The electret microphone has recently appeared as a feasable alternative to the carbon, piezoelectric or dynamic microphone. An electret is an insulator which has a quasi-permanent static electric charge trapped in or upon it. The electret operates in a condenser fashion which uses a set of biased plates whose motion, caused by air pressure variations, creates a changing capacitance and accompanying change in voltage. The electret acts as the plates would, and being charged, it requires no bias voltage. A low voltage provided by a battery used for an FET impedance converter is the only power required to produce an audio signal.

Electrets traditionally have been susceptable to damage from high temperatures and high humidity. New materials and different charging techniques have lowered the chances of damage, however. Only in extreme conditions (such as 120 degrees F at 90 percent humidity) are problems present. The output level of a typical electret is higher than that of a standard dynamic microphone.

VOLTAGE AMPLIFIERS

The important characteristics of a voltage amplifier are its **voltage gain**, maximum undistorted **output voltage**, and its **frequency response**. The voltage gain is the voltage-amplification ratio of the stage. The output voltage is the maximum af voltage that can be secured from the stage without distortion. The amplifier frequency response should be adequate for voice reproduction; this requirement is easily satisfied.

The voltage gain and maximum undistorted output voltage depend on the operating conditions of the amplifier. The output voltage is in terms of *peak* voltage rather than rms; this makes the rating independent of the waveform. Exceeding the peak value causes the amplifier to distort, so it is more useful to consider only peak values in working with amplifiers.

Resistance Coupling

Resistance coupling generally is used in voltage-amplifier stages. It is relatively inexpensive, good frequency response can be secured, and there

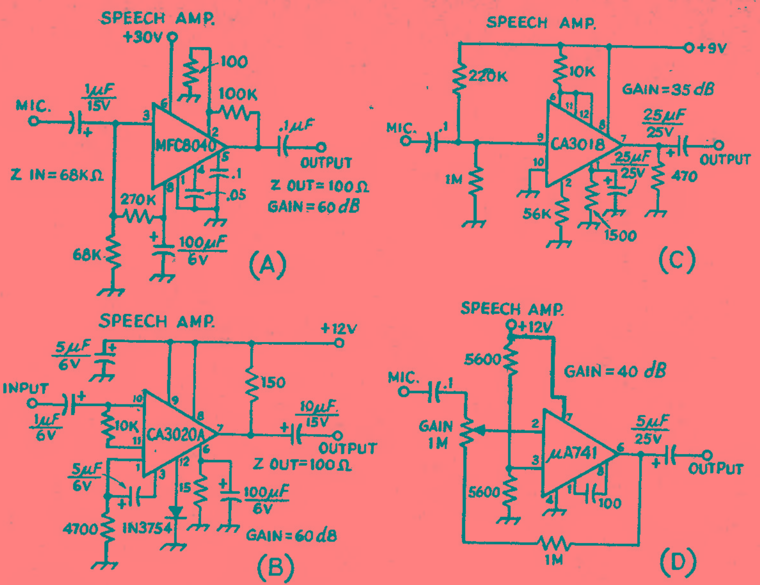


Fig. 13-13 — Typical speech amplifier using integrated circuits.

is little danger of hum pick-up from stray magnetic fields. It is the most satisfactory type of coupling for the output circuits of pentodes and high-μ triodes, because with transformers a sufficiently high load impedance cannot be obtained without considerable frequency distortion. A typical circuit is given in Fig. 13-11.

Phase Inversion

Push-pull output may be secured with resistance coupling by using phase-inverter or phase-splitter circuits as shown in Fig. 13-12. In this circuit the voltage developed across the emitter resistor of Q1 is equal to, but 180 degrees out of phase with, the voltage swing across the collector resistor. Thus, the following two stages are fed equal af voltages. The gain of Q1 will be quite low, if indeed the stage exhibits any gain at all.

Transformer Coupling

Transformer coupling between stages ordinarily is used only when power is to be transferred (in such a case resistance coupling is very inefficient), or when it is necessary to couple between a single-ended and a push-pull stage.

Several types of ICs have been developed for use in speech amplifiers. The Motorola MFC8040 features very low noise, typically 1 μV, (Fig. 13-13A), while the RCA CA3020 has sufficient power output – 500 mW – to drive low-impedance loads (Fig. 13-13B). A transistor IC array can also be put to work in a speech amplifier, as shown in Fig. 13-13C. This circuit uses an RCA CA3018, with a high-gain Darlington pair providing high gain and high input impedance. A second transistor within the IC functions as an emitter follower, for low-impedance output. Most of the operational-amplifier ICs will work as high-gain speech amplifiers, using a minimum of external parts as shown in Fig. 13-13D. The μA741 has internal frequency compensation, but the popular 709 series of operational amplifiers requires external frequency compensation to prevent self-oscillation.

Gain Control

A means for varying the over-all gain of the amplifier is necessary for keeping the final output at the proper level for modulating the transmitter. The common method of gain control is to adjust the value of ac voltage applied to the base or grid of one of the amplifiers by means of a voltage divider or potentiometer.

The gain-control potentiometer should be near the input end of the amplifier, at a point where the signal voltage level is so low there is no danger that the stages ahead of the gain control will be overloaded by the full microphone output. In a high-gain amplifier it is best to operate the first stage at maximum gain, since this gives the best signal-to-hum ratio. The control is usually placed in the input circuit of the second stage.

Remote gain control can also be accomplished with an electronic attenuator IC, such as the Motorola MFC6040. A dc voltage varies the gain of the IC from +6 dB to -85 dB, eliminating the need

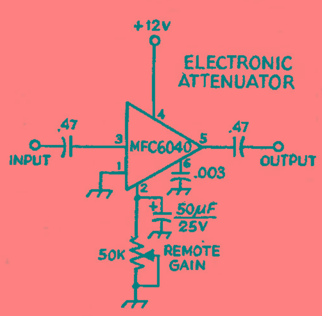


Fig. 13-14 — A dc voltage controls the gain of this IC, eliminating the need for shielded leads to the gain control.

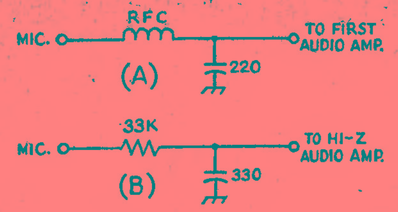


Fig. 13-15 – Rf filters using *LC* (A) and *RC* (B) components, which are used to prevent feedback caused by rf pickup on the microphone lead.

for shielded leads to a remotely located volume control. A typical circuit is shown in Fig. 13-14.

Speech-Amplifier Construction

Once a suitable circuit has been selected for a speech amplifier, the construction problem resolves itself into avoiding two difficulties – excessive hum, and unwanted feedback. For reasonably humless operation, the hum voltage should not exceed about 1 percent of the maximum audio output voltage – that is, the hum and noise should be at least 40 dB below the output level.

Unwanted feedback, if negative, will reduce the gain below the calculated value; if positive, is likely to cause self-oscillation or "howls." Feedback can be minimized by isolating each stage with decoupling resistors and capacitors, by avoiding layouts that bring the first and last stages near each other, and by shielding of "hot" points in the circuit, such as high-impedance leads in low-level stages.

If circuit-board construction is used, high-impedance leads should be kept as short as possible. All ground returns should be made to a common point. A good ground between the circuit board and the metal chassis is necessary. Complete shielding from rf energy is always required for low-level solid-state audio circuits. The microphone input should be decoupled for rf with a filter, as shown in Fig. 13-15. At A, an rf choke with a high impedance over the frequency range of the transmitter is employed. For high-impedance inputs, a resistor may be used in place of the choke.

When using paper capacitors as bypasses, be sure that the terminal marked "outside foil," often indicated with a black band, is connected to ground. This utilizes the outside foil of the capacitor as a shield around the "hot" foil. When paper or mylar capacitors are used for coupling between stages, always connect the outside foil terminal to the side of the circuit having the lower impedance to ground.

DRIVER AND OUTPUT STAGES

Few ssb transmitting mixers have sufficient output to properly drive an output stage of any significant power level. Most modern-day linear amplifiers require at least 30 to 100 watts of exciter output power to drive them to their rated power input level. It follows, then, that an intermediate stage of amplification should be used between the mixer and the pa stage of the exciter.

The vacuum-tube mixers of Chapter 8 will provide 3 to 4 peak volts of output into a high-impedance load. Since most AB_1 exciter output stages need from 25 to 50 volts of swing on their grids for normal operation, it is necessary to employ a driver stage to amplify the mixer output. There are several high-transconductance pentode tubes that work well as drivers. Among them are the 6CL6, the 12BY7, the 6EH7, and the 6GK6. Since all of these tubes are capable of high gain, instability is sometimes encountered during their use. Parasitic suppression should be included as a matter of course, and can take the form of a low-value noninductive resistor in series with the grid, or a standard parasitic choke installed directly at the plate of the tube. Some form of neutralization is recommended and is preferred to resistive loading of the tuned circuits. The latter method lowers the tuned-circuit *Q*. This in turn lowers the stage selectivity and permits spurious responses from the mixer to be passed on to the following stage of the exciter.

A typical driver and PA stage for modern exciters is shown in Fig. 13-16. The PA is set up for AB_1 amplification. The AB_1 mode is preferred because it results in less distortion than does the AB_2 or Class-B modes, and because driving power is not needed for AB_1 operation. A 6146 tube is used but an inexpensive TV sweep tube may be employed if a higher level of IMD is permissible.

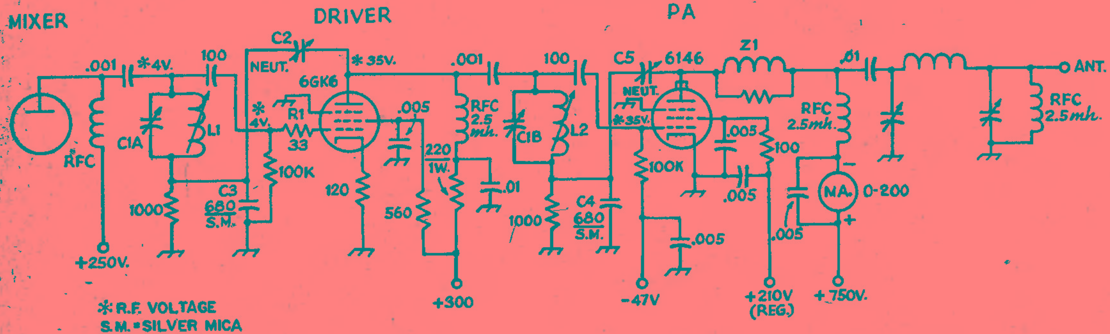


Fig. 13-16 – Schematic diagram of a typical driver and final stage for ssb exciter. Neutralization and parasitic-suppression circuits have been included.

Some sweep tubes are capable of producing less IMD than others, but if not overdriven most of them are satisfactory for ham use. Among the sweep tubes useful as AB_1 amplifiers are the following: 6DQ5, 6GB5, 6GE5, 6HF5, 6JE6, 6JS6, 6KD6, 6KG6, 6LF6 and 6LQ6.

A Practical Circuit

In the circuit of Fig. 13-16, a 6GK6 and a 6146 are shown in a typical driver-amplifier arrangement. Each stage is stabilized by means of R1 in the driver grid, and Z1 in the PA plate, both for parasitic suppression. C2 and C5 are neutralizing capacitors and can take the form of stiff wires placed adjacent to, and in the same plane as the tube anode. Varying the spacing between the neutralizing stubs and the tube envelopes provides the adjustment of these capacitors. Parallel dc feed is used in the mixer and driver stages to prevent the tuned-circuit *Q* from being lowered by dc current flow through L1 and L2. C1A and C1B are ganged, and slug-tuned inductors are used at L1 and L2 to permit tracking of the mixer and driver plate tanks. C3 and C4 form part of the neutralizing circuits. The values shown are suitable for operation on 3.5 MHz but may require modification for use on the other bands. Regulated dc voltage is recommended for the screen grids of the driver and rf stages. Typical rf voltages (measured with a diode rf probe and VTVM are identified with an asterisk. A circuit of this type is capable of up to 60 watts PEP output. For more information on linear amplifiers for sideband service, see Chapter 6.

POWER RATINGS OF SSB TRANSMITTERS

Fig. 13-17 is more or less typical of a few voice-frequency cycles of the modulation envelope of a single-sideband signal. Two amplitude values associated with it are of particular interest. One is the *maximum peak amplitude*, the greatest amplitude reached by the envelope at any time. The other is the *average amplitude*, which is the average of all the amplitude values contained in the envelope over some significant period of time, such as the time of one syllable of speech.

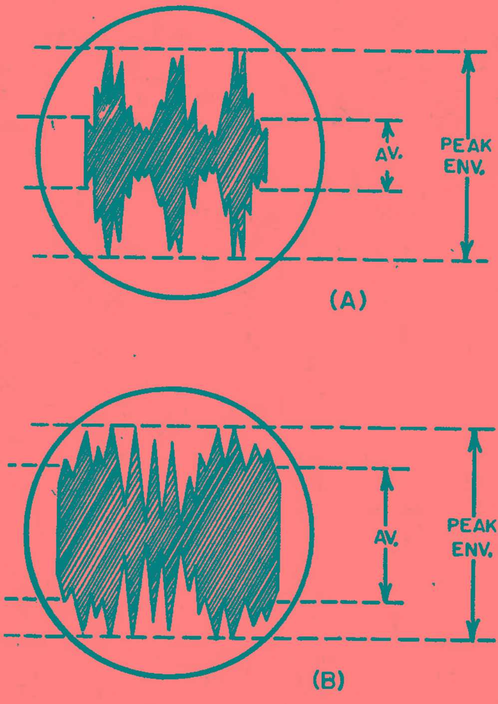


Fig. 13-17 — (A) Typical ssb voica-modulated signal might have an envelope of the general nature shown, where the rf amplitude (current or voltage) is plotted as a function of time, which increases to the right horizontally. (B) Envelope pettern after speech processing to inerease the average level of power output.

The power contained in the signal at the maximum peak amplitude is the basic transmitter rating. It is called the *peak-envelope power*, abbreviated PEP. The peak-envelope power of a given transmitter is intimately related to the distortion considered tolerable. The lower the signal-to-distortion ratio the lower the attainable peak-envelope power, as a general rule. For splatter reduction, an *S/D* ratio of 25 dB is considered a border-line minimum, and higher figures are desirable.

The signal power, *S*, in the standard definition of *S/D* ratio is the power in *one* tone of a two-tone test signal. This is 3 dB below the peak-envelope power in the same signal. Manufacturers of amateur ssb equipment usually base their published *S/D* ratios on PEP, thereby getting an *S/D* ratio that looks 3 dB better than one based on the standard definition. In comparing distortion-product ratings of different transmitters or amplifiers, first make sure that the ratios have the same base.

When the output of an ssb transmitter is viewed on a spectrum analyzer, the display shows the power in the two tones separately, so that the level of distortion products is 6 dB below the level of either tone. However, commercial analyzers usually have a scale over the display tube which is calibrated directly in dB below a single-tone test. Readings may be converted to dB below the PEP level by subtracting 6 dB from the indicated distortion levels.

Peak vs. Average Power

Envelope peaks occur only sporadically during voice transmission, and have no direct relationship with meter readings. The meters respond to the amplitude (current or voltage) of the signal averaged over several cycles of the modulation envelope. (This is true in practically all cases, even though the transmitter rf output meter may be *calibrated* in watts. Unfortunately, such a calibra-

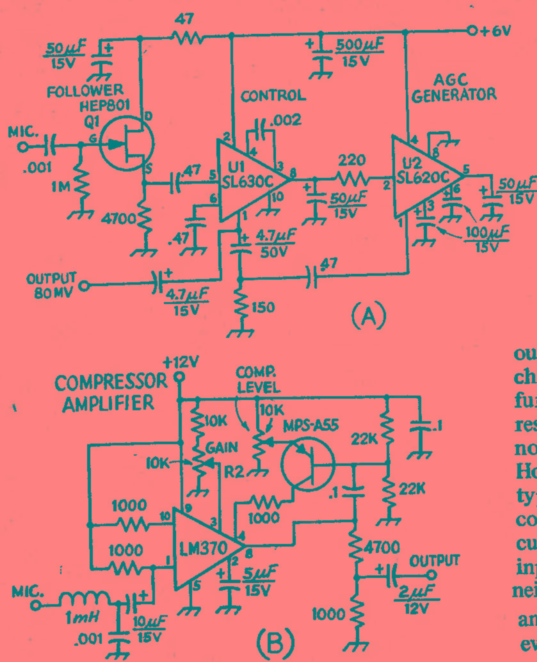


Fig. 13-18 — Typical solid-stata compressor circuit.

tion means little in voice transmission since the meter can be calibrated in watts only by using a sine-wave signal – which a voice-modulated signal definitely is not.)

The ratio of peak-to-average amplitude varies widely with voice of different characteristics. In the case shown in Fig. 13-17A the average amplitude, found graphically, is such that the peak-to-average ratio of amplitudes is almost 3 to 1. The ratio of peak *power* to average *power* is something else again. There is no simple relationship between the meter reading and actual average power, for the reason mentioned earlier.

DC Input

FCC regulations require that the transmitter power be rated in terms of the dc input to the final stage. Most ssb final amplifiers are operated Class AB_1 or AB_2, so that the plate current during modulation varies upward from a "resting" or no-signal value that is generally chosen to minimize distortion. There will be a peak-envelope value of plate current that, when multiplied by the dc plate voltage, represents the instantaneous tube power input required to produce the peak-envelope output. This is the "peak-envelope dc input" or "PEP input." It does not register on any meter in the transmitter. Meters cannot move fast enough to show it – and even if they did, the eye couldn't follow. What the plate meter *does* read is the plate current averaged over several modulation-envelope cycles. This multiplied by the dc plate voltage is the number of watts input required to produce the *average* power output described earlier.

In voice transmission the power input and power output are both continually varying. The power input peak-to-average ratio, like the power-output peak-to-average ratio, depends on the voice characteristics. Determination of the input ratio is further complicated by the fact that there is a resting value of dc plate input even when there is no rf output. *No exact figues are possible.* However, experience has shown that for many types of voices and for ordinary tube operating conditions where a moderate value of resting current is used, the ratio of PEP input to average input (during a modulation peak) will be in the neighborhood of 2 to 1. That is why many amplifiers are rated for a PEP input of 2 kilowatts even though the maximum legal input is 1 kilowatt.

PEP Input

The 2-kilowatt PEP input rating can be interpreted in this way: The amplifier can handle dc peak-envelope inputs of 2 kw, presumably with satisfactory linearity. But it should be run up to such peaks if – and *only* if – in doing so the dc plate current (the current that shows on the plate meter) multiplied by the dc plate voltage does not at any time exceed 1 kilowatt. On the other hand, if your voice has characteristics such that the dc peak-to-average ratio is, for example, 3 to 1, you should not run a greater dc input during peaks than 2000/3, or 660 watts. Higher dc input would drive the amplifier into nonlinearity and generate splatter.

If your voice happens to have a peak-to-average ratio of less than 2 to 1 with this particular amplifier, you cannot run more than 1 kilowatt dc input even though the envelope peaks do not reach 2 kilowatts.

It should be apparent that the dc input rating (based on the *maximum* value of dc input developed during modulation, of course) leaves much to be desired. Its principal virtues are that it can be measured with ordinary instruments, and that it is consistent with the method used for rating the power of other types of emission used by amateurs. The meter readings offer no assurance that the transmitter is being operated within linearity limits, unless backed up by oscilloscope checks using *your* voice.

It should be observed, also, that in the case of a grounded-grid final amplifier, the 1-kilowatt dc input permitted by FCC regulations must include the input to the driver stage as well as the input to the final amplifier itself. Both inputs are measured as described above.

SPEECH PROCESSING

Four basic systems, or a combination thereof, can be used to reduce the peak-to-average ratio, and thus, to raise the average power level of an ssb signal. They are: compression or clipping of the af wave before it reaches the balanced modulator, and compression or clipping of the rf waveform after the ssb signal has been generated. One form of rf compression, commonly called alc (automatic level control) is almost universally used in amateur ssb transmitters. Audio processing is also used to increase the level of audio power contained in the sidebands of an a-m transmitter and to maintain constant deviation in an fm transmitter. Both compression and clipping are used in a-m systems, while most fm transmitters employ only clipping.

Volume Compression

Although it is obviously desirable to keep the voice level as high as possible, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use automatic gain control that follows the *average* (not instantaneous) variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the rectified and filtered dc to a control electrode in an early stage in the amplifier.

A practical example of an audio compressor circuit is shown in Fig. 13-18A. Q1 is employed as an impedance converter, providing coupling between a high-impedance microphone and the input terminal of the Plessey SL630C audio-amplifier IC. Low-impedance microphones can be connected directly to the input of the SL630C. U1 has an agc terminal which allows logarithmic control of the output level with a variable dc voltage. High-frequency cutoff is accomplished by connecting a .002-µF capacitor between pins 3 and 4. Manual gain control is effected by applying a dc voltage to pin 8.

Agc voltage for U1 is developed by the SL620C. A suitable time constant for voice operation is established by the capacitors connected to pins 3, 4 and 6, respectively. The IC provides a fast-attack, slow-decay characteristic for the agc voltage when voice signals are applied and a short burst of agc voltage when a short noise burst occurs. Twenty transistors and four diodes are used in U2.

The compressor will hold the output level constant within 2 dB over a 40-dB range of input signal. The nominal output level is 80 mV; the microphone used should develop at least 3 mV at the gate of Q1.

Fig. 13-18B shows an IC audio compressor circuit using the National Semiconductor LM-370. This IC has two gain-control points, pins 3 and 4; one is used for the input gain adjustment while the other receives agc voltage whenever the output level exceeds a preset norm. R2 establishes the point at which compression starts.

Speech Clipping and Filtering

In speech wave forms the average power content is considerably less than in a sine wave of the same peak amplitude. If the low-energy peaks are clipped off, the remaining wave form will have a considerably higher ratio of average power to peak amplitude. Although clipping distorts the wave form and the result therefore does not sound exactly like the original, it is possible to secure a worthwhile increase in audio power without sacrificing intelligibility. Once the system is properly adjusted *it will be impossible to overdrive* the modulator stage of the transmitter because the maximum output amplitude is fixed.

By itself, clipping generates high-order harmonics and therefore will cause splatter. To prevent this, the audio frequencies above those

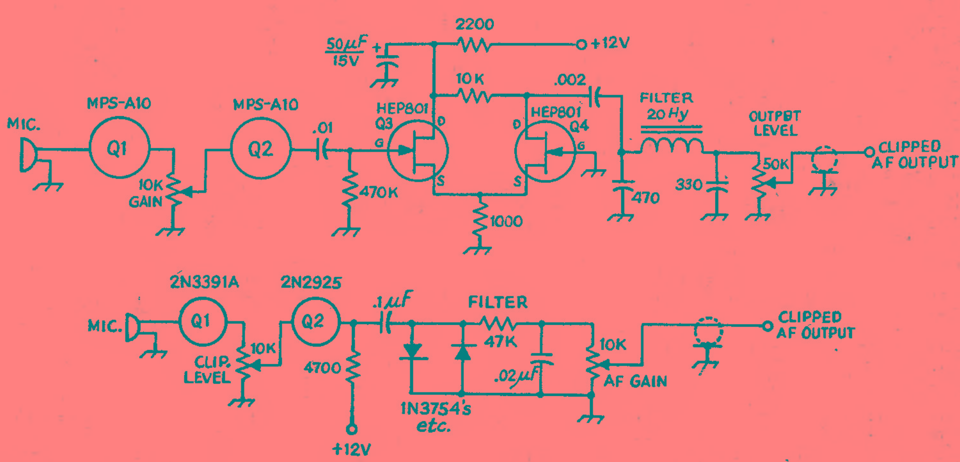


Fig. 13-19 — This drawing illustrates use of JFETs or silicon diodes to clip positive and negative voice peaks.

needed for intelligible speech must be filtered out, *after* clipping and *before* modulation. The filter required for this purpose should have relatively little attenuation below about 2500 Hz, but high attentuation for all frequencies above 3000 Hz.

The values of L and C should be chosen to form a low-pass filter section having a cutoff frequency of about 2500 Hz, using the value of the terminating resistor load resistance. For this cutoff frequency the formulas are:

$$L1 = \frac{R}{7850} \text{ and } C1 = C2 = \frac{63.6}{R}$$

where R is in ohms, L1 in henrys, and C1 and C2 in microfarads.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 to 12 dB) there is almost no change in "quality" but the voice power is increased considerably.

Before drastic clipping can be used, the speech signal must be amplified several times more than is necessary for normal modulation. Also, the hum and noise must be much lower than the tolerable level in ordinary amplification, because the noise in the output of the amplifier increases in proportion to the gain.

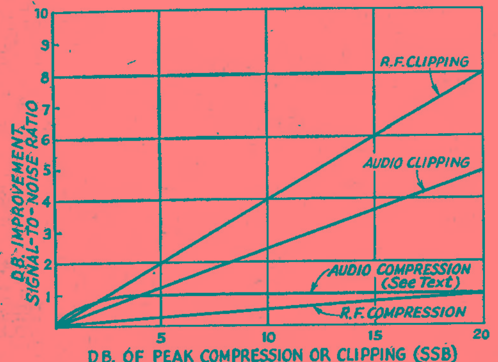


Fig. 13-20 — The improvement in received signal-to-noise ratio achieved by the simple forms of signal processing.

In the circuit of Fig. 13-19B a simple diode clipper is shown following a two-transistor preamplifier section. The 1N3754s conduct at approximately 0.7 volt of audio and provide positive- and negative-peak clipping of the speech wave form. A 47,000-ohm resistor and a .02-μF capacitor follow the clipper to form a simple R-C filter for attenuating the high-frequency components generated by the clipping action, as discussed earlier. Any top-hat or similar silicon diodes can be used in place of the 1N3754s. Germanium diodes (1N34A type) can also be used, but will clip at a slightly lower peak audio level.

SSB SPEECH PROCESSING

Compression and clipping are related, as both have fast attack times, and when the compressor release time is made quite short, the effect on the

(A)
(B)
(C)
(D)

Fig. 13-21 — Two-tone envelope petterns with various degrees of rf clipping. All envelope patterns are formed using tones of 600 and 1000 Hz. (A) At clipping threshold; (B) 5 dB of clipping; (C) 10 dB of clipping; (D) 15 dB of clipping.

wave from approaches that of clipping. Speech processing is most effective when accomplished at radio frequencies, although a combination of af clipping and compression can produce worthwhile results. The advantage of an outboard audio speech processor is that no internal modifications are necessary to the ssb transmitter with which it will be used.

To understand the effect of ssb speech processing, review the basic rf waveforms shown in Fig. 13-17A. Without processing, they have high peaks but low average power. After processing, Fig. 13-17B, the amount of average power has been raised considerably. Fig. 13-20 shows an advantage of several dB for rf clipping (for 20 dB of processing) over its nearest competitor.

Investigations by W6JES reported in *QST* for January, 1969, show that, observing a transmitted signal using 15 dB of audio clipping from a remote receiver, the intelligibility threshold was improved nearly 4 dB over a signal with no clipping.

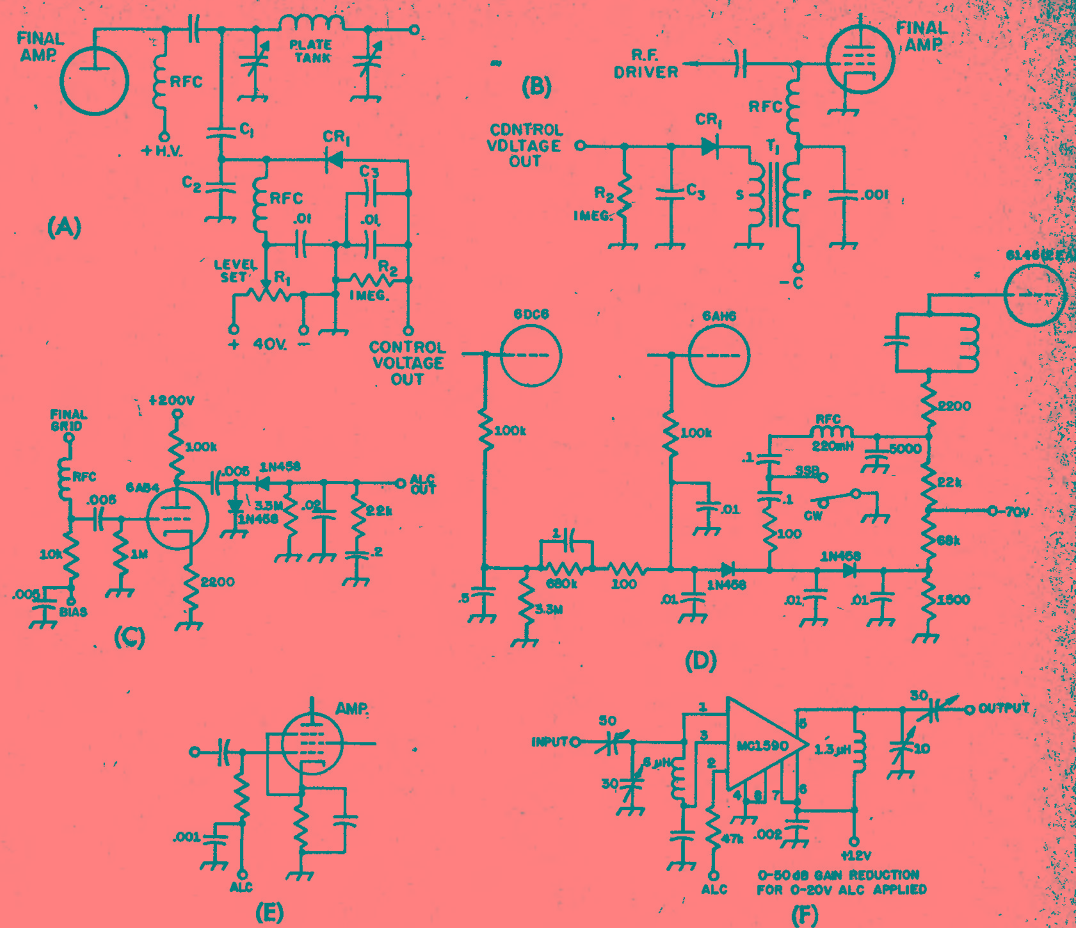


Fig. 13-22 — (A) Control voltage obtained by sampling the rf output voltage of the final amplifier. The diode back bias, 40 volts or so maximum, may be taken from any convenient positive voltage source in the transmitter. R1 may be a linear control having a maximum resistance of the order of 50,000 ohms. CR1 may be a 1N34A or similar germanium diode.
(B) Control voltage obtained from grid circuit of a Class AB_1 tetrode amplifier. T1 is an interstage audio transformer having a turns ratio, secondary to primary, of 2 or 3 to 1. An inexpensive transformer mey be used since the primary and secondary currents are negligible. CR1 may be a 1N34A or similar; time constant R2C3 is discussed in the text.
(C) Control voltage is obtained from the grid of a Class AB_1 tetrode amplifier and amplified by a triode audio stage.
(D) Alc system used in the Collins 32S-3 transmitter.
(E) Applying control voltage to the tube or (F) linear IC controlled amplifier.

Increasing the af clipping level to 25 dB gave an additional 1.5 dB improvement in intelligibility. Audio compression was found to be valuable for maintaining relatively constant average-volume speech, but such a compressor added little to the intelligibility threshold at the receiver, only about 1-2 dB.

Evaluation of rf clipping from the receive side with constant-level speech, and filtering to restore the original bandwidth, resulted in an improved intelligibility threshold of 4.5 dB with 10 dB of clipping. Raising the clipping level to 18 dB gave an additional 4-dB improvement at the receiver, or 8.5-dB total increase. The improvement of the intelligibility of a weak ssb signal at a distant receiver can thus be substantially improved by rf clipping. The effect of such clipping on a two-tone test pattern is shown in Fig. 13-21.

Automatic level control, although a form of rf speech processing, has found its primary application in maintaining the peak rf output of an ssb transmitter at a relatively constant level, hopefully below the point at which the final amplifier is overdriven, when the audio input varies over a considerable range. These typical alc systems, shown in Fig. 13-22, by the nature of their design time constants offer a limited increase in transmitted average-to-PEP ratio. A value in the region of 2-5 dB is typical. An alc circuit with shorter time constants will function as an rf syllabic compressor, producing up to 6 dB improvement in the intelligibility threshold at a distant receiver. The Collins Radio Company uses an alc system with dual time constants (Fig. 13-22D) in their S/Line transmitters, and this has proven to be quite effective.

Heat is an extremely important consideration in the use of any speech processor which increases the average-to-peak power ratio. Many transmitters, in particular those using television sweep tubes, simply are not built to stand the effects of increased average input, either in the final-amplifier tube or tubes or in the power supply. If heating in the final tube is the limiting factor, adding a cooling fan may be a satisfactory answer.

AN AUDIO SPEECH PROCESSOR

Over the years, different speech processing schemes have been employed, with varying degrees of success, to raise the average-to-peak power ratio of a voice signal and improve communications effectiveness. The various methods generally fall into one of two categories – compression or clipping. Described here is a processor which represents a departure from these standard approaches. Processing is done at audio frequencies, but in a unique fashion. The unit is used between the microphone and the transmitter so that no modification to the transmitter is required.

Technical Description

Operation is a consequence of the fact that speech energy resembles an amplitude-modulated signal. The speech waveform represents multiplication of a slowly varying envelope containing energy below 100 Hz with a voice-frequency signal contained mostly between 300 and 3000 Hz. In an analagous fashion, a conventional a-m modulator multiplies an amplitude-varying low-frequency signal (the applied modulation) with a constant-amplitude higher-frequency carrier. Thus the speech waveform envelope corresponds to the a-m modulation and the voice-frequency portion to the carrier. Note that the voice carrier actually varies continuously in frequency, unlike the conventional fixed-frequency a-m case, but is constant in amplitude. The object of this speech processor is to reproduce only the carrier portion of speech. The voice envelope is separated from the voice carrier, and because their respective frequency spectrums are nonoverlapping, the envelope can be filtered leaving only the carrier. (See Fig. 1.)

To separate the envelope and carrier, the speech signal is passed through a logarithmic amplifier which performs the mathematical operation of taking logarithms. By analogy to the a-m model, this signal can be represented as the mathematical product EV, where E represents the envelope and V, the voice carrier, both of which are functions of time. Taking the logarithm produces log EV, but log EV = log E + log V (a well known property of the logarithm). The envelope and carrier components are then separated in terms of their logarithms and it is now a relatively simple matter to process the two components independently. This is something which could not be done up to this point. A high-pass filter with an appropriately chosen cutoff frequency attenuates the envelope waveform but passes the higher-frequency voice carrier. The remaining signal is log V. It goes through an inverse-logarithm amplifier which produces at its output the signal V. The result is the desired constant-amplitude voice carrier.

Circuit Description

Some additional issues arise when one tries to implement the preceding scheme. These will be considered now in a stage-by-stage operational description of the processor. The reader is referred to the block diagram given in Fig. 2 and the circuit shown in diagram Fig. 3. Speech amplifier U1 first boosts the incoming audio signal to a convenient and usable level. Before taking logarithms, however, output from U1 must be full-wave rectified to be all positive since the logarithmic amplifier works only for positive signal input. The logarithm of a number is defined only for positive numbers. U4 and U5 serve as a full-wave rectifier and precedes the logarithmic amplifier, U6 and U7. Matched silicon diodes are recommended for CR1 and CR2. If none are available, individual 1N914 diodes may be substituted. The logarithmic stage separates the voice-frequency and envelope components of the speech waveform, as described above, and the envelope is filtered by an active *RC* high-pass filter, U8.[1] A two-pole Butterworth configuration is used with the lower half-power frequency set at approximately 50 Hz. Those who are experimentally inclined may wish to try lower or higher cutoff frequencies. The expression for cutoff frequency, f_c, in terms of the filter components is:

$$f_c = \frac{1}{2\pi\sqrt{R3\,R2\,C1\,C2}}$$

[1] Because the rectification and logarithmic operations performed upon the original speech signal are nonlinear, the frequency spectrums of the actual envelope and voice carrier signals are, strictly speaking, not exactly the same as those of the signals appearing at the output of the logarithmic amplifier. The main result of these operations is to introduce additional higher-frequency components not present in the original signal. It has been determined, however, that the logarithm of the rectified speech envelope is still primarily low-frequency in nature (mostly far below 100 Hz). This is sufficient to allow the processor to operate as originally described.

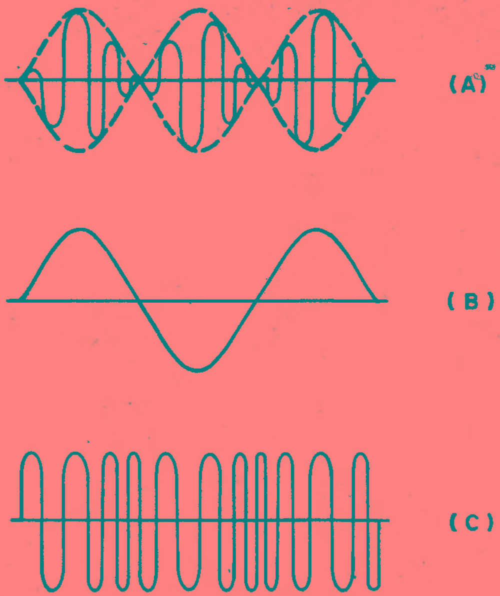


Fig. 1 — A voice signal can be represented as an a-m waveform, which results from multiplication of a relatively slowly varying envelope (B) with a carrier, (C). Note the carrier peak amplitude is constant. The speech processor separates components B and C, and filters out B, leaving the carrier portion only.

where it is required that C1 = C2 and R3 = 2 R2 for proper filter response. Varying the cutoff frequency corresponds to changing the compression level setting on a conventional speech compressor. Lower cutoff frequencies result in reduced "compression." In the original model of this processor, it was found that a filter cutoff frequency of about 400 Hz or higher produced essentially constant-amplitude output from the processor. Harmonic distortion was quite noticeable, however. Thus 50 Hz was chosen as a compromise between maximum "compression" and minimum distortion. The distortion that is inherent in this unit occurs for signals that have considerable energy in the neighborhood of the high-pass filter cutoff frequency. With a setting of 50 Hz, the distortion is quite low. The filtered signal proceeds to an exponential amplifier, U9 and U10. As with the logarithmic amplifier, matched diodes for CR3 and CR4 will produce the best results, but individual 1N914's will serve satisfactorily. The signal at the output of U10 is still in rectified form (all positive). To be converted back to its bipolar form, the signal is multiplied by the correct sign information (either positive or negative). The effect is to invert (make negative) portions of the signal which should be negative, leaving the remaining parts positive. The correct sign information is obtained by hard-limiting the voice signal at the processor input. Output from U1 is further amplified by U2 and then limited by a diode clipper, CR5 and CR6. Because of the very high gain of the U1-U2 cascade, the clipper produces almost pure square-wave output. Thus, any positive input to U1 produces a level of approximately one volt at the output of U2, and any negative input produces a level of about minus one volt. The square-wave output is multiplied with the signal from the exponential amplifier by an analog multiplier, U14. The LM1595 used at U14 produces an output voltage equal to the mathematical product of its two input signals, which in this case are the signals from U2 and U10. The result, then, is to multiply the rectified signal from the exponential amplifier by plus or minus one volt to produce the desired bipolar signal. Output is taken from buffer amplifier U11. The processed signal is passed through a low-pass filter with sharp cutoff above 3 kHz to eliminate unwanted high-frequency energy.

Because the processor is inherently sensitive to even the smallest input signals, undesired background noise or induced ac hum will be processed along with the speech and will appear as a loud disturbance at the output. To help eliminate this problem noise blanker U3 is included in the design. It consists of a free-running multivibrator with square-wave output at about 20 kHz, which is beyond audibility. When this signal is added to the output of the speech amplifier, the effect is to mask, before processing, any noise which is lower in amplitude than the 20-kHz signal.

An audio amplifier, U13, at the output, provides a convenient means of monitoring the processed audio output with low-impedance (eight-ohm) headphones. If high-impedance headphones are to be used, T1 may be omitted and output can be taken directly from pin 6 of U13 through a 5 μF coupling capacitor.

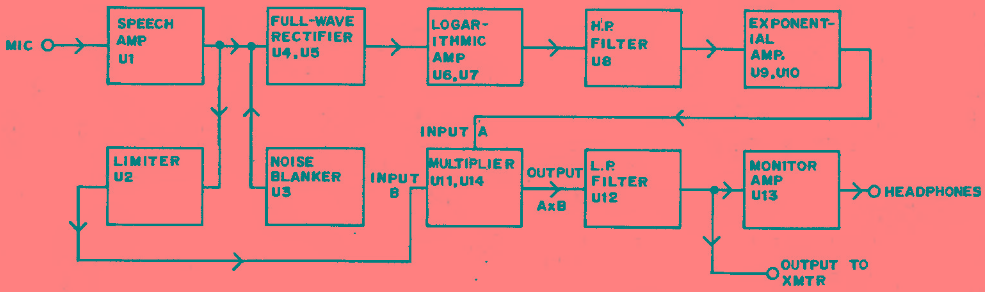


Fig. 2 — Block diagram for the processor.

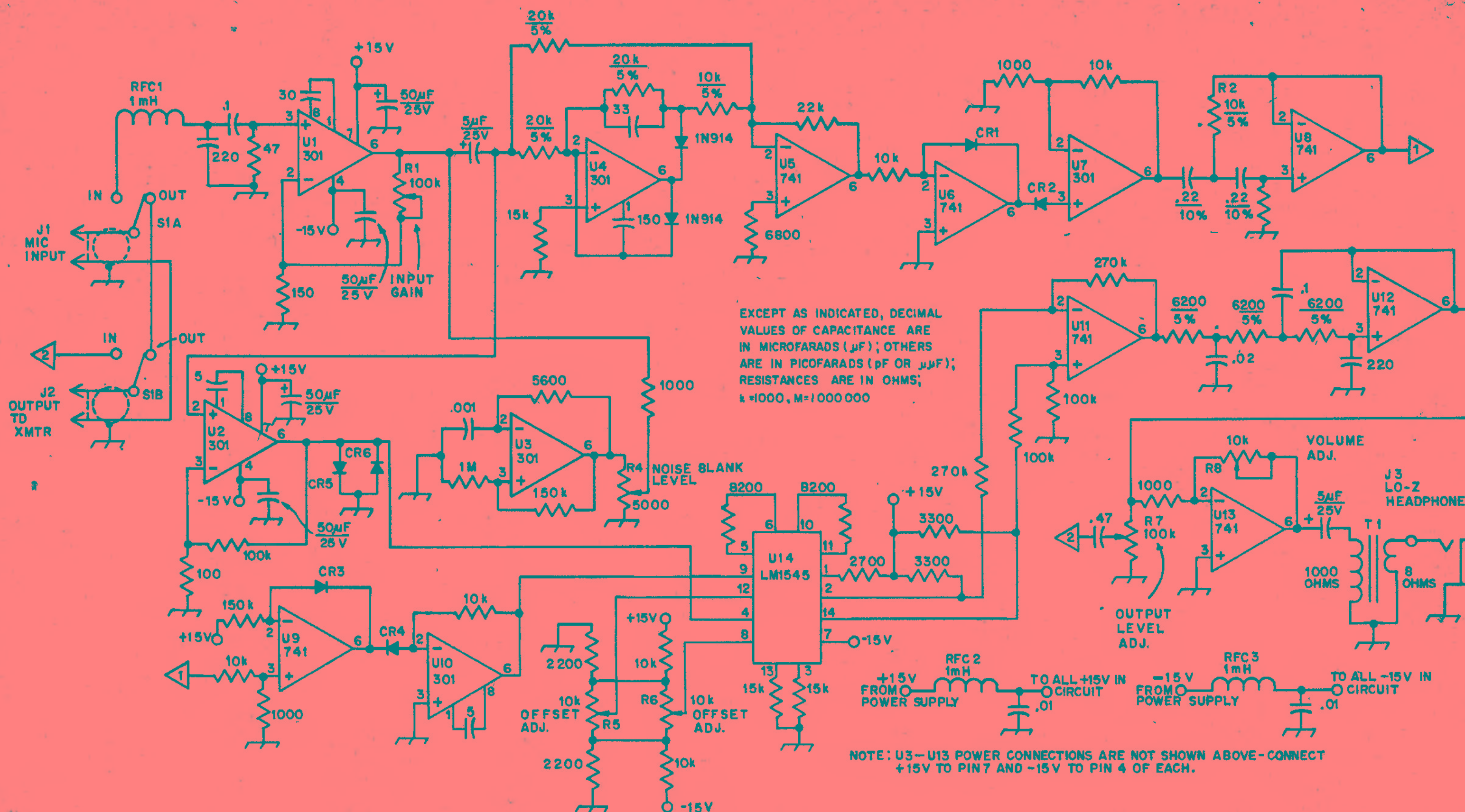


Fig. 3 — Circuit diagram for the speech processor.
CR1, CR2 — pair of matched silicon diodes (see text).
CR3, CR4 — Same as CR1, CR2.
R1, R4, R5, R6, R7 — Use circuit-board type pots.
S1 — Dpdt toggle switch.
T1 — 1000-ohm to 8-Ohm audio transformer, 250 milliwatts.
U1, U5, U6, U8, U9, U11, U12, U13 — 741 op amp (Fairchild µA741, National Semiconductor LM741, Motorola MC1741, or equivalent), 8-pin mini DIP package.
U2, U3, U4, U7, U10 — National Semiconductor LM301 op amp, 8-pin mini DIP package.
U14 — National Semiconductor LM1595 (or M torola MC1595), 14-pin DIP package.

Construction Information

An etched circuit board template pattern is available from ARRL, 225 Main Street, Newington CT 06111. Please include 50 cents and a self-addressed, stamped envelope. Builders who use this layout should experience no problems. Those attempting their own layout, however, should be cautioned that because of the relatively large number of active devices, some operating with high gain, the potential for instability (oscillation) exists in a haphazard layout. Arrangement of circuit components should be generally in a straight line from input to output. The shortest possible leads should be used in all cases. Particular attention should be paid to the positions of U1 and U2 with respect to each other. Because of the very high gain the input of U1 should be kept physically as far apart as possible from the output of U2. Mounting the circuit board inside a metal chassis, such as a suitable Minibox, is recommended for rf shielding.

Procurement of parts should present no particular problems. As of the time of this writing, the 741 and LM301 operational amplifiers used in the circuit can be purchased from mail order houses for about 30 cents apiece. The LM1595 integrated circuit, probably the single most expensive item in the processor, was bought for under two dollars. Matched diodes for CR1, CR2, CR3 and CR4 cost less than a dollar.[2]

The circuit is powered by a dual dc power supply that provides plus and minus 15 volts, as is typically used with most operational amplifiers. Current consumption is approximately 50 mA from each side of the supply.

Initial Adjustments

If an oscilloscope and audio sine wave generator are available, the following alignment procedure should be followed: Set R4 to minimum resistance. Connect a microphone to J1 and the oscilloscope probe to pin 6 of U1. Adjust R1, the input gain control, while speaking into the microphone so that the voice peaks viewed on the oscilloscope are slightly below the output clipping level of U1 (approximately 14 volts peak). Remove the microphone and connect the signal generator to J1. Set the generator frequency to about 1000 Hz and adjust its output level to produce about 10 volts peak at pin 6 of U1. Place the oscilloscope probe on pin 6 of U12. Adjust offset controls R5 and R6 for the best-looking sine wave. It should be possible to produce a nearly perfect sine wave. Disconnect the generator, reconnect the microphone, and plug headphones into J3. Advancing volume control R8, one should now be able to hear himself talking, although background noise and ac hum will probably be very high. Adjust noise blanking control R4 for the desired degree of background noise suppression.

Those who do not have access to test equipment may do the following: Set R4 to the center of its range. Connect a microphone to J1 and headphones to J3. Speaking into the microphone, advance input gain control R1 and monitor volume control R8 to the point where the speech becomes audible in the headphones. Adjust offset controls R5 and R6 for minimum distortion as monitored in the headphones. The final setting of R1 is not critical. It should be high enough so that the circuit functions properly (if set too low, the audio output will sound broken up and "grainy") but not so high that the speech amplifier itself distorts the signal by clipping. Adjust R4 to suppress background noise as desired.

Finally, connect the processor output at J2 to the transmitter's microphone jack. Switch the processor out of the line by means of S1. If a Monitorscope is available to view rf output, speak into the microphone and note the level of the voice peaks. Switch the unit "in" and adjust output level control R7 for the same peak voice output level. If a monitorscope is not available, the transmitter's alc meter readings may be used. With the processor switched "out," speak into the microphone and increase the transmitter's microphone gain control until the alc meter starts to deflect. Note the peak readings. Switch the processor "in" and adjust R6 to obtain the same peak reading.

[2] One source for this item: Tri-tek, Inc., Box 14206, Phoenix, AZ 85031.

SINGLE-SIDEBAND TRANSCEIVERS

A "transceiver" combines the functions of transmitter and receiver in a single package. In contrast to a packaged "transmitter-receiver," it utilizes many of the active and passive elements for both transmitting and receiving. Ssb transceiver operation enjoys widespread popularity for several justifiable reasons. In most designs the transmissions are on the same (suppressed-carrier) frequency as the receiver is tuned to. The only practical way to carry on a rapid multiple-station "round table" or net operation is for all stations to transmit on the same frequency. Transceivers are ideal for this, since once the receiver is properly set the transmitter is also. Transceivers are by nature more compact than transmitter-receivers, and thus lend themselves well to mobile and portable use.

Although the many designs available on the market differ in detail, there are of necessity many points of similarity. All of them use the filter type of sideband generation, and the filter unit furnishes the receiver i-f selectivity as well. The carrier oscillator doubles as the receiver (fixed) BFO. One or more mixer or i-f stage or stages will be used for both transmitting and receiving. The receiver S meter may become the transmitter plate-current or output-voltage indicator. The VFO that sets the receiver frequency also determines the transmitter frequency. The same signal-frequency tuned circuits may be used for both transmission and reception, including the transmitter pi-network output circuit.

Usually the circuits are switched by a

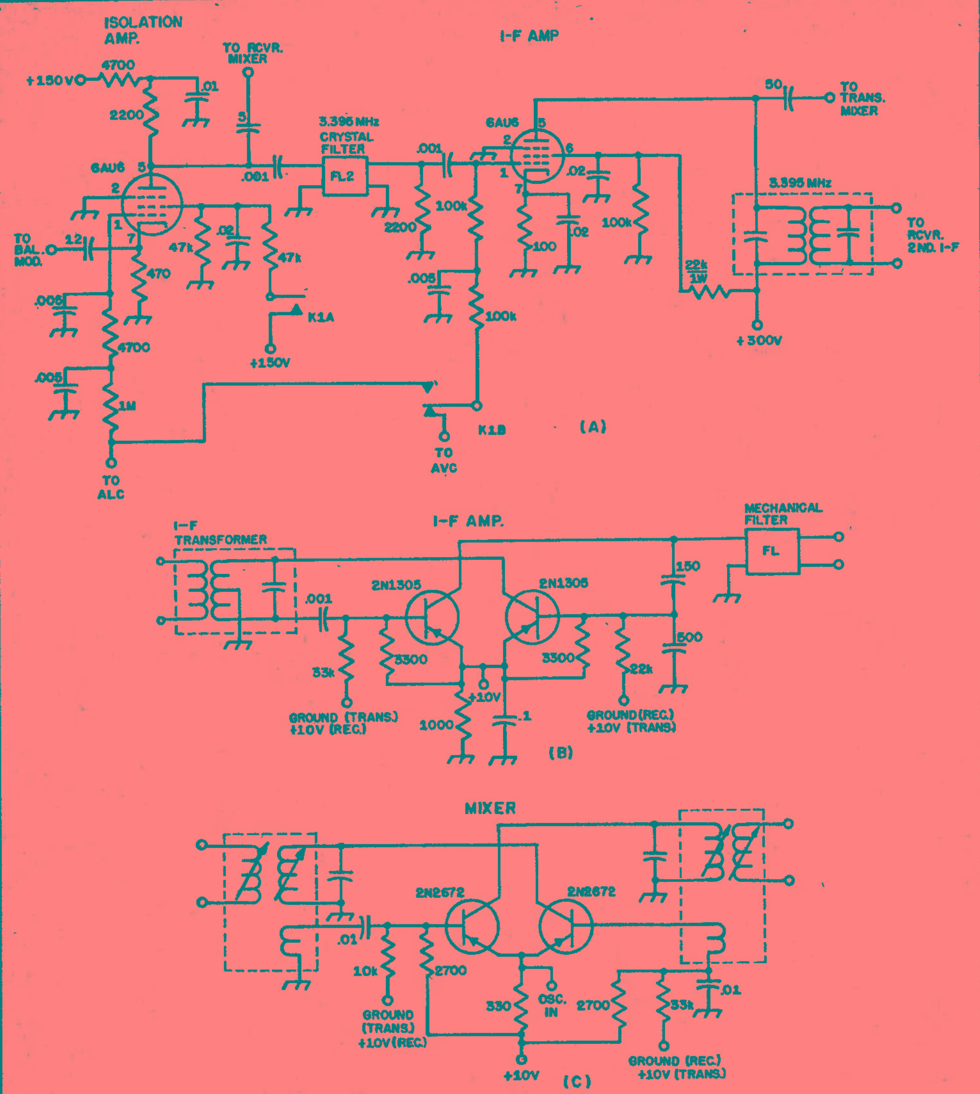


Fig. 13-28 — Transceiver circuits whare a section is made to operate on both transmit and receive. See text for details.

multiple-contact relay, which transfers the antenna if necessary and also shifts the biases on several stages. Most commercial designs offer VOX (voice-controlled operation) and MOX (manual operation). Which is preferable is a controversial subject; some operators like VOX and others prefer MOX.

Circuits

The use of a filter-amplifier combination common to both the transmitter and receiver is shown in Fig. 13-28A. This circuit is used by the Heath Company in several of their transceiver kits. When receiving, the output of the hf mixer is coupled to the crystal filter, which, in turn, feeds the first i-f amplifier. The output of this stage is transformer coupled to the second i-f amplifier. During transmit, K1 is closed, turning on the isolation amplifier that links the balanced modulator to the band-pass filter. The single-sideband output from the filter is amplified and capacitance-coupled to the transmitter mixer. The relay contacts also apply alc voltage to the first i-f stage and remove the screen voltage from the second i-f amplifier, when transmitting.

Bilateral amplifier and mixer stages, first used by Sideband Engineers in their SBE-33, also have found application in other transceiver designs. The circuits shown in Fig. 13-28B and C are made to work in either direction by grounding the bias divider of the input transistor, completing the bias network. The application of these designs to an amateur transceiver for the 80-10 meter bands is given in the 5th Edition of *Single Sideband for the Radio Amateur.*

The complexity of a multiband ssb transceiver is such that most amateurs buy them fully built and tested. There are, however, some excellent designs available in the kit field, and any amateur able to handle a soldering iron and follow instructions can save himself considerable money by assembling an ssb transceiver kit.

Some transceivers include a feature that permits the receiver to be tuned a few kHz either side of the transmitter frequency. This consists of a voltage-sensitive capacitor, which is tuned by varying the applied dc voltage. This can be a useful device when one or more of the stations in a net drift slightly. The control for this function is usually labled RIT for *receiver independent tuning.* Other transceivers include provision for a crystal-controlled transmitter frequency plus full use of the receiver tuning. This is useful for "DXpeditions" where net operation (on the same frequency) may not be desirable.

SSB Bibliography

Single Sideband for the Radio Amateur, by the American Radio Relay League, 5th Edition, 1970.

Hennebury, *Single Sideband Handbook*, Technical Material Corporation, 1964.

Pappenfus, Bruene and Schoenike, *Single Sideband Principles and Circuits*, McGraw-Hill, 1964.

Amateur Single Sideband, by Collins Radio Company, 1962.

TESTING A SIDEBAND TRANSMITTER

There are three commonly used methods for testing an ssb transmitter. These include the wattmeter, oscilloscope, and spectrum-analyzer techniques. In each case, a two-tone test signal is fed into the mic input to simulate a speech signal. From the measurements, information concerning such quantities as PEP and intermodulation-distortion-product (IMD) levels can be obtained. Depending upon the technique used, other aspects of transmitter operation (such as hum problems and carrier balance) can also be checked.

As might be expected, each technique has both advantages and disadvantages and the suitability of a particular method will depend upon the desired application. The wattmeter method is perhaps the simplest one but it also provides the least amount of information. Rf wattmeters suitable for single-tone or cw operation may not be accurate with a two-tone test signal. A suitable wattmeter for the latter case must have a reading that is proportional to the actual power consumed by the load. The reading must be independent of signal waveform. A thermocouple ammeter connected in series with the load would be a typical example of such a system. The output power would be equal to I^2R, where I is the current in the ammeter and R is the load resistance (usually 50 ohms). In order to find the PEP output with the latter method (using a two-tone test input signal), the power output is multiplied by 2.

A spectrum analyzer is capable of giving the most information (of the three methods considered here), but it is also the most costly method and the one with the greatest chance of misinterpretation. Basically, a spectrum analyzer is a receiver with a readout which provides a plot of signal amplitude vs. frequency. The readout could be in the form of a paper chart but usually it is presented as a trace on a CRT. A sweep voltage which is applied to the horizontal-deflection amplifiers of the CRT is also used to control the frequency of the LO for the first mixer (there may be other mixers but these are fixed-frequency types) in the analyzer. (See the block diagram in Fig. 1.) In order to give a meaningful output waveform, the first mixer has to have a broadband and "flat" response. It also has to have very good IMD suppression characteristics, otherwise, the mixer may generate spurious signals itself. Unfortunately, these signals fall on the same frequencies as those being measured in the transmitter output and it may be difficult to tell whether or not the spurious products are being generated in the transmitter or in the spectrum analyzer. Other precautions that should be taken would be to ensure that good RFI-prevention practices are observed. In effect, the problem is similar to trying to listen to one's own signal in the station receiver. Quite often, a signal may not be as bad *or as good* in the latter as it is at a distant station.

Two-Tone Tests and Scope Patterns

A method which is a very practical one for amateur applications is to use a two-tone test signal

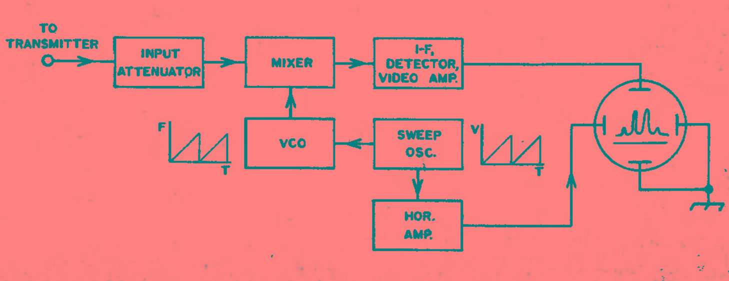


Fig. 1 — Block diagram of a spectrum analyzer.

(usually audio) and sample the transmitter output. The waveform of the latter is then applied directly to the vertical-deflection plates in an oscilloscope. An alternative method is to use an rf probe and detector to sample the waveform and apply the resulting audio signal to the vertical-deflection amplifier input.

If there are no appreciable nonlinearities in the amplifier, the resulting envelope will approach a perfect sine-wave pattern (see Fig. 2A). As a comparison, a spectrum-analyzer display for the same transmitter and under the same conditions is shown in Fig. 2B. In this case, spurious products can be seen which are approximately 30 dB below the amplitude of each of the tones.

As the distortion increases, so does the level of the spurious products and the resulting waveform departs from a true sine-wave function. This can be seen in Fig. 2C. One of the disadvantages of the scope and two-tone test method is that a relatively high level of IMD-product voltage is required before the waveform seems distorted to the eye.

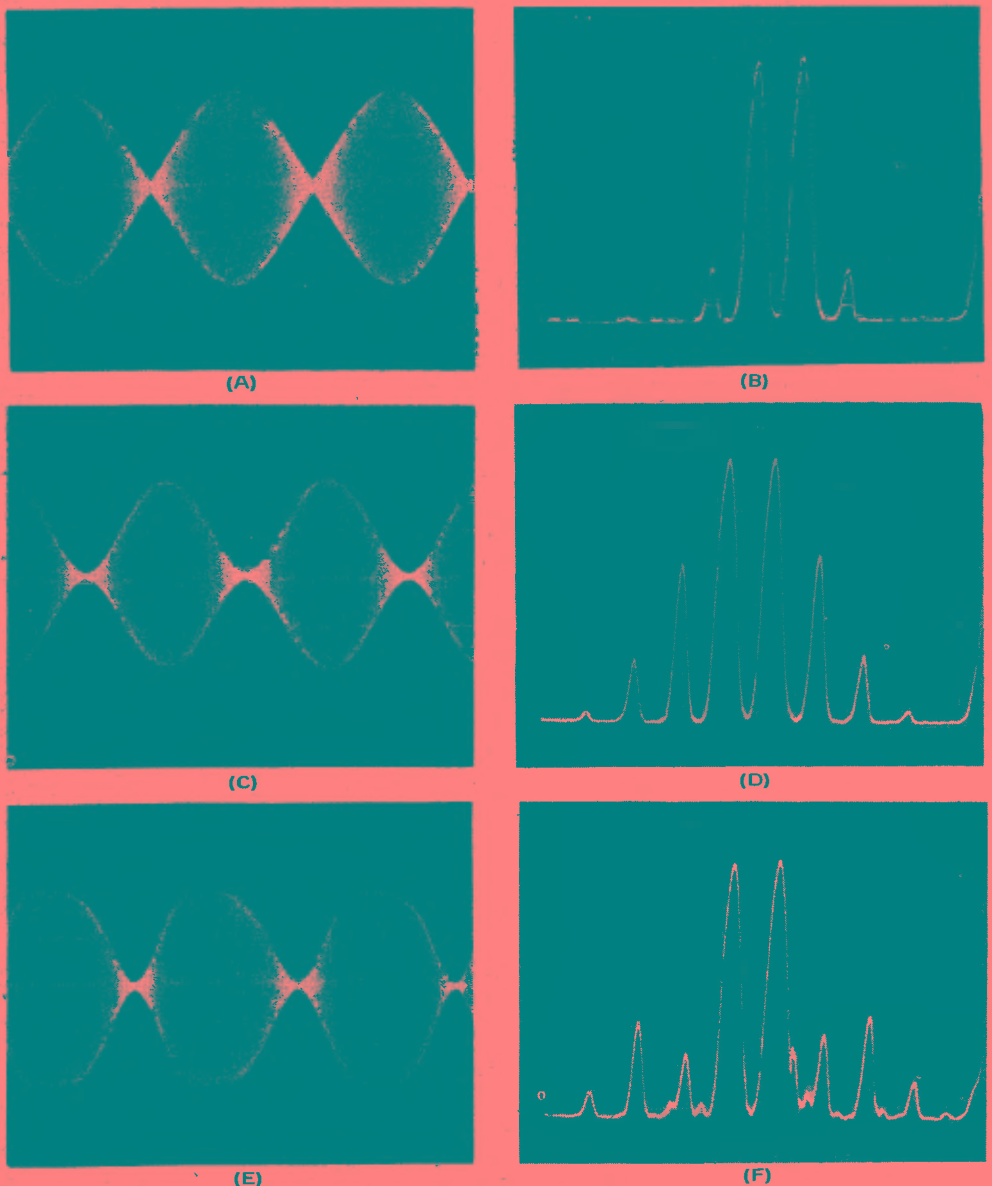


Fig. 2 — Scope petterns for a two-tone test signal and corresponding spectrum-analyzer displays. The pattern in A is for a properly adjusted transmitter and consequently the IMD products are relatively low as can be seen on the analyzer display. At C, the PA bias was set to zero idling current and considerable distortion can be observed. Note how the pattern has changed on the scope and the increase in IMD level. At E, the drive level was increased until the flattopping region was approached. This is the most serious distortion of all sinca the width of the IMD spectrum increases considerably causing splatter (F).

For instance, the waveform in Fig. 2C doesn't seem too much different than the one in Fig. 2A but the IMD level is only 17 dB below the level of the desired signal (see analyzer display in Fig. 2D). A 17- to 20-dB level corresponds to approximately ten-percent distortion in the voltage waveform. Consequently, a "good" waveform means the IMD products are at least 20 dB below the desired tones. Any *noticeable* departure from the waveform in Fig. 2A should be suspect and the transmitter operation should be checked.

The relation between the level at which distortion begins for the two-tone test signal and an actual voice signal is a rather simple one. The maximum deflection on the scope is noted (for an acceptable two-tone test waveform) and the transmitter is then operated such that voice peaks are kept below this level. If the voice peaks go above this level, a type of distortion called "flattopping" will occur and the results are shown for a two-tone test signal in Fig. 2E. IMD-product levels raise very rapidly when flattopping occurs. For instance, third-order product levels will increase 30 dB for every 10-dB increase in desired output as the flattopping region is approached, and fifth-order terms will increase by 50 dB (per 10 dB).

Interpreting Distortion Measurements

Unfortunately, considerable confusion has grown concerning the interpretation and importance of distortion in ssb gear. Distortion is a very serious problem when high spurious-product levels exist at frequencies removed from the passband of the desired channel but is less serious if such products fall within the bandwidth of operation. In this former case, such distortion may cause needless interference to other channels ("splatter") and should be avoided. This can be seen quite dramatically in Fig. 2F when the flattopping region is approached and the fifth and higher order terms increase drastically.

On the other hand, attempting to suppress in-band products more than necessary is not only difficult to achieve but may not result in any noticeable increase in signal quality. In addition, measures required to suppress in-band IMD often cause problems at the expense of other qualities such as efficiency. This can lead to serious difficulties such as shortened tube life or transistor heat-dissipation problems.

The two primary causes of distortion can be seen in Fig. 3. While the waveform is for a single-tone input signal, similar effects occur for the two-tone case. As the drive signal is increased, a point is reached where the output current (or voltage) cannot follow the input and the amplifier saturates. This condition is often referred to as flattopping (as mentioned previously). It can be prevented by ensuring that excessive drive doesn't occur and the usual means of accomplishing this is by alc action. The alc provides a signal that is used to lower the gain of earlier stages in the transmitter.

The second type of distortion is called "cross-over" distortion and occurs at low signal levels. (See Fig. 3.) Increasing the idling plate or collector

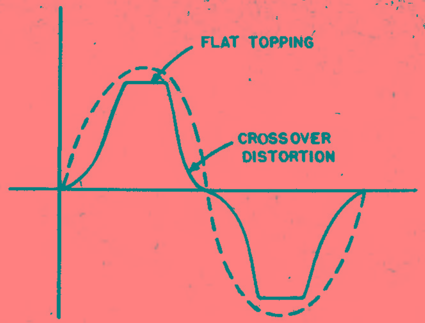


Fig. 3 — Waveform of an amplifier with a single-tone input showing flattopping and cross-over distortion.

current is one way of reducing the effect of cross-over distortion in regards to producing undesirable components near the operating frequency. Instead, the components occur at frequencies considerably removed from the operating frequency and can be eliminated by filtering.

As implied in the foregoing, the effect of distortion frequencywise is to generate components which add or subtract in order to make up the complex waveform. A more familiar example would be the harmonic generation caused by the nonlinearities often encountered in amplifiers. However, a common misconception which should be avoided is that IMD is caused by fundamental-signal components beating with harmonics. Generally speaking, no such simple relation exists. For instance, single-ended stages have relatively poor 2nd-harmonic suppression but with proper biasing to increase the idling current, such stages can have very good IMD-suppression qualities.

However, a definite mathematical relation does exist between the desired components in an ssb signal and the "distortion signals." Whenever nonlinearities exist, products between the individual components which make up the desired signal will occur. The mathematical result of such multiplication is to generate other signals of the form (2f1 – f2), (3f1), (5f2 – f1) and so on. Hence the term intermodulation-distortion products. The "order" of such products is equal to the sum of the multipliers in front of each frequency component. For instance, a term such as (3f1 – 2f2) would be called a fifth-order term since 3+2 is equal to 5. In general, the 3rd, 5th, 7th, and similar "odd order" terms are the most important ones since some of these fall near the desired transmitter output frequency and can't be eliminated by filtering. As pointed out previously, such terms do not normally result from fundamental components beating with harmonics. An exception would be when the fundamental signal along with its harmonics is applied to another nonlinear stage such as a mixer. Components at identical frequencies as the IMD products will result.

When two equal tones are applied to an amplifier and the result is displayed on a spectrum analyzer, the IMD products appear as "pips" off to the side of the main signal components (Fig. 2). The amplitudes associated with each tone and the

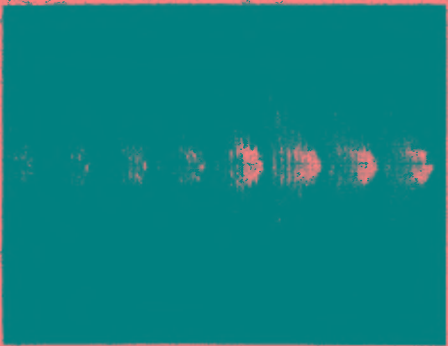


Fig. 4 — Speech pattern of the word "X" in a properly adjusted ssb transmitter.

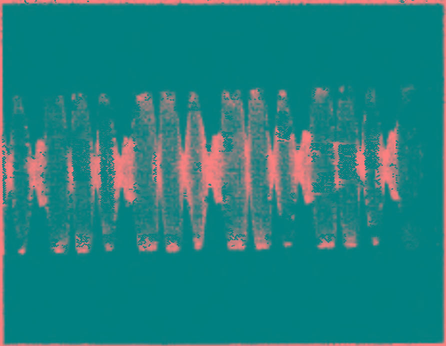


Fig. 5 — Severe clipping (same transmitter as Fig. 4 but with high drive and alc disabled).

IMD products are merely the dB difference between the particular product and one tone. However, each desired tone is 3 dB down from the *average* power output and 6 dB down from the *PEP* output.

Since the PEP represents the most important quantity as far as IMD is concerned, relating IMD-product levels to PEP is one logical way of specifying the "quality" of a transmitter or amplifier in regard to low distortion. For instance, IMD levels are referenced to PEP in Recent Equipment reviews of commercially made gear in *QST*. PEP output can be found by multiplying the PEP input by the efficiency of the amplifier. The input PEP for a two-tone test signal is given by:

$$PEP = E_p\, I_p\, [1.57 - 0.57\frac{I_0}{I_p}]$$

where E_p is the plate voltage and I_p is the average plate current. I_o is the idling current.

Generally speaking, most actual voice patterns will look alike (in the presence of distortion) except in the case where severe flattopping occurs. This condition is not too common since most rigs have an alc system which prevents overdriving the amplifiers. However, the voice pattern in a properly adjusted transmitter usually has a "Christmas tree" shape when observed on a scope and an example is shown in Fig. 4.

Mathematical Relation Between Amplifier Nonlinearity and IMD Products

The term intermodulation-distortion product is often used and the following derivation shows how it is related to amplifier nonlinearity. The output of an amplifier can be related to the input by means of a power series of the form:

$$S = A + Br + Cr^2 + Dr^3 + Er^4 + \ldots\ldots$$

where *s* represents some parameter such as output voltage or current and *r* represents some input quantity (voltage or current). *A*, *B*, *C*, and other constants are primarily determined by the amplifier nonlinearity. *A* represents a dc term and can be neglected. In an ideal amplifier with no distortion, *C*, *D*, and the constants for the higher exponent terms will be zero and only the constant for the "linear" term, *B*, will exist. Consequently, the output quantity will be an exact replica of the input.

If the output is plotted against the input, a straight line will result, hence the term "linear operation." On the other hand, if distortion is present, the *C*, *D*, and other constants will not be zero. The values of the constants will be such that as *r* increases, the higher order terms will add (or subtract) so that *s* follows the input-output curve of the amplifier.

For a two-tone test signal, *r* can be represented by the following formula:

$$r = Ro\,(Cos\, w_1 t + Cos\, w_2 t);\quad w_1 = 2\pi f_1,\quad w_2 = 2\pi f_2$$

where *f*1 and *f*2 are the frequencies of the two tones. If this equation for *r* is substituted into the power series, many terms will result and the algebra involved to find each one would be rather tedious. However, the purpose here is only to show how the IMD products come about. For instance, some terms will yield products such as:

$$(Cos^2\, w_1 t)\,(Cos\, w_2 t)$$

The squared term can be expanded by a trigonometric formula:

$$Cos\, 2\, w_1 t = \frac{1 + Cos\, 2\, w_1 t}{2}$$

This gives rise to a term $Cos 2\, w_1 t\, Cos\, w_2 t$ which can be expanded by another trigonometric formula to give:

$$Cos\, 2\, w_1\, t\, Cos\, w_2\, t = \frac{1}{2}[Cos\,(2\, w_1 + w_2)\, t + Cos\,(2\, w_1 - w_2)\, t]$$

The second term in the bracket represents a third-order IMD "product" which falls close to the ssb passband. Notice that the exponents of the product functions which gave rise to this term are 2 and 1, respectively, hence the term "third-order" product. The manner in which the terms increase will depend upon the distortion curve but generally speaking, the amplitude will follow a law which is proportional to *r* raised to a power *x*, where *x* is the order of the term.

TRANSISTORIZED VOX

Voice-operated relay (VOX) provides automatic transmit-receive switching. It is a useful accessory, and one that can add to the pleasure of operating. Owners of commercially made transmitters that have been designed only for push-to-talk operation, and home constructors who are "rolling their own" rigs, will find that this unit, shown in Fig. 13-36, provides excellent VOX operation and that it can be used with their existing station equipment.

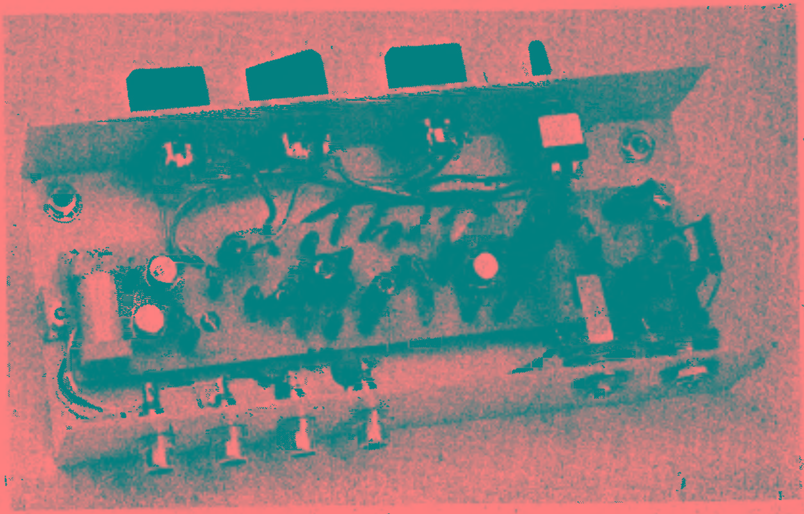
The Circuit

Operation of a VOX circuit is not complicated. A JFET transistor, Q1 in Fig. 13-36, operates as the first audio amplifier. The high input impedance of this type of transistor is desirable, because the use of high-impedance microphones is nearly universal in the amateur service. Q2 and Q3 provide additional amplification of the audio signal. The gain of these two stages is high. But, if additional gain is needed, bypass capacitor C7 may be added across the emitter resistor of Q2. With all but the low-output dynamic microphones, however, this capacitor should not be necessary. The audio output from Q3 is rectified by CR1 and CR2.

The dc output from the audio-signal rectifier is amplified by Q5 and fed to Q6. With no signal on its base, Q6 draws no collector current, holding the voltage on the base of Q7 near zero until the input signal reaches a sufficient level to turn the transistor on. Q7 will then turn on, drawing collector current through the relay coil, closing K1. The transistor that operates the relay is protected by CR5 from transient spikes generated as the current changes in the coil of K1. Provision is made for turning K1 on with a front-panel switch, S1, which holds the relay closed for a period of transmitter tuning or other adjustments.

A delay circuit, borrowed from ON5FE, is included to hold K1 closed for a short time after the audio-signal input ceases. This delay keeps the relay from chattering or opening during the short pauses between words or syllables. The length of the time delay is determined by the value of C15 and the setting of the DELAY control, R22. The advantage of ON5FE's circuit is that a relatively low value of capacitance can be used. Other circuits, which use delay capacitors of 50- to 200-μF, have slow turn-on action because series resistances used in the circuits prevent the large-value delay capacitor from charging instantaneously. A slow turn-on time is definitely undesirable, as it results in clipping of the first word spoken.

Audio output from a station receiver can key the VOX; to prevent this problem, an anti-VOX circuit is included. A sample of the receiver audio is amplified by Q4 and rectifed by CR3 and CR4. The output of this rectifier is negative in polarity and opposes the positive voltage developed by CR1 and CR2. Thus, when controls R19 and R20 are correctly set, any pickup from the speaker does not activate the VOX, as the positive and negative voltages cancel, and Q5 does not operate. A short time constant is desirable on the output of the anti-VOX rectifier; C11 provides this function. Receivers with 4- to 16-ohm speakers require amplification of the audio signal sampled across the speaker leads. If the receiver audio is taken



Interior view. With the exception of the controls, connection jacks, and rf bypass capacitors, all components are mounted on an etched-circuit board.

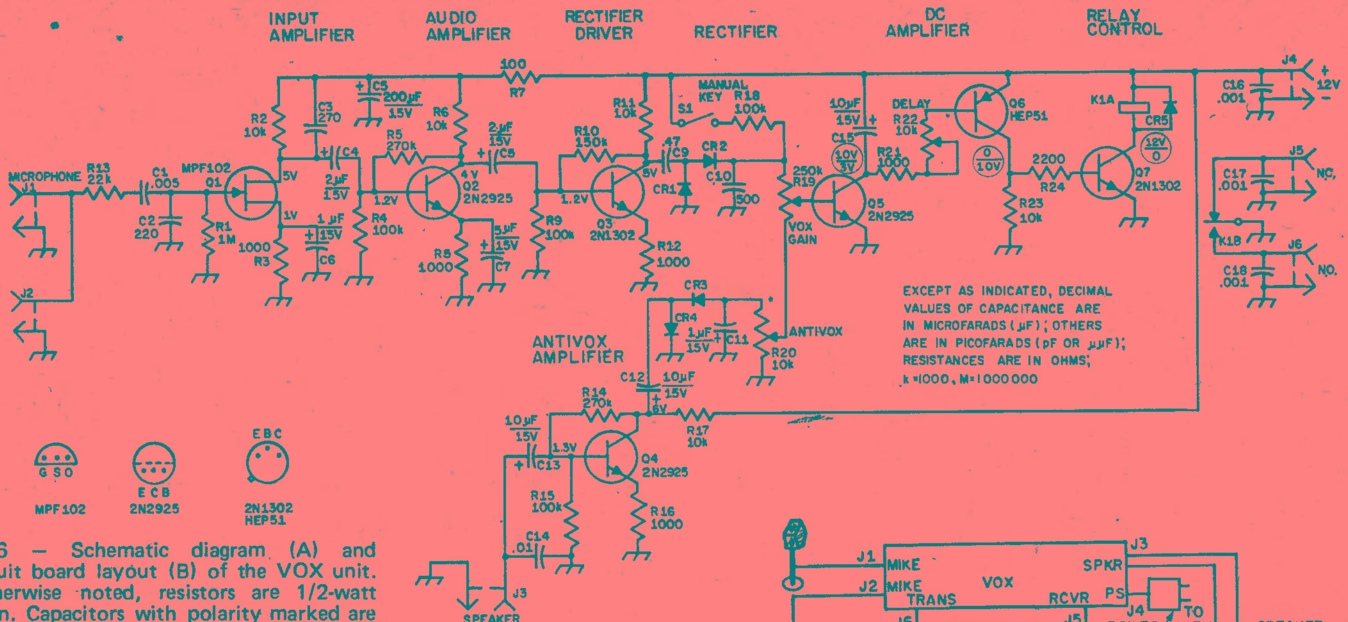


Fig. 13-36 — Schematic diagram (A) and etched circuit board layout (B) of the VOX unit. Unless otherwise noted, resistors are 1/2-watt composition. Capacitors with polarity marked are electrolytic; others are disk ceramic. Numbered components that are not listed below are for circuit-board reference.

C9 — Mylar or other low-leakage type.
CR1-CR4, incl. — Germanium diode, 1N67A or similar.
CR5 — Silicon, 50 PRV or more.
J1, J2 — Phone jack, panel mount.
J3-J6, incl. — Phono type.
K1 — Reed relay, spdt contacts, 12-V coil (Magnecraft W104MX-2).
R19 — Linear-taper carbon control (Mallory MLC254L).
R20, R22 — Linear-taper carbon control (Mallory MLC14L).
S1 — Miniature toggle (Radio Shack 275-1546 or 275 326).

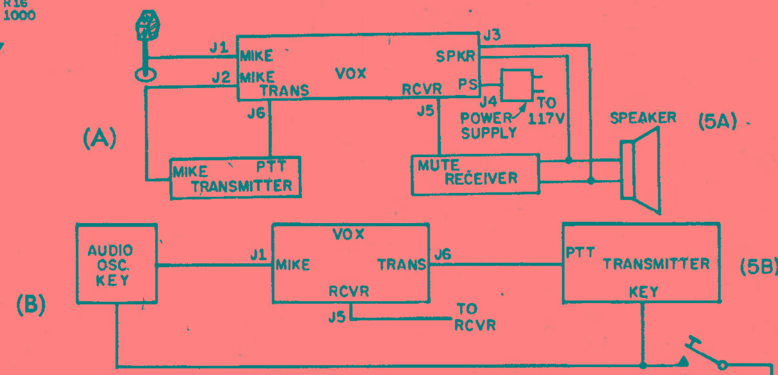


Fig. 13-37 — Typical connections for the VOX adaptor when used for (A) phone and (B) cw operation.