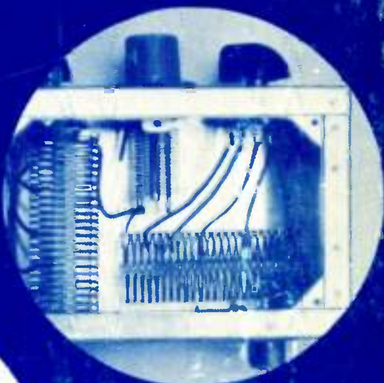
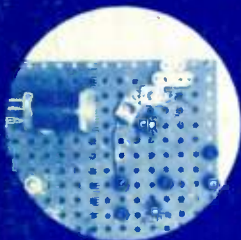
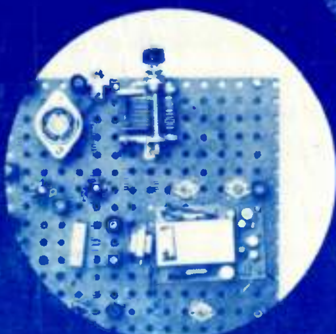
RADIO TRANSMITTER

PRINCIPLES and PROJECTS


by
EDWARD
M. NOLL

W3FQJ

RADIO TRANSMITTER PRINCIPLES AND PROJECTS

by Edward M. Noll, W3FQJ

Radio Transmitter Principles and Projects has been written for the student and the practicing communications technician, for the radio amateur, for the transmitter experimenter, and for those studying for the various grades of amateur or commercial FCC license examinations.

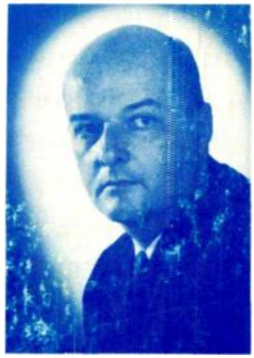
Devoted entirely to the subject of radio transmitters, **Radio Transmitter Principles and Projects** covers basic and advanced subjects including radio-frequency oscillators, multipliers, and amplifiers. The experimenter can learn about all modes of modulation—cw, a-m, dsb, ssb and f-m—as well as all of the important electron devices—bipolar transistors, FETs, tubes, integrated circuits, voltage variable capacitors—plus how to tune and test transmitters.

Radio Transmitter Principles and Projects does not stop short with principles only. You will find projects using all types of electron devices and various modes of modulation.

You can build, test, and observe these circuits with ease. All are low-power circuits that can serve as laboratory experiments in technical schools.

The projects provide an opportunity for the radio amateur to build his own gear and really gain practical experience. The QRP buff will find much to his liking.

Radio Transmitter Principles and Projects is a helpful gathering of modern transmitter principles, ideas, circuits, techniques, and learn-by-doing projects.



ABOUT THE AUTHOR

In addition to being an accomplished author of technical books, lessons, articles, and instruction manuals, Ed Noll is also a consulting enginer and lecturer. His other books include:

73 Dipole and Long-Wire Antennas

73 Vertical, Beam, and Triangle Antennas

Ham and CB Antenna Dimension Charts

First-Class Radiotelephone License Handbook

Second-Class Radiotelephone License Handbook

Radar License Endorsement Handbook

Radio Operators License Handbook

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PREFACE

The explanations of principles and projects in this book are intended for both students and professional technicians. Each chapter begins with basic principles and advances to more detailed information. The projects are based on the basic principles and are designed to further the reader's understanding through actual experience.

The first three chapters contain information on electron devices —the FET, bipolar transistor, and the vacuum tube. Various modes of modulation—cw, a-m, f-m, dsb, and ssb—are discussed in other sections. Chapter 4 describes hybrid transmitter circuits using tubes and transistors. Double-sideband and single-sideband generation and circuits are included in Chapter 5. There is a chapter on linear amplifiers and mixers; another explains integrated circuits. The final three chapters detail vhf circuits, frequency modulation, and transmitter testing.

The radio transmitter projects are all low-power circuits that can serve as laboratory experiments in schools. The radio amateur will gain practical experience with all types of circuits and systems. It is hoped that the radio technician will find this book to be a helpful guide to modern transmitter principles, ideas, circuits, and techniques.

EDWARD M. NOLL, W3FQJ

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CHAPTER 1

FET CW/A-M Transmitter Circuits

"Radio Transmitter Principles and Projects" has been written for the beginner, the oldtimer, and the experienced technician. Its approach can be summed up as "learn-and-do." Principles are detailed and followed by construction projects that firm your knowledge of transmitter circuit operation. In the early fundamental chapters the projects are of step-by-step detail and include a series of suggested experimental procedures to follow which teach the underlying principles. The coverage provides an excellent introduction to solid-state and combined solid-state, vacuum-tube technology.

In the first three chapters the basic electron devices (FET, bipolar transistor, and vacuum tube) are covered. First you learn the principle of the device and then the fundamentals of the radio-frequency circuit with which the device can be associated. Then you will learn how the device can be amplitude-modulated. All sorts of devices in addition to the basic three are covered such as diode, voltage-variable capacitor diode, linear integrated circuits, digital integrated circuits, etc. Oscillators and all classes of amplifier operation A, AB, B, and C are covered. The transmitters include all modes of modulation, cw, a-m, dsb, ssb, and f-m.

The projects include small test units on up to complete transmitters that serve well as laboratory experiments or as actual parts of an amateur radio station. Power levels range from QRPP to QRP (very low power to low power) and on up to a maximum of about 50 watts.

In the early fundamental chapters the projects are in the form of experiments that follow along with the text. Even though you do not perform the experiments, read right along with the copy, because the experiments are an inherent part of the chapter continuity. Performing the experiments, however, gives you that extra and very valuable practical experience with new devices and circuits.

FIELD-EFFECT TRANSISTOR

A field-effect transistor (FET) is a three-element device consisting of source, gate, and drain (Fig. 1-1). The source and drain elements

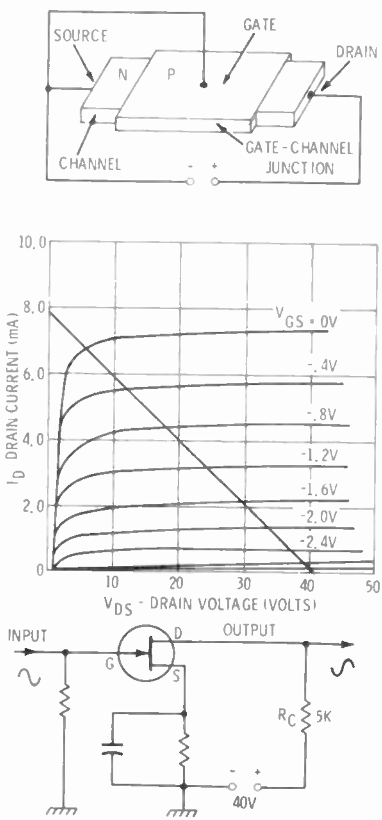


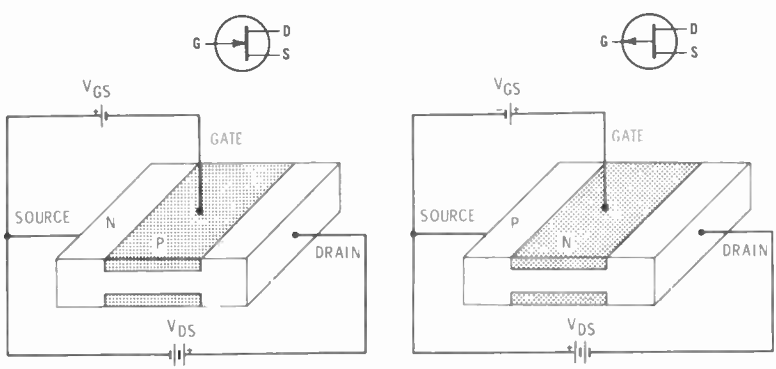
Fig. 1-1. Drawing of basic field-effect transistor.

are positioned at the ends of a continuous semiconductor channel. The charge motion along the channel from source to drain depends on the conductivity of that channel. Conductivity is regulated by a third element, the gate, which surrounds all or part of the channel. A single semiconductor junction is present at the boundary between the gate and the channel. Normally this junction is reverse-biased just as the grid of a vacuum tube is biased negatively.

The gate actually has a capacitive influence on the channel. This capacitive effect is such that a charge depletion area extends into the channel. The amount of reverse biasing of the gate determines the extent of the depletion activity and therefore the conductivity of the channel. In turn, the conductivity of the channel determines the charge motion (current) between source and drain.

As the gate reverse bias is increased, the depletion area increases and lowers the conductance of the channel, thus increasing the channel resistance or decreasing the channel conductance. As a result the channel and output drain currents decrease. If the reverse-bias voltage is made to vary, there results a substantial change in the drain current. This substantial drain-current variation in the common-source circuit of Fig. 1-1 produces a substantial voltage variation across the output. Since the drain-voltage variation is greater than gate-voltage variation, the common-source stage has voltage gain.

A typical family of curves is also shown in Fig. 1-1, along with a load line. Note how similar the set of curves is to those of a pentode



(A) N-channel.

(B) P-channel.

Fig. 1-2. N- and P-channel FETs.

vacuum tube. The lower the gate bias is, the greater is the drain current. Also over a substantial voltage range the drain current remains essentially constant for a given gate bias as the drain voltage is increased. This is characteristic of a high-resistance device. In fact, the input and output resistances of a field-effect transistor are high like those of a vacuum tube and are unlike the low resistances at the input and output of a conventional bipolar transistor. The field-effect transistor in simple circuits is less prone to distortion and the generation of spurious signal components compared to a bipolar transistor.

There are two fundamental types of junction field-effect transistors, those with a p-channel and those with an n-channel (Fig. 1-2). An n-channel has a charge motion determined by free electrons. The gate which surrounds an n-channel is a p-type semiconductor material with its electric charge motion determined by free positive charges, or holes. The second type uses a p-type channel and an n-type gate. The only difference is the polarity of the biasing. Note that the drain of an n-channel type is positive with a negative voltage required at the gate to establish reverse-biasing of the junction. Oppositely the p-channel type operates with a negative drain voltage. In this case a positive bias is applied to the gate to reverse-bias the junction.

RF AMPLIFIER

A typical FET class-B or class-C rf amplifier is shown in Fig. 1-3. External bias can be used. Biasing according to class of operation is shown in the transfer characteristic. For class-A operation the biasing is at the center of the linear portion of the transfer curve. To operate the amplifier class B it is necessary that the level of external bias match the cutoff bias of the transistor. A class-C stage is, of course, biased beyond cutoff.

The input impedance of a FET rf amplifier can be kept at its highest by making certain that the peak of the positive alternation of the input signal does not extend to the near-zero bias level that results in gate current. However, for more convenient and efficient operation and a somewhat lower input impedance, the stage may be operated in such a manner that the gate current is drawn at the peak of the positive alternation. In such an arrangement one can use gate current to establish the required class-B or class-C biasing

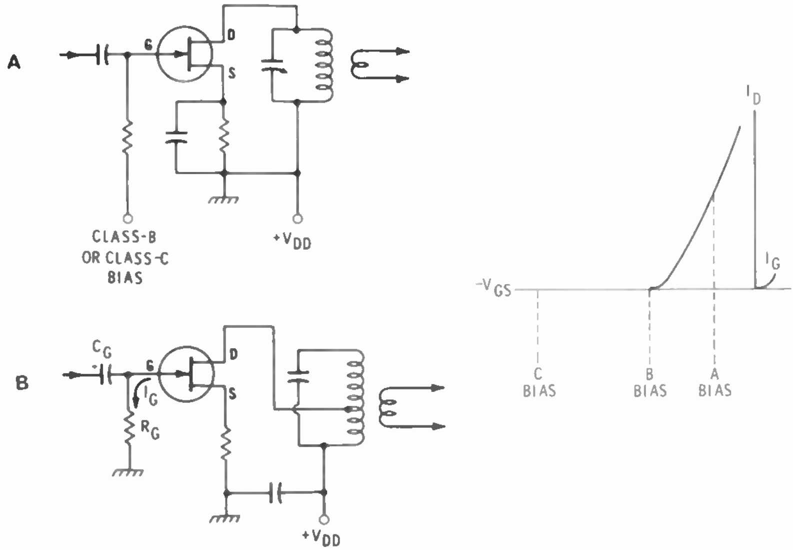


Fig. 1-3. FET class-B and class-C amplifiers.

using the circuit arrangement of Fig. 1-3B. The direction of the gate current is such that a negative bias is developed on the gate capacitor. With a proper time constant this capacitor charge remains constant and serves as the dc gate bias for the stage. As mentioned, the input impedance is now lower, and somewhat more input power is required.

The field-effect class-C amplifier is similar to a vacuum-tube stage in still another way. When the rf excitation is removed from its input, the drain current rises just as the plate current of a similarly designed vacuum-tube stage. It is advisable to use a protective source resistor which limits the drain current to a safe value in case rf excitation is lost. In vacuum-tube practice one often uses a cathode resistor as a safety device.

Approximate class-C waveforms are shown in Fig. 1-4. During the portion of the input wave that causes drain current there is a strong but short-duration burst of drain current. This burst of current contributes power to the output resonant circuit. In fact, at this time the tank capacitor is charged to a negative peak and the drain voltage is at minimum value.

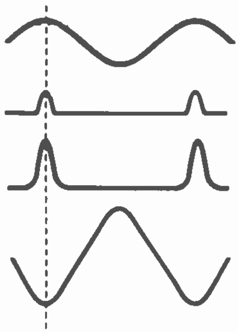


Fig. 1-4. FET class-C waveforms.

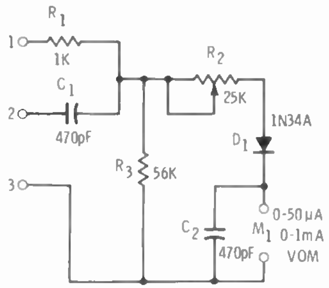
It is important to recognize that with a strong input signal the peak drain current rises from zero to a high peak value. Consequently there can be a substantial change in drain voltage and the power delivered to the drain resonant circuit is much greater than the signal power level delivered to the gate of the transistor. The energy delivered to the tank circuit is stretched out because of the energy-storage capability. As a result a sinusoidal rf voltage is developed across the tuned circuit.

The gate current is present only for a very short interval of time that coincides with the positive peak of the input voltage. Just enough gate current is drawn to charge the bias capacitor.

PROJECT 1: OUTPUT INDICATOR

Before starting construction of your first transmitter circuits, you should build a small piece of test equipment that is helpful in working with solid-state rf amplifiers. This output indicator is shown schematically in Fig. 1-5. It is nothing more than a simple diode detector circuit which develops a dc output voltage that corresponds to the amplitude of an rf signal. Useful readings are obtained for rf power levels considerably lower than 100 mW. A potentiometer is included, and permits the measurement of higher power levels. When measuring across a low-impedance source, connect the signal to be evaluated between inputs 1 and 3. In fact, the indicator can be left connected across the transmitter output to a low-impedance antenna when the transmitter is in operation and feeding power to the antenna.

If the transmitter is to be terminated in a specific impedance, that value can be connected between terminals 1 and 3. For example,



C_1,C_2	470-pF capacitors	R_1	1K ½-watt resistor
D_1	1N34A diode	R_2	25K potentiometer
M_1	Current meter, 0-50 μA or 0-1 mA		5 binding posts

Fig. 1-5. Rf indicator.

if the transmitter is to supply power to a 50-ohm load, one can connect such a 50-ohm resistive termination (of proper power-handling capability) across terminals 1 and 3 and the transmitter will see a proper matched load.

If the signal is to be evaluated at a high-impedance point, use terminals 2 and 3. The presence of the capacitor also permits ob-

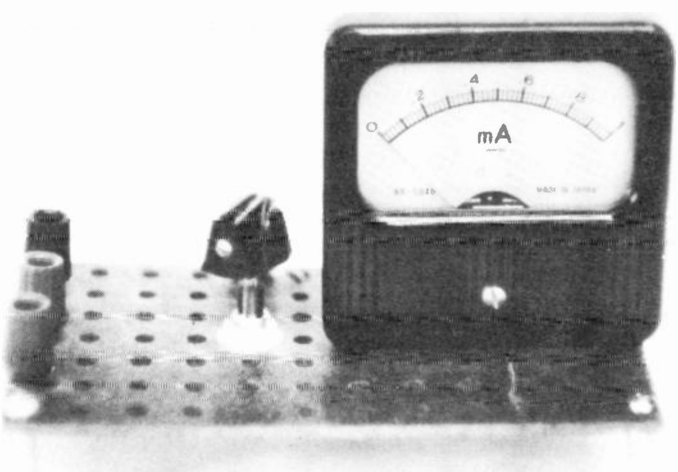


Fig. 1-6. Rf indicator with 0-1 mA meter.

servation where there is a dc as well as an rf component. The capacitor blocks the dc voltage from the detector circuit.

The indicator can be 0 to 1 milliammeter or one with an even higher sensitivity. A volt-ohm-milliammeter (vom) can also be used and provides some additional versatility because of a choice of current scales. The unit can be mounted on a small 3″ × 6″ *masonite* pegboard. The five binding posts provide convenience in use. Such an arrangement with a mounted milliammeter is shown in Fig. 1-6.

PROJECT 2: FET CRYSTAL OSCILLATORS

Most projects of this chapter will be built on the single pegboard arrangement shown in Fig. 1-7. First mount four transistor sockets

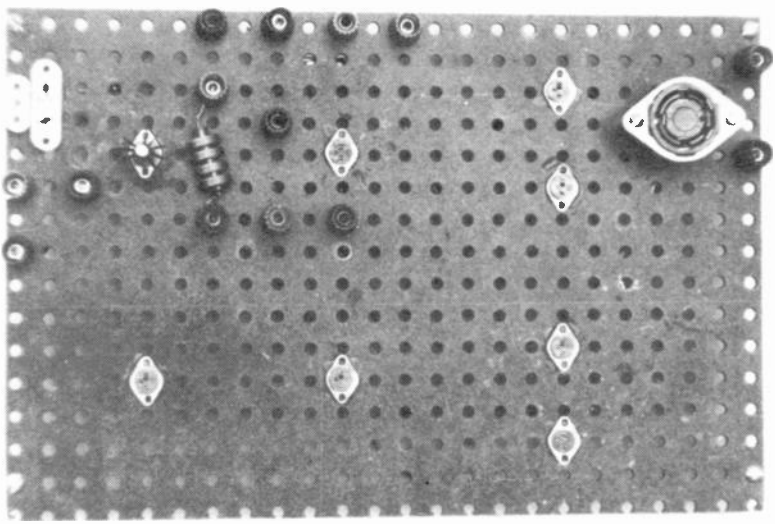
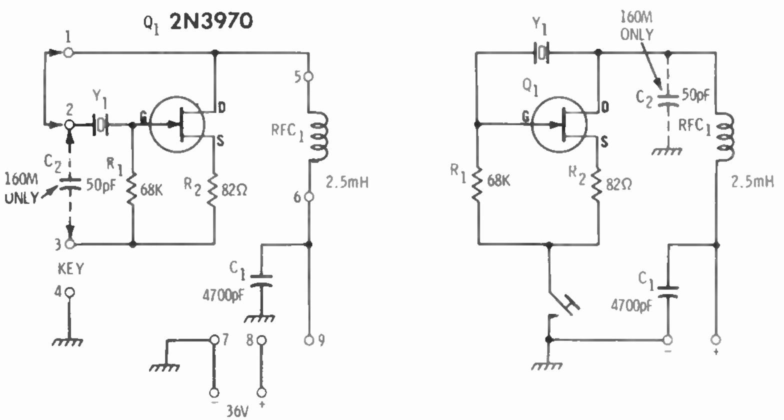


Fig. 1-7. Basic arrangement of pegboard for oscillator circuits.

as per the top half of arrangement. Experiments on crystal oscillator circuits will be made in the area at the top left. A FET and its heat sink are shown along with two crystal sockets to accommodate the two common size crystal plugs. Various binding posts have been mounted in positions that permit ease in making circuit changes. A five-prong tube socket has been mounted at the top right. In later projects, plug-in coils will be used in the tube socket.

20-40-80-160 Pierce Oscillator

Assemble the circuit of Fig. 1-8. Note that nine binding posts are used. A Pierce crystal oscillator circuit is formed by connecting a



C_1	4700-pF disc capacitor	RFC	2.5-mH rf choke
C_2	220-pF disc capacitor		Miscellaneous Parts
Q_1,Q_2	2N3970 transistors	Pegboard	
R_1	68K ½-watt resistor	Transistor sockets	
R_2	82-ohm ½-watt resistor	Binding posts	

Fig. 1-8. Pierce crystal oscillators.

jumper between binding posts 1 and 2 and a 2.5-mH radio-frequency choke between posts 5 and 6.

Drain current can be measured by inserting a 0-100 milliammeter between terminals 8 and 9. When measurements are not being made a jumper can be inserted between binding posts 8 and 9.

The output of the crystal oscillator can be evaluated with the output indicator. Connect binding post 5 to binding post 2 of the indicator and binding post 3 of the indicator to one of the common (ground) binding posts of the oscillator (Fig. 1-9).

Operation—Wire the oscillator carefully to set up the circuit of Fig. 1-8B. Plug in a 40- or 80-meter crystal. Close the key or connect a jumper between posts 3 and 4 to turn on the oscillator. Note the drain current and the output indicator reading on 40 and 80 meters. Tune in the signal on your receiver.

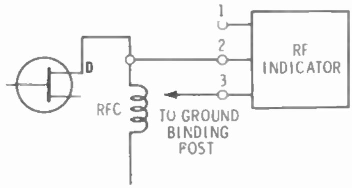


Fig. 1-9. Connection of the output indicator.

Insert a 160-meter crystal. The oscillator may be sluggish in starting on this frequency. A greater capacitive load on the drain output circuit will overcome the problem. Connect approximately a 50-pF capacitor from drain to common whenever operating on 160. One way of doing this is to connect the capacitor to binding post 3 or 4 from either binding post 1 or 2. It can be removed when operating on other bands.

Try a 20-meter crystal in the oscillator. Output reading will be about the same on all bands. Typical output readings fall between 0.35 and 0.5 (maximum sensitivity) milliamperes. Drain current falls to between 25 and 30 milliamperes and somewhat higher on 160 meters.

Drive Reading—Connect the gate circuit of what will be the second stage of the transmitter. This circuit consists of capacitor C_2 and resistors R_3, R_4, and R_5 (Fig. 1-10). The rf drive into this

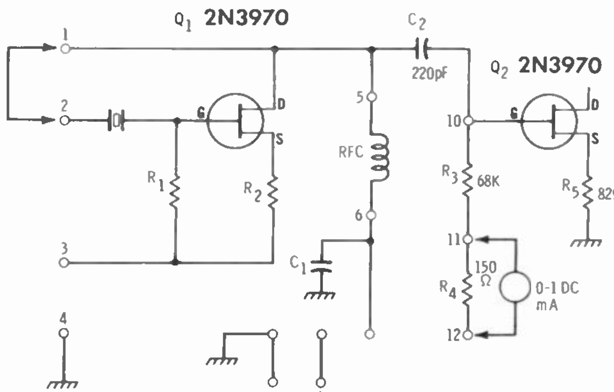


Fig. 1-10. Measurement of drive to next stage using gate current.

stage will cause a flow of gate-circuit charges (gate current). This is a rectified component and a dc meter connected across resistor R_4 will give a relative indication of the rf drive. The dc is of course a function of the peak amplitude of the rf drive. The small value resistor (R_4) is used because it is low enough not to affect circuit operation whenever the dc gate-current meter is disconnected. You can now disconnect the output indicator.

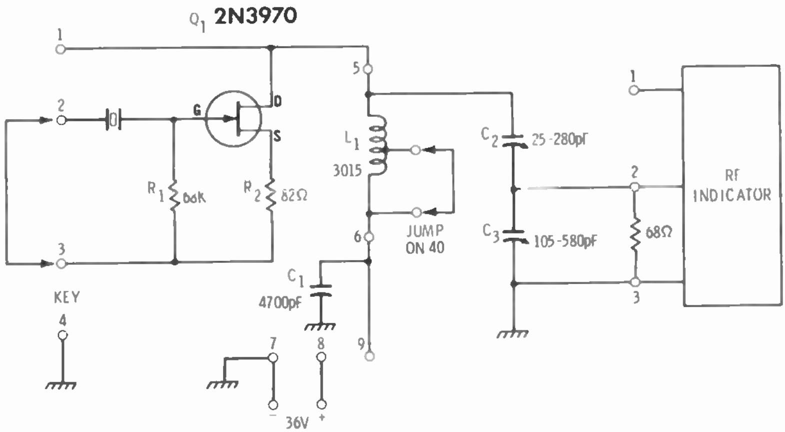
Insert an 80-meter crystal and connect a 0-1 milliammeter (or appropriate current scale of vom) between posts 11 and 12. Turn on the crystal oscillator and note the gate-current meter reading. Check on all bands. Readings should average around 0.2 milliampere.

Remove the crystal from its socket. What happens to the gate current? Radio-frequency drive is removed and there is no rectified gate current.

What happens when the FET of the second stage is removed? There is no gate current again and no rectified current that the meter can record. Remove the second stage FET from its socket while you test still another crystal oscillator circuit.

Miller Crystal Oscillator

A Miller tuned crystal oscillator (Fig. 1-11) can be assembled quickly by connecting the jumper between posts 2 and 3 rather than



C₁ 4700-pF disc capacitor
C₂ 25- to 280-pF trimmer capacitor (ARCO 464)
C₃ 105- to 580-pF trimmer capacitor (ARCO 467)
L₁ Miniductor, Barker & Williamson 3015

Q₁ Siliconix 2N3970 FET
R₁ 68K ½-watt resistor
R₂ 82-ohm ½-watt resistor

Miscellaneous Parts

Pegboard
Crystal socket
Binding posts

Fig. 1-11. Miller crystal oscillator.

between 1 and 2. The resonant coil is connected between posts 5 and 6. A split trimmer capacitor arrangement connects between posts 5 and common (ground). This split-capacitor tunes the resonant circuit and also serves as an impedance-matching combination.

Connect the indicator across the output again. A 70-ohm antenna system can be synthesized by connecting a 68-ohm carbon resistor between indicator post 2 and 3. Connect the indicator across capacitor C_3 as shown in Fig. 1-11.

Insert an 80-meter crystal and turn on the oscillator. Adjust trimmer capacitor C_2 for resonance as indicated by a maximum reading on the output indicator and a near-minimum dip of the drain current.

Observe the drain current as you tune the oscillator through resonance with capacitor C_2. There is greater stability on the high-frequency side of resonance while on the low-frequency side the oscillator drops out of oscillation very quickly. Response is similar to that of a vacuum-tube crystal oscillator. Jockey the C_2 and C_3 adjustments back and forth until maximum output is obtained. This occurs on the high-frequency side of the setting that produces maximum drain-current dip. Optimum drain current again falls somewhere between 25 and 35 milliamperes. Output indicator reading approaches 0.1 milliampere.

Insert a 40-meter crystal and short out a portion of the resonant coil by inserting the tap into binding post 6. Adjust capacitor C_2 for resonance and then vary both trimmers until maximum output indication is obtained.

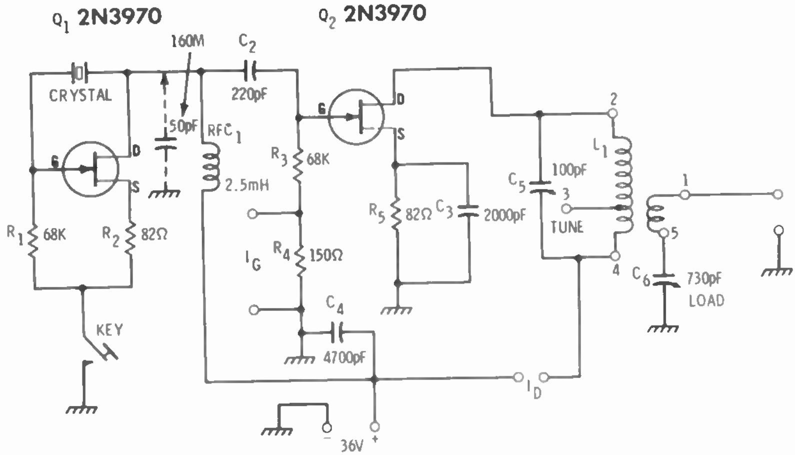
If a 70-ohm antenna system is now connected across capacitor C_3, this oscillator can be used as a very-low-power QRPP transmitter. Dc input power approaches 1 watt and the power output is several hundreds of milliwatts.

Restore the 80-meter coil. Disconnect the output indicator. Connect the 0-1 milliammeter between posts 11 and 12 (Fig. 1-10). Insert the second stage FET into its socket.

Turn on the oscillator. Tune the output resonant circuit for maximum gate current. Note that it is significantly higher than that obtained from the Pierce crystal oscillator. A tuned crystal oscillator will usually give you more drive to the succeeding stage. It has the disadvantage of requiring an additional tuning adjustment.

PROJECT 3: ONE-WATT, TWO-STAGE MULTIBAND FET TRANSMITTER

Two 2N3970 FET's connected as a crystal oscillator and follow-up class-C amplifier can develop approximately ½-watt output on bands 20 through 160 meters. Only a single resonant transformer is



C_1	50-pF disc capacitor	R_2,R_5	82-ohm ½-watt resistor
C_2	220-pF disc capacitor	R_4	150-ohm ½-watt resistor
C_3	2000-pF disc capacitor	RFC$_1$	2.5-mH rf choke
C_4	4700-pF disc capacitor		Miscellaneous Parts
C_5	100-pF variable capacitor		Pegboard (Project 2)
C_6	2-section variable capacitor, 365 pF per section		Crystal socket
			Transistor sockets
L_1	See coil table		5-prong tube socket
Q_1,Q_2	2N3970 FET's		12-volt lantern batteries
R_1,R_3	68K ½-watt resistor		Binding posts

Fig. 1-12. Two-stage FET transmitter.

needed when a Pierce crystal oscillator is used (Fig. 1-12). The class-C amplifier is self-biased by the signal from the crystal oscillator. Excessive drain current is avoided by the use of a source resistor. Only the oscillator needs to be keyed for cw operation.

Construction and Circuit

The basic pegboard is used as shown in Fig. 1-13. The drain TUNE capacitor is a 100-pF variable; the LOAD capacitor which tunes the secondary of the coupled coils is a two-gang 365-pF variable.

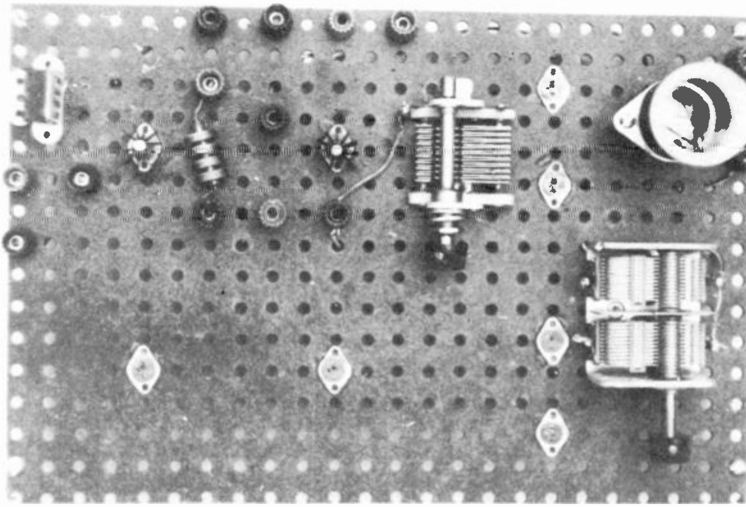


Fig. 1-13. Layout of the two-stage transmitter.

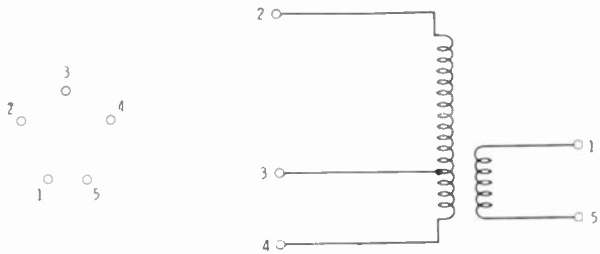
The two FET's and their TO-18 heat sinks can be seen in the photograph.

The coils for the various bands plug into a five-prong tube socket. The coil forms are 1⅛ inch in diameter. The winding is unusual with the secondary bifilar-wound between the same number of turns of the primary starting from the bottom of the coil.

Detail is shown in Fig. 1-14. For example, on 80 meters, the primary has a total of 40 turns. The total number for the secondary is 13 turns but these are wound in between the bottom 13 turns of the primary. The primary is tapped at the 13th turn from the bottom and the tap is connected to pin 3 of the coil form. This tap is not used in the transmitter just described but it will be useful in many of the projects that follow. While you are winding the coils you may as well bring out the tap.

As shown in Fig. 1-12, pin 2 connects to the drain and pin 4 to the drain supply voltage. Pin 1 is the high side of the secondary output while pin 5 connects to the stator side of the variable load capacitor.

The coils are close wound of appropriate size enameled copper wire. This type of construction provides efficient coupling between primary and secondary and the effective transfer of power from the drain circuit to the load.



Regular

BAND	PRI.	TAP	SEC.	WIRE SIZE
160	65 turns	20 turns	20 turns	No. 26 enam.
80	40 turns	13 turns	13 turns	No. 24 enam.
40	21 turns	7 turns	7 turns	No. 22 enam.
20	11 turns	4 turns	4 turns	No. 22 enam.
15	8 turns	3 turns	3 turns	No. 20 enam.
10	5 turns	2 turns	2 turns	No. 20 enam.

Fig. 1-14. Coil data.

Construct the circuit of Fig. 1-12. Use the necessary binding posts to permit the measurement of gate current as in the previous project and also the drain current as well. The output indicator with a 68-ohm termination is connected across the transmitter output.

Tuning FET Transmitter

Insert the 80-meter coil into the socket. Plug in an 80-meter crystal and place the oscillator in operation. Check to see that the gate current is normal according to the previous project.

Set the load capacitor for minimum capacitance. Observe the drain meter as the TUNE capacitor is varied through resonance. Note the drain dip which is similar to that of a vacuum tube. There will also be an output reading on the meter of the rf output indicator.

Vary the LOAD capacitor to peak the indicator current reading. Retune the TUNE capacitor for an output peak. Jockey the two controls back and forth until a maximum output is indicated.

Recheck the gate current. It will be considerably lower than that obtained when there is no load on the output stage.

The transmitter output can now be supplied to a 50- to 70-ohm antenna system. The 68-ohm resistor of the output indicator has been functioning as a dummy antenna load. This resistor should now be disconnected, but not the indicator. The rf indicator can now be used to tune up your transmitter to the antenna. Refer to Fig. 1-15.

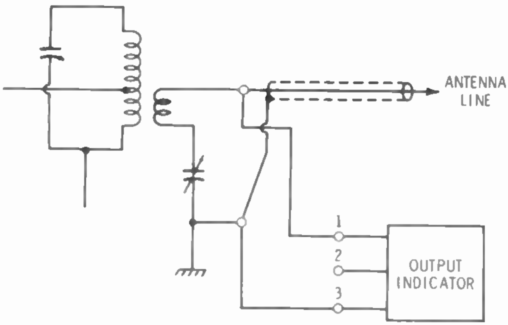


Fig. 1-15. Using the rf indicator to tune transmitter to an antenna.

In placing a load on the transmitter it is apparent that the d-c drain current drawn by the FET class-C amplifier rises. When the transmitter is matched to an antenna or to the dummy load, read the drain-current meter reading and the supply voltage. What is the dc input power to the class-C stage?

When using 36 volts a typical drain-current reading might be 45 milliamperes. In this case the dc input power is:

$$\text{Dc input} = V_{CC} \times I_D = 36 \times 0.045 = 1.62 \text{ watts}$$

Decrease the supply voltage to 24. Retune the TUNE and LOAD capacitors. What happens to the output?

Output falls because of the decrease in drain voltage and drain current. Just as in vacuum-tube practice, the drain voltage can be varied by an audio signal to modulate the rf output.

Restore the 36-volt operation. Retune the transmitter for maximum output. Measure the drain current. What happens to the drain current when the crystal is removed from its socket? There is a small drop in drain current with a loss of excitation. With the crystal out of its socket, momentarily shunt a 47-ohm resistor across the 82-ohm source resistor. Explain what happens.

In this case the source resistance is not high enough and under no excitation the drain current rises to an unsafe value—a value that might exceed the safe dissipation limit of the FET.

Operation on Other Bands

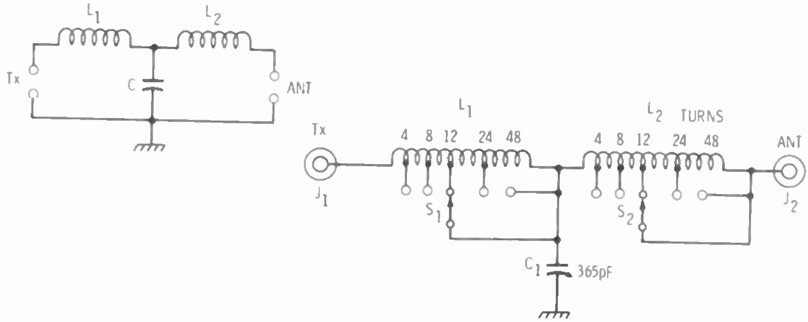
The transmitter can be operated on any one of the four bands. Construct the appropriate coils. A suitable crystal must be available for each band. Fundamental crystals perform well on each band 20 through 160 meters. On 40, 80, and 160 meters obtainable outputs are ½ watt and higher. Less output is obtained on 20 meters. Operating drain currents fall between 35 and 60 milliamperes.

PROJECT 4: LINE TUNER AND SWR METER

The purpose of a line tuner is to provide the most favorable transmitter loading although the impedance looking into the transmitter end of the antenna transmission line is not optimum. Such a line tuner also permits a given antenna to be used at a frequency removed from the limited frequency range for which an antenna presents optimum load conditions for a transmitter. It also permits the loading of a random length of antenna wire or permits a given antenna type to be operated on more than one amateur band. Such a tuner adds convenience and versatility to experimental work with solid-state, vacuum-tube, and low-power transmitters.

There is still another application, and a very important one, when using solid-state circuits. Solid-state devices are low-impedance ones and, especially when operated at other than very low power, are inclined to generate harmonics and other spurious signals. A tuner is a frequency-selective device and is very effective in removing, or substantially reducing, harmonics and other high-frequency components. Thus a transmitter ideally matched to an antenna system should be augmented with such a tuner because of its ability to restrain spurious radiations.

The T-network forms a simple and useful tuner because of its ability to accommodate a wide range of antenna system resistances as well as random length wires. Its also able to match a considerable range of transmitter output impedances. In general (Fig. 1-16) inductor L_2 at the transmission line (antenna) end of the tuner matches the antenna system impedance to the tuner. Conversely, inductor L_1



C_1	365-pF variable capacitor	S_1,S_2	Rotary switches, 1-pole, 5-position
J_1,J_2	Coaxial receptacles, SO-239		6" x 4" x 3" metal case
L_1,L_2	Miniductors, Barker & Williamson 3015		

Fig. 1-16. T-network antenna tuner.

establishes the matching between the tuner and the transmitter and tunes out reactive components reflected from the antenna system. Although there is some interaction between the two sections, this can be minimized by mounting the two inductors at right angles.

Precise matching is accomplished by first selecting the two most favorable tap positions and then adjusting the variable capacitor for minimum SWR.

Tuner Construction

Inasmuch as the tuner is to be used with QRP power not in excess of 75 watts, it was built into a 6 × 4 × 3 aluminum box. The T-network capacitor is a small 365-pF variable. The two coils are Barker and Williamson type 3015 miniductors with appropriate soldered taps as shown in Fig. 1-16. Two five-position nonshorting switches provide optimum tuning conditions for 6 through 80 meters.

The mounting of the coils, switches, and variable capacitor are apparent in Fig. 1-17. The *Ten-Tec* SWR meter is bolted to the top of the tuner (Fig. 1-18). This SWR meter responds to rf outputs of only a fraction of 1 watt, and is ideal for checking out and matching both QRP and QRPP transmitters to an antenna system.

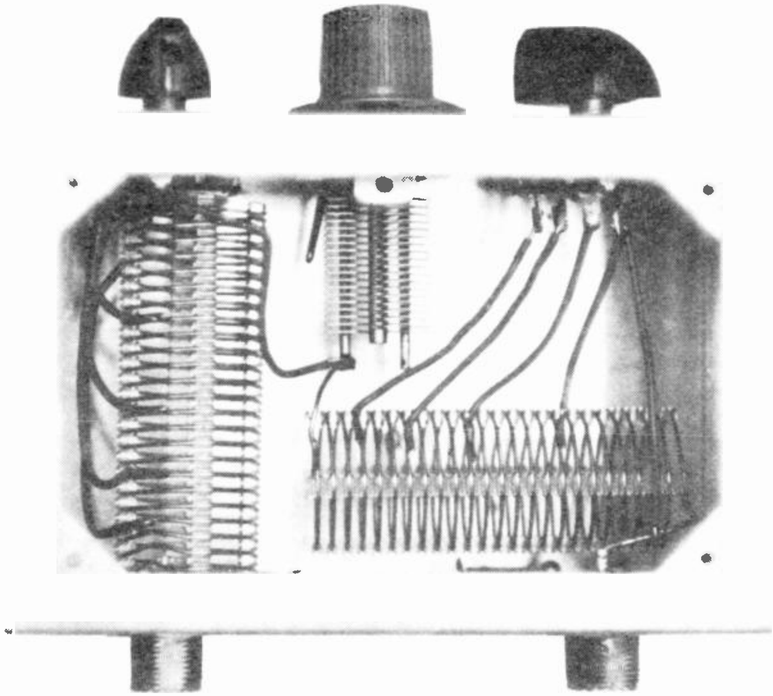


Fig. 1-17. Interior layout of T-network tuner.

Operation

You can gain experience using the tuner and SWR meter by connecting the output indicator with the 68-ohm termination to the output of the tuner. The SWR meter connects between the transmitter and the tuner as shown in Fig. 1-19. Set the tuner switches for maximum inductance on each side. Insert a 40-meter crystal and tune up the transmitter. Adjust the transmitter TUNE and LOAD for maximum reading on the indicator.

Put the SWR meter in the FORWARD position and adjust its sensitivity control for maximum deflection along the SWR scale but not in excess of maximum (don't pin the meter). Go to the REFLECTED position and adjust the tuner capacitor for maximum dip. Retune the transmitter for maximum output. Check the FORWARD position of the SWR meter and lower its sensitivity if necessary to obtain no more than maximum-scale deflection. Change over to REFLECTED position and adjust the variable capacitor of the tuner for minimum SWR.

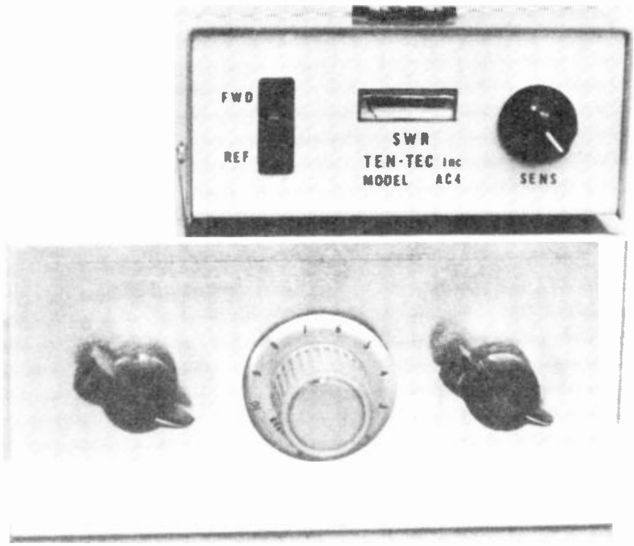


Fig. 1-18. Tuner with Ten-Tec SWR meter.

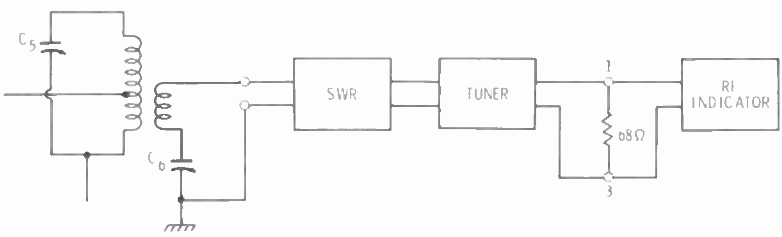


Fig. 1-19. Connection plan to gain experience with tuner and SWR meter.

Try the next lower position for each switch. Note that the SWR does not swing to as low a position and the output is not as great.

Use the same procedure to match the transmitter to an antenna. In doing so disconnect the dummy termination resistance of the rf indicator. Depending on the antenna characteristics, you may have to retune the tuner capacitor. You may have to reset the input or output switch to the next lower position to attain maximum indication on the meter and minimum SWR. Try the combination on other bands.

PROJECT 5: MULTISTAGE, MULTIBAND TRANSMITTER

A three-stage cw transmitter can be constructed on the pegboard as shown in Fig. 1-20. Schematically (Fig. 1-21) it consists of a

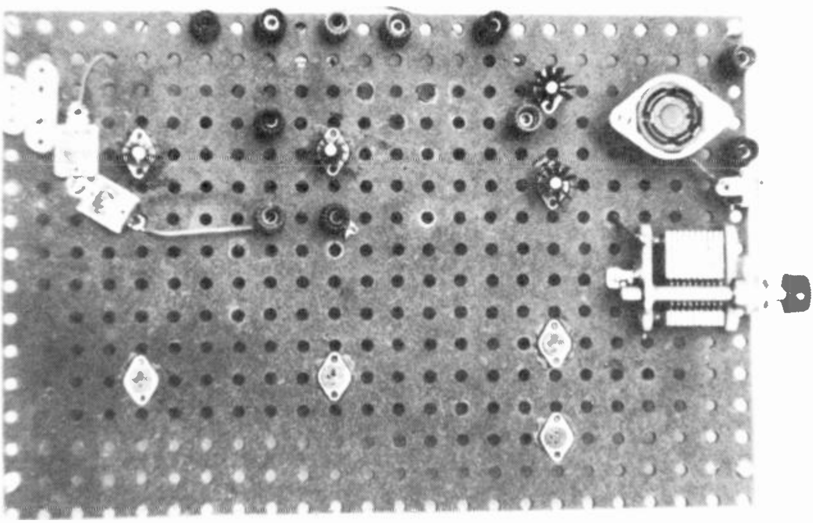


Fig. 1-20. Three-stage cw transmitter, pegboard layout.

Pierce crystal oscillator, buffer, and a final rf amplifier using two FET's in parallel. The buffer stage provides isolation between the oscillator and amplifier and will permit amplitude modulation (a-m) of the final using the modulator of Project 6. The transmitter has a one-watt output capability on 40, 80, and 160 meters.

The Pierce oscillator includes two trimmers, C_1 and C_2. Trimmer C_1 can be adjusted for maximum output and easy crystal starting. Capacitor C_2 can be used to regulate the drive and ensures maximum output along with the development of a pure sine wave (minimum harmonic output). The buffer stage is a source follower presenting a high impedance to the crystal oscillator output and a low impedance feed to the amplifier. It provides a high order of isolation and minimizes the influence of any modulation on the crystal frequency.

The output impedance of the two parallel FET's is quite low and efficient tank-circuit operation is obtained by using the tap as a connection point for the drains. Coil data for the primary and bifilar-connected secondary are given in Project 3. The tuning capacitor is a 100-pF variable. The secondary load capacitor is a small trimmer.

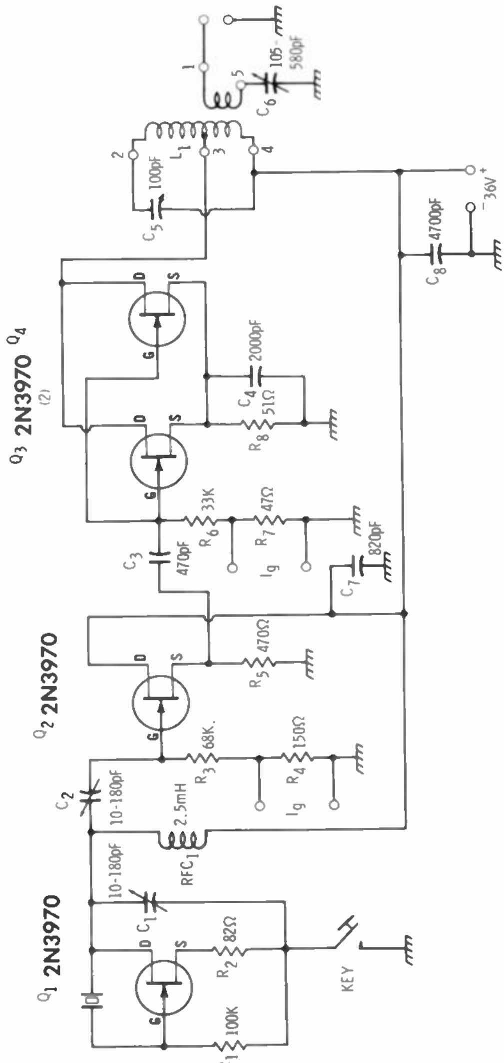


Fig. 1-21. Three-stage cw transmitter schematic.

Operation

Insert the 160-meter crystal and the 160-meter coil. Set trimmers C_1 and C_2 for maximum capacitance and C_6 about midway.

Insert the drain-current meter (0-100 milliamperes approximately). Connect the output indicator across the transmitter output using the 68-ohm termination.

Turn on the transmitter and adjust capacitors C_5 and C_6 for maximum output. Jockey the two controls back and forth to obtain peak rf output.

Slowly decrease the setting of trimmer capacitor C_2 and observe that the output does increase. Retune the output tune and load capacitors as you keep decreasing the value of capacitor C_2. Keep adjusting for maximum output.

Now adjust capacitor C_1 for maximum output. It may be necessary to jockey back and forth between C_1 and C_2 to obtain the best setting. Finally retune C_5 and C_6.

If you have an ordinary service-type oscilloscope or better, connect it to the output. A good-quality sine wave can be observed. To do so, it is necessary to set your horizontal oscilloscope sweep to its highest frequency setting. Also increase the oscilloscope horizontal gain for maximum. Now adjust the sweep vernier on the oscilloscope until a group of sine waves can be observed. Disconnect the oscilloscope.

What is the dc input power? It will probably be in the 2- to 2.75-watt range.

Disconnect the drain-current meter. Key the transmitter and listen on your receiver. If the keying is sluggish, increase the setting

Parts List for Fig. 1-21.

C_1,C_2	Trimmer capacitor, 10 to 180 pF, ARCO 463	R_3	68K ½-watt resistor
C_3	470-pF disc capacitor	R_4	150-ohm ½-watt resistor
C_4	2000-pF disc capacitor	R_5	470-ohm ½-watt resistor
C_5	100-pF variable capacitor	R_6	33K ½-watt resistor
C_6	Trimmer capacitor, 105 to 580 pF, ARCO 467	R_7	47-ohm ½-watt resistor
C_7	820-pF disc capacitor	R_8	51-ohm 2-watt resistor
C_8	4700-pF disc capacitor	RFC_1	2.5-mH rf choke
L_1	Coil set (Project 3)		Miscellaneous Parts
Q_1,Q_2,Q_3, Q_4	2N3970 FETs, (Siliconix)		Pegboard (Project 2)
R_1	100K ½-watt resistor		Crystal socket
R_2	82-ohm ½-watt resistor		5-prong tube socket
			Crystals 20,40,80,160
			2 12V lantern batteries
			Binding posts

of your capacitor C_1 a limited amount. Rf power output should not decline more than 5 to 10%. There should be a noticeable improvement in transmitter keying.

Load up the transmitter on an antenna and you will be ready for some 160-meter QRP contacts. In loading up on the antenna be certain to disconnect the 68-ohm termination. The output rf indicator can be connected in the circuit permanently. Readjust tune and load capacitors C_5 and C_6 for maximum reading.

PRINCIPLES OF AMPLITUDE MODULATION

A field-effect transistor can be modulated in much the same way as a vacuum tube. Drain, gate, or source can be amplitude-modulated. A basic drain-modulation system is shown in Fig. 1-22. It is

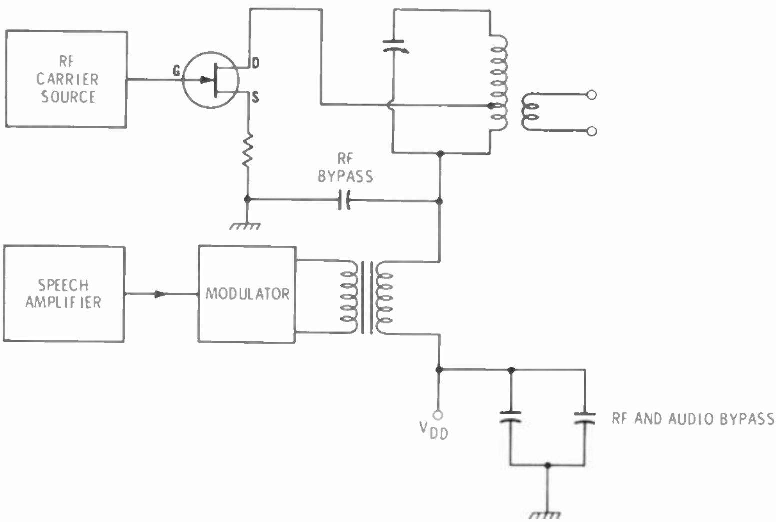


Fig. 1-22. Drain-modulation system.

comparable to the plate modulation of a vacuum tube. The dc drain voltage is made to vary with the modulating signal by applying it through the secondary of the modulation transformer, where it is increased or decreased by the audio (modulating) signal.

The unmodulated radio-frequency carrier is fed to the gate circuit. The mixing of the two signals in the rf amplifier FET's results in the formation of a modulated rf signal across the drain tank circuit.

In the rf stage the tank-circuit voltage varies linearly with the drain supply voltage. Thus, doubling of the supply voltage doubles the tank rf voltage. If the modulated amplifier is operated from a 20-volt supply, the instantaneous drain supply voltage can be made to vary above and below 20 volts. If the modulating wave is a pure sine wave, this variation can be symmetrical.

The rate of variation of the supply voltage depends on the frequency of the modulating wave; the magnitude of the change depends on the amplitude of that wave. In other words, the change in the drain supply voltage becomes a replica of the modulating wave. If the variations on each side of the zero ac axis of the modulating wave are identical, the average dc drain voltage remains fixed at 20 volts.

What influence does the change in drain voltage have on the operation of the modulated amplifier? As the drain voltage changes in a class-C amplifier, so does the peak drain current drawn for each radio-frequency cycle. In fact, the peaks of the latter are a copy of the modulating wave as shown in Fig. 1-23.

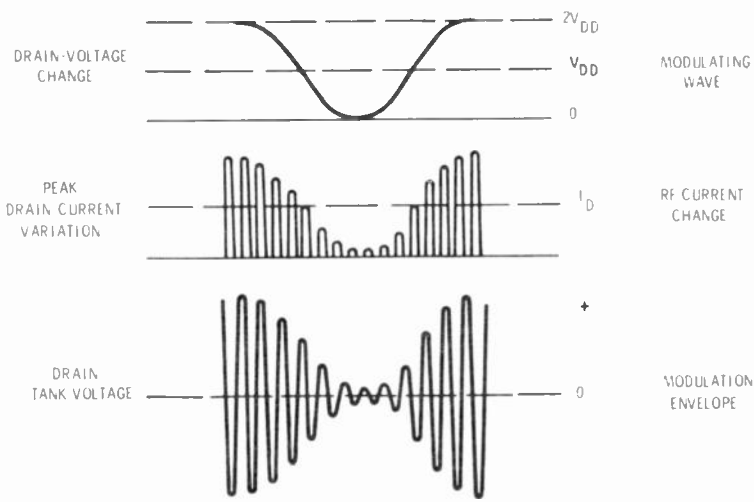


Fig. 1-23. Drain-modulation waveforms showing 100-percent modulation.

The peak drain current, in turn, determines the magnitude of the rf voltage developed across the drain tank circuit. The higher the maximum drain current the higher is the amplitude of the corresponding rf voltage cycle. It follows then that the rf drain voltage waveform is also a copy of the modulating wave.

The extent of the change in the amplitude of the rf output relative to the unmodulated carrier value is usually expressed as a modulation percentage. When the amplitude of the rf voltage swings to twice the amplitude of the unmodulated signal and falls to zero on negative peaks, the carrier is said to be 100% modulated as in Fig. 1-23.

When a carrier is modulated less than 100%, the rf voltage does not rise to twice the unmodulated value, nor does it fall to zero. The equation most often used to calculate modulation percentage is:

$$\%\ \text{modulation} = \frac{E_{max} - E_{min}}{2E_c} \times 100$$

where,

E_{max} is the maximum amplitude of the modulation envelope,
E_{min} is the minimum amplitude of the modulation envelope,
E_c is the amplitude of the unmodulated carrier.

Practical signals are not of constant amplitude so modulation is not normally maintained at 100%. Only on occasion do voice signal peaks rise to the maximum value. Nevertheless an amplitude-modulation system must be adjusted to maintain as high a modulation percentage as possible, but the voice peaks should not exceed 100%.

The result of the modulation process is a composite waveform; each rf cycle differs from the preceding and following one. The resultant signal is no longer a single rf carrier but, in the case of modulation by a single sine wave, a carrier plus upper and lower sidebands. The carrier, despite modulation, remains constant in magnitude and frequency. The only difference has been the addition of two sideband components.

The foregoing relationships indicate that the carrier power is unchanged with modulation, and the rise in total output power is contributed by the two sidebands. Furthermore the greater the magnitude of the two sidebands, the greater is the increase in the total rf output. In the case of the drain-modulation system this extra power must come from the modulator.

In fact, the rf power output of the modulated amplifier is increased by 50% when the carrier is modulated 100% by a sine wave. This additional power comes from the modulator. Since only occasional voice peaks extend to 100% modulation, the average power output of the modulator is less than 50% of the dc power input to the modulated amplifier.

For 100% modulation the peak audio voltage must be approximately equal to the supply voltage. Under this condition the drain voltage swings to twice the drain supply voltage on peaks. On the negative sweep the drain voltage drops to zero. To produce the desired peak audio voltage at the desired power, the audio must be developed across a specific impedance. The impedance seen by the modulator output equals the dc resistance of the modulated-amplifier input. This dc input resistance is related to the dc component of drain current and drain voltage:

$$R_{in} = \frac{V_{DD}}{I_D}$$

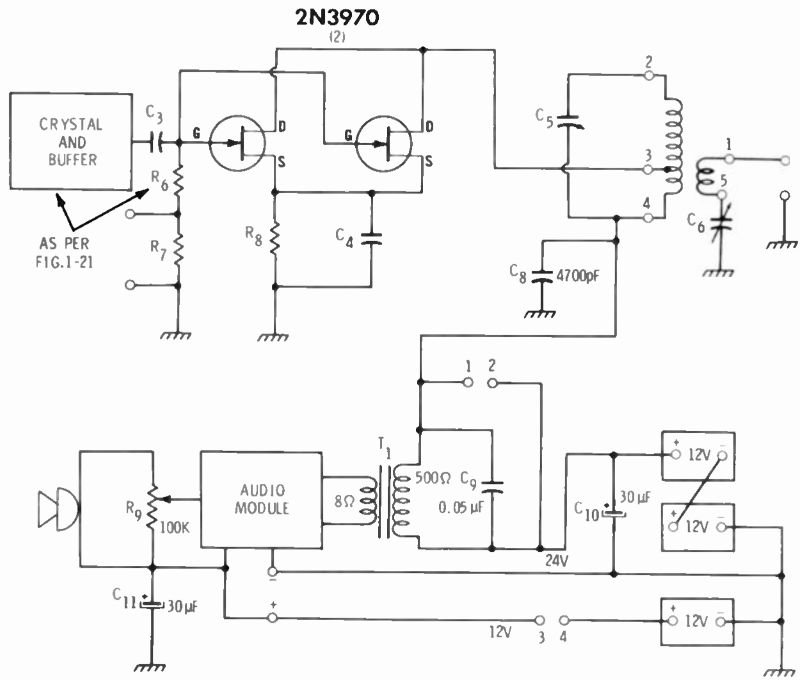
PROJECT 6: SOLID-STATE A-M MODULATOR

A variety of low-cost audio modules are available from electronic supply houses. They are used mainly to drive small speakers and have a low-impedance output. However, they can be used as modulators by including a modulation transformer between their output and the modulation input of the modulated amplifier, (Fig. 1-24). In this case, an audio module with output power of 2 to 3 watts is more than adequate. A small modulation transformer is inserted between its output and the modulated amplifier. This can be the type used widely in vacuum-tube Citizens Band transceivers. Other audio transformers can be used too if they provide the appropriate impedance match.

The audio module, modulation transformer, audio gain control, and microphone binding posts are mounted on the pegboard as shown in Fig. 1-25.

To attain adequate modulation, the supply voltage of the FET transmitter is reduced to 24 volts. If the dc drain current is 50 milliamperes when the transmitter is loaded properly, the input resistance that must be matched by the modulator is:

$$R_{in} = \frac{24}{0.05} = 480 \text{ ohms}$$



C_9 0.05-µF disc or paper capacitor
C_{10},C_{11} 30-µF 50-volt electrolytic capacitors
R_9 100K potentiometer
T_1 Modulation transformer 500 ohms to 8 ohms (Lafayette 99-6132)

Audio Module, 2-3 watts (Philmore AA900, Lafayette 99-91324, etc.)
Microphone
3 12-volt lantern batteries (Eveready 732)

Fig. 1-24. Modulator for FET transmitters.

Thus the modulation transformer should be such that 8 ohms is matched to approximately 500 ohms for the audio module used in Fig. 1-24. This match can be obtained with the recommended modulation transformer by connecting its 8-ohm secondary to the output of the audio module and its 500-ohm primary to the modulated amplifier as shown. Ratios can be off as much as two or more to one and adequate modulation can be obtained if the audio module makes available a little extra power. Under a matched condition the actual audio power required directly at the modulated amplifier input is only 0.6 watt ($24 \times 0.05/2$). The output of the audio module must be greater than this because of some additional loss in the modula-

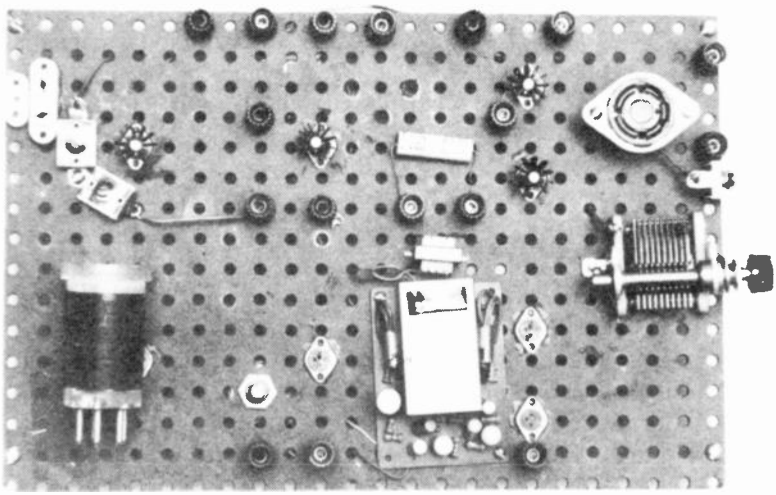


Fig. 1-25. Pegboard layout of FET transmitter and solid-state modulator.

tion transformer and the likelihood of not obtaining a precise match.

Operation

Assemble the modulator on the pegboard. Make arrangements with switches or binding posts and jumpers to take the audio modulator out of the circuit for initial tune up and for cw operation. In the pegboard arrangement of Figs 1-24 and 1-25, the audio system is deactivated by inserting a jumper between binding posts 1 and 2 and removing the jumper from between binding posts 3 and 4. Do this initially in placing the transmitter in operation.

Insert a 160-meter crystal and the 160-meter coil. Turn on the rf section and tune for maximum rf output as described in Project 5. Now remove the jumper from between binding posts 1 and 2. Retune the transmitter slightly for maximum output.

Tune in the signal on your receiver. Turn on the modulator. Turn up the microphone gain and speak into the microphone. Good voice reproduction should be obtained. As you speak into the microphone there should be a slight kick in the dc drain current and in the meter reading of the rf indicator. There should not be a drastic downward kick however.

Better modulation capability is obtained when the rf output is reduced about 10 to 20 percent by increasing the capacitance of the loading capacitor (C_6). It also helps to add additional capacitance at C_2, if available. When C_2 is adjusted it may be necessary to re-adjust C_1 for maximum oscillator output. Increasing C_2 provides some additional drive to the modulated amplifier and improves its ability to handle modulation peaks (good upward modulation).

An oscilloscope, even if it is only of the service type, can give you a better idea of the modulation quality when it is connected across the rf output of the transmitter under load. You can also make a better observation if your audio input is a sine-wave tone rather than a voice signal. A 1000-Hz tone signal of good sine-wave form is useful in obtaining a detailed picture of modulation activities.

Repeat the process on 40 and 80 meters. Load up to an antenna system and give QRP a-m transmission a whirl.

CHAPTER 2

Bipolar CW and A-M Transmitter Circuits

In this chapter you are introduced to the operating characteristics of bipolar transistors in transmitter circuits. The bipolar transistor is now used widely in modern commercial transmitters as well as some amateur transmitters. In fact for vhf-uhf commercial applications transistors have largely replaced vacuum tubes.

The projects of this chapter permit you to build up a variety of bipolar oscillator and class-C amplifier stages. In addition to crystal-controlled oscillators you can also work with a bipolar variable-frequency oscillator (vfo). A multiband bipolar transmitter can be constructed. Higher-powered bipolar amplifiers can be added on to the basic multiband cw transmitters.

BIPOLAR TRANSISTOR

A bipolar transistor is a three-element device consisting of emitter, base, and collector, (Fig. 2-1). The emitter and collector elements are positioned on each side of the base. In this position the base is able to control the motion of charges that flow between emitter and collector. Actually the bipolar transistor has two pn junctions, emitter and collector. The emitter junction between the emitter and the base is forward-biased; the collector junction, between base and collector is reverse-biased. When the emitter junction is

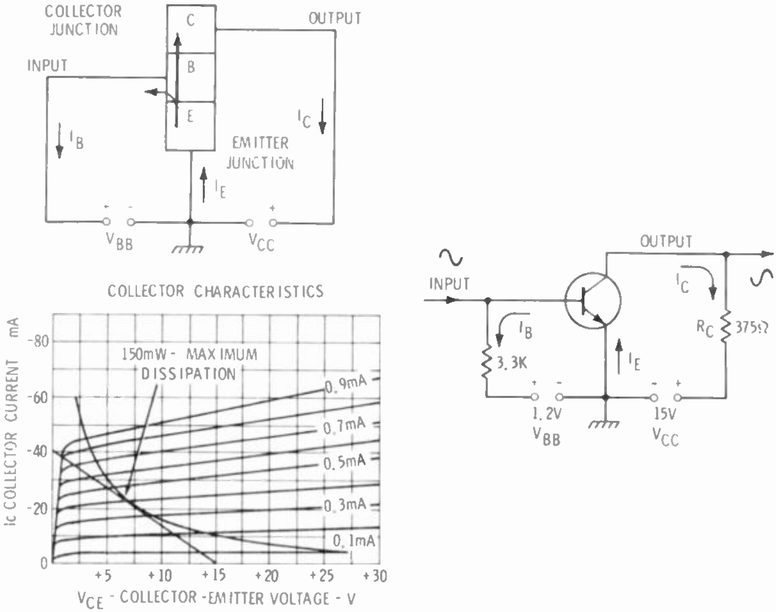


Fig. 2-1. Basic operation of a bipolar transistor.

forward-biased there is a motion of charges (current) across the junction. These charge carriers diffuse through the base and cross the collector junction even though it is reverse-biased. Their charge is such that they are attracted across the junction by the potential of the collector. A substantial current results.

The current division is as follows:

$$I_E = I_B + I_C$$

where,

I_B is base current,
I_C is collector current,
I_E is emitter current.

The ratio between the emitter current and collector current is known as the transistor alpha (α):

$$\alpha = \frac{I_C}{I_E}$$

The alpha is always less than 1.

Note that the collector current is greater than the base current. The ratio between these two currents is known as the *figure of merit* or beta (β) of the transistor.

$$\beta = \frac{I_C}{I_B}$$

The collector current of a bipolar transistor is related to the collector voltage and the base current as shown in the family of collector characteristic curves. As the base bias current is increased with more emitter-junction forward bias, the collector current rises. If the base current is made to vary with signal there results a like and amplified change in the collector current. In the case of the common-emitter circuit of Fig. 2-1, there is also a substantial voltage change across the output. This is greater than the voltage change applied to the input of the circuit. Therefore the common-emitter stage has voltage gain.

The input emitter junction of a bipolar transistor is forward-biased. This junction has a high conductance and a significant base current is present. As a result the input resistance of a bipolar circuit is low, a factor that must be taken into consideration in terms of the loading influence it has on the preceding stage. The interstage coupling system must be designed to make a suitable match to the low input resistance of a bipolar stage. Likewise the preceding stage must supply a signal of adequate power level if the stage is to operate in a normal manner. Conversely, a field-effect transistor or a vacuum tube has an input system that is reverse-biased. There is only an insignificant amount of grid or gate current. This type of stage has a high input resistance and requires less driving power.

There are two fundamental types of bipolar transistors; those with a p-type base and those with an n-type base (Fig. 2-2).

In transmitter rf stages the npn type is more common than the pnp type. For proper operation of an npn transistor a positive voltage is applied to the collector. As a result the collector junction is reverse-biased, the base being negative with respect to the collector. The emitter junction must be forward-biased. As a result the emitter must be made negative relative to the base. A positive bias on the base forward-biases the emitter junction.

Since the base is made positive with respect to the n-type emitter,

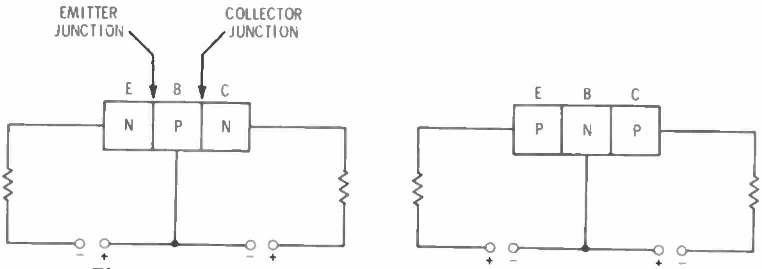


Fig. 2-2. Biasing of npn and pnp bipolar transistors.

the electron charges of the emitter cross the junction into the base. These electrons diffuse through the base and are attracted by the positive voltage of the collector. Therefore they cross the collector junction and flow into the output circuit.

Conversely, the pnp transistor must be arranged in such a manner that a negative voltage is applied to the collector. In so doing the collector junction is reverse-biased. Likewise, the base must be made negative relative to the emitter in order to forward-bias the emitter junction. In this case the charge carriers in the emitter are *holes*, or positive charges. They are attracted by the negative biasing of the base and cross the emitter junction into the base. The positive charges diffuse through the base and are attracted by the negative potential of the collector. They cross the collector junction into the output circuit.

BIPOLAR CLASS-B and CLASS-C AMPLIFIERS

Bipolar transistors, like vacuum tubes, can be operated in class A, class B, or class C. When a bipolar transistor is forward-biased class A, the bias point is usually centered on the linear portion of the transfer characteristic (Fig. 2-3). In class-B biasing the bias point is collector current cutoff. In class-C biasing the base-emitter junction is reverse-biased beyond cutoff.

Except for class B push-pull operation, a low-frequency amplifier biased class B or class C produces a distorted output. The output wave is no longer a good replica of the input wave. The above distortion restriction does not apply to tuned radio-frequency amplifiers and it is possible to bias a bipolar transistor amplifier class B or class C and still obtain an output wave that is undistorted. This is

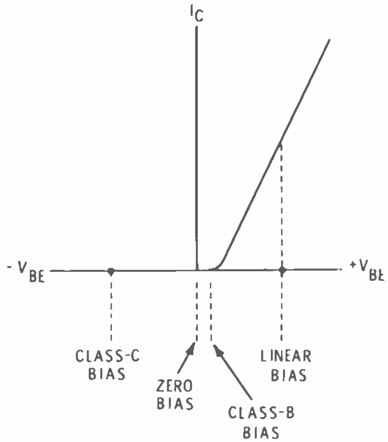


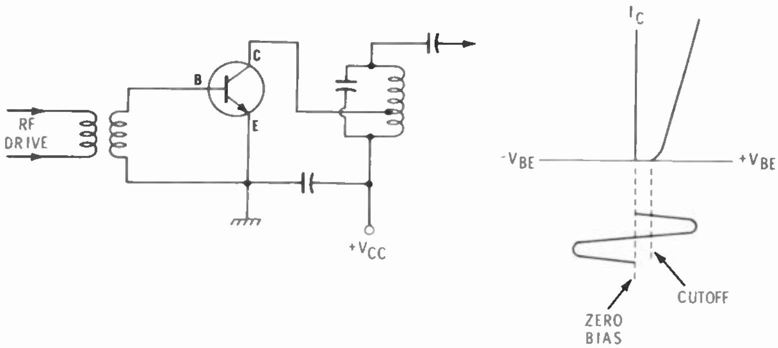
Fig. 2-3. Class biasing of a bipolar transistor.

a result of the energy-conserving ability of the output resonant circuit. Class-B and class-C amplifiers are more efficient than class-A types with obtainable efficiencies of 65% and higher in well-designed circuits.

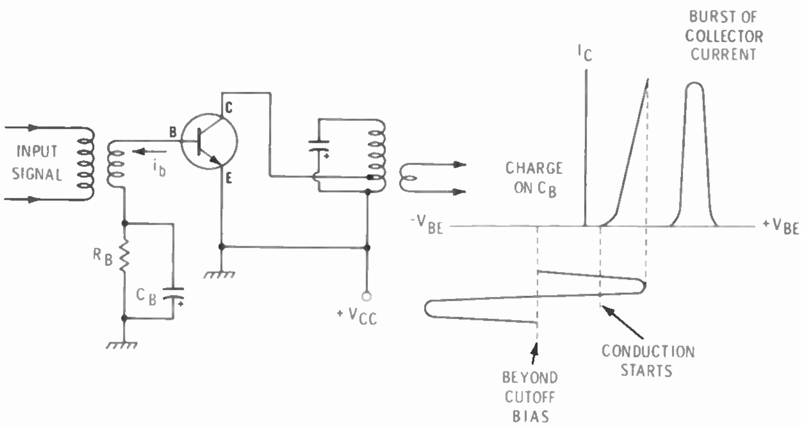
When a bipolar is biased class B or class C the collector output current is present for only a portion of one alternation of the input rf wave. Thus, the output current variation is a very much distorted version of the input rf signal. However, the presence of an output resonant circuit of appropriate characteristics results in an rf output voltage that is a reasonably undistorted version of the input voltage. Two practical bipolar class-C amplifiers are shown in Fig. 2-4. When a bipolar transistor is connected with no forward bias present at its emitter junction it is already biased class C. There is no collector current without the application of at least a small amount of forward bias. For germanium and silicon transistors the required forward-biasing is approximately 0.2 volt and 0.7 volt respectively.

In the circuit arrangement of Fig. 2-4A the stage is biased slightly class C by simply using no base-to-emitter junction bias whatsoever. The amount of biasing is only slightly beyond the value of the class-B cutoff bias. This technique is used frequently in bipolar class-C power amplifiers.

In the arrangement of Fig. 2-4B there is a base circuit resistor-capacitor combination that establishes the cutoff biasing. In the case of the npn stage shown, the more positive portion of the input



(A) No base-to-emitter bias.



(B) Base circuit resistor-capacitor combination.

Fig. 2-4. Typical class-C amplifiers.

wave forward-biases the junction and results in a base current. The direction of the base current is such that the charge placed on the capacitor is a back bias and the emitter junction is reverse-biased significantly beyond cutoff value.

Approximate class-C waveforms are shown in Fig. 2-5. During the portion of the input wave that biases the emitter junction in a forward direction there is a strong but short interval of collector current. This burst of current contributes power to the output tank circuit. At this time, when the tank capacitor is charged to a negative peak, the collector voltage is at minimum value.

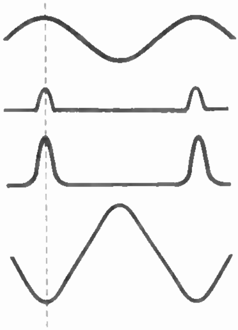


Fig. 2-5. Class-C waveforms.

It is important to recognize that with a strong input signal the peak collector current rises from zero to a very high value. Consequently there can be a substantial change in the collector voltage and the power delivered to the oscillating resonant circuit is much greater than the signal power level delivered to the base of the transistor. The energy delivered to the tank circuit is stretched out because of the energy-storing capability of the resonant circuit. As a result a sinusoidal rf voltage is developed across the tuned circuit.

The Q of the resonant circuit is of significance. The unloaded value should be as high as possible to obtain efficient operation of the tank circuit and to obtain maximum transfer of power from the tank to the load. The loaded Q of the resonant circuit should be relatively low to permit the efficient transfer of power, but it must not be too low or the output waveform becomes distorted (strong harmonics).

The attainment of a high unloaded Q is often difficult when using a power transistor because of the inherent low output impedance of the transistor. The problem can be circumvented by the use of an appropriate output resonant circuit.

In the examples of Fig. 2-4, note that the collector is connected to a low-impedance point of the output resonant circuit. Therefore the transistor itself has a much lower loading effect. Likewise the energy for the low-impedance output is taken off by a step-down transformer arrangement in the form of a few-turn secondary winding closely coupled to the low-impedance end of the resonant circuit. Other types of output circuits can be used to provide this low-impedance matching and efficient transfer of power, such as an L-filter, a pi-network, a combination pi and L, etc.

Quite often an emitter resistor is used to avoid thermal runaway and/or prevent too high peak collector current or excessive power dissipation. Time constant of any base resistor-capacitor combination must be long in comparison to the period of the rf wave. In so doing the necessary dc component of class-C bias is developed. Resistive values are lower and power ratings higher, the higher the power dissipation capability of the transistor is.

Most important, recognize that with no bias applied to a bipolar transistor class-C amplifier there is no collector current. This is quite different from vacuum-tube or FET circuits which draw high current when the rf excitation is removed and there is no biasing. Removal of the rf excitation from the input of a bipolar class-C stage, of the type shown in Fig. 2-4, causes the collector current to drop to zero. This is a definite advantage in conserving power and in coming up with a simple way of keying a bipolar cw transmitter. If the oscillator circuit is keyed, for example, each succeeding class-C stage draws no current when the key is up. The oscillator itself can be designed to draw just minimum current with the key up or no more than it draws when it is oscillating.

PROJECT 7: BIPOLAR CRYSTAL OSCILLATORS

The projects of this chapter can be built on a separate pegboard or you can rearrange the pegboard used for the projects of Chapter 1. Again crystal oscillator experiments can be confined to the top left of the board. Later a class-C power amplifier is mounted to its right. An individual rf output indicator and a modulator will also be added. A vfo can be mounted bottom left.

All-Band Pierce Oscillator

Assemble the circuit of Fig. 2-6. The transistor is a 2N3553 which operates well on all bands 10 through 160 meters. It also performs well on 2 and 6 meters, and is quite versatile for radio amateur use.

The circuit is quite simple and uses but a few components. Collector current (dc) can be measured with a meter inserted between posts 3 and 4. Otherwise a jumper can be connected between these two. The load placed on the crystal oscillator is inserted between binding posts 1 and 2. This load can either be resistive or

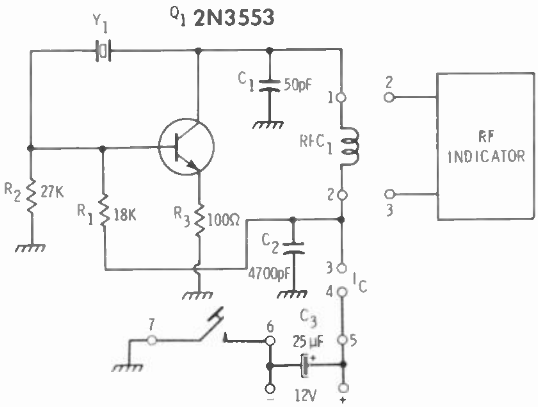


Fig. 2-6. Pierce crystal oscillator.

inductive. The oscillator can be keyed by inserting the key between binding posts 6 and 7. Otherwise a jumper can be inserted between these two points.

To oscillate, a bipolar transistor must be forward-biased slightly. Otherwise it will remain at cutoff and oscillations will not start. This is a function of the base-bias divider resistors (R_1 and R_2). Some stabilizing emitter bias is supplied by resistor R_3.

Operation

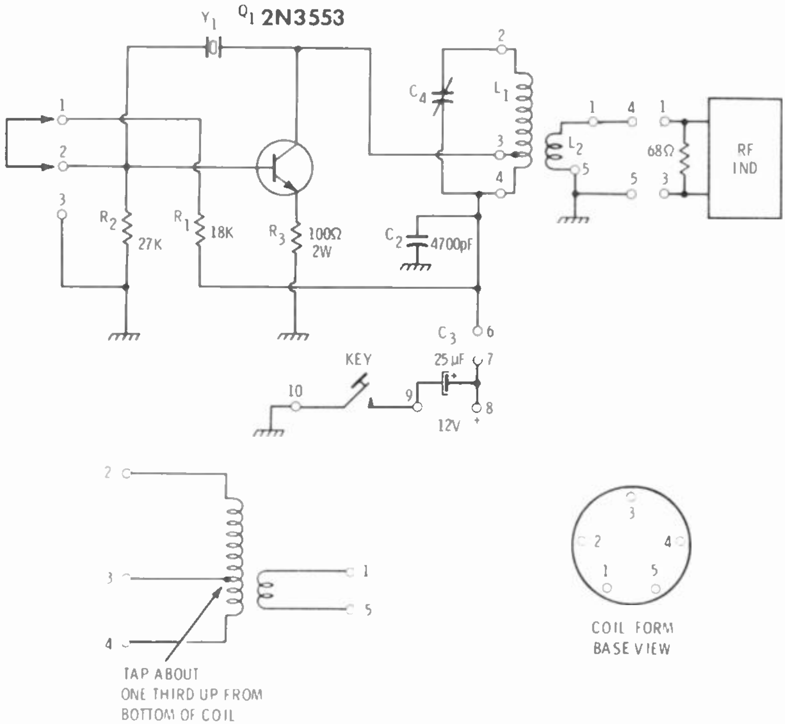
Plug in a 40- or 80-meter crystal. Close the key or connect a jumper between posts 6 and 7. Note the collector current and the output indicator reading on 40 and 80 meters. Tune in the signal on your receiver. Try both bands.

Insert a 160-meter crystal. If it does not start, place an additional 100-pF capacitor across the 50-pF capacitor (C_1) now in the circuit.

Try operation on 10, 15, and 20 meters using the Z-14 radio-frequency choke. This choke presents a more suitable reactance for operating on the high bands. Fundamental crystals must be used in the Pierce circuit.

Modified Pierce Oscillator

The power bipolar is a low-impedance device. More output can be obtained by using an output resonant circuit and an impedance-matching capability. Such a circuit is shown in Fig. 2-7. The resonant cuit for maximum rf-indicator reading. Observe the collector cur-



Closewound on 1¼" Diameter Coil Form

	L_1	L_2
160	70 turns #26 enam. (tap at 20 turns)	8 turns #20 enam.
80	45 turns #22 enam. (tap at 15 turns)	6 turns #20 enam.
40	21 turns #22 enam. (tap at 7 turns)	4 turns #20 enam.
20	11 turns #22 enam. (tap at 4 turns)	3 turns #20 enam.
15	8 turns #20 enam. (tap at 3 turns)	2 turns #20 enam.
10	5½ turns #20 enam. (tap at 2 turns)	2 turns #20 enam.

Fig. 2-7. Modified Pierce with output resonant circuit.

lector circuit. A low-impedance secondary link provides matching to a low-impedance load. This can be an antenna system, the oscillator making available approximately 100-500 milliwatts of output using coils of Fig. 2-7.

The emitter circuit includes a resistor that limits the nonoscillating collector current to a proper value during the key-up position.

A bipolar transistor must be forward-biased to start oscillations. This is the function of the base divider-resistor pair (R_1 and R_2). Binding posts are provided for additional versatility. When operating in bipolar fashion the jumper between binding posts 1 and 2 must be closed. This switches in the desired forward-biasing. If the stage is to operate as an amplifier the crystal is removed from its socket and the drive signal is applied between binding posts 2 and 3. In this case the jumper is removed from between posts 1 and 2. It is also possible to use the 2N3970 FET in this oscillator circuit. To do so the jumper between binding posts 1 and 2 must be removed.

Wire the circuit of Fig. 2-7 permanently because it will be used as a drive source for the complete transmitter. A five-prong tube socket is used for the coils. Coil data is also given. This information is given for each individual band and the coils can be used again in later projects. The 15-meter coil itself permits operation on the 10, 15, and 20 meter bands. The 20-meter coil provides coverage of both the 20- and 40-meter bands. Instead of a variable, a small trimmer is used for the capacitive leg of the output resonant circuit. On 160, shunt it with an additional 100-pF fixed capacitor.

A dc collector-current meter is inserted between binding posts 6 and 7; key, between binding posts 9 and 10. The rf indicator is

Parts List For Bipolar Crystal Oscillators

C_1	50-pF disc capacitor	Q_1	2N 3553 bipolar transistor
C_2	4700-pF disc capacitor	R_1	18K ½-watt resistor
C_3	25-µF 25V electrolytic capacitor	R_2	27K ½-watt resistor
		R_3	100-ohm 2-watt resistor
C_4	140-pF variable capacitor	RFC_1	2.5-mH or Z-14 rf choke (see text)
L_1, L_2	See coil table		

Miscellaneous Parts

Pegboard 8″ x 12″	5-prong tube socket
Crystal socket	5-prong coil forms
Crystals, 20,40,80,160 fundamentals	magnet wire (see coil table)
Transistor socket	12-volt lantern battery
5-prong coil forms	Binding posts

connected across binding posts 4 and 5 (Fig. 2-7). Use the 68-ohm termination resistor across the input of the rf indicator.

Insert a 40- or 80-meter crystal. Tune the output resonant circuit for maximum rf-indicator reading. Observe the collector current with key down and key up.

Tune in the signal on your receiver. Check the keying. It may be necessary to retune capacitor C_1 slightly for best keying.

You can now put the oscillator on the air. Remove the 68-ohm termination of the rf indicator, leaving the rf indicator connected into the output circuit. Also attach the antenna between terminals 4 and 5. It may be necessary to retune capacitor C_1 slightly for good keying and maximum output.

Try the oscillator on other bands. This particular oscillator with its tuned output circuit will function with both fundamental and overtone crystals.

Remove the jumper from between binding posts 1 and 2. Insert a 2N3970 FET into the transistor socket. Check the operation by using a 40- or 80-meter crystal. With a 12-volt battery the rf output will be several hundreds of milliwatts. More output can be obtained by increasing the supply voltage from 12 to 24 volts. Restore the crystal circuit to bipolar operation.

For higher-powered operation of the oscillator you can insert the close-coupled coils of Fig. 1-14. Outputs up to 1-watt plus can be obtained depending on band and crystal activity. Use fundamental crystals. Tuning is more critical and spurious oscillations can develop. Check your bandwidth and keying before going on the air. Retuning of C_1 can make a difference. Use a tuner between transmitter and antenna. Refer to Project 4.

PROJECT 8: BIPOLAR TWO-STAGE TRANSMITTER

The two-stage transmitter is shown schematically in Fig. 2-8. The crystal oscillator is fundamentally the same as that shown in Fig. 2-7. Data for the oscillator coils are given in Fig. 2-7; amplifier coils in Fig. 1-14.

The secondary winding is connected directly into the base-emitter circuit of the rf amplifier. Binding posts 8 and 9 permit you to check the output of the oscillator and input to the amplifier. Both the oscillator and amplifier collector currents can be measured by inserting a meter across the appropriate binding posts. When not measuring collector currents, jumpers are placed across these posts.

Binding posts 10 and 11 have been included in the emitter circuit of the amplifier to permit a change in the emitter resistor value. Depending on the characteristics of the individual transistor the 2.2-ohm resistor is a good optimum. However, if the oscillator tends to self-oscillate strongly or generates spurious signals and noise it is advisable to increase the value of resistor R_4. On 160 meters it may even be necessary to use a value of resistor as high as 12 ohms.

Separate batteries or power supplies for oscillator and amplifier are suggested. The oscillator should not be operated higher than 12 volts as a crystal safety precaution. The amplifier should be tuned up on 12 volts by connecting a jumper between binding posts 14 and 16. Then the jumper can be connected between posts 14 and 15 to increase the amplifier voltage and possible power output. Suggested maximum voltage is 18. Depending on frequency and transistor characteristic, the voltage on the final can be increased with care to as high as 24 volts. Watch out for instability and possible self-oscillation if the two resonant circuits are not effectively isolated from each other. A shield between stages is advisable. An increase in the value of emitter resistor R_4 helps with instability problems with a sacrifice in power output.

The transmitter includes its own rf output indicator. It is in the form of a crystal diode rectifier and a resistor-capacitor filter combination (C_8 and R_5). A small amount of drive for the indicator is obtained by way of the low value capacitor C_7. An increase in the value of C_7 or a decrease in the value of resistor R_5 steps up the level of the rectified current in case more meter deflection is wanted.

Operation

Construct the two-stage transmitter on the pegboard and insert the 80-meter crystal. Supply power to the crystal oscillator after connecting the rf indicator between binding posts 8 and 9. Tune trimmer capacitor C_1 for maximum output.

Connect a 51-ohm two-watt resistor across the output of the amplifier stage. Connect a 0-1 mA milliammeter between binding posts 18 and 19. Also meter the collector-current circuit. Current scale should be approximately 0-600 mA.

Turn on the oscillator power only. Note that there is a very tiny reading on the 0-1 milliammeter. Adjust trimmer capacitor C_1 for maximum. This indicates maximum drive to the amplifier. Now apply 12-volts to the amplifier by connecting the jumper between

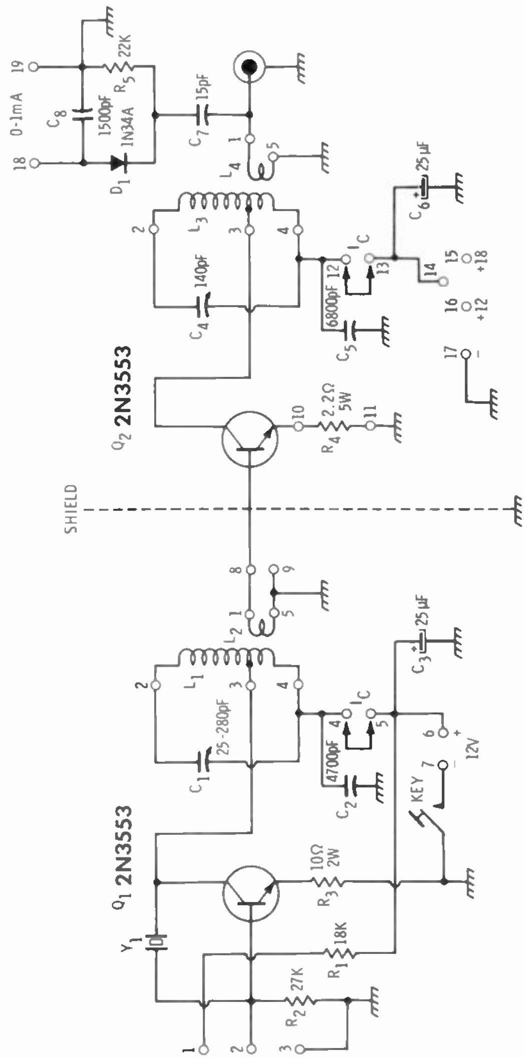


Fig. 2-8. Two-stage bipolar transmitter.

posts 14 and 16. Adjust capacitor C_4 for maximum indication on the 0-1 mA. This may not necessarily be the proper setting for capacitor C_4. Spurious oscillations are a particular trouble with bipolar rf amplifiers.

Usually in varying capacitor C_1 over its range you will find a number of points that indicate output. The one point you are looking for is that one under the control of the crystal oscillator. Usually it will be a smoother response, though not necessarily higher, than the improper ones. Self-oscillations produce unstable and more abrupt jumps in output readings.

The best plan is to use some sort of calibrated absorption wavemeter to determine the signal frequency in the tank circuit. An alternative approach is to use a receiver with an S-meter that has been previously set exactly to the crystal frequency (using the oscillator alone). When the absorption meter rises or the S-meter shows an increase as capacitor C_1 is brought to one of the peaks it is likely to be a correct one.

After the transmitter has been tuned properly momentarily remove the crystal from its socket. What happens to the amplifier collector current? This proves that when excitation is removed from the amplifier all forward-biasing is lost and the collector current is cut off. In fact, you probably noted that as capacitor C_1 or C_4 are peaked on the output meter, the collector current rises when going into the resonant settings.

After the transmitter has been tuned properly you can now increase the supply voltage to the amplifier. You must peak the output

Parts List for Fig. 2-8.

C_1	25-280 pF trimmer capacitor (ARCO 464)	D_1	1N34A diode
C_2	4700-pF disc capacitor	L_1, L_2, L_3, L_4	See coil table (Fig. 1-14)
C_3, C_6	25-µF 25-volt electrolytic capacitor	Q_1, Q_2	2N3553 bipolar transistors
C_4	140-pF variable capacitor	R_1	18K ½-watt resistor
C_5	6800-pF disc capacitor	R_2	27K ½-watt resistor
C_7	15-pF disc capacitor	R_3	10-ohm 2-watt resistor
C_8	1500-pF disc capacitor	R_4	2.2-ohm 5-watt resistor
		R_5	22K ½-watt resistor

Miscellaneous Parts

Pegboard (project 7)	5-prong tube socket
Crystal socket	2 sets of coils (see coil table)
Crystals 20-40-80-160 fundamentals	0-1 mA meter
3 12-volt lantern batteries	51-ohm test resistor
	Binding posts

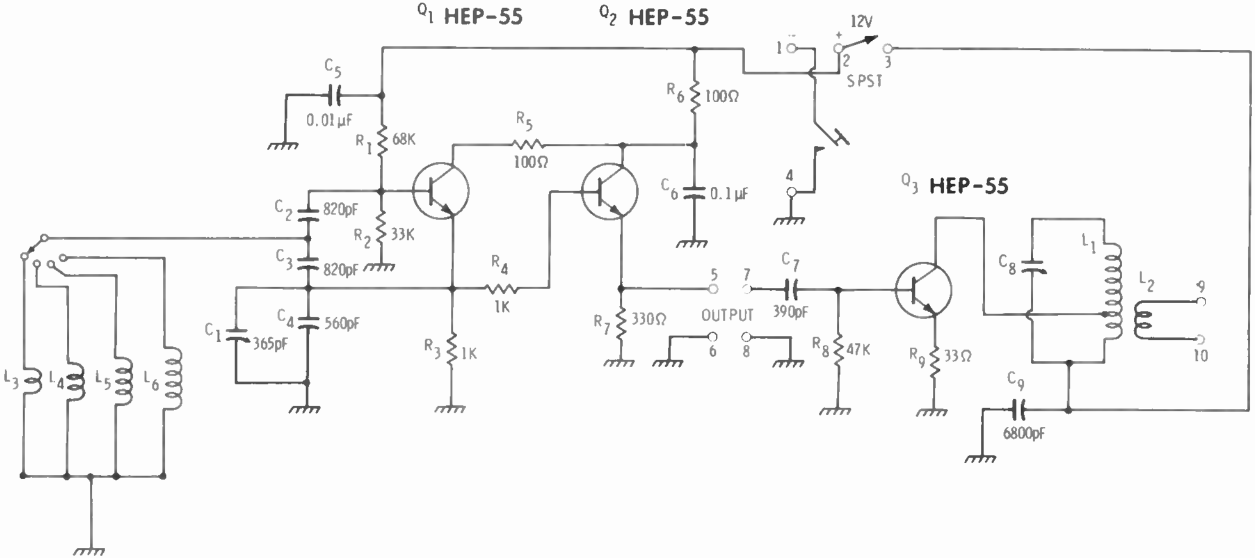


Fig. 2-9. VFO and amplifier.

Parts List for Fig. 2-9.

C_1	365-pF variable capacitor	L_5	2.38- to 3.96-μH slug-tuned inductor (J.W. Miller 21A 336 RB1)
C_2,C_3	820-pF silver-mica capacitors		
C_4	560-pF silver-mica capacitor		
C_5	0.01-μF disc capacitor	L_6	10.8- to 18-μH slug-tuned inductor (J.W. Miller 21A 155 RB1)
C_6	0.1-μF paper or disc capacitor		
C_7	390-pF disc capacitor		
C_8	100-pF variable capacitor	Q_1,Q_2,Q_3	HEP-55 transistors
C_9	6800-pF disc capacitor	R_1	68K ½-watt resistor
L_1,L_2	Coil set (Fig. 2-7)	R_2	33K ½-watt resistor
L_3	0.088- to 0.12-μH slug-tuned inductor (J.W. Miller 20A 107RB1)	R_3,R_4	1K ½-watt resistor
		R_5,R_6	100-ohm ½-watt resistor
		R_7	330-ohm ½-watt resistor
L_4	0.735- to 0.984-μH slug-tuned inductor (J.W. Miller 20A 827RB1)	R_8	47K ½-watt resistor
		R_9	33-ohm ½-watt resistor
		S	Spst switch

Miscellaneous Parts

Pegboard 8″ x 5″	5-prong tube socket
3 transistor sockets	Binding posts

by readjusting capacitors C_1 and C_4 slightly. Amplifier output is near 2 watts using 12 volts and approaches 3 watts with 18 volts. Typical drain current is 220 milliamperes. Therefore:

$$P_{DC} = V_{CC} \times I_C$$
$$= 18 \times 0.22 = 3.96 \text{ watts}$$

Efficiency is about 70% which is quite good.

Try operation on 40 meters. Similar results are obtained. Do the same for 20-meter operation. Efficiency begins to fall off and better results can be obtained by placing a short between posts 10 and 11. Connect the 2.2-ohm resistor across resistor R_3 in the emitter circuit of the crystal oscillator. A good output can now be available on 20 meters, usually above two watts using 18 volts on the amplifier. Somewhat less output is made available on 15 meters.

This little transmitter puts out a good signal and a nice clean one with proper tuning. Remember that a good absorption meter and/or receiver used as a tunable field-strength meter are of definite help in the tuning process. Also the tuner of Project 4 aids matching and rejects harmonics and other spurious signals.

PROJECT 9: VARIABLE-FREQUENCY OSCILLATOR AND AMPLIFIER

The variable-frequency oscillator (vfo) is a useful signal source in experimenting with solid-state and vacuum-tube circuits. One that operates on a number of bands using a simple switching arrangement is particularly helpful. The variable-frequency oscillator has advantages for QRP operation. One can move on top of a station calling CQ or terminating a QSO. It is also possible to move out from under QRM and high-powered stations.

It is important that a vfo be stable so that it will not drift or jump frequency. Two factors have a great influence on stability; these are electrical design and mechanical rigidity. The electrical stability requires a well-regulated power supply or a battery source and the use of one of the high-stability oscillator circuits such as Clapp, Vackar, Seiler, and others.

Mechanical stability depends on the selection of parts and a firm and rigid construction. A good quality variable capacitor is essential. Proper shielding and an insulated tuning shaft are important in reducing hand-capacity effects.

The vfo of Fig. 2-9 uses bipolar transistors. The first stage is a Seiler-type oscillator with separate inductors for 160, 80, and 40, and 20 meters. Coil data is given in the parts list. The output stage is connected as an emitter follower to present a light load to the oscillator and a low-impedance output. A battery provides a stable d-c power source.

With the component values shown the variable capacitor is able to tune over the entire 20, 40, 80 and 160 meter bands using appropriate coils. Proper band coverage is obtained by adjusting the slugs of the four inductors. If you wish to tune finely over only a segment of one of the bands an additional 20-pF variable capacitor can be connected in parallel with the 365-pF variable C_1. Output is removed at binding posts 5 and 6.

Amplifier

A low-power class-C amplifier using the same type bipolar transistor can be added. Vfo and amplifier together can be operated as a very low power (QRPP) transmitter. Output power is 50 to 100 milliwatts.

An additional advantage of the amplifier is that it permits the vfo to serve as an all-band frequency source by using the amplifier as a frequency multiplier. For example with the vfo operating on 40 meters the amplifier can be made to double to 20 meters or triple to 15 meters.

Output is weak but nevertheless you will have a source of signal that can drive succeeding amplifiers. With the vfo operating on 20 meters the amplifier can be made to operate as a doubler to obtain 10-meter output. Use the coils detailed in Fig. 1-14.

A somewhat greater power output can be obtained by operating the amplifier at 18 or 24 volts. Do not use more than 12 volts on the oscillator for stability reasons. A pegboard mounting of oscillator and amplifier is shown in Fig. 2-10.

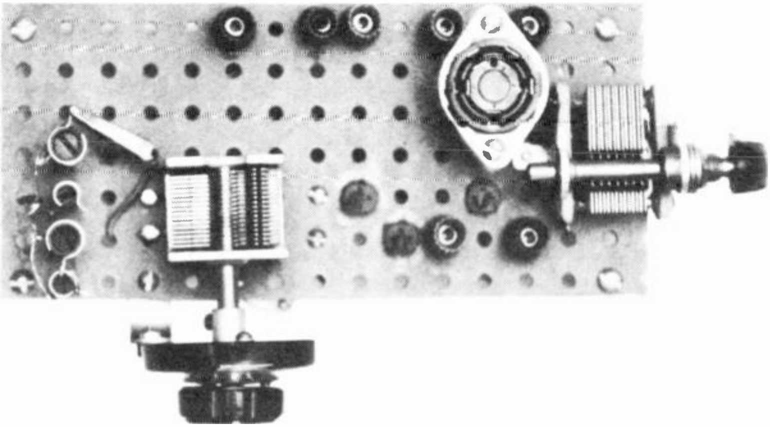


Fig. 2-10. Vfo and amplifier mounted on pegboard.

Operation

Set the oscillator bandswitch for 80-meter operation. Set capacitor C_1 to midposition. Connect your rf indicator between posts 5 and 6. Turn on the oscillator only. If the oscillator is operating, there will be an output reading.

Turn on your receiver and set its frequency to approximately 3.75 MHz. Adjust the slug in the 80-meter vfo coil until you hear the signal in the receiver. Vary capacitor C_1 and follow the signal up and down the band.

Set the vfo to some frequency around 3.75 MHz. Tune the receiver to zero beat on this frequency. Do not touch vfo or receiver. Note the high electrical stability of the oscillator.

Put your hand around the vfo coil and in the vicinity of the variable capacitor. Frequency changes. Tap the oscillator lightly. There is also a change in frequency.

The hand-capacity effect and mechanical instability can be corrected by mounting the vfo rigidly inside of a metal case. Such a construction is shown in Fig. 2-11. You may wish to do this after you have finished your experimentation with the vfo. The amplifier is not included in this arrangement because of the need for plug-in coils.

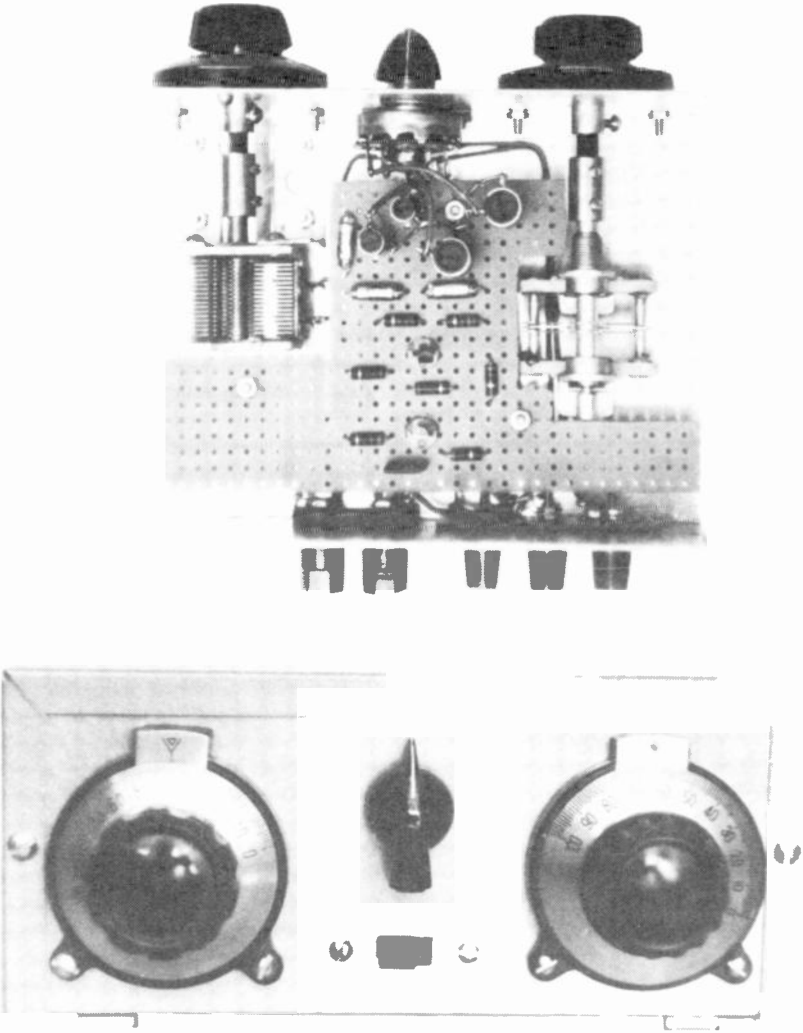


Fig. 2-11. Vfo mounted in metal case.

Place the amplifier in operation by connecting a jumper between binding posts 2 and 3. An alternative plan is to apply 18 to 24 volts positive to binding post 3 for higher powered operation. Plug in the 80-meter coil. Connect the rf indicator to the output of the amplifier. Connect the output of the oscillator to the input of the amplifier. Tune capacitor C_5 for maximum output. Try the operation of the oscillator on the other three bands. Go through the same procedure.

Restore operation on 80 meters. Turn on oscillator and amplifier. Tune capacitor C_5 for maximum output. Use the 68-ohm terminating resistor. A rather good output is obtained.

You can make a local 80-meter contact by first detaching the terminating resistor and connect an 80-meter dipole across the output. Use the rf indicator to peak the output signal with capacitor C_5.

The oscillator can be keyed by putting the power switch to its off position and connecting the key between binding posts 1 and 4. At a later time you may wish to use the vfo as a part of an a-m or ssb transmitter. In this case the push-to-talk microphone switch can be connected between binding posts 1 and 4.

Set the vfo for 40-meter operation and plug in the 40-meter amplifier coil. Adjust the oscillator to approximately 7.1 MHz using your receiver. Tune capacitor C_5 for maximum output. Now remove the 40-meter coil from the amplifier and substitute the 20-meter coil. Tune capacitor C_5 for maximum output. Use an absorption wavemeter or your receiver to verify the frequency doubling to 14.2 MHz.

In using a vfo-multiplier circuit of this type one must always be certain to verify the harmonic to which the amplifier resonant circuit is tuned. Substitute the 15-meter coil in the amplifier and tune for a maximum output. Again check frequency with absorption wavemeter or receiver. If you have the amplifier tuned to operate as a tripler; the output frequency should be 21.3 MHz (7.1×3). Be certain you are tuned to the third harmonic and not the second. It is very possible that the resonant circuit will cover both frequencies and the third harmonic is the peak that occurs at a low capacitance setting of capacitor. C_5

Set the oscillator on 20 meters. Use the 20-meter plug-in coil of the amplifier. Set the oscillator frequency to about 14.3 MHz.

Adjust capacitor C_8 for maximum output. Verify the output frequency with an absorption wavemeter and/or a receiver.

Now substitute the 10-meter coil in the amplifier. Tune capacitor C_8 for maximum output. If the output circuit is tuned to the

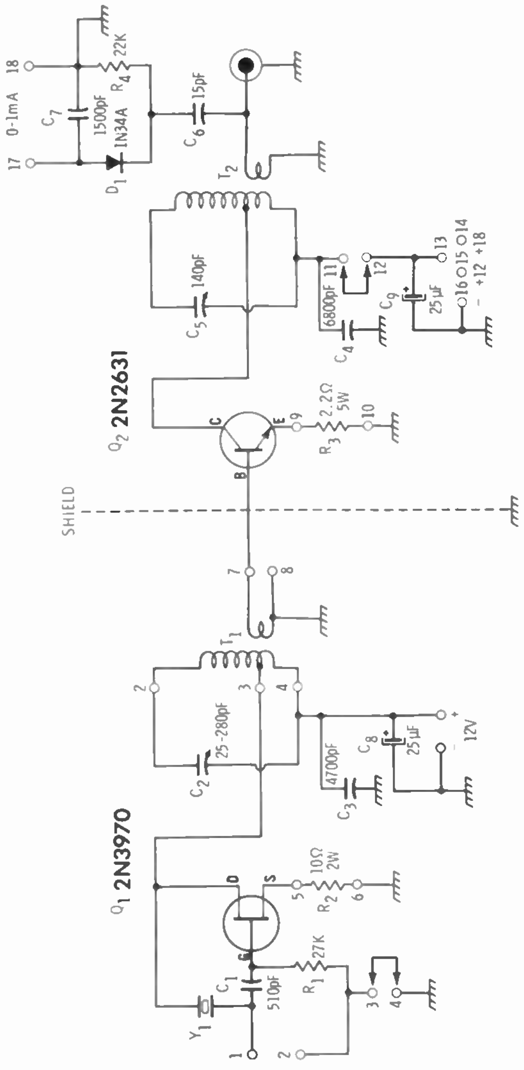


Fig. 2-12. Two-stage hybrid transmitter.

second harmonic of the oscillator frequency you will have an output frequency of 28.6 MHz. Again verify your results with an absorption wavemeter and/or a receiver tuned to the 10-meter band.

PROJECT 10: FET-BIPOLAR HYBRID TRANSMITTER

A combination of FET crystal oscillator and bipolar amplifier performs well as a 20-, 40-, 80-, and 160-meter cw transmitter. Rf power output is in the two- to four-watt range. The circuit of Fig. 2-12 can be assembled by revising that of Fig. 2-8. Transformer T_1 uses the close-coupled coils detailed in Fig. 1-14. Transformer T_2 will give you valuable experience constructing and working with toroid coils and transformers. Resistor R_3 in the emitter circuit of the amplifier is 2.2 ohms. This resistor is used for bands 40, 80, and 160 meters. On 20 meters connect a shunt between binding posts 9 and 10.

Toroid Coils

Toroid coils are popular in solid-state transmitters. Such a ring core has a closed magnetic loop which ensures good coupling between primary and secondary windings. A few turns of wire result in a coil of high inductance and, therefore, fewer turns are needed to obtain a given resonant frequency when using a given amount of tuned-circuit capacitance. A very small and compact resonant transformer is then possible, making the toroid coil especially at-

Parts List for Fig. 2-12.

C_1	510-pF silver-mica capacitor	Q_1	2N3970 FET (Siliconix)
C_2	25-to 280-pF trimmer capacitor (ARCO 464)	Q_2	2N2631 bipolar transistor
C_3	4700-pF disc capacitor	R_1	27K ½-watt resistor
C_4	6800-pF disc capacitor	R_2	10-ohm 2-watt resistor
C_5	140-pF variable capacitor	R_3	2.2-ohm 5-watt resistor
C_6	15-pF disc capacitor	R_4	22K ½-watt resistor
C_7	1500-pF disc capacitor	T_1	Coil set (Fig. 1-14)
C_8,C_9	25-μF 25-volt electrolytic capacitors	T_2	Toroid coil set (Fig. 2-13)
D_1	1N34A diode	Y_1	Crystals 20-40-80-160 fundamentals

Miscellaneous Parts

Pegboard (Project 8)	0-to 1-mA meter
Crystal socket	3 12-volt lantern batteries
2 Five-prong tube sockets	Binding posts

tractive for compact solid-state rf stages. Toroid coil data are given in Fig. 2-13.

Toroid

BAND	MFG. TYPE	PRI.	TAP	SEC.	WIRE SIZE
160	Permacor 13⁄16″ 57-1541	68 turns	20 turns	20 turns	No. 26 enam.
80	Permacor 13⁄16″ 57-1541	52 turns	15 turns	15 turns	No. 24 enam.
40	Amitron 0.68″ E	40 turns	12 turns	12 turns	No. 24 enam.
20	Micrometals T-50-2	30 turns	9 turns	9 turns	No. 24 enam.
15	Amitron 0.5″ SF	14 turns	4 turns	4 turns	No. 24 enam.
10	Amitron 0.5″ SF	9 turns	3 turns	3 turns	No. 24 enam.

Fig. 2-13. Toroid coil data.

The primary is tapped again to obtain a low-impedance connection point for the collector or drain of the device. The secondary is bifilar wound between turns of the lower segment of the primary and matches the low-impedance antenna or input of the succeeding stage. Toroid cores are available at low cost from a number of sources. Sources of cores, transistors, and other special components are listed in Appendix I.

For multiband operation, wire and mount the toroidal transformer to a five-prong CP plug, (Fig. 2-14). Thus, toroid and conventional core constructions can be used interchangeably. An alternative approach is to wire five binding posts into the circuit to match the leads of the various toroidal coils you construct.

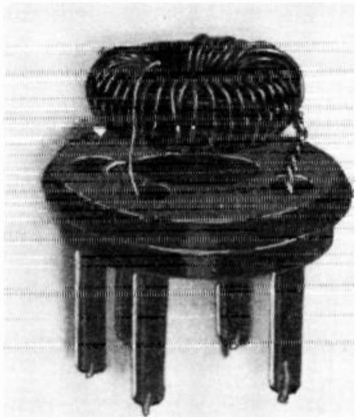


Fig. 2-14. Toroid construction and mount on five-prong plug.

Operation

Insert the 80-meter coils and crystal. Supply power to the crystal oscillator. Adjust capacitor C_2 for maximum reading on the 0-1 milliammeter with no power supplied to the amplifier. Now adjust capacitor C_5 for maximum. Again no power is supplied to the amplifier. This step places the two resonant circuits near proper frequency.

Supply 12 volts to the amplifier by connecting the jumper between binding posts 13 and 15. Tune capacitor C_5 for maximum output. Inasmuch as additional capacitance has been added to the circuit this means that the true tuning point will be at a lower capacitance setting. Retune capacitor C_2 for maximum output and retune capacitor C_5 slightly for peak output. Again check settings with an absorption wavemeter and/or a receiver tuned to the crystal frequency.

Increase the amplifier supply to 18 volts. Retune capacitors C_5 and C_2. Again check the quality of your signal. Use a line tuner when you go on the air.

Go through the tune up procedure on other bands. For 160-meter operation connect an additional 200-pF fixed capacitor across C_2. On 20 meters connect a jumper between binding posts 9 and 10.

VFO Drive of the Hybrid Transmitter

The hybrid transmitter can be used as a tunable-frequency transmitter by removing the crystal from the crystal socket and applying the vfo output between binding posts 1 and 2, (Fig. 2-15). The vfo covered in Project 9 supplies adequate drive to the transmitter. In

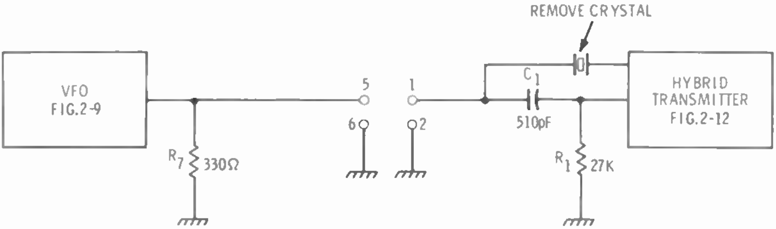


Fig. 2-15. Connecting the vfo to the hybrid transmitter.

this case binding posts 5 and 6, of the vfo, (Fig. 2-9), connect to binding posts 1 and 2 of the hybrid transmitter. The amplifier of Fig. 2-9 is deactivated because adequate signal can be obtained from the oscillator alone. In fact, the shielded vfo of Fig. 2-11 does an excellent job. It will supply adequate drive signal on 20, 40, 80, and 160 meters.

Before connecting the vfo to the transmitter tune up the transmitter on 80 meters using an 80-meter crystal. Tune in the signal on your receiver. Turn off the transmitter and remove the crystal.

Connect the output of the vfo to the input of the transmitter. Turn on the vfo and adjust its frequency until it can be heard in the receiver which has been set previously to the crystal frequency. Turn on the first stage of the transmitter (stage previously used as a crystal oscillator). There will be some slight reading on the 0-1 milliammeter connected between binding posts 17 and 18. Peak this reading with capacitor C_2 of the hybrid transmitter.

Operate the amplifier at low voltage by connecting the jumper between binding posts 13 and 15. Tune capacitor C_5 for maximum output. Jockey back and forth between C_2 and C_5 tuning for maximum output.

You can now operate the final amplifier with full voltage by connecting the jumper between binding posts 13 and 14. Retune capacitor C_5 and capacitor C_2 for maximum output.

Vary the frequency of the oscillator ±25 kHz above and below tune-up setting. You will find you can change the frequency of the transmitter a reasonable amount without having to retune the two amplifier stages.

Gain experience by tuning up the vfo and hybrid transmitter combination on the other three bands. The combination serves as a tunable-frequency cw QRP transmitter for four bands.

PRINCIPLES OF A-M MODULATION (BIPOLAR)

A bipolar transistor just like the field-effect transistor of Chapter 1, Project 6, can be amplitude-modulated by varying the supply voltage to one of its elements—base, emitter or collector. A basic collector-modulation system is shown in Fig. 2-16. It is comparable to

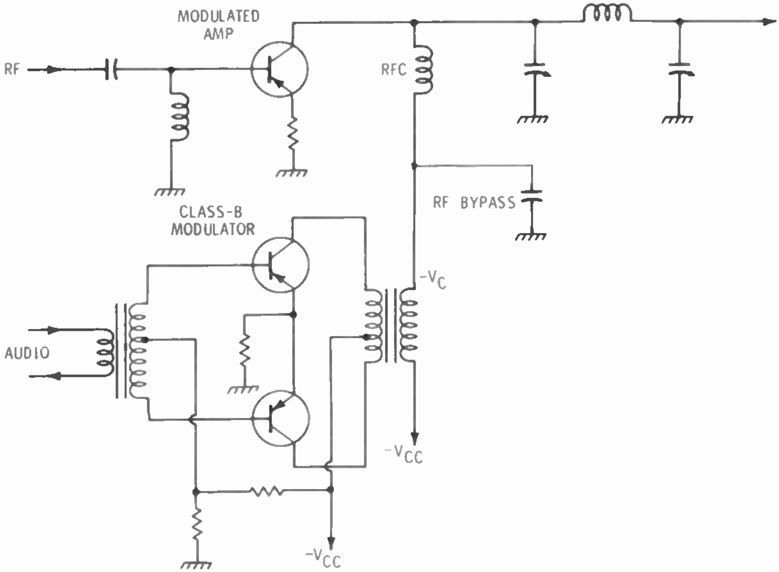


Fig. 2-16. Collector modulation system.

the drain modulation of an FET or plate modulation of a vacuum tube. The dc collector voltage is made to vary with the modulating signal by inserting the secondary of the modulation transformer in the supply voltage path. The collector voltage of the amplifier is increased or decreased by the audio (modulating) signal.

The unmodulated radio-frequency carrier is fed to the base circuit. The mixing of the two signals results in the formation of a modulated rf signal across the collector tank circuit.

In the rf stage the tank-circuit voltage varies linearly with the collector supply voltage, thus doubling of the supply voltage doubles tank rf voltage over a suitable operating range. If the modulated amplifier is operated from a 12-volt supply, the instantaneous collector supply voltage can be made to vary above and below 12 volts.

If the modulating wave is a pure sine wave this variation can be symmetrical. In bipolar transistor stages it is quite difficult to obtain absolutely symmetrical and full modulation. However, satisfactory and reasonably good amplitude modulation is obtainable with simple circuits.

The rate of variation of the supply voltage depends on the frequency of the modulating wave; the magnitude of the change depends on the amplitude of that wave. This change in the collector supply voltage becomes a replica of the modulating wave. If the variations on each side of the zero axis of the modulating wave are identical, the average dc collector voltage remains fixed at 12 volts. Likewise in a perfect bipolar modulation system the dc collector current would also remain constant. This ideal operating characteristic is not always feasible in simple bipolar modulation systems.

The change in the collector voltage with modulation results in a similar change in the peak collector current. In this manner a modulation envelope is formed as described under Project 6. Instead of a peak drain current change, it is a peak collector current change which develops a corresponding rf envelope variation across the output tank circuit.

To develop the desired change in the envelope with full-modulation capability, the collector voltage must be made to swing between twice the collector supply voltage and zero. In practical bipolar modulation systems maximum modulation is usually less than 100 percent.

The impedance seen by the modulator output equals the dc resistance of the modulated amplifier input. This dc input resistance is quite low because of the low collector voltage and high collector current:

$$R_{in} = \frac{V_{cc}}{I_c}$$

PROJECT 11: BIPOLAR MODULATOR

In this experiment a modulator is added to the hybrid two-stage transmitter of Fig. 2-12. A variety of low-cost audio modules are available from electronic supply houses. They are used mainly to drive small speakers and have a low-impedance output. Some of them are tapped with various output impedances made available. The audio module described previously in Project 6 has both 8-ohm and 16-ohm outputs.

The input resistance for a bipolar modulated amplifier, because of low collector voltage and high collector current, is quite low and the addition of a second transistor audio power transformer can provide a good match. For example, a bipolar stage operating at 12 volts and drawing 400 mA has an input resistance of

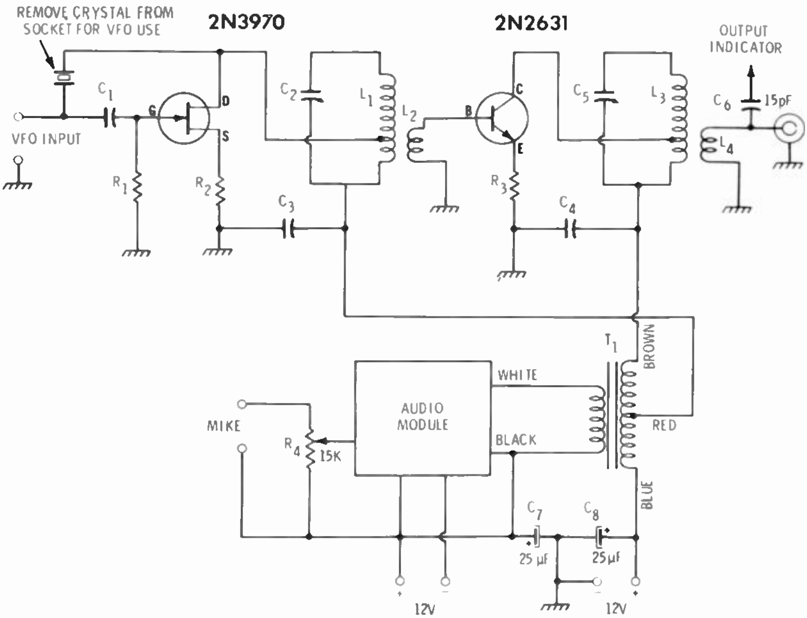
$$R_{in} = \frac{12}{0.4} = 30 \text{ ohms.}$$

An audio module with output of 3 watts and 16 ohms resistance will provide adequate modulator power. An additional transformer can be used to match the 16-ohm output of the module to the input of the modulated amplifier. The circuit arrangement is shown in Fig. 2-17. The 16-ohm side of transformer T_1 is connected to the 16-ohm output of the audio module.

Both the final amplifier and the buffer amplifier (former crystal oscillator stage of the hybrid transmitter) are modulated. The full 48-ohm secondary is connected in the collector supply voltage path to the final amplifier. To obtain good modulation at a reasonably high percentage, the drain of the buffer is also modulated. Note that its drain supply-voltage line connects to the center tap of the secondary of transformer T_1. This technique is used widely in solid-state transmitters to obtain a better modulation characteristic than can be obtained when only the final stage is modulated.

A 15K potentiometer (R_1) is used to regulate the level of the microphone signal at the input of the audio module. Most of the audio modules employ pnp transistors and it is important that the supply voltage is of proper polarity. For the a-m transmitter it is recommended that separate 12-volt batteries or supplies are used for transmitter and module. Make certain that the module is connected to its battery with proper polarity. Recognize that plus is common for the audio module while negative is common for the hybrid transmitter.

The ground of the module is connected to the common of the system with capacitor C_7. (This can be capacitor C_8 of the drain cir-



Hybrid transmitter	From Project 10	T_1	Audio transformer (Knight 6-T-51) Secondary 48 ohms CT, Primary 16 ohms, 10-watt rating
Audio module	Philmore AA900 or Lafayette 99-91324		
Mike	High-level communications microphone	R_4	15K potentiometer

Fig. 2-17. Arrangement for modulating hybrid transmitter.

cuit of the hybrid transmitter, Fig. 2-12). It is disconnected because it must not be in the modulated supply-voltage line to the drain circuit. It can then be used in the position shown in Fig. 2-17 to tie the audio module and hybrid transmitter commons together. Other parts are labeled as in Fig. 2-12.

The source of carrier signal is the two-stage vfo of Project 9. Its output is connected to the vfo input of the hybrid transmitter.

Carrier power output is two to three watts on the various bands and this carrier can be modulated well with the arrangement of Fig. 2-17. It is a very stable arrangement. Crystal control can be employed too but at high modulation level there is some shift in the crystal oscillator frequency because of changing drain voltage. This is not too serious if the modulation level is kept down.

Operation

Insert appropriate coils for hybrid transmitter operation on 80 meters. Turn on the vfo and adjust its frequency to some spot in the 80-meter phone band. Connect the rf indicator to the transmitter output using a 51-ohm or 68-ohm 5-watt dummy load resistor.

Apply power to the transmitter and tune up on 80 meters. You can take a look at the carrier with a service-type oscilloscope connected to the output or with an oscilloscope such as the Heathkit SB-610 inserted between the transmitter output and the dummy load.

Apply power to the audio module after connecting the microphone to its input. Turn up the audio gain control and speak into the microphone. You will be able to observe the modulation envelope on the oscilloscope screen. Adjust for an appropriate depth of modulation.

Note as you speak into the microphone there is only a slight upward kick of the 0-1 milliammeter of the rf indicator. This indicates quite good modulation. Most simple solid-state transmitters show a downward kick (negative carrier shift) with modulation. If a buffer is used, the modulation of both buffer and amplifier is of help in obtaining reasonably good modulation.

Check the modulation quality on a receiver placed some distance away. Avoid excessive distortion which may occur with the audio gain control turned up too high. The arrangement is such that it is not likely that the transmitter will overmodulate seriously. An on-the-air check with a nearby ham will help you establish a proper modulation level where an oscilloscope is not available.

Tune up the transmitter on the 40-meter phone band. Then try 20 and 160 meters. Similar results are obtained on these bands.

PROJECT 12: AMPLIFIER FOR TEN-TEC TX-1 TRANSMITTER

The *Ten-Tec* TX-1 is a popular QRPP transmitter that can be operated on 15, 20, 40, and 80. Output power is 1 watt plus depending on frequency of operation. The unit supplies more than enough drive to a 2N2631 class-C amplifier. Output can be boosted to three or four watts, using 12-volts on the 2N2631. The amplifier can also be used with other solid-state units that supply a half watt or so of output.

The *Ten-Tec* transmitter module is a two-stage affair available at low cost and wired for immediate operation (Fig. 2-18). It con-

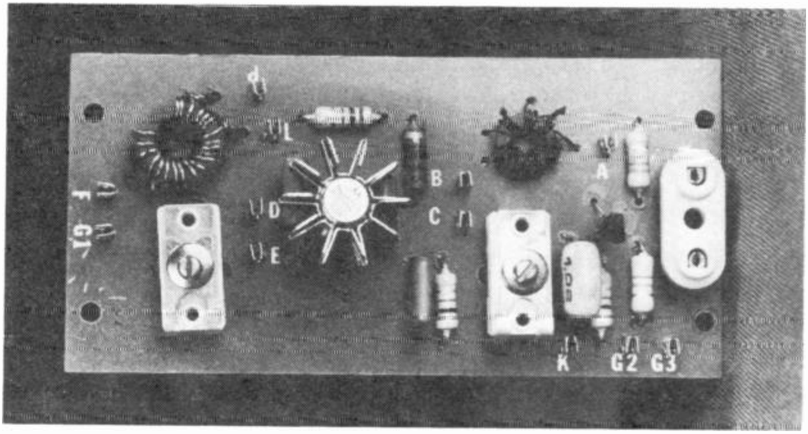


Fig. 2-18. TEN-TEC transmitter module.

sists of a crystal oscillator and power amplifier. Appropriate coil taps permit four-band operation (Fig. 2-19). Connecting terminals E to D and C to B provides coverage of 40 and 80 meters. Coverage on 15 and 20 meters requires that E be connected to L and C to A.

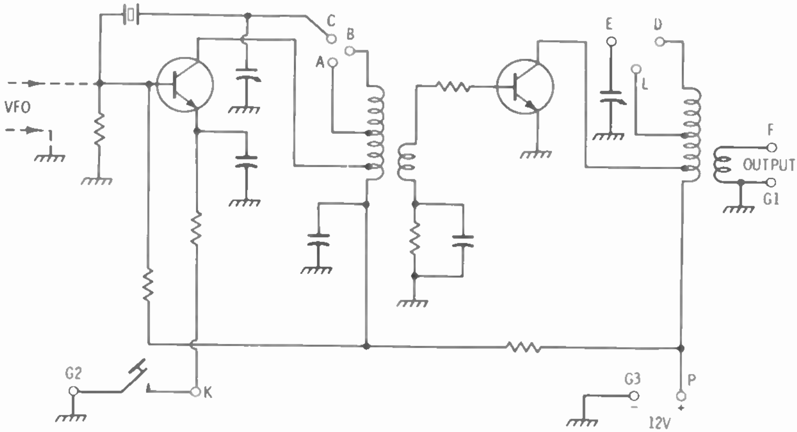
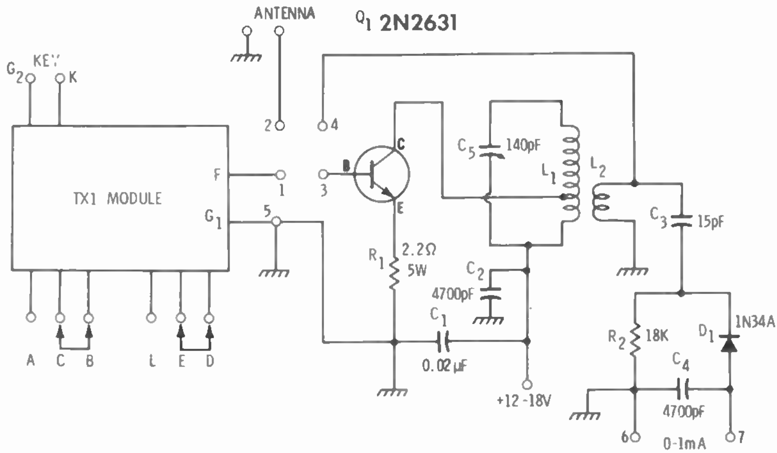


Fig. 2-19. TEN-TEC TX1 circuit.

Trimmer-type capacitors are connected across the tuned circuits. When using the 12-volt battery the dc current is approximately 250 milliamperes when driving a proper load.

The transmitter is keyed in the emitter circuit of the crystal oscillator. The closing of the key between terminals G_2 and K closes the emitter circuit to common. The battery is connected between terminals P and G_3. For ease in operation the module can be fastened to a pegboard and binding posts connected to the various terminals. This permits ease in band changing.

The amplifier can be mounted on the same pegboard (Fig 2-20). It is straightforward and uses the coil data given in Fig. 1-14. Binding posts can be so arranged that the output of the module can either



C_1	0.02-µF disc capacitor	L_1,L_2	Coil set (Fig. 1-14)
C_2,C_4	4700-pF disc capacitor	Q_1	2N2631 transistor
C_3	15-pF disc capacitor	R_1	2.2-ohm 5-watt resistor
C_5	140-pF variable capacitor	R_2	18K ½-watt resistor
D_1	1N34A diode		

Miscellaneous Parts

Pegboard 5″ x 12″	0-to 1-mA meter
TEN-TEC TX1 module	6-volt lantern battery
Transistor socket	12-volt lantern battery
5-prong tube socket	

Fig. 2-20 Add-on amplifier for TEN-TEC TX1.

be supplied to the amplifier or to the antenna. When using the module alone binding post 1 is connected to binding post 2. For amplifier operation binding post 1 connects to binding post 3 and binding post 4 is connected to binding post 2.

Operation

Set up the *Ten-Tec* module for 80-meter operation. Plug in an appropriate crystal. Connect the rf indicator across the output of the module. Use a 68-ohm terminating resistor.

Supply power to the module only. Adjust the oscillator and amplifier trimmer capacitors for maximum output.

Prepare the 2N2631 amplifier for operation on 80 meters. The amplifier has its own output indicator. You need only connect the 0-1 milliammeter across binding posts 6 and 7. Plug in the 80-meter coil.

Turn on the *Ten-Tec* module but do not supply power to the amplifier as yet. Adjust crystal oscillator and module amplifier for maximum reading on the 0-1 milliammeter. Now tune capacitor C_5 for maximum. What you are doing is bringing the amplifier resonant circuit near to its final resonant setting.

Apply power to the amplifier. Adjust capacitor C_5 for maximum. This will require a somewhat lower capacitance than before. Likewise, the tuning capacitor of the TX-1 module amplifier will require slight readjustment.

Use the absorption wavemeter to make certain that the output is on the desired frequency, and you have not adjusted the amplifier to some spurious setting.

Repeat the procedure for each of the four bands. In setting up for 15- and 20-meter operation it is important again that you use the absorption wavemeter to make certain the amplifier resonant circuit is adjusted to the correct frequency and not some harmonic.

CHAPTER 3

Vacuum-Tube CW/A-M Transmitter Circuits

In this chapter you are introduced to the operating characteristics of vacuum tubes in transmitter circuits. The vacuum tube has been the mainstay of transmitter circuits these many years. Only recently have the bipolar transistor and field-effect transistor taken over some of the responsibilities. Vacuum tubes are still used widely in amateur and other low- and medium-power transmitters. High-powered transmitter amplifiers are exclusively vacuum-tube types. Many commercial and some amateur units employ transistors up to the final vacuum-tube power amplifiers.

The projects of this chapter permit you to build up vacuum-tube oscillators, frequency multipliers, and class-C amplifiers. You will also experiment with using a solid-state vfo to drive a vacuum-tube multiplier and amplifier combination. A FET speech amplifier and tube modulator will also be constructed.

THE VACUUM TUBE

There are various types of vacuum tubes according to number of elements and element arrangement. Three common types used in amateur transmitters are triode, pentode, and beam power. A *triode* is a three-element tube consisting of a heater-cathode, control grid,

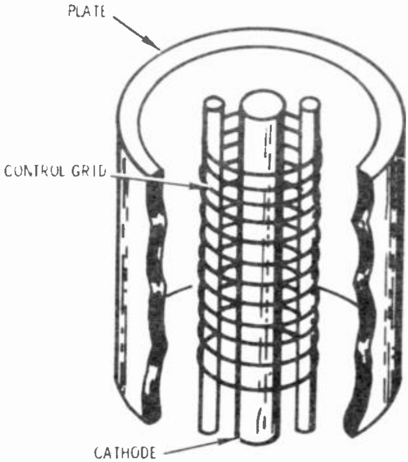


Fig. 3-1. Basic construction of a triode tube.

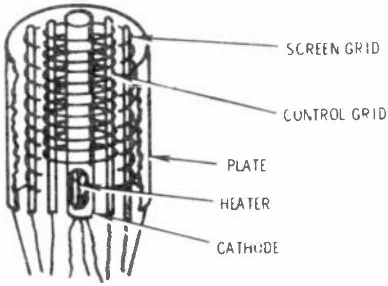
and plate, all mounted within an evacuated glass, metal, or ceramic case (Fig. 3-1). At the center of the structure is the *cathode,* which acts as a source of electrons when heated. The cathode electron-emitting surface can be heated directly or indirectly. In most vacuum tubes there is a filament winding to which is applied an alternating current. Heat of the filament raises the temperature of a separate cathode surface. In a directly heated filament system used in high-powered vacuum tubes and vacuum-tube rectifiers the filament wire itself emits the electrons.

The indirectly heated cathode of the usual vacuum tube is surrounded by a *control grid* made of fine wire mesh. Its grid-like construction is such that the electrons can readily pass through the wires and move on toward the more positive plate. The grid function is to control the number of electron charges moving from cathode to plate.

The *plate* is a solid electrode that surrounds the grid. A positive potential (with respect to the cathode) is applied to the plate, which then attracts the electrons emitted from the cathode. The control grid is able to control the number of these emitted electrons that reach the plate.

A *tetrode* is similar to a triode except for an additional electrode between the grid and the plate (Fig. 3-2). This additional electrode, called a *screen grid,* makes the plate current much more independent of plate voltage. In effect, it acts as a shield between the plate and

Fig. 3-2. Basic construction of a tetrode tube.



grid circuits, minimizing feedback between the input and output circuits by way of the internal tube capacitance. The screen grid, also being a fine wire mesh, permits the easy passage of electrons enroute to the plate. Because it operates at a positive potential the screen grid has a substantial influence on the amount of plate-current flow.

In a tetrode, the plate voltage may not be driven too far in the negative direction. If the plate voltage becomes lower than the screen voltage during a portion of the input cycle, electrons strike the plate and dislodge secondary electrons which are attracted to the screen (the screen grid now has a higher potential than the plate). This limits the linear range over which the plate voltage can be made to change by an applied signal.

The more common *pentode* is a five-element tube. It is similar to the tetrode in construction except that it has an additional electrode *(suppressor grid)* between the screen grid and the plate, Fig. 3-3. The primary function of the suppressor electrode is to counteract secondary emission at the plate.

The inserted suppressor grid, held at or near ground potential, retards the emission of secondary electrons, causing them to fall back on the plate. This retarding action can be anticipated because the suppressor grid is essentially negative with regard to the plate. As a result the positive plate voltage can swing to a value that is substantially lower than the actual screen-grid potential. Secondary electrons are blocked in their attempted movement between plate and screen grid because of the intervening near-ground potential of the suppressor.

Pentodes and tetrodes have a high plate resistance and can be made with a very high G_m (mutual conductance). Thus, a substantial

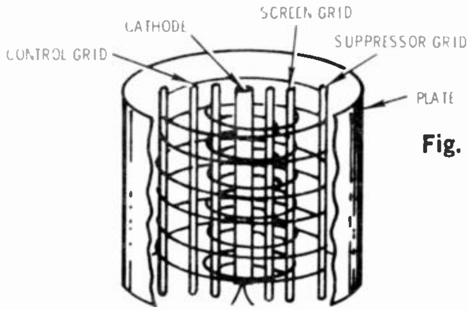


Fig. 3-3. Basic construction of a pentode power tube.

change can occur in the plate current and plate voltage as caused by only a small change in the applied grid signal, a grid signal much smaller, incidentally, than that required to develop a comparable output with a triode.

The *beam-power tube* (Fig 3-4) is a special form of tetrode-pentode with characteristics more like those of a pentode. It provides

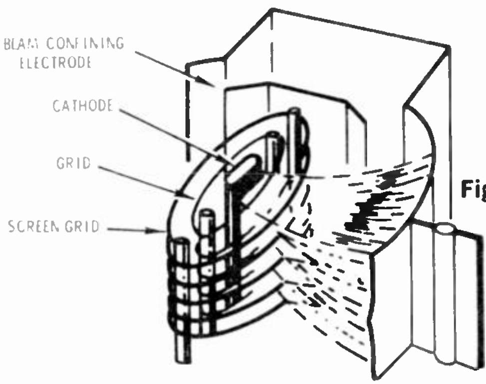


Fig. 3-4. Basic construction of a beam-power tube.

the ultimate in power sensitivity and absolute value of output power. Beam-forming plates between the screen grid and the plate guide electrons in an efficient beam-like manner between the cathode and the plate. Because these plates are at cathode potential their focusing action sets up a simulated suppressor grid between the screen grid and the plate. Hence, even at low plate voltage there is no motion of secondary electrons from the plate to the screen grid. Such a tube has both high plate efficiency and power sensitivity. It is used widely in audio power stages and as radio-frequency amplifiers and frequency multipliers in transmitters.

Basic Amplifier Operation

In a basic vacuum-tube amplifier, as shown in Fig. 3-5, the dc voltage (E_b) supplied to the plate attracts the electrons emitted from

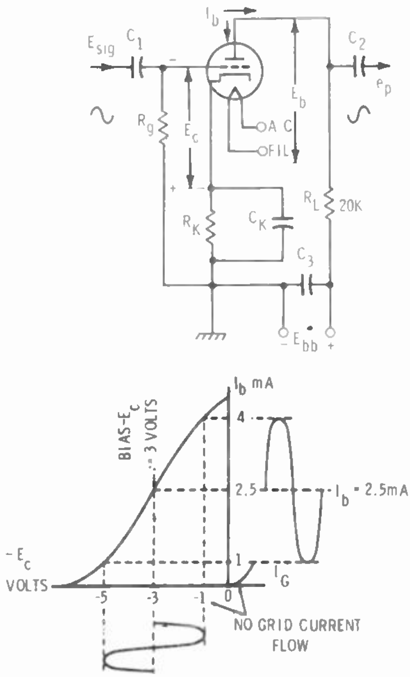


Fig. 3-5. Typical grounded-cathode stage.

the cathode. The number of electrons reaching the plate and flowing through the plate-load resistor (R_L) is determined by the negative voltage supplied to the control grid. This negative potential forms a space-charge of electrons between cathode and grid. In this circuit the plate current (I_b) in flowing through the cathode resistor (R_K) develops a potential of 3 volts dc across the cathode resistor-capacitor combination. The direction of the plate current through the cathode-resistor is such that the grid is made negative with respect to the cathode by 3 volts.

The so-called transfer curve of Fig. 3-5B shows that the flow of the plate current is a function of the negative grid voltage. Notice that over the most linear portion of the curve (between -1 and -5 grid volts), the plate current changes linearly with respect to the grid voltage. The negative dc grid bias is usually at the center of the linear portion of the transfer characteristic (-3 volts in the example). With this amount of negative grid bias supplied to the vacuum-tube stage, the dc component of plate current I_b would be 2.5 mA. This is referred to as the operating-point plate current.

To understand the operation of the triode as an amplifier, assume that an ac sine wave (E_{sig}) of 2 volts peak is being supplied to the control grid. As the sine wave rises on its positive crest, the negative grid voltage decreases and the plate current increases. At the crest of the positive alternation the instantaneous grid voltage is -1 volt and the plate current is 4 mA. The less negative the grid is driven, the weaker the space charge is and the more freely the electrons flow between cathode and plate.

On the negative alternation of the grid-voltage signal, the grid swings in the negative direction. Consequently the plate current decreases because the negative grid has a greater holding power on the space charge electrons. At the crest of the negative alternation, the instantaneous grid voltage is -5 volts and the plate current is only 1 mA.

When the plate current increases with a rise in grid voltage, the plate voltage drops because of the greater voltage drop across the load resistor. Conversely, when plate current falls during a negative swing of the grid, the plate voltage rises. This is to say, that input and output voltages are 180° out-of-phase or opposite in polarity.

A very important fact to recognize is that the plate-current change follows the grid-voltage change. As the grid voltage swings between peaks, it changes between -1 and -5 volts. Instantaneously the plate current also changes between the limit of 4 mA. and 1 mA. Thus the peak ac plate current swing is 1.5 milliamperes, which flows in the 20,000-ohm plate load resistor (R_L).

The plate voltage change can be determined by multiplying the plate current by the ohmic value of the plate-load resistor or:

$$\Delta E_b = \Delta I_b R_L = 0.0015 \times 20{,}000 = 30 \text{ volts}$$

where,

ΔE_b equals change in plate voltage,

ΔI_b equals change in plate current,
R_L equals value of load resistance.

The plate-voltage change is 30 volts peak. Inasmuch as the initial grid voltage change was 2 volts peak, the amplifier has a gain of 15 (30/2).

Classes of Operation

In a class-A amplifier, the output plate current is a replica of the input voltage as demonstrated in Fig. 3-5. Plate current flows during the entire cycle of the input voltage wave (Fig. 3-6). A class-B amplifier is biased at cutoff. Thus, with an applied sine wave the tube conducts only during the positive alternation of the input wave. This is to say that plate current flows for only 180° of the cycle of the input wave.

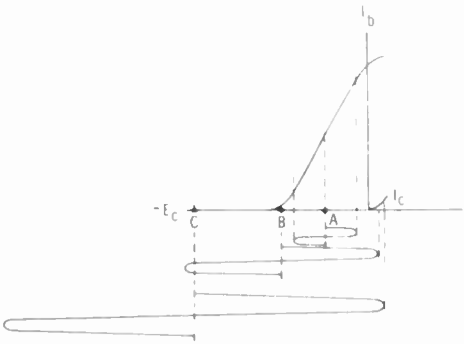


Fig. 3-6. Class-A, -B, and -C bias.

Class-AB bias refers to some level of bias between strictly class A and strictly class B. In this class the plate current flows for something more than 180° but less than 360° of the input cycle, and flows for the entire positive alternation of the input wave and for a portion of the negative alternation.

In class-C amplifier operation the tube is biased beyond cutoff. As a result the plate current flows for only a portion of the positive alternation of the input wave. In a typical class-C amplifier the plate current may flow for only 90° to perhaps 150° of the input cycle.

CLASS-C AMPLIFIER OPERATION

A typical class-C amplifier is shown in Fig. 3-7. It includes input and output resonant circuits, which in the case of straight-through or fundamental frequency rf amplifier are tuned to the same fre-

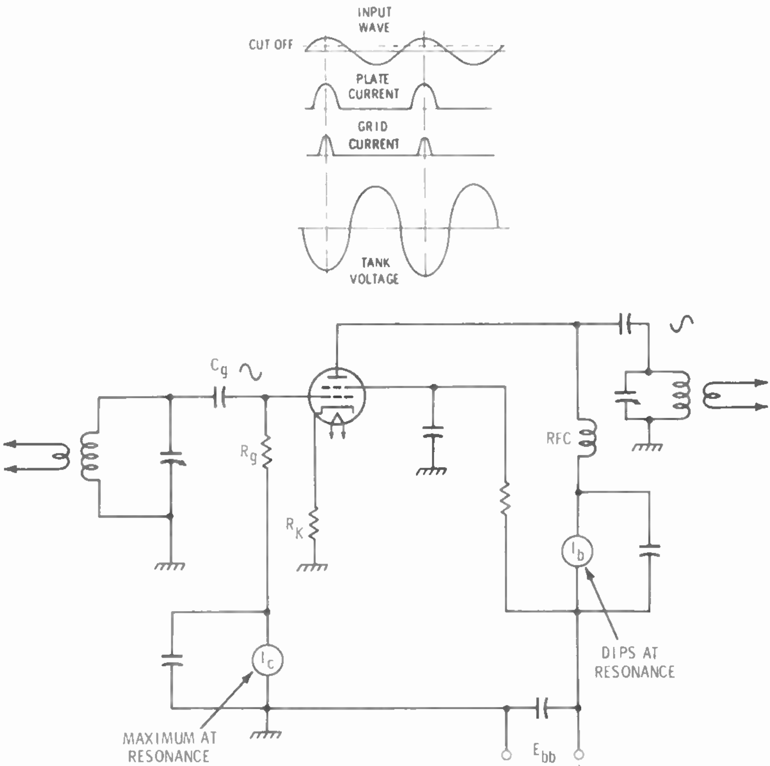


Fig. 3-7. Basic class-C amplifier.

quency. The beam-power tube is the most common rf power amplifier, although a number of pentodes and some triodes are used in this function. In many instances, with appropriate circuit, tube, and shielding, there is no need for neutralizing a beam-power or pentode tube.

In a class-C stage the tube draws plate current only during the most positive portion of the positive alternation of the input wave. Thus the plate current is drawn in bursts of high peak value. Such a burst supplies a lot of electrical energy into the plate resonant tank circuit and the capacitor is charged to a high negative value (sharp drop in plate voltage). However, a resonant circuit has smoothing and energy-storing abilities. When the burst ceases, the charge on the capacitor begins to fall off and energy is transferred to the coil. After the charges reach zero the collapsing field of the coil releases its stored energy back into the tank circuit. An opposite flow of current results and charges the capacitor in the opposite polarity even though the plate current has stopped flowing.

The magnetic field about the coil is now of opposite direction; when the capacitor loses its peak charge, the magnetic field again collapses and a new cycle of original polarity begins.

Soon afterwards, a new burst of plate current is introduced because of the positive swing of the grid waveform and the plate tank-circuit capacitor is again charged to a maximum negative value, initiating a new cycle of operation. Although the rf amplifier operates class-C and plate current does not flow for the full cycle of the input wave, a good sine wave is developed across the plate tank circuit with the proper choice of LC constants and load. This is the so-called "flywheel" effect in class-C rf amplifiers.

In most rf power amplifiers, the class-C bias is developed by the grid current in the grid circuit. On the postive peak of the input wave there is a substantial amount of grid current which develops a negative charge on grid capacitor C. The current is such that the charge placed on the capacitor biases the tube beyond cutoff. If the time constant R_gC_g is high enough, a reasonably constant bias is attained during the entire cycle of the input wave. This is equivalent to a dc bias voltage of a value that is several times beyond cutoff. Of course, the capacitor is restored to full charge by the grid current at the crest of each positive input wave.

The grid current is often measured with a dc meter. Such a meter is useful in tuning a class-C stage. There is maximum dc grid current when the applied input wave is peaked at its maximum amplitude. A dc meter is often used in the plate circuit (or in the cathode circuit) to measure the dc plate current. This reading is also useful in tuning a class-C stage.

When a plate resonant circuit is tuned to resonance it displays a maximum impedance and the dc plate current dips to a minimum. As energy is withdrawn from this tank circuit and delivered to the load, more power must be supplied to the tank circuit, and there will be a rise in the dc plate current reading.

Multiplier Operation of Class-C Stages

A typical harmonic generator or frequency multiplier is shown in Fig. 3-8. Triode or multigrid tubes can be used as multipliers. The

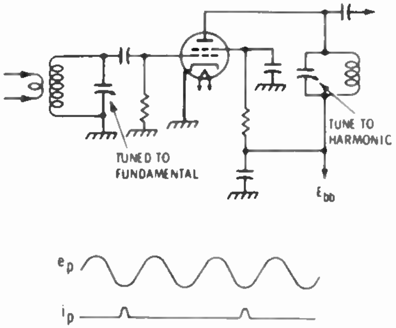


Fig. 3-8. Frequency multiplier.

beam-power tube is very common because of its low driving power and capability of generating a strong harmonic output. No neutralization is necessary because the output plate tank circuit is tuned to a frequency different from the input signal. It is tuned to a second, third, or higher-order harmonic of the input frequency, depending on the desired output frequency.

The frequency multiplier is operated class-C. In fact, to obtain a strong harmonic output it is often operated even further beyond cutoff than a normal straight-through class-C amplifier. To some extent the desired harmonic, second, third, fourth, etc., determines the most favorable operating angle for the stage. (This refers to the number of degrees or portion of the input wave during which there is plate current.) In general, the sharper the plate current burst the stronger is its harmonic content. However, efficiency is also a factor in selecting the proper beyond-cutoff bias.

If the harmonic generator is to operate as a doubler the plate tank circuit is tuned to twice the frequency of the grid resonant circuit. A burst of plate current is drawn through the plate tank circuit each time the grid waveform rises to its positive crest. However, the amount of energy drawn into the plate tank circuit is great enough that the oscillations within the plate tank circuit can go through two cycles before energy must be added. This means that a multiplier operates at a lower efficiency than a straight-through amplifier. Furthermore, the higher the desired harmonic is, the lower is the operating efficiency. Thus the output capability drops off rather quickly as the order of the desired harmonic is increased.

PROJECT 13: VACUUM-TUBE CRYSTAL OSCILLATOR

The vacuum tube, like bipolar and FET transistors, can be connected in a number of oscillator configurations. Three popular types are shown in Fig. 3-9. The Pierce crystal oscillator requires no reso-

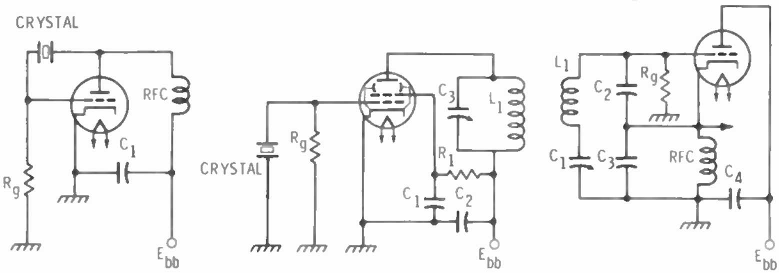


Fig. 3-9. Vacuum-tube oscillator circuits.

nant circuit. It can be made to operate over a wide range of frequencies simply by plugging in an appropriate crystal. More output can be obtained from the Miller configuration of Fig. 3-9B. The beam-power tube type provides better isolation from the output resonant circuit and the crystal and permits a higher output. Vacuum tubes for many years have been used in variable frequency oscillator circuits (vfo) using many configurations. The basic Clapp variable circuit with its circuit loading capacitances is shown in example C.

Over the years there have been a number of popular crystal oscillator circuits that provide fundamental crystal frequency output as well as harmonic outputs. One such circuit is the popular tri-tet circuit of Fig. 3-10. The oscillator itself is a fundamental Pierce type using the screen grid of the vacuum tube as the anode of a triode crystal oscillator. A cathode LC combination is employed which enhances the harmonic output of the crystal circuit. The signal generated by the Pierce section is electron-coupled within the tube to the plate circuit. The plate resonant circuit is tuned to the fundamental or the desired crystal harmonic.

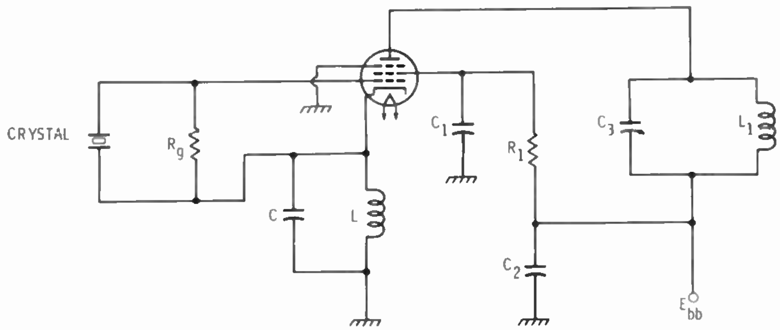


Fig. 3-10. Tri-tet crystal oscillator.

This basic oscillator operates as a fundamental frequency generator using appropriate crystals. In most applications it operates as a fundamental-frequency oscillator on 40, 80, and 160 meters. In its harmonic generating application it employs a 40-meter crystal to obtain output on 20, 15, or 10 meters (second, third, and fourth harmonics of the crystal).

Construction

In this project you will construct an experiment with a tri-tet crystal oscillator. An appropriate power supply is to be mounted on the pegboard. In addition you will prepare for projects number 14 and 15 by also building a class-C amplifier and leaving enough space for the addition of an FET speech amplifier and vacuum-tube modulator.

The circuit diagram is given in Fig. 3-11; a photograph of the complete assembly (all three projects combined) is shown in Fig. 3-12. The crystal oscillator is mounted at the bottom left with the amplifier to its right. Power supply is located at the top left with the speech amplifier and a-m modulator to its right. Included is a parts list for all three projects.

The power supply is simple and uses a silicon bridge rectifier. These rectifiers can be purchased with all four diodes mounted in a small module. The filter is a capacitive-input LC combination. Two switches are included, one in the primary of the transformer and the other in the ground return. The transmitter can be keyed at this point by turning switch S_2 off and connecting the key between binding posts 15 and 16. The switch part of a push-to-talk microphone can also be connected between these two binding posts when a-m operation is desired.

The crystal oscillator employs a 6CL6 pentode. The input circuit includes binding posts 1 and 2 for driving the stage from a variable-frequency oscillator when desired. In this case any crystal must be removed from the crystal socket. Likewise a jumper is connected between binding posts 3 and 4 to short out the cathode LC combination when the stage is vfo-driven.

Slug-tuned coils are used in the plate resonant circuit. The largest coil (L_2) is used for 160 meters. Inductor L_3 is for 80 meters. Inductor L_4 is fundamentally for 40 meters operation but will also tune on 20 meters with proper adjustment of the slug. Inductor L_5 is used for the three high-frequency bands 10, 15, and 20 meters. When you are building the crystal oscillator circuit also wire in the grid-cathode and filament circuits of the class-C amplifier. It is then possible to use the grid circuit as an output indicator for testing the crystal oscillator. Capacitor C_{11} is a small variable trimmer that permits the transfer of a proper drive signal to the grid of a beam-power amplifier. At the same time it provides a reasonable amount of isolation between the amplifier and oscillator. On 10, 15, and 20 meters its capacitance can be reduced if there is any instability and tendency for the amplifier to self-oscillate. Note in the construction that the power supply and crystal oscillator occupy about one-half of the pegboard area.

CAUTION

In Chapters 1 and 2 you worked with low-voltage semiconductor devices. It was very convenient to make changes and adjustments with the operating voltages turned on. The possibility of shock was nonexistent. Do not forget that in vacuum-tube circuits you are dealing with shocking and lethal voltages. Be on guard continuously and turn off the power supply completely when making changes. When making adjustments with the power turned on, work with one hand. Keep the other hand and the rest of your body free of the unit and ground.

Operation

Before applying power connect a small NE-2 neon bulb between the grid side of capacitor C_{11} and ground. Also connect a 0-1 dc mil-

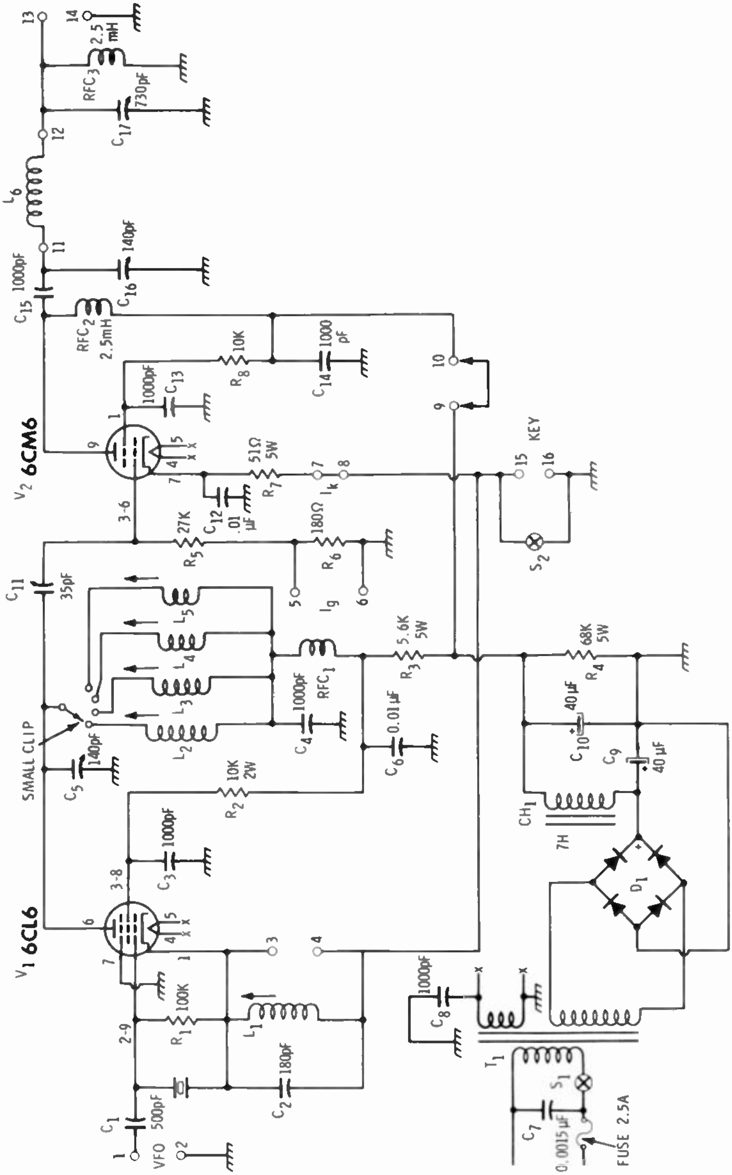


Fig. 3-11. Vacuum-tube cw/a-m transmitter.

liammeter between binding posts 5 and 6. Insert an 80-meter crystal and connect the 80-meter coil (L_3) into the circuit with a small clip.

Apply power to the oscillator but only filament power to the amplifier. Be sure there is no jumper between binding posts 9 and 10 if you have wired the class-C amplifier into the circuit.

Set capacitor C_{11} for maximum capacitance. Adjust capacitor C_5 for maximum glow of the neon bulb. Note the grid-current meter reading. Listen to the oscillator signal on your receiver. A neon bulb is useful for locating rf signals. In this application be certain that you and the neon bulb do not come in contact with high dc voltages.

Turn off the power and transfer the plate connection to the 160-meter coil (L_2). Insert a 160-meter crystal. Adjust C_5 for maximum neon glow and grid-current meter reading.

Parts List for Fig. 3-11.

C_1	500-pF disc or mica capacitor, 500V
C_2	180-pF disc or mica capacitor, 500V
C_3,C_4,C_8, C_{13},C_{14}, C_{15}	1000-pF disc or mica capacitors, 500V
C_5,C_{16}	140-pF variable capacitor
C_6,C_{12}	0.01-µF paper or disc, 500V or greater
C_7	0.0015-µF paper or disc, 125V ac
C_9,C_{10}	40-µF electrolytic capacitor, 450V
C_{11}	35-pF variable trimmer
C_{17}	730-pF variable capacitor; 2-section 365-pF per section
CH_1	Filter choke, 5 to 7H, 75 to 100 mA
D_1	Bridge rectifier, 2A, 800PIV
L_1	1 to 1.8µH slug-tuned inductor (J. W. Miller 21A156RB1)
L_2	73 to 90µH slug-tuned inductor (J. W. Miller 21A825RB1)
L_3	35.6 to 56.4µH slug-tuned inductor (J. W. Miller 21A475RB1)
L_4	10.8 to 18µH slug-tuned inductor (J. W. Miller 21A155RB1)
L_5	0.885 to 1.2µH slug-tuned inductor (J. W. Miller 21A106RB1)
L_6	Pi-network coils; B&W Miniductors, #3013, 3015, and 3016
R_1	100K ½-watt resistor
R_2,R_8	10K 2-watt resistor
R_3	5.6K 5-watt resistor
R_4	68K 5-watt resistor
R_5	27K ½-watt resistor
R_6	180-ohm ½-watt resistor
R_7	51-ohm 5-watt resistor
RFC_1, RFC_2, RFC_3	2.5-mH rf chokes
S_1,S_2	spst toggle switches
T_1	Power transformer; 240V sec., 100mA; 6.3V fil. 2.5A
V_1	6CL6 tube
V_2	6CM6 tube

Miscellaneous Parts

1 Pegboard, 8″ x 12″
2 9-pin miniature sockets
3 Crystals, 160, 80, and 40 meters
1 Crystal socket
1 2.5A fuse and holder
16 Binding posts
1 NE-2 neon bulb (for testing)

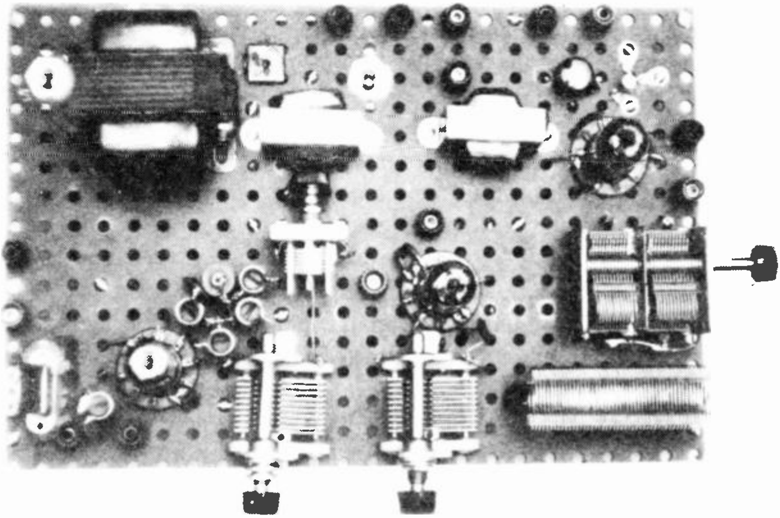


Fig. 3-12. Two-stage vacuum-tube transmitter with modulator.

Turn off the power and transfer the plate circuit to the 40-meter coil (L_4). Insert a 40-meter crystal. Tune capacitor C_5 away from its maximum setting until the neon bulb glows and there is an indication of grid current. Adjust the slug of L_4 to obtain 40-meter output at as near a maximum setting of capacitor C_5 as possible.

Now vary the capacitor toward minimum and note that at near minimum capacitance there will also be a neon indication and grid current. You have now tuned the output to the second harmonic of the crystal frequency, or 20 meters. You can use one 40-meter crystal for operation on both 20 and 40.

Leave capacitor C_5 at the same minimum capacitance that produces 20-meter output. Turn off the power and transfer the plate circuit to the smallest coil, inductor L_5. Turn on the power and retune capacitor C_5 toward maximum. At a reasonably high value of capacitance you will once again obtain 20-meter operation as indicated by the grid-current meter reading. Adjust the slug of L_5 until maximum reading is obtained at as high value of capacitance as possible.

Slow decrease the setting of capacitor C_5. At approximately midposition you will again read a peak in grid current. This will be the third harmonic of the 40-meter crystal. Continue decreasing capacitance and at a position very near to minimum capacitance setting there will again be grid-current meter reading. This will be the fourth harmonic of the 40-meter crystal. It will be somewhat lower in magnitude than the second and third harmonics. Note that you now have significant output capability from one crystal oscillator on all bands 10 through 160 meters.

PROJECT 14: VACUUM-TUBE CLASS-C AMPLIFIER

The class-C amplifier employs a 6CM6 beam-power tube. A pi-network is used in the output to match the tube to a low impedance load. The total plate and screen-grid current can be measured by inserting a 0-100 milliampere meter between binding posts 9 and 10 or 7 and 8. The stage includes a safety cathode resistor R_7. The radio-frequency choke at the output is also a safety device. If capacitor C_{15} shorted, the high dc voltage would appear between binding posts 13 and 14 on any antenna system. The rf choke would place a short across the output in terms of dc voltage and remove the shock hazard.

The coils are B&W type. For 160-meter operation use type 3016; on 40 and 80 meters a 3015 type modified as in Project 2. The tap provides 40-meter operation. A single type 3013 will tune to 10, 15, and 20 meters using the 140-pF variable capacitor C_{16}. These three coils will then provide operation on all bands 10 through 160 meters. Binding posts 11 and 12 permit coil changing.

Construction

Complete construction of the amplifier if you did not finish it while wiring the crystal-oscillator circuit. As shown in Fig. 3-12 the amplifier is mounted to the right of the oscillator. Coupling capacitor C_{11} can be seen at the approximate center of the pegboard. Coil ends are inserted into the binding posts located at the very bottom right.

Operation

Set up the transmitter for operation on 80 meters. Use the 80-meter inductor (L_3) and the combination 40-80 meter coil at L_6. Insert an 80-meter crystal. Turn on the oscillator only, and tune for maximum grid current.

Set capacitor C_{17} to its maximum setting. Connect your output indicator between binding posts 13 and 14. Use a 5-watt 51-ohm resistor as a dummy antenna.

Apply power to the amplifier and tune capacitor C_{16} for maximum output. Now adjust capacitor C_{17} for maximum output. Jockey back and forth between C_{16} and C_{17} until maximum output is obtained. Touch up C_5. A wattmeter connected to the output should measure between four and six watts. The plate current reading should be between 50 and 60 mA with a class-C supply voltage of 200-250 volts depending on power supply regulation. Dc input power will usually measure between 9 and 12 watts. Measure these quantities for your transmitter. Grid-current meter reading will be near to 1 mA.

With a supply voltage to the class-C amplifier of 250 volts and a cathode current of 50 mA, the dc input power would be:

$$P_{in} = E_b I_b = 250 \times 0.05 = 12.5 \text{ watts}$$

What would be the dc input resistance for matching to a modulator?

$$R_{in} = \frac{E_b}{I_b} = \frac{250}{0.05} = 5000 \text{ ohms}$$

If available, insert an 80-meter crystal for the cw end of the band somewhere between 3500 and 3600 kHz. Retune the transmitter. Turn off the power and use the 40-meter tap on output coil L_6. Adjust capacitor C_5, C_{16}, and C_{17} for maximum output. Jockey back and forth between C_{16} and C_{17} for peak output. Note that the amplifier operates very well as a frequency doubler producing a 40-meter output not too much lower in power level than the 80-meter output.

Turn off the power. Connect the 40-meter coil in the plate circuit of the oscillator (inductor L_1). Turn on the unit and peak capacitor C_5 for maximum output. Approximately the same output is obtained. However, in this case, the frequency doubling takes place in the output circuit of the crystal oscillator and the amplifier is being used straight through (40 meter input and output).

Turn off the power. Insert a 40-meter crystal. Retune the transmitter slightly. The entire transmitter is now being operated straight-through on the same frequency. A slightly greater output is obtained with straight-through operation.

Turn off the power and insert the 10-15-20 meter coil into the L_6 position. Tune capacitor C_{16} for maximum output (plates will be about three-quarters meshed). You are now operating the amplifier as a frequency doubler to obtain 20-meter output. Jockey back and forth between C_{16} and C_{17} for a good maximum signal output.

Now tune capacitor C_5 toward a minimum capacitance setting. Very near to minimum capacitance there will be another peak. In this case the oscillator is being used as a frequency doubler while the amplifier functions straight through. In this manner of operation a higher output is obtained.

Turn off the power and connect the plate of the oscillator to 10-15-20 meter coil L_5. Tune capacitor C_5 for maximum output near maximum capacitance setting. Retune C_{16} and C_{17} for maximum output. This is the combination that will provide the highest 20-meter output using the 40-meter crystal.

To obtain 15-meter output set capacitors C_{16} and C_5 to about midposition. Decrease the capacitance of C_{11} until plates are only one-third meshed. Vary capacitor C_5 for maximum grid current. Adjust capacitor C_{16} for maximum output. Jockey between C_{16} and C_{17} to obtain peak output. Momentarily remove the crystal from its socket. If there is a substantial output reading it indicates that the final amplifier is unstable. Decrease the capacitance setting of C_{11}. Retune the transmitter. Find a stable setting at which the crystal can be removed from the socket and the output drop away to zero or near zero. Obtainable output is less than on 20 meters. Some sacrifice in output is made in obtaining high stability.

Repeat the above procedure for 10-meter operation. In this case, both capacitors C_{16} and C_5 are set very near to minimum capacitance. Capacitor C_{11} should be set at minimum capacitance.

More output can be obtained on 10 meters by using the amplifier as a frequency doubler. To do this set capacitor C_{11} for maximum capacitance. Set capacitor C_5 for 20-meter output (near maximum setting). Capacitor C_{16} is set to near minimum capacitance to obtain 10-meter output. Ten-meter output will be near to two watts for this arrangement.

Turn off the power. Connect the plate of the oscillator to the largest coil (L_2). Insert a 160-meter crystal. Insert the 160-meter coil in the L_6 position. Tune up the transmitter on 160 meters.

This little rig will give you a QRP-cw signal on all bands 10 through 160 meters.

VFO Control

The variable-frequency oscillator of Project 9 can supply drive for the transmitter. This will give you a multiband tunable-frequency QRP transmitter.

Remove any crystal from the crystal socket of the oscillator. Place a jumper between bindings posts 3 and 4 to short out the tri-tet cathode LC combination. Connect the output of the vfo between binding posts 1 and 2.

Set the vfo on the desired frequency and band. Use the 40-, 80-, or 160-meter output of the vfo for operating on the 40-, 80-, and 160-meter bands. Excellent results are obtained on these bands.

Two methods of operation are possible on 20 meters. Using the 40-meter output of the vfo, you can tune the output of the former crystal oscillator to 40 meters and the output of the amplifier doubles to 20 meters. Another possibility is to use the 20-meter output of the vfo to operate both the former oscillator (now acting as an amplifier) and the amplifier on 20 meters.

PRINCIPLES OF AMPLITUDE MODULATION

Basically, modulation is a mixing process in which two frequencies—radio and modulating—are combined to produce four. As shown in Fig. 3-13, three high-frequency components form the modulated rf wave. The modulating wave itself, because it is of a much lower frequency, can be readily filtered and does not appear in the output.

In the example, a 1000-hertz modulating wave has been introduced via the plate circuit of the rf power amplifier. In fact, the dc plate voltage to the amplifier is made to vary with the modulating wave. The radio-frequency wave is applied to the grid circuit of the amplifier as shown in Fig. 3-14. The mixing activity produces original and sum and difference components as follows:

Radio-frequency wave—1,000,000 Hz
Upper sideband—1,000,000 Hz plus 1000 Hz
Lower sideband—1,000,000 Hz minus 1000 Hz
Modulating wave—1000 Hz (discarded)

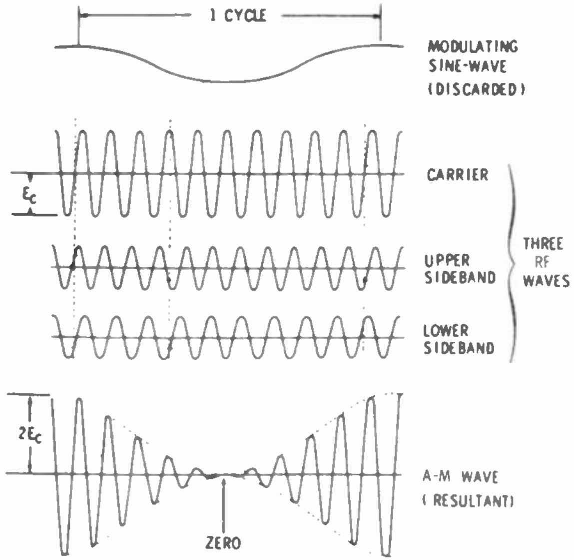


Fig. 3-13. Basic amplitude-modulation principles.

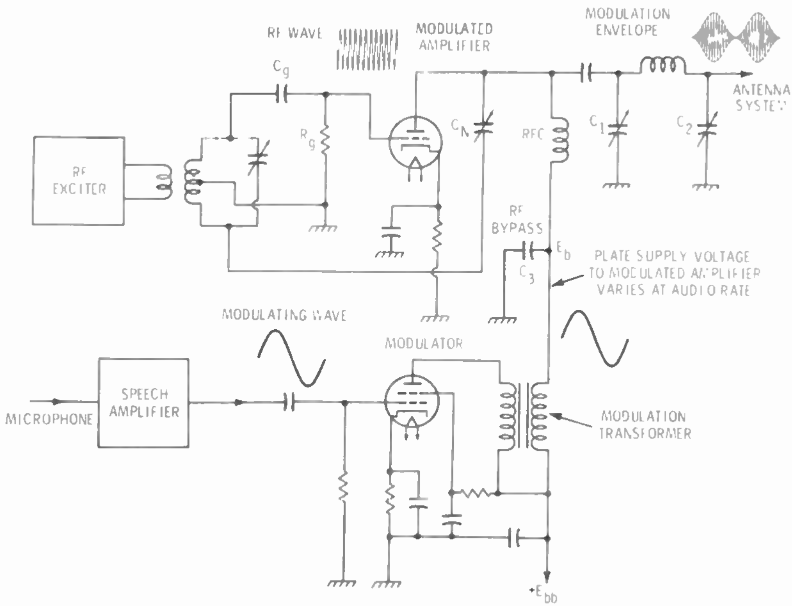


Fig. 3-14. Basic plate-modulation system.

Since the radio-frequency components are generated simultaneously, they continuously add to and subtract from each other. This produces a modulation envelope which varies with the frequency, shape, and instantaneous amplitude changes of the modulating wave, as shown in Fig. 3-13.

In considering class-C you learned that the tank-circuit voltage and power output of the amplifier are closely related to the dc plate-supply voltage. For example, over the linear operating range, any doubling of the supply voltage doubles the tank-circuit voltage. Hence, the rf output voltage will faithfully follow any change in dc plate voltage.

In a plate-modulated system, the plate supply voltage is varied by the modulating wave. If 500 volts is supplied to an amplifier, the modulating wave will cause this plate supply to vary above and below 500 volts. If the modulating wave is a pure sine wave, this variation will be symmetrical. The rate of variation depends on the frequency of the modulating wave, and the extent depends on the amplitude of the wave. In other words, the change in plate supply voltage becomes a replica of the modulating wave. If the variations on each side of the zero ac axis of the modulating wave are identical, the *average* dc plate voltage remains fixed at 500 volts.

What will be the influence of the plate-voltage change on the modulated amplifier? As you learned in the coverage of class-C amplifiers, the plate voltage determines the peak plate current. Thus as the plate voltage changes, so will the peak plate current drawn for each radio-frequency cycle. In fact, the current variation will also be a copy of the modulating wave, as shown in Fig. 3-15.

The peak plate current, in turn, determines the magnitude of the rf voltage developed across the plate tank circuit. The higher the current, the higher is the voltage. It follows then that the rf plate-voltage variation will also be a copy of the modulating wave. Notice in Fig. 3-15 that when the plate voltage drops to its lowest value, the peak plate current is minimum and the rf cycles across the plate tank coil fall. During the crest of the positive alternation of the modulating wave, the plate supply voltage rises to its maximum. Consequently peak plate current is drawn and the peak amplitude of the rf voltage across the tank circuit is also maximum.

An important characteristic of modulation is the modulation linearity. Is the rf peak-voltage change an exact copy of the variations

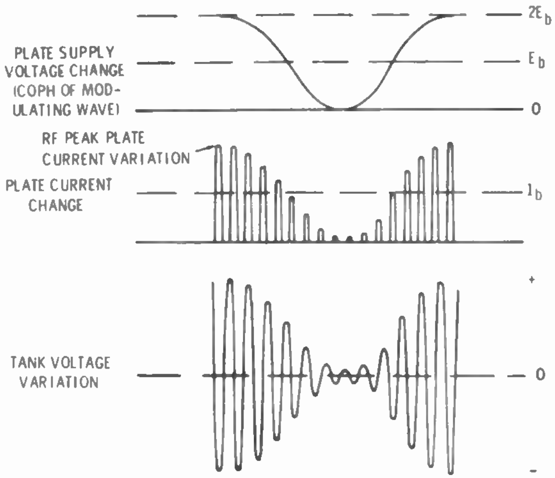


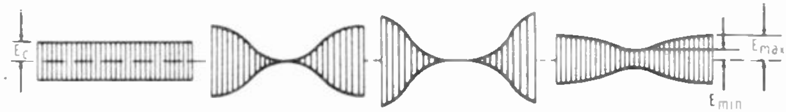
Fig. 3-15. Class-C amplifier modulated waveforms—under 100% modulation.

of the modulating wave? In a class-C amplifier there is a linear relationship between the plate supply voltage and rf output voltage. This is ideal for obtaining linear modulation, and if a class-C amplifier is adjusted properly for use as a modulated amplifier, a very linear modulation characteristic can be obtained.

Note in Fig. 3-13 that the variations of the modulation envelope (on both sides of the zero axis) conform to the shape of the modulating wave. Good linearity is a definite advantage of plate modulation. A modulation envelope can be displayed on an oscilloscope screen. By observing and measuring it, much can be disclosed about the linearity and general operation of modulators and modulated amplifiers.

Modulation Percentage

When the amplitude of the rf voltage swings to twice the unmodulated value on a positive modulating crest and falls to zero on the negative crest, the carrier is said to be 100% modulated. This is shown in Fig. 3-16. Full modulation of the carrier is important in obtaining a good transmission range and maximum demodulated audio at the receiver.



(A) Carrier. (B) 100%. (C) Over 100%. (D) Less than 100%.

Fig. 3-16. Modulation percentages.

To obtain a linear and undistorted modulation characteristic, the amplitude of the positive crest of the modulating rf sine wave must be not more than twice the amplitude of the unmodulated value. Otherwise, the negative crest will flatten off at zero, as shown in Fig. 3-16C. Here the negative and positive crests of the envelope become asymmetrical and distortion is present. On the positive crest, the modulation swings above twice the unmodulated value. During the negative crest, the rf carrier falls away to zero for an extended period of time. This is called *overmodulation*—it produces a distorted envelope and also causes the generation of sideband-frequency components that are widely separated from the carrier frequency. These undesired sideband components, referred to as splatter, can cause interference on adjacent channels.

Fig. 3-16D shows a carrier that is modulated less than 100%. In other words, the rf voltage does not rise to twice the unmodulated value on the positive crest. Nor does it fall away to zero on the negative crest of the modulating wave. Fig. 3-16D represents approximately 50% modulation of the carrier. A formula often used to calculate modulation percentage is:

$$\%\ \text{mod.} = \frac{E_{max} - E_{min}}{2E_{car}} \times 100$$

where,

E_{max} is the maximum amplitude of the modulated carrier,
E_{min} is the minimum amplitude of the modulated carrier,
E_{car} is the amplitude of the unmodulated carrier.

Since a modulating signal is seldom a pure sine wave but rather a combination of various frequencies of differing amplitudes, it is often useful to determine positive and negative modulation percentages separately. Positive modulation refers to the increase in rf-cycle amplitude above the unmodulated value; negative modulation, to the drop in rf-cycle amplitude below the unmodulated value. The positive- and negative-modulation percentage formulas are:

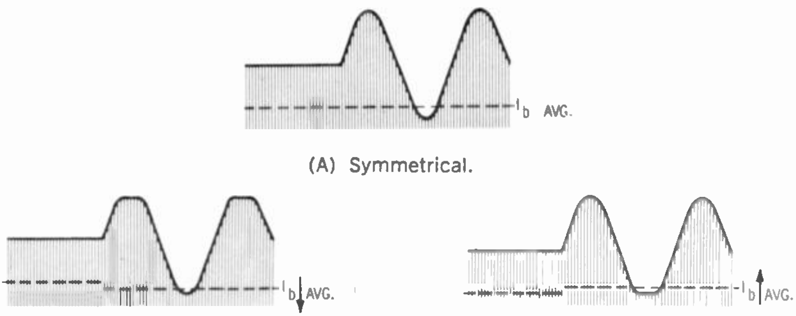
$$\% \text{ mod. (positive)} = \frac{E_{car} - E_{min}}{E_{car}} \times 100$$

$$\% \text{ mod. (negative)} = \frac{E_{max} - E_{car}}{E_{car}} \times 100$$

Many of the amplitude modulation meters include facilities for measuring both positive- and negative-modulation percentages. Overmodulation in the negative direction is to be avoided; it represents a swing in a direction that can cut off the carrier because no plate voltage is being supplied to the modulated amplifier for an extended time. Severe distortion and interference will thus occur.

Most modulating waves are not of the constant amplitude that maintains 100% modulation continuously. The average during voice modulation is considerably lower, and only on an occasional voice peak does the modulation reach 100%. Nevertheless an a-m system must be adjusted so that there will be no overmodulation on voice peaks. Since the voice peaks are only occasional, it is apparent that the average modulation must be considerably lower than 100%.

When the modulating wave is symmetrical and the modulation is linear, there is a like swing of the rf voltage on each side of the



(A) Symmetrical.

(B) Downward.

(C) Upward.

Fig. 3-17. Upward and downward modulation and carrier shift.

unmodulated value, as shown in Fig. 3-17. Likewise the peak plate-current variation is the same on each side. Thus, regardless of the percentage of modulation (up to 100%), the average dc component of plate current is unchanged and is equal to the unmodulated value. Hence, the reading of a dc plate-current meter, inserted into the plate-supply voltage line, does not change with modulation because the meter reads the average dc component of plate current.

If modulation is asymmetrical, the plate current will swing up or down with respect to its unmodulated value whenever the stage is modulated. This condition can be caused by nonlinearity in the actual modulation process or by a distorted modulating signal.

Asymmetrical modulation of a carrier is sometimes referred to as upward or downward modulation. When the peak plate current and rf voltage of the modulation envelope do not rise as much on the positive peaks as they dip during negative peaks for sine-wave modulation, the condition is referred to as downward modulation. Sometimes it is referred to as carrier shift because the actual dc component of plate current (in the case of a plate-modulated system) falls with modulation. This is shown in Fig. 3-17B.

When the positive peaks rise higher than the negative peaks fall, as shown in Fig. 3-17C, the condition is referred to as upward modulation. Notice that the average dc component of plate current rises with modulation.

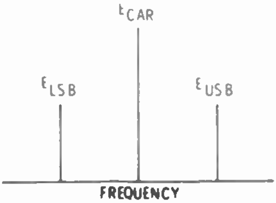
Carrier and Sideband Power

As mentioned previously, the modulation envelope is a composite waveform. During the modulation each rf cycle differs from the preceding and following ones. They are no longer pure sine waves. This is another way of stating that the resultant signal is no longer composed of a single wave (rf carrier) but, in the case of modulation by a single sine-wave tone, a carrier plus upper and lower sidebands. Actually the carrier, despite modulation, remains of constant magnitude and frequency. The only difference has been the addition of two sideband components.

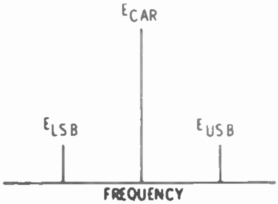
The foregoing relationships indicate that the carrier power is unchanged with modulation. Any net addition in power is contributed by two sidebands.

The instantaneous power of the modulation envelope varies continuously during the modulating cycle. At the peak of the envelope, assuming 100% modulation, the rf voltages are twice as high as during the unmodulated period. This doubling of the voltage represents a four fold increase in power because the rise in power is equal to the square of the voltage. At the crest of the negative modulation, the resultant envelope falls to zero and the instantaneous power is zero.

When the power is averaged over the complete sine-wave cycle for 100% modulation, it is 50% higher than the unmodulated carrier power. One hundred percent modulation occurs whenever the amplitude of each sideband is one-half the amplitude of the unmodulated carrier, as shown in Fig. 3-18A. This means that each side-



(A) 100% modulation.



(B) 50% modulation.

Fig. 3-18. Carrier and sideband voltage levels.

band contains one-fourth as much power as the carrier. Since there are two sidebands, they contain one-half the power of the carrier.

The modulating information to be transmitted is contained in the sidebands, each of which contain the same information. The carrier itself contains none of the information to be transmitted.

Let us now reconsider the power relationship in terms of the total power output. The total output (envelope power) is the sum of the power contained in the carrier and sidebands or:

$$P_o = P_{car} + P_{usb} + P_{lsb}$$

where,

P_o is the total output (envelope) power,
P_{car} is the carrier power,
P_{usb} is the upper sideband power,
P_{lsb} is the lower sideband power.

Assuming carrier power as unity:

$$P_o = 1 + \frac{1}{4} + \frac{1}{4} = \frac{3}{2}$$

Since the sidebands contain only one-half the power of the carrier, they represent only one-third the total transmitted power. The carrier represents the remaining two-thirds:

$$\frac{P_{car}}{P_o} = \frac{1}{3/2} = \frac{2}{3}$$

$$P_{car} = \frac{2}{3} P_o$$

$$\frac{P_{sb}}{P_o} = \frac{\frac{1}{2}}{\frac{3}{2}} = \frac{1}{3}$$

$$P_{sb} = \frac{1}{3} P_o$$

It is apparent that a-m is a rather wasteful method of conveying information via a radio wave. All the information to be sent can be packed into one of the sidebands, representing one-sixth the total transmitted power. The relationship is even more exaggerated for modulation of less than 100%—the prevailing condition in the usual voice communications. An example of 50% modulation is shown in Fig. 3-18B.

For 50% modulation the sideband amplitude is only one-quarter the carrier amplitude. Therefore each sideband contains but one-sixteenth the power in the carrier. Hence the power in both sidebands is only one-eighth the carrier power, a very small fraction of the total transmitted power. This relationship proves why a high average modulation percentage is important in maintaining an effective transmission range and strong audio-frequency components at the receiver.

The efficiency of the average well-designed class-C modulated amplifier is approximately 70% and some higher-powered ones operate with efficiencies up to 80%. The efficiency factor is the quotient of the rf power output and dc power input, as follows:

$$\text{Efficiency factor} = \frac{\text{rf power output}}{\text{dc input power}}$$

$$= \frac{P_o}{I_b E_b}$$

The dc power input remains constant, whether the carrier is modulated or unmodulated. Likewise the carrier power output is a constant, and the rf carrier output is supplied by conversion from the plate supply of the modulated amplifier.

However, the rf power output increases with modulation. What is the source of this additional power? You learned that the additional power output with modulation is contained in the two sidebands. The sidebands are due to introduction of the modulating wave into the plate supply line, via the modulator. Consequently, the rf sideband power is contributed by the audio-output system or modulator.

Modulator Power

With 100% sine-wave modulation, the rf power output of the transmitter is increased by 50%. To make this much additional power available, the audio power output of the modulator must be one-half the dc power input to the modulated amplifier. However, since only occasional voice peaks extend to 100% modulation, the average power output of the modulator can be less than 50% of the dc power input.

For 100% modulation, the audio peak voltage must be approximately equal to the supply voltage. Only under this condition will the plate voltage of the modulated amplifier swing to twice the plate supply voltage, a necessity for obtaining 100% modulation at the crest of the positive audio cycle. The negative sweep of the same modulating wave reduces the plate supply voltage to zero to produce the 100% modulated envelope trough.

To produce the desired peak audio voltage at the required power, the audio must be developed across a specific impedance. The impedance seen by the modulator output must equal the dc input resistance to the modulated amplifier. This dc input resistance is related to the dc component of plate current and plate voltage, or:

$$R_{in} = \frac{E_b}{I_b}$$

The modulation transformer or the output system of the modulator must be designed to match this impedance, just as an audio output transformer must have the proper turns ratio to match the impedance of a speaker voice coil.

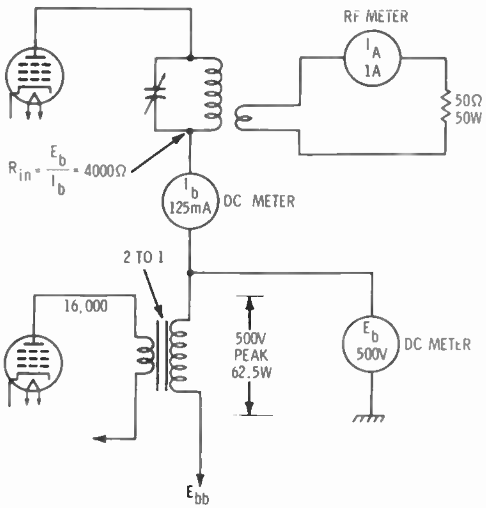


Fig. 3-19. Mathematical relationships.

The example in Fig. 3-19 will demonstrate this important relation. Let us assume that the modulated amplifier delivers a 50-watt carrier into a 50-ohm antenna load. Let us further assume the class-C amplifier has an efficiency of 80%.

What is the dc power input to the modulated class-C amplifier? If the efficiency factor is 0.8, the input is as follows:

$$P_{in} = \frac{P_o}{0.8}$$

$$= \frac{50}{0.8}$$

$$= 62.5 \text{ watts}$$

What is the dc plate current? Since 500 volts is supplied, it will be:

$$P_{in} = I_b E_b$$

$$I_b = \frac{P_{in}}{E_b}$$

$$I_b = \frac{62.5}{500}$$

$$I_b = 125\text{mA}$$

The dc input resistance load into which the modulator must work is calculated as follows:

$$R_{in} = \frac{E_b}{I_b}$$

$$= \frac{500}{.125}$$

$$= 4000 \text{ ohms}$$

How many watts of audio must be supplied to obtain a sine-wave modulation of 100%? The available audio power must be one-half the dc power input, or:

$$P_{audio} = \frac{P_{in}}{2}$$

$$= \frac{62.5}{2}$$

$$= 31.25 \text{ watts}$$

What is the output voltage of the modulator? Since 31.25 watts must be developed across 4000 ohms, the rms peak audio voltage must be:

$$P_{audio} = \frac{E^2}{R}$$

$$E_{rms} = \sqrt{PR}$$

$$= \sqrt{31.25 \times 4000}$$

$$= 353.5 \text{ volts}$$

What is the peak audio voltage? The conversion between rms and peak voltage is as follows:

$$E_{peak} = E_{rms} \times 1.414$$

$$= 353.54 \times 1.414$$

$$= 500 \text{ volts}$$

Notice that this 500 volts peak matches the dc supply voltage. Hence the instantaneous plate-supply voltage swings between zero and 1000 volts, a condition of 100% modulation.

What must the turns ratio of the modulation transformer be if the modulator tubes must work into an impedance of 16,000 ohms? In the example, the modulator must match 4000 ohms. Consequently the step-down transformer ratio must be as follows:

$$\text{Ratio} = \frac{N_P}{N_S}$$
$$= \sqrt{\frac{Z_P}{Z_S}}$$
$$= \sqrt{\frac{16{,}000}{4000}}$$
$$= 2$$

or a 2-to-1 turns ratio (primary to secondary) step-down transformer.

Let us next consider the power delivered to the load, as well as the load or antenna current. Assume the antenna system presents a 50-ohm load to the output of the transmitter. As established previously, the carrier output power is 50 watts and the efficiency of the modulated amplifier is 80%.

What is the unmodulated rf antenna current? Since the power and impedance are known, the antenna current can be calculated as follows:

$$I_A = \sqrt{\frac{P}{R_A}}$$
$$= \sqrt{\frac{50}{50}}$$
$$= 1 \text{ ampere}$$

Under 100% modulation, what is the sideband power delivered to the 50-ohm load? As calculated previously for 100% modulation the required audio power is 31.25 watts. Since the efficiency of the modulated amplifier is 80%, the audio rf sideband power is:

$$P_{sb} = \text{efficiency} \times P_{audio}$$
$$= .8 \times 31.25$$
$$= 25 \text{ watts}$$

Notice that the sideband power is one-half the carrier power, as it should be for 100% sine-wave modulation.

What is the antenna current with 100% modulation? Since the rf power output increases with modulation, one can anticipate an increase in the rf antenna current. Its value will be:

$$I = \sqrt{\frac{P}{R}}$$

$$= \sqrt{\frac{75}{25}}$$

$$= 1.224 \text{ ampere}$$

What is the percentage of increase in antenna current with 100% modulation? The antenna current increases from 1 ampere to 1.224 amperes, or:

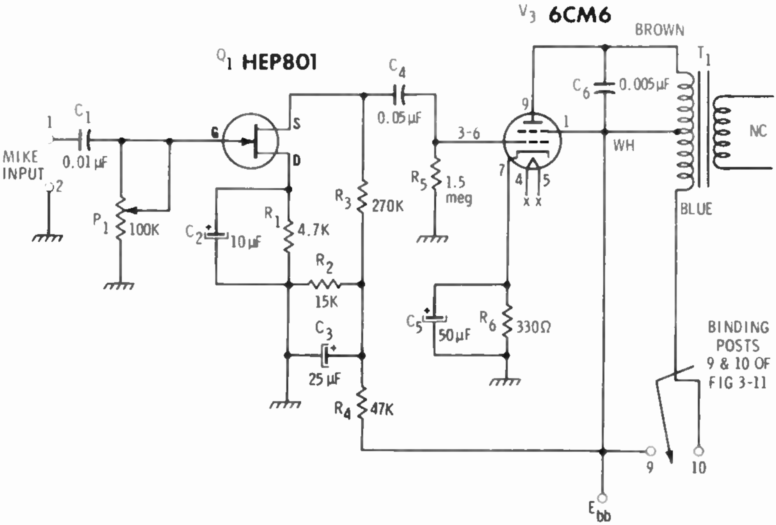
$$\% \text{ increase} = \frac{1.224 - 1}{1} \times 100$$

$$= 22.4\%$$

This figure suggests that with modulation, one can expect an increase in the antenna current. In fact, when 100% modulation is attained, the antenna current increases by 22.4%. The antenna-current meter provides a good way of monitoring the performance of an a-m transmitter, and assists in making the modulation adjustments necessary to obtain an adequately high level of modulation. Ordinary voice modulation contains only a few peaks that extend to 100%. Consequently the antenna current increases but does not rise as much as 22.4% with normal conversation. An antenna-current increase of 10 to 15% is more reasonable and indicates proper modulation.

PROJECT 15: VACUUM-TUBE MODULATOR

A two-stage amplifier-modulator will provide sufficient audio power to modulate the transmitter of Project 14. Audio power output need be only 4 to 5 watts peak. A simple speech amplifier and modulator is shown in Fig. 3-20. The speech amplifier is a field-effect transistor; the modulator, a beam-power 6CM6 (same type tube used in



C_1	0.01-µF paper or disc capacitor	R_1	4.7K ½-watt resistor
C_2	10-µF 12-16V electrolytic capacitor	R_2	15K ½-watt resistor
C_3	25-µF 25V electrolytic capacitor	R_3	270K ½-watt resistor
C_4	0.05-µF paper or disc capacitor	R_4	47K ½-watt resistor
C_5	50-µF 25V electrolytic capacitor	R_5	1.5-megohm ½-watt resistor
C_6	0.005-µF paper or disc capacitor, 500V or higher	R_6	330-ohm 2-watt resistor
P_1	100K potentiometer	T_1	Modulation choke or transformer (1 to 1) see text
Q_1	HEP-801 FET	V_3	6CM6 tube

Miscellaneous Parts

Pegboard (same board as transmitter)	1 Transistor socket
1 9-pin miniature tube socket	2 Binding posts

Fig. 3-20. Speech amplifier and modulator.

the output stage of the transmitter). The modulator operates as a class-A audio power amplifier. Choke modulation is employed. One of the autotransformer chokes used in vacuum-tube citizens band transceivers can be employed. An alternative is a modulation transformer that will match the modulator to the input of the modulated class-C amplifier.

In Project 14 we calculated that the approximate input resistance R_{in} for the modulated amplifier was 5000 ohms. Typical output impedance for the 6CM6 as a class-A amplifier is also about 5000 ohms. Therefore, either a choke inductor can be employed or a 1-to-1 (5000-ohms to 5000-ohms) modulation transformer will do the job.

The field-effect transistor is operated from the plate supply voltage. Resistors R_2 and R_1 act as a voltage divider. Resistor R_3 is of high value and the combination of the three resistors keep the drain voltage at a safe value and develop a good ac output voltage for driving the grid of the modulator tube. Source bias is developed by the resistor-capacitor combination R_1C_2 while cathode bias is established by resistor R_6 and capacitor C_5. Capacitor C_6 limits high-frequency components to those required for good speech intelligibility.

The field-effect transistor and speech gain control can be seen at the very top right of Fig. 3-12. The modulation choke is mounted at top center with the audio modulator tube to its right. Binding posts 9 and 10 are mounted near to the modulator choke.

For cw operation the modulator tube should be removed from its socket and a jumper connected between binding posts 9 and 10. For a-m modulation the jumper is removed and the modulation tube inserted in its socket. Binding posts 9 and 10 are shown in both Figs. 3-11 and 3-20.

Wire the speech amplifier and modulator. Check your wiring carefully.

Operation

Remove the modulator tube and place the jumper between binding posts 9 and 10. Connect a 51-ohm dummy antenna load. Insert an 80-meter crystal. Tune up the transmitter. After tune-up remove the jumper; there may be some slight drop in output.

Now plug in the modulator tube. Power output may fall away because of the additional demand made on the power supply. Retune capacitors C_{16} and C_{17} slightly. Insert the microphone and set the gain control to near maximum.

Connect a service-type oscilloscope across the dummy load or a modulation oscilloscope, such as the *Heathkit SB-610*, in the line between the transmitter output and the dummy load. Speak into the microphone and observe the modulation process. Whistle into the microphone and try and stop the scope pattern. Peak modulation will be close to 100%. In some instances, it may be necessary to set your gain control to maximum to obtain near 100% modulation.

Modulation linearity will be acceptable and there is little danger of any significant overmodulation. Repeat the above procedure tuning up to an 80-meter dipole antenna. Have a local ham check your modulation quality.

Tune up your transmitter on other bands. Make similar checks. If you wish, you can now build up the transmitter permanently on a chassis and mount it in a metal case. It will give you a multiband cw/a-m capability for low-power local operation, as well as facility for some longer distance QRP operations.

CHAPTER 4

Transistor/Tube Transmitter Circuits

Many modern transmitters employ both solid-state devices and vacuum-tubes. Often, such a unit is completely solid state except for the final rf power amplifiers. There are a number of transmitters that employ but a single vacuum-tube. One popular commercial broadcast transmitter is so designed.

Vacuum tubes have advantages in high-powered rf power circuits because they are easy to drive (such as from a small solid-state rf exciter), tune up easily, and are more free of distortion and spurious frequency components.

Their disadvantages are the need for a high-voltage power supply; filament power; and, at high-power levels, adequate ventilation and, at times, forced-air cooling.

The solid-state and vacuum-tube combination is attractive for experimental and home-brew ham transmitters. A beam-power rf tube can be driven fully from a simple solid-state exciter, developing a good output. In the case of amplitude modulation, one encounters a little less trouble in obtaining linear and full modulation of a vacuum tube. The modulator itself can be solid state. If silicon rectifiers are employed in the power supply, a one tube transmitter is possible.

In this chapter you will experiment with solid-state/vacuum-tube combinations. For one transmitter the exciter for the modulated rf-amplifier tube is a single integrated circuit. A 10-15-20 version incorporates a field-effect vfo and amplifier that will drive a vacuum-tube final amplifier. Vacuum-tube and FET linear amplifiers complete the chapter.

PROJECT 16: IC VFO

There are many applications for integrated circuits in ham radio. Effective and instructive circuits can be constructed for ham transmitters. There are a quite number of IC devices that perform well in versatile modulator and demodulator systems for the various modes of modulation. Some of these will be covered in later chapters.

Integrated circuits of considerable power level have been used in audio services. A number of these devices, although their use has been confined mainly to audio frequency work, have a relatively good high-frequency response. One such device is the RCA CA3020

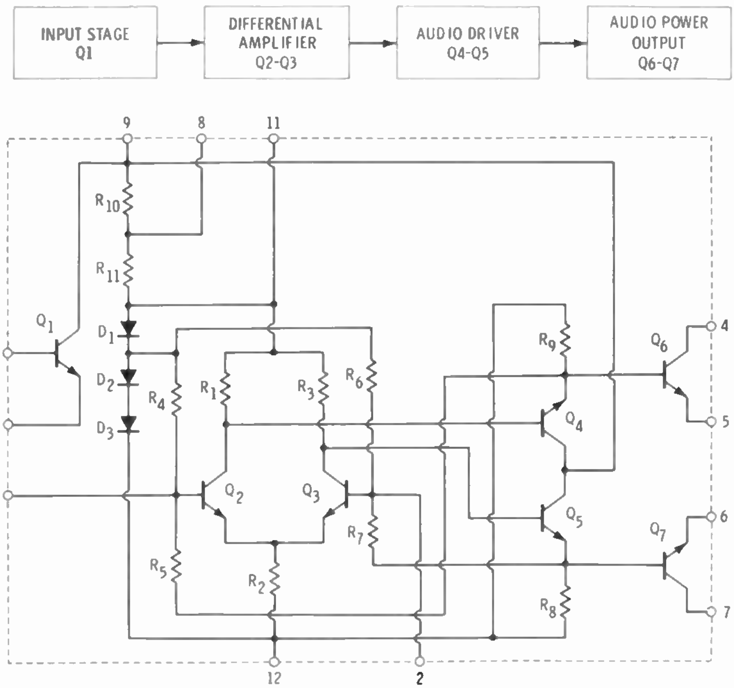
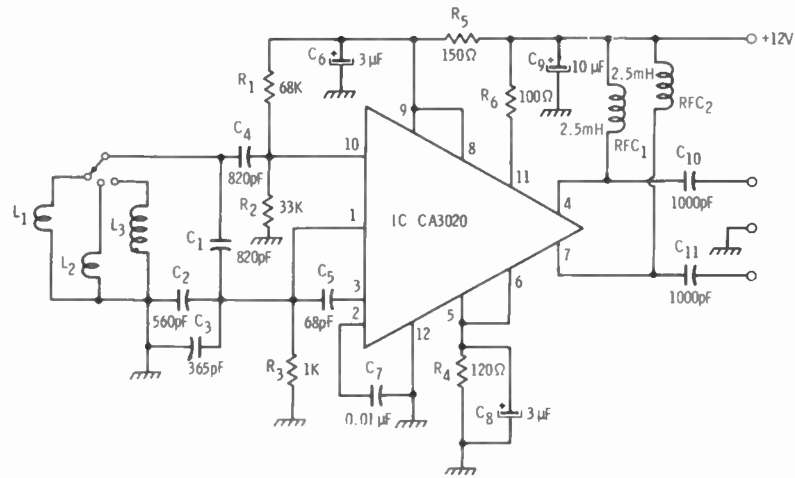


Fig. 4-1. Internal plan of CA3020 integrated circuit.

(Fig. 4-1). It is known mainly as an audio power amplifier capable of delivering several watts of audio output. However, this unit does function well in amplifier and oscillator circuits up to 8 MHz. It will also operate as a linear rf power amplifier when biased properly.

Internally it consists of an input stage, differential amplifier, driver, and power-output configuration as shown in Fig. 4-1. In rf application the input stage (Q_1) can be used as an oscillator. The remainder of the IC will build up the power level to several hundreds of milliwatts of rf output. For this manner of operation the emitter of Q_1 is coupled to the bases of the differential amplifier by joining terminals 1 and 3. If the unit is to be employed strictly as an amplifier,



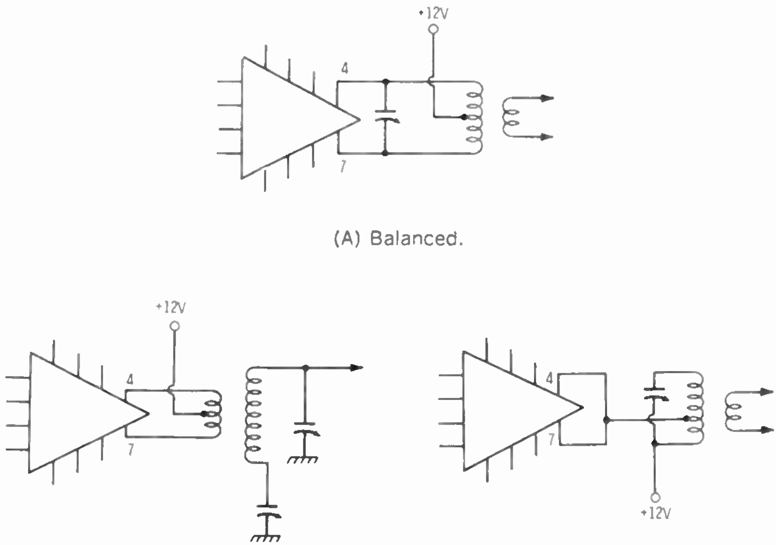
C_1	820-pF polystyrene-film or silver-mica capacitor
C_2	560-pF polystyrene-film or silver-mica capacitor
C_3	365-pF variable capacitor
C_4	820-pF polystyrene-film or silver-mica capacitor
C_5	68-pF polystyrene-film or silver-mica capacitor
C_6	3-μF 15V electrolytic capacitor
C_7	0.01-μF paper or ceramic disc capacitor
C_8	3-μF 15V electrolytic capacitor
C_9	10-μF 15V electrolytic capacitor
C_{10},C_{11}	1000-pF ceramic-disc capacitors.
IC_1	CA 3020 integrated circuit
L_1,L_2,L_3	Coils for frequency desired. J. W. Miller 21A225RB1 for 160 meters
RFC_1, RFC_2	2.5-mH rf chokes
R_1	68K ½-watt resistor
R_2	33K ½-watt resistor
R_3	1K ½-watt resistor
R_4	120-ohm 1-watt resistor
R_5	150-ohm ½-watt resistor
R_6	100-ohm 1-watt resistor

Fig. 4-2. IC vfo and amplifier.

transistor Q_1 can be deactivated and the signal applied directly to the input of the differential amplifier.

A typical circuit arrangement is shown in Fig. 4-2. The input transistor is used as a Seiler vfo. For multiband operation a switch and appropriate coils can be incorporated. Note that the emitter output of the oscillator is coupled to the input of the differential amplifier by capacitor C_5 which connects between terminals 1 and 3. Supply voltages arrive by way of terminals 8, 9, and 11 while common is connected to terminal 12 to complete the circuit. Unbalanced or push-pull balanced outputs can be derived at the collectors of the output transistors (terminals 4 and 7). The output load in this case is a pair of rf chokes. In this arrangement the integrated circuit is being used to supply drive voltage to a succeeding stage. This can be a push-pull stage if the balanced outputs are used. Only one side is employed if signal is to be applied to an unbalanced input.

Resonant outputs are also possible using any one of the arrangements of Fig. 4-3. The remainder of the circuit is identical except for the manner of connecting the output transistors to the resonant circuit. Fig. 4-3A shows a balanced arrangement, while Fig. 4-3B provides drive to an unbalanced succeeding stage or antenna in the



(A) Balanced.

(B) Unbalanced.

(C) Push-push doubler.

Fig. 4-3. Other output plans.

case of a QRPP transmitter. Output transistors can also be connected in a push-push doubler as shown in Fig. 4-3C. Note that collectors 4 and 7 are joined together.

Construction

In this initial project a CA3020 is used as a 160-meter variable-frequency oscillator and amplifier. Output will be unbalanced and will supply drive to the grid circuit of a beam-power vacuum tube. This project will deal only with the integrated circuit. However, the vacuum-tube stage will be added in the next project. The complete rf section of a 160-meter transmitter including power supply, can be seen in Fig. 4-4.

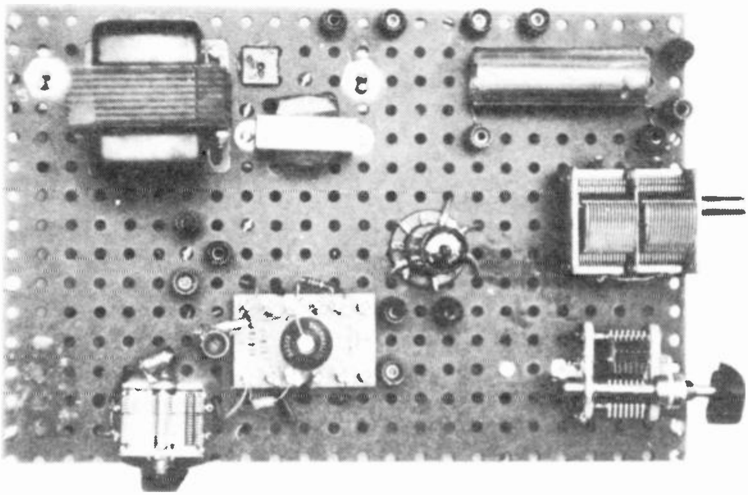


Fig. 4-4. IC VFO and vacuum-tube amplifier.

You will note that the integrated circuit is plugged into a *Vector* 12-pin socket for a TO-5 mount. The IC terminals of this mount are brought out to spring-loaded connection terminals. These are ideal for experimental work with integrated circuits.

A complete schematic diagram of the 160-meter transmitter including vacuum-tube rf amplifier is given in Fig. 4-5. The integrated circuit vfo is identical to that given in Fig. 4-2 except for a few component values and the use of an unbalanced output.

Build up the unit around the IC board socket. The capacitor

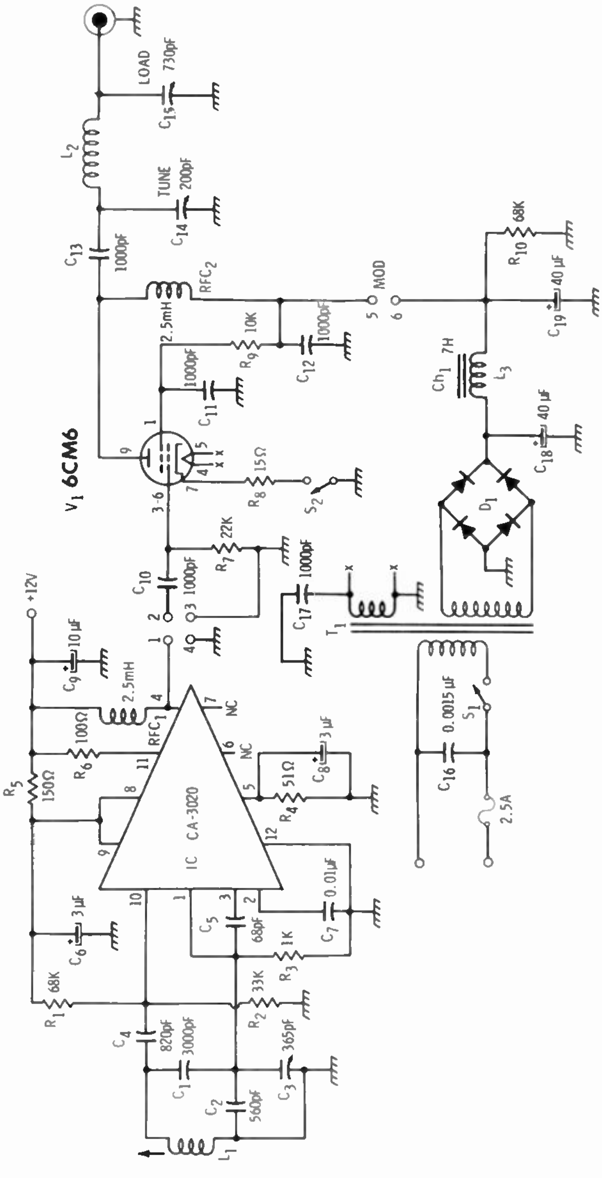


Fig. 4-5. 160-meter transmitter.

and coil of the oscillator are attached to the main pegboard. Use a heat sink on the integrated circuit.

Operation

Set the variable capacitor of the oscillator to midposition. Tune your receiver to the center of the legal segment of the 160-meter band for your area. Apply power to the oscillator. Adjust the coil slug until the signal is heard in the receiver.

If an oscilloscope is available observe the oscillator signal at terminal 3 of the IC. Also observe the signal at terminal 4 which is the IC output. The waveform at this point is pulse like in appearance because no output resonant circuit is employed. This waveform, however, will provide the necessary drive for the vacuum-tube class-C amplifier.

Keep the integrated circuit in operation. Zero-beat the signal on the receiver. Notice if there is any frequency drift of the oscillator

Parts List for Fig. 4-5.

C_1	3000-pF polystyrene-film or silver-mica capacitor
C_2	560-pF polystyrene-film or silver-mica capacitor
C_3	365-pF variable capacitor
C_4	820-pF polystyrene-film or silver-mica capacitor
C_5	68-pF polystyrene-film or silver-mica capacitor
C_6	3-μF 15V electrolytic capacitor
C_7	0.01-μF paper or ceramic-disc capacitor
C_8	3-μF 15V electrolytic capacitor
C_9	10-μF 15V electrolytic capacitor
C_{10},C_{11}, C_{12},C_{13} C_{17}	1000-pF disc capacitors
C_{14}	200-pF variable capacitor
C_{15}	730-pF variable capacitor (dual 365-pF)
C_{16}	0.0015 paper capacitor
C_{18},C_{19}	40-mF 450 electrolytic capacitors

Ch_1	5-7 henry filter choke
D_1	Bridge rectifier 2A 800 PIV
IC_1	RCA CA3020 integrated circuit
L_1	16.2 to 26.4 μH slug-tuned coil (J.W. Miller 21A225RB1)
L_2	B&W 3015 or 3016 miniductor (see text)
R_1	68K ½-watt resistor
R_2	33K ½-watt resistor
R_3	1K ½-watt resistor
R_4	51-ohm 1-watt resistor
R_5	150-ohm ½-watt resistor
R_6	100-ohm 1-watt resistor
R_7	22K ½-watt resistor
R_8	15-ohm 5-watt resistor
R_9	10K 5-watt resistor
R_{10}	68K 5-watt resistor
RFC_1 RFC_2	2.5-mH rf chokes
S_1,S_2	SPST toggle switches
T_1	Power transformer; 240V sec. @100mA, 6.3V fil. @2.5A
V_1	6CM6 vacuum tube

Miscellaneous Parts

2.5A fuse and holder
Pegboard 8″ x 12″
IC socket, Vector 570-F
9-pin miniature tube socket
12 binding posts
12V battery (Eveready 732)

over a one-half hour or one hour time period. Determine the possible frequency drift. This test can only be made accurately when the receiver itself does not drift in frequency.

Turn off the oscillator. After a 15-minute interval turn it back on. Is the oscillator on the same frequency? It may be off slightly but it should quickly zero-beat again.

PROJECT 17: 160-METER IC AND VACUUM-TUBE TRANSMITTER

The rf amplifier is a single-tube beam-power 6CM6. In the circuit of Fig. 4-5 it will deliver about 4 or 5 watts of carrier output when driven directly by the output of the IC. There will be adequate isolation between the oscillator transistor and the vacuum-tube stage to permit amplitude modulation of the vacuum tube.

The circuit is a straightforward class-C amplifier with no neutralization required. Two pairs of binding posts are used to link the output of the IC to the grid input of the amplifier. This will permit the vacuum-tube stage to be excited by different forms of solid-state or vacuum-tube exciters if desired.

The grid resistor and capacitor combination establishes the class-C operating bias. A 15-ohm cathode resistor acts as a safety device in case of loss of rf excitation. The output circuit is a pi-network using a B&W 3016 miniductor.

The supply voltage is shunt-fed through an rf choke. Two binding posts (5 and 6) in this path permit metering of the plate-screen current and as a point for the insertion of the modulation wave.

Operation

Wire the amplifier stage and connect the output of the integrated circuit to the input of the stage. Turn on the IC rf exciter. Set its frequency to the midpoint of the usable 160-meter spectrum for your area. Heat up the vacuum-tube filament.

Connect a 50- or 68-ohm termination to the output along with the rf output indicator. Connect a 0-100 mA dc meter in the supply line (between binding posts 5 and 6). Apply screen and plate power to the amplifier. Adjust the tune and load capacitors for maximum output. The output power should be a minimum of four watts.

What is the d-c input power? With an approximate plate voltage of 300 and plate current of 30 mA it would be 9 watts. A neon bulb glows brightly when held near to the plate side of the pi-network inductor. An oscilloscope connected across the ouput will display a pure 160-meter sine wave.

The amplifier can also be used as a frequency doubler to obtain two to three watts of carrier on 80 meters. To do so, one need only insert an 80-meter coil (B&W 3015 miniductor). Tune up the transmitter on this band.

PROJECT 18: UNIVERSAL MODULATOR

A variety of solid-state higher-powered audio modules can be purchased at reasonable cost. These modules are designed for use in record or tape reproducing systems, public address amplifiers, and commercial audio systems. Usually two or more separate output impedances are included. Units with power output levels of 10 watts and higher are available.

They are ideal for modulator application in low-power ham transmitters. Audio transformers can be connected to their output to obtain any required impedance, *stepup or stepdown.*

A typical unit is the 10-watt Amperex PC-8-36 shown in Fig. 4-6. It consists of a two-transistor input circuit followed by a pair of

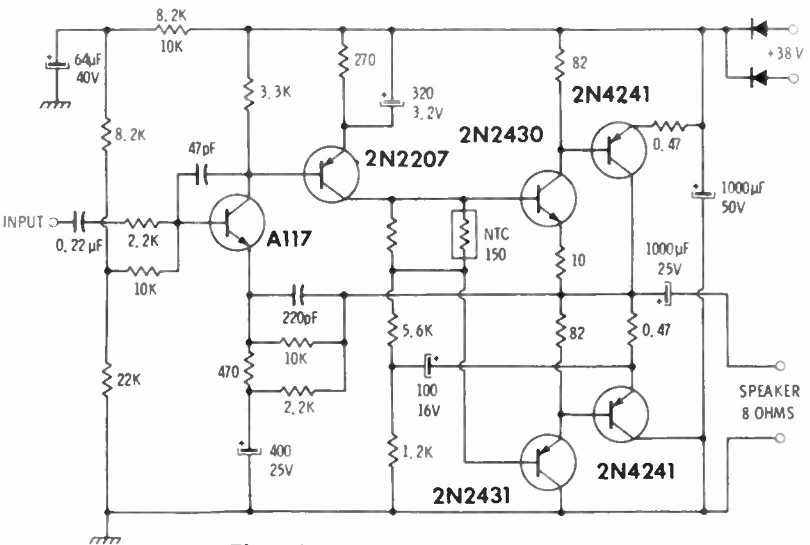


Fig. 4-6. Schematic of audio module.

complementary drivers and push-pull output transistors. It is a transformerless output stage with an output impedance of 8 ohms. This makes it convenient to use a high-power transistor output transformer as a modulation transformer for solid-state circuits. The 8-ohm secondary of such a transformer is connected to the audio module output while the transformer primary connects to the modulation input of a higher-powered solid-state transmitter. A universal transistor output transformer provides a variety of match possibilities.

To match a vacuum-tube modulated amplifier an audio power output transformer can be employed that normally matches a vacuum-tube audio power stage to an 8-ohm loudspeaker load. Again the 8-ohm secondary of the transformer is connected to the output of the audio module. The higher impedance primary winding is connected in the plate-supply voltage line to the modulated stage. This latter arrangement is used in this project to modulate the vacuum-tube rf amplifier of the previous project.

Construction

The audio module along with modulation transformers and power supply can be mounted on a single pegboard as shown in Fig. 4-7. In

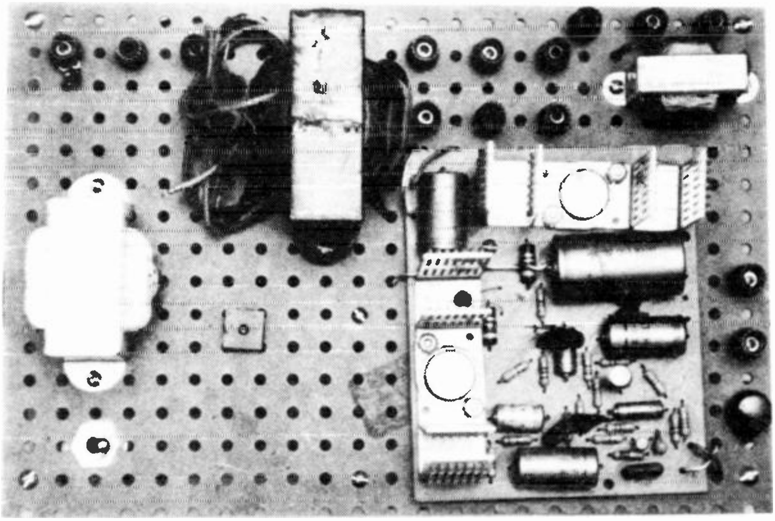
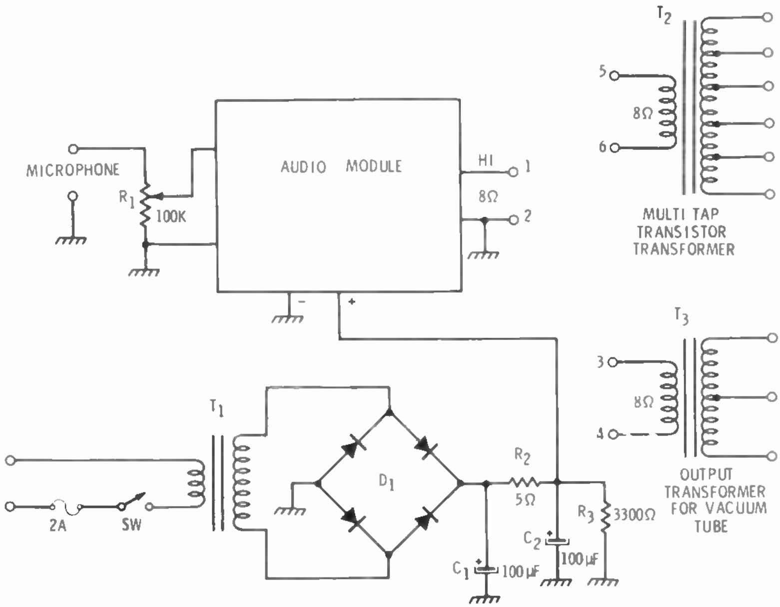


Fig. 4-7. Modulator mounted on pegboard.

the arrangement two transformers and appropriate binding posts are wired into the circuit. One transformer is a 10-watt multitap transistor output transformer; a vacuum-tube type, provides matching to the higher-impedance modulated tube amplifier. A pair of jumpers connect the desired transformer to the module output.

Some audio modules do include a gain control; many do not. In the example, a gain-control potentiometer is connected across the input of the module as shown in Fig. 4-8. A complete parts list is included.

The audio module can be operated with a supply as high as 38 volts. In the example a 24-volt filament transformer was employed.



Audio Module	Amperex PC-8-36, 20-watt	S_1	spst power toggle switch
C_1,C_2	100-μF 50V electrolytic capacitors	T_1	24-volt transistor-type power transformer
D_1	Diode bridge rectifier, 2A 200PIV	T_2	Multitap transistor output transformer, 10W (Lafayette 33-75011)
R_1	100K potentiometer	T_3	Universal tube-type output transformer, 8-ohm secondary
R_2	5-ohm 5-watt resistor		2.5A Fuse and holder
R_3	3.3K 2-watt resistor		20 Binding posts

Fig. 4-8. Universal modulator.

A bridge rectifier and resistor-capacitor filter provide adequate hum removal. Assemble the universal modulator following the schematic of Fig. 4-8.

Operation

The addition of the modulator to the circuit of Fig. 4-5 will give you a complete low-power 160 meter a-m transmitter. Excellent results can be obtained on this band with a suitable antenna. Nighttime results are often exceptional.

Connect the 8-ohm winding (binding posts 3 and 4) of the vacuum-tube audio transformer to the output of the module (Posts 1 and 2). Connect one-half of the high-impedance winding of the transformer between binding posts 5 and 6 of the class-C amplifier. Do not turn on the modulator.

Place the rf section of the transmitter in operation, feeding the appropriate dummy load. Connect a service-type oscilloscope to the output, or a modulation oscilloscope such as the Heathkit SB-610, between the output and the load.

Turn on the modulator. Speak into the microphone and notice the modulation envelope. Increase the microphone gain control until the voice peaks approach 100% modulation. With normal modulation there will be a slightly upward kick of the rf output meter with modulation. Likewise the dc plate current will just kick up very slightly on modulation peaks.

Connect your 160-meter antenna to the transmitter output. Give it a try on the air. Modulation quality and linearity are quite good. After experimentation you can mount the unit in a small cabinet and you will have a QRPP phone transmitter for 160-meter work.

The transmitter can also be operated on 80 meters using the appropriate pi-network coil. You will have to operate the microphone gain a bit lower to prevent overmodulation of the transmitter on this band.

PROJECT 19: VACUUM-TUBE CRYSTAL OSCILLATOR AND UNTUNED RF AMPLIFIER

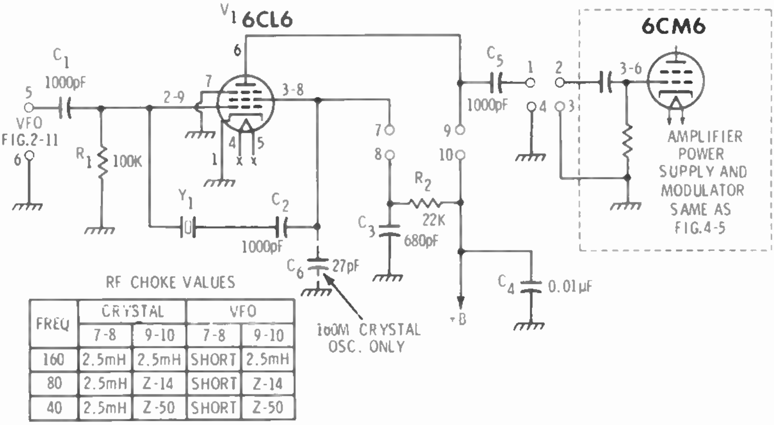
The output of the average solid-state vfo is usually not great enough to drive the 6CM6 beam-power tube directly. An alternative approach is to add an untuned rf amplifier between the vfo output and the modulated rf amplifier. This can be done without any additional tuning control.

Furthermore with the use of jumpers or a switch the added tube can be made to operate as a crystal oscillator when desired. The stage operates as an electron-coupled crystal oscillator with the screen grid serving as the plate of a Pierce oscillator.

Construction

The circuit arrangement is shown in Fig. 4-9. For crystal controlled operation the 2.5-mH choke is inserted between binding posts 7 and 8. The plate choke is inserted between binding posts 9 and 10. Its value depends on the desired band.

When a vfo signal is applied the crystal is removed from its socket and a short circuit is placed between binding posts 7 and 8.



FREQ	CRYSTAL 7-8	CRYSTAL 9-10	VFO 7-8	VFO 9-10
160	2.5mH	2.5mH	SHORT	2.5mH
80	2.5mH	Z-14	SHORT	Z-14
40	2.5mH	Z-50	SHORT	Z-50

C_1, C_2, C_5 1000-pF ceramic disc capacitors
C_3 680-pF ceramic disc capacitor
C_4 0.01-µF ceramic disc capacitor
C_6 27-pF ceramic disc capacitor
R_1 100K ½-watt resistor
R_2 22K 2-watt resistor

Miscellaneous Parts

1 pegboard (projects 16 and 17)
1 9-pin miniature tube socket
1 Crystal socket
Crystals for 40-80-160
1 2.5-mH rf choke
1 44-µH rf choke (Ohmmite Z-14)
1 7-µH rf choke (Ohmmite Z-50)
1 6CL6 vacuum tube
1 VFO

Fig. 4-9. Vacuum-tube untuned rf amplifier for 40-80-160 a-m transmitter.

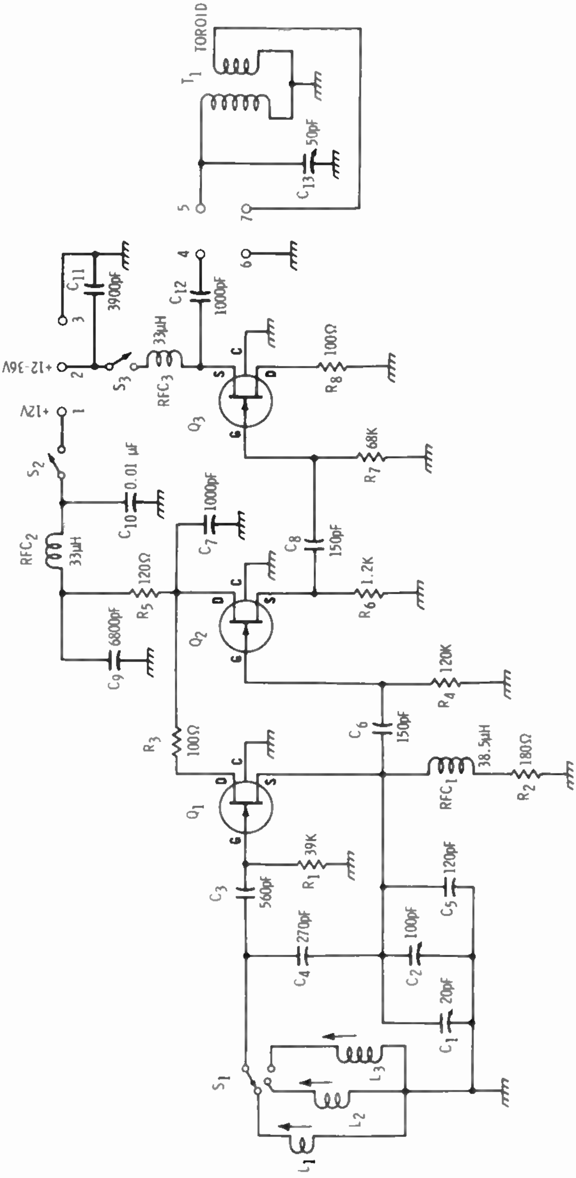


Fig. 4-10. 10-15-20 FET vfo.

Now the stage operates as an untuned pentode rf amplifier. An appropriate choke is connected between binding posts 9 and 10, depending on band of operation.

The stage is added to the pegboard after the 160-meter integrated-circuit vfo and amplifier are removed. Provide means for adding a small capacitor between screen grid and ground to ensure easy crystal starting on 160 meters.

Operation

Set up the stage for crystal operation on 80 meters. Insert the appropriate rf chokes. Turn on the unit. It is a two-control rf section; only the tune and load capacitors of the output stage need be adjusted.

Apply modulation as in Project 18. Good upward 100% modulation is obtained; any attached rf wattmeter will kick upward with modulation. Remove the crystal from its socket and place a shunt between binding posts 7 and 8. Connect the vfo to the input binding

Parts List for Fig. 4-10.

C_1	20-pF variable capacitor
C_2	100-pF variable capacitor
C_3	560-pF polystyrene-film or silver-mica capacitor
C_4	270-pf polystyrene-film or silver-mica capacitor
C_5	120-pF polystyrene-film or silver-mica capacitor
C_6,C_8	150-pF polystyrene-film or silver-mica capacitors
C_7,C_{12}	1000-pF ceramic disc capacitors
C_9	6800-pF ceramic disc capacitor
C_{10}	0.01-uF ceramic disc capacitor
C_{11}	3900-pF ceramic disc capacitor
C_{13}	50-pF variable capacitor
L_1	.088 to 0.12-μH slug-tuned inductor (J. W. Miller 20A107RB1)
L_2	0.238 to 0.39-μH slug-tuned inductor (J. W. Miller 20A337RB1)
L_3	0.735 to 0.984-μH slug-tuned inductor (J. W. Miller 20A827RB1)
Q_1,Q_2,Q_3	HEP-801 transistors
R_1	39K ½-watt resistor
R_2	180-ohm ½-watt resistor
R_3,R_8	100-ohm ½-watt resistor
R_4	120K ½-watt resistor
R_5	120-ohm ½-watt resistor
R_6	1.2K ½-watt resistor
R_7	68K ½-watt resistor
RFC_1	38.5-μH rf choke (J. W. Miller RFC-21)
RFC_2, RFC_3	33-μH rf chokes (J. W. Miller 74335A1)
S_1	Single-pole 3-position water switch
S_2,S_3	Spst toggle switches
T_1	Toroid transformer, micrometals T-50-2 core wound with No. 24 enameled wire(see text)

Miscellaneous Parts

1 Pegboard 6″ x 12″
3 Transistor sockets
12 Binding posts
2 12-volt lantern batteries

posts (5 and 6). Set the vfo to the same frequency as the crystal. Note that rf power and modulation characteristics are similar.

Repeat the above for the 40- and 160-meter bands. Be certain to use the proper rf chokes in the amplifier. You have experimented with a transmitter that uses two vacuum tubes, a bipolar vfo, and a bipolar speech amplifier and modulator.

PROJECT 20: 10-15-20 FET VFO

High-frequency stability is a major assignment in the design and construction of a low-cost and practical vfo for the high-frequency bands. Long-term frequency drift can be kept to a practical value that is satisfactory for the push-to-talk or cw QSO. Rigid mechanical construction is essential. Use good quality coils and capacitors.

Construction

A two-stage vfo and follow-up amplifier are shown in Fig. 4-10. It is a Seiler variation followed by an amplifier that can be operated tuned or untuned. Tuned operation permits a stronger output but requires another tuning adjustment. The amplifier can also be operated at a higher supply voltage to obtain even greater output.

Three separate coils that can be preset on each band are used in the oscillator circuit. A single toroid transformer serves the amplifier. For an untuned output, signal is removed between binding posts 4 and 6. If the tuned output transformer is to be employed connect a jumper between binding posts 4 and 5 and remove the output between binding posts 6 and 7. Enough output can be obtained with the tuned output circuit to drive a small vacuum-tube class-C amplifier. The toroid transformer is a Micrometals T50-2 core wound with No. 24 enameled wire. There are 15 primary turns and 7 secondary turns. Adjust the primary spacing so that bands 10, 15, and 20 can be tuned with the 50-pF variable.

For experimental work you can breadboard the vfo and amplifier. A permanent case mounting for the combination is shown in Fig. 4-11.

Operation

Turn on the two-stage vfo—not the amplifier. Use the 20-meter coil and set the variable capacitor to midposition. Set your receiver

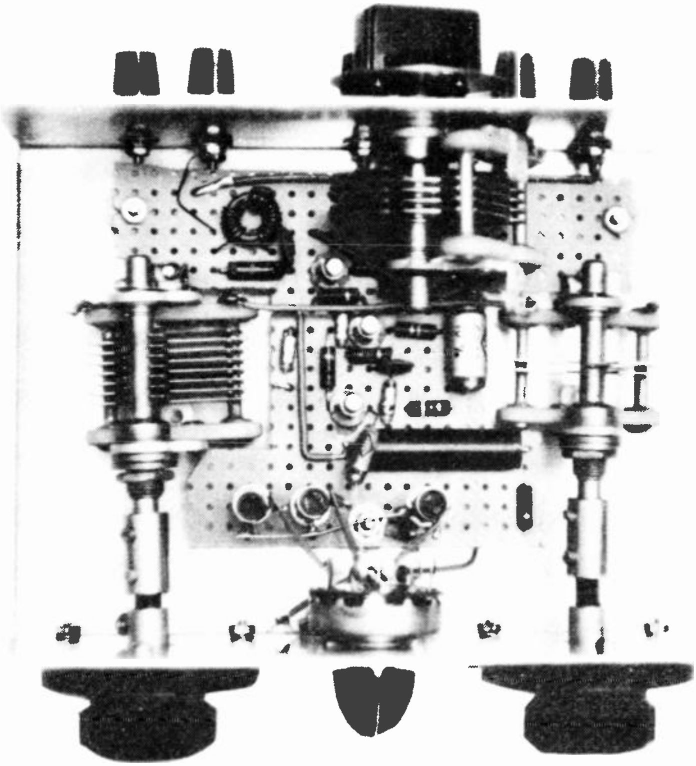


Fig. 4-11. 10-15-20 VFO

on about 14.2 MHz. Now vary the large variable until the vfo signal can be heard in the receiver. Adjust the slug in coil L_3 and the large capacitor (C_2) in such a manner that resonance is obtained at as high a maximum setting of C_2 as possible. This can be done with your receiver tuned to 14 MHz. Capacitor C_1 is used for bandspread tuning over a segment of the band.

Connect the rf indicator between binding posts 4 and 6. Turn on the amplifier. There will be only a slight output reading on the indicator meter.

Connect the jumper between binding posts 4 and 5 and the rf indicator between binding posts 6 and 7. Use a 68-ohm termination. Adjust capacitor C_{13} for maximum output.

Repeat the above steps for both 10 and 15 meters. Remember after you have the slug of the coil set for a given band you must tighten its lock-nut. Mechanical stability is especially important in operating a vfo on these higher-frequency bands.

To minimize frequency drift, ground the cases of all three HEP-801's. If you desire a very minimum drift, try several different HEP-801's in the vfo and find optimum settings for the coil slugs and capacitor C_2.

PROJECT 21: 10-160 VACUUM-TUBE RF AMPLIFIER

An all-band rf amplifier is shown in Fig. 4-12. In this arrangement the input stage is a tuned one to obtain adequate drive for the final on 10, 15, and 20 meters. Four coils cover all six bands, one of the coils being tunable to both 10 and 15 and a second to both 20 and 40 meters.

To reduce any tendency toward self-oscillation on these high-frequency bands a low value grid resistor is used in the final. A bit more driving power is needed in the interest of maintaining higher stability. Otherwise, the final rf amplifier is the same as that used in previous projects of this chapter. The final can be 100% modulated on all six bands, using the universal modulator.

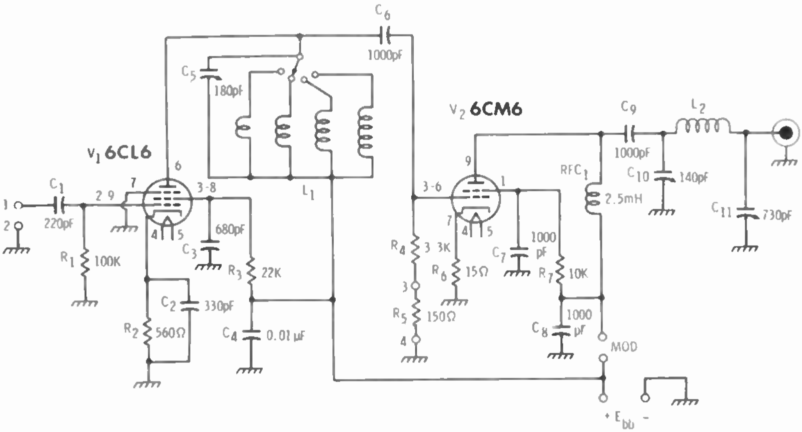
A source of drive signal can be the high-frequency vfo of Project 20 (with amplifier) or the low-frequency vfo of Project 9 (without amplifier).

Construct the amplifier on the pegboard used previously in this chapter for vacuum-tube stages. The very same power supply can be employed. The major change will be the addition of an input-stage resonant circuit.

Operation

Connect the dummy load and output indicator to the output of the final amplifier. Use the 20-meter coils. Connect the output of the high-frequency vfo to the input of the first stage. Connect a vacuum-tube voltmeter or a dc milliammeter across binding posts 3 and 4 in the grid circuit of the amplifier.

Set the vfo to an appropriate frequency in the 20-meter band using a receiver to monitor the frequency. Turn on the input stage but not the plate and screen voltages of the final amplifier. Adjust capacitor C_5 for maximum grid current. Also adjust the output capacitor of the vfo for maximum grid-current reading.



C_1	220-pF ceramic disc capacitor
C_2	330-pF ceramic disc capacitor
C_3	680-pF ceramic disc capacitor
C_4	0.01-μF ceramic disc capacitor
C_5	180-pF variable trimmer capacitor
C_6,C_7,C_8 C_9	1000-pF ceramic disc capacitors
C_{10}	140-pF variable capacitor
C_{11}	730-pF variable capacitor (2 gang, 365-pF per section)
L_1	Slug-tuned coil, 10 and 15 meters—J. W. Miller 20A337RB1; 20 and 40—J. W. Miller 21A156RB1; 80 meters—J. W. Miller 21A225RB1; 160 meters—J. W. Miller 21A825RB1
L_2	B & W miniductors 3013, 3014, 3015, 3016 (see text)
R_1	100K ½-watt resistor
R_2	560-ohm ½-watt resistor
R_3	22K 2-watt resistor
R_4	3.3K ½-watt resistor
R_5	150-ohm 5-watt resistor
R_6	15-ohm 5-watt resistor
R_7	10K 2-watt resistor
RFC_1	2.5-mH rf choke
V_1	6CL6 tube
V_2	6CM6 tube

Miscellaneous Parts

1 Pegboard 8" x 12"
2 9-pin miniature tube sockets
1 Power supply (Project 17)
1 Universal modulator (Project 18)

Fig. 4-12. 10 through 160 rf amplifier.

Apply power to the final amplifier. Adjust capacitors C_{10} and C_{11} for maximum output. Readjust capacitor C_5 for maximum output and retune C_{10} and C_{11}. Output power will be 2 to 4.5 watts. The higher output can be obtained by operating the amplifier of the vfo with 18 volts instead of 12.

Connect the modulator to the output stage. Modulate the transmitter on 20 meters. Good modulation quality can be obtained. Operate the transmitter on both 10 and 15 meters.

Disconnect the high-frequency vfo and connect the low-frequency vfo. Repeat the above steps to check operation on 40, 80, and 160 meters. Good performance can be obtained on each band. You can obtain fine results with this a-m QRP transmitter on each band.

CHAPTER 5

Basic Principles of SSB-DSB Generation

In a standard a-m modulation system, a carrier and two sidebands are transmitted. As mentioned previously, the carrier remains fixed in amplitude and carries none of the modulation, while the two sidebands carry identical modulating information. As far as conveying information is concerned one could well abandon the carrier. If this is done and both sidebands are transmitted, the method of modulation is called *double-sideband* modulation (dsb). Since both sidebands carry identical modulation, it is also possible to dispose of one of the sidebands. Thus carrier and one sideband is removed and all of the desired data can be conveyed by the single remaining sideband. This is single-sideband modulation (ssb).

The advantages of single-sideband modulation over standard amplitude modulation (a-m) are narrower bandwidth, better signal-to-noise ratio, better signal-to-interference ratio, and more economical use of available power. The last advantage facilitates the design of compact and lightweight gear for a given rf sideband output.

Two steps are involved in the generation of a dsb signal (Fig. 5-1). The carrier itself must be suppressed and the two sidebands then converted in frequency and amplified to a desired power level. Linear mixing and amplification are a necessity because the double-sideband envelope must not be distorted. Class-C amplifiers cannot be employed. Class-A, AB, and B linear amplifiers are required.

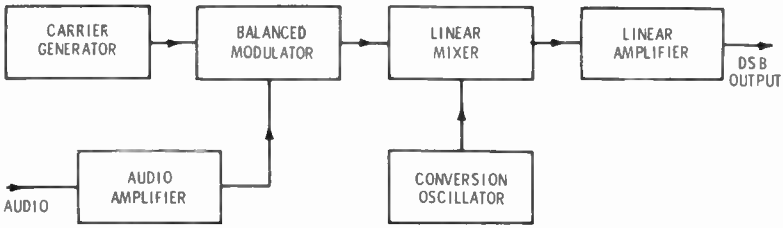


Fig. 5-1. Basic dsb transmitter.

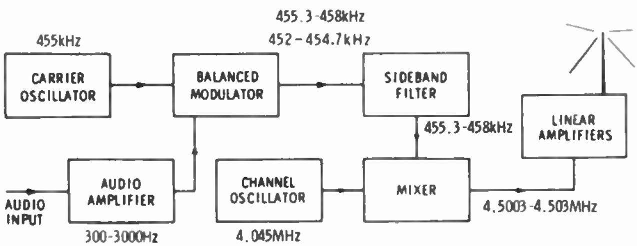


Fig. 5-2. Filter type single-sideband transmitter.

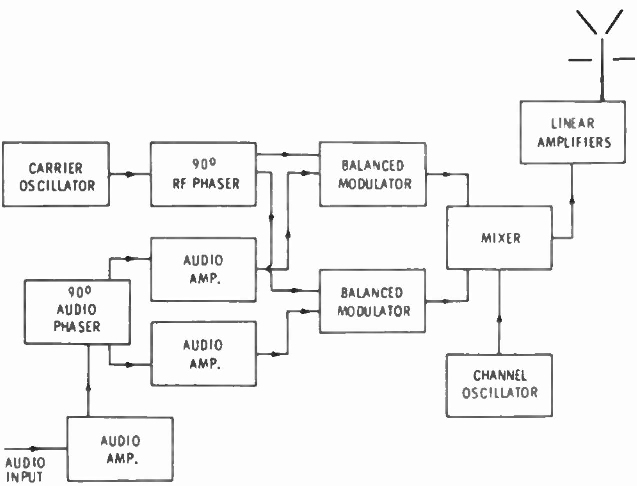


Fig. 5-3. Phasing-type single-sideband transmitter.

Three specific processes are involved in the generation of an ssb signal. The carrier itself must be suppressed, the undesired sideband removed, and the desired sideband converted in frequency and then amplified to a desired power level.

Two basic ssb generating systems are employed; these are the filter and phasing methods as shown in Figs. 5-2 and 5-3. A carrier generator, usually crystal controlled, is used in both systems. Each system employs balanced modulators to remove the carrier. A balanced modulator is fundamentally an a-m modulator with the exception that the circuit arrangement cancels out the carrier and only the two pairs of sidebands remain. A double-sideband suppressed-carrier signal appears at the output of the balanced modulator.

The second process in the single-sideband generation is removal of one sideband. It does not matter which is removed because both carry like information. Choice depends on the basic system planning and, in some instances, freedom from interference on one side or the other of the carrier signal and the possible generation of annoying heterodynes (birdies).

In the filter method of sideband removal, a carefully designed inductor-capacitor filter, mechanical filter, or crystal filter is employed. The latter type of filter is most popular in modern amateur transmitters and transceivers. Next in popularity is the mechanical filter.

The purpose of the filter is to display a high attenuation to the undesired sideband frequencies. The desired sideband frequencies are passed with a minimum of attenuation.

After the sideband signal has been formed it must be converted to the transmit frequency. Inasmuch as it carries modulation, it must be mixed and amplified by linear circuits. Frequency multipliers and class-C amplifiers are out. Linear mixers and amplifiers are in.

A crystal or vfo oscillator generates an rf signal which, on mixing with the ssb signal, produces an output frequency in the desired transmit band. In fact, the output can be either the sum or difference frequency, depending on whether the desired transmit frequency is above or below the generated sideband spectrum. An example of frequency relationships is given in Fig. 5-2.

In the phasing method of sideband generation, Fig. 5-3, one of the sidebands is removed by cancellation rather than by a sideband filter. Two balanced modulators are used. Correctly phased rf and audio signals are applied to the two balanced modulators, the outputs of which are combined. (This type of circuit is often called a double-balanced modulator.) The two sidebands generated by the two balanced modulators are identical except that they are either in phase or out of phase. For example, when the upper sideband is to be generated the two upper-sideband components in the common output circuit are in phase, while the two lower-sideband components are exactly out of phase and thus cancel each other. Oppositely, the phasing arrangement can be such that the lower sidebands appear in phase, while the two upper components are out of phase and cancel.

It is important to recognize that an identical type of signal is produced regardless of which method of generation is employed. In the filtering method the sideband transmitted depends on the frequency of the filter, while in the phasing method the sideband generated depends on phasing relationships.

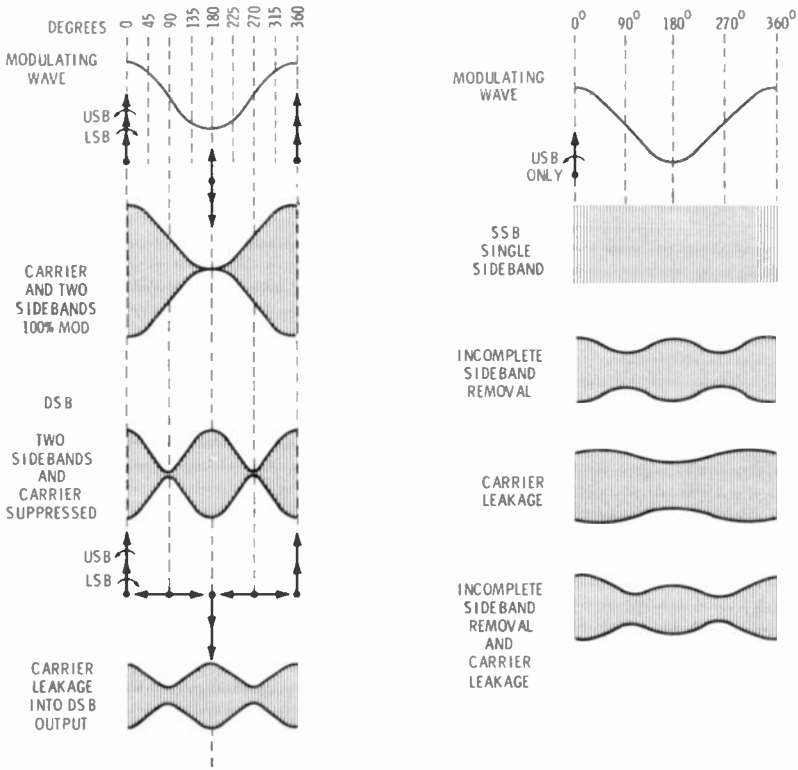
VECTOR RELATIONS

A clearer understanding of the make-up of various a-m waves can be obtained by considering Fig. 5-4. Such knowledge permits a better understanding of ssb modulation envelopes, which are important in checking and adjusting ssb equipment.

As you know, the standard a-m wave, when modulated by a single sine-wave tone, is a composite of three rf waves—a constant-amplitude carrier and two side-frequency components. At 100% modulation the two side frequencies are one-half the amplitude of the carrier.

In the vector comparison of Fig. 5-4A, all three components are in phase at the crest of the modulation envelope, and the peak amplitude builds up to twice the unmodulated value. Since the two side frequencies are not the same and also differ from the carrier frequency, the phase relationship among all three components changes throughout the modulating cycle.

Figs. 5-4A and 5-5 establish the carrier as a zero reference phase. On the positive crest of the modulating wave, the carrier and both sidebands momentarily are all in phase. Hence the resultant modulation-envelope signal rises to twice the unmodulated-carrier value. When the modulating sine wave is passing through its zero, the two side frequencies are related 90° to the carrier and 180° to each other. Therefore, the side frequencies cancel and the net amplitude of the amplitude-modulated wave is of the same value as its carrier.



(A) Double sideband. (B) Single sideband.

Fig. 5-4. Typical double- and single-sideband waveforms.

On the negative peak of the modulating wave, the two side frequencies are again in phase with each other but 180° out of phase with the carrier. Now there is complete cancellation and the net output drops to the zero point of the modulation envelope.

In summary, it is the changing phase relationship among the three components that causes the amplitude variations in their resultant amplitude-modulated envelope. Of course, the intermediate levels of the modulation envelope also depend on the instantaneous phase relationship among the three waves. Actually, as shown in Figs. 5-4 and 5-5, the side frequencies, in following the amplitude changes of the modulating wave, can be considered to rotate vectorially around the zero reference phase set by the carrier. They rotate in opposite

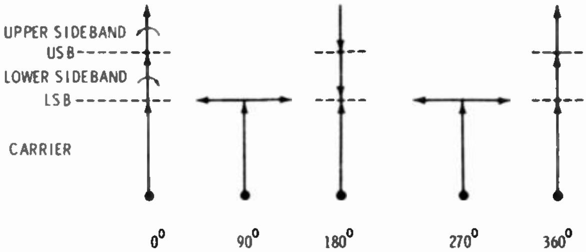


Fig. 5-5. Vector relationships for standard a-m wave through one cycle of modulating wave.

directions, one being above the carrier frequency and one being below the carrier frequency.

Let us next consider what type of a resultant modulation envelope is obtained when the carrier is suppressed and the two side frequencies remain. As shown in Fig. 5-4A, the side-frequency components are out of phase relative with each other. Even though the carrier magnitude is zero, the side frequencies can be considered to revolve in phase with respect to the carrier reference phase of zero. As the

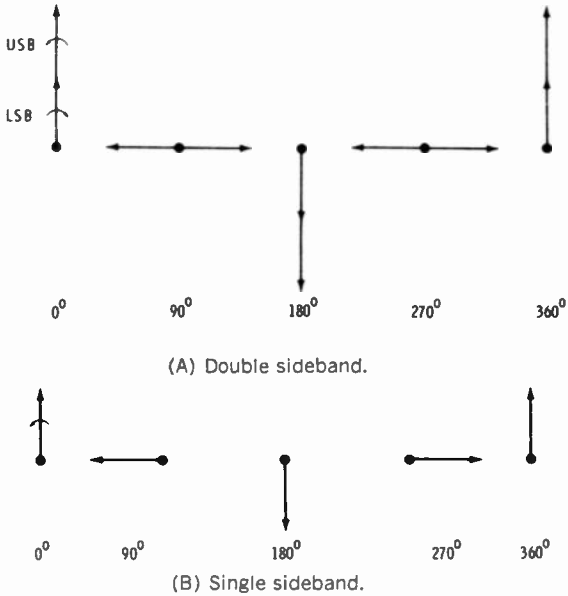


Fig. 5-6. Vector relationships for dsb and ssb components.

modulating wave goes through its cycle, the side frequencies at times become out of phase and exactly in phase during the modulating wave period. Actually, the modulation envelope goes through two cycles during the one cycle of the modulating wave. Vector magnitudes are shown in Fig. 5-6A. The peak amplitude of the envelope rises to twice the individual sideband magnitude. It falls to zero when the two vectors are exactly out of phase.

Figs. 5-4B and 5-6B show the magnitudes and vector relationships for a single-sideband signal. Now there is only one rotating vector. In fact, only one rf component is generated as the modulating wave goes through its cycle. Hence a constant-amplitude resultant is produced.

The dsb and ssb modulation envelopes are important in judging the performance of a sideband system. The nature of any distortion on these envelopes points up various characteristic defects and maladjustments.

POWER RELATIONSHIPS

In addition to the reduced bandwidth, compared with that of a double-sideband system, a single-sideband system also conserves power. A mathematical example will prove this saving. A standard a-m transmitter with a rated carrier of 1000 watts radiates 250 watts per sideband with 100% modulation. The carrier power is 1000 watts and the total sideband power is 500 watts. Hence the average transmitter power rating is 1500 watts. Peak power is of course four times the carrier value, or 4000 watts, because the rf output voltage is twice the unmodulated-carrier voltage during 100% modulation peaks. Nevertheless the sideband power output is only 250 watts.

In generating a 250-watt sideband using the single-sideband modulation method, the peak power rating of the transmitter need only be 250 watts. For the same sideband power output, this represents a substantial saving in the size, weight, and rating of both radio-frequency and power components. Thus the single-sideband technique is particularly adaptable to the design of mobile and other compact equipment. Furthermore, the ssb system lends itself to transceiver construction, with some of the circuits and components, such as the sideband filter and various oscillator stages, used interchangeably for transmission and reception.

It is customary to rate single-sideband transmitters in terms of peak envelope power (PEP). The PEP of a single-sideband transmitter is defined as the rms power developed at the modulation crest. The ssb transmitter described previously would then have a PEP rating of 250 watts. At lower levels of modulation, of course, the power output would be less.

It has become standard to make power comparisons between standard a-m and ssb transmitters on the basis of signal-to-noise ratio. In a single-sideband transmitter with a PEP of 500 watts, the signal produced at the receiver output has the same signal-to-noise ratio as a standard a-m transmitter with a 1000-watt carrier output. Under the above condition, the single-sideband output would be 500 watts and the total power in both sidebands of the standard a-m transmitter would also be 500 watts.

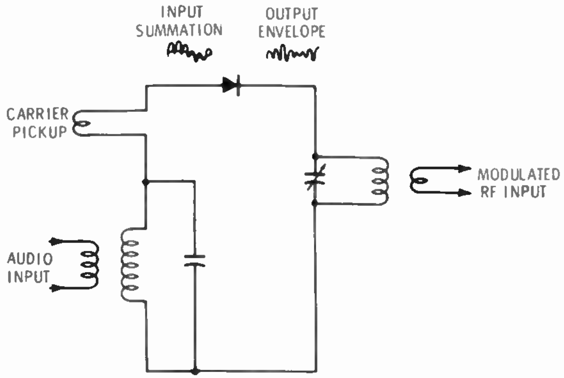
Some transmitters are designed for both ssb and standard a-m. Their ratings are usually given in terms of PEP for sideband and carrier power output for standard a-m. Usually the rated output for standard a-m falls between one-third to one-half the permissible PEP rating of the same transmitter.

CARRIER SUPPRESSION

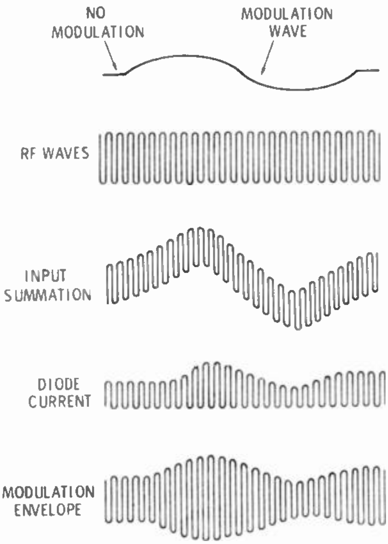
The first step in the generation of a single-sideband signal is the simultaneous formation of sidebands and the removal of the carrier. Diodes, grid-type vacuum tubes, and transistors can be used in balanced-modulator circuits. Considerable help in understanding the operation of a balanced modulator can be gained by first reviewing the operation of a diode as a modulator. The diode is used frequently as an a-m detector or demodulator. In this application the modulated signal is applied to its input. By rectifying the radio-frequency cycles, the lower-frequency modulating wave can be recovered in its output circuit.

A diode can be made to operate in the opposite fashion; that is, a weak audio signal and a stronger radio-frequency signal can be supplied to its input. The same rectifier action (nonlinear operation) causes sum and difference frequency components to be developed in the output circuit. These are the upper and lower side frequencies.

The waveforms of Fig. 5-7 demonstrate the mixing or modulation process. When the radio-frequency signal alone is applied to the



(A) Modulator circuit.



(B) Diode modulator circuit waveforms.

Fig. 5-7. The diode modulator.

diode modulator, diode current flows during positive alternations of the input cycle. Hence positive bursts of current flow through the load. Inasmuch as the load is usually a resonant circuit, a sine wave is reconstructed in the output circuit. This action is similar to the influence of the resonant circuit in the output of a class-B or class-C rf amplifier.

When an audio wave is applied to the input, along with the radio-frequency wave, the combined appearance of the two components (algebraic summation) appears as shown in Fig. 5-7B. The peak amplitudes of the diode-current bursts now depend on the net diode plate voltage at the crest of each radio-frequency cycle. This peak voltage varies up and down with the modulating wave. As a result, the peak diode current varies correspondingly; this peak diode current change is comparable to the peak plate-current variation of a modulated class-C amplifier. Again the resonant circuit, because of its energy-storing ability, reconstructs the negative alternation of the output variation, forming the familiar amplitude-modulated envelope.

The input wave is the simple combining of two separate signals—a high-frequency radio wave and a low-frequency audio wave. However, the output wave results from nonlinear mixing or heterodyning and is composed of three radio-frequency components—the carrier plus two side frequencies. The audio wave is filtered out by low impedance of the output circuit.

Diode modulation has poor efficiency. However, at a low signal level, combinations of diodes in a number of circuit arrangements can serve admirably as stable balanced modulators. Diodes have a low order of modulation linearity. Nevertheless by using a modulating wave of much lower amplitude than the radio-frequency wave (low-percentage modulation), a linear output can be obtained (modulation-envelope variation corresponds faithfully to the modulating wave).

Now that you understand that a diode can be used as a modulator, let us consider how two or more diodes can serve as modulators, at the same time they suppress the radio-frequency carrier.

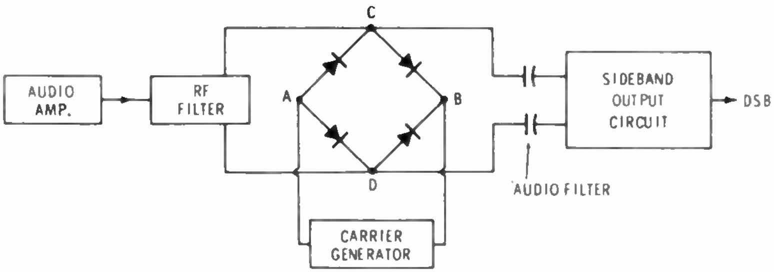


Fig. 5-8. Bridge balanced modulator.

Two basic and common diode balanced modulators are the bridge and ring circuits shown in Figs. 5-8 and 5-9, respectively. Since

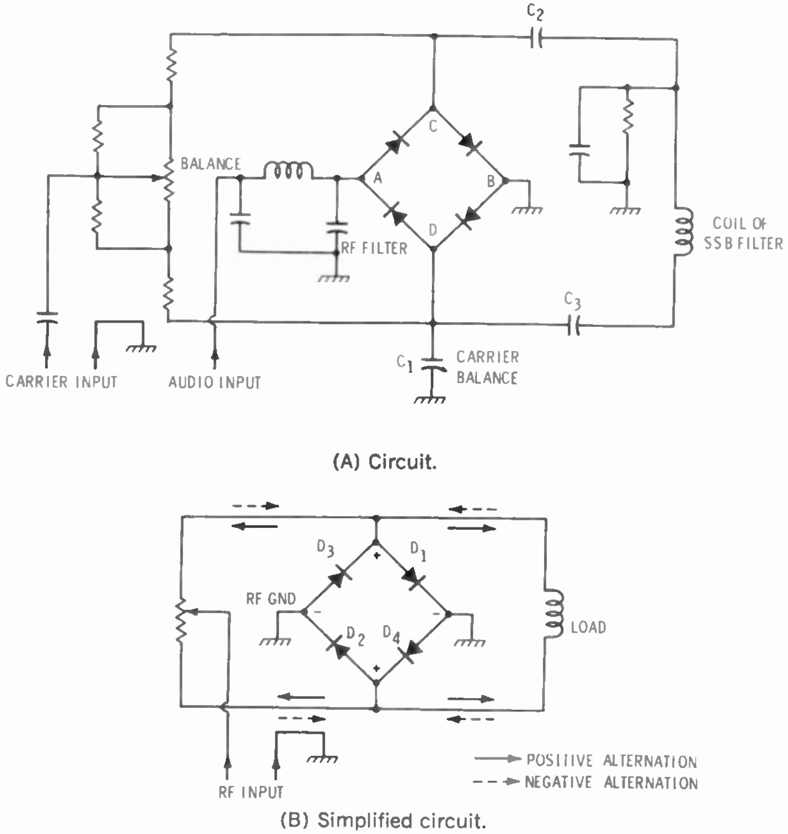


Fig. 5-9. A ring balanced modulator.

radio-frequency and audio waves are supplied to the various diodes, modulation takes place. Thus, carrier and sideband currents flow. However, the carrier-current components cancel with relation to the output systems.

Let us consider the operation of the bridge circuit when only the radio-frequency carrier is applied. When the radio-frequency cycle swings on its negative alternation, point A on the bridge is made negative with respect to point B. Consequently the four diodes are back-biased and no current flows. If the four diodes have the same reverse resistance, the bridge will be balanced and there will be no difference of potential between points C and D. Consequently no current flows in the output circuit.

When the radio-frequency wave swings positive, all four diodes become forward-biased. Hence their resistances are low and current will flow. However, if they all have the same resistance, the bridge will remain balanced. Therefore, no difference of potential will develop between points C and D, and again no carrier current is in the output circuit.

It is apparent that neither the positive nor the negative alternations of the radio-frequency wave will cause an output voltage, and so the carrier frequency is effectively suppressed. However, the radio-frequency wave does appear across each of the four diodes, and rectifier action takes place—the diode currents being switched on and off by the alternations of the radio-frequency wave.

Let us next consider the diode activities when a modulating wave is applied along with the radio-frequency carrier. Inasmuch as the audio wave is made substantially lower in amplitude than the radio-frequency alternation of the radio-frequency carrier. The amount of current flow, as in the simple diode modulator, depends on the instantaneous sum of the radio-frequency and audio-frequency waves combined. As a result the peak radio-frequency current will vary with the modulating wave, as it does in any form of amplitude modulation.

Of equal importance is the influence of audio voltage on the bridge. This voltage is applied between points C and D. As a result the bridge is unbalanced, the degree of unbalance depending on the magnitude of the audio signal. The unbalance permits side-frequency currents to flow in the output circuit. It is apparent, then, that the bridge—although balanced at the carrier frequency—is unbalanced at off-carrier frequencies. As a result, the sideband frequencies develop in the output circuit. Audio-frequency components are blocked from the output circuit by the small series capacitors and the low shunt impedance of the output tank circuit at audio frequencies.
output of other circuits. As shown in the simplified version, radio-

The ring balanced modulator (Fig. 5-9) delivers up to twice the frequency carrier current would flow in the output circuit, except that the two paths are opposing. As a result, the currents cancel and produce no output.

On the positive alternation, diodes D_1 and D_2 conduct. However the current path is down through the output coil for diode D_1 and up through the output coil for diode D_2. If the ring circuit is balanced exactly, no carrier current will flow in the output circuit. During the negative alternation of the carrier, diodes D_3 and D_4 will conduct. The carrier currents that would tend to flow in the output circuit are again equal and opposite. The rf carrier has been suppressed.

The presence of the audio wave unbalances diode operation. As mentioned previously, the carrier wave does appear across each diode, and the application of audio causes a heterodyning activity. The side frequencies are produced and developed across the output because of the unbalance which the audio wave introduces into the ring circuit. In the ring balanced modulator, side frequencies are developed during both alternations of the rf wave. This is unlike the bridge circuit described previously, where current flow coincided only with the positive alternations. In diode balanced-modulator circuits, carrier suppression of 40 dB and higher can be attained by careful design and adjustment.

In the circuit of Fig. 5-9A the carrier is applied between points C, D, and ground, while the audio signal from the low-impedance cathode follower is applied between points A and B. The radio-frequency carrier is introduced through a resistor network which includes a carrier balance control. Any slight performance differentials in the ring circuit can be compensated for with this control. Thus it is possible to adjust the circuit for a very minimum of carrier output. Stray and differential capacitances may also disturb the ring balance. Consequently, an additional carrier balance control is included in the form of an adjustable capacitor (C_1).

The audio wave is introduced via a low-pass filter which blocks the radio-frequency components from entering the audio system. In the output circuit, capacitors C_2 and C_3 have very high reactance at audio frequencies. Consequently the audio components are blocked from the follow-up sideband filter.

A very simple and popular modulator is the two-diode arrangement (Fig. 5-10). It is also a bridge affair with two equal-value resistors replacing two of the bridge diodes. The potentiometer serves as a balance control on the rf input side. Audio is applied across the other two points of the bridge. Two diode types can give as much as 40 dB carrier suppression.

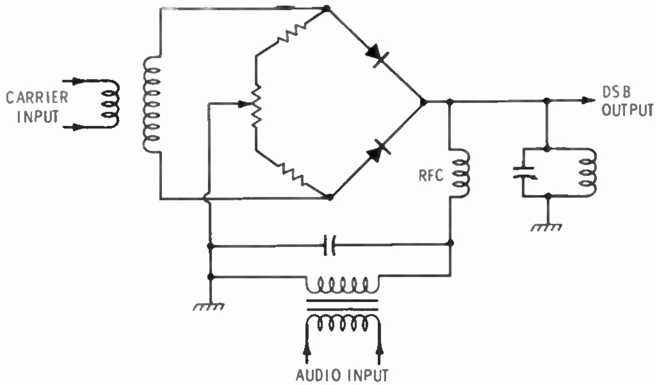


Fig. 5-10. Two-diode balanced modulator.

TRANSISTOR AND TUBE BALANCED MODULATORS

Bipolar transistors can be used in balanced modulator circuits. Most popular are the bipolar configurations associated with integrated circuits. In the design and manufacture of an integrated circuit it is possible to obtain almost perfectly balanced combinations. This is a very attractive advantage in terms of carrier suppression. Two of the projects of Chapter 7 employ integrated circuits as balanced modulators.

The dual field-effect transistor with its extraordinary balance provides a simple circuit with high carrier suppression. A basic dual-

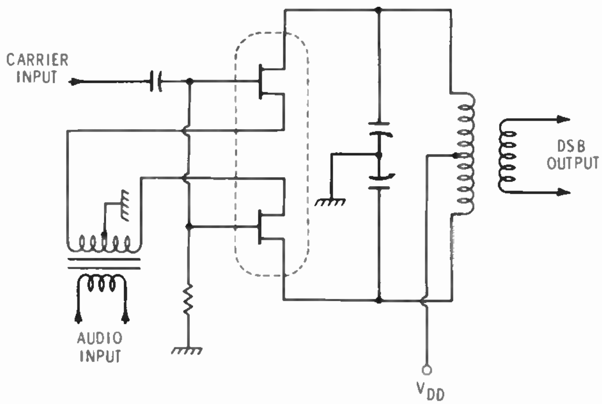


Fig. 5-11. Dual-FET balanced modulator.

FET balanced modulator is shown in Fig. 5-11. The two gates are fed in parallel with carrier signal. The drain output circuit is in push-pull, producing carrier cancellation. Audio is applied in push-pull to the low-impedance source circuit. Both FET's are amplitude-modulated with side frequencies developing in the output.

A similar circuit using two pentode vacuum tubes is shown in Fig. 5-12. Carrier signal is applied to the two grids in parallel. Plates are connected in push-pull to obtain carrier cancellation. Audio is

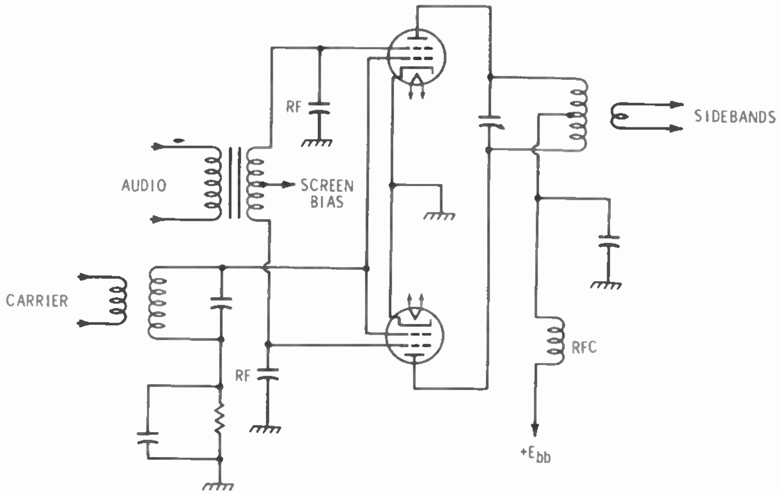


Fig. 5-12. Pentode balanced modulator.

applied to the screen grids in push-pull. Screen-grid modulation results and side frequencies are developed in the output, while the carrier is cancelled.

BEAM-DEFLECTION MODULATOR

A beam-deflection tube functions as an excellent balanced modulator capable of 60 dB of carrier suppression. In this tube (Fig. 5-13), a beam of electrons is directed outward toward a pair of plates. These plates are angled away from the straight-line direction and include deflection electrodes. If there is no difference of potential between the deflection electrodes, the concentrated beam of electrons moves in a straight line and does not strike the plates. How-

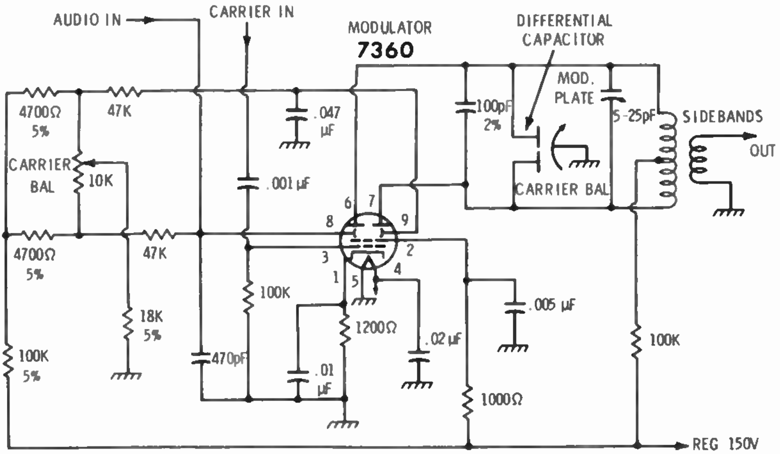


Fig. 5-13. Beam-deflection modulator.

ever, if a difference of potential exists across the deflection electrodes, the beam will be bent in the direction of the more positive potential. Thus the beam will be guided toward the plate associated with the positive deflection electrode. The beam can be directed from one plate to the other by changing the polarity of the deflection voltage. The portion of the beam current collected by each plate depends on the magnitude of the potential difference between the focusing electrodes. This difference is made to vary with the applied audio.

In the double-sideband generator shown in Fig. 5-13, the radio-frequency carrier is supplied to the control grid. If the difference of potential between focusing electrodes is maintained at zero, no plate current flows because the electron beam will not be pulled toward either plate, despite the presence of radio-frequency current. Thus no carrier current flows in the output circuit.

An equal amount of dc voltage is supplied to the two deflection electrodes. The audio signal is applied to only one deflection electrode (pin 8). Thus this electrode will vary plus and minus with respect to the electrode at pin 9 and the beam will change over between the two plates in accordance with the audio variation. Modulation takes place because the electron beam carries a radio-frequency current variation. Thus sideband currents flow in the output circuit.

A carrier balance control, associated with the beam deflection electrode, equalizes the relative influence of the two tube segments on the electron beam. A capacitive balance control (in the form of a split-stator differential capacitor) is connected across the output tank circuit.

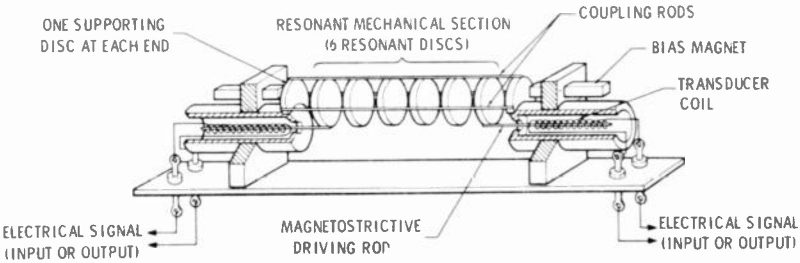
SIDEBAND REMOVAL

In the filter-type ssb generator, the filter follows the balanced modulator. The sideband filter removes the undesired sideband. This can occur at low, medium, or high frequencies. Low-frequency filters operate in the range between 50 and 60 kHz. Some have been used as low as 20 kHz. In telephone services, even lower-frequency sideband filters are employed. The medium-frequency filter usually operates in the 450- to 470-kHz spectrum; many do so at 455 kHz. High-frequency filters generally operate in the 9-MHz region, although some have been used at frequencies up to and beyond 60 MHz.

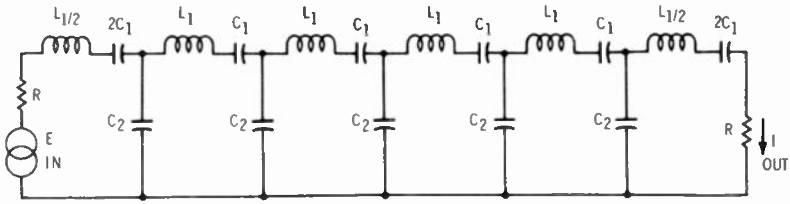
Lumped-constant LC filters using stable capacitors and high-Q inductors are employed for low-frequencies. Such filters usually consist of multisection bandpass units composed of a group of series- and parallel-resonant circuits. They are precision made and are usually mounted in sealed enclosures. A typical low-frequency filter may have a bandpass that positions the carrier at 50 kHz. It will then pass a range of frequencies that extend from approximately 47 kHz, up to 50 kHz. Above 50 kHz the frequency response will drop off very abruptly, making certain the upper sideband is attenuated sharply. Below 47 kHz the response will drop off more gradually, being almost 50 dB down at 45 kHz. These are seldom used in amateur gear.

In modern single-sideband equipment, a common filter is the medium-frequency mechanical filter. In the mechanical filter, stable resonant conditions establish a most favorable bandpass characteristic because the mechanical vibrations are entirely a function of the mechanical dimensions of the filter elements.

Such a mechanical filter and its electrical equivalent are shown in Fig. 5-14. Its response and a typical circuit diagram appear in Fig. 5-15. In addition to its high stability, the mechanical filter has an extremely high Q. Thus a filter with a flat bandpass and, at the same



(A) Filter construction.



(B) Equivalent circuit.

Fig. 5-14. Collins mechanical filter.

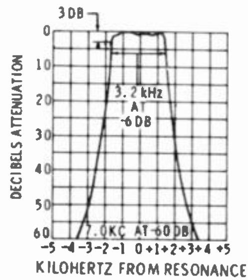
time, very steep skirts is possible. This is, of course, an ideal condition for a single-sideband filter if the desired sideband is to be passed with minimum loss, and the undesired sideband is to be attenuated sharply.

The mechanical filter takes signals of the passband to which it is resonant and transfers them, with minimum attenuation, from one vibrating disc to another (input to output). Any signal applied off the passband is incapable of setting the transducer element and succeeding discs into vibration. Thus such a signal is not conveyed from input to output.

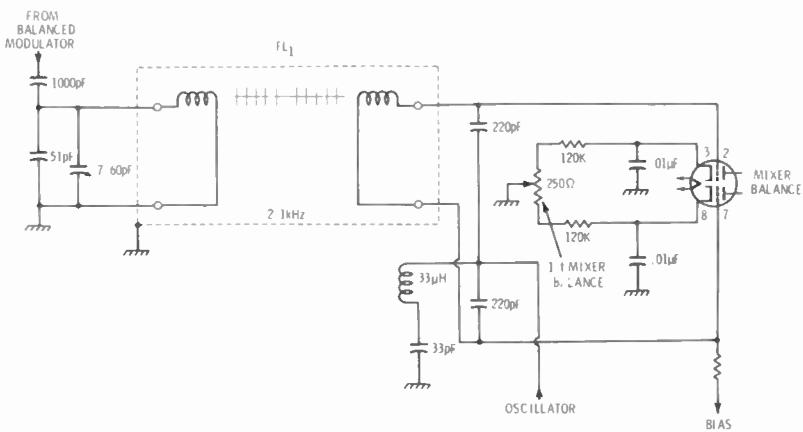
One of the important characteristics of a sideband filter pointed up in the characteristic curve of Fig. 5-15 is bandwidth. The curve indicates that the bandwidth of the filter is 3.2 kHz between −6 dB points on each side of the preferred bandpass. The peak-to-valley ratio is 3 dB (this is the ratio of maximum to minimum response over the desired bandpass). The insertion loss refers to the loss the signal encounters in being transferred from input to output; losses vary from 2 to 16 dB, depending on design and frequency.

Skirt selectivity is given in terms of shape factor. This is the ratio between the bandpass at −60 dB below peak over the bandpass at

−6 dB below peak. In the example, the shape factor is about 2.2 (7/3.2).



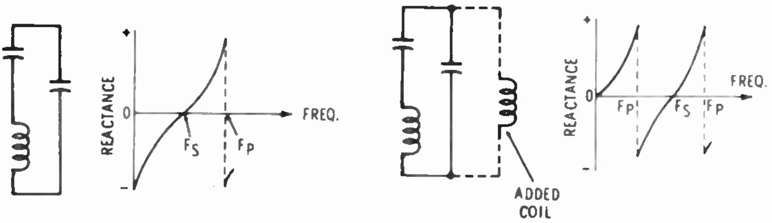
(A) The bandpass response.



(B) Typical filter circuit.

Fig. 5-15. Collins mechanical filter response and circuit.

Crystal filters in various configurations are used in single-sideband equipment. A piezoelectric crystal has the characteristics of a resonant circuit. Fig. 5-16A shows the equivalent circuit of a crystal and its resonant response. A crystal has both a parallel- and series-resonant frequency; at series resonance, the impedance of the crystal is minimum. This is indicated by frequency F_s on the zero-reactance line (in a series-resonant circuit, the net reactance falls to zero). At a somewhat higher frequency the capacitive and inductive reactances become high and equal. This is the parallel or antiresonant frequency F_p. In practice, the two resonant frequencies are several hundred cycles apart.



(A) Parallel (F_p) and series (F_s) response reactances.

(B) Effect of a coil added in parallel with crystal.

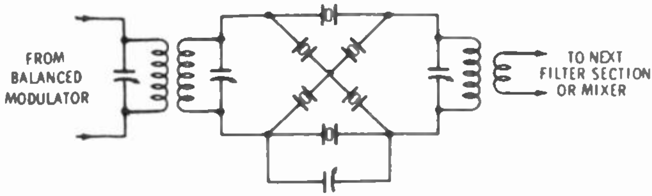
Fig. 5-16. Crystal circuit equivalents and response.

The spread between the two resonant frequencies can be controlled by insertion of an additional inductor. This inductance will also introduce a second parallel-resonant condition, as shown in Fig. 5-16B. This point can also be used to advantage in establishing a desired bandwidth and response. Actually the external inductances can be coils associated with the resonant circuit at the sideband filter input and output.

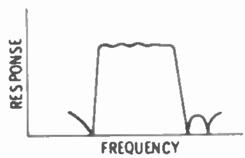
Since a crystal displays both parallel and series resonances, by careful choice of several crystals with regard to their resonance properties, a very effective sideband filter can be constructed. A practical filter can consist of as few as two crystals. More elaborate filters, with more rigid response requirements, can have as many as eight or more crystals.

A practical type of crystal sideband filter is shown in Fig. 5-17, along with a typical bandpass response curve. The series-connected crystals are set to one frequency, and the shunt pair 2 or 3 kilocycles away. By careful selection and adjustment, the series-resonant frequency of one set of crystals can be made to correspond to the parallel-resonant frequency of the other pair. Thus the crystals can be made to display a minimum series opposition to the transfer of frequencies in the desired bandpass and a maximum shunt impedance. Oppositely, off the desired bandpass the series-connected crystals will display a high series impedance, and the shunt-connected units a minimum shunt impedance. Therefore there will be maximum attenuation and a sharp skirt selectivity at these frequencies.

So-called "half-lattice crystal filters" are shown in Fig. 5-18. They are used for less critical sideband applications. As in the full-lattice filter, the bandwidth is determined by the separation between

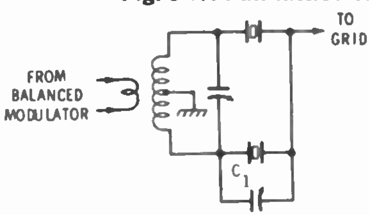


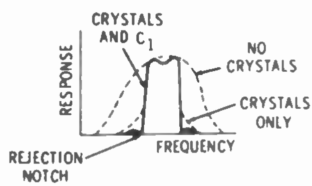
(A) Circuit.



(B) Response.

Fig. 5-17. Full-lattice crystal filter and response.





(A) Circuit. (B) Response.

Fig. 5-18. Half-lattice filter and response.

the two crystal frequencies, and is approximately twice the frequency separation between the series- and parallel-resonant frequencies of these crystals.

In the half-lattice arrangement the two crystals sharpen the regular response curve of the associated resonant circuit. Thus they permit a sharp skirt response.

The addition of a tunable capacitor across one of the crystals introduces still another resonant condition; two rejection notches are formed as shown. Additional half-lattice combinations can be used in cascade to further improve the bandpass response.

PHASING METHOD

The phasing method of sideband generation suppresses the carrier and removes one sideband without using a sideband filter. Two balanced modulators, an audio phase shifter, and a radio-frequency phase shifter are required. A functional block diagram is given in Fig. 5-19.

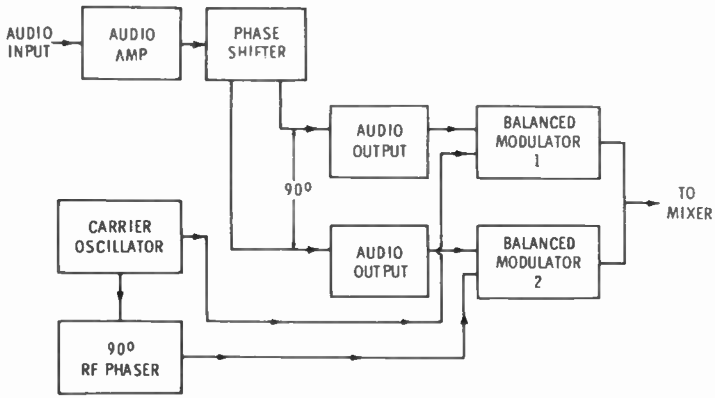


Fig. 5-19. Phasing method of sideband generation.

In the vector analysis of a standard amplitude-modulated signal, the phase of each side frequency can be considered to rotate with relation to the reference zero phase of the carrier. Since there is one side-frequency component above and one below the carrier frequency equivalent to the amount of the modulating frequency, the sideband vectors rotate in opposite directions. This is shown in Fig. 5-20.

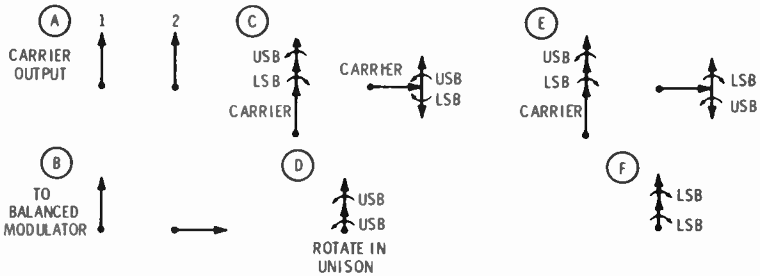


Fig. 5-20. Vector relations for phasing method of sideband generation.

In the phasing method of ssb generation, two separate double-sideband signals with suppressed carriers are generated. One pair of sidebands occurs in phase in the output circuit, and the opposite pair cancels. Two balanced modulators are used, driven by two radio-frequency carriers and two modulating waves of the proper relative phases.

Notice that the two carrier components are in phase as they leave the carrier generator. One component, however, passes through a 90° phase shifter. As a result, the two balanced modulators are excited by 90°-related carrier components. This is shown in Fig. 5-20B.

The modulating wave is also supplied to a phase-shift network. At the network output there are two audio waves exactly 90° out of phase. These supply audio signal to the separate balanced modulators. The audio phase-shift network must be designed to maintain an exact 90° shift over the desired audio band. Lower and higher frequencies should be removed before the audio signal is applied to the phase-shift network. In this way, less distortion and fewer spurious signal components are introduced in the modulation process.

The above phasing relations will always keep one pair of sidebands in phase and the opposite pair out of phase. Vector diagram C (Fig. 5-20) shows the instantaneous angles obtained when the carrier and two sidebands of balanced modulator 1 (Fig. 5-19) are in phase. At the very same instant in balanced modulator 2, the two sidebands are out of phase and 90°-related to the carrier. Note that at this particular instant, the two upper sidebands are in phase but the two lower sidebands are exactly out of phase. When the two combine in the common output circuit, the upper sidebands add and the lower sidebands cancel. Furthermore, each balanced modulator removes its own carrier. Therefore the only output is the two upper sidebands (vector D).

One important relationship should be understood from vector C. Both upper-sideband components are the same frequency. Thus they will rotate in phase and remain in phase, regardless of the instantaneous amplitude of the modulating wave. Likewise the two lower-sideband components are of the same frequency, and their vectors will continue to rotate in and out of phase relationship throughout the modulating cycle. In other words, vector C shows only one instantaneous relationship, although the in-phase upper-sideband and out-of-phase lower-sideband relationships are maintained throughout the cycle.

Vectors E and F of Fig. 5-20 shows instantaneous relationships after the phase of one of the modulating audio waves has been shifted

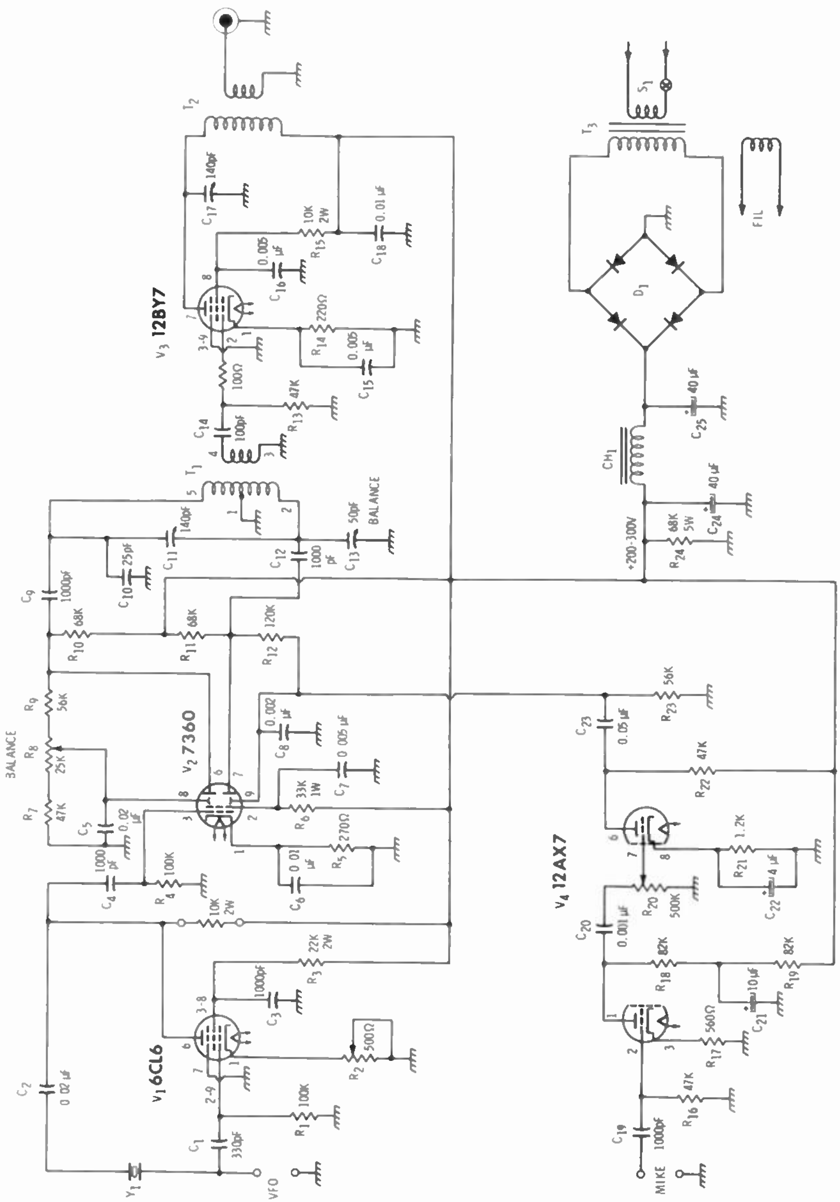


Fig. 5-21. 40-, 80-, and 160-meter dsb generator.

through 180°. The vector relationships in balanced modulator 1 remain unchanged. However, those in balanced modulator 2 have been rotated through 180° with respect to the sidebands. Now it is the two lower sidebands that are in phase, and the two upper sidebands that are out of phase. The lower sideband is the one developed in the common output circuit. Thus switching from upper- to lower-sideband operation is a simple procedure.

PROJECT 22: 40-80-160 DSB GENERATOR

A four-tube double-sideband generator is shown in Fig. 5-21. It functions well as a QRP double-sideband transmitter, delivering about a one-watt peak envelope power output. It can be constructed with ease on a 12" x 2" x 7" chassis.

Parts List for Fig. 5-21.

C_1	330-pF disc capacitor
C_2,C_5	0.02-μF disc capacitors
$C_3,C_4,C_9,C_{12},C_{19},C_{20}$	1000-pF disc capacitors
C_6,C_{18}	0.01-μF disc capacitors
C_7,C_{15},C_{16}	0.005-μF disc capacitors
C_8	0.002-μF disc capacitor
C_{10}	25-pF disc capacitor
C_{11},C_{17}	140-pF variable capacitors
C_{13}	50-pF trimmer capacitor
C_{14}	100 pF disc capacitor
C_{21}	10-μF electrolytic capacitor, 450V
C_{22}	4-μF electrolytic capacitor, 450V
C_{23}	0.05 paper or disc capacitor
C_{24},C_{25}	40-μF electrolytic capacitors
CH_1	5 to 7H, 50 to 100mA filter choke
D_1	Diode bridge rectifier, 2A 800PIV
R_1	100K 1-watt resistor
R_2	500-ohm 2-watt potentiometer
R_3	22K 2-watt resistor
R_4	100K 1-watt resistor
R_5	270-ohm 1-watt resistor
R_6	33K 1-watt resistor
R_7,R_{13},R_{16}	47K 1-watt resistors
R_8	25K potentiometer
R_9,R_{23}	56K 1-watt resistors
R_{10},R_{11}	68K 1-watt resistors
R_{12}	120K 1-watt resistor
R_{14}	220-ohm 1-watt resistor
R_{15}	10K 2-watt resistor
R_{17}	560-ohm 1-watt resistor
R_{18},R_{19}	82K 1-watt resistors
R_{20}	500K potentiometer
R_{21}	1.2K 1-watt resistor
R_{22}	47K 2-watt resistor
R_{24}	68K 5-watt resistor
S_1	Spst power switch
T_1	Set of balanced coils (Fig. 5-22)
T_2	Set of output coils (Project 7)
T_3	Power transformer, 240V sec., 6.3V fil.
V_1	6CL6 tube
V_2	7360 beam-deflection tube
V_3	12BY7 tube
V_4	12AX7 tube

Miscellaneous Parts

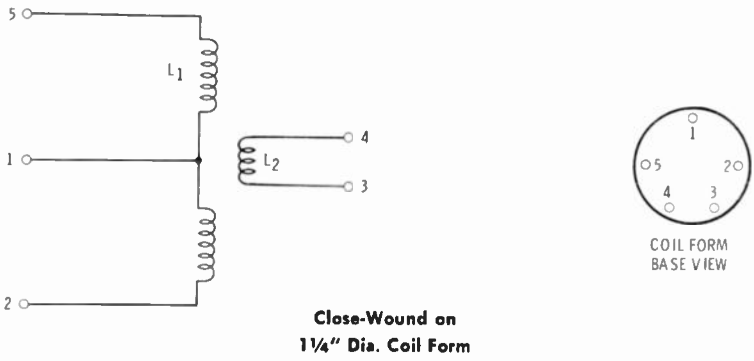
Chassis, 11" x 7" x 2"
4 9-pin tube sockets
2 5-pin tube sockets (coil sockets)
1 crystal socket
1 Ac line cord
Binding posts

Circuit Description

The balanced modulator is a 7360 beam-deflection type. Audio is applied to the deflector plates and carrier to the control grid. Its double-sideband output is increased in level by a 12BY7 pentode amplifier. Output level is such that a weak but usable pattern can be observed on a monitor scope such as the Heath SB-610.

A 6CL6 pentode is the source of the carrier frequency. When a crystal is inserted it operates as a crystal oscillator. With the crystal removed the stage can be driven by an external vfo. Adequate drive can be obtained from the vfo covered in Project 9.

The balanced modulator includes two balance controls: the 25K potentiometer in the deflector-plate circuit, and a balance variable capacitor in the output circuit. The coils of rf transformers T_1 and T_2 are wound on plug-in coil forms to permit multiband operation. The coil data given in Project 7 is appropriate for transformer T_2. The coil construction of transformer T_1 must be balanced and the push-pull coil data of Fig. 5-22 is appropriate.



Band	L_1	L_2
160	60 turns #26 enam. center-tapped	12 turns #26 enam.
80	45 turns #22 enam. center-tapped	8 turns #26 enam.
40	21 turns #22 enam. center-tapped	4 turns #20 enam.
20	11 turns #22 enam. center-tapped	3 turns #20 enam.
15	8 turns #20 enam. center-tapped	2 turns #20 enam.
10	5½ turns #20 enam. center-tapped	2 turns #20 enam.

Fig. 5-22. Data for balanced coils.

The output stage operates as a linear rf amplifier. This mode is essential if the modulated sideband output of the balanced modulator is not to be distorted. Proper bias level is set by the cathode resistor-capacitor combination.

Microphone signal is built up in level by a two-stage triode audio amplifier. The mike gain control is located between stages. The double-sideband generator operates with a supply voltage of 200 to 300 volts. A simple arrangement of a bridge rectifier and inductor-capacitor filter functions well.

Operation

The operation of the double-sideband generator is best observed on 160 meters because a service-type oscilloscope connected to the output permits a good display of the sideband pattern. The resonant sections of transformers T_1 and T_2 can be set in an appropriate spot in the 160-meter band using a dip oscillator. Insert a 160-meter crystal into the oscillator circuit and connect an oscilloscope across the output. Use a 50- to 300-ohm terminating resistor. It is now possible to jockey back and forth between two balance controls and the output tune capacitor of the balanced modulator to obtain minimum carrier level at the output. The output capacitor of the linear amplifier is tuned for maximum output. Now go back and readjust the balance control for minimum output.

A 1000-Hz audio signal can now be applied to the mike input of the sideband generator. Set the peak amplitude of the input sine wave to a level that corresponds to the peak output of the microphone. The characteristic double-frequency, double-sideband pattern should be displayed. Refer to Fig. 5-4. Adjust the audio gain until there is just a trace of pattern flat topping.

Connect the microphone to the input. When speaking into the microphone the pattern on the scope screen should be kept to approximately the same peak level. If not, make the necessary change in the setting of the mike gain control.

PROJECT 23: SINGLE-SIDEBAND GENERATOR

The circuit of the previous project can be rearranged to obtain a single-sideband signal as shown in Fig. 5-23. It involves the addition of a 455-kilohertz discriminator transformer and a mechanical

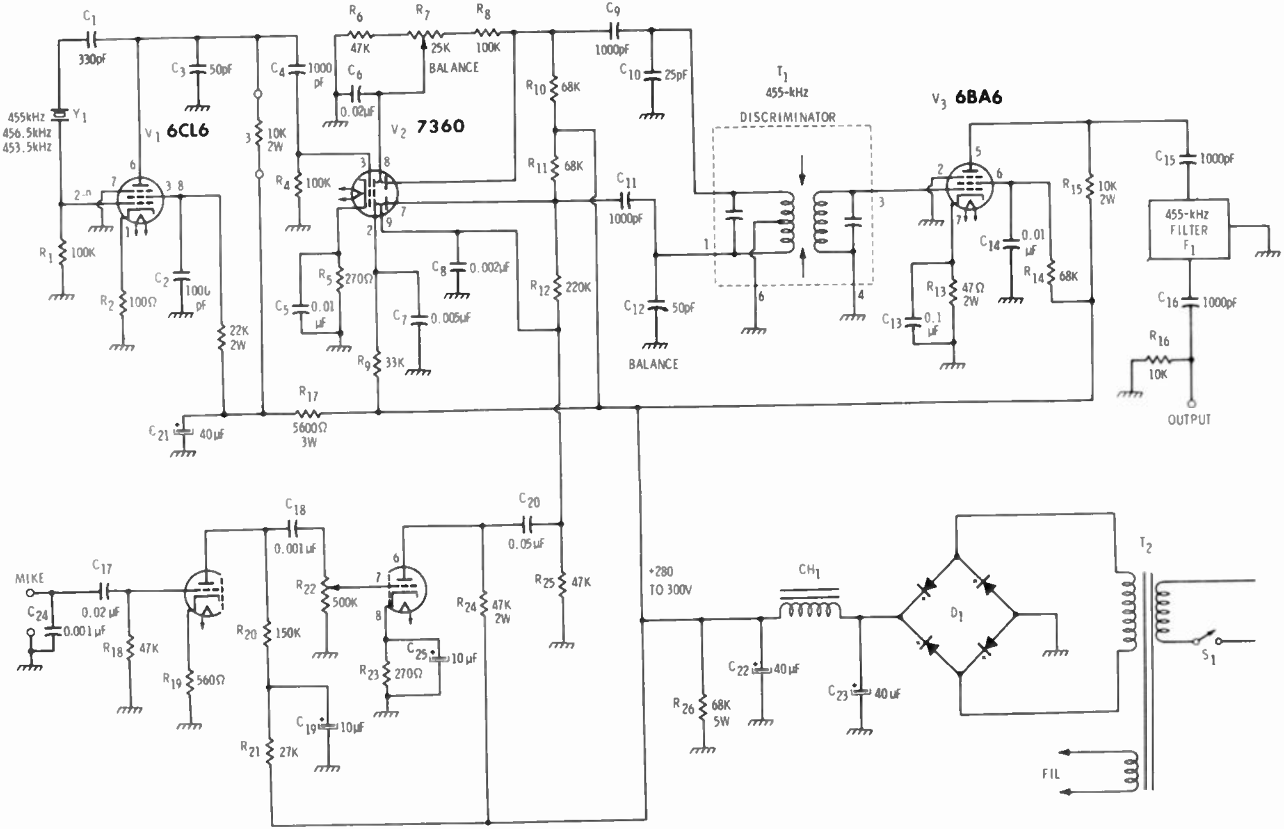


Fig. 5-23. Single-sideband generator

sideband filter. An appropriate crystal must be employed, the frequency of which depends on the desired transmit sideband, upper or lower. The 455-kilohertz sideband signal must be supplied to a succeeding linear mixer and oscillator combination to up-convert the signal to the desired transmit band.

Circuit Arrangement

The crystal-oscillator circuit is basic and similar to that of the previous project. However, the crystal frequency must be considerably lower to match the frequency of the mechanical filter. The balanced-modulator circuit is similar too except for value changes and an

Parts List for Fig. 5-23.

C_1	330-pF disc capacitor
$C_2, C_4, C_9, C_{11}, C_{15}, C_{16}, C_{18}$	1000-pF disc capacitors
C_3	50-pF disc capacitor
C_5	0.01-μF paper or disc capacitor
C_6	0.02-μF paper or disc capacitor
C_7	0.005-μF paper or disc capacitor
C_8	0.002-μF paper or disc capacitor
C_{10}	25-pF disc capacitor
C_{12}	50-pF trimmer capacitor
C_{13}	0.1-μF paper or disc capacitor
C_{14}	0.01-μF disc capacitor
C_{17}	0.02-μF paper or disc capacitor
C_{19}	10-μF electrolytic capacitor, 450V
C_{20}	0.05-μF paper or disc capacitor
C_{21}, C_{22}, C_{23}	40-μF electrolytic capacitors, 450V
Ch_1	5-7 henry filter choke
D_1	Diode bridge rectifier, 2A, 800PIV
F_1	455-kHz mechanical filter
R_1, R_4, R_8	100K 1-watt resistors

R_2	100-ohm 1-watt resistor
R_3	10K 2-watt resistor
R_5, R_{23}	270-ohm 1-watt resistors
R_6, R_{18}, R_{25}	47K 1-watt resistors
R_7	25K potentiometer
R_9	33K 1-watt resistor
R_{10}, R_{11}	68K 1-watt resistors
R_{12}	220K 1-watt resistor
R_{13}	47-ohm 2-watt resistor
R_{14}	68K 1-watt resistor
R_{15}	10K 2-watt resistor
R_{16}	10K 1-watt resistor
R_{17}	5600-ohm 3-5 watt resistor
R_{19}	560-ohm 1-watt resistor
R_{20}	82K 1-watt resistor
R_{21}	27K 1-watt resistor
R_{22}	500K potentiometer
R_{24}	47K 2-watt resistor
R_{26}	68K 5-watt resistor
S_1	Spst toggle switch
S_2	Spst slide switch
T_1	455-kHz discriminator transformer
T_2	Power transformer, 240V sec., 6.3V fil.
V_1	6CL6 tube
V_2	7360 beam-deflection tube
V_3	6BA6 tube
V_4	12AX7 tube
Y_1	Crystals, 455, 456.5, and 453.5 kHz

Miscellaneous Parts

Chassis 11" x 7" x 2"	Crystal socket
3 9-pin miniature tube sockets	Ac line cord
1 7-pin miniature tube sockets	Binding posts

output discriminator transformer. A discriminator transformer permits a balanced connection with its center tap grounded. The output resonant circuit is unbalanced and supplies the double-sideband signal to the grid of the 6BA6 linear amplifier.

The frequency response of a mechanical sideband filter is given in Fig. 5-24. Note that the flat part of the response is centered about

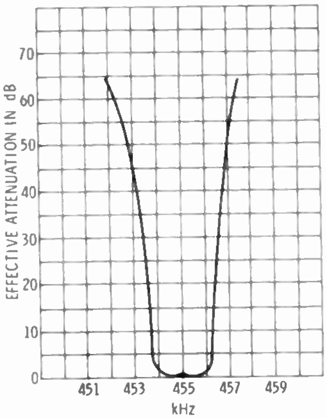


Fig. 5-24. Response of mechanical filter.

455 kHz (about ± 1 kilohertz). On a frequency ± 1.5 kHz on each side of center, the response is down 30 dB. For upper-sideband output the carrier frequency is located on 453.5 kHz; lower-sideband emission, on 456.5 kHz. These are the crystal frequencies indicated in Fig. 5-23. A 455-kHz crystal is also useful in tuning up the sideband generator because it is centered in the bandpass of the mechanical filter.

A consideration in selecting the crystal is whether the mixer-oscillator frequency is to be located above or below the transmit frequency. Let us use 160 meters as an example. Assume that a lower-sideband signal is to be transmitted in 1850 kHz. If the oscillator is to operate on the low side of the transmit frequency, proper sideband is obtained by using a 456.5-kilohertz carrier crystal. Therefore the oscillator frequency would have to be 1393.5 kHz (1850 − 456.5). With a modulating tone of 1500 Hz, the sideband output would be 1848.5 kilohertz (1393.5 + 455). Note that this is on the low frequency side of the 1850-kHz carrier frequency forming a lower side frequency.

In transmitting a lower sideband signal with the oscillating frequency on the high side of the transmit frequency it is necessary to use a carrier crystal frequency of 453.5 kHz. In this case the oscillator frequency would have to be 2303.5 kHz (1850 + 453.5). Now a modulating 1500-hertz tone produces a sideband on 1848.5 kHz (2303.5 — 455). Again the sideband output is on the low-frequency side of the 1850-kHz carrier frequency.

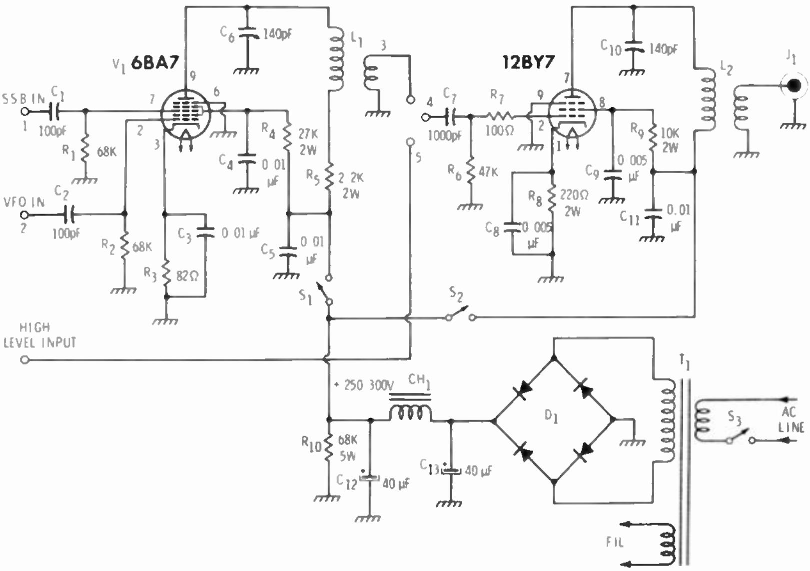
Operation

Tuneup of the single-sideband generator is quite similar to the double-sideband version. Connect an oscilloscope to the output of the filter through an isolating probe. Plug in a 455-kHz tuneup crystal. The two windings of the discriminator transformer and the two built-in adjustments (input and output resonance) of the mechanical filter are now made to peak the output. Next insert the appropriate carrier crystal. Adjust the balance controls (R_7 and C_{12}) and the balance winding of the discriminator transformer for minimum output. This setting of the discriminator winding is indicated when there is a rise in output on each side of the correct minimum, or null, adjustment.

Apply the 1500-Hz tone to the mike input. Again set its level to correspond with the normal output of the microphone. A single-tone pattern should now appear on the oscilloscope screen. Normal and abnormal conditions are shown in Fig. 5-4. Replace the tone signal source with a microphone and check the output with voice. Output should rise to the same level on voice peaks. You are now ready to supply output signal to a succeeding amplifier and mixer oscillator. A simple circuit is detailed in Project 24.

PROJECT 24: UTILITY LINEAR AMPLIFIER AND MIXER

The circuit arrangement of Fig. 5-25 can be used as a single-stage or two-stage linear amplifier The input stage can also be used as a mixer. Such a utility device is useful in building up the signal level of sideband generators, particularly the low-level outputs of solid-state devices. Output level is such that it can be displayed conveniently on an oscilloscope screen or used in conjunction with an rf monitor scope such as the Heath SB-610. Peak envelope power output is in the 1-to-2 watt PEP level.



C_1,C_2	100-pF disc capacitors	R_4	27K 2-watt resistor
C_3,C_4,C_5,	0.01-µF paper or disc capacitors	R_5	2.2K 2-watt resistor
C_{11}		R_6	47K 1-watt resistor
C_6,C_{10}	140-pF variable capacitors	R_7	100-ohm 1-watt resistor
C_7	1000-pF disc capacitor	R_8	220-ohm 2-watt resistor
C_8,C_9	0.005-µF disc capacitors	R_9	10K 2-watt resistor
C_{12},C_{13}	40-µF electrolytic capacitors	R_{10}	68K 5-watt resistor
CH_1	5-7 henry filter choke	S_1,S_2	Spst slide switches
D_1	Diode bridge rectifier,	S_3	Spst power toggle switch
	2A 800 PIV	T_1	Power transformer 240V sec.,
J_1	Coaxial receptacle		6.3V fil.
L_1,L_2	Coil sets (Project 17)	V_1	6BA7 tube
R_1,R_2	68K 1-watt resistors	V_2	12BY7 tube
R_3	82-ohm 1-watt resistor		

Fig. 5-25. Utility linear amplifier and mixer.

Circuit

Two vacuum-tube stages are used in conjunction with a solid-state bridge rectifier that supplies 250-300 volts for the tube electrodes. The input stage can be used as a mixer or an amplifier and uses a pentagrid 6BA7. In the linear amplifier mode, the signal can be applied to either grids 2 or 7. Signal is further amplified by the follow-up 12BY7 stage. If a high-level input signal is to be amplified, use the output stage only and appropriate jumper from 4 to 5.

Plug-in coils are used and permit multiband operation. Coil data are given in Fig. 2-7 of Project 7.

Operation

The operation of the linear amplifier and mixer can be checked out using the 455-kHz sideband generator of the previous project. Its output is supplied to grid 7 of the mixer. To obtain lower sideband output on the 160-meter band, insert the 453.5-kHz crystal. The vfo frequency must be in the 2.3 megahertz range. The vfo, as detailed in Project 9, can be adjusted to operate in this frequency range. It may be necessary to readjust the appropriate 160-meter coil slug.

The plate resonant circuits must be tuned to the desired 160-meter frequency. If operation is desired on 1820.5 kilohertz, the vfo should be tuned to 2274 kHz (1820.5 + 453.5). When using 1500-hertz tone modulation of the sideband generator the resultant sideband is displayed on the oscilloscope screen. Despite the low output, a visible pattern will also appear on the screen of a monitor scope.

To operate between 1800-1850 kHz, the vfo must operate between 2255-2305 kHz. On 80 meters (3800-4000 kHz), the vfo must operate between 4255-4455 kHz. On 40 meters (7200-7300 kHz), the vfo must operate between 7655-7755 kHz. The vfo of Project 9 can be adjusted to cover these ranges by selection of the proper oscillator inductor.

PROJECT 25: DUAL-FET BALANCED MODULATOR

The dual field-effect transistor consists of two FETs of identical characteristics mounted in the same case. Designed mainly for use in differential amplifiers, they also operate ideally in balanced modulator and demodulator circuits. They function well as balanced modulators for both double-sideband and single-sideband transmitters, some units up to several hundred MHz.

Circuit

A circuit that operates double-sideband 10 through 160 meters is given in Fig. 5-26. The paralleled gates are supplied with carrier; the drains are connected in push-pull to obtain carrier cancellation. Modulating signal is applied to a push-pull connected source circuit. A balanced resonant primary circuit and an untuned secondary form the output transformer. Balanced coil data is given in Project 22. Transformer T_1 data is given in Fig. 1-14.

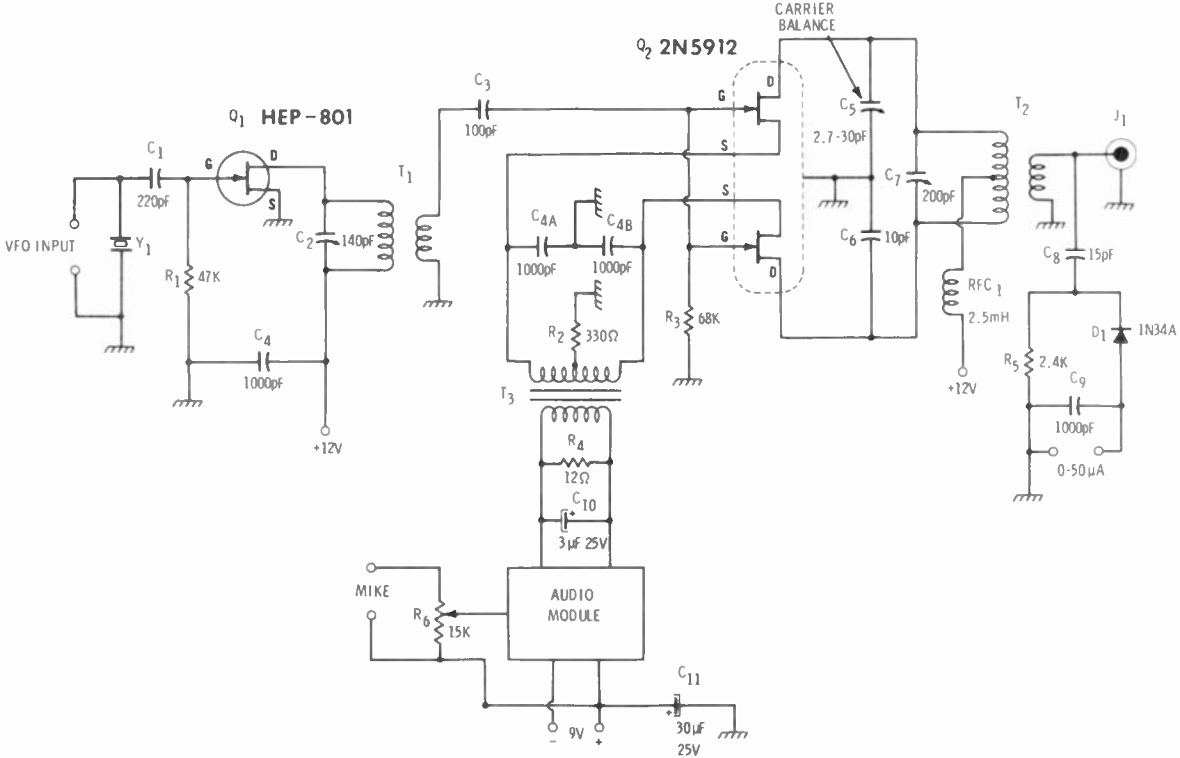


Fig. 5-26. FET balanced modulator.

The output circuit is balanced with a small trimmer capacitor (C_5). An output indicator is attached and provides a means of adjusting the carrier balance. Indicator deflection follows the modulation too. An acceptable reading can be obtained on a 0- to 1-mA meter although better meter deflection is possible if a more sensitive meter such as a 0- to 50-μA is available.

The source of carrier signal is an FET Miller-connected crystal oscillator. This stage can also be used as an amplifier simply by removing the crystal from its socket. In this case it can be driven by either the vfo of Project 9 or Project 20.

The source of audio signal is a small audio module. An input potentiometer provides an audio gain control. A transistor output transformer permits matching between the very low output impedance of the module and the somewhat higher input impedance of the source circuit of the balanced modulator. Usually the audio module will provide more output than is necessary and its output can be loaded with a resistor. This can be done experimentally until proper audio drive is obtained and nondistorted output is obtained from the module. In our example using a 3-watt-output type it was necessary to connect a 12-ohm load resistor (R_4) across its output. To limit the

Parts Lists for Fig. 5-26.

C_1	220-pF disc capacitor	Q_1	HEP 801 transistor
C_2	140-pF variable capacitor	Q_2	2N5912 FET, Siliconix
C_3	100-pF disc capacitor	R	47K ½-watt resistor
C_4	1000-pF disc capacitors	R_2	330-ohm ½-watt resistor
C_4A,B,C_9		R_3	68K ½-watt resistor
C_5	2.7- to 30-pF trimmer capacitor	R_4	12-ohm 2-watt resistor
		R_5	2.4K ½-watt resistor
C_6	10-pf disc capacitor	R_6	15K potentiometer
C_7	200-pF variable capacitor	RFC	2.5-mH rf choke
C_8	15-pF disc capacitor	T_1	Set of coils from Project 3
C_{10}	3-μF electrolytic capacitor, 25V	T_2	Set of coils from Project 22
		T_3	Transistor output transformer, pri 100 ohms C.T., sec 3.2/8/16 ohms.
C_{11}	30-μF electrolytic capacitor, 25V		
D_1	1N34A diode	Y_1	Crystal for appropriate band
J_1	Coaxial output receptacle		

Miscellaneous Parts

Pegboard 8″ x 12″
2 Crystal sockets
1 Transistor socket
7 Soldering lugs to mount 2N5912
1 Audio module, 2- to 3-watt
1 12-volt lantern battery
1 9-volt transistor battery
2 Five-prong coil sockets

modulating frequency to the voice range, a large capacitor (C_{10}) was also shunted across the output. With this value of capacitor the high-frequency response begins to drop off rather sharply at 2500 hertz.

Most audio modules operate with a positive ground, and capacitor C_{11} maintains dc isolation, but joins the two commons for the audio and radio frequencies.

Operation

Connect a 270-ohm terminating resistor across the output. Insert a 160-meter crystal. Connect a 0-1 milliammeter or preferably a 0-50 microammeter across the indicator binding posts. Connect an oscilloscope across the output. Set capacitor C_5 to near maximum value.

Turn on the oscillator and balanced modulator. Adjust capacitor C_2 for maximum output then adjust capacitor C_7 for a greater maximum output. If your oscilloscope has a high enough horizontal-sweep frequency, it is possible to observe the quality of the 160-meter sine wave.

Adjust capacitor C_5. Note that the oscilloscope pattern as well as the indicator reading falls. Note at one point that the scope vertical deflection and the meter reading fall to a low minimum value. This demonstrates the importance of the carrier-balance control in minimizing the carrier level at the output.

Use the oscilloscope to observe the level of carrier signal at both gates and both drains of the balanced modulator. Observe that the carrier level at the gate is higher than at the output. Also both gates are supplied with equal-amplitude carrier signal level. Even greater signal appears at the drains, although both drains provide equal-amplitude outputs, they are of a relative phase that produces cancellation.

Connect the output of an audio generator to the microphone input. Set the audio oscillator frequency to approximately 1500 Hz. Slowly increase the audio gain setting until the pattern appears on the oscilloscope screen. Adjust the horizontal-sweep control of the oscilloscope until four cycles are displayed.

Observe the influence of the microphone gain-control setting. Adjust the setting for maximum vertical deflection up to the point at which the peak begins to round off, indicating peak clipping and the generation of distortion components.

Connect the oscilloscope across the output of the audio module. Only two sine waves are displayed verifying that the rf output observation displays the double-frequency, double-sideband modulation pattern.

Decrease the audio frequency to 750 Hz, displaying just two cycles on the scope screen. Increase the audio frequency to 3000 Hz. At this frequency eight cycles are displayed. Observe that the pattern is now of lower amplitude indicating the dropoff of the high-frequency response, the preferred manner of sideband operation. Reset the audio oscillator frequency to 1500 Hz.

Connect a 50-pF capacitor across C_5. By so doing the bridge is unbalanced and a carrier component develops in the output. In fact, as your oscilloscope shows, you are displaying two cycles of a conventional amplitude-modulated signal. This is known as carrier insertion and some sideband modulators include this facility in one form or another. Notice that the amplitude-modulation percentage, although not 100%, is relatively high. This affords a simple means of switching between conventional a-m and double-sideband carrier-suppressed modulation.

Disconnect the audio oscillator and attach a microphone across the microphone input binding posts. Turn on the balanced modulator and audio module. Observe that when you speak into the microphone there is vertical deflection of the scope pattern and an upward kick of the indicator meter. To obtain maximum output and, at the same time, avoid distortion it is necessary to set the microphone gain control properly. If you sound a sustained "ah" or speak and extend the numeral "four" any flattening of peaks indicates distortion. This is known as peak clipping.

As far as the output indicator reading is concerned it should deflect up to the same level, or perhaps, just a bit higher than the fixed reading obtained when using the 1500-Hz audio test.

The linear amplifier of Project 24 can be connected to the output of the FET balanced modulator. Signal level will then be built up to about 1 watt PEP. On some bands the modulator output is high enough that only the output stage is needed.

Turn off the audio modulator and balanced modulator. Remove any crystal. Connect the output of the vfo of Project 9 to the input of the carrier amplifier. Turn on the vfo, set it to 160 meters, and tune it to the desired frequency. Turn on the balanced modulator. There should be no need to readjust the carrier balance although resetting of C_2 may be necessary for maximum output. Turn on the audio module. Note that the performance is similar to that obtained with crystal operation. Pattern shape and output level are identical to those for crystal-controlled operation.

Shunt the 50-pF capacitor across C_5. Note that the carrier comes up on the oscilloscope display, and also note the increase in the indicator reading. Speak into the microphone. Now a normal amplitude-modulation envelope with speech modulation is displayed on the screen. The indicator reading is reasonably constant except for upward kick with modulation peaks. This is the normal condition for a-m.

Repeat the above procedures for each of the radio amateur bands 10 through 80 meters. Similar results are obtained on each band. For 10-meter operation it was necessary to decrease values of capacitors C_2 and C_7 to 50 pF when using the ten-meter coils of Projects 3 and 22.

CHAPTER 6

Basic Principles of Linear Mixers and Amplifiers

A single-sideband signal is seldom transmitted on the frequency at which it is generated when using a filter-type sideband generator. Instead it is heterodyned to the transmit frequency by up or down frequency conversion. A sideband signal generated by the phasing method is often transmitted on the frequency at which it is generated and can be transmitted on a lower or higher frequency with a conversion system. A double-sideband signal is often transmitted on the generated frequency. However, it too can be transmitted on higher or lower frequencies using up or down frequency converters.

When a mechanical type sideband filter in the 455-kHz range is employed, up-frequency conversion is employed, Fig. 6-1. If we assume the response of the sideband filter is centered about 455 kHz, Fig. 5-15, the carrier generator frequency determines whether the upper or lower sideband is to be transmitted. Typical carrier-oscillator frequency for lower-sideband transmission would then be 456.5 kHz: upper-sideband transmission 453.5 kHz. For example, if the carrier oscillator is set to 456.5 kHz, a lower sideband component would appear at exactly 455 kHz when modulated with a 1500-hertz tone. This would be in the center of the passband of the 455-kHz mechanical filter. The high-frequency sideband with 1500-hertz modulation would be 458 kHz (456.5 + 1.5). This would be out of the bandpass of the filter and would be attenuated.

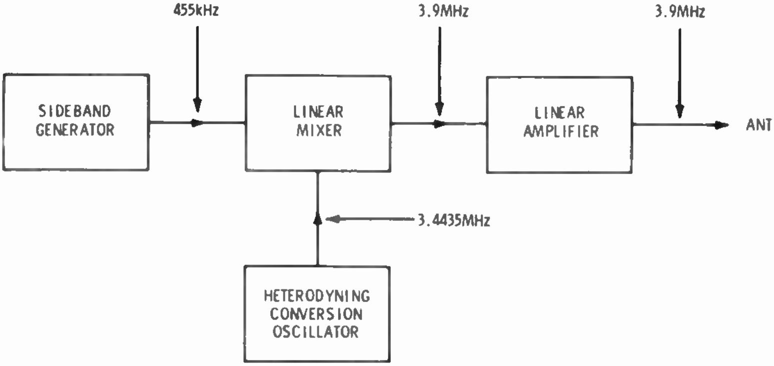


Fig. 6-1. Linear conversion and amplification plan.

If the higher-frequency sideband is to be passed to the output of the filter, it would be necessary to use a carrier oscillator frequency of 453.5 kHz. When modulated with a 1500-Hz component there would be a high-frequency sideband at 455 kHz (453.5 + 1.5). The low-frequency sideband would appear at 452 kHz or off the band-pass of the filter, and would be attenuated.

Let us assume that the low-frequency sideband is being made available at the output of the sideband filter. This is applied to the input of a mixer. What would the frequency of the conversion oscillator have to be to obtain a low-frequency sideband signal at a carrier frequency of 3.9 MHz.

It would be necessary to use up-conversion and the conversion oscillator frequency would have to be the difference frequency between the desired transmit frequency and the carrier oscillator frequency or 3.4435 MHz (3.9 − 0.4565). When modulating with a 1500-Hz tone, the lower sideband would appear at 3.8985 MHz (3.4435 + 0.455).

If the transmitter is to be made tunable between 3.8 and 4 MHz, the conversion oscillator would have to tune between 3.3435 and 3.5435 MHz.

It must also be stressed that whether an upper- or lower-sideband frequency is transmitted also depends on the frequency of the conversion oscillator. Were the conversion oscillator set on the high-frequency side of the desired transmit frequency, the 456.5-kHz carrier oscillator would produce an upper-sideband signal rather than a lower-sideband signal. In this case the conversion oscillator frequency would be 4.3565 MHz (3.9 + 0.4565). The transmit frequency is now the difference frequency between the conversion oscillator frequency and the carrier oscillator frequency. With a modulating frequency of 1500 Hz, the sideband component would be transmitted on 3.9015 MHz (4.3565 − 0.455) which is on the high-frequency side of the transmit carrier frequency. Conversely using a 453.5-kHz carrier oscillator frequency would result in the transmission of a lower-sideband signal when the conversion-oscillator frequency is set to the high-frequency side of the desired transmit frequency.

In the radio amateur services the most common carrier-oscillator frequency is in the 9-MHz range. When this plan is used as shown in Fig. 6-2, a down-conversion technique is often employed to ob-

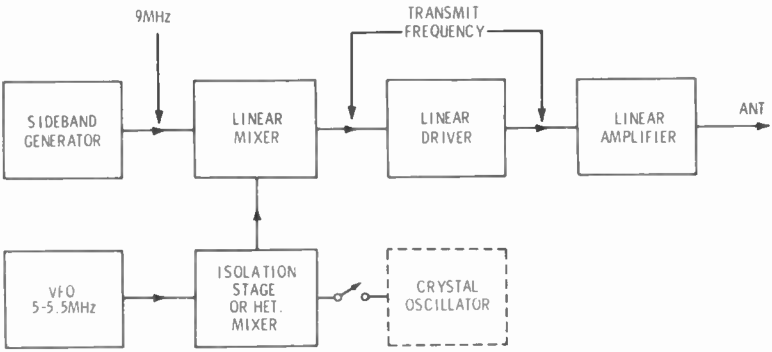


Fig. 6-2. Linear mixing and amplifying plan with up or down conversion.

tain output on the 40-, 80-, and 160-meter bands. In the example, a variable-frequency oscillator tunes over a frequency range of 5 to 5.5 MHz. If this frequency is applied to a linear mixer tuned for down conversion, there is an 80-meter (3.5 to 4 MHz) difference frequency. When the same linear mixer is operated as an up-converter, the summation of the two frequencies produces 20-meter output (14.0 to 14.5 MHz). A 9.0015-MHz carrier crystal provides lower sideband operation on 20 or 80 meters; an 8.9985-MHz carrier crystal, produces upper-sideband operation on either band using this conversion technique.

A second mixer step can be inserted between the vfo and the linear mixer. Along with an appropriate crystal oscillator and bandswitch it can be used to obtain operation on other bands. For example a 25-MHz crystal when beat with the 5-MHz vfo signal can produce a 30-MHz sum frequency at the output of the second heterodyne mixer. The difference frequency at the output of the linear mixer will then be the 15-meter band (30MHz − 9MHz).

LINEAR MIXERS

The linear mixer circuits in single-sideband and double-sideband systems must be designed and adjusted to minimize the level of the injection frequency that can develop in the output. The injection frequency is of course the means of obtaining up or down conversion but it is a component that must be rejected in the output. Mixer circuits must be designed to emphasize the desired sum or difference frequency that is to be developed in the output. Furthermore, if the sum frequency is to be used, the output system must reject the difference frequency. If the difference frequency is to be developed in the output, the output circuit must reject the sum frequency. The nearer the desired and undesired frequencies are, the more difficult it becomes to reject the undesired component.

Balanced mixers are popular. They are very similar to balanced modulators and can provide a high order of rejection of undesired oscillator-mixing frequencies. Solid-state mixers perform well because of the high-order of balance that can be obtained in integrated circuits and in special dual FET and bipolar combinations. Multigrid converter tubes also are popular because of the isolation between output and input and the low level of oscillator injection signal required. A diode bridge or ring configuration serves well as a sideband mixer when the diodes have identical characteristics. Hot-carrier diodes have been particularly popular in these circuits because of their linearity and high ratio of back-to-forward resistances.

The balanced configuration of Fig. 6-3 applies the sideband signal between points A and C. The mixing oscillator component is applied in balanced fashion between points B and D. The output is a single-ended arrangement with signal being removed between center tap and common. The output resonant circuit selects either the sum or difference frequency. The double-balance arrangement results in the cancellation of the input sideband frequencies as well as the injection oscillator frequency.

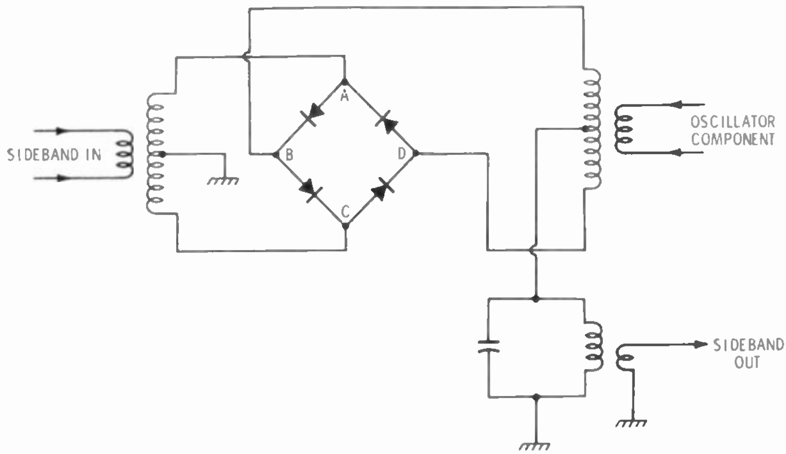


Fig. 6-3. Diode balanced mixer.

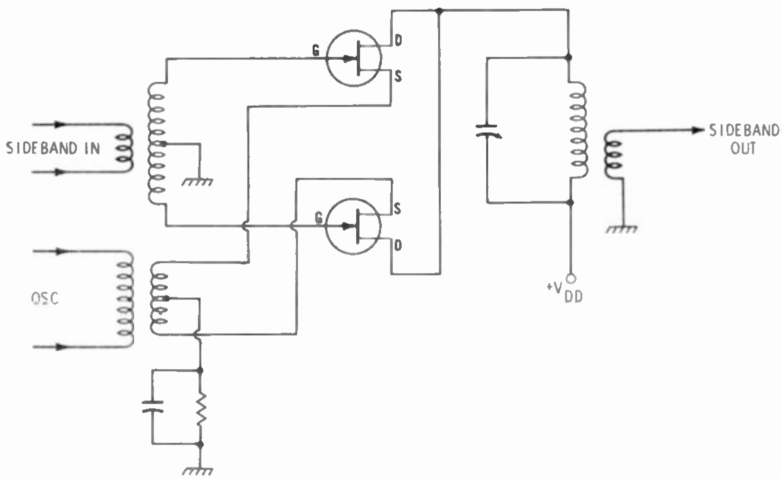


Fig. 6-4. FET balanced mixer.

Two identical field-effect transistors function well in balanced mixer configurations. Sideband signal can be applied in push-pull between gates, Fig. 6-4. This connection results in a very light loading of the sideband signal source. The stronger injection oscillator component is applied in push-pull between the two sources. Again there is balanced feed of both mixing components and cancellation in the parallel-connected drain circuit. By tuning the drain resonant circuit to either the sum or difference frequency the side-band signal frequency can be converted up or down.

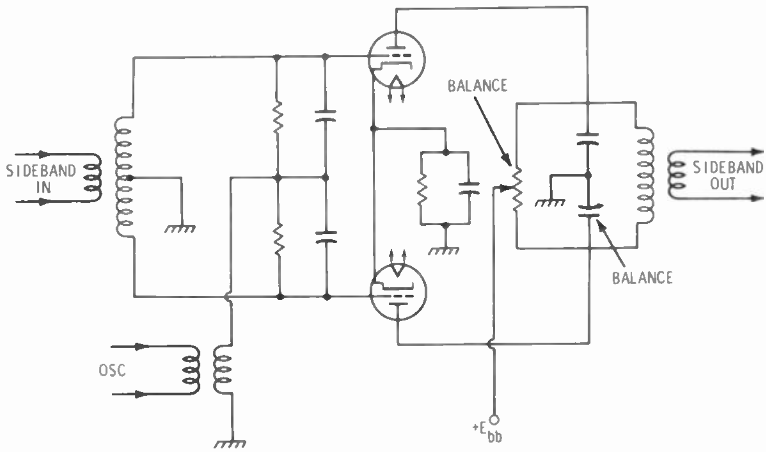


Fig. 6-5. Triode balanced mixer.

A vacuum-tube arrangement is shown in Fig. 6-5. Oscillator injection signal is applied to the grids in parallel. Therefore there is cancellation in the push-pull connected plate circuit. Sideband signal

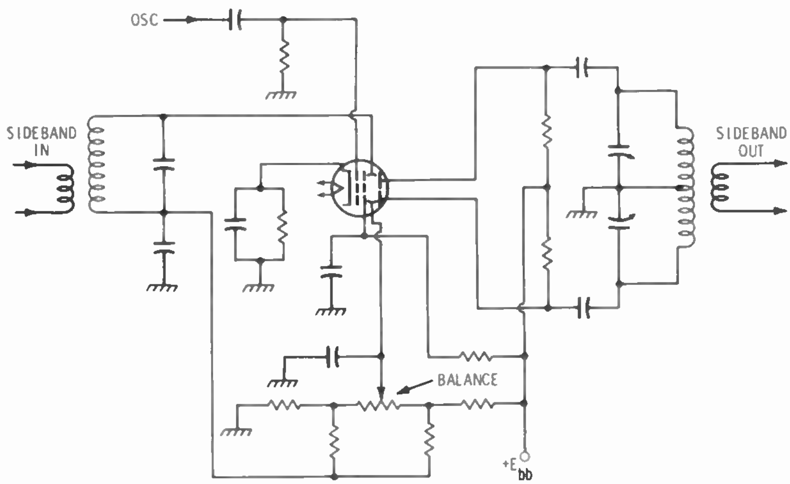


Fig. 6-6. Beam-deflection linear mixer.

is applied from grid to grid. Amplitude and phase balance is obtained with the plate circuit potentiometer and trimmer capacitor respectively.

The good balance and effective isolation of the beam-deflection tube makes it effective as a sideband mixer, Fig. 6-6. In this arrangement the sideband signal is applied to the deflection electrode while the stronger injection oscillator signal is applied to the control grid. The balanced plate-to-plate circuit cancels the oscillator component. The associated resonant circuit is tuned to the sum or difference frequency and the sideband signal is converted up or down in frequency. The same resonant circuit acts as a low-impedance shunt to the original sideband frequencies and they do not develop in the output.

When there is adequate separation in frequencies, a pentagrid tube performs well as a mixer (Fig. 6-7). In this case the sideband

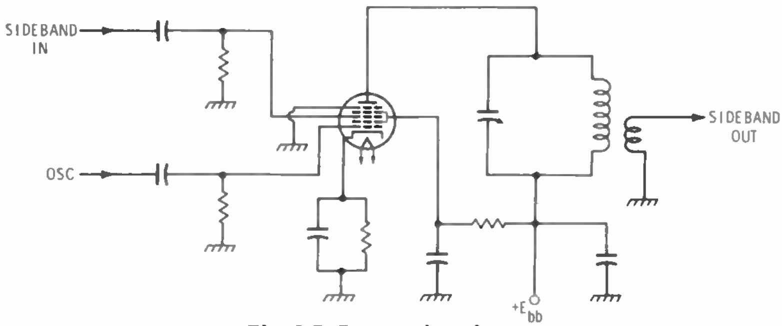


Fig. 6-7. Pentagrid mixer.

signal is applied to the second signal grid while the stronger injection-oscillator component is applied to the control grid. The plate output circuit must have a high Q and be tuned exactly to the desired sum or difference frequency.

LINEAR AMPLIFIERS

Linear amplifiers must be used to build up the level of any type of a-m signal—be it standard a-m, double-sideband suppressed carrier (dsb), or single sideband. Linear amplifiers are essential in single-sideband systems. After the sideband signal has been converted to the transmit frequency it must be built up in power level. Also a linear amplifier is often used to give additional power boost to the signal made available at the output of amplitude-modulated transmitters.

Certain vacuum tubes perform well as linear amplifiers if they are biased properly. Power type field-effect transistors show great promise as linear amplifiers. Even bipolar transistors with suitable circuits, and at lower efficiency, can be used in linear service. In fact, several bipolar transistors are now available which are designed specifically as linear amplifiers with sideband outputs of 150 watts PEP and higher.

The rf and i-f amplifiers of a receiver are operated class-A linear because the received amplitude-modulated signals must not be distorted. Some of the low-level stages of sideband transmitters also operate class-A. However in transmitter power amplifiers class-AB or

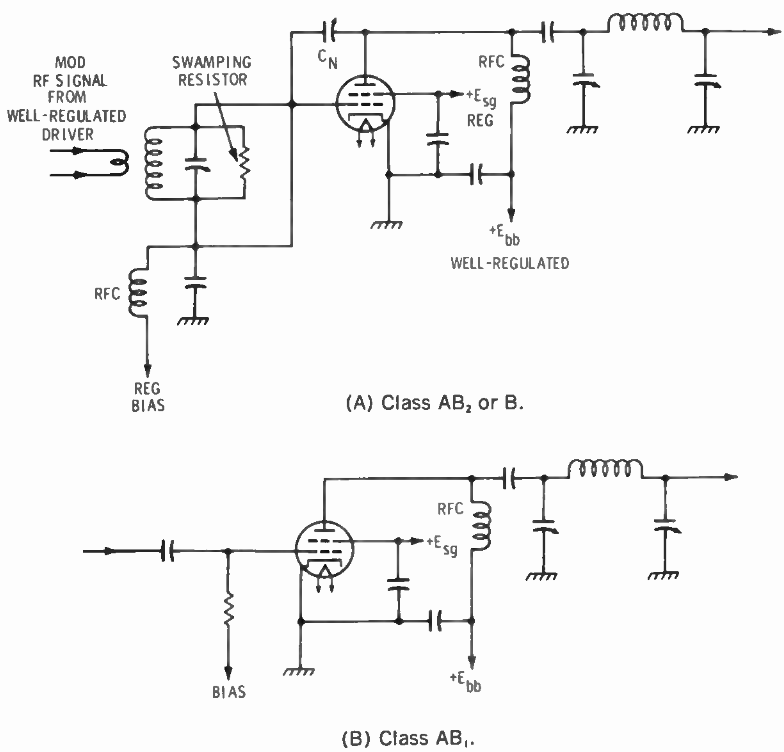


Fig. 6-8. Typical linear amplifiers.

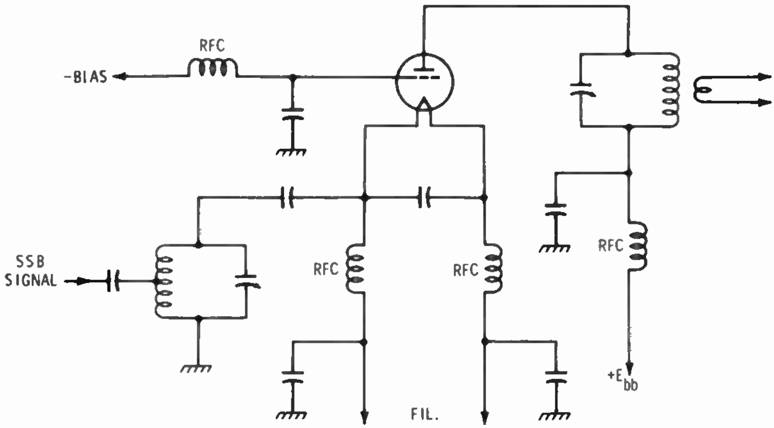
class-B, either single-ended or push-pull, provide higher efficiency and higher power output. When the output circuits are resonant, single-ended class-AB or class-B stages can be operated linearly because of the smoothing action of a resonant tank circuit. Nonetheless, the class-AB or class-B linear amplifier must be well designed, carefully balanced, and properly tuned if favorable operating characteristics are to be obtained.

Schematically the linear amplifier differs very little from the appearance of the regular class-C rf power amplifier. Operating voltages and load impedances however are more nearly like those of the same tube used as a class-AB or class-B audio modulator. External bias is required to set the operating bias properly. Good linear operation requires that the bias and plate voltages, as well as the screen-grid voltages for tetrodes, be well regulated.

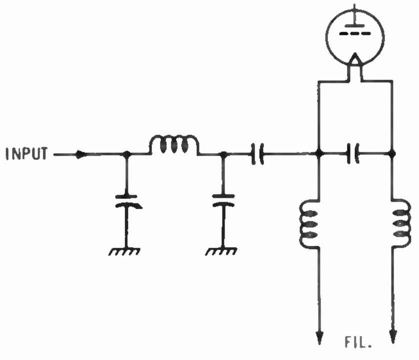
Typical linear-amplifier circuits are shown in Fig. 6-8 and 6-9. Most tetrode stages are operated class-AB_1 or AB_2 (very close to class-B operation). Class-AB_2 operation permits a somewhat higher output and lower resting plate current. In class-AB and class-B the resting (no-signal) plate current is low and rises with signal. The plate current varies with the changes in the incoming modulation envelope. Be it a standard a-m, or a sideband signal, the nearer the amplifier is biased class-B, the lower the resting plate current and the greater the plate-current swing with modulation.

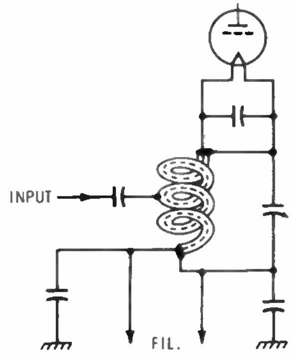
In the case of the tetrode linear amplifier, class-AB_2 or class-B operation (Fig. 6-8A) imposes stringent operating conditions. Since a class-AB_2 or class-B stage draws grid current, its input impedance varies throughout the radio-frequency cycles that comprise the modulation envelope. To maintain good linearity, both the rf driver and dc grid-bias source must be well regulated. In fact, a swamping resistor is often used across the input tank circuit. This is done even though the tank circuit itself is given a high C-L ratio for improved regulation. The swamping resistor must have a low ohmic value so that the load it places on the driver is better than five times the load placed by the input impedance of the linear amplifier. For class-AB_2 and class-B operation, a dc bias voltage must be supplied via a radio-frequency choke to minimize the influence of grid-current flow on the bias value.

The input system for class-AB_1 operation is less critical. Often a higher power output can be obtained by using several tubes in par-



(A) Tuned cathode-filament circuit.





(B) Pi-network input.

(C) Bifilar-tank input.

Fig. 6-9. Grounded-grid linear amplifiers.

allel in class-AB_1, in preference to the higher-power operation of a single tube in class-AB_2 or class-B mode. Tetrode tubes are commonly used for class-AB_1 amplifier operation because they are capable of delivering high peak plate current even though no grid current flows.

As shown in Fig. 6-8B, the absence of grid-current flow simplifies the input system. The driver need not supply power. Grid bias can be applied via a resistor. Because of the constant high input impedance, an untuned grid circuit is often used. For low-frequency operation with a suitable tube, no neutralization is required.

For truly linear operation amplifiers must be perfectly neutralized and be absolutely free of parasitics. Nonlinear operation of the amplifier produces two basic distortion components—radio-frequency harmonics of the carrier, and intermodulation distortion components related to the frequency of the modulating wave. The resonant tank circuits and output coupling system do much to eliminate harmonic components.

The intermodulation components are basically responsible for the need for absolutely linear operation. The sideband frequency is displaced from the reference carrier frequency by the frequency of the modulating wave. However, when nonlinearity is present, sideband components are generated that are separated from the reference carrier frequency by two, three, and more times the audio frequency. Such a condition places spurious frequency components, or "splatter," throughout the desired bandpass, causing distortion. Additional components are found on the adjacent channels, causing interference. No wonder the choice of operating voltages and conditions are so important to good linear-amplifier operation. In some circuits rf feedback paths are used to improve linearity, just as audio feedback improves the loading and linearity of high-power audio amplifiers.

The grounded-grid triode stage (Fig. 6-9) is a popular linear amplifier. The low input impedance does not change over as wide a range as the tube current changes with modulation. Furthermore, the power delivered to the cathode circuit by the driver is also fed in series with the output, contributing additional useful output. The power dissipated across a swamping resistor in a grounded-cathode circuit is wasted.

In a well-designed grounded-grid stage, there are few neutralization and parasitic problems. Although an untuned cathode circuit may be used with a grounded-grid amplifier, splatter and distortion components are considerably less when the cathode circuit is tuned. The energy-storing ability compensates for the influence of the changing cathode load relative to the positive and negative alternations of the driving signal.

Two additional recommended input systems are given in Fig. 6-9. A pi-network tank circuit provides impedance matching between the driver and the low impedance of the cathode circuit. Radio-frequency chokes provide a means of feeding the filament voltage while the cathode is maintained at a high rf potential. A bifilar tank coil serves as the inductor of the resonant circuit, as well as a means of applying the filament voltage, thus keeping the filamentary cathode at a high rf potential and, at the same time, keeping rf out of the filament supply lines.

Zero-bias rf tubes are particularly adaptable to linear-amplifier operation because they require no grid-bias supply. Input-impedance variations are not so great because grid-current flow and constant

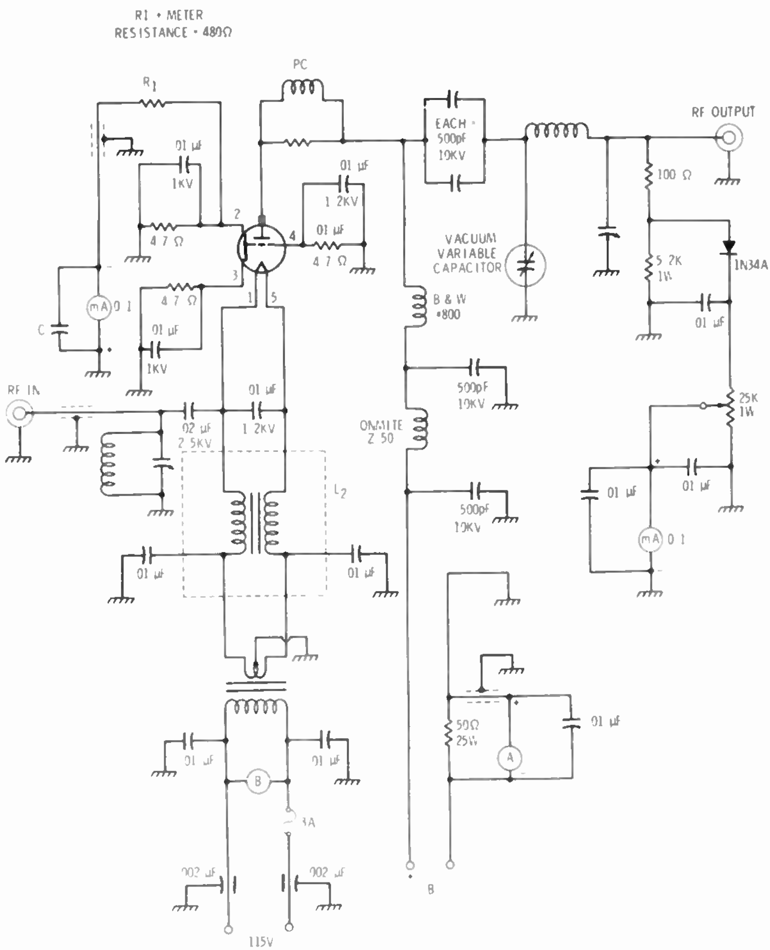


Fig. 6-10. A grounded-grid linear amplifier using a zero-bias tube.

loading persist over a substantial portion of the radio-frequency input cycle. A typical circuit using a zero-biased triode is shown in Fig. 6-10. A power gain as high as 20 can be obtained from a plate potential of 2500 volts. For single-sideband operation, a PEP plate input power of 2 kilowatts can be run. A 65-watt ssb generator can supply adequate drive.

The tube has been designed for high-frequency use, operating with full ratings up to 110 megahertz. For this reason, the control-grid system must have a low and uniform rf impedance. The rf drive is monitored by the dc meter in the control-grid circuit.

The plate-current meter is connected in the −B lead and it peaks at about 400 milliamperes with voice modulation. The plate-supply voltage is shunt-fed through high and low radio-frequency chokes. The output circuit is a pi network which includes an rf diode and dc metering circuit to record the relative rf power output.

The performance of a linear amplifier should be evaluated and careful adjustments made if distortion and splatter are to be avoided. The oscilloscope is the most effective indicator for check-out and adjustment of double-sideband and single-sideband gear. Audio- and radio-frequency vtvm's also have their place, as do the dc current meters of the linear power amplifier. The latter is less effective than an oscilloscope in the precise adjustment of a linear amplifier; however, once the amplifier is set for optimum operation, it is useful in monitoring its operation.

Pure sine-wave tones are important to the alignment of a single-sideband generator. Thus a good audio generator is almost a necessity. Inasmuch as the simultaneous use of two pure sine-wave tones is helpful, a simple two-tone audio generator that can form two sine waves simultaneously can be very useful.

FET LINEAR POWER AMPLIFIER

High input impedance and lack of secondary breakdown make the field-effect transistor attractive in linear amplifier applications. The high input impedance as compared to a bipolar type simplifies interstage coupling and matching. There is no thermal runaway. In class-AB power amplifier operation the basic square-law transfer characteristic over the entire input signal range reduces intermodulation distortion and the possibility of splatter.

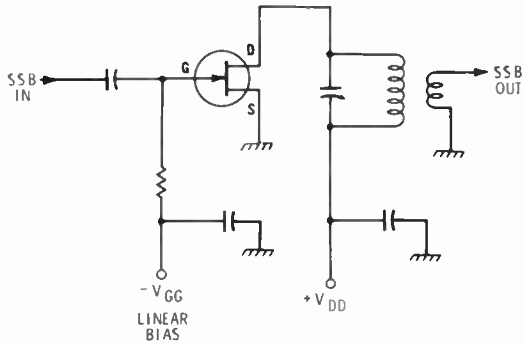


Fig. 6-11. FET linear amplifier circuit.

A typical low-power FET linear amplifier is shown in Fig. 6-11. The drain circuit is tuned; the input circuit, untuned. Class-AB bias is applied to the gate. In a typical practical amplifier this bias is set at one-half the pinch-off voltage of the particular FET. To prevent the flow of gate current the peak gate voltage may be no greater than the dc gate bias.

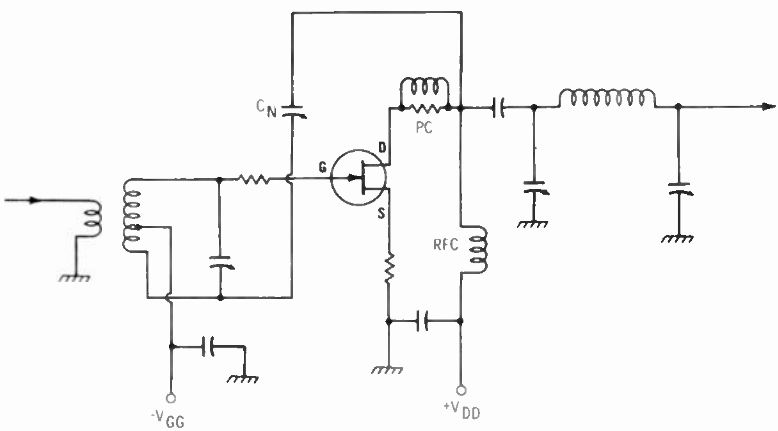


Fig. 6-12. Neutralized FET linear amplifier.

A higher-powered linear amplifier is shown in Fig. 6-12. This stage employs tuned input and output circuits with neutralization facility. Again the biasing level is set to approximately one-half the pinch-off voltage or a bit higher depending on the desired resting drain current and the amplitude of the input signal. As in the case of high-powered vacuum tubes there are hazards of self-oscillation and parasitics. The low value parasitic resistor in the gate circuit and/or a parasitic choke directly off of the drain reduces this problem. Parasitic oscillations can build up to the point at which the base current becomes excessive and the transistor is destroyed.

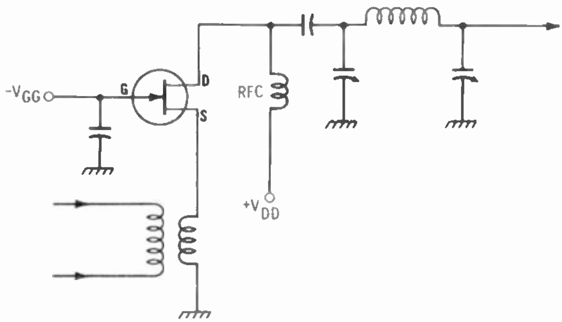


Fig. 6-13. Grounded-gate amplifier.

The grounded-gate circuit of Fig. 6-13 is attractive because the stage can be operated at high power levels without the need for neutralization. Of course, source of signal should have adequate drive for the lower input impedance of the source circuit. However, this added power also contributes to the effective power output of the amplifier in the grounded-gate configuration. Such an amplifier is constructed in one of the projects of this chapter.

BIPOLAR LINEAR AMPLIFIERS

Bipolar transistors can also be operated in linear-amplifier circuits at lower efficiency. Input impedance is of course quite low, and to obtain class-AB operation they must be forward-biased, causing a continuous flow of base and collector currents during much of the input wave. Base and collector currents are low with no modulation; both rise with modulation.

In the circuit of Fig. 6-14 there are tuned input and output circuits. Neutralization is not required because of the low input and output impedances of a bipolar transistor. A resistive voltage divider and potentiometer are used to set the class-AB bias.

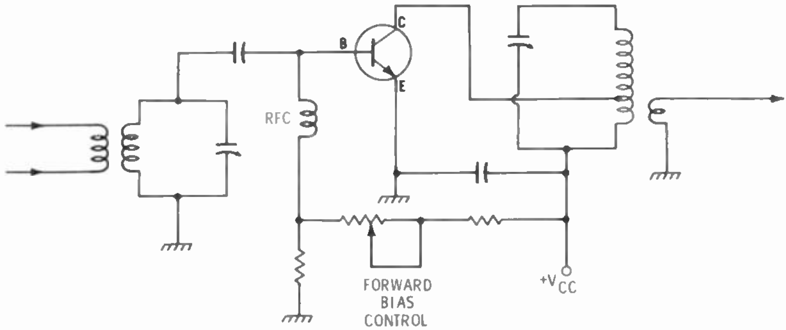


Fig. 6-14. Bipolar linear amplifier.

Many bipolar rf power transistors have been designed with class-C operation in mind. Forward-biasing of many of these types away from class-C to class-AB operation increases the secondary breakdown

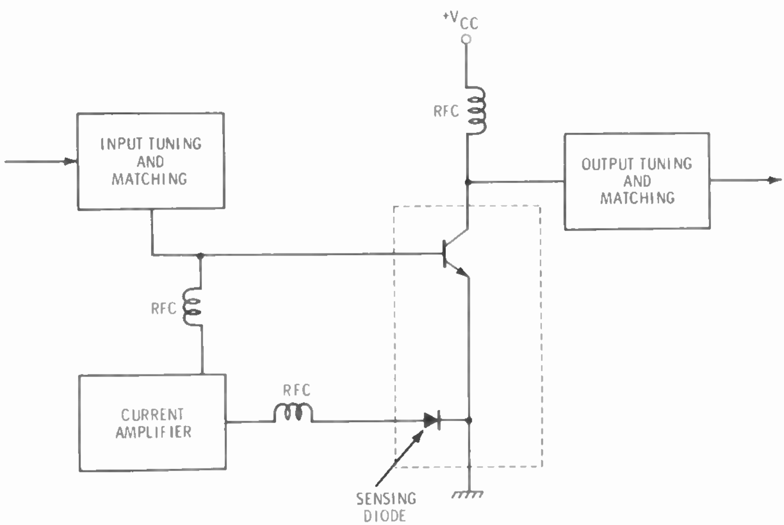


Fig. 6-15. RCA 2N6093 linear bipolar circuit.

susceptibility. Many manufacturers are now paying attention to the design of bipolar types to be used reliably and with good performance in linear rf power amplifier circuits. Types with special internal constructions and compensation diodes are now available that can offer PEP outputs of 150-watts and higher.

An example is the special bipolar circuit of Fig. 6-15, using an RCA 2N6093. A single unit of this type can provide 75-watt PEP on 30 MHz and, in a push-pull configuration, an output capability of 150-watts PEP from 2 through 30 MHz.

This transistor type provides safe operating conditions in the required dc operating region. It can be forward-biased safely into class-AB because of the special subdividing of emitter surface and an appropriate resistive ballast. Added resistance improves stabilization and permits low distortion operation. Included is an internally mounted temperature-sensing diode for bias compensation and runaway protection. It ensures a bias voltage that varies with temperature in the same manner as the base-emitter voltage of the transistor. An external current amplifier builds up the level of the diode sensing, which, in turn, amplifies the adjusting bias current for the base.

PROJECT 26: LINEAR AMPLIFIERS

The construction and tests of a vacuum-tube linear amplifier and mixer are the subjects of Projects 26 and 27. The amplifier and mixer are constructed on a large chassis for ease in construction and making changes. The power supply is built on a separate smaller chassis with a multiconductor cable linking the two segments. PEP input power ranges between 30 and 45 watts. When the units of Projects 22 and 25 are used as a signal source, double-sideband output can be obtained on any band 15 through 160 meters. In Project 27 a

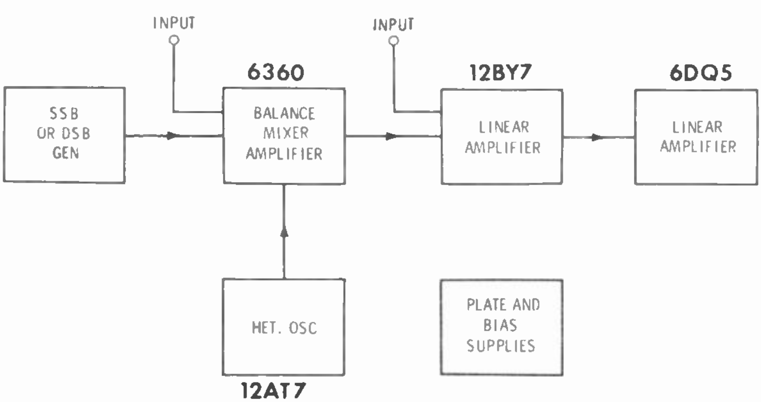
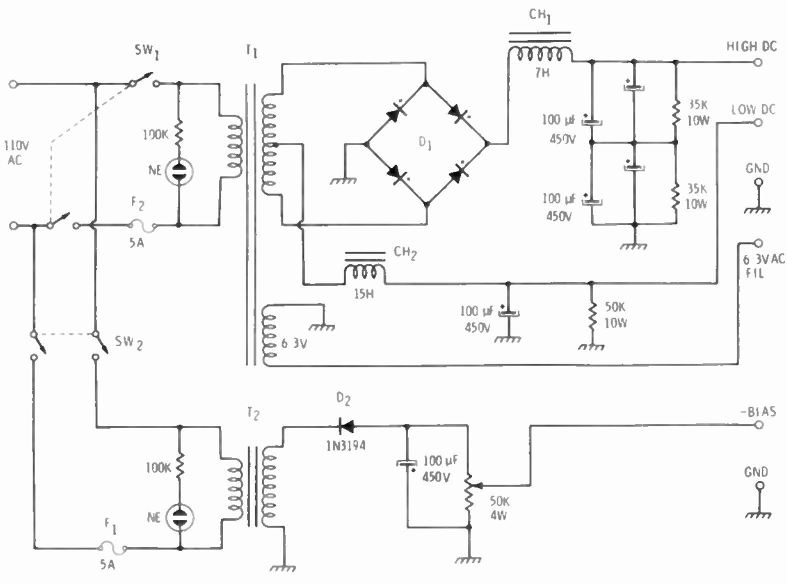


Fig. 6-16. Functional plan of a linear amplifier and mixer.

vfo and mixer are added for single-sideband operation on 160 meters. Later on you may wish to use the units covered in Projects 29 and 30 as sources of signal for the linear amplifier.

A functional block diagram is given in Fig. 6-16. There are two linear amplifier stages using a 12BY7 and a 6DQ5. The 6360 can be used either as a linear amplifier or a balanced mixer. Either two or three linear amplifier stages can be used to match the level of the input signal. The 12AT7 functions as a heterodyne oscillator for 160-meter sideband operation.

A two-section power supply is needed (Fig. 6-17). The plate sup-



Ch_1	7-henry, 75 mA (Triad C8X)	T_2	125V, 50-mA power transformer (Standard PA8421)
Ch_2	15-henry, 75 mA (Standard C-1002)	1	8" x 12" x 3" chassis
D_1	Bridge rectifier, 800 PIV	1	4-terminal power strip.
D_2	IN3194 diode	1	2-terminal power strip.
Ne	Neon panel lights.		
SW_1,SW_2	Dpst power switches		
T_1	270-0-270V, 120-mA power transformer; 6.3V 5A filament. (Standard PC8405).		

Fig. 6-17. Power supply circuit. Resistor and capacitor values are shown on the schematic.

ply voltage section uses a bridge rectifier and a choke-input filter. It generates the high voltage needed by the plate of the final 6DQ5 linear amplifier. When using a bridge rectifier, a one-half total voltage output can be obtained at the center tap of the secondary where a choke-input filter is again used. The transformer also makes available the 6.3-volt filament power.

A second supply employs a simple half-wave rectifier. It develops the negative dc bias voltage for the grid of the final linear amplifier. This bias level can be adjusted by the potentiometer.

The complete circuit for Projects 26 and 27 is given in Fig. 6-18. For two-stage operation only the top two circuits are made active. This operation is controlled by switches S_1, S_2, and S_3. A pentode input stage is employed and includes an adjustable cathode potentiometer that can be used to regulate the drive level to the grid of the final linear amplifier.

No neutralization system is employed. However, binding posts are included at the input of the final to permit you to insert loading resistors that prevent self-oscillation of the final. You can experiment with different values as indicated, adjusting output up to the highest PEP output without self-oscillation. A value of 10K is a good optimum. Include a second set of binding posts in the cathode circuit of the final. A 100-mA dc meter can be inserted at this point. This meter is helpful in adjusting the final bias and in indicating suitable rise in plate current with sideband modulation.

The 6360 tube can be used either as a linear amplifier or a balanced mixer. Switch S_1 must be closed when this amplifier is to be in operation. Its output is applied to the input of the 12BY7 by connecting a jumper between bindings posts 3 and 1. A balanced single-sideband or double-sideband signal is applied between binding posts 4 and 6 for three-stage amplification.

A 12AT7 vfo is included. It tunes to the 2.3-MHz range when 160-meter sideband operation is desired. The vfo plate supply voltage is held constant by an OC3A voltage regulator tube.

Plug-in coils provide multiband operation. The coil information of Project 7 is appropriate for the resonant coil (L_1). The 6360 employs a balanced output and the push-pull coils of Project 22 are appropriate. The pi-network coil (L_2) of the final amplifier is connected across two standoff insulators. Barker and Williamson *miniductor* coils can be used here.

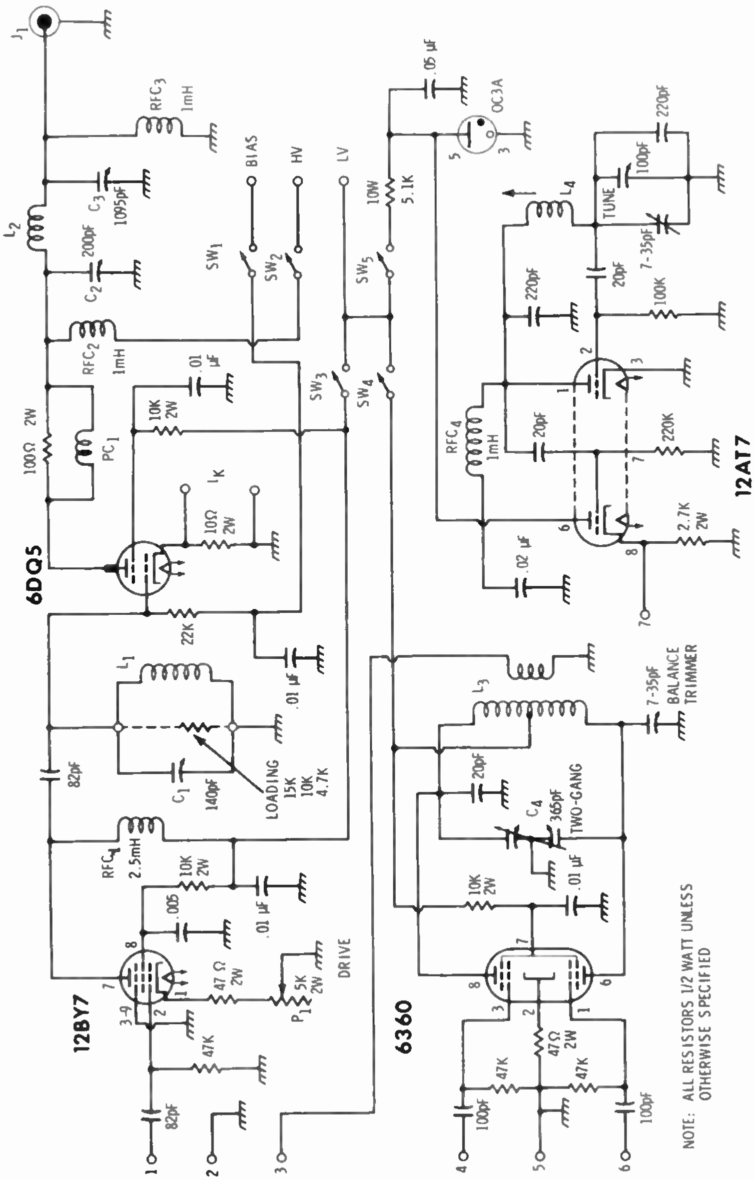
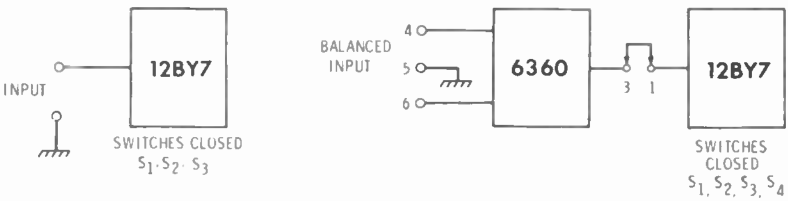


Fig. 6-18. Linear amplifier and mixer. Resistor and capacitor values are shown on the schematic.

Operation

The FET double-sideband generator of Project 25 is a good signal source. Double-sideband operation on any band 15 through 160 meters is possible. Three-stage amplification is required and the 6360 is used as a linear amplifier. The connections for two- and three-stage operation are given in Fig. 6-19. The output of the FET double-



(A) Two-stage. (B) Three-stage.

Fig. 6-19. Input connections for linear operation.

sideband generator is arranged to supply a balanced feed to the 6360. The ungrounded secondary of the double-sideband output transformer of Project 25 is connected to binding posts 4 and 6 of the linear amplifier. A ground jumper must also be connected between the two units; that is, the ground of the double-sideband generator should be connected to binding post 5 of the linear amplifier.

Connect a 50-ohm dummy load to the output and a 100 mA dc meter into the cathode circuit of the final amplifier. On the lower-frequency bands the double-sideband generator supplies good output and the 12BY7 drive control must be adjusted accordingly. Output level is not as high on 15 and 20 meters and full drive setting may be necessary. An oscilloscope or an rf monitor scope such as the

Parts List for Fig 6-18.

J_1	Coaxial receptacle	$SW_{1,2,3,4,5}$	Spst switches.
L_1	Set of coils from Project 7	1	8″ x 3″ x 15″ Chassis
L_2	Set of B&W miniductors, 3016, 3015, 3014, 3013	2	5-prong sockets
L_3	Set of coils from Project 22	2	Octal sockets
L_4	Slug-tuned inductor, J. W. Miller 21A825RB1	3	9-pin miniature sockets
		2	1-inch standoff insulators for L_2
P_1	5K 2-watt potentiometer.		
PC_1	Parasitic choke, 4 turns No. 14 enameled wire wound on a 2W 100-ohm resistor.	2	Power strip terminals, 4-terminal and 2-terminal

Heath SB610 should also be connected to the output. A regular service-type oscilloscope is suitable for monitoring operation on 80 and 160 meters.

Insert an 80- or 160-meter crystal into the double-sideband generator and adjust this circuit in accordance with the information of Project 25 using tone modulation. Next apply supply voltage to the 6360 stage alone. Attach the oscilloscope at binding post 3, resonate the 6360 output, and retouch the dsb generator adjustments.

Now move the oscilloscope to the antenna output. Connect a jumper between binding posts 1 and 3. Turn off the supply voltage to the linear amplifier. Use a grid-dip meter to resonate the C_1-L_1 combination to the desired 80- or 160-meter frequency of the crystal. Do the same for the output resonant circuit with capacitor C_2 after setting C_3 to its maximum position. Insert the 10K loading resistor across C_1L_1.

Switch off the modulated signal at the amplifier input and apply the low and high voltages to the three stages. Adjust the bias potentiometer of the power supply to obtain a cathode current reading of 20mA for the final. There should be no rf output. Apply modulation, slowly increasing the level of the double-sideband signal at the input of the linear amplifier. There should be some indication of output, and the three resonant circuits should be peaked using capacitors C_4, C_1, and C_2. Adjust the output of the double-sideband generator for normal peak output without clipping. Use a 1000- to 1500-hertz tone such as is made available from the Heath SB610 waveform monitor or an audio oscillator.

The various controls must be adjusted to obtain good dsb pattern. If there is severe flattopping it may indicate that there is too much drive to the final and an adjustment of the drive control is in order. If your dummy load includes a wattmeter, a normal reading would fall between 5 and 8 watts output with no severe flattopping.

At output levels higher than this there is usually substantial flattopping and a much greater tendency to self-oscillation. Self-oscillation is indicated by a severe breakup of the normal pattern plus high output and high plate current readings.

Go through the same procedure on other bands. Each band is likely to require some readjustment of the drive control and, perhaps, the value of the loading resistor. Normal cathode-current deflection with modulation is about 50 to 70 mA on peaks.

PROJECT 27: LINEAR MIXER AND HETERODYNE OSCILLATOR

The vfo (Fig. 6-18) has already been constructed on the chassis. It is a 12AT7 dual triode designed to operate on the high-frequency side of the 160-meter band. When mixed with a 455-kHz sideband signal in the 6360 balanced mixer, a 160-meter sideband signal develops in the resonant output circuit of the mixer. The connections for this manner of operation are given in Fig. 6-20. Note that the output binding post (7) of the vfo supplies parallel signals to the grid input circuit of the 6360 through two 50-pF capacitors. The secondary of the mechanical sideband filter is connected across pins 1 and 3, thus supplying a push-pull signal to the mixer input. A ground connection must also be run between the sideband generator of Project 23 and binding post 2.

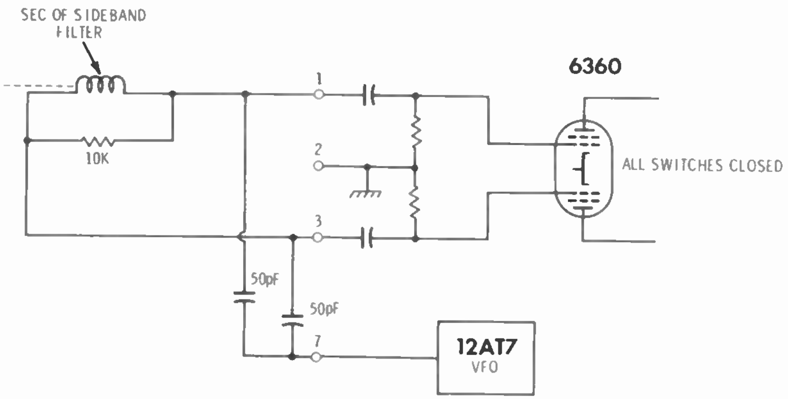


Fig. 6-20. Single-sideband connections.

Inasmuch as the heterodyne oscillator signal is applied in parallel to the two grids it will cancel in the output, provided the output resonant circuit is balanced correctly. A balance control is included to minimize as much as possible any undesired feedthrough of carrier from the oscillator.

The vfo operates on the high-frequency side of the 160-meter band. Lower-sideband operation is customary on 160 meters. Thus the required sideband crystal frequency is 453.5 kHz.

For example, if operation on 1.825 MHz is desired, the heterodyne oscillator frequency must be 2.2785 MHz (1.825 + 453.5). If a 1500-cycle modulating tone is now applied the output of the sideband filter is 455 kHz (453.5 + 1500 hertz). With the heterodyne oscillator on frequency, the output frequency on 160 meters would be 1.8235 *MHz* (2.2785 – 0.455) on the low-sideband side of 1.825 MHz. It is this frequency range that will appear in the output of the balanced mixer. However, the 2.275 component must be rejected in this output circuit.

Operation

Prepare the linear amplifier for operation on 160 meters according to the information given in Project 26. Connect the 6360 in accordance with Fig. 6-19. Use a dip oscillator to adjust the three linear circuits to the desired frequency in the 160-meter band. This should correspond to the 160-meter frequency assignments in your area. On the east coast the 160-meter band is quite narrow extending between 1.8 and 1.85 megahertz. A good adjustment frequency would then be 1.825 MHz.

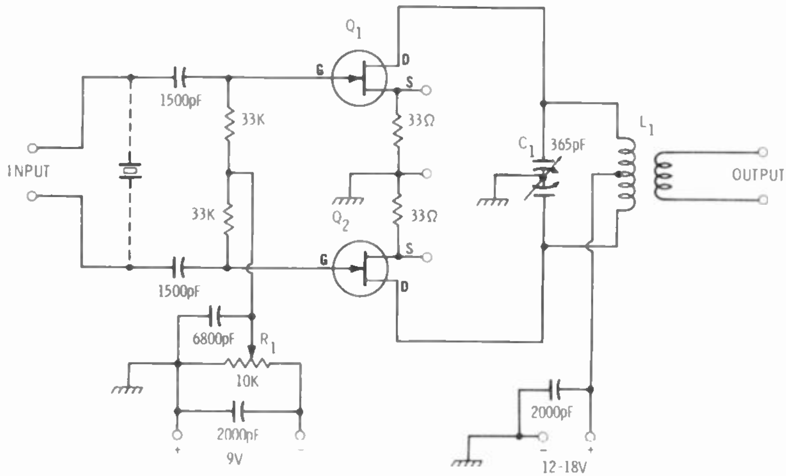
Proper adjustment of the resonant circuits to the desired frequency is important because they could be incorrectly tuned to the heterodyne oscillator frequency, or on the high side of the heterodyne oscillator frequency rather than the low. For example, it could be possible to generate an illegal signal on 2.732 MHz (2.2785 + 0.4535). A reasonably accurate grid-dip meter avoids this possibility.

Turn on the heterodyne oscillator and balanced mixer alone. Adjust your 455-kHz sideband generator for normal operation in accordance with Project 23. Use two-tone modulation rather than the single-tone operation of Project 26. The Heath SB-610 monitor scope provides a two-tone test signal. Apply a properly adjusted two-tone r-f signal from your sideband generator to the 6360 mixer. Connect the oscilloscope between binding post 3 and common. Adjust capacitor C_1 for maximum output. Adjust the balance control for minimum carrier in the output.

After a good two-tone signal is obtained transfer the oscilloscope to the output and connect the jumper between binding posts 3 and 1. Place the two-stage linear amplifier in operation and go through the same adjustment procedure covered in the previous project until you obtain a satisfactory pattern.

PROJECT 28: FET LINEAR AMPLIFIER AND UTILITY OSCILLATOR-AMPLIFIER

The field-effect transistor like a vacuum tube is a high-impedance device which operates well in push-pull circuits. In addition, the gates can be biased toward cutoff to permit linear operation just as you would a vacuum tube. In coming years one can expect the development of higher- and higher-powered FET's. A typical circuit that can be used as an oscillator, class-C amplifier, or rf linear amplifier is shown in Fig. 6-21.



C_1	Dual 365-pF variable capacitor	1	Small circuit board
L_1	Set of plug in coils from Project 22.	2	Transistor sockets
		1	5-Prong tube socket
R_1	10K 2-watt potentiometer	1	Crystal socket
Q_1,Q_2	Siliconix 2N3970 FETs	10	Binding posts

Fig. 6-21. Linear amplifier and utility amplifier-oscillator. Capacitor and resistor values are given in the schematic.

Input and output circuits are balanced. A balanced input signal is a necessity. Typically the outputs of balanced modulators are balanced and the low-powered FET stage can be used to build up signal level after the modulator. The output resonant circuit is a balanced one and the push-pull coils of Project 22 are ideal.

Separate source resistors are used. It may be necessary to make a slight readjustment in the ohmic value of one of the source resistors so as to equalize the drain currents. All but a few of a number of Siliconix 2N3970 FET's provided currents that were reasonably close and two 33-ohm resistors were satisfactory.

Somewhat higher outputs can be obtained using lower value source resistors. In fact, if the two FET's are reasonably well balanced the sources can be grounded to obtain maximum output. However, a heat sink is then recommended. Two 12-ohm resistors provide a good output and the FET's do not run too hot.

Drain supply voltage can be 12 to 18 volts. A 9-volt transistor radio battery supplies bias by way of a 10-K potentiometer. This facility is not required when the amplifier is to be operated as an oscillator or a class-C amplifier.

A crystal socket is connected across the input. This does not interfere with the operation of the unit as an amplifier when no crystal is inserted. When a fundamental crystal is inserted, the stage will operate as a low-power oscillator. Output is one-half to three-quarters of a watt, providing a simple QRP transmitter.

The low-power HEP-801 as well as the high-power U222 FET's operate in the circuit.

Operation

The FET dsb generator of Project 25 is an excellent source of signal. The combination is an all-band FET dsb QRPP transmitter. PEP output swings up close to ¾ watt.

The amplifier can be used to build up the level of signal made available at the output of the 160-meter phasing-type sideband generator covered in Project 30 of Chapter 7. This unit has a balanced output and is ideal for driving the linear.

The output of the linear amplifier should be connected to a 50-ohm dummy load. An oscilloscope can be connected across the same load. The linear bias potentiometer is adjusted to obtain maximum output without any flattopping as seen on the scope.

When the stage is to be operated as a class-C amplifier, the bias battery is disconnected and the arm of the potentiometer is run over to the ground side. When the circuit is to be operated as a crystal oscillator it is only necessary to insert the appropriate crystal. Using the coils of Project 22 and fundamental crystals, it is possible to obtain a good output from 10 through 160 meters using the 2N3970 FETs.

CHAPTER 7

Integrated-Circuit Fundamentals

The integrated circuit is but an extension of the solid-state science of packing active devices (diodes and transistors) into a smaller space. So much reliability and versatility have been built into these devices that there appears to be an infinite number of external circuits and systems to be tried. Wide application in all types of transmitters is a certainty.

Diodes, transistors, and resistors are the only components used in most integrated circuits. A limited number of integrated circuits may include an occasional capacitor or coil. Capacitors are not common because they take up considerable space. It is customary to use a type of internal circuit that does not require capacitance for its operation.

It is difficult to fabricate a resistor of some precise value into an integrated circuit. On the other hand, there is no great problem in including two or more resistors of exactly the same value, even though a certain absolute value is difficult to attain. Hence, internal circuitry employs balanced configurations that require equal-value resistors. but are lenient relative to absolute resistance values. For these reasons the most common integrated circuit is the balanced dc amplifier.

BASIC DIFFERENTIAL AMPLIFIER

The differential amplifier is the mainstay of integrated circuits. It is shown in discrete form in Fig. 7-1 and is basically an emitter-

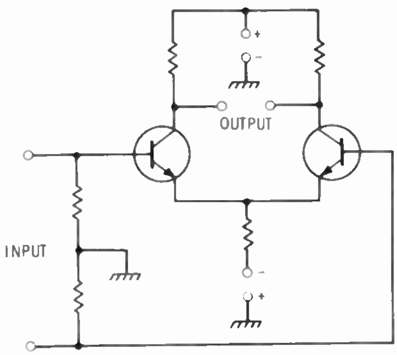


Fig. 7-1. Basic differential amplifier.

coupled configuration. As a dc amplifier it has fine stability and good rejection of undesired signal components. Being a direct-coupled amplifier, no interstage coupling capacitors are needed with the great amount of space they would require.

Ideal differential operation requires that the two collector resistances be the same and the characteristics of the two transistors be identical. In terms of discrete component circuits this is a disadvantage because perfectly matched transistors and resistors are necessary. However, in IC production these conditions are met quite readily and at low cost.

In basic operation the differential amplifier emphasizes the signal difference that exists between base inputs, developing equal-amplitude and out-of-phase collector signals. It is stated that a differential-mode input signal is being applied. In practice it is done by applying the desired ac signal to just one of the base inputs. Since no signal is applied to the opposite base, the difference voltage between the two equals the magnitude of the signal applied to the one base.

When two equal-amplitude but same-polarity signals are applied to the base inputs, the ac signals across the common emitter resistor are subtractive. In a situation of perfect balance the differential amplifier performs in bridge-like manner, and there is no output observed from collector to collector and very reduced output from each collector and common. Such an applied signal is referred to as a *common-mode* input signal. This is usually in the form of an undesired signal such as hum and interference components which are to be eliminated.

In the *difference-mode* operation a signal applied to base 1 appears at the collector output of transistor 1 and also across the common emitter resistor. The latter signal components serves as the input signal for transistor 2. As a result an opposite-polarity signal variation appears at the collector of transistor 2. The differential amplifier acts as a phase splitter, developing two equal-amplitude but opposite-polarity signal components at the output.

The differential amplifier has a high order of dc stability, reducing the influence of supply voltage changes, temperature, etc. A more stable dc amplifier results and it is very practical to construct a multistage affair using the difference concept. A differential amplifier or a group of them connected in special cascade arrangements are the most common circuit configurations built into ICs.

It is interesting that not only are the interstage coupling capacitors eliminated, but the emitter bypass capacitors as well. In making a comparison between ICs and discrete circuits, it is to be noted that an integrated circuit has fewer passive components (resistors, capacitors, and coils), and more active components (transistors and diodes) than a comparable amplifier made of discrete active and passive components.

STABILITY

It is true that in a perfectly balanced differential amplifier there is stable amplification with changes in dc operating conditions and temperatures. A change in leakage current and/or gain in one side of the differential circuit is balanced out by a like change in the second side. Such balance and ability to compensate for any imbalance that does arise sets the operating limits of the differential amplifier in practical usage.

In the reduction of common-mode signals one depends on the degenerative effects of the common-emitter resistor. Of course, the higher the ohmic value of this resistance, the greater is the rejection. Such increase is limited by supply voltage requirements and the greater difficulty involved in including high-value integrated resistors.

CONSTANT-CURRENT SOURCE

The answer to this problem is to include a constant-current emitter source composed of an additional active component rather than a high value resistance. The fundamental arrangement is shown in Fig. 7-2. In this circuit the combination of the transistor and its low-

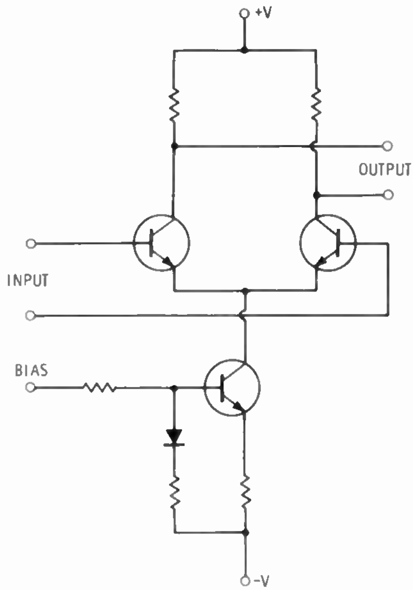


Fig. 7-2. Constant-current bias source with temperature-compensating device.

value emitter resistor acts as a high-resistance constant-current source. The presence of a common-mode signal on the differential transistors will affect base voltages and junction resistances. However, emitter and collector currents are held constant by the constant-current emitter source. In fact, the undesired voltage change appears totally across the constant-current source circuit which is highly degenerative. Thus the differential gain of the amplifier in terms of common-mode signals is greatly reduced.

The diodes associated with the base circuit of the constant-current source provide temperature compensation. Exact compensation is obtained when characteristics of the base-emitter junction of the constant-current transistor and the diode junction are identical. With a rise in temperature there is an increase in the conductance of the base-emitter junction. Inasmuch as the compensating diode junction is physically near to the transistor there is a similar change in its conductance and a compensating change is made in the base bias keeping the collector-emitter current constant. The circuit of Fig. 7-2 is a very common integrated circuit configuration.

DARLINGTON CONFIGURATION

The differential amplifiers of Figs. 7-1 and 7-2 have low input impedances. High input impedances, if desired, can be made a part of ICs by using Darlington circuitry which involves the addition of two more active elements. A simplified Darlington combine is shown in Fig. 7-3; and are associated with an integrated circuit differential amplifier in Fig. 7-4.

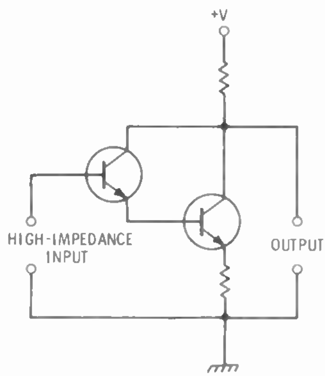


Fig. 7-3. Basic Darlington pair.

In the normal operation of a transistor the emitter junction (base-emitter junction) is forward-biased and conducts. Resistance is low and approximates the product of beta times the emitter resistance. To some degree the input resistance can be increased by increasing the ohmic value of the emitter resistance at a sacrifice in gain. A better approach is to use the input resistance of a second transistor as the emitter resistance of the first transistor. The input stage then operates with a highly degenerative emitter circuit and consequent high input resistance. Both stages contribute output with a gain figure that is comparable to that obtained using a single transistor of the same type but operating with a much lower input resistance. Two such identical circuits are needed for the two separate inputs of a differential amplifier.

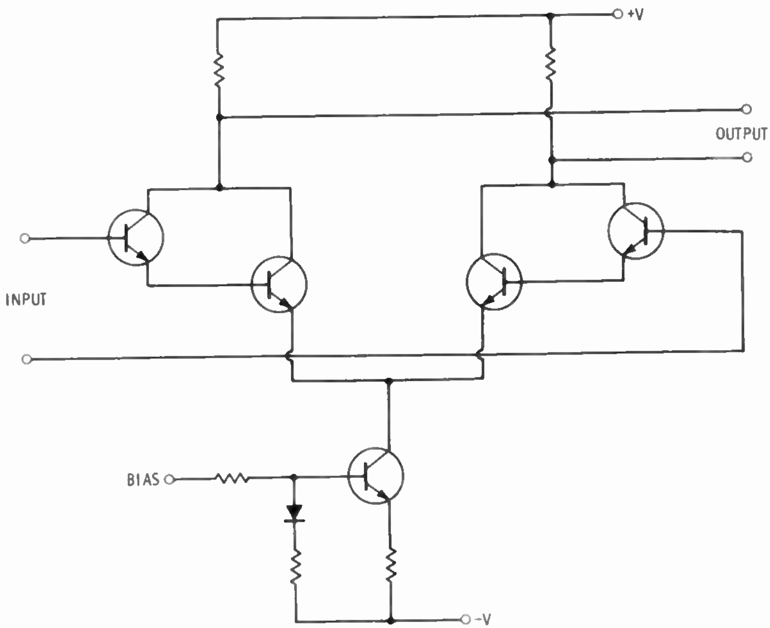


Fig. 7-4. Differential amplifier with Darlington pairs.

A BASIC IC

The Motorola HEP-580 is a simple low-cost integrated circuit composed of six resistors and four transistors. Internally the transistors are connected in pairs with separate base inputs (Fig. 7-5). All emitters are joined together at pin 4. It is a basic differential amplifier configuration using paired transistors instead of singles. If desired, an external stabilizing constant current source can be added at pin 4. Collector load resistances of equal value (3.6K) are internal. Series base resistances increase the input resistance, reduce the tendency toward parasitic oscillations, and also provide additional isolation.

Two ways that integrated circuits are depicted schematically are shown in Fig. 7-6. In A, a differential circuit is arranged around base pin designations of the IC. The triangular arrangement of B is more common and more instructive because the circuit layout can be set down with well defined input and output sides regardless of pin numbers.

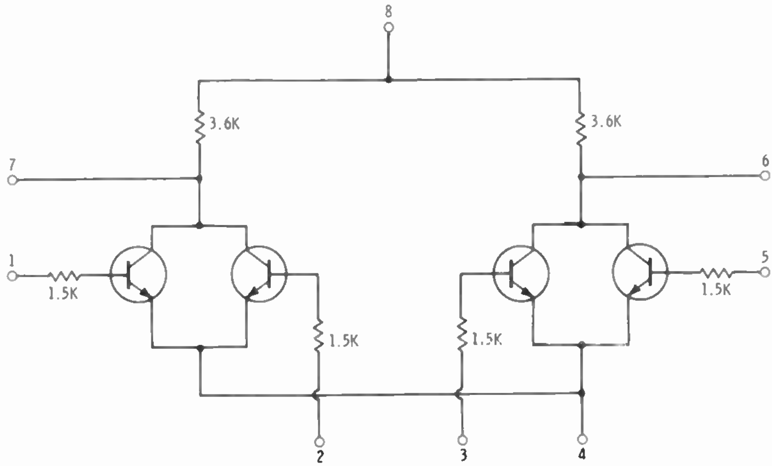


Fig. 7-5. The Motorola HEP 580 integrated circuit.

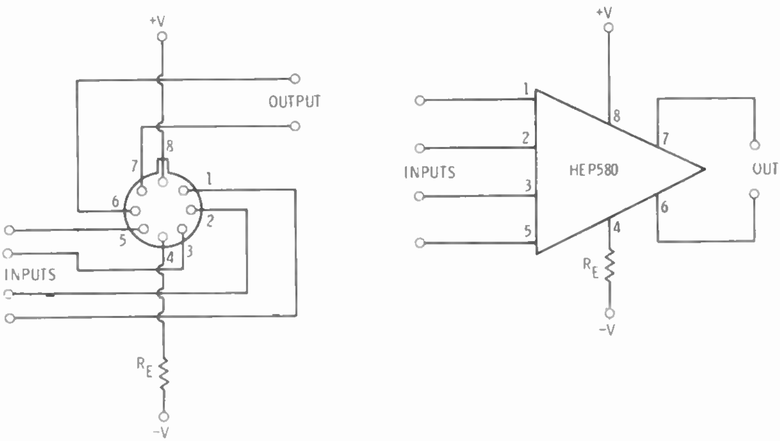


Fig. 7-6. Two methods of representing the HEP 580.

100-MILLIWATT IC RF GENERATOR

A 100 milliwatt QRP transmitter can be constructed from two HEP-580's (Fig. 7-7). The first section of one of the ICs operates as a crystal oscillator; the second section as a phase inverter. Choke output is employed and approximately equal-amplitude and opposite-polarity rf signals are available for driving the output IC. The output IC operates as a push-pull amplifier. It will draw 20 to 30 milliam-

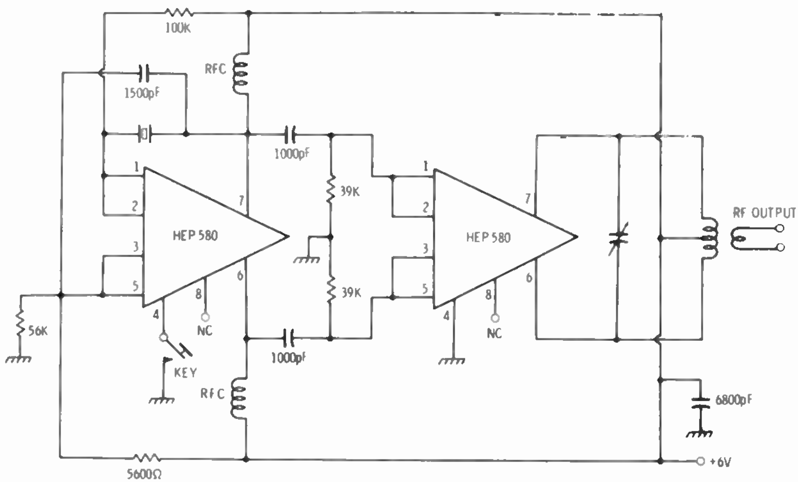


Fig. 7-7. Integrated-circuit transmitter has output of 100 milliwatts on 40 meters.

peres from a six-volt lantern battery. Dc input power is 120 milliwatts or higher.

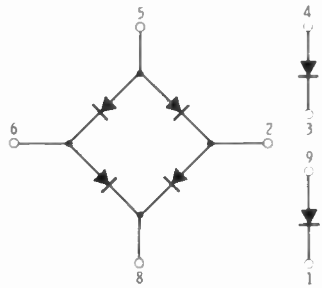
MODULATORS AND DEMODULATORS

Integrated circuits hold great promise in the realms of modulation and demodulation. The modern transmission modes of single sideband, double sideband and suppressed carrier, and frequency modulation, plus phase-lock receiving systems depend a great deal on balanced circuits. Integrated circuits with their arrays of identical diodes, bipolars, and FETs provide this balance.

The diode balanced modulator and diode balanced mixer are popular transmit circuits. Diode-array integrated circuits provide the experimenter with a small and well-balanced diode assembly. One example is the RCA CA3019 (Fig. 7-8). It consists of four diodes connected in a bridge plus two additional diodes all in a TO-5 10-pin case. A conventional balanced modulator and a ring modulator circuit are shown in Fig. 7-9. Diode capacitances are low (1.8 pF) and uniform, thus ensuring good high-frequency balance.

An attractive single-sideband generator with good carrier suppression and without any associated resonant circuit is the Motorola

Fig. 7-8. Diode configurations of RCA-CA3019.



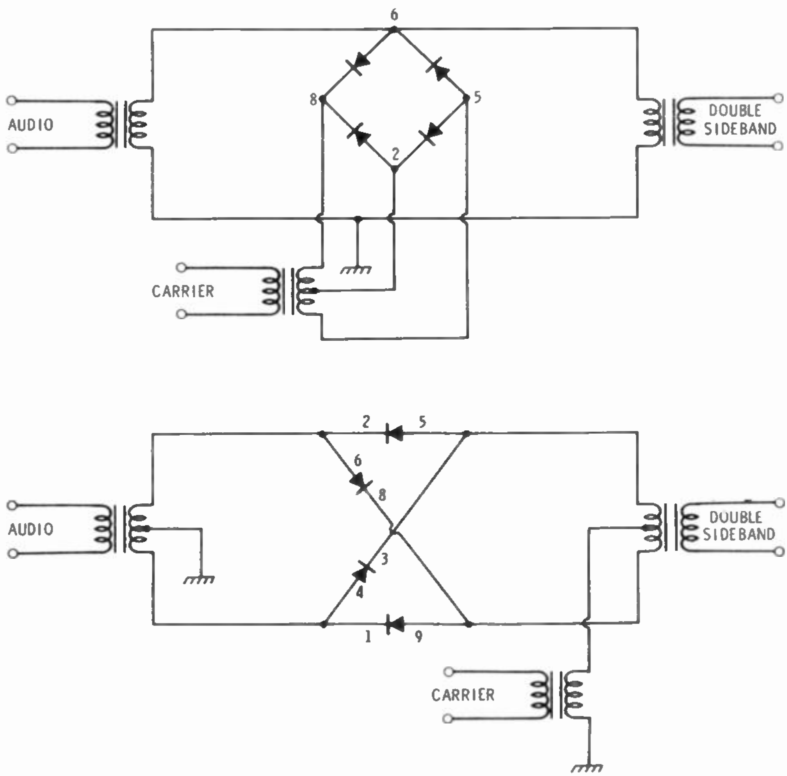
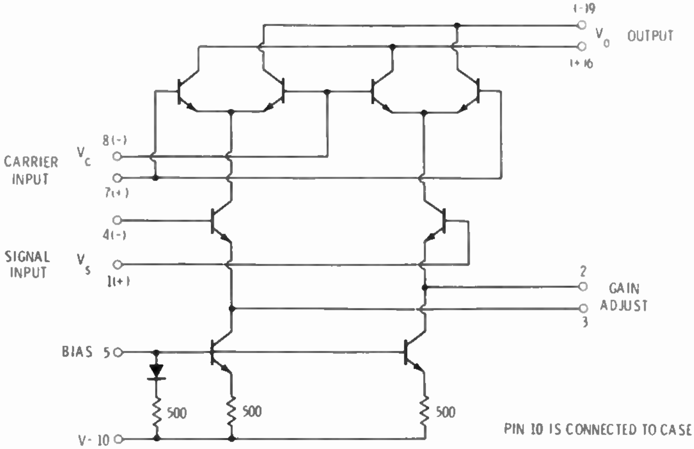
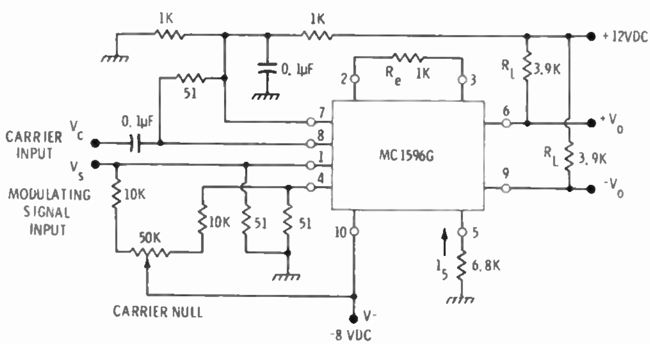


Fig. 7-9. Diode balanced-modulator circuits.

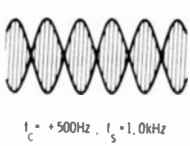
MC-1596G shown in Fig. 7-10. As a function of modulating-signal amplitude, sideband outputs of ½ to 1 volt rms can be obtained with an input carrier level of 100 mV. A solid-state vfo or crystal oscil-



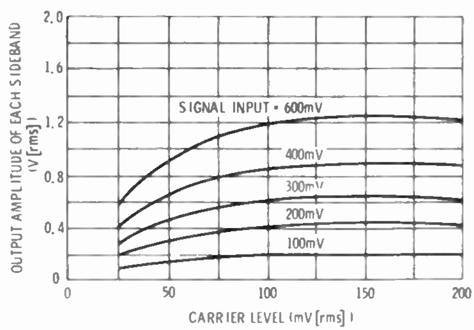
(A) Internal components of MC-1596G.



(B) Double-sideband suppressed-carrier generator.



(C) Suppressed-carrier waveform.



(D) Single sideband output level.

Fig. 7-10. Balanced modulator in an integrated circuit.

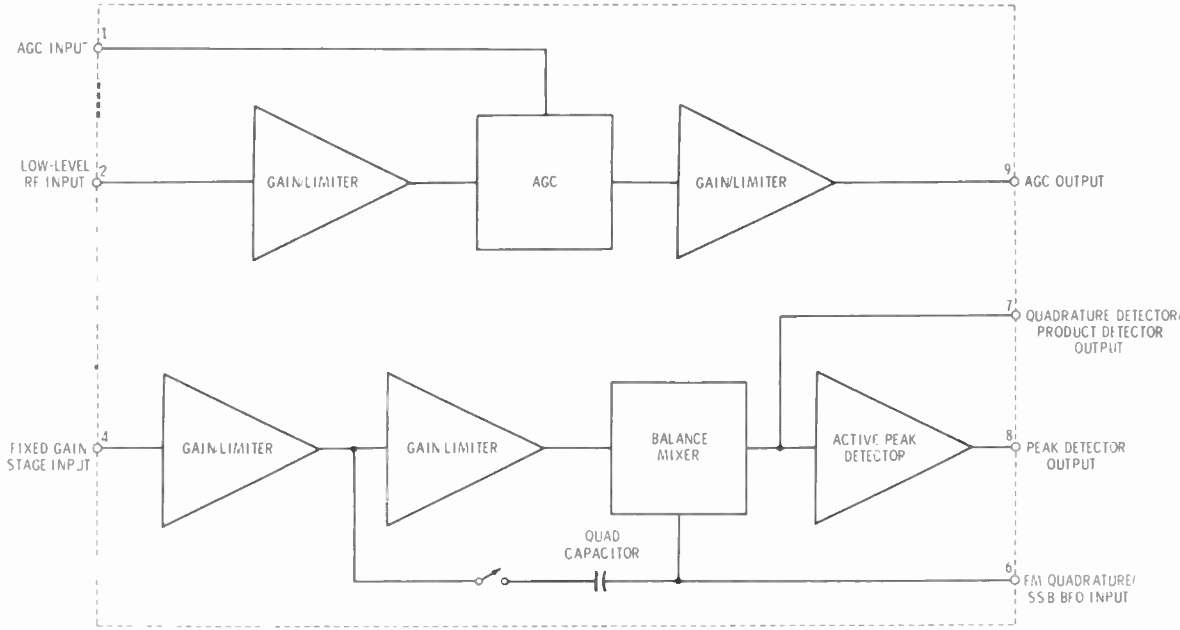


Fig. 7-11. Functional diagram of LM 373.

lator will supply more than adequate drive. Required audio drive is about ½ volt rms for the circuit suggested by Motorola.

Internal circuit configuration and external circuit plan for a double-sideband suppressed carrier generator are given in Fig. 7-10. Two differential amplifier pairs are included and incorporate individual transistors in their common-emitter circuits to supply constant-current bias. A second transistor is included in each leg for injecting the modulating signal. Carrier is applied in differential-mode

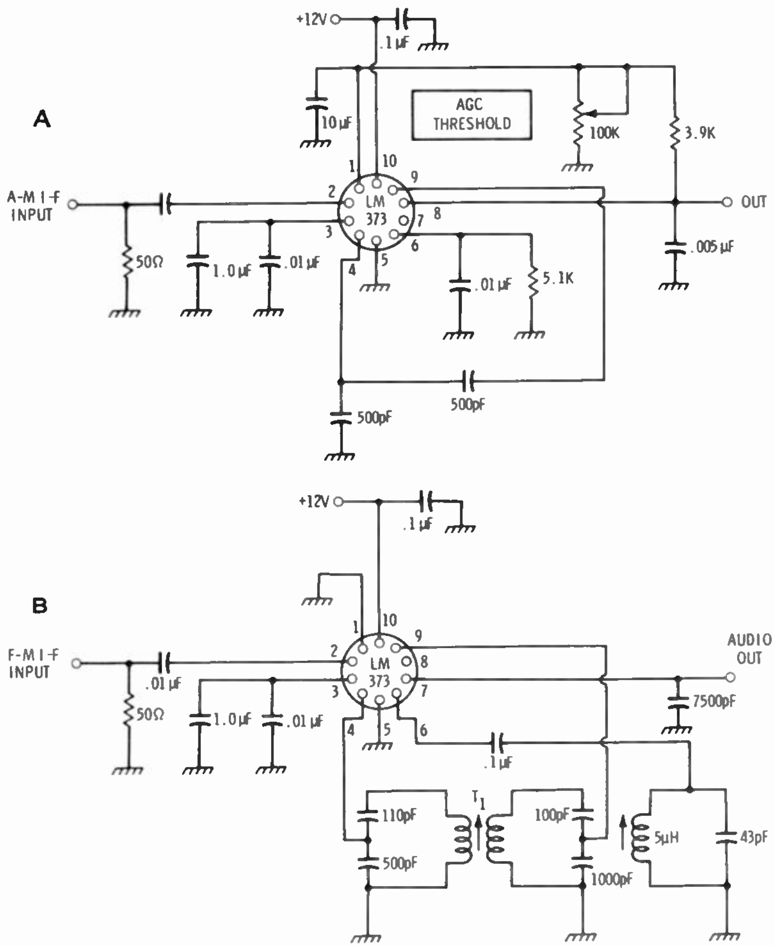


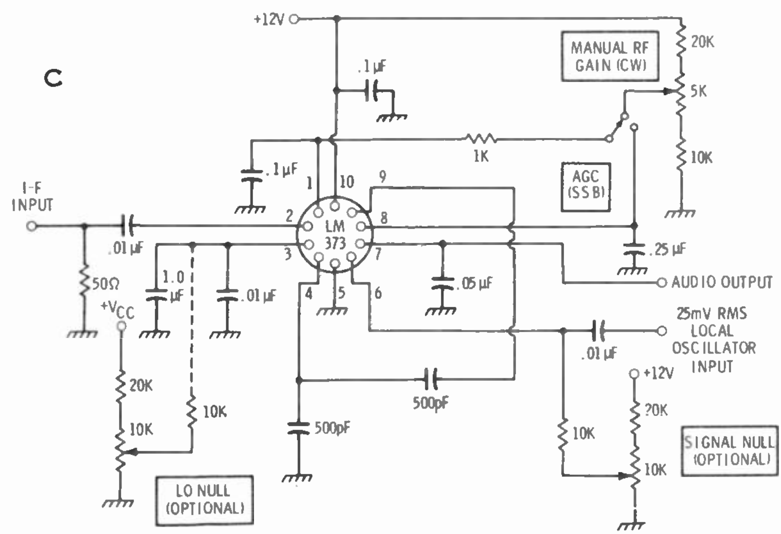
Fig. 7-12. Demodulation

fashion to pairs of differential transistors. Outputs of the differential pairs are out of phase and under true balance the net carrier voltage is zero. Out-of-phase audio is applied across the bases of the two modulation-insert transistors located in the emitter legs of the differential pairs. Upper and lower side frequencies develop across the output while the modulating wave cancels.

Note in the external circuit that the carrier is applied between pins 8 and 7. The modulating signal appears between pins 1 and 4. Biasing for these latter two transistors is obtained from the −8 volt source connected to the arm of the carrier null potentiometer. This biasing, because of its link to the emitters of the differential pairs sets their bias and permits an appropriate adjustment for balancing out the carrier.

A balanced output is available between pins 6 and 9; single-ended output can be derived from between either pin and common. Excellent carrier rejection is obtained using the proper levels of carrier and modulating wave.

Some of the most intriguing IC devices are those that contain an i-f amplifier plus a capability of various modes of demodulation. They lend themselves to the construction of small and versatile receiving



modes of the LM 373.

systems. With a suitable front end and switching facility a receiver could be evolved for direct conversion or conventional superhet operation. A compact all-band (1.8 MHz to 144 MHz and perhaps higher), all-mode (cw, a-m, f-m, and sideband) receiver becomes feasible.

One such device is the National Semiconductor LM373 a-m/f-m/ssb i-f strip. Simple external connections can switch over the IC from one demodulation mode to another. Bandpass shaping may be accomplished externally over a frequency range from audio up to 15 MHz. Here is the possibility for either a direct-conversion or conventional i-f amplifier receive selection. Agc capability is also included.

A functional block diagram of this IC (Fig. 7-11) shows a row of stages at the top which includes amplifier/limiter, agc facility, and a second gain circuit. The signal to be demodulated is applied to pin 2. Output is removed at pin 9 and is, in most applications, reintroduced at pin 4, passing through an external filter circuit on the way. More gain stages follow with the output being applied to a balanced mixer and/or peak detector. These latter two stages with proper external circuitry can be switched for demodulating a-m, f-m, or ssb. Output voltage extends between 50 and 120 millivolts rms depending on the receive mode.

Circuit connections for the various demodulation modes are given in Fig. 7-12. An a-m i-f signal is applied to pin 2 (Fig. 7-12A), removed at pin 9 and reintroduced at pin 4. In this case demodulated output is removed from the peak detector at pin 8.

The f-m detector mode is shown in Fig. 7-12B and uses a quadrature demodulator. Again the f-m i-f signal is applied to pin 2, is picked up at pin 9, and through selectivity transformer T_1 is reapplied to the IC at pin 4. The quadrature LC circuit connects at pin 6 and demodulated audio is removed at pin 7 at the output of the balanced mixer.

The ssb/cw mode is given in Fig. 7-12C. Intermediate frequency signal follows the same path, being reintroduced at pin 4 after amplification. The balanced mixer now operates as a product detector with demodulating carrier being introduced at pin 6. Output is again taken off at pin 7. An appropriate switch permits selection of agc action for single-sideband operation or manual gain control for cw reception.

IC RF DIVIDERS

Such devices as flip-flops, divide-by-12 counters, divide-by-16 counters, 4-bit binary counters, etc. have found their way into transmitters, both a-m and f-m.

A popular a-m broadcast transmitter uses two two-to-one binaries (Fig. 7-13). The crystal oscillator frequency can be set to the stable

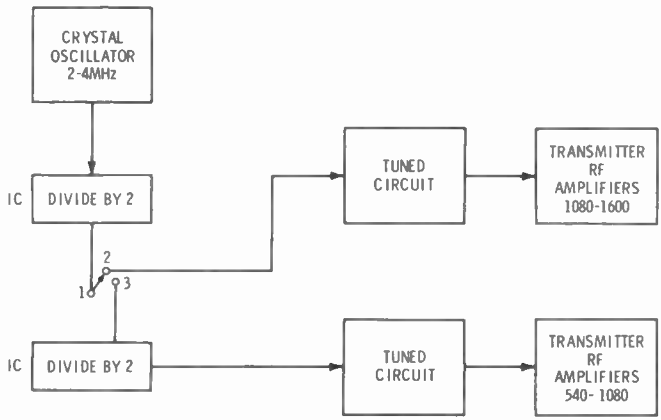


Fig. 7-13. Circuit for using IC frequency dividers in a-m broadcast transmitters.

frequency range in the 2- to 4-MHz spectrum. If the broadcast transmitter is to transmit between 540 and 1080 kHz, two 2-to-1 counters provide a net count down of four. If the transmitter must operate between 1080 and 1600 kHz, a single 2-to-1 counter is inserted to divide down the frequency.

Digital IC's of this type generate square waves but this is no problem. An output resonant circuit of appropriate *Q* can change over the square wave to a pure sine wave. The input to the counter should also be squared and this can be accomplished with a simple limiter circuit at the output of the crystal oscillator.

Some of the modern f-m transmitters use elaborate phase-locked digital counter chains (Fig. 7-14). Such arrangements permit the frequency-modulated oscillator to operate at the fundamental frequency of the f-m broadcast station. An automatic-frequency phase-lock system keeps the oscillator on the assigned frequency well within the

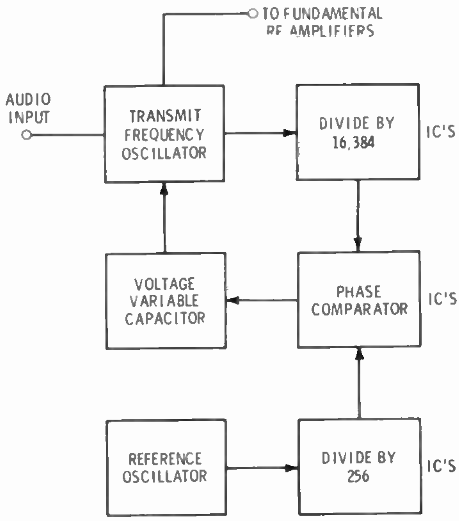


Fig. 7-14. Integrated circuits in a phase-lock system determine the operating frequency of f-m broadcast transmitter.

FCC tolerance. (This tolerance is ±2000 Hz on the f-m broadcast band.)

In one f-m broadcast transmitter an output from the basic oscillator is applied to a 16,384-to-1 counter and then to a phase comparator. A crystal reference oscillator operates in the 1.5- to 2-MHz range depending on the assigned frequency of the f-m broadcast station. This frequency is divided down by 256-to-1 and applied to the phase comparator. The phase comparator develops a dc reference voltage that keeps the transmit oscillator on frequency.

In such a divider system, digital IC's must be selected that operate at very high frequencies. Such units are available at low cost. What are some of the possibilities for communications transmitters? We are accustomed to using multipliers that developed frequencies related harmonically to a crystal oscillator or vfo. A digital divider system permits us to drop down in frequency. For example, a divide-by-two counter following a 40-meter crystal oscillator or vfo that tunes between 7 and 8 MHz can give us an 80-meter output between 3.5 (7/2) and 4 (8/2) MHz. Two 2-to-1 dividers would provide a net division of 4 and put us into the 160-meter band (Fig. 7-15). Digital IC's that can be connected as stable multivibrators can also be made

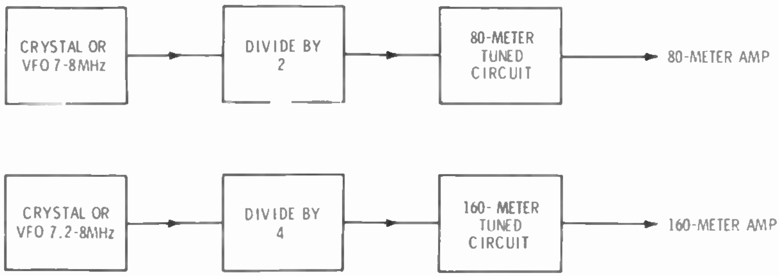


Fig. 7-15. IC dividers in amateur equipment.

to multiply. Is it possible that we could find a substitute for the common tube multiplier chain? The arrangement of Fig. 7-16 then might be feasible as a source of an all-band signal or, perhaps, as an accurate all-band signal source and calibrator. Twice a 7.5-MHz master-oscillator frequency provides a WWV check point at 15 MHz.

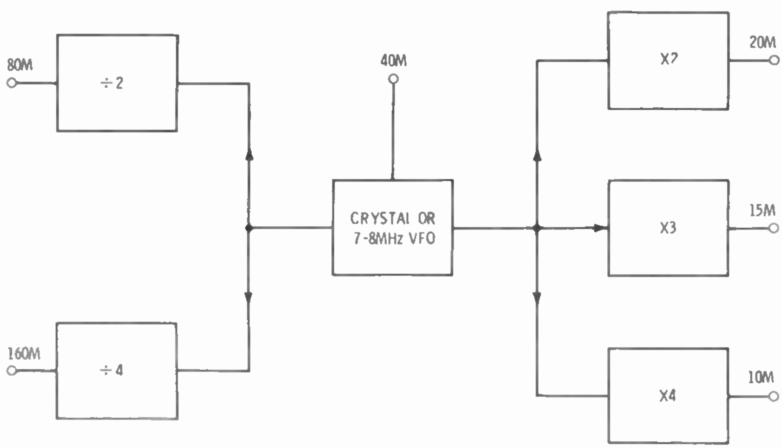


Fig. 7-16. All-band signal source using 7-MHz vfo.

Reconstruction of a sine wave from the square-wave outputs is no great undertaking with the use of resonant circuits or multisection integrators. One can expect some problems. We would be dealing with pulse waveforms and their multi-frequency makeup. Therefore proper shielding and grounding is important as in any computer system.

Features of the plan are attractive for all-band QRP operations. It matches various types of QRP modes such as cw, a-m, dsb, and phase-type ssb. There is also merit in terms of the direct-conversion receiver because one can work down frequency and up frequency from a single high stability receive oscillator.

Relative to vhf-uhf operations, IC dividers present the possibility of having the transmit oscillator operate on the transmit frequency, doing away with the multiplier chain. Possibilities for 2- and 6-meter operation are shown in Fig. 7-17. A divide-by-eight system would bring a fundamental 6-meter signal down in the 6- to 7-megahertz range where it could be compared with a crystal oscillator, vxo or very stable vfo. The two signals would be applied to a phase com-

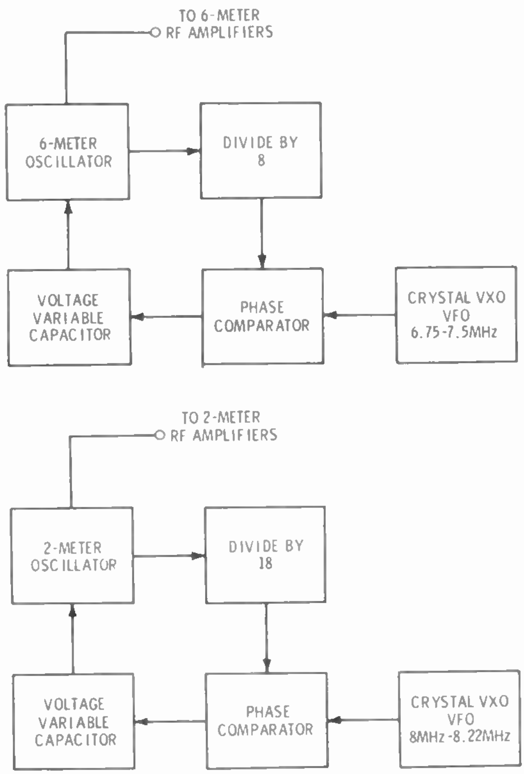


Fig. 7-17. Using IC frequency dividers in frequency-determining system for 6 and 2 meters.

parator (also available as a digital IC). A dc afc component is developed for application to the fundamental-frequency oscillator, holding it on frequency. The fundamental-frequency oscillator can be a FET or bipolar circuit and employ a voltage-variable capacitor diode that responds to the phase-lock afc system.

In terms of 2-meter operation an 18-to-1 divider brings the fundamental frequency down to the 8-MHz range for comparison. Otherwise the system would be the same as for 6-meter operation. Again the plan is adaptable to various modes of modulation. Even frequency modulation can be accommodated. However, division must be greater to prevent the modulating frequencies from affecting the phase-lock operation.

SIMPLE RADIO-FREQUENCY DIVIDER

The plan is proven with a low-cost plastic-case MDTL clocked R-S flip-flop, wired externally to operate as a J-K flip-flop. The frequency divider was inserted into a FET circuit.

The output of the Pierce crystal oscillator (Fig. 7-18) is fed directly to the clock input of the flip-flop. A shaping diode is connected between terminal 2 and common. The flip flop is biased with a 9-volt transistor radio battery.

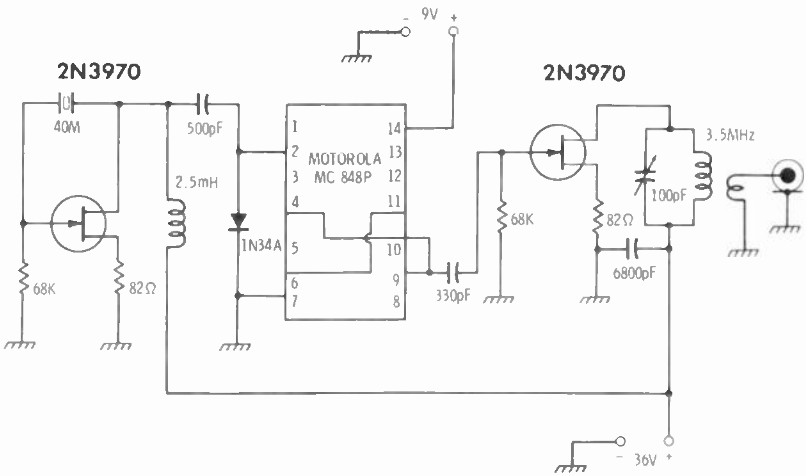


Fig. 7-18. QRP transmitter for 80 meters uses a 40-meter crystal and an IC 2:1 frequency divider.

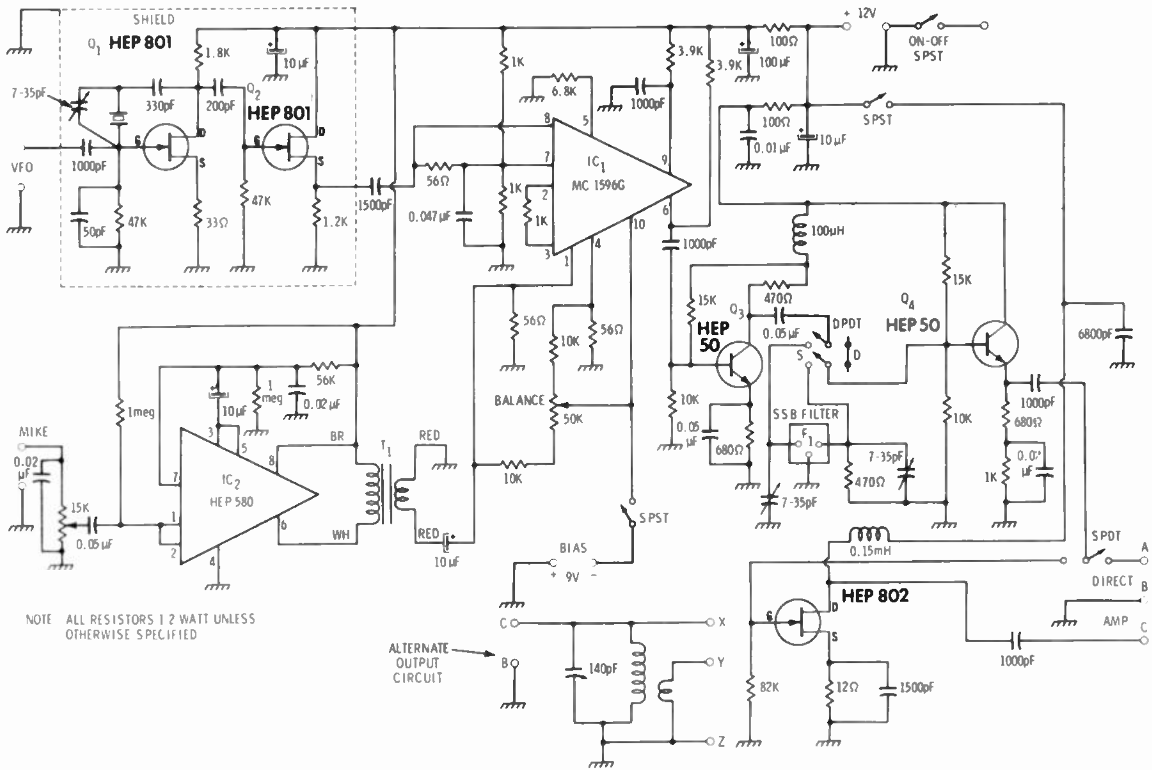


Fig. 7-19. IC dsb-ssb generator. Capacitor and resistor values are shown on the schematic.

A tuned r-f amplifier follows. It is operated class-A because of the limited drive from the flip flop. Nevertheless approximately 200 milliwatts of rf output was developed. More importantly a beautiful 80-meter sine wave is observed at the output. A local 80-meter QRP contact using a 40-meter crystal is something a bit different.

PROJECT 29: IC DSB-SSB GENERATOR

The integrated-circuit generator of Figs. 7-19 and 7-20 is a versatile exciter for both double-sideband and single-sideband operation. It will operate as a double-sideband generator on 10 through 160 meters, either crystal or vfo controlled. It also provides a 9-MHz single-sideband output that can be used as an exciter for a sideband transmitter. This exciter plus the mixer and linear amplifier combination of Projects 26 and 27 can give you an all-band sideband transmitter.

The shielded FET crystal oscillator and buffer are shown at the top left of Fig. 7-19. One need only plug in an appropriate crystal for supplying a carrier component for double-sideband generation. For vfo operation, be certain no crystal is inserted. The carrier oscillator signal is applied to pin 8 of the balanced modulator.

When operating sideband an appropriate 9-MHz crystal must be inserted. A small 7-35 trimmer capacitor should be placed across the crystal, permitting you to set the crystal oscillator precisely on frequency to correspond with the characteristics of your 9-MHz crystal sideband filter.

The microphone signal is amplified by an integrated circuit composed of two differential amplifiers. An interstage transistor audio transformer conveys the modulating component to pin 3 of the balanced modulator. A carrier balancing circuit is incorporated between pins 1 and 4. Included are a bias battery and switch.

The output of the balanced modulator is applied to a two-stage

Parts List for Fig. 7-19.

1 Chassis 7" x 10" x 2"
1 10-pin in-line IC socket
5 Transistor sockets
1 Minibox (for oscillator)
F_1 9-MHz sideband filter (Spectrum international XF-9A)
T_1 Modulation transformer, Pri. 3000 ohms, Sec. 500 ohms center-tapped. (Lafayette 99P-61327)
14 Binding posts
3 Sideband and carrier crystals (8998.5, 9000.0, and 9001.5 kHz).

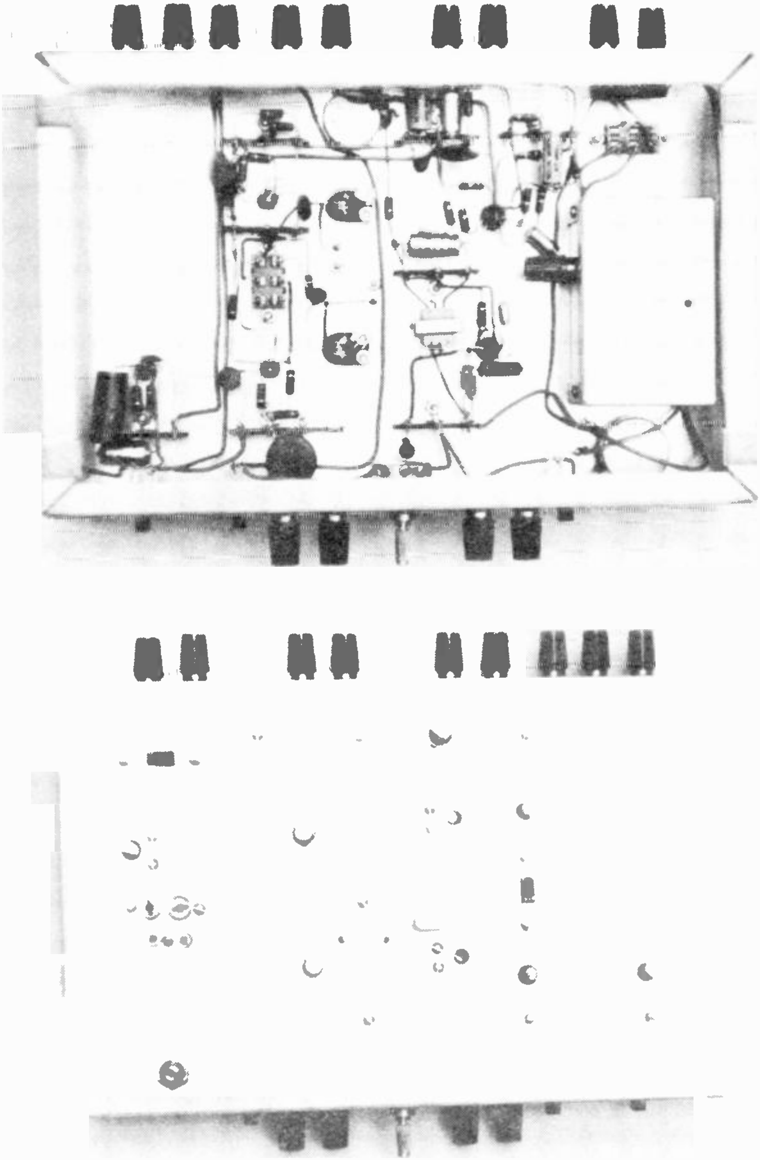


Fig. 7-20. IC ssb generator.

bipolar transistor amplifier. The crystal sideband filter is located between these two stages. It includes input and output trimmer capacitors for resonating the filter to exactly 9 MHz.

A double-pole, double-throw switch located between stages permits the selection of either single-sideband or double-sideband operation. This facility permits you to switch around the sideband filter when double-sideband operation is desired.

A succeeding FET amplifier follows the sideband filter circuit. This amplifier can be switched in or out of the circuit. Since there is some attenuation in the filter this stage is helpful in providing good drive when operating on the sideband mode. For double-sideband operation the output can be taken directly from the second HEP-50 and, the HEP-802 amplifier can be turned off.

An addition of a resonant circuit and output coupling provides a higher output (Fig. 7-19). Coil data of Figs. 1-14 and 2-7 are appropriate. High impedance drive can be taken off the primary; low impedance, from across the secondary. The latter permits feeding an antenna system with a QRPP dsb signal. The resonant circuit can also be used to step up the level of a 9-MHz ssb signal when required. Regular untuned output can be obtained simply by removing the coil and setting the capacitor to minimum capacitance.

Operation

The unit is not difficult to tune. Plug-in an appropriate crystal when double-sideband operation is desired. Apply no audio, and connect the oscilloscope to the direct output of the unit. The oscilloscope must be able to amplify at the crystal frequency. Adjust the balance control for minimum carrier output. True setting is of course the minimum between a rise in carrier output on either side.

You can now apply a tone signal to the input or a microphone signal. The characteristic double-sideband output can now be observed and you can adjust the audio gain control to the point at which peak flattening begins.

Several additional adjustments are required in tuning up the generator for single-sideband operation. A 9.000-MHz carrier crystal is helpful. Plug in the carrier crystal. Tune in the signal on a receiver that has been set exactly to 9 MHz using a crystal calibrator and WWV signal. Adjust the 7-35 pF trimmer to zero-beat the crystal oscillator on exactly 9 MHz.

Adjust the balance control for minimum output and the two trimmer capacitors of the crystal filter for maximum output. In this adjustment one must make certain not to place a capacitive load of too high a value on the output of the second HEP-50. In order to make sure that it is not too high, connect your oscilloscope across the amplifier output rather than the direct output.

Remove the carrier and insert either the 8.9985-MHz or 9.0015-MHz crystal, depending on whether upper or lower sideband operation is desired. Remove the audio for the moment and adjust the balance control a little bit to make certain that the carrier output is minimum. Reapply an audio tone to the input. Output should now be of constant amplitude with little or no ripple; refer to Fig. 5-4. A two-tone test signal is particularly helpful in adjusting sideband equip-

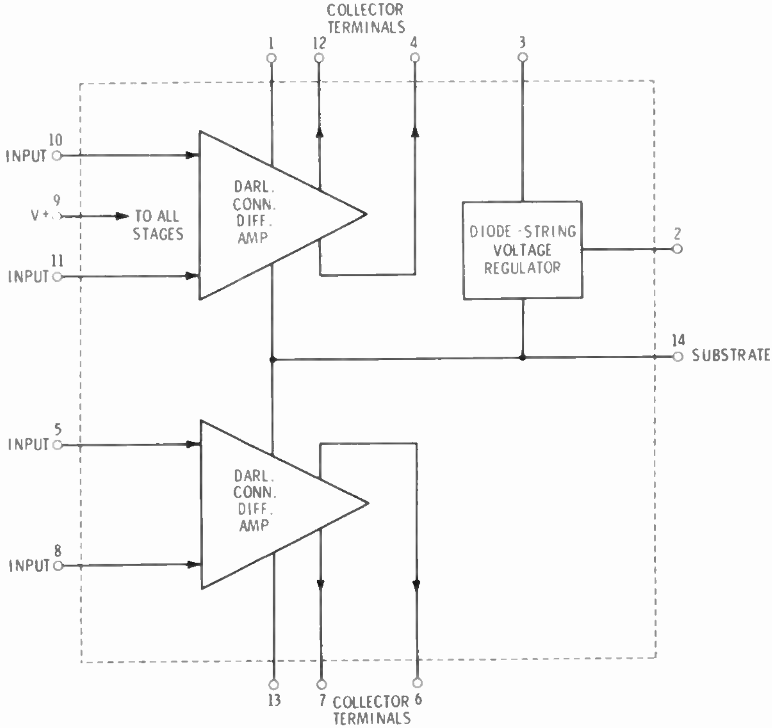


Fig. 7-21. Dual differential amplifier can be used as a doubly balanced modulator.

ment. If you own or can borrow a SB-610 Heath monitor scope you will find that both single-tone and two-tone test signals are made available. Refer to Chapter 10 for further discussion and typical waveforms.

PROJECT 30: 160-METER PHASING SSB GENERATOR

The RCA CA3050 linear integrated circuit consists of two differential amplifiers with Darlington inputs (Fig. 7-21). Such a combination functions ideally as a double-balanced modulator capable of generating a single-sideband signal using the phasing method. A complete sideband generator can be built in conjunction with the RCA linear integrated circuit CA 3018 (Fig. 7-22) which can provide the 90°-related audio drive for the phasing method. The lower set of transistors in Fig. 7-22 are connected as a Darlington pair by joining

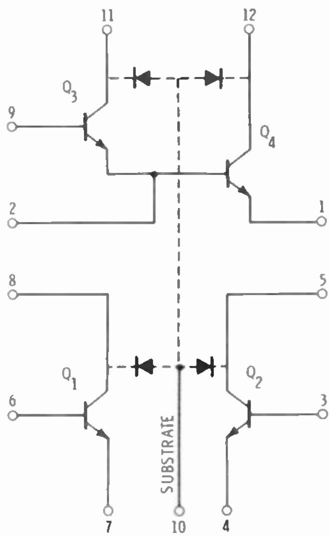


Fig. 7-22. Transistor array.

the emitter of one (pin 4) to the base of the second (pin 6). This makes the top and bottom pairs identical. The complete schematic is shown in Fig. 7-23.

The source of audio signal is a 1-watt audio module. Its 8-ohm or 16-ohm output can be connected directly to the audio input of the sideband generator. Make certain that it is a balanced drive and that none of the secondary output leads are grounded. Potentiometer

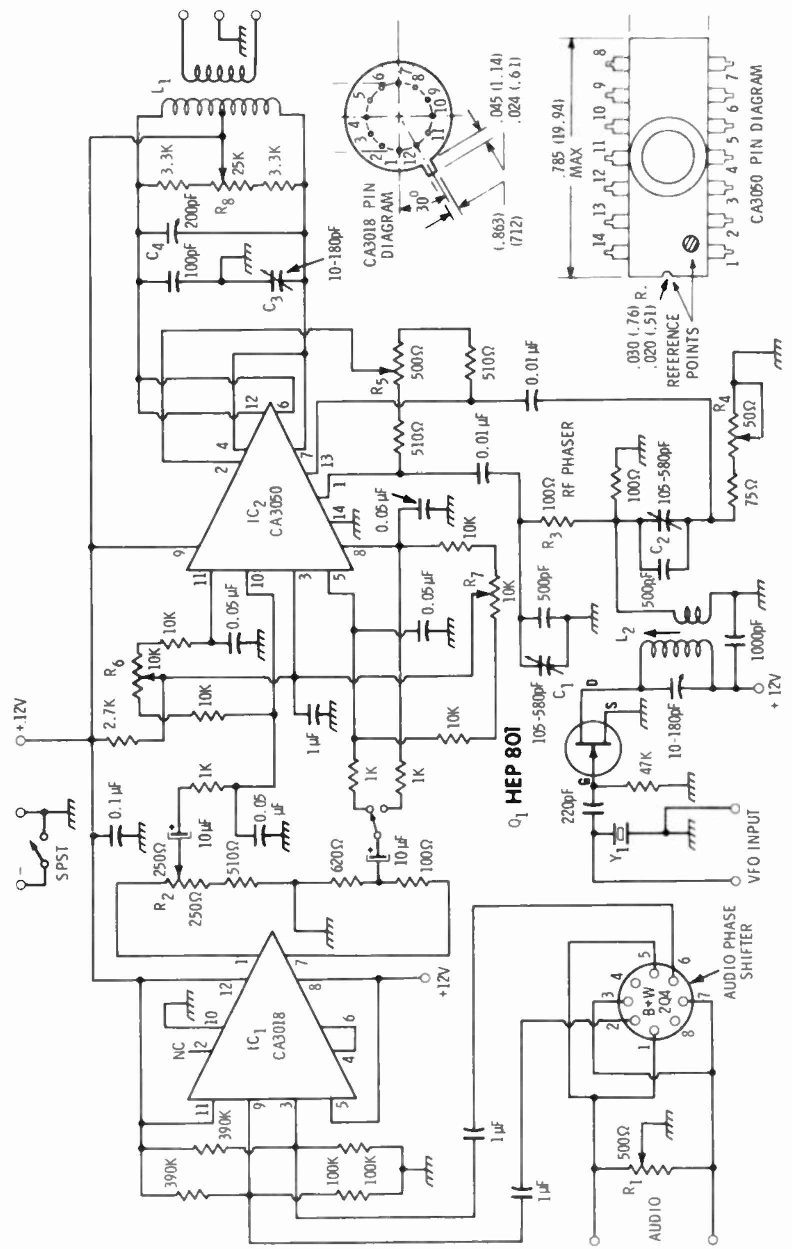


Fig. 7-23. 160-meter phasing-type sideband generator. Resistor and capacitor values are shown on the schematic.

R_1 is included at the input of the audio phase-shift network to make certain that equal-amplitude but opposite-polarity audio components are applied to the Barker and Williamson 2Q4 audio phase-shift network. Output of the phase-shift network is supplied to the dual Darlington amplifiers in the CA3018 integrated circuit.

The balanced but 90°-related outputs are applied through a suitable resistor-capacitor network to the sideband modulator. Potentiometer R_2 permits you to equalize the audio levels at the inputs. Included is an upper- and lower-sideband switch that permits you to apply audio of proper phase for appropriate sideband choice.

The 90°-related carrier components are applied to pins 1 and 13 (Figs. 21 and 23). Two resistor-capacitor networks at the input provide the required 90° phase shift. One network is a leading 45° combination; the other, a lagging 45° combination. The net phase shift is 90°. Component values have been selected for 160-meter operation. A 45° phase shift is obtained when the capacitive reactance equals the resistance of a series resistor-capacitor combination. This relationship can be used to calculate the required capacitor values for other bands.

The two pairs of paralleled-collectors of the Darlington pairs are applied to a balanced output resonant circuit. This manner of connection too is required for single-sideband generation using the phasing method. Carrier cancellation results from this connection. One pair of sidebands are additive; the other, subtractive because of the previous 90° relationships established for both audio and carrier injections. A variety of controls permit you to adjust for excellent carrier rejection and unwanted sideband attenuation.

A crystal oscillator using a field-effect transistor provides carrier signal. A selection of eight to ten crystals can provide versatile opera-

Parts List for Fig. 7-23

1	Chassis 7" x 11" x 2"
1	14-pin in-line IC socket
1	Octal tube socket (for phase shifter)
1	B&W 2Q4 audio phase-shift network
1	Crystal socket
1	Toroid coil form (Permacor 57-1541)
1	Spool No. 26 magnet wire for L_1
Y_1	160-meter crystal
L_2	73.5 to 98.4-μH slug-tuned coil (J. W. Miller 21A825RB1)
IC_1	RCA CA3018 integrated circuit
IC_2	RCA CA3050 integrated circuit
Q_1	HEP 801 FET

tion over a 25-kilohertz segment of the band. If desired, you can use the vfo of Project 9 as a carrier-signal source. When operating with a vfo the FET stage acts as amplifier; remove any crystal from the crystal socket. For operation on other bands it would also be necessary to use an appropriate drain resonant circuit for that band.

For 160-meter operation the crystal-oscillator resonant circuit consisted of 10-180 pF trimmer capacitor and a J. W. Miller (21A825RB1) slug-tuned coil. A secondary winding of 12 turns of #26 enameled copper wire was wound on the same coil form below the regular winding. This winding served as the secondary for the transformer.

The sideband output transformer was wound on a Permacor 57-1541 toroid core. The primary consists of 80 turns of #26 enameled copper wire, centertapped, while the secondary is 50 turns of #26 overlapping the centertapped area of the primary.

Operation

A good first step in putting the sideband generator in operation is to adjust the audio and rf phase-shift networks. Place a 100-ohm terminating resistor across the output of the sideband generator. Both networks operate at low frequency and signal can be observed readily on a service-type oscilloscope. The audio input potentiometer (R_1) is first set to midposition as are potentiometers R_3 and R_4. Adjust potentiometer R_1 for equal-amplitude signals at pins 3 and 9 of the audio IC.

Now observe the magnitude of the audio signal at pins 4 and either 5 or 8 of the double-balanced modulator. Potentiometer R_2 can now be adjusted for like audio levels at these two points.

The objective of the rf-phase shift adjustments are to attain equal-level and 90°-related components at pins 1 and 13 of the double-balanced modulator. The 90° phase shifts are handled by two separate 45° networks.

When resistance and capacitive reactance are equalized for such a series combination there is a 45° relationship established. Not only that, this relationship also equalizes the magnitude of the two rf voltage components developed across the reactance and resistance. Thus an oscilloscope can be used to determine the voltage levels across reactance and resistance and, you will know, that the phase shift is near to 45° when the two voltages are made equal.

Initially the 50-ohm potentiometer (R_1) is set to midposition. Capacitor C_1 is adjusted for equal-voltage levels across R_3 and C_1. Likewise the trimmer capacitor of the C_2-R_4 combination is adjusted for equal levels across C_2 and the series combination of R_4 and its associated 75-ohm fixed resistor.

The next step is to adjust the balance controls for minimum carrier output. The audio must be disconnected for this procedure. Open up capacitor C_3 several turns away from its maximum setting. Adjust the tune capacitor (C_4) for maximum output. However, try to locate a point at which the output dips slightly and then rises on each side. Recall that at exact resonance, there should be carrier cancellation and therefore a dip in the output. Capacitor C_3 can then be readjusted to bring down the carrier to a very minimum level. It is necessary to jockey back and forth between capacitors C_3 and C_4 to obtain the best results.

Now potentiometer R_8 can be set for minimum carrier. Do the same for potentiometer R_5. Then go to potentiometers R_6 and R_7 to obtain the very minimum of carrier output. It may be necessary to again retouch the other controls working for the very minimum of carrier output.

Now apply a 1000- to 1500-hertz tone to the input. Tune in the modulation on a receiver. The sideband switch can now be checked and you can note that a changeover is made between upper and lower sideband emission. The dominant sideband, of course, moves from one side of the carrier frequency to the other when you change the switch position. With the switch set for lower-sideband operation, tune the receiver to the upper sideband. Adjust potentiometer R_4 for minimum level of unwanted sideband. A very slight adjustment of potentiometer R_2 may also aid in cutting back on the undesired sideband. Of course, an oscilloscope display would be of definite help in setting the various controls (refer to Fig. 5-4).

This sideband generator can be used on 160 meters in association with the linear amplifiers of Projects 24 and 26. When used with the low-powered amplifier of Project 24, the combination serves as a QRPP sideband transmitter of about 1-watt PEP output. Driving the linear amplifier of Project 26 raises you to the QRP class of sideband output power.

CHAPTER 8

VHF/UHF Transmitter Circuits Principles

Very high frequency (vhf) and ultrahigh frequency (uhf) circuits and techniques differ somewhat, but not drastically, from those employed on the lower frequencies. Lumped-constant LC circuits are common in low-powered stages. Of course, resonance is attained with a small coil of few turns and a low-value capacitor with a very low minimum value. Hairpin loops and other special physical configurations serve as the inductance part of many resonant circuits on 2 meters and below.

In higher-powered applications, tuned sections of transmission line serve as complete resonant circuits (Fig. 8-1). In a practical circuit the linear length of the transmission line section is substantially shorter than a physical quarter wavelength. In tune with the net capacitance of the circuit it forms a parallel-resonant tank. In some instances the line becomes an electrical three-quarter wavelength instead of one-quarter wavelength.

Such a line can be resonated in one of two ways, or a combination of both as shown in Figs. 8-1A and 8-1B. The inductance of the resonant circuit is varied by moving a shorting bar along the length of the line at the far end. It can also be resonated by using a small trimmer capacitor across the open end of the line.

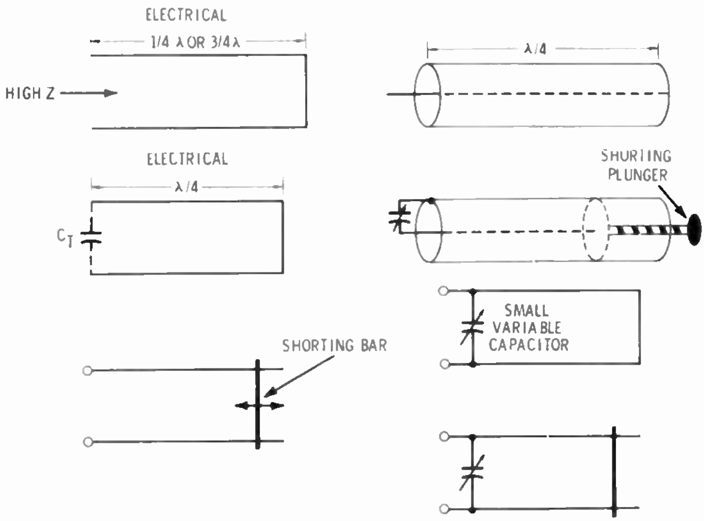


Fig. 8-1. Sections of transmission line as parallel-resonant circuits.

Coaxial tank circuits (Fig. 8-1C) are also popular. Similar methods of tuning are employed. The coaxial type of resonant circuit is well shielded because of its grounded outer conductor which completely surrounds the tuned circuit.

The common electron devices with appropriate physical and electrical characteristics serve well. These can be tube, bipolar transistor, field-effect transistor, and various types of integrated circuits. In addition there are two special diode types that have been popular in vhf-uhf transmitters. These are voltage-variable capacitor diodes and the varactor diodes.

The voltage-variable capacitor diode displays a capacitance that is a function of the voltage between its anode and cathode. One of the effects of changing voltage across a semiconductor diode is a change in capacitance. By suitable semiconductor design this change in capacitance can be emphasized and made reasonably linear with respect to voltage change. This special diode can be used in frequency-modulated circuits, automatic frequency control systems, and other configurations where a capacitor or resonant-circuit frequency is to be varied with a dc or ac voltage.

The voltage-versus-capacitance relation of a typical solid-state voltage-variable capacitor is shown in Fig. 8-2. If such a capacitor

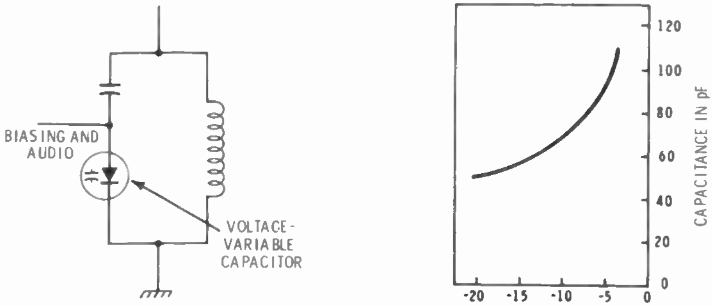


Fig. 8-2. Voltage-variable capacitor diode.

is made a part of a resonant circuit a change in the voltage across the diode can change the frequency of the resonant circuit. Such capacitors are common in f-m transmitters and are discussed in greater detail in Chapter 9.

The varactor is a higher-powered special version of the voltage-variable capacitor diode that can be used as an efficient frequency multiplier. Again a reverse bias forms a depletion layer across the junction and its capacitance is determined by the value of the reverse voltage. In general the capacitance is inversely proportional to the square root of the applied voltage.

To operate as a frequency multiplier, the biasing voltage of the usual varactor circuit is such that the varactor voltage swings with signal between full conduction and reverse-breakdown potential. It is not driven completely to the breakdown potential but is permitted to swing for a very short time into the forward-conduction region. This latter swing aids in improving the harmonic content of the resultant varactor current.

Since the varactor capacitance changes with the input signal there is a nonlinear relation between the varactor voltage and the input signal. Thus, the varactor current has other than a fundamental-frequency component. In addition when the junction is forward-biased for a short interval, at the very crest of the input wave, there is a movement of carriers across the junction. However, before these injected minority carriers have time to be combined with majority carriers, the applied voltage swings quickly into the nonconduction region. As a result the carriers quickly move into their former positions. This condition also aids in the production of a pulse-like varactor current that is rich in harmonics.

Typical varactor frequency multipliers are shown in Fig. 8-3. Inductor L_1 along with capacitor C_1 and the average capacitance of the diode form a series-resonant circuit at the fundamental frequency. Thus a resonance current exists in the input side. However, the volt-

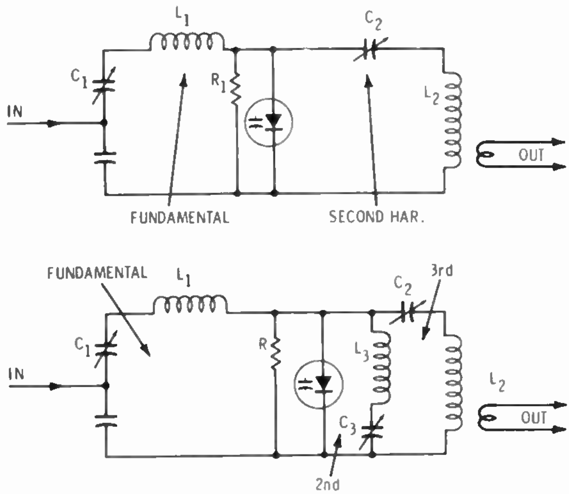


Fig. 8-3. Varactor doubler and tripler circuits.

age variation across the varactor is rich in harmonics. The second harmonic is made dominant in Fig. 8-3A with the second-harmonic resonant circuit (L_2C_2) although even higher order harmonics are present. The L_2C_2 combination accents the flow of second-harmonic components in the output circuit.

There is a small amount of forward current in the varactor circuit because some recombining takes place during the short interval that the signal biases the junction in the forward direction. Like grid current, this current, in association with an external resistor, can be used to set the average bias of the diode. Therefore it is not necessary to use an external source of dc bias. The input signal itself swings the diode to a forward-bias value at its crest. Bias is held constant by the circuit capacitance and the biasing resistor (R_1).

A basic tripler circuit is shown in Fig. 8-3B. Input resonance is again at the fundamental frequency. However, the L_2C_2 output circuit is tuned to the third harmonic, transferring this signal component to the load.

A third resonant circuit (L_3C_3) is tuned to the second harmonic. Therefore a fundamental component and a small second-harmonic component are present in the varactor. Inasmuch as the varactor is, in itself, a nonlinear device, these two current components produce a sum component which is, in effect, a third-harmonic component. This second-harmonic circuit is called an idler tank circuit that takes advantage of the mixing activity of the varactor to give a further boost to the third-harmonic output of the varactor circuit.

BASIC TRANSMITTER PLANS

The beginning signal of most transmitters that operate in the 6, 2, 1½, and ¾ meter bands starts at a much lower frequency. In fact, the basic signal for many originates somewhere in the 5- to 13-MHz frequency range. A series of multipliers step up the frequency to the desired transmit range.

The most common method is that shown in Fig. 8-4. A multiplier count of 6 along with an oscillator frequency of 8.5 MHz would

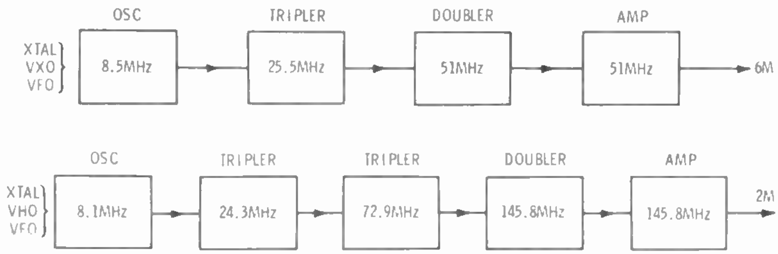


Fig. 8-4. Multiplier chains for 2 and 6 meters.

produce a transmit frequency in the 6-meter band at 51 MHz. A tripler and a doubler produce the required multiplication of 6.

A multiplication of 18 would be necessary to obtain 2-meter output with the oscillator operating in the same frequency range. An output of 145.8 MHz is obtained with an oscillator frequency of 8.1 MHz and a total count of 18. The 18 multiplier can be obtained with a doubler and two tripler stages.

Three optional oscillator plans are given. The first is simply crystal control. The second plan is known as a vxo—variable crystal control. In this arrangement the fundamental crystal oscillator circuit includes a means for a slight change in the frequency of operation of the crystal oscillator. Though this change might be quite limited on the fundamental frequency it does produce a substantial change at the final transmit frequency.

Frequency change is multiplied by the same ratio as the absolute frequency. For example, when the oscillator is varied 10 kHz, the resultant change in the transmit frequency for the 2-meter case is 180 kHz (10 kHz × 18).

A third plan uses a vfo. Vfo operation requires high frequency stability because frequency drift is multiplied by the same factor. For example, if the vfo has a 500-Hz drift at 8.1 MHz, the drift in the 2-meter band would be 9000 Hz.

Another method of changing the final transmit frequency from a lower originating frequency is shown in Fig. 8-5. This system uses

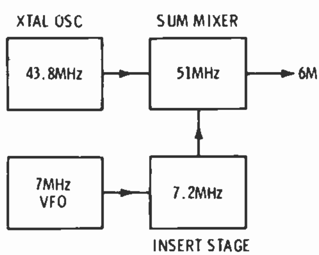


Fig. 8-5. Up-conversion technique.

the up-conversion technique covered previously in Chapter 6. In the example, a 7.2-MHz signal is mixed with a crystal-controlled frequency of 43.8 MHz. The output of the mixer is tuned to 51 MHz in the 6-meter band. A simple overtone crystal oscillator circuit can provide the oscillator injection frequency. Other combinations can be used to provide up-conversion to any one of the vhf-uhf bands.

A third method places the oscillator on the transmit frequency (Fig. 8-6). No frequency multiplier is required. Succeeding stages simply build up the signal to the final transmit power. Frequency drift is a problem when operating at this high frequency. A phase-lock technique can provide high stability. In a typical example a series of dividers can divide the oscillator frequency down to a very low value at which it can be compared with a stable frequency source. Assume the transmit oscillator frequency is 51.3 MHz and there is a pair of three-to-one dividers producing a net division of

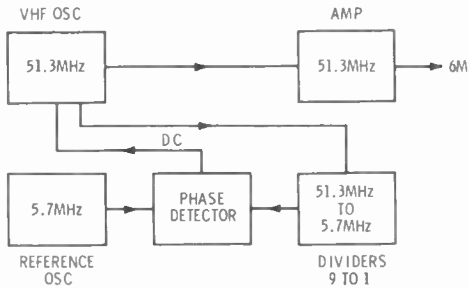


Fig. 8-6. Phase-locked loop method.

9 (3 × 3). The divider output frequency is 5.7 MHz. This can be compared with a 5.7-MHz crystal oscillator in a phase comparator circuit. The error voltage can then be used to keep the transmit oscillator on frequency, using a voltage-variable capacitor.

If the standard crystal oscillator is made a vxo, its frequency can be varied, and there will be a follow up change in the frequency of the transmit oscillator. All of this can be done with rather simple integrated-circuit packages.

MODULATION METHODS

All forms of modulation are popular on the vhf-uhf frequencies. Cw is perhaps the least common; conventional a-m is perhaps the most popular. To a great extent it is giving way to frequency modulation, especially on 2 meters. Sideband modulation is rising in popularity because next to cw it is best for DXing.

The type of modulation has an influence on the plan used to attain the final transmit frequency. The multiplication technique of Fig. 8-4 can be used for cw and a-m. For a-m it assumes that the modulation takes place after the final transmit frequency has been reached. The method of Fig. 8-4 can also be used for frequency modulation. Usually the fundamental-frequency oscillator is frequency modulated. An advantage of f-m is that frequency multiplier stages which operate class C can be used after the modulation process has taken place. This is not so for any type of amplitude modulation (conventional a-m, dsb, or ssb).

In fact, the plan of Fig. 8-4 is not usable for single-sideband transmission because the class-C multiplier and amplifiers would distort the modulation.

If a single-sideband is to be built up in frequency the heterodyning process of Fig. 8-5 is employed. The same plan can also be used for the two other forms of modulation, conventional a-m and dsb. Modulation can then take place at low power levels.

The phase-lock plan of Fig. 8-6 has interesting possibilities for the various modulation modes. Amplitude modulation can be handled in a succeeding amplifier. Frequency modulation can take place at the transmit oscillator frequency when using a suitable high-ratio divider chain. A simple switching arrangement might permit the choice of either f-m or a-m.

A balanced modulator following the oscillator can be used to generate a double-sideband signal. Succeeding amplifier stages would then have to operate as linears.

The oscillator followed by an up-conversion mixer would provide for single-sideband transmission. If a 9-MHz sideband generator is used, the transmit oscillator frequency would have to be dropped in frequency by 9 MHz to produce a signal on the same transmit frequency. Again it would be necessary to use linears after the mixer.

The plan of Fig. 8-7 could be the core for an all-modulation-mode vhf-uhf transmitter. The balanced circuit following the oscillator could operate as a balanced mixer for ssb and other a-m modes. For cw, conventional a-m, and f-m you would switch around this

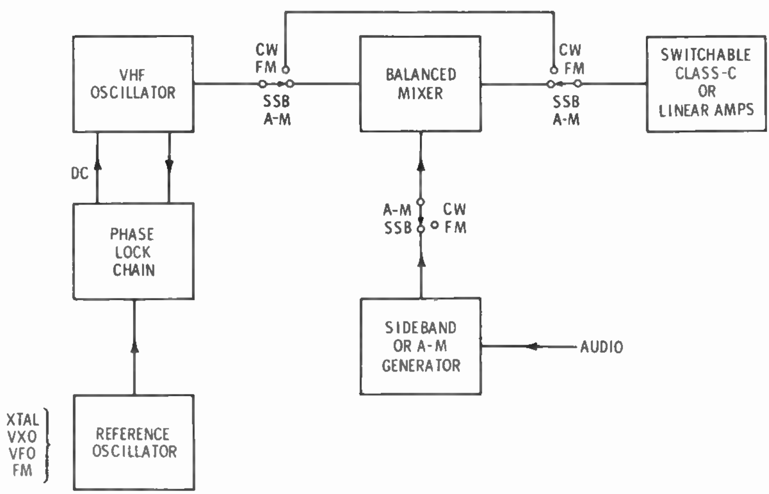


Fig. 8-7. Multimode modulation method.

stage and switch off the linear bias of the amplifier. For a very simple dsb modulator, a simple switching plan could permit the balanced mixer to also operate as a balanced modulator. Audio would be applied directly to the balanced circuit.

TYPICAL CIRCUITS

When the oscillator of a vhf-uhf multiplier chain operates below 20 MHz, regular fundamental crystal oscillators are employed. These are the same as those covered in earlier chapters. At higher frequencies, up to approximately 60 MHz, overtone crystals are more commonly employed. A typical circuit is shown in Fig. 8-8. In this arrangement a third-overtone crystal is operated in its series-resonant mode. A vacuum-tube counterpart is shown in Fig. 8-9. The oscillator of Fig. 8-10 can be operated above 60 MHz and up into the 2-meter band. Fifth and seventh mechanical overtones of the crystal develop the usable output.

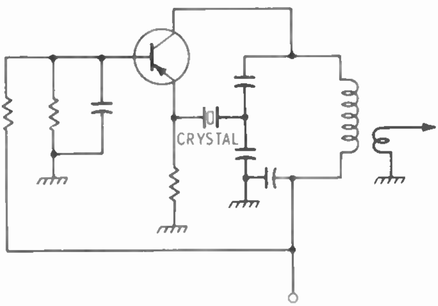


Fig. 8-8. Bipolar overtone oscillator.

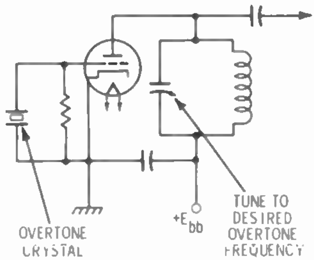


Fig. 8-9. Vacuum-tube overtone oscillator.

The various crystal oscillators can be arranged for vxo operation. They can also be frequency-modulated; typical circuits of which are given in the following chapter. A bipolar vxo (of the type shown

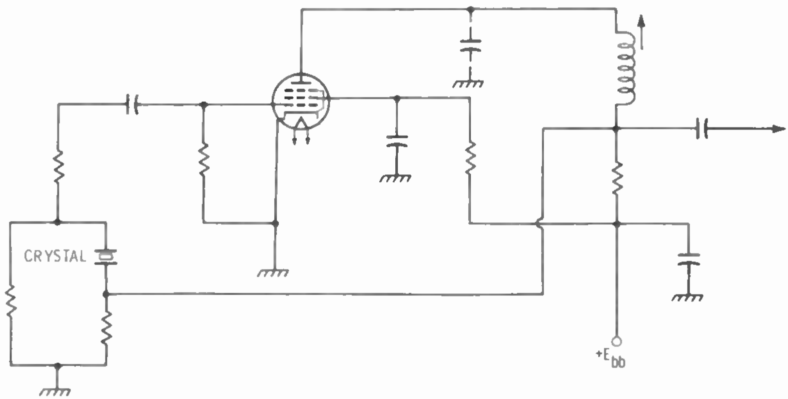


Fig. 8-10. Fifth- and seventh-overtone oscillator.

in Fig. 8-11) will pull the crystal oscillator frequency approximately 5 kHz with stability. The high value variable capacitor lowers the operating frequency below that of the crystal as its capacitance is increased. The limit of the frequency change is set by the associated inductor. It must be such that the frequency change is not too great or a sacrifice in oscillator stability must be made. Nevertheless, with

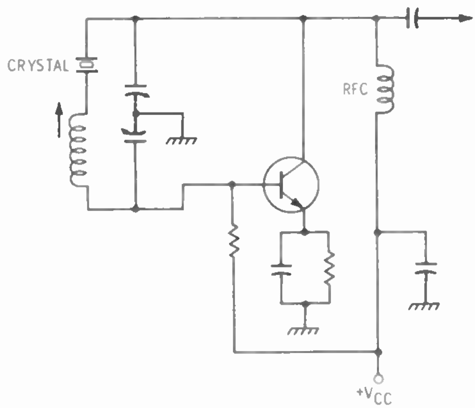


Fig. 8-11. Bipolar vxo.

a crystal in the 8-MHz range, the possible frequency change on 6 meters is 30 kHz and nearly 100 kHz on 2 meters.

FET and vacuum-tube versions are shown in Fig. 8-12. Again the split-stator variable is used for tuning and the adjustable inductor for controlling the maximum amount of stable frequency change.

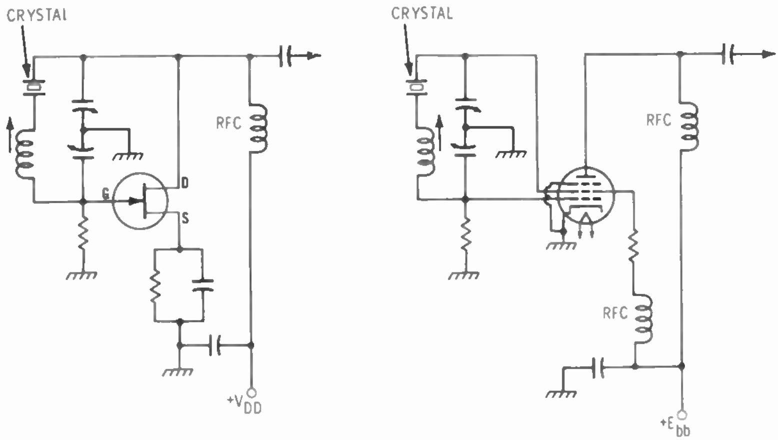


Fig. 8-12. Basic FET and vacuum-tube vxo circuits.

All of the various electron devices—tubes, bipolars and FETs—can be used as frequency multipliers. Doublers and triplers are used extensively in the frequency chains that permit operation in the vhf-uhf bands. Frequency doublers were covered earlier in the book. One additional type that is often used in multiplier chains because of its higher efficiency is the push-push doubler of Fig. 8-13. In the push-push doubler the input signal is applied to the grids in push-pull. However, the plates are connected in parallel. During the positive alternation the top tube conducts and draws a burst of current into the plate tank circuit. The negative alternation causes the lower tube to conduct. As a result another burst of plate current hits the tank circuit. Thus each alternation of the input signal (two per cycle) results in plate current. Inasmuch as the plate tank circuit is oscillating at twice the frequency of the incoming signal there is plate current in the tank circuit once during each harmonic cycle. Thus in the push-push arrangement plate current coincides with the negative peak of each output cycle, thus increasing the output and efficiency.

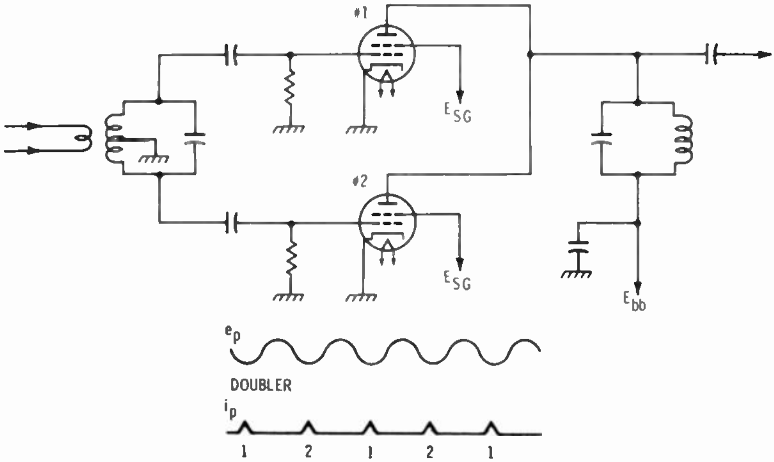


Fig. 8-13. Push-push frequency doubler.

Field-effect transistors are fine multipliers. They can be operated as doublers and triplers, producing a good output with very little input drive.

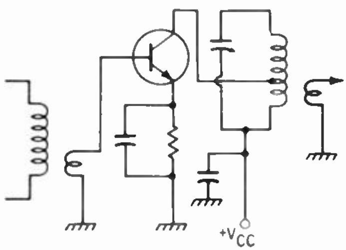
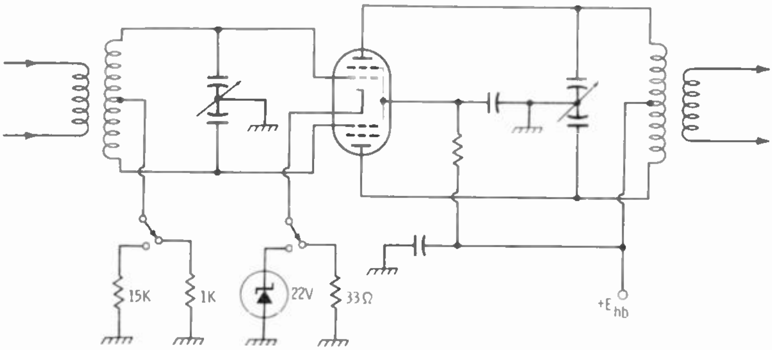
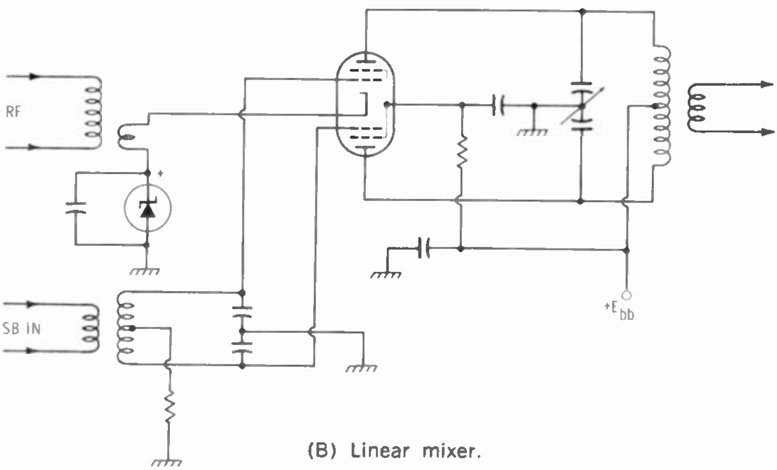


Fig. 8-14. Bipolar multiplier stage.

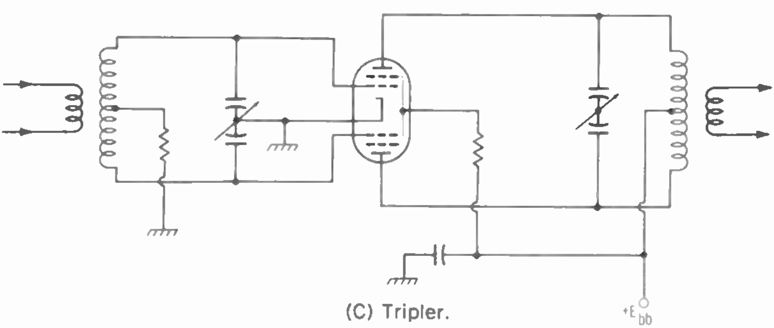
A bipolar frequency multiplier is shown in Fig. 8-14. Bipolar circuits are inherently good harmonic generators. However, they must be tuned carefully to minimize fundamental and undesired harmonic feedthrough. Input and output resonant circuits are preferred. In the example, there is a parallel-resonant input tank with its low-impedance winding that matches the low-impedance input of the bipolar transistor. The output resonant tank is tuned to the desired harmonic (second or third) and is tapped. Thus a match is made between the collector of the bipolar transistor and the resonant tank circuit. At



(A) Typical amplifier.



(B) Linear mixer.



(C) Tripler.

Fig. 8-15. Vhf and uhf balanced rf circuits.

the same time the tank circuit can be made to have a good Q. A low-impedance secondary permits feed to a succeeding stage.

Tube Circuits

One of the most popular vhf-uhf tubes is the 6360. It operates well as a frequency multiplier, fundamental amplifier, and linear mixer or amplifier. These three possibilities are shown in Fig. 8-15. It is a dual tetrode in a single envelope and, consequently, lends itself to the push-pull balanced circuits so ideal for the design of balanced and trouble-free vhf-uhf circuits.

A typical 2-meter amplifier is shown in A. A properly balanced circuit using this tube requires no neutralization. Note the balanced input and output coils and the split-stator variable capacitors.

A simple switching arrangement permits the amplifier to be used class C or class-AB linear. When using a 1K grid resistor and a 22-volt zener diode in the cathode circuit the stage is biased for linear operation. For class-C operation use a 15K grid resistor and replace the cathode zener with a 33-ohm resistor.

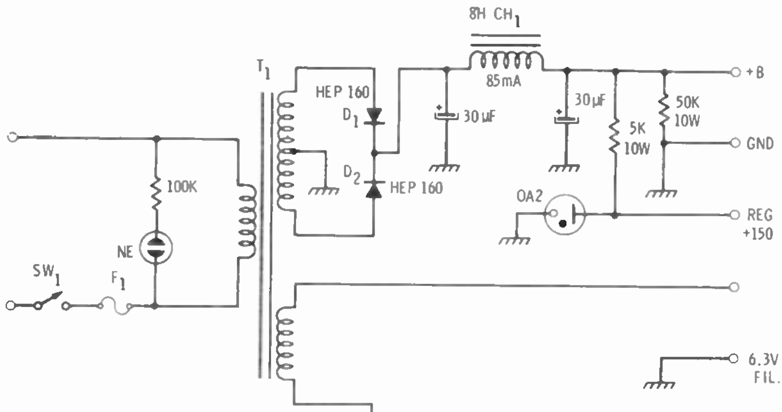
The linear biasing technique is unique in that the cathode current and plate resistances of the two sections of the tube are such that a constant 22-volt drop is maintained across the zener. Of course, regular linear biasing could be used. In this case a 33-ohm resistor would be used in the cathode circuit and the 1K grid resistor instead of being grounded would be connected to a −22 volt source.

Circuit B shows the same tube being used as a linear mixer. Linear biasing is obtained again from a zener diode. The stage operates as a grounded-grid linear. Heterodyning oscillator injection is made in the cathode circuit. This component cancels in the output because of the push-pull connection of the plates. The sideband signal is applied to the grids in push-pull.

A tripler stage is shown in C. The balanced input circuit is resonant at the input frequency; the balanced plate circuit is tuned to its third harmonic.

PROJECT 31: SIX-METER VACUUM-TUBE TRANSMITTER

Most vhf-uhf transmitters operate with low-frequency oscillators (crystal, vxo, or vfo) and a series of stages that multiply the signal up to the transmit frequency. In this project you can construct a three-tube 6-meter transmitter that employs 8- to 9-MHz fundamental crystals. A separate power supply furnishes filament and high voltage as well as regulated 150-volts for the oscillator. The crystal oscillator is a vxo type which permits an approximate 100 kHz of frequency movement on the 6-meter band. The vxo is followed by a tripler and a doubler. The final stage is a 6360 six-meter push-pull amplifier.



T_1	Power transformer, 270-0-270 volts @ 125 mA., 6.3V. 3.5A. filament (Standard PC 8405)	D_1, D_2	HEP-160 silicon rectifiers
			Chassis 1″ x 2″ x 7″
			Fuse and holder

Fig. 8-16. Power supply for vhf transmitter. Resistor and capacitor values are shown on the schematic.

The power supply schematic is given in Fig. 8-16; it can be constructed on a 11″ × 2″ × 7″ chassis. This unit can serve as a utility power supply for numerous other vacuum-tube projects you may have in mind. The regulator section is particularly helpful in maintaining the frequency stability of various oscillator types.

The radio-frequency section of the transmitter is shown in Fig. 8-17. The exciter employs two 6CX8 triode-pentode combination tubes followed by the 6360 amplifier. The crystal oscillator includes a two-gang broadcast variable and a suitable inductor (L_1) which permits the crystal oscillator frequency to be varied over a satisfactory frequency range with good stability. This stage uses a choke output followed by a pentode tripler stage. Next there is a triode doubler to obtain a total multiplication of 6 (3 × 2). An 8.4-MHz crystal in the circuit would result in an output frequency of 50.4 MHz (8.4 × 6).

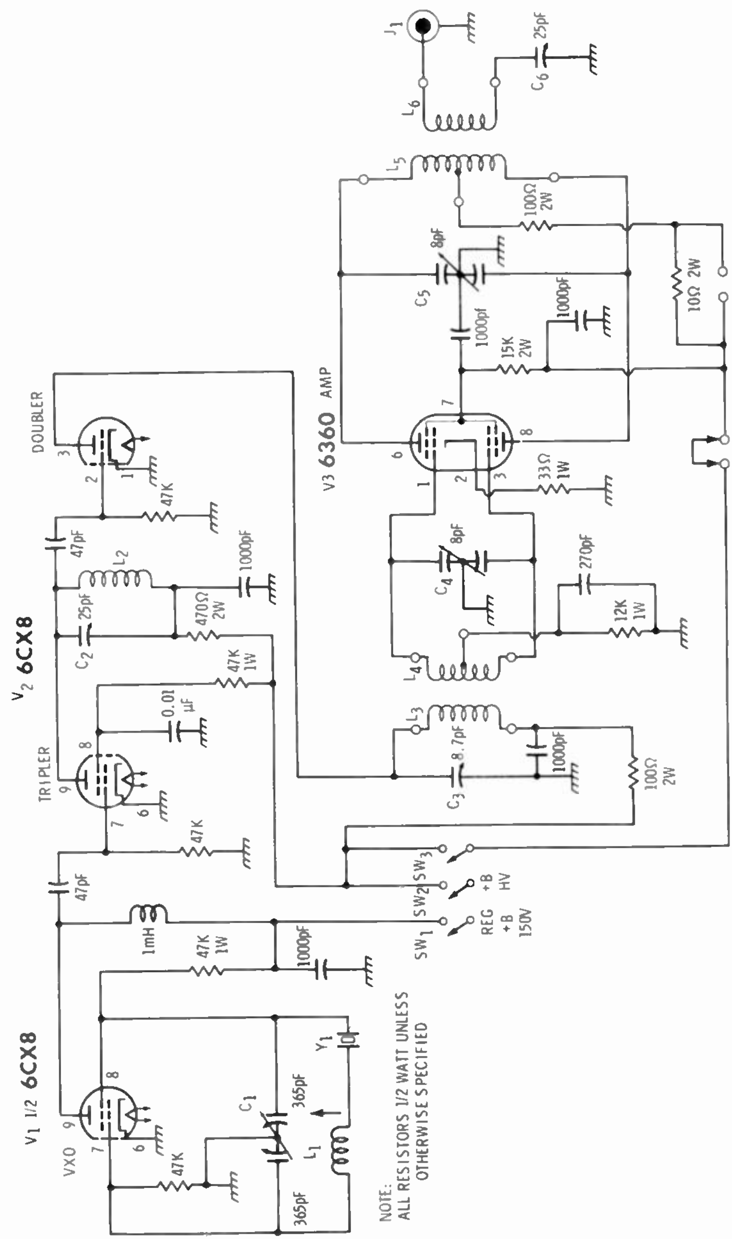


Fig. 8-17. Six-meter rf section. Resistor and capacitor values are shown on the schematic. Rig is built on a 7" x 3" x 15" chassis.

Parts List for Fig. 8-17.

C_1	Two-gang 365-pF variable capacitor	L_4	7 turns B&W 3011 miniductor, centertapped (spacing between L_3 and L_4 is 1" center to center)
C_2,C_6	25-pF variable capacitors	L_5	7 turns B&W 3011 miniductor, centertapped
C_3	8.7-pF variable capacitor, (Johnson 160-104)	L_6	2 turns hookup wire wound around center of L_5
C_4,C_5	8-pF per section butterfly capacitor (Johnson 160-208)		15 binding posts
J_1	Coaxial chassis receptacle		1 power terminal strip
L_1	16- to 24-µH slug-tuned inductor (J. W. Miller 21A225RBI)		3 9-pin tube sockets
L_2	10 turns B&W 3014 miniductor		2 crystal sockets
L_3	6 turns B&W 3011 miniductor	$SW_{1,2,3}$	Spst toggle switches

Construct the first two stages using standard vacuum-tube practices. Mount the oscillator components rigidly and note that it is the only tube supplied with regulated 150 volts. A separate power supply switch permits you to power the oscillator without activating the rest of the transmitter. This is useful in permitting you to zero-beat on a specific frequency without radiating a signal.

From the plate circuit of the tripler to the output use short leads and proper isolation between input and output circuits. It is important to minimize stray capacitances to ground. Note in the photograph of Fig. 8-18 that small perforated boards are useful in mini-

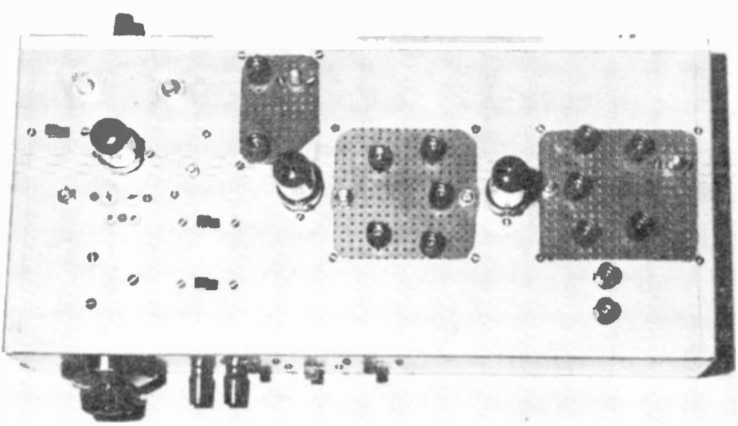


Fig. 8-18. Top view of 6-meter transmitter.

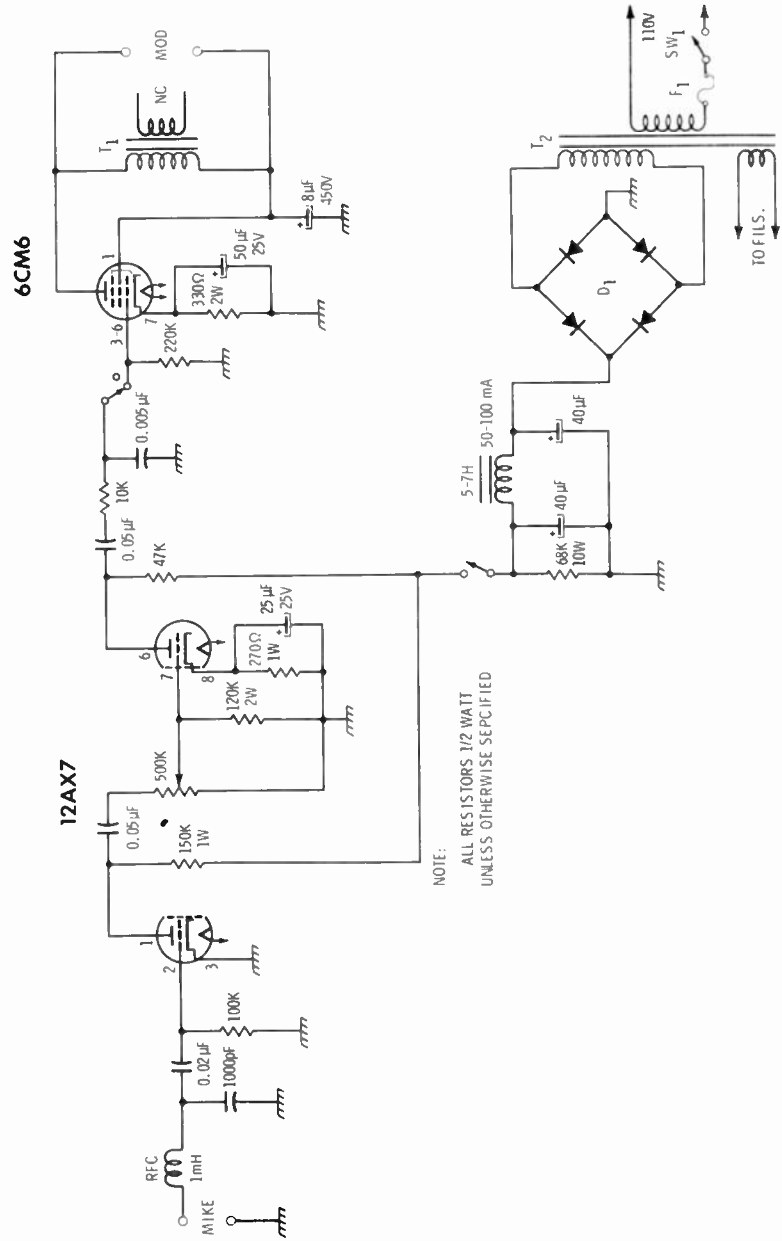


Fig. 8-19. Speech amplifier and modulator. Resistor and capacitor values are shown on the schematic. Modulator is built on a 11" x 7" x 2" chassis

mizing these stray capacitances. The coils themselves are supported by binding posts which permit complete versatility and the insertion of appropriate coils for 6, 2, or even lower wavelengths. Coil data are given for 6 meters in the parts list.

The 6360 output stage is a balanced push-pull arrangement that does not require neutralization. It employs two special split-stator variables called butterfly capacitors. Their construction permits good balance and a low minimum capacitance and consequent efficient resonant-circuit operation on these high frequencies. Wind your coil carefully to obtain the very best balance.

An amplitude modulator for the transmitter is built on a separate 7″ × 2″ × 11″ chassis and includes its own power supply. The schematic diagram for the modulator is given in Fig. 8-19. It is a two-

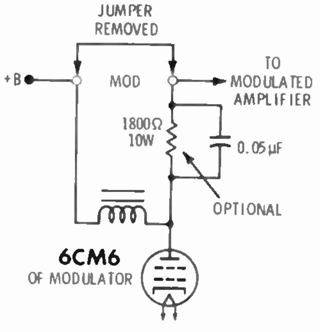


Fig. 8-20. Application of modulation to 6-meter final amplifier.

tube affair consisting of a dual-triode voltage amplifier and a beam-power tube modulator. Output power is 3 to 5 watts which is capable of obtaining near 100% modulation using a simple Heising choke modulator. If 100% modulation is desired an optional parallel resistor-capacitor combination can be inserted in the rf amplifier supply voltage line to the modulator (Fig. 8-20).

The usual high-impedance communications microphone will supply enough voltage at the input of the first input stage. Rf feedback into the microphone line is avoided by the input radio-frequency

Parts Lists for Fig. 8-19.

D_1	Diode bridge rectifier, 2A,800 PIV	T_2	Power transformer, 240V. sec. 6.3 V fil.
T_1	Modulation choke or use primary of audio output transformer	SW_1	Power switch, spst toggle switch
			2 9-pin tube sockets

1 Fuse and holder

choke. A resistor–capacitor low-pass filter between the voltage amplifier and the modulator tube holds the overall response to the voice-frequency range.

When used with the r-f section of the transmitter the plate and screen voltage of the modulator tube is supplied by the power supply associated with the r-f section. The audio unit can also be used as a utility modulator, and, in some applications, an output transformer can be inserted rather than a modulator choke. The plate and screen voltage of the modulator can then be supplied by the power supply built into the audio unit.

Note that a switching arrangement is inserted between the second voltage amplifier and the input of the modulator. This permits the two input stages to be used for frequency modulating a transmitter. In fact, in one of the projects of the next chapter an f-m oscillator and amplifier will be added to this same chassis. This will give you a utility a-m and f-m modulator for a variety of possible applications in the teaching laboratory or ham shack.

Operation

In adjusting the transmitter, first supply power to the oscillator and check out its operation on a communications receiver. You may wish to pick up the second or third harmonic of the crystal oscillator. If a receiver is available to tune to the third harmonic you will have a useful monitor for checking out the operation of the following tripler stages. An important first step to take, however, is to use a dip meter to set each of the resonant circuits on the desired frequency. The tripler resonant circuit should be set on the *third* harmonic of the crystal frequency, while the output of the doubler and the two amplifier resonant circuits should be set on the *sixth* harmonic of the crystal frequency. This preliminary adjustment is not only useful in preventing you from tuning one of the stages to an improper harmonic, but also keeps you away from settings that can cause self-oscillation within the transmitter. Self-oscillation of the triode can be a problem if input and output resonant circuits are inadvertently set to the same frequency. The stage will oscillate by itself.

Insert a 0-100 mA dc meter across the 10-ohm resistor in the plate supply line of the final amplifier. Connect a 50-ohm dummy load (preferably with a wattmeter) across the r-f output. If you now supply power to the transmitter a substantial, or at least an observable output, should be recorded, depending on how well the resonant circuits have been set on proper frequency. You can now retouch the controls beginning at the tripler. However, tune each control very finely and don't take it too far off the resonant frequency. Adjust for an output peak reading. Do the same for the triode output circuits and then the push-pull input and push-pull output tuning adjustments. A dc input voltage of about 300 volts should result in about 5 to 8 watts of rf output and a dc plate current reading of 40 to 60 milliamperes. The transmitter will tune up well over the entire 6-meter band using appropriate crystals for the vxo.

Connect the a-m modulator to the transmitter as shown in Fig. 8-20. Do so without the optional parallel resistor-capacitor combination. First tune up the transmitter using the jumper across the modulator binding posts. Now remove the jumper and connect the modulator into the circuit before applying power to the modulator. There will be some decrease in output after power has been supplied to the audio section.

Connect a 1000-Hz audio tone to the microphone input. Set the speech control of the audio amplifier to about midposition. If available connect an oscilloscope to the output. Increase the audio level of the 1000-Hz tone and observe the rf envelope display. Note that there is flattening of the envelope before 100% modulation is obtained. However, good modulation of 85 to 90% is obtainable. The normal modulation characteristics are such that there is a slight downward shift of output power with modulation. For best quality do not permit this to become excessive. Avoid any serious flattening as indicated by the waveform display.

A higher level of modulation can be obtained by inserting the parallel resistor-capacitor combination. There is a sacrifice in output power. However, the plate-modulation system responds in more normal fashion when this technique is employed. There will be an upward kick of the power output reading with modulation.

Remove the audio tone and connect your microphone across the audio input. Sound an "oh" into the microphone and observe the waveform pattern. Speak normally and observe the display making a suitable adjustment of the audio gain control until deflection corresponds to about that obtained previously with the 1000-Hz tone level adjusted for about 85- to 90-percent modulation. Check the

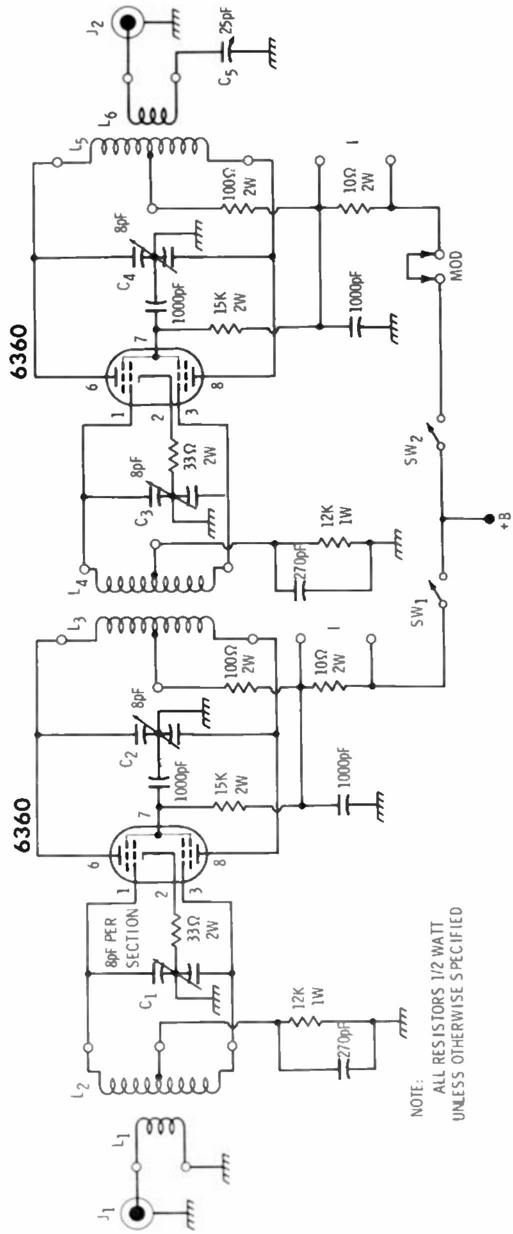


Fig. 8-21. Tripler and 2-meter amplifier. Capacitor and resistor values are shown on the schematic.

voice quality on a receiver located nearby or obtain a quality report from another ham several miles away.

PROJECT 32: TWO-METER TRANSMITTER

The six-meter transmitter of Project 31 plus two additional 6360 stages permits two-meter operation. The 6360 input stage operates as a tripler and obtains its signal at the output of the transmitter of Project 31. Circuit arrangement is shown in Fig. 8-21. Individual stages are identical to the output stage of the previous transmitter except for the coils. Three of the resonant circuits are tuned to 2 meters and consist of two turns of B&W 3013 miniductor coil. Each is center-tapped. The input resonant circuit must be tuned below six meters and is made identical to the output coil of the six-meter transmitter.

This two-meter tripler and amplifier can be operated from the same power supply as built for Project 31. Input and output coupling is handled by single turns of hookup wire wound around the center of the input and output coils. Interstage resonant coils are positioned near to each other without touching.

Operation

For two-meter operation the six-meter transmitter of Project 31 is operated at a somewhat lower frequency. To obtain 2-meter output it is necessary that a crystal in the 8-MHz range be employed. For example a 8.1-MHz crystal produces an output on 145.8 MHz ($3 \times 2 \times 3$).

The first step in tuning the transmitter for two meter operation is to use a dip meter to set each of the resonant circuits on frequency. The tripler output resonant circuit is set on 24.3 MHz in our example, while the output of the doubler must be tuned to 48.6 MHz. Tune up the transmitter in accordance with the instructions of Project

Parts List for Fig. 8-21.

$C_{1,2,3,4}$	8-pF butterfly variable capacitors (Johnson 160-208)	L_2	7 turns B&W 3011 miniductor, centertapped
C_5	Coaxial connectors	$L_{3,4,5}$	2 turns B&W 3013, centertapped
J_1,J_2	25-pF variable capacitor	L_6	1 turn hookup wire around center of L_5
L_1	1 turn hookup wire around center of L_2	SW_1,SW_2	Spst slide switches

31. If any of the resonant circuits do not tune low enough in frequency you may need to add a turn or two.

Supply power and tune up the transmitter on this new frequency.

Now use the dip-meter on the two-stage multiplier and amplifier. The input resonant circuit must, of course, be tuned to the same frequency as the output of the exciter section. The output of the tripler and the grid and plate resonant circuits of the amplifier are all tuned to 145.8 MHz. You can now supply power to both the exciter and the multiplier-amplifier. The tripler stage itself has a good two-meter output because the push-pull connection operates well as an odd harmonic generator. Tune the final amplifier. Output should be only a bit less than that obtained on six meters. You can amplitude-modulate the transmitter by following the procedure covered in Project 31.

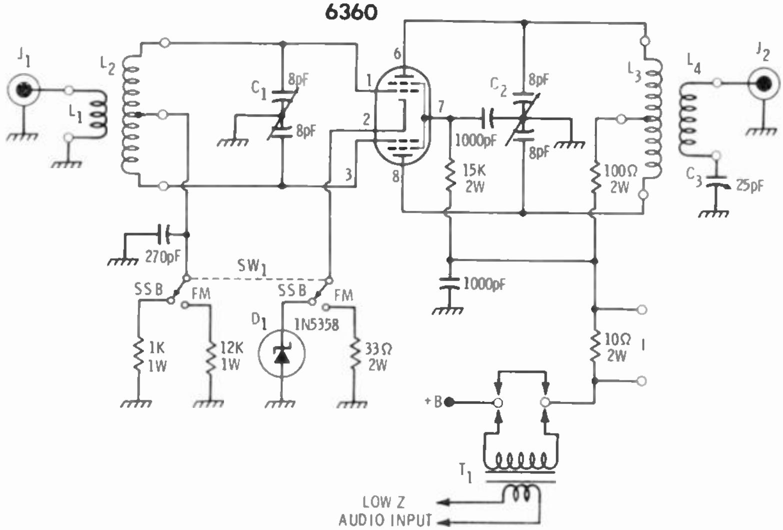
PROJECT 33: UTILITY VHF AMPLIFIER

The basic 6360 design covered in Projects 31 and 32 can be elaborated upon to provide you with a versatile amplifier for two and six meters. Appropriate coils permit operation on 10 and 1¼ meters as well. The circuit arrangement of Fig. 8-22 permits operation as a class-C amplifier, linear rf amplifier, or frequency tripler.

A simple rearrangement of the input circuit as in Fig. 8-23 permits operation as a push-push frequency doubler. If binding posts are included to mount the coils, this can be done very quickly.

A simple switching arrangement permits straight class-C or linear operation. Class-C is the mode to use for fm or CW transmission. If amplitude modulation is desired, the class-C amplifier can be plate-modulated. For convenience a modulation transformer is included. This can be a 5- to 10-watt output transformer with an approximate 3000- to 6000-ohm primary and a low-impedance 3- to 8-ohm secondary. However, the transformer is reverse connected with high-impedance side connected in the supply voltage line to the class-C amplifier. The audio input impedance is, of course, very low and matches the low-impedance output of any type of audio or PA amplifier normally used to drive a speaker.

The utility amplifier can also be used to build up the magnitude of a single-sideband, double-sideband, or a-m input signal. When switched to the linear mode, the zener diode in the cathode circuit sets linear bias for the stage. An input signal of a couple of watts or



C_1,C_2	8-pF butterfly variable capacitors (Johnson 160-208)
C_3	25-pF variable capacitor
D_1	1N5358 Zener diode, 22V
J_1,J_2	Coaxial connectors
L_1	2 meters, 1 turn hookup wire around L_2 6 meters, 2 turns hookup wire around L_2
L_2,L_3	2 meters, 2 turns B&W 3013 miniductor 6 meters, 7 turns B&W 3011 miniductor
L_4	2 meters, 1 turn hookup wire around L_3 6 meters, 2 turns hookup wire around L_3
SW_1	Dpdt slide or toggle switch
T_1	5- to 10-watt audio output transformer

Fig. 8-22. Utility amplifier for vhf. Resistor and capacitor values are shown on the schematic.

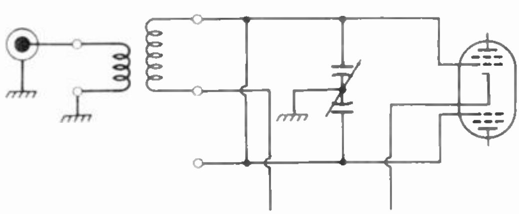
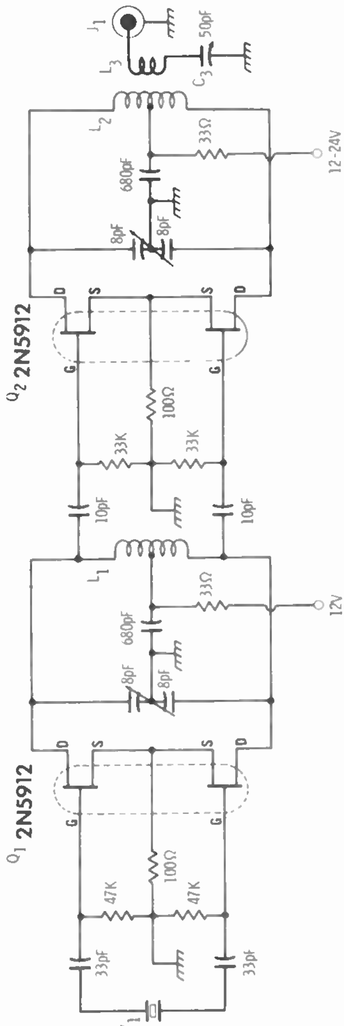


Fig. 8-23. Push-push input connections.

less can be given a good boost by the utility amplifier. It is a good way of giving a little extra boost to a low-power solid-state unit.



Parts List for Fig. 8-24.

C_1,C_2	8-pF butterfly capacitors (Johnson 160-208)
C_3	50-pF variable capacitor
L_1,L_2	7 turns B&W 3011 miniductor, centertapped
L_3	2 turns hookup wire around center of L_2
Q_1,Q_2	2N5912 dual FETs (Siliconix)

Fig. 8-24. Six-meter QRPP transmitter. Resistor and capacitor values are shown on the schematic. Rig. is built on a 9" x 4" perf board.

Operation

Tune up and operation are similar to the methods described previously in Projects 31 and 32. Again avoid tune-up problems by first setting each of the resonant circuits with a dip meter. You can tune up the amplifier while it is set to the class-C mode. Only a slight readjustment may be needed when changing over to linear operation. Use a jumper across the modulation input terminals when the unit is not to be amplitude modulated. To prevent flat-topping in the linear mode of operation, do not overdrive the stage. An oscilloscopic display is of particular help in adjusting the amplifier for good linear performance.

PROJECT 34: FET SIX-METER QRPP TRANSMITTER

The simple crystal oscillator and amplifier transmitter of Fig. 8-24 uses two Siliconix differential pairs 2N5912 FETs. These balanced pairs are effective in push-pull circuits. The crystal oscillator stage is a push-pull arrangement using an overtone crystal. The oscillator is followed by a similar amplifier.

The resonant circuits are tuned by two butterfly capacitors. Coils L_1 and L_2 consist of seven turns of B&W 3011 coil stock. Each coil is centertapped. The output coil (L_3) consists of two turns of hook-up wire wound around the center of L_2. The transmitter can be built on a perf board. A convenient socket for the FETs is a round 8-pin integrated-circuit socket.

Operation

The transmitter tunes and peaks very simply. Initially apply power to the oscillator only. Listen for the signal on the appropriate frequency on your six-meter receiver. You can peak the oscillator capacitor, using the receiver as an indicator.

Apply power to the amplifier and adjust the amplifier tuning and loading capacitors for maximum output. Some slight readjustment of the oscillator capacitor may be necessary. The output indicator of Chapter 1 can be used as a sensitive dummy load using a 50- to 70-ohm resistor. Without a terminating resistor it can serve as an output indicator when an antenna is connected to the output of the transmitter. The transmitter can be keyed in the common-source path to ground.

CHAPTER 9

Frequency Modulation: General Principles

Frequency modulation has several attractive advantages for radio communication use; it requires no significant audio power and has a very low susceptibility to noise and interference. Since a frequency-modulated wave is employed, the amplitude variations contributed by noise can be limited sharply at the receiver. The ability to eliminate amplitude impulse noise is an important advantage when the equipment must be operated in close proximity to ignition systems and other noise sources which radiate strong amplitude-varying rf signals. In a similar manner the beat variations (amplitude changes) between two interfering rf signals can be reduced. Thus, by using a frequency-modulation system there will be less interference between stations operating on the same or adjacent channels. In f-m systems the stronger station dominates the weaker station and pushes it into the background.

It is significant that f-m transmission requires a wider emission bandwidth. The emission bandwidth is more than twice the highest modulating frequency. The greater the frequency deviation (higher modulation index), the greater is the emission bandwidth per a given modulating frequency. However, bandwidth can be held down by establishing a limit for the highest audio frequency and by using a very low maximum-permissible frequency deviation. In the two-way radio services it is customary to limit the highest modulating audio frequency to 3000 hertz and the maximum deviation to ±5 kHz or ±15 kHz.

In an f-m system the frequency of the resultant rf wave is made to vary with the modulation. The extent of the frequency change, or deviation, varies with the *amplitude* of the modulating wave. The rate at which the frequency changes varies with the *frequency* of the modulating wave.

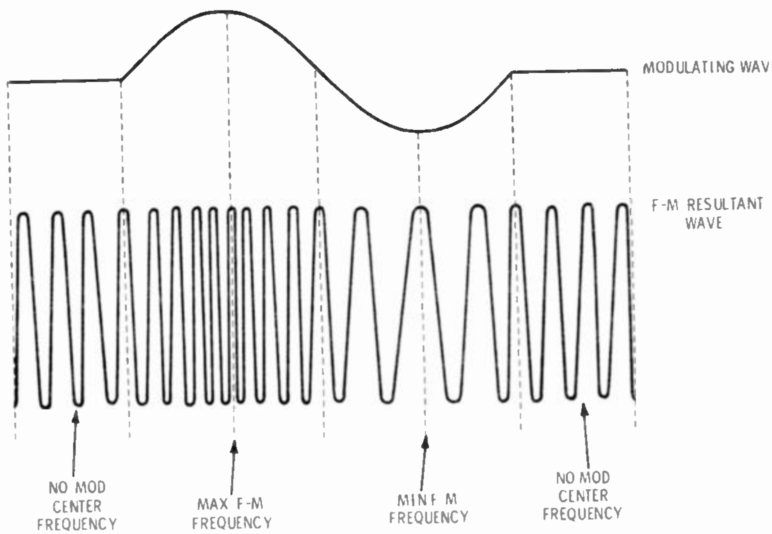


Fig. 9-1. The frequency-modulated wave.

As shown in Fig. 9-1, the actual amplitude of the resultant rf wave remains constant—only its frequency changes with the modulation. This latter characteristic means that any amplitude variations (noise) present in the resultant f-m signal can be safely removed because the desired signal information has nothing whatsoever to do with amplitude change. This elimination of undesired amplitude variations is accomplished in the f-m receiver.

DIRECT AND INDIRECT FREQUENCY MODULATION

The basic theoretical principal of the frequency-modulation process can be demonstrated with a motor-driven capacitor as shown in

Fig. 9-2. If such a capacitor has a linear frequency characteristic, and is placed across the tank circuit of an oscillator the frequency of the oscillator can be made to change with the capacitor rotor rotation. The oscillator frequency would of course be minimum when the capacitor plates are meshed fully and maximum when the capacitor plates are entirely out.

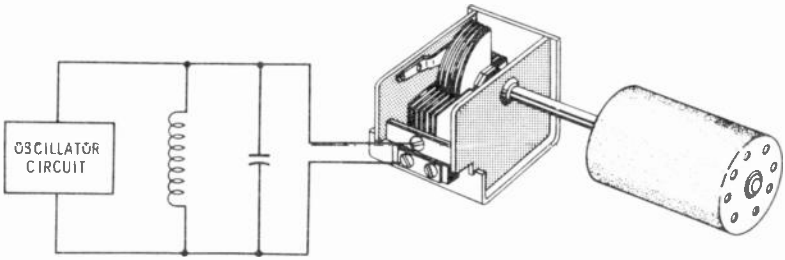


Fig. 9-2. A simple motor-driven method of changing oscillator frequency.

If the associated motor rotates the capacitor rotor at the rate of 100 revolutions per second, the frequency of the oscillator would also vary at the 100 cycle-per-minute rate. If the motor speed were increased or decreased there would be a similar change in the rate of frequency change. This would be the same as changing the frequency of the modulating wave in a practical f-m system.

If the value of the capacitor is reduced, the range over which the frequency of the oscillator could change would be reduced. The lesser deviation would correspond to the application of a weaker modulating wave to the input of a practical f-m system.

Were it practical for a rotating motor to follow the complex variations of speech a circuit of the type shown in Fig. 9-2 could be used to generate an f-m wave for a two-way radio system. However, a similar and practical deviation can be accomplished with the use of a vacuum-tube or transistor stage designed to place a variable capacitance or variable inductance across the tuned circuit of an oscillator. This type of circuit is called a *reactance modulator*.

An f-m wave can also be formed by switching the phase of an rf wave in accordance with the variations of a modulating signal. This technique is used extensively in f-m two-way radio systems and the circuit is referred to as an f-m *phase modulator*. The source of the signal is usually a crystal oscillator, vxo, or vfo.

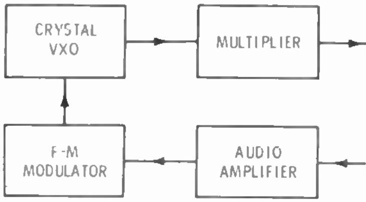
When an f-m wave is formed by deviating the frequency of the oscillator with a reactance device it is called *direct frequency modulation.* The oscillator can also be any one of the above types, crystal or vfo.

In most f-m units the fundamental oscillator operates at a much lower frequency than the assigned transmit channel frequency. A series of multiplier stages then follows the low-frequency oscillator to increase the frequency to the desired transmit channel frequency. As the frequency is multiplied, any deviation of the fundamental oscillator frequency will be increased by a like amount. To obtain a small linear frequency change, the actual oscillator deviation produced by the reactance circuit is quite limited. Assume the 8.5-MHz frequency of an oscillator is deviated by only ±2.5 kHz and there is a multiplication factor of six. The final transmit frequency will be 51 MHz (8.5 × 6) and the maximum deviation ± 15 kHz (6 × 2.5). From this example it can be seen that frequency-multiplier stages increase the small frequency deviation of the fundamental oscillator to the maximum permissible deviation desired.

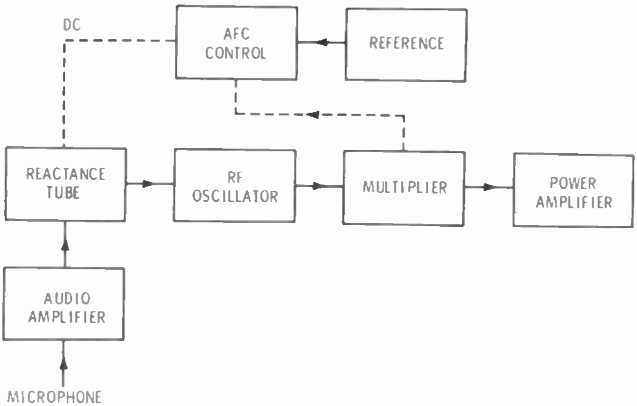
Multiplier stages used in an f-m transmitter can be conventional class-C amplifiers. The f-m wave, which carries the modulating information, is in the form of a changing frequency. There is no danger of distortion because of the nonlinear amplitude characteristic of a class-C stage. The standard a-m and sideband modulation methods require linear amplification, after the modulation envelope has been formed, in order to retain the original modulation. In fact, the limiting activity of a class-C stage in terms of amplitude change is used to advantage in an f-m system. In the f-m process undesired amplitude variations may also occur simultaneously with a desired frequency change. These undesired amplitude variations are eliminated by the class-C multipliers, thus producing a constant amplitude f-m signal output.

Very little audio power is necessary in the modulation of an f-m wave as compared to a standard a-m system. A high-impedance circuit that produces ten to twenty volts of audio will usually suffice. This is to be expected in the frequency-modulation process because no additional rf power must be generated. The power in a resultant f-m wave remains constant with modulation; it is simply distributed in differing amounts among carrier and sidebands, thus producing a changing frequency but a constant amplitude output.

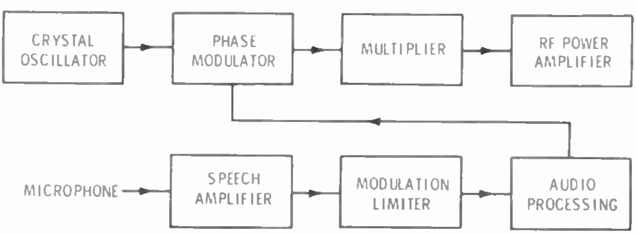
One of the problems with direct frequency modulation is the center-frequency stability (Fig. 9-3A). It must be held constant to



(A) Direct frequency modulation.



(B) An automatic-frequency-control system.



(C) Indirect f-m system.

Fig. 9-3. Functional plans of f-m systems.

maintain the proper operating stability and prevent distortion. Both the oscillator and associated reactance device must be carefully designed for the most stable operation. A stable crystal oscillator or vfo is needed. An alternative is an afc system (Fig. 9-3B) that compares the center frequency of the transmitter to some crystal-controlled standard. A dc error voltage is developed at the output of the afc circuit and, through the reactance device, it alters the center frequency an amount necessary to correct any frequency drift.

A block diagram of an indirect frequency-modulation system is shown in Fig. 9-3C. In this case the rf carrier is phase-modulated (p-m). Phase (or angle) modulation, like direct frequency modulation, can be accomplished in a number of ways. Using a suitable correction arrangement, the phase-modulation process can be used to generate a true f-m wave.

Most indirect systems use a crystal oscillator or vxo. Thus the center-frequency stability is good and an afc circuit is not required. A phase-modulation process is limited in the amount of deviation that can be obtained; however, this is an unimportant consideration in two-way radio systems because they are predominantly narrow-band limited-deviation systems.

In the phase-modulation process used in two-way commercial and radio amateur transmitters the audio variations continuously vary the phase of the rf carrier wave. They do so by placing a changing reactive component in the path of the rf carrier wave. This produces a changing phase shift. The higher the amplitude of the audio wave, the greater is the phase shift and the more the rf wave is delayed during one alternation. Conversely, on the opposite alternation of the audio wave the phase shift is of opposite sign and the rf wave is advanced instead of delayed. If the phase modulator speeds up the rf wave its period is reduced. This means it has a higher frequency (Fig. 9-4A). If the rf wave is delayed, its period is increased and it has a lower frequency (Fig. 9-4B).

In practice the phase shift varies throughout the entire audio cycle reaching the maximum value only at the sine-wave peaks. Positive and negative peaks produce the same phase shift but of opposite sign (lag and lead). The phase shift is zero when the audio wave passes through its zero. Therefore the frequency of the rf wave deviates above and below the carrier value following the amplitude and polarity changes of the audio wave (Fig. 9-4C).

An increase in the modulating frequency increases the frequency deviation of the carrier. The amount of frequency modulation produced indirectly by phase modulation varies directly with the frequency of the modulating signal.

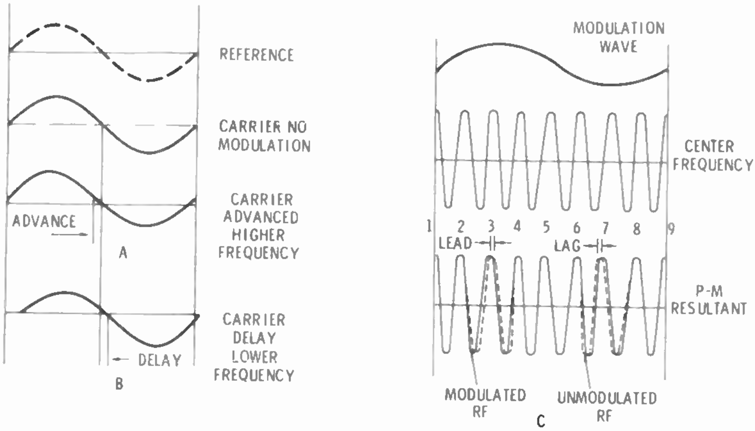


Fig. 9-4. Phase-modulation waveforms.

The amount of frequency deviation produced is relatively small. Usually a low carrier frequency is chosen to be phase modulated. After the f-m wave has been generated by the indirect process it is applied to a series of multipliers. These multipliers increase both the frequency and frequency deviation. As in the direct f-m system, the multipliers increase the deviation to maximum permissible value allowed for the particular service.

There is one fundamental difference between frequency and phase modulation. In a direct f-m system, the *frequency deviation is constant* for a given audio amplitude regardless of the audio frequency. The modulation index formula indicates that a change in audio frequency will cause a change in the modulation index even though the deviation remains fixed.

$$\text{Modulation index} = \frac{\text{frequency deviation}}{\text{modulating frequency}}$$

Thus in a direct f-m system *the modulation index decreases with modulating frequency,* assuming a fixed deviation.

In a phase-modulation process, it is the *modulation index that remains constant* for a given amplitude regardless of audio frequency. Inasmuch as the modulation index is constant the deviation itself must vary with the audio frequency.

$$\text{Frequency deviation} = \text{modulation index} \times \text{modulating frequency}$$

Actually, the higher the audio frequency of a given amplitude, the greater is the frequency deviation when using phase modulation.

It is apparent from the above paragraphs that in generating a true f-m wave using the phase-modulation process some form of compensating circuit must be used ahead of the modulator. Such a circuit is called a predistorter and is inserted between the audio source and the phase modulator.

The predistorter has a response that declines with an increase in frequency. Thus the output of the predistorter at a high audio frequency is less than the output at a lower frequency for the same input amplitude. Its response compensates for the fact that a phase modulator produces a greater deviation for a high-frequency modulating wave than for a lower one of the same amplitude. The compensation introduced by the predistorter is equal and opposite to that of the modulation process. Therefore, with predistortion it is possible to generate a true f-m wave using the phase-modulation process.

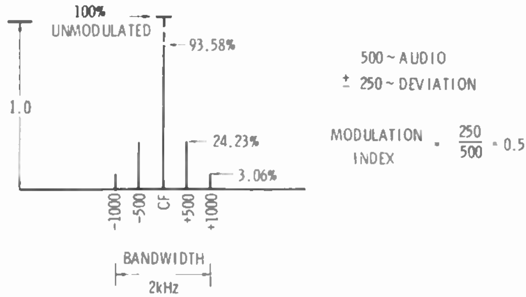
In the usual f-m two-way radio system there is an essentially narrowband audio-frequency range, and any necessary predistortion can be handled very simply by proper control of overall amplifier frequency response. The associated modulation limiters and integrators inherently suppress the high-frequency components and prevent them from causing excessive deviation of the center frequency. Thus it is possible in some cases, to handle predistortion without the actual use of a predistorter network.

COMPOSITION OF F-M WAVE

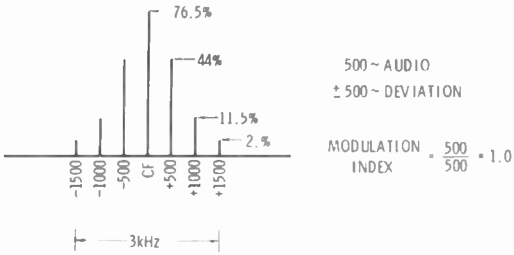
The resultant f-m wave, like an a-m envelope, is the algebraic summation of a number of individual rf waves. In the f-m process, when modulating with a single-frequency tone, a number of sideband pairs are generated according to modulating frequency and deviation. This is unlike the a-m process which generates only a single pair of sidebands when modulation is by a single-frequency tone.

In a standard a-m system, the carrier remains constant in amplitude; however, the algebraic addition of the two sidebands to the carrier produces an amplitude changing resultant (a-m envelope). In an f-m modulation system the amplitude of the resultant wave remains constant while the amplitude of the carrier changes. The carrier amplitude change is such that when added to the various sideband pairs, a constant-amplitude but changing-frequency resultant is produced.

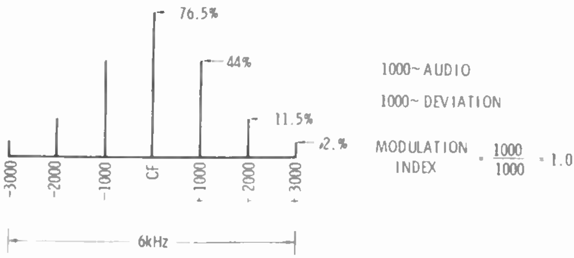
Actually the ratio of the powers in the sidebands and in the carrier changes with the modulating frequency and deviation. However, the resultant power output is always a constant. In fact, under specific conditions of modulating-wave amplitude and frequency, it is possible that the carrier or center-frequency power can even fall to



(A) ±250-Hz deviation, 500-Hz audio modulation.



(B) ±500-Hz deviation, 500-Hz audio modulation.



(C) ±1000-Hz deviation, 1000-Hz audio modulation.

Fig. 9-5. Spectrum distribution of various f-m waves.

zero. In this case, all of the power is in the sidebands. Under other conditions of frequency modulation, most of the power is in the carrier and only a small amount in the sidebands. In fact, the evaluation of shifts in power and relative changes of amplitude among sidebands and carrier can be useful in checking the performance of an f-m system.

The number of sideband pairs generated in the f-m process depends on the modulation index. The emission bandwidth, in turn, depends on the modulation index at maximum permissible deviation of the highest-frequency modulating wave. Some possible frequency distributions are illustrated in Fig. 9-5.

The wave equations and solutions that result when one rf wave is frequency-modulated by another lower-frequency wave are very complex. Such equations can be solved using Bessel functions. Despite the complex solutions very practical mathematical results can be obtained using a table of Bessel functions. Table 9-1 provides a set of Bessel-function solutions that are applicable to the narrowband frequency modulation used in two-way radio. The table gives the

Table 9-1. Bessel Function Chart Applied to Carrier and Side Frequencies of an F-M Wave

MODULATION INDEX	CARRIER (f_c)	$f_c \pm f_m$	$f_c \pm 2f_m$	$f_c \pm 3f_m$	$f_c \pm 4f_m$	$f_c \pm 5f_m$	$f_c \pm 6f_m$
0	1.00	0					
0.1	0.9975	0.0499					
0.2	0.99	0.0995					
0.3	0.9776	0.1483					
0.4	0.9604	0.1960					
0.5	0.9385	0.2423	0.0306				
0.6	0.912	0.2867	0.0437				
0.7	0.8812	0.329	0.0589				
0.8	0.8463	0.3688	0.0758				
0.9	0.8075	0.4059	0.0946				
1.0	0.7652	0.4401	0.1149	0.0196			
1.2	0.6711	0.4983	0.1679	0.0329			
1.4	0.5669	0.5419	0.2073	0.0505			
1.6	0.4554	0.5699	0.257	0.0725			
2.00	0.2239	0.5767	0.3528	0.1289	0.034		
3.00	0.2601	0.3391	0.4861	0.3091	0.1320	0.0430	
4.00	0.3971	0.066	0.3641	0.4302	0.2811	0.1321	
5.00	0.1776	0.3276	0.0466	0.3648	0.3912	0.2611	0.131

relative magnitudes of carrier (center frequency) and side frequencies for various modulation indexes. The unity reference represents the magnitude of the carrier when unmodulated. A sideband frequency is considered significant when its magnitude is more than 2 percent of the magnitude of the unmodulated carrier.

For example, when the f-m modulation is such that the modulation index is 0.5, the magnitude of the center-frequency component is 93.85 percent of the unmodulated center-frequency level, the first pair of sidebands has a magnitude that is 24.23 percent of the unmodulated center-frequency value and finally, a second sideband pair has a magnitude 3.06 percent of the unmodulated center-frequency value. If the modulating frequency were 500 Hz, there would be a pair of sidebands, one at each side of the center frequency, displaced by 500 cycles. Each sideband, upper and lower, would have a magnitude of 24.33 percent of the unmodulated carrier value. Furthermore, there would be a second pair of sidebands displaced by 1000 cycles (2 × 500) from the center frequency. These frequencies would have a magnitude of 3.06 percent. A simple spectrum distribution graph (Fig. 9-5A) can be used to illustrate relative magnitudes among the wave components and the required emission bandwidth for transmitting this f-m signal.

If we increase the magnitude of the modulating wave and thus cause a greater deviation, the emission bandwidth will increase because an additional significant sideband pair will be transmitted. Refer to Fig. 9-5B.

The higher the audio frequency for a given index, the greater is the separation between side-frequency pairs of significance, and the greater is the required emission bandwidth. Compare examples B and C in Fig. 9-5.

Collate these graphs with the information given in the Bessel Function Table. The results prove that the emission bandwidth increases both with frequency and deviation. Therefore, in the narrowband two-way radio services it is necessary to hold down the emission bandwidth by limiting both the extent of the deviation of the center frequency and the highest permissible audio frequency.

The maximum emission bandwidth assigned to a given radio service is a function of the maximum permissible deviation and the highest modulating frequency. The modulation index for this important operating-condition is called the *deviation ratio:*

$$\text{Deviation ratio} = \frac{\text{maximum permissible deviation}}{\text{highest modulating frequency}}$$

Calculations for the two-way radio deviations of ±5kHz and ±15kHz, based on a maximum permissible modulation frequency of 3000 Hz become:

$$\text{Deviation ratio} = \frac{5000}{3000} = 1.67$$

$$\text{Deviation ratio} = \frac{15000}{3000} = 5$$

Under FCC rules and regulations 97.65 (C) on frequencies below 29 MHz and between 50.1 and 52.5 MHz, the bandwidth of an F3 emission (frequency or phase modulation) shall not exceed that of an A3 emission having the same audio characteristics; and the purity and stability of emissions shall comply with the requirements of 97.73.

From the above it is obvious that on the radiotelephone portions of bands 15 through 80 meters, the frequency-modulation emission must not be any broader than a corresponding conventional a-m signal. Theoretically the deviation ratio (maximum bandwidth modulation index) should not exceed 0.5. This means that with an upper audio frequency of 3000 hertz, maximum permissible deviation would only be ±1.5 kHz. This applies on 10 meters to the frequency spectrum below 29 MHz as well as the 6-meter spectrum between 50.1 and 52.5 MHz. Frequency-modulation F3 emission is not permitted on the 160-meter band. In practice a deviation ratio of 1.0 is acceptable because the second pair of sidebands will not be too great and maximum deviation at highest audio frequency does not occur too often.

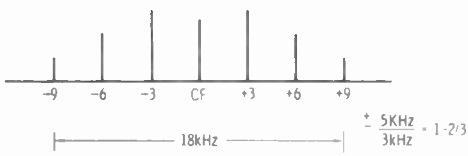
Wider band f-m emission is permitted on the vhf-uhf bands above 52.5 MHz. It is also permitted in the 10-meter spectrum between 29.0 and 29.7 MHz.

There is very little f-m activity below 29 MHz. The most active band for f-m is 2 meters with 6 meters following after. There is some f-m activity on the other vhf-uhf bands and a limited amount at the high-frequency end of the 10-meter band. Most common deviation is ±15 kHz and next ±5 kHz. Some operate with a deviation of

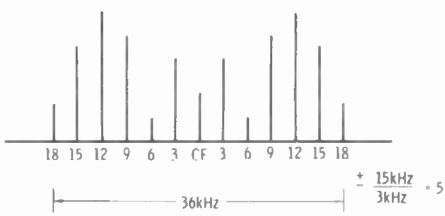
±7.5 kHz so that a reasonable signal can be received whether the receiver operates narrow-band ±5 kHz or the wider band ±15 kHz.

It should be noted from Fig. 9-5A that with a modulation index of 0.5 there is only one significant sideband pair. The second sideband pair is of very low amplitude and drops off to an insignificant value below the 0.5 index. A conventional amplitude-modulated wave also has but a single significant sideband pair. Therefore the emission bandwidth for conventional a-m signal and an f-m signal with a deviation ratio of 0.5 to 1.0 is approximately the same.

The bandwidth for a deviation of 1.67 (±5 kHz) and 5 (±15 kHz deviation) are substantially greater as shown in Fig. 9-6. The emission bandwidth for the greater deviation is approximately twice that of the ±5-kHz deviation.



(A) ±5-kHz deviation, 3-kHz audio modulation.

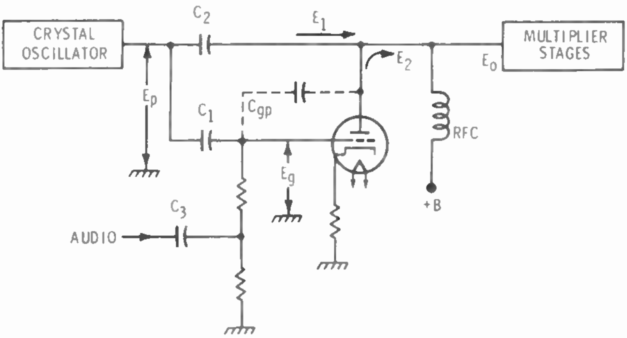


(B) ±15-kHz deviation, 3-kHz audio modulation.

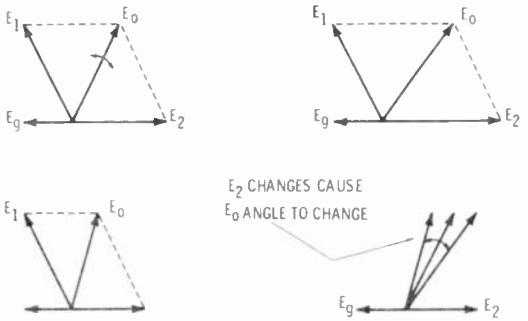
Fig. 9-6. Emission characteristics for maximum deviations at highest modulating frequency.

TYPICAL CIRCUITS

In most f-m transmitters the modulation occurs at low frequency using the basic plan of Fig. 8-4 in the previous chapter. Narrowband deviation is the rule and, consequently, the most common frequency modulators are either the phase modulator or direct frequency modulation of a crystal oscillator using a voltage-variable capacitor diode.



(A) Schematic.



(B) Vectors.

Fig. 9-7. Phase modulator and vectors.

An example of a triode phase modulator is given in Fig. 9-7. Such a phase modulator is inserted in the radio-frequency path between a vxo or stable vfo and the multiplier section of the transmitter. Its input circuit is designed to present minimum loading of the rf signal source, usually a crystal oscillator vxo. This precaution prevents the phase modulator from having any adverse influence on the frequency stability of the source. The purpose of the phase modulator is to change the phase of the rf signal in accordance with audio variations. The oscillator signal, before it reaches the first multiplier stage, will have been frequency-modulated by an indirect means.

The rf oscillator voltage (E_p) follows two paths. One path is through capacitor C_1 to the control grid of the phase modulator. This rf component is labeled E_g in the schematic and associated vector diagram. A second path leads from the oscillator to the plate circuit of the phase modulator by way of capacitor C_2. This rf component is shown as voltage E_1 in Fig. 9-7. The rf component E_g supplied to the control grid appears in the plate circuit of the phase modulator as voltage E_2. Since a tube normally produces a polarity reversal, it would seem that the two components E_1 and E_2 present in the plate circuit would be of opposite polarity or 180° out of phase. However, this is not the case because phase-shifting capacitors C_1 and C_2 are of low value and, along with the other component values, have been selected to establish a fixed and desired phase relation between the two rf components (E_1 and E_2) in the plate circuit of the phase modulator.

The vector diagram shows a practical relation between the direct component E_1 and the component E_2 that is a result of tube operation. One might assume that the E_2 voltage would be much greater in amplitude than the E_1 voltage; however, the capacitor values and the operating conditions of the phase modulator have been chosen to keep the two components at a comparable voltage level. A small amount of degenerative feedback, which results from the unbypassed cathode resistor, assists in equalizing the two rf output voltages.

In the output circuit, the direct voltage (E_1) remains constant in amplitude. Likewise the rf signal component (E_g) applied to the grid of the phase modulator is constant. However, the g_m of the tube varies with the applied audio modulating wave. Therefore the output voltage (E_2) varies in amplitude with the modulating wave. As shown by the vectors, the amplitude of voltage E_2 varies with respect to the constant amplitude of voltage E_1.

The vector diagram shows how the phase modulation is caused. The net rf output voltage (E_o), or resultant f-m wave, is the vector sum of E_1 and E_2. This resultant voltage changes in phase (angle $\emptyset$) as the amplitude of the E_2 component varies with the applied audio. The higher the amplitude of the applied audio, the greater is the phase deviation and the greater is the resultant frequency modulation. The higher the frequency of the applied audio, the more often the angle swings between two extremes.

When the amplitude of the E_2 component increases, the phase angle of resultant voltage E_o swings toward the E_2 vector. When the E_2 voltage decreases, the phase angle of E_o swings away from the E_2 vector. The angular deviation of the E_o vector stretches out and compresses the cyclic periods of the rf wave, producing frequency modulation. The last vector diagram shows how the angle deviates with the applied audio.

It should also be noted that the magnitude of the resultant vector (E_o) also changes with the modulation. Thus, initially, some amplitude modulation may also be present. However, the succeeding class-C multipliers remove the amplitude variation and a constant-magnitude and changing-frequency (f-m) wave is eventually produced. The frequency multiplier stages, of course, multiply both the center frequency and the deviation.

The operation of the transistor phase modulator, shown in Fig. 9-8, is similar to its vacuum-tube counterpart. The radio-frequency

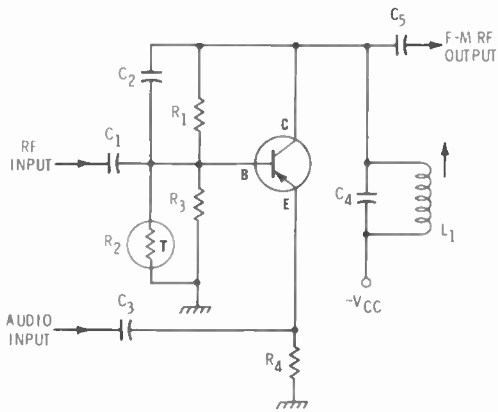
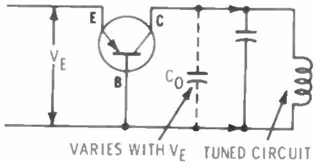


Fig. 9-8. Transistor phase modulator.

signal from the center-frequency oscillator is supplied to the base by way of capacitor C_1. Likewise there is a component fed into the collector output circuit by way of capacitors C_1 and C_2. The audio component is supplied to the emitter circuit. The base-divider bias circuit includes a thermistor to stabilize the operating point against temperature change.

The rf component fed into the collector output circuit via capacitor C_2 is of constant amplitude and phase. However, the rf component that develops in the collector circuit through transistor operation is varied in magnitude by the audio signal applied to the emitter. A change in the resistance (mutual conductance) of the path between the base and collector with emitter voltage change causes a variation in the magnitude of the rf component developed in the output circuit. The relative change between this rf output and the steady rf component produces the phase-modulated resultant.

Transistors can also be used as variable reactance devices. The output capacitance of a transistor, the capacitance of the collector junction (C_o) specifically, can be made to vary quite readily. The collector junction is reverse-biased and the depletion area of the junction depends on the junction voltage. When a higher reverse voltage is applied to the junction the depletion area increases and reduces the junction capacitance. An audio signal applied to the base or emitter circuits can cause this capacitance to change. If the capacitance is shunted across a tuned circuit there will be a corresponding change in its resonant frequency (Fig. 9-9A). In fact, the transistor can be used as an oscillator as shown in Fig. 9-9B. The oscillator is



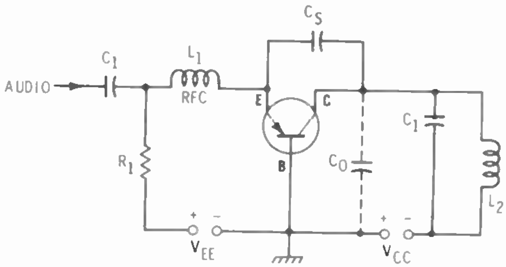


Fig. 9-9. Self-reactance modulator-oscillator system.

connected in a common-base configuration with the small capacitor C_s providing the necessary feedback between collector and emitter circuit to sustain self-oscillation. The audio signal is supplied to the emitter and the resultant change in the collector-to-base voltage will cause the transistor-output capacitance (C_o) to vary and cause a change in the frequency of oscillation. This type of circuit is referred to as a self-reactance f-m oscillator.

VOLTAGE-VARIABLE CAPACITOR

One of the effects of a changing voltage across a semiconductor diode is a change in capacitance. By suitable semiconductor design and doping, the change in capacitance can be emphasized and made to have a capacitance-versus-junction voltage characteristic that will permit it to operate as an efficient voltage-to-frequency converter.

A voltage-variable capacitor, placed across an oscillator tank circuit as shown in Fig. 9-10, can be used to control or vary the reso-

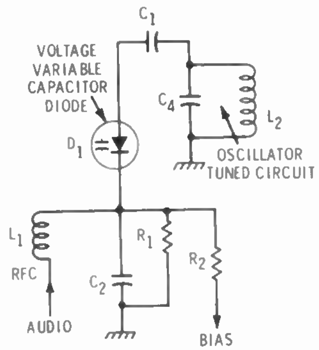


Fig. 9-10. Using voltage-variable capacitor diode to vary frequency of a resonant circuit.

nant frequency of the tuned circuit. As shown the modulating voltage is supplied to the diode by way of an isolating rf filter. A voltage source sets the dc bias of the diode at a point on its voltage-capacitance transfer characteristic that ensures a linear deviation of the oscillator signal. As this bias point is made to vary with the positive and negative excursions of the modulating wave there is a corresponding change in diode capacitance and in the resonant frequency of the tuned circuit with which the diode is associated. A capacitive divider arrangement (capacitor C_1 and voltage-variable capacitor D_1) may be used to regulate the extent of the capacitance deviation for a given magnitude of audio input signal. Such an adjustment may be used to ensure a linear frequency deviation for a given maximum level of audio input signal.

The variable-capacitance diode can be used to either frequency- or phase-modulate an rf signal. In the simple phase modulator of Fig. 9-11A the variable-capacitance diode (D_1) is a part of a reso-

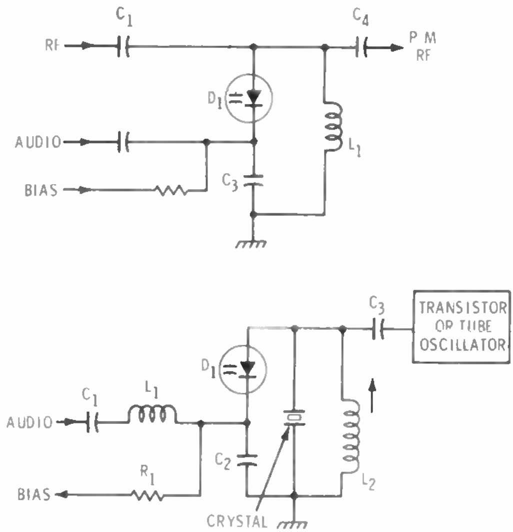


Fig. 9-11. Variable-capacitance diode as a frequency or phase modulator.

nant tank circuit that follows the oscillator. The coupling capacitor (C_1) from the oscillator to the resonant circuit has a rather high reactance at the output frequency. Thus, the audio variation that is applied across the voltage-variable capacitor is able to change the resonant frequency of the tank circuit but not the oscillator frequency. However, the change in the resonant frequency causes a change in the phase angle of the rf signal delivered by the oscillator as it is developed across the tank circuit. This is to be anticipated because with a change in the frequency of the tuned circuit away from the frequency of the oscillator, the tuned circuit will present an impedance that now has a reactive component at the oscillator frequency.

The unmodulated resonant frequency of the tuned circuit depends on the LC constant and the capacitance contributed by the variable-capacitance diode (at the established dc bias). In the case of no applied modulation the oscillator frequency and the resonant frequency of the tuned circuit are one and the same. However when an audio signal is applied to the variable-capacitance diode there will be a corresponding change in the diode capacitance. This will cause the resonant frequency of the tuned circuit to change. In effect the tuned circuit will now present a changing reactive component and the individual rf cycles will be shifted in phase. As in any form of phase modulator this change in phase will compress and expand the periods of individual rf cycles and frequency modulation will have been produced by an indirect means.

It is also possible to connect the variable-capacitance diode directly into the crystal circuit of an oscillator as shown in Fig. 9-11B. Often an overtone-crystal oscillator is used. The resonant circuit consists of inductor L_2, capacitor C_3, and the variable-capacitance diode (D_1). An applied audio signal will cause a change in the capacitance and a limited deviation of the crystal oscillator frequency will result. By using an overtone oscillator, this change in frequency can be made substantial and narrowband frequency modulation will result. The deviation, of course, can be increased by using succeeding multiplier stages.

A practical variable-capacitance frequency modulator used in conjunction with a Clapp-type crystal oscillator is shown in Fig. 9-12.

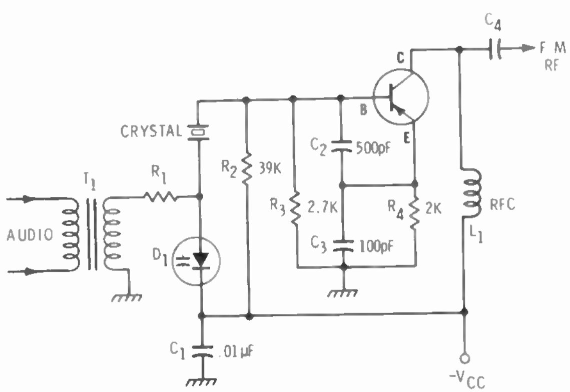


Fig. 9-12. Crystal oscillator with variable-capacitance diode f-m modulator.

The crystal operates in its series-resonant mode. Capacitors C_2 and C_3 are of high value and act as swamping capacitors. Their values are high enough to swamp out any capacitive changes that may be contributed by the transistor due to heating or other circuit changes.

The frequency of the oscillator is determined by the resonant characteristics of the crystal and the capacitance added in series by the variable-capacitance diode. In effect, the variable-capacitance

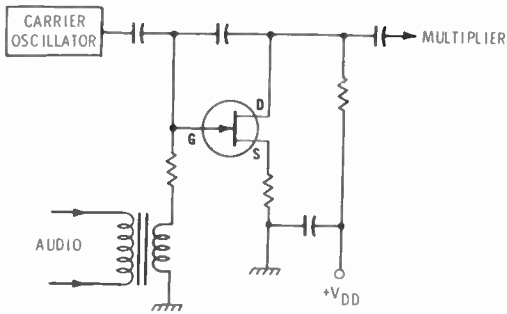


Fig. 9-13. FET phase modulator.

diode is adding additional series capacitance to the series-resonant characteristic of the crystal when it is oscillating in its series mode. It is apparent, then, that any changes in the capacitance of D_1 with the application of an audio wave will cause a corresponding deviation in the frequency of the crystal oscillator.

The field-effect transistor and its vacuum-tube-like characteristics permit its use as either a direct reactance modulator (oscillator and voltage-variable capacitor diode combination) or a phase modulator. A typical phase modulator circuit is shown in Fig. 9-13. Again the rf component to be modulated is applied directly to the gate and also through a very low value capacitor to the drain circuit. The audio signal is applied through an output transformer to the gate. Signal source can be a vxo crystal oscillator. The phase-modulated component in the drain circuit is applied to a follow-up amplifier and multiplier stages.

An example of a bipolar crystal oscillator that is frequency-modulated with a voltage-variable capacitor diode is shown in Fig. 9-14. A fundamental-frequency crystal is used but the oscillator tank circuit is tuned to the third harmonic. Inductor L_1 permits the oscillator to be set precisely on a desired f-m channel.

The voltage-variable capacitor completes the crystal circuit to common. Audio is introduced by way of the radio-frequency choke.

PROJECT 36: VACUUM-TUBE F-M MODULATOR

There are various methods of frequency-modulating a crystal oscillator. One simple technique which is not used too often is to

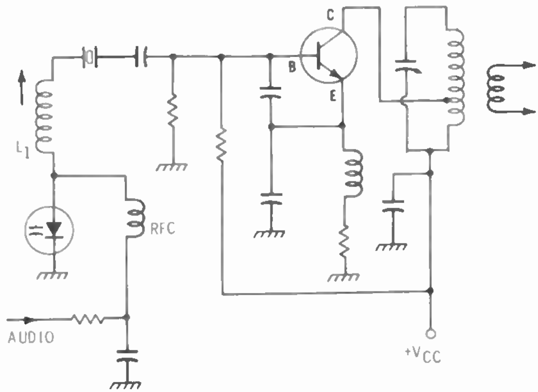
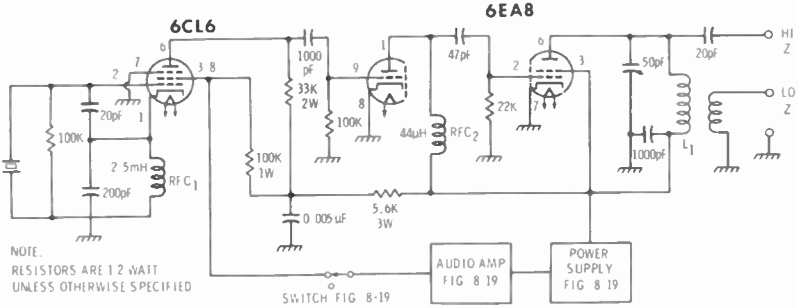


Fig. 9-14. Direct frequency modulation of a bipolar transistor using voltage-variable capacitor.

apply the modulating wave to the screen grid of a pentode crystal oscillator. The input capacitance of the tube is a function of the screen-grid voltage. If the voltage is made to vary, there will be a like change in input capacitance. Inasmuch as this capacitance is felt across the crystal, there will be a limited deviation of the crystal frequency. Such limited deviation of an 8-MHz crystal oscillator results in suitable narrowband f-m on the 6- and 2-meter bands.

Some amplitude modulation also occurs but this is removed by limiting action in the succeeding stages. Good isolation between stages is important to prevent unfavorable loading of the modulated stage and shift of the crystal center frequency.



L_1	10-meter coil set (Project 7)	1 Audio amplifier and power supply from Fig. 8-19.
RFC_2	44-μH rf choke (Ohmmite Z-14)	

Fig. 9-15. Vacuum-tube f-m exciter. Resistor and capacitor values are indicated on the schematic.

A simple and effective circuit is shown in Fig. 9-15. This is a two-stage affair with only a single resonant circuit. There is only one resonant tuning adjustment required and it is well isolated from the frequency-modulated stage itself.

The output can be applied to the input of the second stage of the 6-meter transmitter of Fig. 8-17. In this case the vxo stage of that transmitter is inactivated.

The output of the f-m exciter is in the 24-MHz range and is capacitively coupled to the grid of the 6CX8 tripler of the circuit shown in Fig. 8-17. In this project the tripler is used as a straight-through amplifier. The audio amplifier uses a dual triode and is identical to the input voltage amplifier of the modulator shown in Fig. 8-19. The 10-meter coil of Fig. 2-7, Project 7, tunes to the 24-MHz range.

Operation

The first step again is to use a dip meter to set all the resonant circuits on frequency. The output resonant circuit of the exciter, for example, will tune to the second and third harmonics of the 8-MHz crystal. Use the third harmonic for driving the 6-meter transmitter. The 6-meter transmitter is tuned in accordance with the instructions given in Project 31. If you insert an 8.4-MHz crystal, the output frequency will be 50.4 MHz. The transmitter can also be used on 2 meters using the additional data given in Project 32.

Apply power but no modulating signal to the transmitter and peak all of the resonant circuits by touching up slightly each control. Recheck the transmitter stages using your dip meter as an absorption wavemeter. Your dummy load should indicate a power output of 5 to 9 watts on 6 meters.

While power is being supplied to the dummy load, tune in signal on an f-m receiver. The signal can also be demodulated with an a-m receiver if you slope-tune the receiver. Apply an audio tone to the input of the exciter and increase the speech gain control until you obtain a strong receiver output. If your test receiver is designed for narrowband f-m reception the output will distort if your speech gain is set too high.

Deviation can also be checked on a spectrum oscilloscope or using the carrier-zero technique and a selective receiver as discussed earlier in this chapter. The latter measurement can be made on any one of the harmonics as well as at the crystal frequency. Remember, however, that the modulation index is always the quotient of the desired deviation and the modulation frequency. For example, if the maximum desired deviation on 50.4 MHz is ±10 kHz, the desired deviation at 25.2 MHz would only be ±5 kHz. Furthermore the required deviation of the crystal oscillator would only be ±1⅔ kHz. Thus the modulation indices would be:

$$\text{Mod. index}(50.4) = \pm 10/\text{audio freq.}$$
$$\text{Mod. index}(25.2) = \pm 5/\text{audio freq.}$$
$$\text{Mod. index}(8.4) = \pm 1\tfrac{2}{3}/\text{audio freq.}$$

The first carrier-zero modulation index is 2.405. A carrier zero at this modulation index with the application of a 2000-hertz tone indicates a deviation of ±4.81 kHz (2.405 × 2).

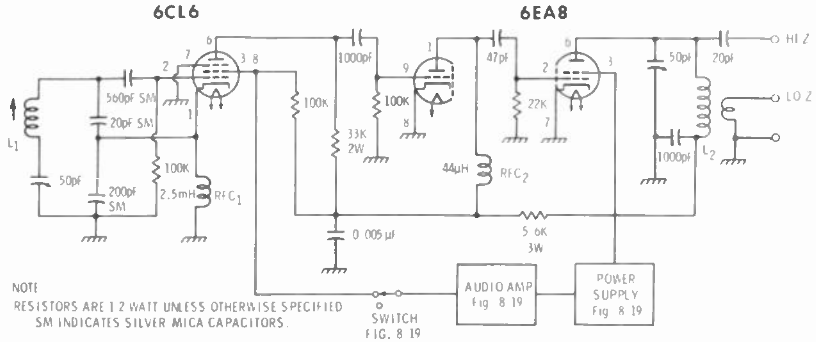
If possible tune in the f-m signal on 25.2 MHz (third harmonic of the crystal frequency). Start from zero level and increase the strength of the 2000-hertz audio tone until the carrier dips toward zero. At this level of applied audio you are obtaining a deviation near ±5 kHz. Therefore on 50.4 MHz you are obtaining a deviation of ±10 kHz, the deviation multiplying with the frequency.

If you now measure the magnitude of the 2000-hertz tone being applied to the input of the exciter you will know that a speech signal of the same peak amplitude produces ±10 kHz deviation on the 6-meter band.

Not as much audio is required for ±10 kHz deviation on the 2-meter band because of the additional threefold multiplication. You know from Project 32 that there is a total multiplication of 18 rather than 6. Consequently a deviation at the crystal frequency need only be some 800 hertz to obtain a full 10 kHz deviation on 2 meters (10 kHz/12).

PROJECT 37: VFO FM EXCITER AND UTILITY MODULATOR

The circuits of Figs. 8-19 and 9-16 can be combined to provide a convenient f-m exciter and a-m modulator. The switching facility and a-m modulation capability were discussed in Project 31. The oscillator is a vfo and operates over the 7- to 15-MHz range. Therefore it accommodates all of the basic frequencies normally used for 6 meters, 2 meters, and shorter wavelengths. It can also generate the



L_1	Oscillator coil, 2.38- to 3.96-µH slug-tuned (J. W. Miller 21A105 RBI)	RFC_2	44-µH rf choke (Ohmmite Z-14)
L_2	Coil set from Project 7	1	Audio amplifier and power supply from Fig. 8-19

Fig. 9-16. Utility modulator and vfo f-m exciter. Resistor and capacitor values are indicated on the schematic.

basic signal (either 7.4 or 14.8 MHz) for f-m operation on the 10-meter band.

Audio signal is applied to the screen grid of the variable-frequency oscillator. The oscillator is stable and good frequency deviation is obtained. Good drive is obtained when the output resonant circuit is tuned to the same frequency as the oscillator. When the transmitter is adjusted for normal narrowband f-m modulation, amplitude variations of the output signal are virtually eliminated. Coil data can be obtained from Fig. 2-7 of Project 7. Lower level but reasonable output can also be obtained on the second and third harmonics of the oscillator frequency.

Operation

The first adjustment is to set the oscillator into the desired frequency range. This can be done by adjusting the slug in the coil to a setting which will permit the capacitor to tune over a preferred frequency range. Operation in the 8- to 9-MHz spectrum permits six-meter operation with a multiplier of 6 and two-meter operation with a multiplication of 18.

If the coil slug is set properly, the capacitor tuning can be made to fall between 8 and about 14 MHz. Thus you can also operate between 12 and 13.5 MHz, obtaining six-meter output with a multiplication of 4, and two-meter output with a multiplication of 12.

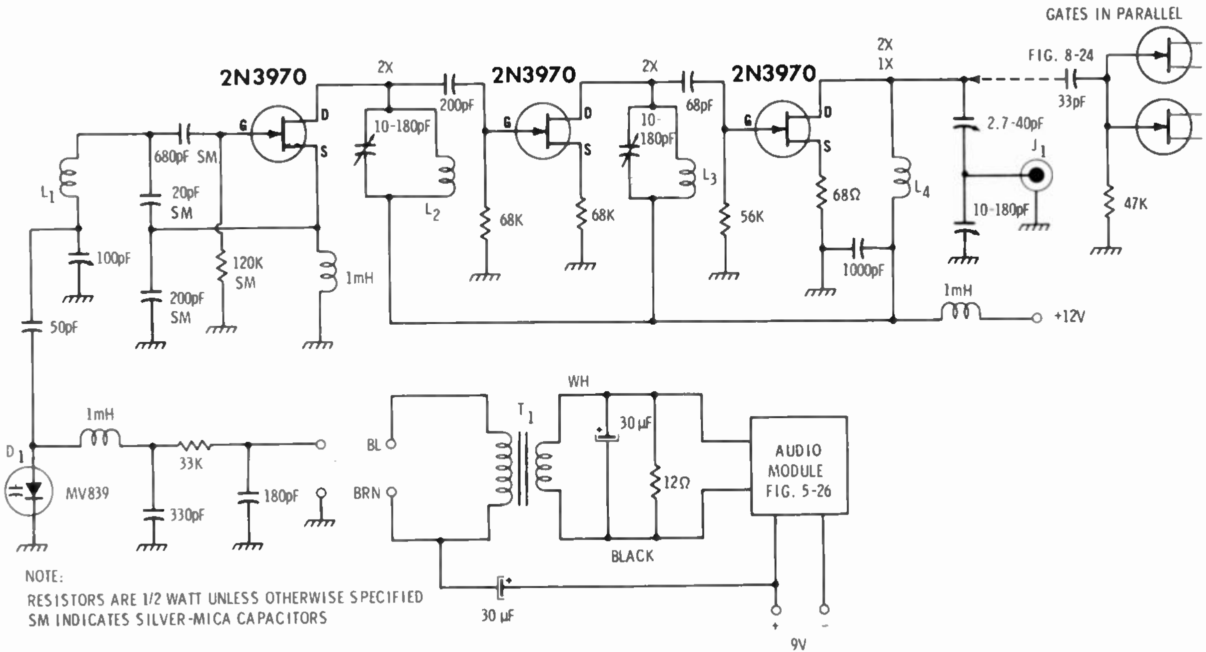


Fig. 9-17. FET f-m exciter. Resistor and capacitor values are shown on the schematic. Unit is built on a 9" x 4" perforated board.

The ten-meter f-m band is above 29.5 MHz. To operate on 29.6 MHz for example, the vfo would have to be set to 14.8 MHz for a multiplication of 2, or 7.4 MHz for a multiplication of 4. Adequate f-m deviation can be obtained even though the multiplication factor is only two.

The two tuning adjustments set the oscillator on frequency and peak the output resonant circuit. A high-impedance output provides drive to any high-impedance grid circuit. The low-impedance link output is used when the input resonant transformer of the multiplier includes a low-impedance input primary. In this arrangement a coaxial cable can be used to transfer the signal over a considerable distance between the vfo f-m exciter and the multiplier chain of the transmitter.

PROJECT 38: FET F-M EXCITER

Field-effect transistors are ideal for use in f-m exciter chains because of their vacuum-tube-like characteristics and simple circuitry. Reasonably high-*Q* resonant circuits are easy to obtain and good harmonic generation is possible. The 2N3970 performs well in vfo circuits and can be frequency-modulated with ease using a voltage-variable capacitor diode. Choices of oscillator and harmonic frequencies are many depending on the objectives of the builder. In the example of Fig. 9-17 an oscillator frequency range between 6 and 7.5 MHz was selected. By so doing a multiplication of 4 will set you up in the 10-meter f-m band; a multiplication of 8, in the 6-meter f-m spectra. The final stage is operated as a straight amplifier for ten-meter operation and a doubler for 6-meter output. If two-meter operation is desired the output stage operates as a doubler with the vfo operating in the 6-MHz range. Output of the exciter is then in

Parts List for Fig. 9-17.

D	MV839 Voltage-variable capacitor diode
L_1	10.8- to 18-μH slug-tuned inductor (J. W. Miller 21A336RBI)
L_2	1.08- to 1.8-μH slug-tuned inductor (J. W. Miller 21A156RBI)
L_3,L_4	.885- to 1.2-μH slug-tuned inductor (J. W. Miller 21A106RBI)
J_1	Coaxial output connector
1	Audio module and transformer (Fig. 5-26)
T_1	Transistor output transformer, pri 100 ohms C.T., Sec. 3.2/8/16 ohms

the 48-MHz spectrum and can be tripled by additional stages to the two-meter f-m band.

The vfo deviates readily with low-level audio. There is no need to bias the voltage-variable capacitor. The source of audio was the same arrangement used previously in Fig. 5-26 using one of the readily available audio modules plus an additional transformer that can be used to step up the impedance and voltage level ahead of the low-pass audio filter and voltage-variable capacitor diode.

Small trimmer capacitors and adjustable coils are used throughout. The adjustable slug is of aid in making each resonant circuit accommodate a desired frequency range. The output circuit is arranged to provide both low- and high-impedance outputs. The low-impedance output is used when there is a significant separation between the exciter and succeeding amplifiers. In one arrangement we placed the exciter very near to the 6-meter transmitter of Fig. 8-24. High-impedance drive was used and what was previously the crystal stage was connected as a push-push doubler as shown in the optional diagram of Fig. 9-17. In this case the final 2N3970 was operated as an amplifier. The first stage of the dual-FET 6-meter transmitter was operated as a doubler and supplied good drive to the final dual-FET amplifier which was operated straight-through to provide a good QRPP output level.

Operation

The tune-up procedure is straightforward. Again it is important to emphasize that a reasonably accurate dip meter and absorption wavemeter is essential. The vfo is operating at quite a low frequency and spurious harmonics with only 6- or 7-MHz separation are present. Therefore each stage should be preset on its desired frequency using a dip meter. After the transmitter is activated, the absorption wavemeter part of the dip meter is used to resonate each tuned circuit and make certain that it is operating on the correct harmonic.

Modulation level can again be checked on an f-m receiver. More precise measurement can be made on one of the harmonics that you can tune in with your high-frequency communications receiver. The carrier-zero technique can then be used to determine deviation, making certain that your receiver is adjusted for as narrow a receive bandpass as is possible.

CHAPTER 10

Test Equipment and Procedures

Several test instruments should be considered essential in testing radio transmitters; still others are helpful in deriving peak performance from transmitters. A means of measuring frequency rather accurately is most important. One can get by without accurate measurement if crystal control is employed. However, when operating under close tolerance or near band edges, or when using harmonics of a crystal, a means of frequency measurement is needed.

Power input and power output measurements are of significance especially when operating near FCC maximums. A dc voltmeter and current meter for the final amplifier should be incorporated when operating near the legal limit.

$$\text{Input power} = E \times I$$

where,

E equals voltage to the final amplifier,

I equals final amplifier current.

A power output meter, especially with a dummy load is a definite assist in tuning and judging the operating efficiency of the transmitter. An in-line wattmeter, or at least an SWR meter can give you that important clue as to how well your transmitter is delivering power to the antenna system.

Another important consideration is modulation. Be it cw, a-m, sideband, or f-m it is helpful to have some instrument that can tell you the degree of modulation and the quality of modulation. In the case of amplitude modulation there should not be overmodulation. In the case of sideband transmission there should be no flat topping. In the case of frequency modulation there should not be excess deviation. A transmitted cw signal should be a clean one.

Various other instruments such as dip oscillator, relative output indicators, absorption wavemeters, etc. are of definite benefit. These can help in tuning a transmitter and locating spurious signals as well as give you help in peaking adjustments and setting resonant circuits to the proper harmonic or fundamental frequencies.

FREQUENCY METER

A good receiver with a reasonably well-calibrated dial and a crystal calibrator can serve well as the station frequency meter. There are various types of actual frequency meters on the market but unless you pay a high price for them they are usually not as accurate as a well-calibrated receiver.

Many receivers include a built-in crystal calibrator. If your receiver does not have an accessory it is important that you purchase or build one. Such a calibrator can usually serve as a good secondary frequency standard once the crystal has been set precisely on one of the WWV frequencies. Of course, this calibration should be checked regularly.

Most frequency calibrators permit you to locate 100-kHz marker points over the amateur bands. Some include additional dividers that can be switched on to obtain 50-kHz, 25-kHz, and even 10-kHz calibration points. Such a calibrator is of great importance because of the several subdivisions of the amateur band such as cw, phone, novice, general, advanced, and extra segments.

TUNING INDICATORS

The basic radio-frequency indicator and frequency meter is the absorption wavemeter shown in Figs. 10-1 and 10-2. It is hardly more than a calibrated resonant circuit which, when coupled near the source of radio-frequency energy, can withdraw some of the energy. This meter will absorb maximum energy when tuned to the same resonant frequency as the source.

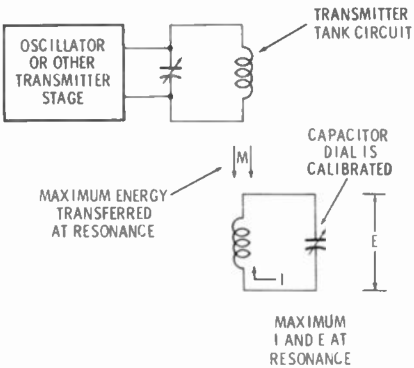


Fig. 10-1. Principle of the absorption wavemeter.

Some form of radio-frequency indicator such as a pilot lamp, neon bulb, or crystal-and-dc meter combination can be used to indicate the relative strength of the energy absorbed by the resonant circuit of the wavemeter. When the wavemeter is tuned to the frequency of the radio-frequency source, the indicator will read maximum; but if tuned to either side of the resonant-energy frequency, the meter reading or lamp brilliance will decrease.

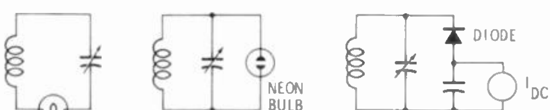


Fig. 10-2. Indicators for absorption wavemeters.

If the resonant frequency of the wavemeter tank circuit is known from the dial setting, the frequency of the radio-frequency energy can be determined. Usually the capacitor dial of the wavemeter is calibrated in frequency. Therefore, the setting of the pointer on the calibrated scale, when the capacitor is tuned for maximum rf indication, shows the frequency of the rf signal being measured. In other absorption wavemeters, the dial is calibrated from 0 to 100, and the resonant frequency of the absorption tank circuit is determined from a chart.

Wavemeters are usually equipped with replaceable coils so they can be made to operate on different frequency bands, with a separate dial or calibration chart for each band.

Energy from the tank circuit of the unit under test is coupled into the resonant circuit of the wavemeter. Use loose coupling to minimize loading the circuit under test, and couple the absorption wavemeter only close enough to the radio-frequency source to permit a usable rf indication (lamp glow or dc-meter reading).

To provide minimum loading and additional convenience, some absorption wavemeters include a low-impedance pickup coil. The main meter coil remains part of the wavemeter proper and is not coupled close to the rf energy being measured. The pickup loop is coupled to the source of energy and transfers the small amount of energy needed into the absorption-wavemeter circuit (Fig. 10-3).

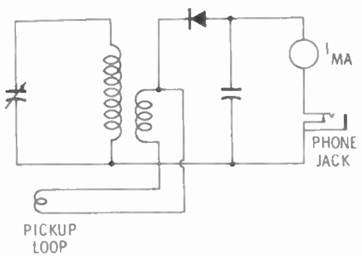


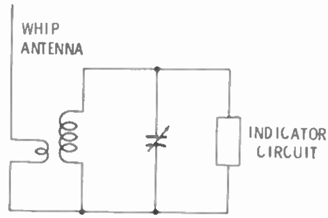
Fig. 10-3. Absorption wavemeter with pickup loop and phone jack.

The absorption wavemeter is important in making approximate frequency measurements while a transmitter is tuned, and, in particular, for checking multiplier harmonic outputs. It also serves as a good indicator of the strength of the radio-frequency energy. When set at a fixed position from the energy source, the influence of tuning on the radio-frequency output can be noticed immediately.

An absorption wavemeter can include a phone jack into which a pair of headphones can be plugged and any amplitude modulation on the rf signal can be heard. Hence, when the absorption wavemeter is used to check out rf amplifiers that convey a modulated rf signal, the quality of that modulation can be tested approximately, using the detector circuit that is part of many absorption wavemeters. For proper demodulation of an f-m radio-frequency signal, some form of f-m detector would have to be associated with the wavemeter.

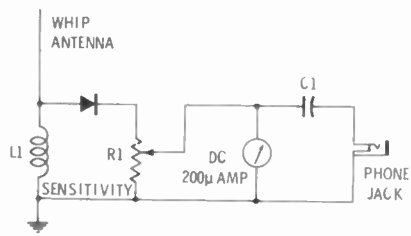
The wavemeter principle can also be used as a radio-frequency output indicator. In fact, it is no problem to attach a small antenna to a wavemeter (as shown in Fig. 10-4) to make it even more sensitive to the radio-frequency output of a transmitter.

Fig. 10-4. Use of the wavemeter principle in an rf indicator.



If the wavemeter is tuned to the operating frequency of the transmitter and placed somewhere in its immediate field, a strong rf indication can be obtained. It is, in effect, a field-strength indicator. With the meter positioned a fixed distance from the transmitter, various tuning adjustments can be made to maximize the rf output. In fact, the wavemeter, if placed in the field of the transmitter antenna, permits a rather good indication of the tuning of the antenna system and the effectiveness of energy transfer from transmitter to antenna.

Fig. 10-5. A typical rf-field indicator.



A nonresonant rf indicator is shown in Fig. 10-5. This is a very popular instrument for field checking the power output and tuning of communication transmitters. It consists of a crystal diode and a sensitive dc meter. The indicator is a simple rectifier and filter circuit. Capacitor C_1, inductor L_1, and potentiometer R_1 function as a filter to smooth out the unidirectional detector pulses. Thus the steady current flow through the dc meter is a function of the signal strength at the input. To improve the sensitivity of the device, particularly when hf and vhf signals are being checked, a short antenna can be attached. Sensitivity can be adjusted with potentiometer R_1. The amount of resistance inserted into the circuit influences the amplitude of the current flowing through the dc meter.

A tuning meter of this type, when placed near the transmitter or its antenna, will indicate the relative power output of the transmitter, and the effect of any transmitter adjustment on the power output can be noted.

It must be stressed that this type of rf meter, being untuned, is sensitive to an extremely wide frequency range. For this reason, a meter of this type cannot be used to check the transmitter frequency. The nonresonant indicator of Fig. 10-5 is excellent for checking the output of low-power transmitters and individual stages.

DIP OSCILLATOR

The dip oscillator is a useful test instrument for checking transmitter circuits. Most dip oscillators can also be used as absorption wavemeters.

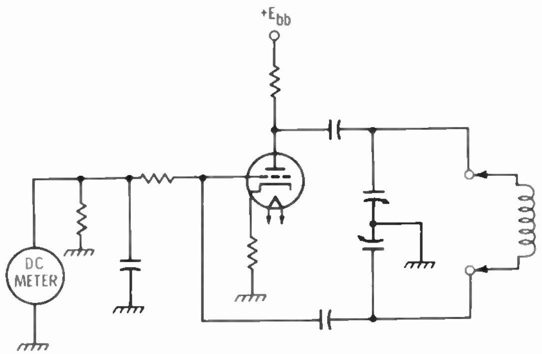
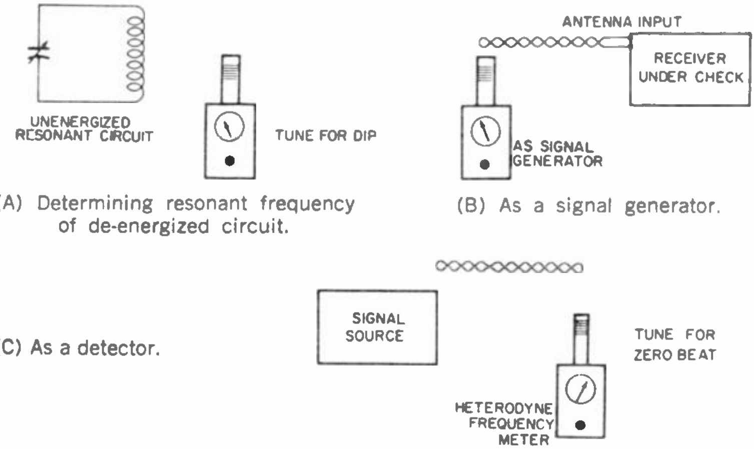


Fig. 10-6. Basic vacuum-tube dip meter.

The dipper contains a frequency-calibrated oscillator. The Colpitts oscillator, such as shown in Fig. 10-6, operates over a frequency range between 1.7 and 300 MHz. Seven plug-in coils are employed to cover overlapping frequency ranges between these two extremes. A sensitive dc meter is positioned in the oscillator grid circuit for measurement of the current there. The variable capacitor of the grid-dip oscillator is accurately calibrated and is used, in conjunction with an accompanying chart, to set the oscillator to a specific frequency.

The dip oscillator can be used to determine the resonant frequency of a deenergized radio-frequency tuned circuit. When a resonant circuit is brought close to the oscillating tank circuit, the tuned circuit under measurement absorbs some of the energy from the oscillator tank circuit (refer to Fig. 10-7). Inasmuch as energy is removed from the tank circuit of the grid-dip oscillator, the oscillator feedback into the grid circuit decreases. The grid-current meter reading drops.



(A) Determining resonant frequency of de-energized circuit.

(B) As a signal generator.

(C) As a detector.

Fig. 10-7. Using a grid-dip meter.

When the grid-dip oscillator is set on the exact frequency of the resonant circuit under measurement, the meter will dip to its minimum value. This setting indicates that the grid-dip oscillator is generating a signal of the same frequency as the resonant circuit under check.

It is apparent that the dip oscillator is very helpful in determining the resonant frequency of various types of tuned circuits. Be it a small, weak-signal resonant circuit of a receiver or the larger, higher-powered resonant circuit of the transmitter final amplifier, the resonant frequency can be determined without any signal present. In transmitter circuits the grid-dip oscillator can also be used to track down spurious resonant conditions that cause parasitic oscillations.

A dip oscillator can also be employed as an oscillating detector, similar in action to a heterodyne-frequency meter. In this application, as shown in Fig. 10-7C, the grid-dip oscillator is brought near the source of rf energy. A pair of headphones, inserted into its phone jack, will pick up a beat note when the frequency of the grid-dip oscillator is brought near the signal-source frequency. If this note is zero-beat, the grid-dip oscillator is set to the same frequency as the source of the signal. In this application the grid-dip oscillator has been used for determining the frequency of an unknown signal source.

The oscillating-detector principle of the grid-dip meter can in itself be used to calibrate the meter accurately, as shown in Fig. 10-8.

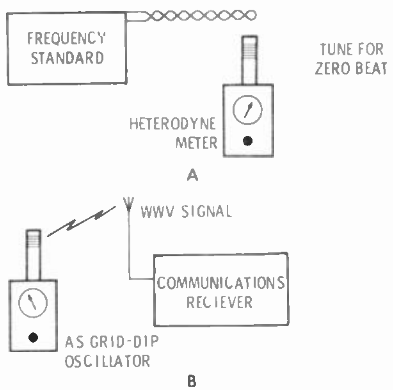


Fig. 10-8. Calibrating a grid-dip meter.

In this check, it is coupled near a crystal-controlled frequency standard. Whenever the grid-dip oscillator is tuned to the crystal frequency or to one of its harmonics, a beat note will be heard. At the beat-note position, the dip calibration should be checked. Some dip oscillators provide a calibration control so the oscillator can be reset if it has drifted.

Still another method of checking the calibration of a grid-dip oscillator is to tune a communications receiver to a WWV frequency (2.5, 5, 10, 15, 20, and 25 MHz), as in Fig. 10-8B. The grid-dip oscillator is then tuned to the same frequency. When a zero beat is heard at the receiver output, the grid-dip oscillator has been set to the same frequency as the incoming WWV signal. The dial calibration can be checked at this point.

When the oscillator is turned off, the dip oscillator becomes an absorption wavemeter, because with no supply potential supplied to it, the tube functions as a diode and the meter becames part of the diode-load circuit. When coupled near and tuned to the frequency of an rf energy source, the meter will display a current increase. The operation is similar to that discussed for absorption wavemeters.

The grid-dip oscillator has many applications in testing and tuning transmitter circuits. The obvious applications are for tuning tank circuits and checking the output. Other applications include use as an rf indicator in neutralizing various stages of the transmitter, or in tracking down troublesome parasitic oscillations. It is also helpful for peaking various types of resonant traps used in transmitters to prevent parasitic oscillations, or for preventing the transfer of strong harmonic components from stage to stage and from the transmitter to the antenna system. Finally, the grid-dip oscillator is handy for checking antenna performance. It can be used in bringing the antenna system to resonance and in minimizing standing waves on the transmission line.

POWER OUTPUT AND ANTENNA METERS

To minimize interference and to be fair to all users, transmitters must adhere to certain performance standards, in the form of specific FCC technical rules and regulations. These standards cover frequency of operation, frequency stability, modulation level, and power output. Suitable test instruments are available for testing these transmitter characteristics.

An rf power-output meter is helpful. It permits one to determine if a given transmitter does function efficiently, and helps him tune up the transmitter. One with a dummy load permits tune up without putting a signal on the air.

Power output can be measured by attaching a dummy load to the transmitter output. The dummy load should have the same impedance as the antenna system into which the transmitter normally

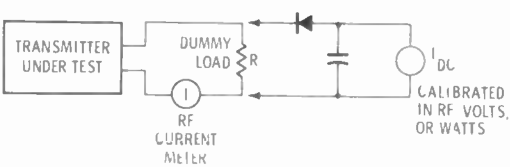


Fig. 10-9. Rf power measurement.

works. An rf ammeter (Fig. 10-9) can be inserted into the dummy-load circuit. Power output will be:

$$P = I^2R$$

where,

P equals power output,
I equals rf current,
R equals resistance of the dummy load.

A calibrated rf voltmeter can also be used to measure the rf voltage across the dummy load. In this case the rf power output is:

$$P = \frac{E^2}{R}$$

An rf voltmeter is generally a crystal diode rectifier and a sensitive dc meter calibrated to measure rf voltage. One can build up a matched dummy load by using paralleled resistors, a coaxial fitting, and a mount, as shown in Fig. 10-10. The 50-ohm dummy can be

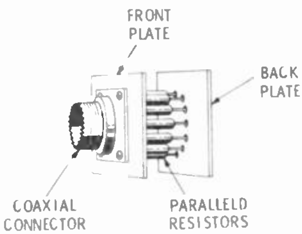


Fig. 10-10. Home constructed dummy load.

plugged into the antenna-output fitting of the transmitter. Resistors should have proper ratings for dissipating the output power of the transmitter.

A problem in making rf power measurements is that the circuit operating conditions change when the test instruments are inserted across the transmitter output or into the transmission-line path between the transmitter and antenna system. Insertion-type or in-line

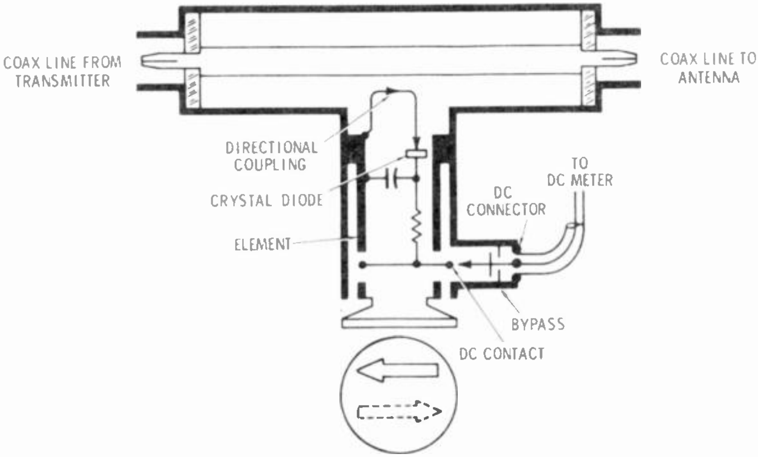


Fig. 10-11. Basic in-line wattmeter.

rf wattmeters provide the answer to this problem. As shown in Fig. 10-11, it is inserted between the transmitter output and the coaxial transmission line going to the antenna system.

The instrument responds to wave direction and is capable of measuring the power in the forward wave traveling between transmitter and antenna, as well as the power contained in the reflected wave traveling back toward the transmitter because of a mismatch in the antenna system. It is apparent that the rf wattmeter is helpful not only in making power measurements, but also in tuning the transmission line and antenna system for a minimum standing-wave ratio to ensure the most efficient transfer of power from transmitter to antenna.

The rf energy, passing between transmitter and antenna, is sampled by a pickup element and applied to a crystal diode and rectifier. A rectified and filtered dc is then linked by a meter cable to a sensitive current meter. A complete coupling circuit consists of the pickup loop, crystal diode, and associated components.

The pickup element of the rf wattmeter functions as a directional coupler. As shown in Fig. 10-12, the rf energy is conveyed to the

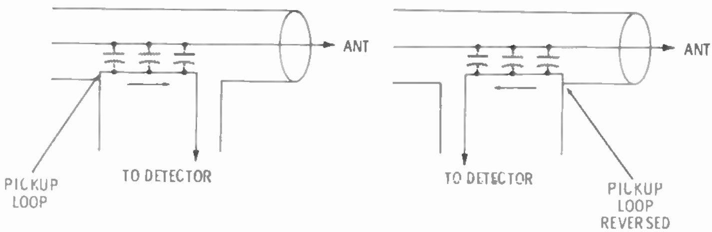


Fig. 10-12. Principle of a reflectometer.

crystal diode over two paths. One path is by way of capacitive coupling from the center conductor, which is part of the coaxial feedthrough arrangement. Radio-frequency energy is also inductively coupled into the pickup probe. By careful design, the two components are of equal amplitude and are either additive or subtractive.

The phase of the inductive component is a function of the direction of wave travel. In the arrangement of Fig. 10-12A, the energy picked up from the wave traveling between transmitter and antenna adds to the capacitive component and produces a reading on the calibrated meter. Any reflected energy returning along the transmission line induces a voltage out of phase with the direct-coupled component from the reflected wave. Consequently, the influence of the reflected wave is canceled and its strength is not recorded.

To read the reflected power, the coupling element is inserted in the opposite direction, as shown in Fig. 10-12B. The reflected power induces into the coupling circuit a voltage in phase with the capacitively coupled component. Hence, the meter will read the reflected power. Insofar as the direct wave moving from the transmitter to antenna is concerned, the coupling loop is so connected that the induced component and the capacitive component are equal and opposite in polarity. The net voltage is zero; therefore, the direct power is not recorded on the meter.

This type of power meter, usually referred to as a reflectometer, has the ability to measure direct and reflected power. The power delivered to the antenna or other load can be calculated by subtracting the reflected power from the direct power.

A standing wave ratio (SWR) meter is a more simplified reflectometer arrangement as compared to the in-line wattmeter. It provides a relative reading of forward and reflected energies. In a typical arrangement using coaxial line, Fig. 10-13, an inductive-capac-

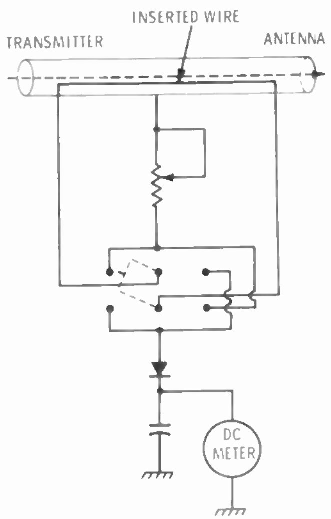


Fig. 10-13. Basic SWR meter.

itive pickup loop is in the form of enameled wire that is fed in and out of the coaxial braid. A switch determines whether the diode current is the result of the forward or direct rf voltage carried by the short section of coaxial line.

A sensitivity control is included to adjust for the absolute level of the power being carried by the line. In normal operation, with the instrument measuring forward voltage the sensitivity control is set for maximum meter deflection. When the switch is set to the reflected voltage position the meter reading will drop to near zero when the standing-wave ratio on the line is low.

Such a meter inserted at the antenna, or at a suitable position, along the transmission line, provides a good measure of the standing wave on the line and the effectiveness with which the transmitter power is transferred to the antenna.

The SWR meter and/or reflectometer arrangement are used in checking out and monitoring antenna systems. Resonant antenna cuts can be made with the proper insertion of an SWR meter designed for the specific impedance of the transmission line. (For the usual SWR meter, optimum performance is obtained with 50- or 70-ohm coaxial lines.)

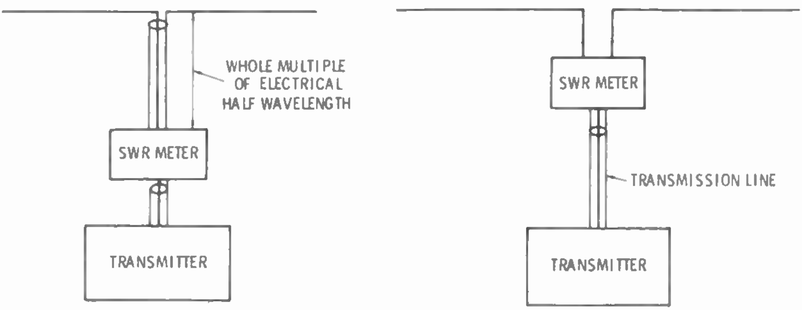


Fig. 10-14. Measurement of SWR and antenna resonance.

Two preferred arrangements are shown in Fig. 10-14. True SWR measurements can be made by inserting the meter right at the antenna. Usually this is not a convenient arrangement. An alternative is to insert the meter one electrical half-wavelength away from the antenna terminals or at some part of the line that is a whole multiple of an electrical half wavelength. The latter plan permits the SWR meter to be located near the transmitter. However, the very best accuracy in terms of the SWR reading and in determining the resonant length of the antenna is feasible only when the exact length of line between the antenna and meter is a whole multiple of electrical half wavelengths. Under this condition the antenna terminal conditions are reflected to the meter and the reactive effects of the transmission line are reduced. The equation for determining the physical length of an electrical half-wavelength line for a given frequency is as follows:

$$\text{Line length} = VF \times \frac{492}{f_{\text{MHz}}} \times \text{whole multiple of } \lambda/2$$

The SWR measurement technique requires the use of a signal source (transmitter operated at low power level or a signal generator with an output capable of supplying adequate signal level to the SWR device). Because of transmitter designs, it is sometimes necessary to operate the transmitter at normal output power level, so that its operating conditions are favorable for matching into 50 ohms.

The usual procedure for operating your SWR meter is employed. In most cases the antenna will be cut long and to a resonant frequency lower than that which is desired. Therefore if you tune your transmitter to the desired frequency and make an SWR measurement it will be higher than that which can be ultimately obtained. As you tune the transmitter lower in frequency the SWR reading drops. The actual minimum may be found considerably lower in frequency than desired.

The antenna may now be trimmed as you watch the SWR minimum move up toward the desired operating frequency. The resonant-frequency indication and the SWR readings using this technique are reasonably accurate, and are more indicative of operating conditions than is indicated by random insertion of an SWR meter into a transmission line.

MODULATION METERS

Today's radio services employ various modulation modes. The measurement technique used depends on the type of modulation.

A-M Modulation Meter

The basic plan of a simple but effective modulation meter is given in Fig. 10-15. The meter assumes that the transmitter is to be terminated in 50-ohms although other impedance values can be used by

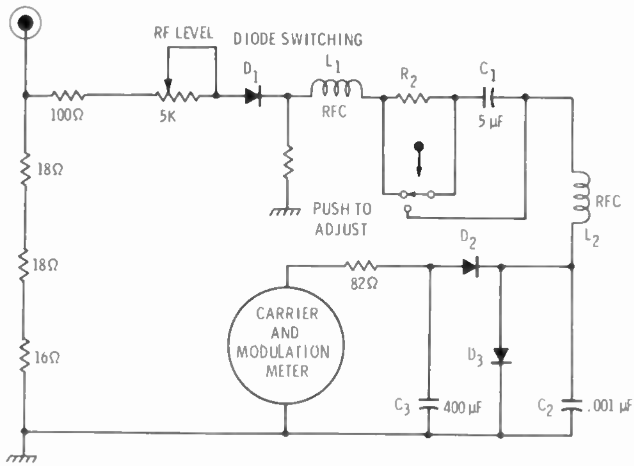


Fig. 10-15. Modulation-measurement portion of tester.

changing the ohmic value of the terminating resistor. The signal applied to the modulation meter is fed through a series resistor and level control to an a-m detector diode (D_1).

The dc component of carrier current is supplied through a radio-frequency choke and switch to the metering circuit. The level control can then be adjusted until the meter is deflected full scale or to a specified preset level. When the switch is set to the *read* position, the dc path to the modulation meter is switched out. The demodulated audio is now transferred to the metering circuit by way of capacitor C_1. The negative alternation of the demodulated wave causes diode D_2 conduction and the resultant peak diode current is recorded on the meter. The time constant is short and the meter responds to the peak of the modulation. Capacitor C_3 does provide some meter damping. This results in a slower drop off of the meter needle, preventing erratic meter movement with modulation.

A-M Oscilloscope Observations

An oscilloscope is excellent for checking a-m and sideband modulation characteristics.

There are two basic types of a-m displays—waveform envelope and trapezoidal. In the envelope type of display, some of the modulated radio-frequency energy is removed from the transmitter and supplied directly to the vertical-deflection plates of the oscilloscope.

The horizontal section of the oscilloscope operates in a normal manner, using the time-base oscillator and horizontal amplifier.

The conventional service-type oscilloscope can be used for checking amplitude modulation. Most oscilloscopes have direct-access terminals that can be used to supply signal to the vertical-deflection plates. Quite often this is in the form of a back plate, which must be removed whenever direct application of signal is to be made. Inasmuch as a high dc voltage is supplied to the deflection plates from the oscilloscope power supply, resistor-capacitor coupling combinations are included between the back plate terminals and the vertical deflection plates of the oscilloscope. Check the schematic of the oscilloscope to make certain that this is so. If not, such components should be connected externally. Complete rf modulation scopes are available such as Heathkit SB-610.

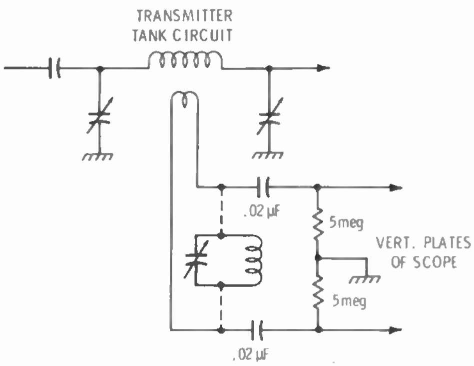


Fig. 10-16. Rf takeoff method for oscilloscope modulation display.

As shown in Fig. 10-16 a small pickup loop, placed near the plate tank circuit of the modulated stage, can withdraw enough energy for suitable vertical deflection on the scope screen. For a low-power transmitter, a resonant circuit (shown in dotted lines) can be inserted to raise the rf voltage.

Some high-power transmitters include modulation-monitor test terminals; it is only necessary to run a connecting line between the terminals and oscilloscope. Other transmitters are so well shielded that it is sometimes difficult to make a convenient internal pickup of the modulated rf signal. Often enough energy can be picked up by bringing a loop near the transmission line or making a capacitive connection to the line with a small alligator clip. Some of the dummy loads used for transmitter checking also provide a convenient means of taking off rf signal for an oscilloscope.

A simple resonant circuit in association with a small pickup antenna positioned near the transmitter, dummy load, or transmitter antenna usually develops enough radio-frequency energy for driving the vertical-deflection plates. An effective arrangement for a low-power transmitter is shown in Fig. 10-17. The constants of the reso-

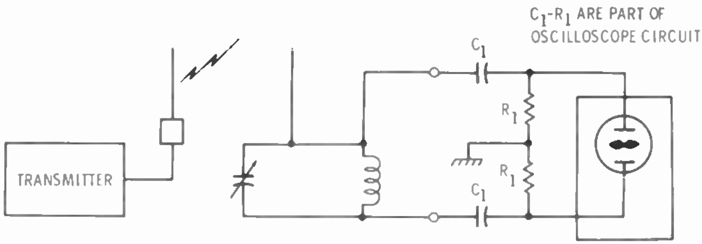


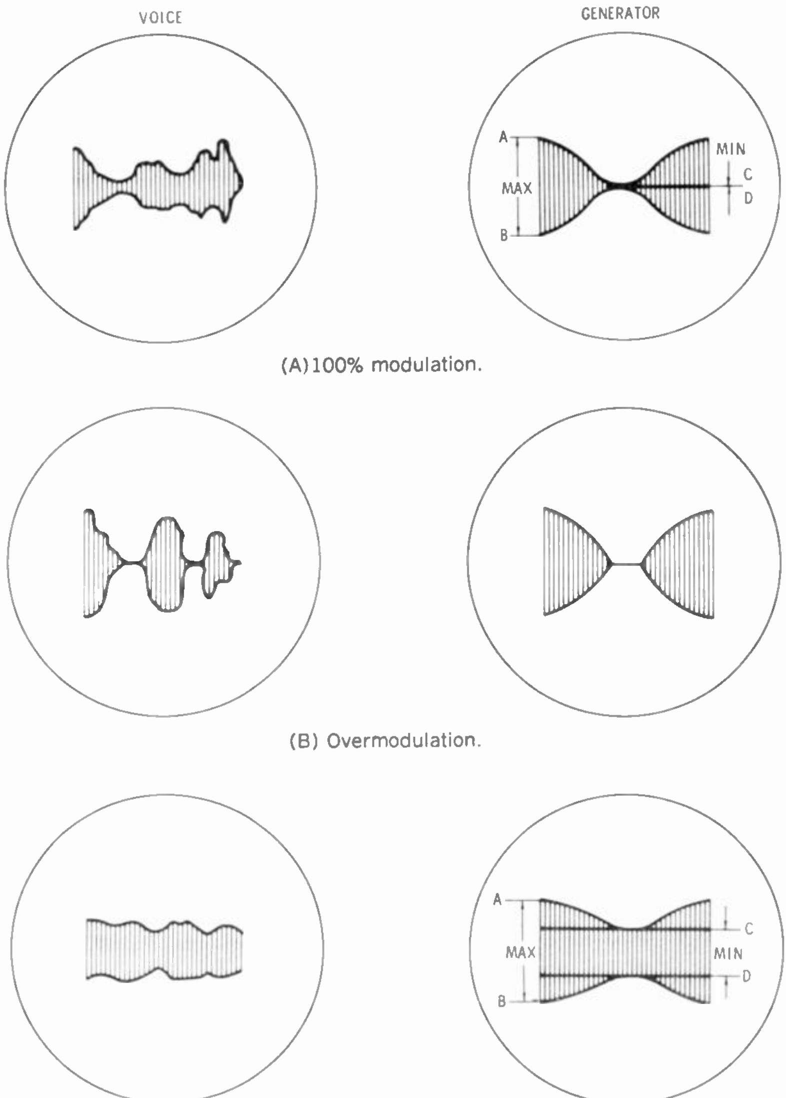
Fig. 10-17. Direct pickup of signal.

nant circuit are chosen according to transmit frequencies. Enough energy from a low-power transmitter can be picked up when the two antennas are separated by several feet. The resonant circuit can be mounted on the rear, near the back plate, of the oscilloscope. A short length of stiff wire serves as the antenna.

A pickup of this type permits the transmitter to be operated in a normal fashion without placing any significant load on it. Nor is there any need for gaining access to any of the internal circuits of the transmitter. Modulation percentage, linearity, and the operation of modulation limiter circuits can be observed.

The depth of the modulation can be observed by speaking into the microphone in the normal fashion. When the modulation is normal, the waveform should appear as shown in Fig. 10-18A. There should be no overmodulation (Fig. 10-18B). However, the average modulation should be high, not weak as in Fig. 10-18C.

More precise modulation measurements can be made by supplying signal from an audio oscillator or generator to the microphone input. The actual voltage depends on the transmitter design and the type of microphone. Try always to set the input-signal level so that it is comparable to the input-voltage requirement of the transmitter audio system.



(A) 100% modulation.

(B) Overmodulation.

(C) Undermodulation.

Fig. 10-18. Speech and sine-wave envelope displays.

A 1000-hertz sine-wave test signal is common for most transmitter modulation checks. The horizontal-sweep frequency of the oscilloscope should be set to display a modulation envelope composed of two or three sine-wave cycles. Typical patterns are shown in Fig. 10-18. When the amplitude of the audio signal is varied over a voltage range comparable to the output limits of the microphone, there should be a high average modulation of the transmitter but no overmodulation. In fact it should be possible to increase the input-signal level somewhat above the maximum output made available by the microphone without signs of overmodulation.

The actual modulation percentage at specific levels can be calculated using a simple formula and the vertical-scale division of the oscilloscope:

$$\% \text{ mod.} = \frac{AB - CD}{AB + CD} \times 100$$

where,

A, *B*, *C*, and *D* are the number of scale divisions measured at the points indicated in Fig. 10-18C.

Distortion, as a result of nonlinearity or introduction of harmonics, is also indicated on the modulation envelope by the flattening of

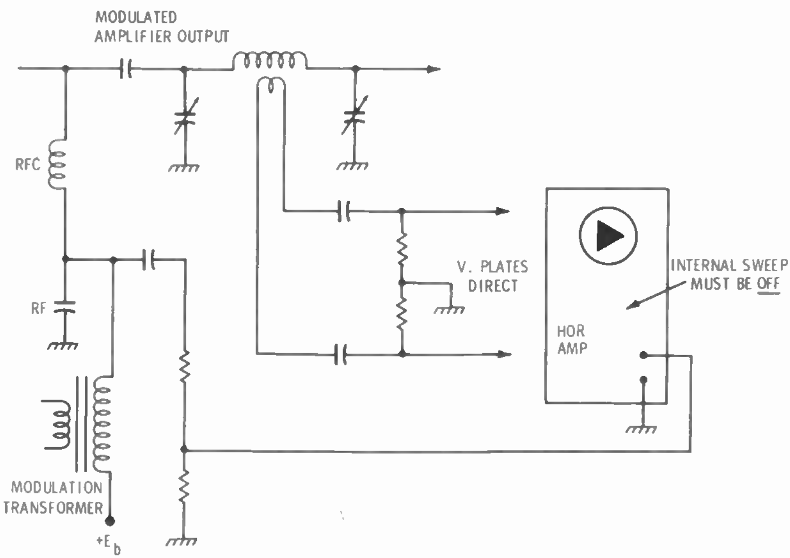
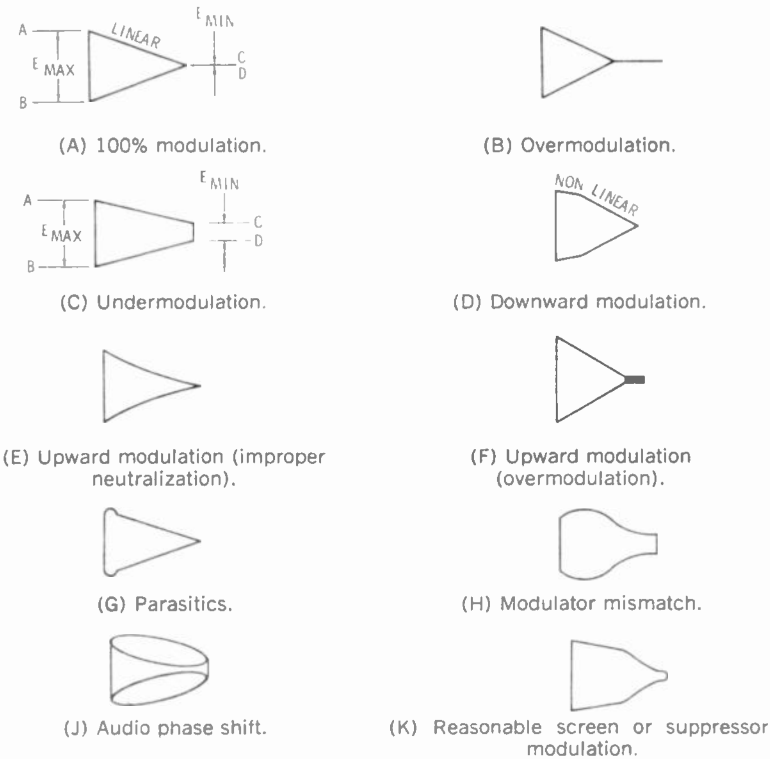


Fig. 10-19. Interconnection plan to obtain a trapezoidal display.

one or both modulating wave peaks or other distortion of the modulation envelope. The trapezoidal modulation display provides an even more critical evaluation of the linearity of the modulation.

In a trapezoidal modulation display (Fig. 10-19), the rf modulation envelope is again supplied to the vertical-deflection plates. The modulating wave itself is supplied to the horizontal-deflection plates; the internal time-base oscillator of the oscilloscope is turned off. A component of the modulating wave is derived from the modulator output or elsewhere in the modulator stage. Again the service-type oscilloscope can be used for the trapezoidal amplitude modulation display.

As in many types of oscilloscopic linearity checks, a straight line is formed whenever the input and output signals are linear. Inasmuch as the modulation envelope consists of both a negative and positive



(A) 100% modulation.
(B) Overmodulation.
(C) Undermodulation.
(D) Downward modulation.
(E) Upward modulation (improper neutralization).
(F) Upward modulation (overmodulation).
(G) Parasitics.
(H) Modulator mismatch.
(J) Audio phase shift.
(K) Reasonable screen or suppressor modulation.

Fig. 10-20. Trapezoidal modulation patterns.

component, the two traces form a trapezoid. Furthermore, if the modulation envelope swings to twice the unmodulated carrier level and to zero (as in 100% modulation) the triangle shown in Fig. 10-19 is formed. Additional trapezoidal patterns are shown in Fig. 10-20.

If the transmitter is modulated with a sine wave as described previously, the actual modulation percentage can be calculated:

$$\% \text{ mod.} = \frac{AB - CD}{AB + CD} \times 100$$

If the sides of the trapezoid are other than a straight line, this indicates poor modulation linearity. Typical fault patterns, are also given in Fig. 10-20. The pattern in Fig. 10-20J, however, is an interconnection problem. It indicates only that the polarity of the modulating wave should be reversed so that it is in phase with the modulation on the rf envelope. Correction can be made by removing the modulating wave at some point in the modulator section where the phase of the modulating wave is of opposite polarity.

Again as in the case of the simple envelope display, upward or downward modulation characteristics can be uncovered. The presence of parasitics or improper neutralization can also be disclosed. Notice that linearity defects are more obvious because of the straight-line linearity characteristic of the trapezoidal pattern.

Sideband Modulation—The same technique can be used to obtain oscilloscopic patterns for double sideband and single sideband transmission. The same manner of interconnection is employed. The appearance of the patterns differ, but in general, are evaluated in the same manner. Typical double-sideband and single-sideband patterns were given previously in Fig. 5-4.

F-M Modulation—Frequency deviation can be measured with a meter or by spectrum analysis. The usual f-m deviation meter is a high quality f-m receiver that employs a calibrated f-m discriminator and output circuit that responds to the peaks of the modulating wave (Fig. 10-21). A balanced discriminator is used that can be set precisely to deliver *zero* output at the center frequency. The demodulated f-m signal is applied to a crystal diode circuit that responds to the peak of the demodulated wave. (This peak of course corresponds to maximum deviation.) A time constant in the output of the detector holds this peak charge constant and applies it through a suitable calibration system to a dc amplifier and associated metering cir-

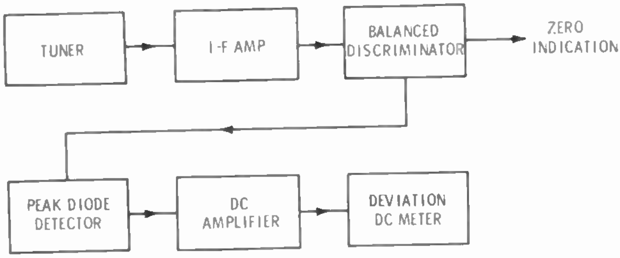


Fig. 10-21. Functional plan of f-m modulation meter.

cuit. A switching arrangement is usually included to permit measurement of both above-carrier frequency and below-carrier frequency maximum deviation.

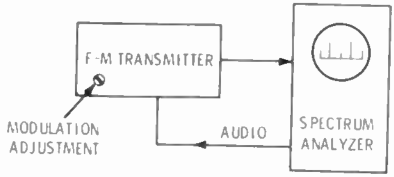
Bessel-Function Method—The resultant f-m wave is composed of a center frequency and a number of sideband pairs. The number of sideband pairs and their relative magnitude among center frequency and sidebands are a function of the modulation index. At certain index values the carrier level itself reduces to near zero. The first five such null points are:

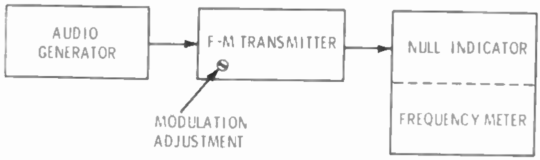
1. 2.405
2. 5.52
3. 8.654
4. 11.792
5. 14.931

These test points can be used to advantage in making frequency deviation checks. To make use of the check points it is necessary that the emission bandwidth be displayed on the screen of a spectrum analyzer, or that a frequency meter or receiver be used that contains sharply tuned resonant circuits that are able to delineate the carrier from the first pair of sidebands.

Two basic systems can be used to obtain a carrier null as shown in Fig. 10-22. A spectrum analyzer can be used. This type of oscilloscope display shows the center frequency and individual sidebands at their proper relative magnitudes. The display for the first two carrier-null conditions would appear on the oscilloscope screen exactly as depicted in Fig. 10-23. An accurate and highly selective frequency meter can also be used to observe a center-frequency null (Fig. 10-22B). The frequency meter *must* have sharply tuned and high-Q resonant circuits to be able to distinguish among carrier and

(A) Connection for a spectrum analyzer.





(B) Connection for a frequency meter.

Fig. 10-22. Two carrier-null methods of adjusting frequency modulation.

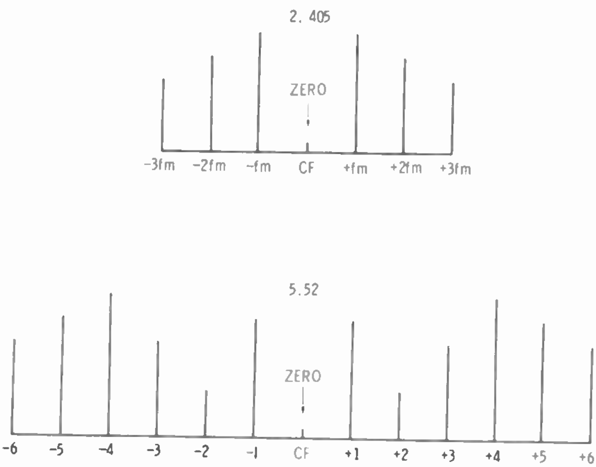


Fig. 10-23. Approximate levels of sidebands and carrier/zero for first two carrier-null points.

sidebands. In this way it is possible to bring the center frequency to a null without being impeded by the first pair of sidebands.

In checking out the performance of an f-m two-way radio unit that is assigned a maximum permissible deviation of ±5 kilohertz, the first and second carrier nulls can be used. What deviation is obtained, for instance, when the modulation index is 5.520 and the modulating frequency is 880 cycles? A simple calculation shows:

$$\text{Deviation} = \text{mod. index} \times \text{mod. frequency}$$
$$= 5.52 \times 880$$
$$\text{Deviation} = \pm 4.97 \text{ kilohertz}$$

Note that the deviation obtained is just short of the ±5-kilohertz limit and can serve as a good adjustment indication. The 880-hertz audio tone can be obtained with great precision, if desired, because it is the second harmonic of the 440-hertz modulation of WWV. A twice-frequency lissajous pattern on the oscilloscope, as in Fig. 10-24,

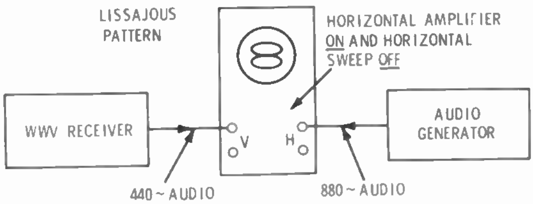


Fig. 10-24. Using WWV signal to set audio generator on 880 hertz.

would indicate that the audio generator applied to the horizontal amplifier is set to 880 hertz with the 440-hertz modulation recovered from WWV applied to the vertical amplifier.

The first carrier null point can also be used to advantage for indicating when the magnitude of the modulating wave applied to a two-way radio is high enough to cause a deviation greater than ±2 kilohertz. Mathematically this is:

$$\text{Deviation} = 2.405 \times 880$$
$$\text{Deviation} = \pm 2.12 \text{ kilohertz}$$

If a frequency meter instead of a spectrum analyzer must be used to establish the carrier null, its selectivity will have to be such that it can null the carrier without interference from the adjacent sidebands spaced only 880 cycles on each side of the center frequency. A greater separation between the center frequency and first pair of sidebands can be obtained by increasing the modulating frequency. A 2000-hertz modulating tone (4th harmonic WWV 500-hertz tone) and the first carrier null position represents a good combination. In this case it figures as follows:

$$\text{Deviation} = 2.405 \times 200$$
$$\text{Deviation} = \pm 4.81 \text{ kilohertz}$$

It is important that you realize the modulating frequency selected must fall within the linear bandpass of the transmission system. Using modulating frequencies above or below the linear bandpass will result in improper adjustment of the modulating level for the transmitter.

An appropriate modulating frequency can also be selected in conjunction with a carrier null point to measure the maximum permissible deviation of ±15 kilohertz. Since the audio modulating frequency band is quite narrow for two-way radio, a higher order of carrier null is used. For example with a modulating frequency of 1200 hertz and a modulation index of 11.79 the deviation is in excess of 14 kilohertz:

$$\text{Deviation} = 11.792 \times 1200$$
$$\text{Deviation} = \pm 14.15 \text{ kilohertz}$$

A precise 1200-hertz tone can be obtained because the audio generator can be set to the 1200-hertz second harmonic of the 600-hertz WWV modulation. When using a higher order of carrier null, one must, of course, be careful to locate the correct null. For example, by increasing the magnitude of the 1200-hertz audio modulating wave from zero there will be carrier nulls at modulation indexes of 2.405, 5.52, 8.65 and finally the desired one of 11.79.

PROJECT 38: DIGITAL FREQUENCY CALIBRATOR

Three low-cost digital integrated circuits can be wired into a versatile frequency calibrator with calibration points separated by as much as 100 kHz and as little as 1000 hertz. The calibrator uses an MRTL dual two-input NOR gate and two 7490N decade counters, Fig. 10-25.

The entire unit can be mounted on one side of a 4½" by 5½" vector board, Fig. 10-26. A vhf calibrator is to be mounted on the right side of the board as described in Project 39.

The HEP-580 internal circuit is given in Fig. 10-27 along with the logic wiring diagram for the 100-kHz crystal oscillator. Note that only five external components are required—two 100K resistors, crystal socket, calibration trimmer and 0.02-μF feedback capacitor. Actually, the circuit is a crystal-controlled multivibrator that produces a square-wave pulse output that is rich in harmonics.

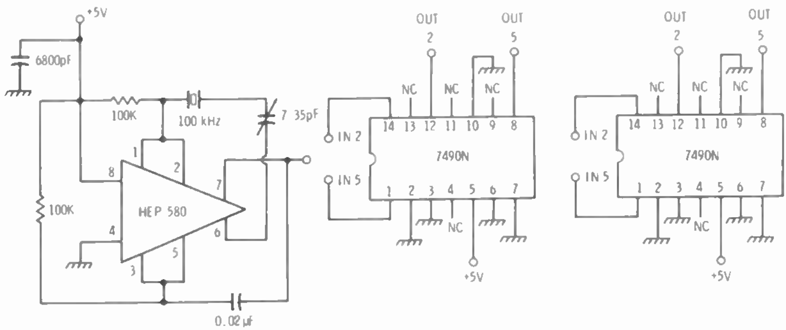


Fig. 10-25. Digital IC calibrator. Resistor and capacitor values are shown on schematic. Unit is built on a 4½" x 5½" VECTOR board.

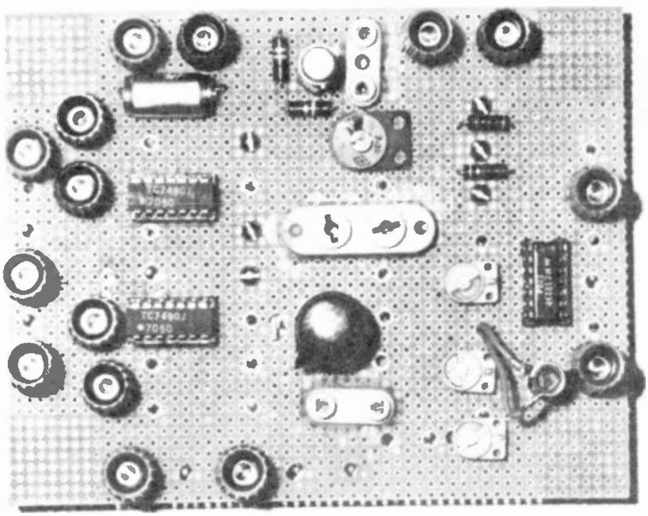


Fig. 10-26. Hf-vhf Calibrator

The 7490 decade counter consists of a 2-to-1 and 5-to-1 counters mounted in the same case. The two counters are connected in cascade to obtain an overall count of 10-to-1. The wiring of the 7490 IC for each of the three applications is given in Fig. 10-28.

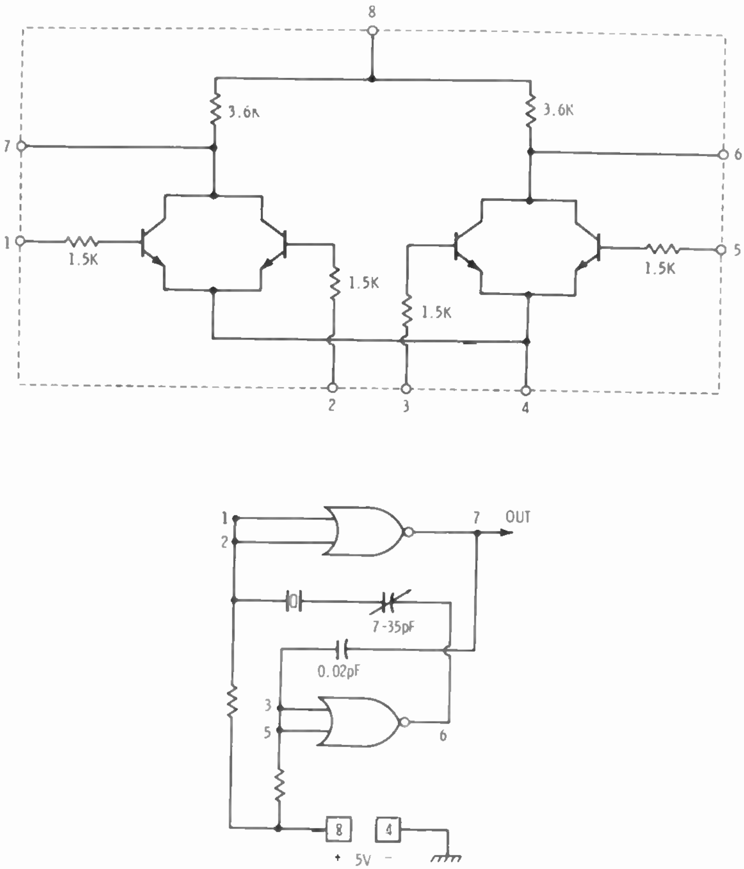


Fig. 10-27. NOR gate crystal MV circuit using HEP-580

For operation as a simple two-to-one counter the input is applied to pin 14 and removed at pin 12. In the 5-to-1 mode the input signal is applied to pin 1 and removed at pin 8. For all three cases the supply voltage is connected to pin 5 and ground to pin 10. The four reset inputs are also connected to ground.

When divide-by-ten operation is desired, the output of the 5-to-1 counter at pin 8 is applied to the input of the 2-to-1 counter at pin 14. The divide-by-ten output is then taken off at pin 12. Input signal is applied to pin 1, the input of the 5-to-1 counter.

If the output of the oscillator is terminated at a binding post and separate input and output binding posts are used for the two decade

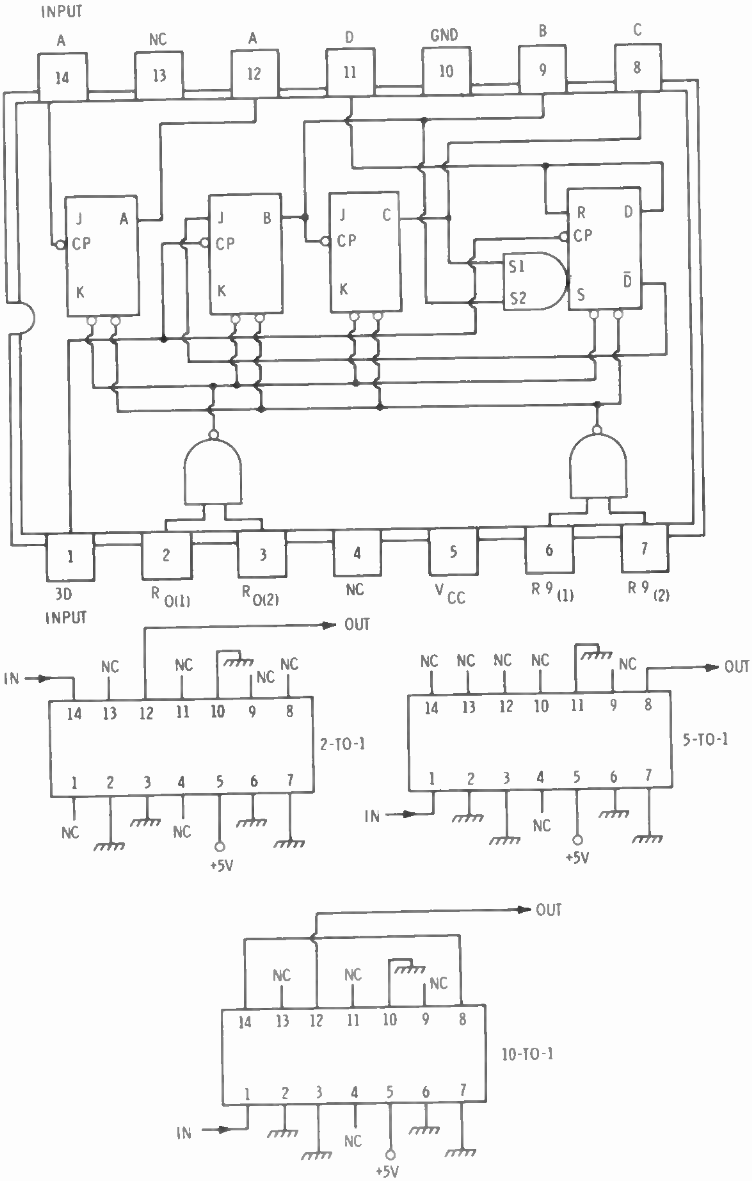


Fig. 10-28. Decade IC and count possibilities.

counters many versatile count terminations can be set up using appropriate jumpers, Fig. 10-25. The example of Fig. 10-29A shows

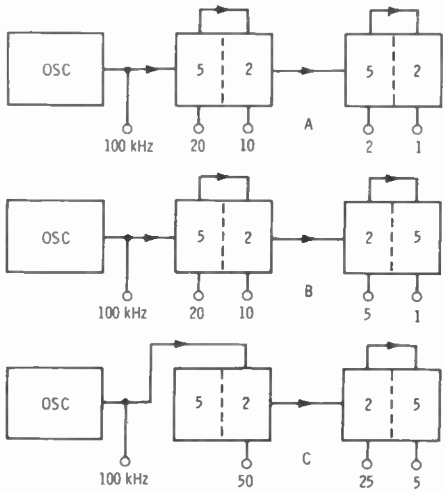


Fig. 10-29. Several count possibilities.

how outputs of 100 kHz, 20 kHz, 10 kHz, 2000 Hz and 1000 Hz are made available.

If you prefer one 5000-hertz output instead of 2000 hertz, one need only change around the last decade counter. In this case the output of the first decade counter is applied to the 2-to-1 counter input of the last decade, pin 14. This makes a 5000-hertz output available at pin 12. Pin 12 can be jumped to pin 1, the input of the 5-to-1 counter of the last 7490 decade, making the 1000-hertz output available at pin 8.

Of course, many combinations can be set. Any combination you choose can be wired permanently, in which case only output binding posts are required. Other combinations can be established by using only three of the counters instead of the complete set of four. For example, using two 2-to-1 and one 5-to-1, outputs of 100 kHz, 50 kHz, 25 kHz and 5 kHz are possible.

Operation

The steep-sided waveforms generated by digital IC's have a high harmonic content. In fact, useable markers are obtained on frequencies as high as the 2-meter band. In fact, if you are using a 2-meter receiving system that is finely calibrated it is possible to delineate 1000-hertz markers over the entire 2-meter band. Calibrator output can be supplied to the receiver input with inductive coupling obtained by wrapping several turns of wire around receiver transmission line or by direct capacitive coupling to the antenna termination by way of a low-value isolating capacitor.

The accuracy of the calibration also depends on the precise setting of the 100-kHz crystal oscillator. With the circuit completely wired and in operation, adjust the calibrating capacitor for zero beat by receiving WWV on the highest possible reception frequency for WWV in your area.

PROJECT 39: DIGITAL IC HF-VHF OSCILLATOR/CALIBRATOR

High-frequency digital integrated circuits perform well as high-frequency oscillators and the high harmonic content of their outputs permit them to serve well as calibrators up into the vhf-uhf frequency ranges. Two NOR gates suitably interconnected provide a calibration signal of low impedance, high stability, and strong harmonics. A good performing device is the MECL MC 1023P which is a dual NOR-gate type, which will operate as an oscillator up into the 100 MHz range using overtone crystals.

For medium- to high-frequency operation, with fundamental crystals, no resonant circuit is required as shown in the simple circuit of Fig. 10-30A. A small trimmer capacitor (C_1) permits one to adjust the oscillator frequency precisely. The only additional components consist of an isolating resistor and bias components. For operation above 20 MHz and up to 100 MHz a rather broadly tuned resonant circuit is employed as shown in Fig. 10-30B.

A load capacitor (C_2) establishes the proper phase relationship for feedback. The two capacitors must be set rather carefully to obtain easy starting of the oscillator. Once set, however, the oscillator will take off over a rather wide frequency range. One setting permitted operation between 50 and 80 megahertz.

The second gate of the dual pair functions as a phase splitter and isolates the output from the oscillator section. Two square-wave outputs of opposite polarity are available.

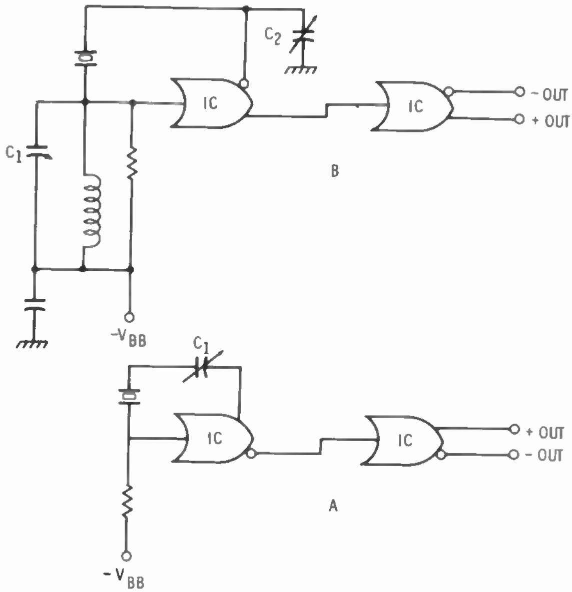


Fig. 10-30. Vhf and hf oscillator circuit.

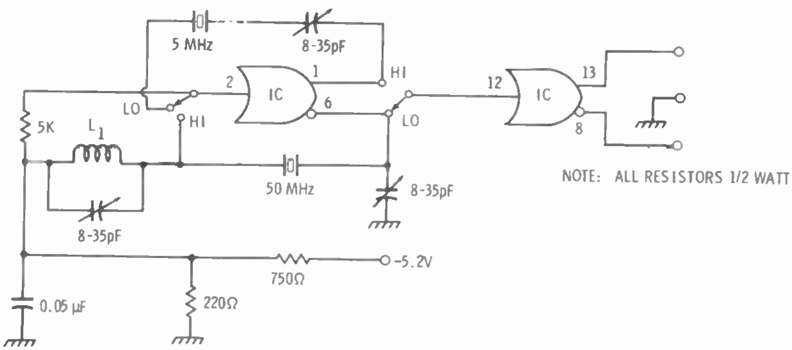


Fig. 10-31. Vhf-hf switched oscillator. Component values are shown on the schematic. IC is MC 1023. Unit is built on a 4½" x 5½" VECTOR board. L₁ is J. W. Miller 20A337RBI slug-tuned inductor.

The circuit of Fig. 10-31 shows how a simple dpdt switch can be incorporated to change over between vhf harmonic-mode operation and lower-frequency fundamental-mode operation. This is an ideal facility for calibration work in the vhf-uhf range.

Operation

The wiring diagram, on the basis of a dual in-line socket, is given in Fig. 10-32. Note that the unused input gates are tied to-

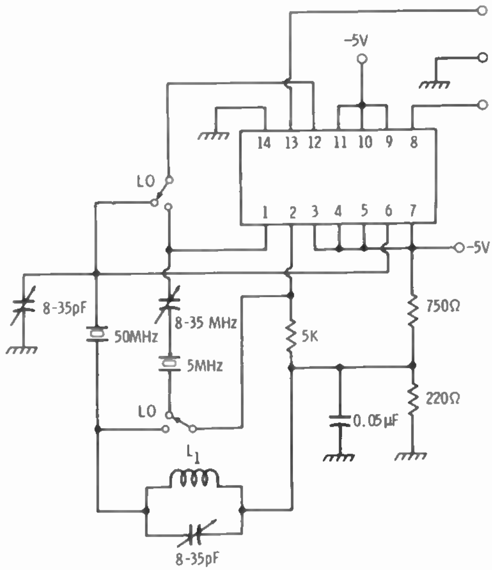


Fig. 10-32. Oscillator pin-out schematic.

gether and connected to the −5 volt point. The two crystals can be inserted into their sockets and kept there without any adverse influence on the two-frequency capability. The crystal oscillator is shown at the right of the calibrator board of Fig. 10-26.

In our own calibration work we used a 50-MHz overtone crystal and three fundamental crystals of 5, 2, and 1 MHz. Additional overtone crystals were also employed as will be mentioned later.

A 1-MHz and 50-MHz combination performed well on 6 meters. The fifth harmonic of the 1-MHz crystal can be calibrated on the WWV 5-MHz or 10-MHz signal. Likewise the 100-kHz crystal can also be calibrated on either one of these WWV frequencies.

The 1-MHz crystal produces strong harmonics in the 6-meter band. In addition, at weaker level you will also find the 100-kHz calibration points. Harmonics from both of the previous crystals fall at exactly 50 MHz where they can make a check of the calibration of the 50-MHz overtone oscillator circuit. The 50-MHz oscillator, of course, produces a very strong signal component at the low end of the 6-meter band. This same crystal also provides a strong marker on 450 MHz, the high-frequency end of the three-quarter meter band. Also 1-, 2-, or 5-MHz markers can be heard on the three-quarter meter band as well as the 1¼-, 2-, and 6-meter bands.

Much activity seems to be concentrated at the low-frequency ends of the various vhf-uhf bands. Overtone-crystal frequencies can be chosen to obtain very strong marks at each low-frequency edge. The 50-MHz crystal provides this service on the 6-meter band.

A 72.5-MHz crystal provides a strong mark at 145 MHz on the 2-meter band. Here it can be calibrated with a harmonic of the 5-MHz crystal. Of course, a 72-MHz crystal will give you a strong mark at exactly 144-MHz, the low-frequency end of the 2-meter band. The 2-MHz crystal will provide a WWV calibration point at this frequency. The low edge of the 1¼-meter band can be located by using a 55-MHz overtone crystal. At this 220-MHz frequency a WWV calibration point can be inserted with the 5-MHz crystal.

The combination of Projects 38 and 39 provide all-band calibration capability. Strong outputs are an asset and often no direct coupling is needed. Just place the calibrator in the vicinity of the receiver or converter. It can be of particular help when trying to locate rather exact spot frequencies; building and testing direct-conversion receivers, vhf-uhf preamplifiers, and converters; and other types of receivers and tuners.

PROJECT 40: DIRECT-CONVERSION TEST RECEIVER

Direct-conversion receivers have become increasingly popular, especially among homebrewers. Good performance, simplicity, and low cost are the advantages. The technique lends itself to the construction of small receivers that have utilitarian application around the ham station. Such receivers can be tunable or crystal-controlled on any number of fixed test frequencies. Although the receiver of this project is very simple and its application is mainly as a transmitter test unit, it can be pressed into service as a spare receiver because it performs well on both the cw and single-sideband modes for local and more distant contacts. The zero-beat technique can be used to demodulate standard a-m signals. The addition of a diode detector permits more convenient a-m detection.

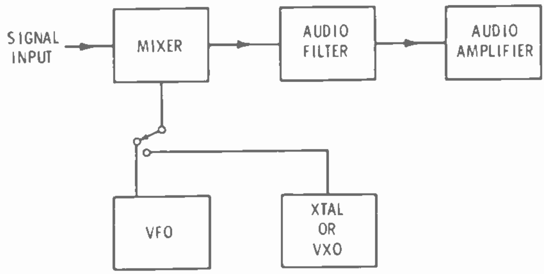


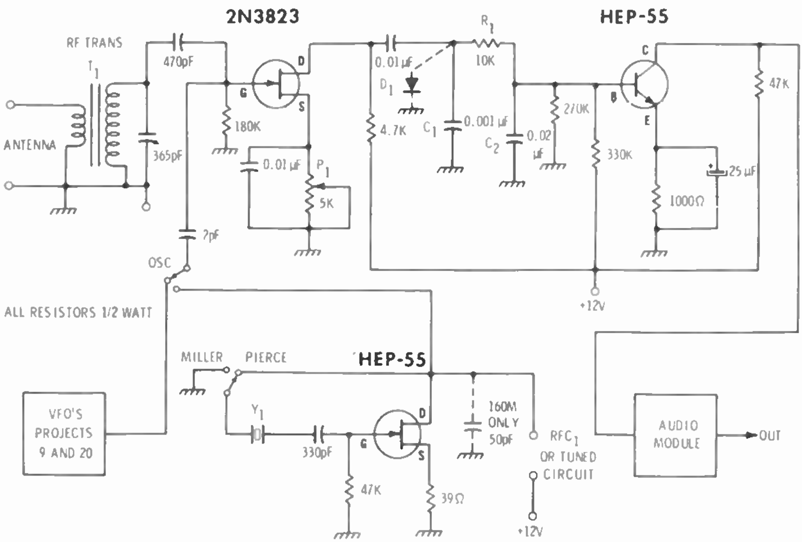
Fig. 10-33. Basic direct-conversion technique.

In the direct-conversion process (Fig. 10-33) the incoming signal is applied to the mixer. It is beat against a locally generated signal supplied by a vfo, crystal oscillator, or vxo set to the same frequency. A direct conversion is made between incoming signal frequency and the demodulated audio frequency. The output of the mixer is tuned to the difference frequency which is in the audio-frequency range. This is handled by an audio filter. Its component values are such that the radio-frequency components are removed. The output of the audio filter is supplied to an audio amplifier and on to a speaker or headset.

The selectivity of the direct-conversion receiver is determined mainly by the audio filter which is designed to have a cutoff frequency just above the high end of the voice frequency range. The filter characteristics can be sharpened further if only cw reception is desired. The sensitivity of the device depends very much on the noise characteristics of the mixer and the input audio amplifier. A low-noise, high-gain audio amplifier is important, because it determines the overall gain of the receiver and the receiver's ability to keep the desired signal above the background noise.

The schematic diagram of the test receiver is given in Fig. 10-34. The mixer stage is a low-noise FET (Siliconix 2N3823) which performs well as a mixer up into the vhf-uhf frequency range. A tuned input transformer is employed. Any number of these can be mounted on the pegboard depending upon the desired frequency ranges. The three rf transformers recommended cover between 1.7 and 36 megahertz. A source potentiometer permits you to set the mixer for optimum mixing depending on the level of the input signal.

The heterodyning oscillator component can be supplied by the vfo's of Projects 9 and 20 which will permit reception on all bands



D_1 1N34A diode
P_1 5K potentiometer
RFC_1 2.5-mH rf choke
T Rf transformer; 1.7-5.5 MHz, (J. W. Miller B-5495), 5.5-18.0 MHz, (J. W. Miller C-5495), 12.0-36.0 MHz, (J. W. Miller D-5495)
1 Audio module
1 Vfo from Projects 9 and 20

Fig. 10-34. Direct-conversion test receiver.

10 through 160 meters. The receiver also includes a crystal-controlled oscillator for fixed-frequency operation of the test receiver. This can be connected as a simple Pierce crystal oscillator using fundamental crystals on the 20-, 40-, 80-, and 160-meter bands. Tuning of the oscillator, in this case, is not required and the drain load can be a 1-mH radio-frequency choke. A 50-pF capacitor should also be connected between drain and ground for 160-meter operation only. For operation on 15, 10, and the vhf bands it is necessary to use overtone crystals. In this case the oscillator can be connected as a Miller circuit and an appropriate resonant circuit must be inserted in the drain circuit.

The low-pass audio filter consists of capacitors C_1 and C_2 plus resistor R_1. These constants will give an adequate audio response, filter out the radio frequencies, and provide an adequate selectivity. A bipolar audio stage follows, developing an output signal of reasonable magnitude and low impedance for supplying signal to the low-impedance input of the bipolar audio module board. This can be any one of the low-cost units available with power output ratings of 1 to 3 watts.

The convenient demodulation of a strong a-m signal can be accomplished by inserting a diode detector (D_1) across the input capacitor (C_1) of the audio filter. The resistors that follow then serve as load resistors across which the demodulated audio is developed. In this manner of operation power must be removed from the heterodyning oscillator. The receiver then consists of an input rf amplifier, diode detector, and audio amplifier. This connection is advantageous when checking out an a-m transmitter.

Operation

Construct the receiver on the pegboard or build it into a small cabinet the same size as used for vfo's of Projects 9 and 20.

Connect the 80-meter input transformer into the circuit. Insert an 80-meter crystal into the heterodyning oscillator circuit set to its Pierce position. Turn on the unit. Tune the input variable capacitor for maximum noise output, indicating the input resonant circuit is tuned to the same frequency as the crystal.

Set the transmitter under test on its "oscillator-tune" position and adjust to the same frequency. You will hear the heterodyne as the transmitter vfo is tuned through the crystal frequency. Load your transmitter into its dummy load and modulate it with cw or sideband. Carefully adjust the transmitter vfo until demodulated output can be heard at the output of the receiver.

If a vfo is being used as the receiver heterodyne oscillator, the transmitter can be preset to a specific frequency and tuned in on the receiver by adjusting the vfo main and bandspread tuning capacitors.

While transmitter supplies power to a dummy load it can be modulated with single-tone or two-tone signal. The direct conversion receiver can then be moved a considerable distance from the transmitter and the quality of the demodulation checked. In this application you are getting the same results as applying power to the transmitting antenna and making a receiver check several miles away.

The receiver can be used in making the same type of check because sensitivity is quite good as you will learn when you tune over an amateur band with an adequate antenna attached. Of course, you cannot tone-modulate your transmitter when it is supplying power to an antenna and you will have to enlist someone else's help when making an on-the-air check.

Turn on an amplitude-modulated transmitter and tune in the signal using the zero-beat technique. Again, you can use a fixed-frequency heterodyne oscillator to tune the transmitter vfo, or use a tunable vfo for the heterodyne oscillator and set the transmitter on a specific frequency.

Turn off the heterodyne oscillator and insert the crystal detector into the circuit. Heterodyning is not necessary and a strong audio component is demodulated for analysis. This manner of operation will also permit you to demodulate an f-m signal by tuning the input resonant circuit until the f-m carrier is centered on the skirt of the response curve of the input transformer. In fact, with the receiver you can check the various multiplier frequencies on up to the transmit frequency using suitable resonant input transformers.

Appendix

ADDRESSES OF SUPPLIERS

HEP Transistors
ALLIED RADIO
100 North Western Avenue
Chicago, IL 60680

Siliconix Transistors
2201 Laurelwood Road
Santa Clara, CA 95054

Sideband Filter
Spectrum International
Box 87
Topsfield, MA 01983

Permacor
9540 Tulley Avenue
Oak Lawn, IL 60453

Lafayette Radio Electronics
111 Jericho Turnpike
Syosset, L.I. New York 11791

Micrometals
72 East Montecito Avenue
Sierra Madre, CA 91024

Ami-Tron Asso.
12033 Otsego Street
North Hollywood, CA 91607

Ten-Tec Inc.
Sevierville, TN 37862

International Crystal Mfg. Co.
10 North Lee
Oklahoma City, OK 73102

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