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TRANSISTOR TRANSMITTERS for the AMATEUR

by Donald L. Stoner, W6TNS

Complete details, including circuit descriptions, construction techniques, parts lists, and operating instructions, for 12 useful transistor devices. Valuable projects for hams, CB'ers, students, and experimenters.

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Preface

Amateur radio has served to introduce thousands of enthusiasts to the field of electronics. For many, amateur radio is the start of a rewarding profession; for others, it is a helpful supplement to their formal education in engineering.

Constructing your own equipment is a satisfying and educational segment of the hobby. Also, you can keep abreast of the latest developments in circuits and components at little expense.

Construction projects for the beginner, as well as for the experienced ham, are described in this book. Included are projects for novice, technician, general-class, and Citizens-band licensees.

For example, the "Novice Powerhouse" in Chapter 5 is an ideal rig for the beginner or the QRP general-class operator. The modulator described in Chapter 6 can be added to the "Novice Powerhouse" for phone operation. The 10-meter rig in Chapter 8 can be used to communicate over distances of thousands of miles under proper conditions.

The latest design and construction techniques are incorporated throughout, including transistor circuitry and circuit-board layouts.

Every piece of equipment has been built and thoroughly tested on the air. I hope that you will enjoy building and operating these projects as much as I did preparing them for this book.

DONALD L. STONER, W6TNS

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Transistor Transmitters Are Different

There is always an element of glamor and just a little extra thrill when operating a short-wave transmitter that employs no vacuum tubes. This is particularly true when you are using it to communicate with someone several hundred or even a thousand miles away.

Before plunging into the construction and operation of these solid-state "peanut whistles," let's review a bit of theory to obtain a clearer understanding of how they work. Even if you are an "old pro" in building rock-crushing tube rigs, transistor transmitters *are* different. Let's see why.

THE HISTORY OF COMMUNICATIONS TRANSISTORS

Although it may not be immediately obvious, much credit should be given to the computer industry for the great progress which has been made in the field of communications transistors. In the mid-1950's there were few manufacturers who had the foresight to see the vast potential of the transistor for eliminating vacuum tubes in communications equipment. They seemed content to fabricate noisy, delicate devices suitable for use in raspy-sounding portable radios.

Early transistor devices proved to be a poor second compared to the vacuum tubes used in communications transmitters. For several years it appeared that high-power, high-frequency transistors were theoretically impossible to fabricate. Fortunately for the communications industry this has not proved to be the case.

The computer industry, however, was enthusiastic about this new electronics infant. They immediately visualized the possibility of eliminating the thousands of heat-generating and inefficient tubes used in a typical computer. Some of the "dreamers" went so far as to predict table-top computers (Fig. 1-1) which could be operated by an average secretary or bookkeeper.

Fig. 1-1. Transistor IBM 1620 computer.

The transistor was able to replace many of the tubes in the slower and less "intelligent" computers, but the designers soon found that the transistor had many shortcomings. When instructed to "turn on," the process was accomplished rather slowly. Once the delay in following instructions was overcome, it was reluctant to obey the "turn-off" signal and continued to plod along for several microseconds. The time characteristics are illustrated in Fig. 1-2.

The designers could not tolerate such sluggish performance in equipment which had to make hundreds of thousands of "decisions" every second. Transistor manufacturers were asked to make faster and faster switching devices in order to raise the IQ of the next model computer. Devices were also needed

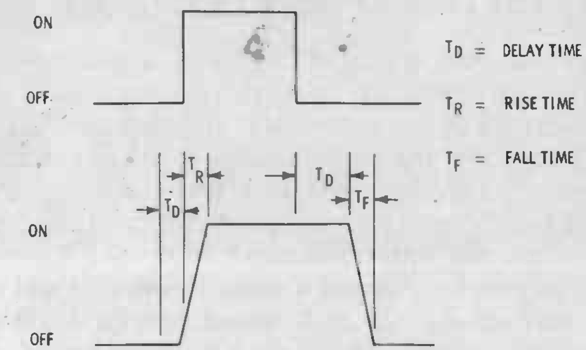


Fig. 1-2. Rise, fall, and delay time characteristics.

which would deliver watts of power, rather than milliwatts or microwatts, in order to drive memory cores in the complex electronic cerebral cortex of the computer.

This constant pressure on transistor manufacturers for faster and more powerful devices brought about a revolution in the semiconductor industry. Vast sums were invested in research programs to perfect new ways of making better transistors for less money. The infant industry progressed from the humble *junction* transistor, through the *drift field*, the *diffused*, and the *mesa*, to the present-day *planar* transistor, a miracle of semiconductor technology.

Modern transistors are able to operate at the speed of light. A planar computer transistor is capable of switching from the *off* state to the *on* state in less time than it takes the light reflected from this page to reach your eyes (remember that light travels at the speed of 186,000 miles per second). Today it is possible to purchase devices which deliver 20 watts or more of radio-frequency energy while switching on and off 50 million times (50 megacycles) each second. Although these transistors are very expensive, they are available. Further pressures on the semiconductor industry, and the ever increasing competition within it, will result in greatly reduced prices of communications transistors. If the present rate of progress continues, it will soon be possible to buy a device which will perform the same as a 6146 tube for less than $10.

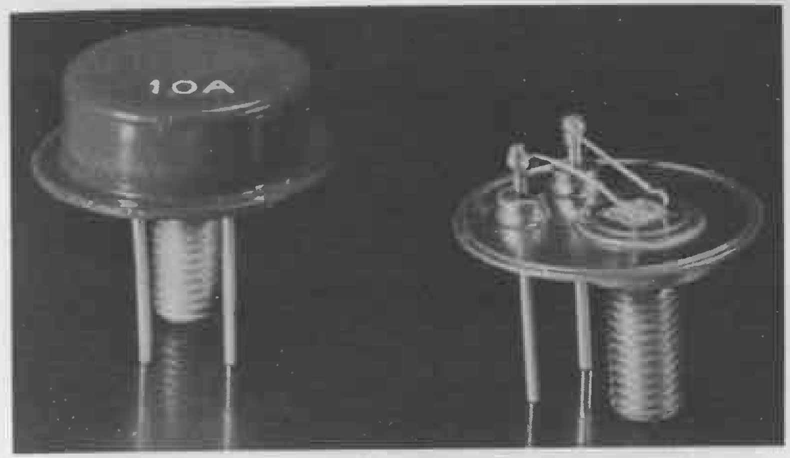
POWER AND HEAT DISSIPATION

If a manufacturer wishes to make a more powerful vacuum tube, he employs larger elements and encloses them in a larger bottle. Although the interelectrode capacitance may increase, generally speaking, the performance will be the same as that of a physically smaller tube, except that the larger tube will be able to handle more power without damage. Unfortunately this is not the case with transistors.

Since the transistor is not a lossless device, it will always produce heat when a current is passed through it. Its design, which minimizes these losses through improved manufacturing techniques, and its ability to get rid of the heat once generated, determine the power-handling ability of the transistor.

Making the amount of active material and the junction larger decreases losses in the transistor, but it also impairs its high-frequency performance. Special techniques must be employed to keep capacitances low while maintaining a large junction to dissipate heat. Often this conflict results in some rather weird internal structures, such as the "comb," the "star," and the "snowflake."

If you have ever torn apart a large power transistor (Fig. 1-3) you may have been surprised to see the tiny piece of semiconductor material inside. The large metal slug which comprises the case (usually copper) is designed to draw heat from the junction and radiate it through the heat sink on which the transistor is mounted. The temperature will always tend to equalize and, in effect, the heat will tend to be drawn out of the junction into the cooler metal to which it is attached. Where operating life and reliability are concerned, the junction temperature of the transistor is the single most important consideration. When building or using transistor transmitters, particular importance should be placed on the operating temperature of the device to keep it as low as possible. As a "rule of thumb," the transistor should be mounted on at least 10



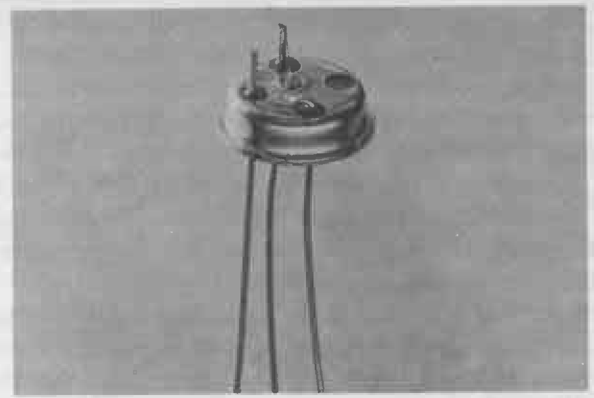


Fig. 1-3. Construction of a typical transistor.

square inches of copper or aluminum for every watt of power input. It is preferable to use commercial heat sinks (Fig. 1-4) which compress 100 square inches into a space of approximately 2″ × 3″ × 4″. Never run the transmitter at its maximum power level in an effort to get the last possible watt into the antenna. Remember, if the power output is reduced by half, it only causes a loss of one "S" unit on a distant receiver. Experimental evidence indicates that for every 10°C. that the junction temperature is reduced, the expected life span of the device will be doubled.

How hot can a transistor get? That depends entirely on the manufacturer's rating, which is determined by the package the

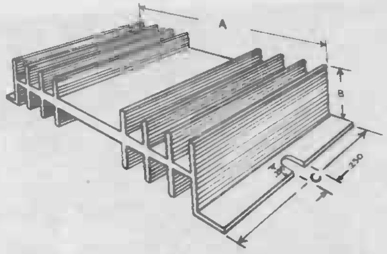


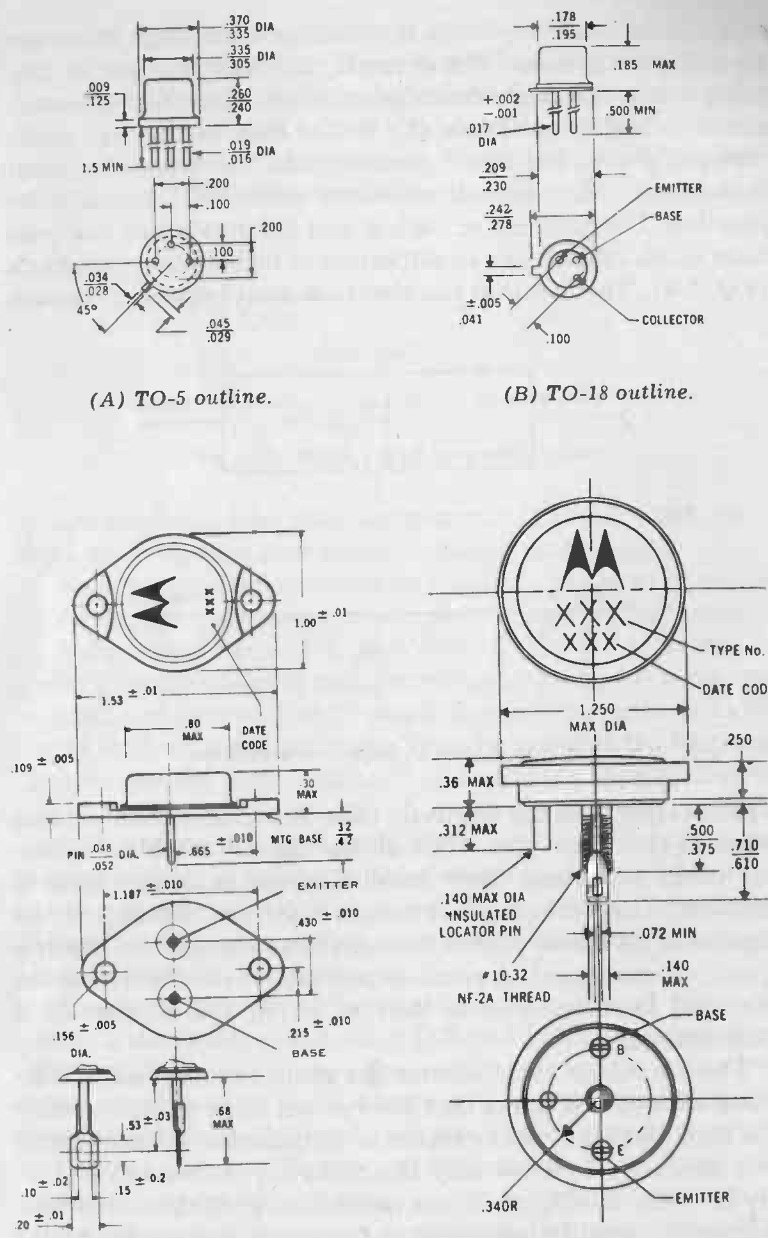
Fig. 1-4. Commercially built heat sink.

transistor chip is mounted in, and on the type of material it is composed of. Generally speaking, one should not feel heat in a germanium device in either the TO-5 or TO-18 case (Fig. 1-5). Germanium devices in the diamond or round case (TO-3 and TO-36, respectively) should never be painful to touch, although they are often operated safely around 130°F., which is just below the threshold of pain. Silicon transistors are capable of operation at much higher temperatures than are germanium transistors. Silicon transistors in the TO-5 and TO-18 package can safely be operated at 130°F. or higher. Silicon transistors in packages which can be mounted on a heat sink often are operated beyond the temperature point where they are painful to touch.

Whenever you use transistors near their maximum temperature rating, always remember that it is excessive junction temperature that destroys the transistor, and the junction itself will always be at a higher temperature than the case. More important, the junction temperature can rise suddenly and destructively long before the increased temperature reaches the outside of the package.

HIGH-FREQUENCY OPERATION

Vacuum tubes are excellent amplifiers of high-frequency radio energy. The *Nuvistor*, for example, is capable of very impressive performance at low cost. The primary limiting factors of high-frequency performance in vacuum tubes are the interelectrode capacitances and transit time; that is, the time it takes the electron to reach the anode after leaving the cathode.



(A) TO-5 outline.

(B) TO-18 outline.

(C) TO-3 outline.

(D) TO-36 outline.

Fig. 1-5. Typical transistor cases.

A similar condition exists in transistors, although it is compounded many times. For example, the junctions are in intimate contact, being physically attached, and they generally exhibit a high capacitance (C_c on the data sheets). To make matters worse, the signal injected into the transistor must pass through the electrode resistance (called R_b) to reach the junction. The presence of both R and C forms a tiny low-pass filter which inhibits the amplification of high-frequency signals (Fig. 1-6). The fact that the electrons must negotiate through

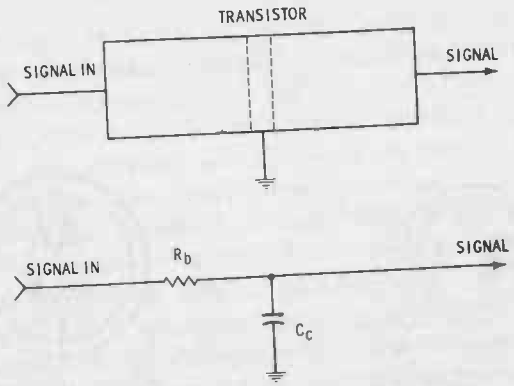


Fig. 1-6. R_b and C_c form low-pass filter.

a solid rather than the relatively effortless confines of a vacuum tends to slow them down. Not all the carriers reach the collector at the same time; they travel different paths and tend to disperse. This storage effect accounts for the inability of the transistor to follow instructions instantaneously. As the frequency is increased, a point is reached where the transistor does not know whether to turn on or off, and as a result it does nothing.

The transistor experimenter generally becomes quite confused with all the terms that have arisen in an effort to define the high-frequency performance of a transistor. At one time it was necessary to know only the *alpha-cutoff frequency* (f_{ab}). By knowing this figure, it was possible to predict roughly how a transistor would work at a given frequency, and one transistor could be compared with another. Alpha cutoff is simply defined as the frequency at which the current gain is down 3 db from

a much lower frequency of, say, 1,000 cycles. The term expresses the loss of gain in the common-base configuration and is not valid for the more useful common-emitter circuit. Further, unless input and output impedances are known, it is not possible to determine power gain. The alpha-cutoff frequency curve for a typical transistor is shown in Fig. 1-7. It should be

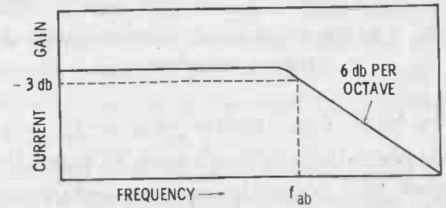


Fig. 1-7. Alpha-cutoff frequency curve.

pointed out that frequency is shown on a log scale which makes the high-frequency gain appear to fall off quite rapidly.

Another term which expresses the high-frequency operation of a transistor is f_t, the *gain-bandwidth product frequency*. It is a convenient figure, for, as a "rule of thumb," it reveals the gain of the transistor at any particular frequency. For example, a transistor with an f_t of 60 would have a power gain of 2 at 30 mc, 4 at 15 mc, 8 at 7.5 mc, etc. Thus, by knowing the frequency and the desired power output it is possible to approximate the driving power required.

A third term, f_{max}, will probably become the industry standard; this term is defined as *the theoretical or computed maximum frequency at which the transistor is capable of oscillation.* Therefore at f_{max} the power gain is equal to unity and the device is not capable of amplification.

The high-frequency characteristics of a typical transistor, such as the Philco Micro Alloy Diffused *(MADT)*, is shown in Fig. 1-8. Note that it has a flat portion where gain is relatively constant. However, as the frequency is increased, a point is reached where the internal workings of the transistor combine to oppose the amplification of high frequencies. Above this point the power falls off at a rate equal to 6 db/octave. Each time the applied frequency is doubled, the power gain drops by 6 db. The point where the rolloff curve intersects the unity-gain line is known as f_{max}.

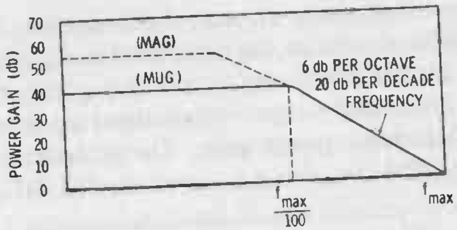


Fig. 1-8. High-frequency characteristics of typical transistor.

If one works back from unity gain or f_{max}, it would seem possible to achieve infinitely high gain at some low frequency. As Fig. 1-8 shows, this is not the case; a modern high-frequency transistor is easily capable of more than 50 db of gain. *Maximum available gain* (MAG in Fig. 1-8) is a function of the device current gain and the operating impedances. However, 50 db of radio-frequency gain concentrated in so small a space tends to be highly unstable and virtually impossible to tame without extremely complex neutralization and shielding schemes. For this reason gains are usually held down to 40 db/stage by impedance mismatching in the input or output circuit, or in both. Thus the 40-db figure is usually considered to be the *maximum useful gain* (MUG). Although some power gain is lost, the increased stability and reduced circuit complexity make it a small price to pay.

Once a 40-db, flat-gain line is assumed and f_{max} (which is given on the manufacturer's data sheet) is known, it is possible to predict the maximum power gain for the transistor at any frequency. The rolloff is more or less constant at 6 db/octave or 20 db/decade. Assuming 40-db maximum usable gain, or two decades, the frequency of the rolloff knee can be calculated simply by dividing the f_{max} figure by 100. If the device f_{max} is 100 mc, the knee occurs at 1 mc. At any frequency above this point the power gain will diminish.

RUGGEDNESS

Newer transistors can withstand high external temperatures. Most modern germanium transistors may be soldered into a circuit without heat-sinking the leads or the case itself. The

leads can be heated, as in dip-soldering of printed-circuit boards, for as long as 10 seconds without internal damage. All silicon transistors can be soldered into a circuit without damage.

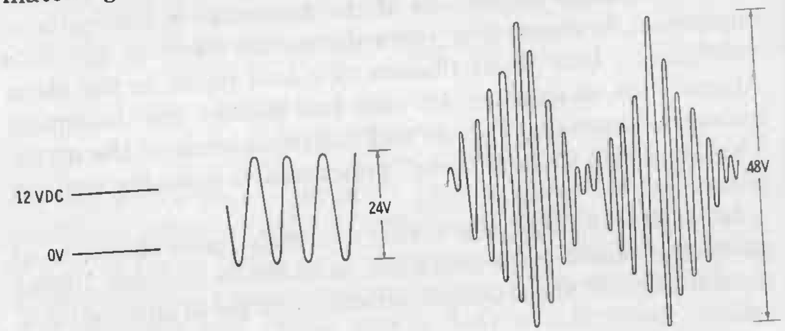
The electrical ruggedness of the transistor is not quite so impressive, however. For years the author used an 807 tube which had a hole about the size of a lead pencil in the plate. At one time or another the tube had flashed over internally because of excessive voltage and vaporized some of the metal. The tube, even with its "lung" punctured, is probably working today.

In contrast to this, the author currently has a collection of transistors which were subjected to excessive voltage. Unlike the durable 807, these transistors are no longer of any electrical value.

Excessive voltage across either of the junctions in a transistor can result in instantaneous destruction due to localized heating. The carriers, propelled at enormous speeds and in great quantities, tend to "gang up" in one area while taking the avenue of least resistance. This causes a sudden temperature rise which virtually melts the transistor and almost instantaneously shorts the junction. The destructive avalanche is much faster than the finest fuse or circuit breaker can cope with.

If a resistance were inserted in series with the device to limit current flow below the destructive heating point, no damage would occur, even if the junction voltage (V_{CE}) were increased far beyond the maximum rating. For example, if a 1-meg resistor were inserted in series with a general-purpose PNP audio transistor and connected to a 500-volt supply, no damage would occur to the junction because the current would be limited to a few microamperes.

Unfortunately, we cannot do this in a transistor transmitter; the same resistor which limits destructive currents also limits the power output below the useful level. Thus the maximum voltage ratings must be closely observed. It should also be stressed that the maximum voltage supplied by the DC source must always be much less than the maximum transistor rating. For example, in a class-C RF power amplifier, the back emf and flywheel action of the coil can generate peak voltages twice the supply value (Figs. 1-9A and B). If the stage is high-level collector modulated, the RF peaks will reach at least four times the supply voltage (Fig. 1-9C). If the antenna loading or matching is improper, the voltage can reach even higher levels,



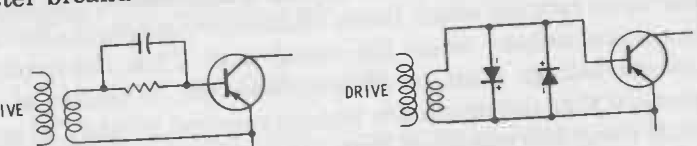
(A) Power-supply voltage of 12 volts on collector. *(B) Increase due to flywheel effect and back emf.* *(C) Increase due to high-level collector modulation.*

Fig. 1-9. Voltage increase due to back emf, flywheel effect and high-level modulation.

and any one of these peaks can have sufficient duration to cause instantaneous destruction of the transistor.

Thus, as a rule of thumb for AM transmitters, always select a transistor with a V_{CC} rating at least four times the supply voltage, or reduce the supply until it is one-fourth the maximum rating. A transistor with a six-times rating would be more conservative and result in a greater safety margin in the presence of transients or voltage surges. In CW, FM, and SSB equipment a device with a two-times supply rating should be selected, and one with three times the rating is preferable.

In certain transistors, notably the mesa types, the base-emitter breakdown voltage is also important. Excessive drive



(A) Drive-limiting resistor. *(B) Peak-voltage limiter.*

Fig. 1-10. Base-emitter breakdown protective circuits.

can cause this junction to break down. Drive-limiting resistors or peak-voltage limiting devices are often used to prevent this form of destruction. These techniques are shown in Figs. 1-10A and B.

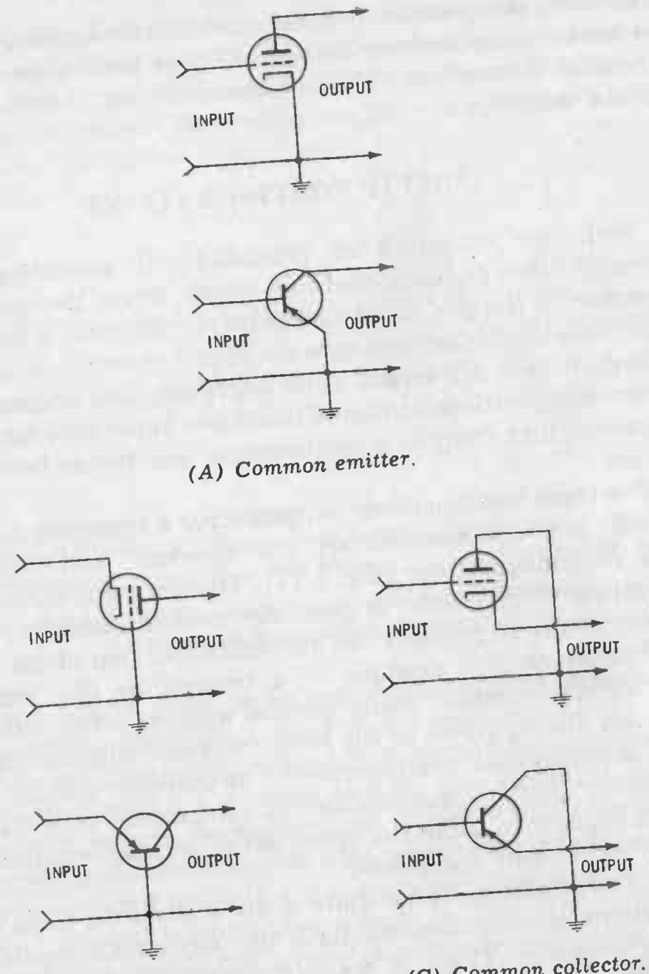
CIRCUIT CONFIGURATIONS

With the exception of grounded-grid amplifiers, most vacuum tubes are employed in a circuit where the cathode is common to the grid-input and plate-output circuit. Likewise, this is the customary arrangement found in transistor circuits. However, there are several other interesting and unusual transistor configurations employed to achieve impedance matching, improved high-frequency performance, and better heat dissipation.

The three basic methods of connecting a transistor into the circuit are *common-emitter*, *common-base* and *common-collector* configurations (Fig. 1-11). The common-emitter configuration (Fig. 1-10A) is generally used in radio-frequency circuits, since it provides the highest power gain of the three configurations. For example, in a transmitter the common-emitter configuration would require approximately one-fifth as much driving power as the same transistor when connected in a common-base configuration. The common-emitter configuration almost approximates the comparable tube circuit, which appears to make the operation of common-emitter circuitry more easily understood.

A third advantage is not quite so obvious. Of the three configurations, the common emitter is the only circuit capable of phase inversion. Thus, in radio-frequency amplifier circuits the output is degenerative (opposite phase) with respect to the input. Inherently the stage becomes more stable, particularly on the flat portion of the transistor response curve, when connected common emitter. Neutralization will result in an increase in power gain of as much as 3 db. Because of increased stability in the common-emitter mode, circuits tend to become more reproduceable; that is, the gain and bandwidth variations from transistor to transistor are greatly reduced.

There is considerable confusion regarding the relative merits of common emitter versus common base in transmitter circuits.



(A) Common emitter.

(B) Common base.

(C) Common collector.

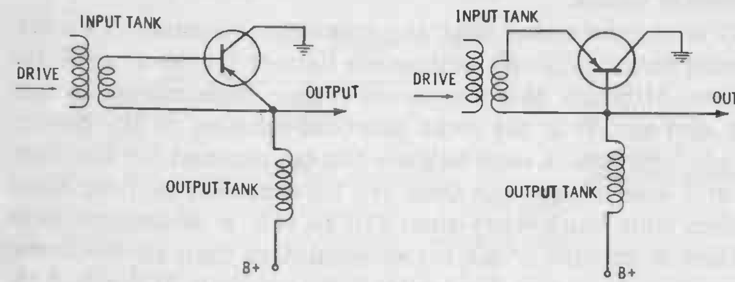
Fig. 1-11. Basic configurations and vacuum tube equivalents.

Actually the f_{max} is the same for all configurations and the actual circuit used would seem to be a matter of choice. However, there are subtle differences which are not immediately obvious. If the frequency of operation is such that the transistor is operating on the flat-gain portion of the frequency-response curve, the common-emitter configuration should be employed for the many reasons just described. However, the

common-base configuration (Fig. 1-11B) would be used on the 6 db/octave slope frequencies. This is because the common-base configuration becomes somewhat regenerative due to the in-phase feedback energy. This regeneration increases the power gain of the stage noticeably and tends to produce more output power for a given amount of driving power. This becomes particularly true and much more pronounced when the transistor is operated near f_{max} and only a few db of gain are available. It is quite usual to find common-base circuitry employed in equipment operating on VHF and UHF bands. Like the grounded-grid, vacuum-tube amplifier, some of the driving power appears in the output tank. This makes the stage appear to have a higher power gain than is actually the case.

The common-collector configuration (Fig. 1-11C) is an impedance-matching scheme and is seldom found in transmitter circuits. Its vacuum-tube equivalent is the cathode follower, and like this circuit, the emitter follower is capable of matching a high impedance to a low impedance. It is not capable of voltage amplification but does have current gain. The emitter follower may occasionally be used for impedance matching into high-power stages when link or capacitive coupling is undesirable. The input impedance of this configuration is determined primarily by the input impedance of the driven stage multiplied by the current gain of the transistor in the driving stage.

In transistor transmitters (particularly those where the transistors operate at elevated temperatures) a variation of



(A) Common emitter, grounded collector.

(B) Common base, grounded collector.

Fig. 1-12. Variations of basic circuits.

these circuits may be employed, as shown in Fig. 1-12. This is done to permit direct grounding of the collector to the heat sink without the use of an insulating mica washer. Eliminating this resistance to heat transfer greatly reduces the transistor junction temperature. It should be stressed that common emitter, for example, is still common emitter even if the transistor collector is grounded. This is why the term *grounded emitter* instead of common emitter can be quite misleading. To illustrate the point, consider a battery, a pilot lamp, and a meter connected in series. Any part in the circuit could be called common and could be grounded to a chassis without affecting the circuit. The same is true in transistor circuits, although they are considerably more complex than the simple circuit just described. Ground or common is strictly an arbitrary term and can be placed at any point so long as the relationship of other components to this common point is faithfully observed.

POWER SUPPLIES

A particular transistor transmitter circuit may require anywhere between 1.5 and 24 volts for proper operation as contrasted to the dangerously high potentials involved in vacuum-tube equipment. This coin has two sides, however, because for a given amount of power considerably more current must flow in the transistor transmitter circuit. This requires heavier wiring, larger meter ranges, and greatly increased bypass capacitor values.

It is recommended that the transistor transmitter experimenter obtain a 12-volt automobile battery for use as a voltage source. Although this may seem rather unglamorous in this day and age, it is the most practical solution to the power-supply problem. A used battery can be obtained for less than $5 and even a new one from the supermarket or drug store seldom runs much more than $10 to $12. A 40-ampere/hour battery is capable of far better regulation than an electronic power supply costing five to ten times as much. If the lead tie lugs are exposed, they may be drilled and tapped for connections to make an adjustable voltage source, as shown in the accompanying photos (Figs. 1-13 or 1-14A and B). The case

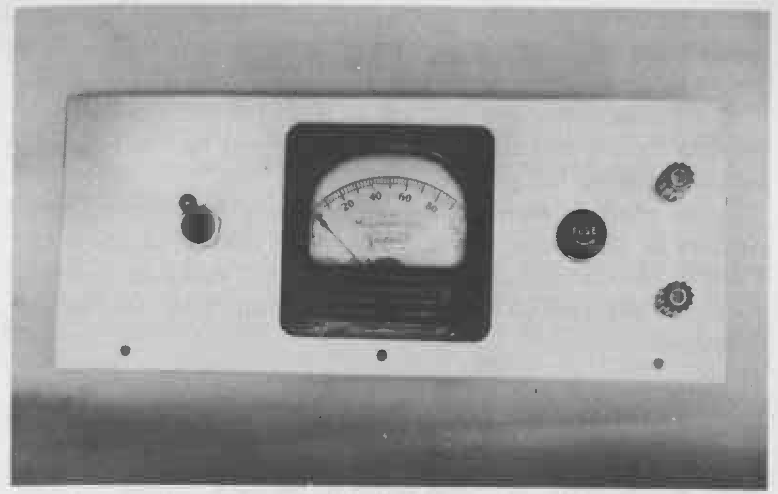
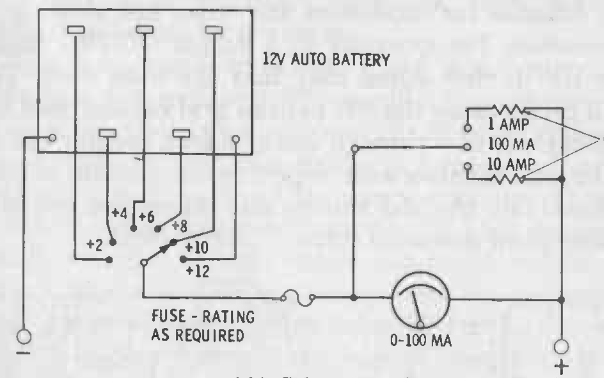
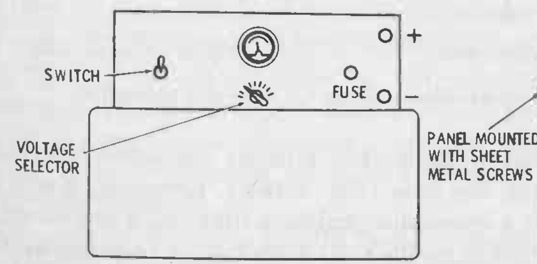


Fig. 1-13. Inexpensive bench voltage source.



(A) Schematic diagram.



(B) Construction details.

Fig. 1-14. Details of bench voltage source.

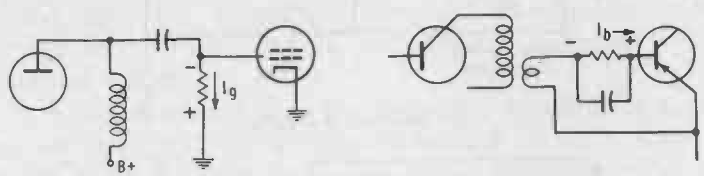
is usually thick enough to permit drilling and tapping for the installation of a control panel containing a current meter or meter jack, a voltage switch, a fuse holder, etc. (Fig. 1-13B). The adjustable-voltage feature is a useful addition since it permits testing and tuning of transmitters at reduced voltages before applying full power.

In an effort to standardize voltages to aid constructors, all circuits in this book are optimized for use at 12 volts DC. Further, all circuits are designed with the common bus connected to negative so that they may be used in an automobile, if desired.

BIAS AND CLASSES OF OPERATION

Although the theory of operation for tubes and transistors is quite different, they are both amplifiers and the classes of operation are usually considered to be approximately the same.

The bias schemes for transistors and tubes are often quite different, however. For example, in a simple 6V6-807 transmitter, the RF driving signal may bias the final tube. The positive half cycles cause the 807 to draw grid current, and the subsequent current flow through the grid-leak resistor causes the grid to become negative with respect to the cathode. Should the drive signal fail, the bias voltage also ceases and the tube will very likely draw excessive current (Fig. 1-15A).



(A) Tube signal bias. *(B) Signal-induced bias.*

Fig. 1-15. Signal-bias systems for tube and transistor.

In most transistor transmitter circuits the signal also supplies some or all of the bias (Fig. 1-15B). However, it will be remembered that a transistor, unlike a tube, does not conduct until a forward bias is applied. In a transistor transmitter the forward bias is also supplied by the drive signal. However, in the absence of signal the forward bias and the collector current

cease. In effect the transistor is self-protecting, much the same as the 6V6-807 transmitter would be if a screen-clamp tube (controlled by grid bias) were employed.

Class-A transistor operation is almost identical to tube operation; that is, the device is biased near the center of the linear curve, and the output is considered to be a true replica of the input waveform.

Class-B transistor operation is similar to zero-bias, class-B tube circuitry. A small amount of forward bias may be applied to the stage. Half of the signal cycle pulses the stage into operation, while the other half cycle drives the stage well into cutoff.

Class-C transistor circuits seldom use the large cutoff bias found in vacuum-tube circuits. The transistor will have neither a forward nor a reverse bias applied. To achieve class C, a resistor is usually inserted in series with the drive signal (Fig. 1-15B). Base current, caused by conduction on one half cycle, causes current to flow in this "base-leak" resistance, which generates reverse bias for the stage.

Class-A operation is seldom found in a transmitter circuit other than for oscillators. This mode is characterized by a pure sine-wave output independent of the flywheel action of a tank circuit. Class-A operation results in a freedom from harmonic generation and improved frequency stability. This condition is difficult to achieve in oscillator circuits because the transistor usually swings violently between saturation and cutoff and the output approaches the appearance of a square wave. The flywheel action of a tank circuit helps restore the sine waveshape, but it is seldom perfect if the original waveshape is badly distorted. To obtain true class-A operation, careful attention must be paid to such details as biasing, impedance matching, and, most important, the drive level of the feedback energy.

Actually, class-A operation is only desirable in variable-frequency oscillators where stability is of paramount importance. It is much less important in crystal oscillators, since the driven stage contributes far more distortion and subsequent harmonic radiation.

Class-B operation is usually found in transmitters employing transistors well down the 6 db/octave slope where adequate drive is lacking or more stage power gain is required. Many power amplifiers which are supposed to be operating class C may actually be in self-bias class B because of insufficient drive to create a large reverse bias. In such cases a small amount of forward bias (just enough to place the operating point at the bottom of the linear portion of the curve) will improve output by overcoming the base barrier potential. This form of class-B biasing also seems to improve the output of a frequency multiplier.

The characteristic of class-C transistor operation is identical with vacuum-tube operation, although bias may be obtained in a slightly different manner. In a typical PNP common-emitter amplifier, for example, the negative peaks drive the stage into conduction while the positive peaks drive the stage beyond cutoff.

ALIGNMENT

A transistor transmitter is considerably easier to align than a vacuum-tube rig. It is only necessary to connect a suitable dummy load to the output and tune everything for maximum power output. Unless the final transistor is severely overloaded because of excessive drive (which is seldom the case), no damage will occur due to excessive loading, out-of-resonance tuning, insufficient drive, etc. The only precaution, which cannot be stressed too often, is *always have the output properly loaded.* If this precaution is not observed faithfully, excessive voltage of sufficient amplitude to damage or destroy the transistor can be generated.

An excellent dummy load can be constructed by employing pilot lamps (Fig. 1-16). Although the brilliance determines the actual impedance of the bulbs (the impedance will be less at reduced brilliance) and tuning changes the brilliance, this technique is very successful for adjusting transmitters. For equipment in the 100-milliwatt (0.1-watt) class, a No. 48 or 49 pilot bulb makes an excellent output indicator. For 1-watt transmitters, a No. 47 pilot is excellent. Four series-parallel No. 47 bulbs can be used for transmitters in the 5-watt class. Above this power No. 44 bulbs can be series-parallel connected to make up the correct impedance.

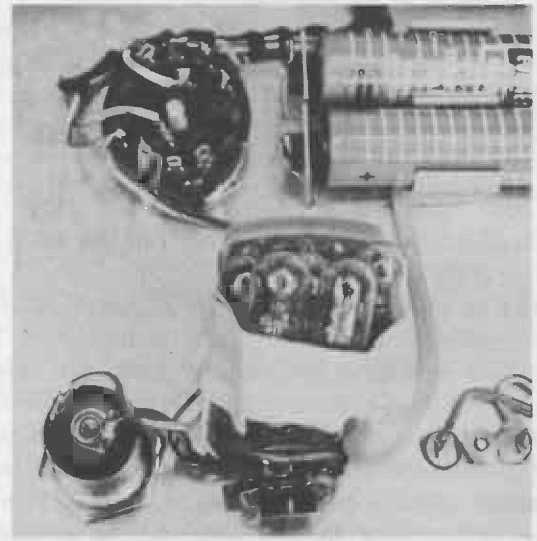


Fig. 1-16. Practical dummy-load system for transmitters.

The impedance of the bulbs may be easily calculated by dividing the voltage rating of the bulb by the current rating. This will equal the bulb resistance (and therefore impedance) at full brilliance. For example, a No. 47 pilot lamp (0.15 ampere at 6.3 volts) has a full-brilliance resistance of 42 ohms, which is a satisfactory match for 50 ohms. A No. 44 pilot, rated at 0.25 ampere at 6.3 volts, has a resistance of 25 ohms, and two in series make a perfect 50-ohm match.

The amount of power required to light the lamp to full brilliance can be determined by multiplying the voltage and current rating. For example, slightly less than 1 watt will fully illuminate a No. 47 lamp, while a No. 44 requires slightly more than 1.5 watts. Combinations of these bulbs require proportionately more power. By knowing these two facts, dummy-load strings can be constructed for any power or impedance. This technique also makes an excellent way of estimating transmitter power output simply by comparing it to a fully lit bulb as illustrated in Fig. 1-17. Power can be measured quite accurately by varying the voltage to a comparison bulb until it is the same brilliance as the dummy-load bulb or bulbs. By knowing the power consumed by the comparison bulb, the power

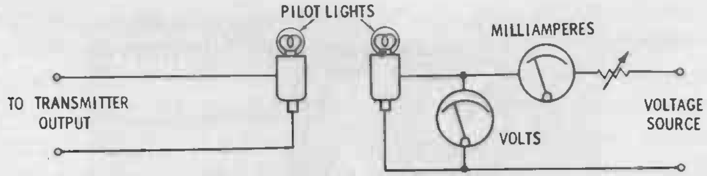


Fig. 1-17. System for estimating power output.

required to light the dummy-load bulb to the same brilliance can be accurately calculated.

It should be pointed out that the comparison method becomes increasingly inaccurate as the transmitting frequency is raised. This is due to the inductance of the spiral-wound lamp filaments. The error introduced by this inductance is not particularly severe below 30 mc, however. There is one other pitfall which almost always traps transistor transmitter experimenters when measuring power output. One common method of determining the power output of a transmitter is to measure the voltage developed across a known value of load resistance. The voltage is then squared and divided by the value of load resistance and the resultant is the power in watts. However, few RF voltmeter probes are accurate, and small errors in voltage produce large errors in power. Even more important, the probes are calibrated to read rms volts. As is often the case, transmitter output may not be a pure sine wave and the voltmeter may indicate something other than the true rms value. The voltage error increases as the transmitter output continues to depart from a true sine wave. For this reason it is strongly recommended that pilot lamps or other heating devices be used as power-output indicators.

Diode voltmeters frequently create other problems. It is often noted, when observing the output of a transistor transmitter on an SWR bridge or field-strength indicator, that modulating the transmitter makes the meter kick downward rather than showing the increased reading customarily associated with upward modulation. However, this is an effect common to germanium diode detectors and can generally be ignored. The most accurate test of upward modulation, other than an oscilloscope, is to observe the dummy-load pilot lamps. They should increase in brilliance a little more than 20% for true 100% upward modulation.

2

Oscillators

Oscillator circuits employing vacuum tubes generally are confined to two or three well-known types. The transistor, however, can be used in several configurations, as well as variations of these configurations. A myriad of circuits result when these configurations are multiplied by the various types of oscillators.

Figs. 2-1 and 2-2 illustrate a few of these circuits. None of these oscillators are crystal controlled; the addition of a

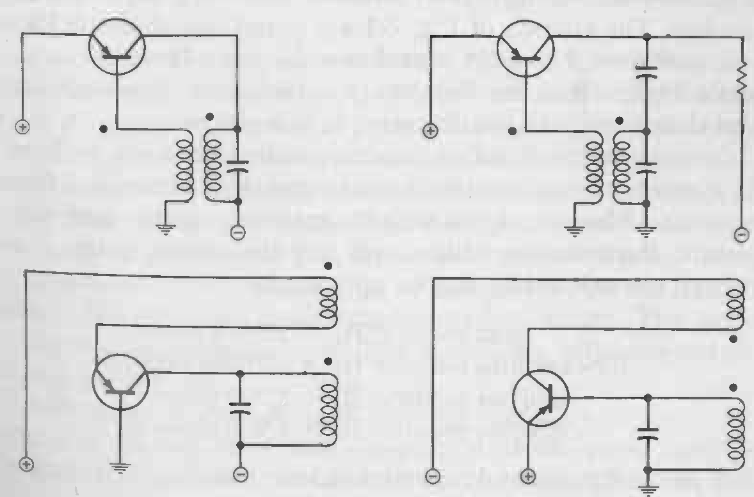
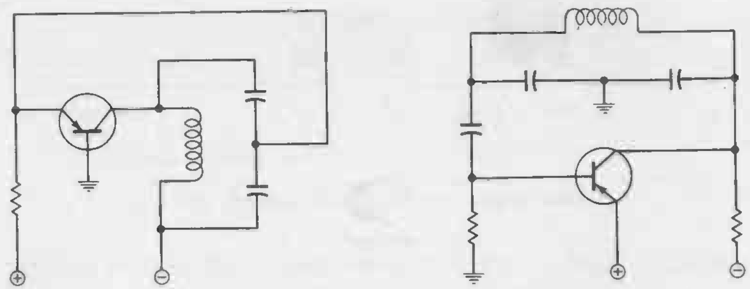
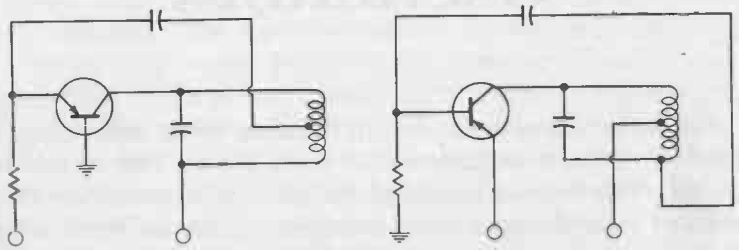


Fig. 2-1. Variations of tickler-coil oscillators.



(A) Basic Hartley circuits.



(B) Basic Colpitts circuits.

Fig. 2-2. Hartley- and Colpitts-type oscillators.

frequency-stabilizing crystal results in another group of similar circuits. The circuits of Fig. 2-1 are variations of the tickler-coil oscillator. Fig. 2-2A represents the basic Hartley circuit, while Fig. 2-2B is the Colpitts-type oscillator. These circuits and their variations are discussed in this chapter.

In the circuits to follow, exact resistor values are replaced by symbols to simplify the schematic and the discussion of their operation. However, if you wish to experiment with a particular circuit, the following values will get the circuit going even though the values may not be optimum:

base resistor R_B = 10,000 ohms
forward-bias resistor R_F = 100,000 ohms
emitter resistor R_E = 1,000 ohms
collector resistor R_C = 4,700 ohms

The particular coil and capacitor values in each circuit depend on the frequency of operation.

Fig. 2-3 represents the "black box" equivalent of an oscillator circuit. By definition an oscillator is simply an amplifying stage connected in such a manner that some of the output

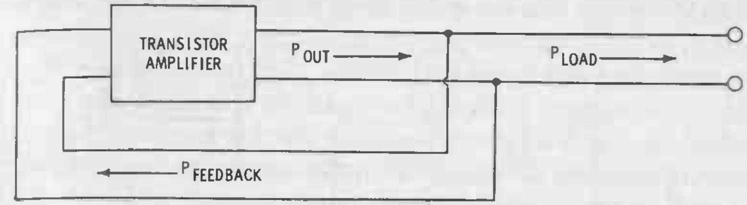


Fig. 2-3. Black box equivalent of an oscillator.

power (P_o) is coupled back into the input circuit. The feedback energy (P_{feed}) must be in-phase or regenerative to sustain oscillations. An important point to remember is that the energy used for feedback is no longer available as useful power output. The circuit must also contain a frequency-selective network to determine the frequency of oscillation. This circuit can take the form of an LC network (a tuned circuit), a quartz crystal, or in the case of audio-frequency oscillators, an RC network. Also necessary, though not shown in Fig. 2-3, are the bias components required to bias the transistor and stabilize the operating point.

The input and output impedances of a vacuum-tube oscillator are both high; therefore, losses in the feedback network seldom cause a serious loss of power output. On the other hand, the common-base oscillator is characterized by a relatively high output impedance but an extremely low input impedance. If some form of impedance matching is not employed, it is necessary to use "brute force" to supply sufficient feedback to sustain oscillations. For all practical purposes, this power is wasted and is not available for driving succeeding stages. The same thing is true for common-emitter oscillators, although not to the same extent.

Although a transistor oscillator will operate well above the alpha-cutoff frequency (see Chapter 1), it can never operate at f_{max}. This is a computed frequency, based partly on R_B and C_C, where the gain of the transistor is zero. There is always loss in the feedback network and the transistor must supply the power to overcome this loss. As a rule of thumb you can successfully use a transistor to approximately one-half of f_{max}, although the power output at this frequency will be very small.

Let's see how a typical radio-frequency amplifier can be used as an oscillator. Fig. 2-4 is the schematic for a PNP RF ampli-

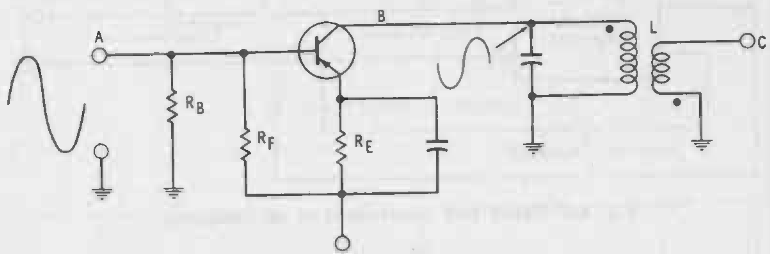


Fig. 2-4. PNP RF-amplifier circuit.

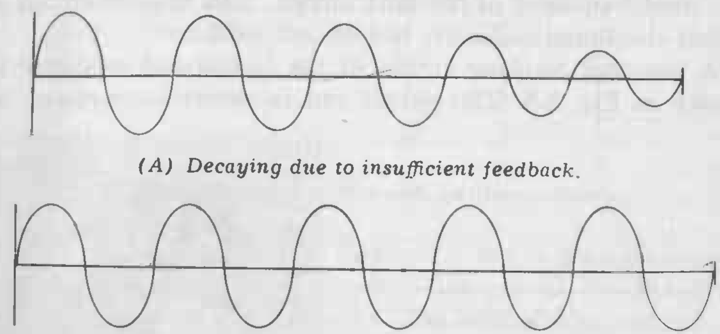
fier. The resistive divider network (R_F and R_B) provides forward bias, while R_E is connected in series with the emitter circuit to provide DC stabilization with varying temperatures and supply voltages. Resistor R_E is bypassed so that the signal current which flows through this resistor does not oppose the input signal and lower the stage gain. An output tank, resonant at the signal frequency, completes the circuit. A signal applied to the input of the stage will appear amplified many times across the tank circuit. As the input signal becomes more positive, it decreases the forward bias, and at the same time the voltage across the tank decreases and is said to be *negative-going*. Thus, in this circuit the output signal is always out of phase with the input signal. If the transistor is operated on the flat portion of its frequency-response curve and the tuned circuit is resonant at the signal frequency, this relationship will remain constant at 180° of phase shift. Signal coupling back through the transistor capacitance lowers the stage gain since it opposes the input signal.

The signal voltage induced in the link winding may be either in phase or out of phase, depending on the winding direction and the polarity of connections. If, for example, the start of both windings is common and they are wound in the same direction, the voltage at point C will be 180° out of phase with respect to point B. Where confusion can arise as a result of

winding polarity, dots are usually included near the coil or transformer to indicate the start of the winding (Fig. 2-1).

Tickler-Coil Oscillator

If the voltage at point C is compared to the signal at point A, the two will be in phase, since the transistor has introduced 180° of phase shift and the coil (connected as just described) contributes another 180°, making a total of 360°. If points C and A are connected, the stage will oscillate immediately. The instant that power is applied to the circuit, forward bias on the transistor causes a surge of current to flow in the tank circuit. This causes a damped wave (Fig. 2-5) of current to flow in the inductance which is coupled to the link winding. If there is insufficient feedback, the waveform decays as in Fig. 2-5A



(A) Decaying due to insufficient feedback.

(B) Sine-wave output with proper feedback.

Fig. 2-5. Damped-wave oscillator output.

and the stage refuses to oscillate. Assuming, however, that all conditions are present to sustain oscillations, the positive-going half cycle is link-coupled to the base or input of the stage. The action of the signal induced in the link winding inverts the phase so that the feedback at this instant is negative-going. This, of course, causes the transistor to draw even more current and drives the stage further into conduction until the transistor saturates and can no longer amplify. Since no further change in collector current can occur, the feedback signal ceases. This represents the peak of the sine wave. At this point the static bias applied to the oscillator takes over and attempts to decrease the collector current. This decreases the current through the coil and causes the direction or polarity of the feedback to reverse so that is is now positive-going at the emitter, thus driving the stage into cutoff. When cutoff is reached, which represents the negative peak in the cycle, the stage can once again no longer amplify and the feedback ceases. As before, the forward bias tries to bring the collector current up to the static-bias point by overcoming the signal-induced reverse bias. This causes the feedback polarity to reverse and the cycle repeats. This second cycle reinforces the damped wave so that the output signal consists of a series of sine waves as in Fig. 2-5B, rather than the decaying waveform in Fig. 2-5A. This oscillation will continue until the feedback path is broken or until the supply voltage is disconnected. The frequency of oscillation is determined almost entirely by the resonant frequency of the tank circuit. This type of circuit is called the *tuned-collector, tickler-coil oscillator.*

A practical working circuit of the tickler-coil oscillator is shown in Fig. 2-6. The circuit can be easily constructed to

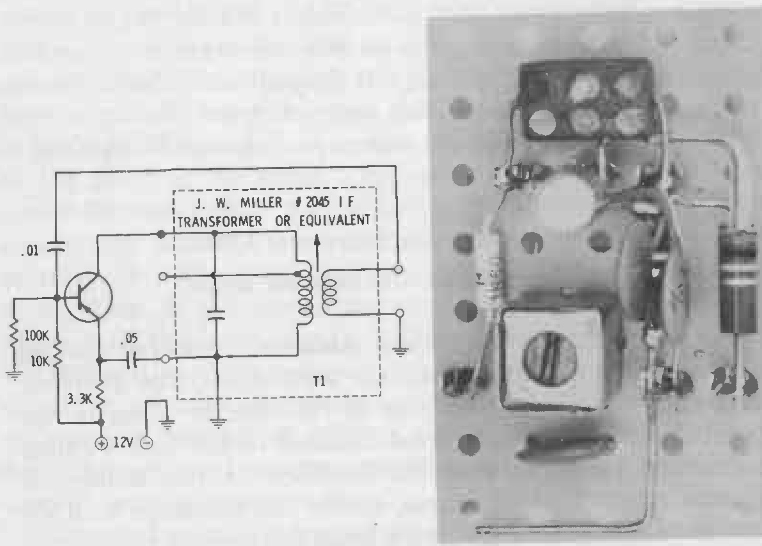


Fig. 2-6. Practical circuit of a tickler-coil oscillator.

investigate the principles just discussed. The coil and capacitor may be resonated at any desired frequency within the operating range of the transistor. There should be approximately

one-tenth the number of turns on the link as are on the primary. The circuit shown is used as a beat-frequency oscillator and the tuned circuit consists of a J. W. Miller No. 2045 IF transformer (T1). If the circuit does not oscillate immediately, reverse the connections to the link winding. Any general-purpose RF PNP transistor can be used with the typical values shown so long as it is capable of a reasonable amount of amplification at the frequency of oscillation.

Fig. 2-7 illustrates another form of tickler-coil oscillator that employs the tuned circuit in the base or input section. A con-

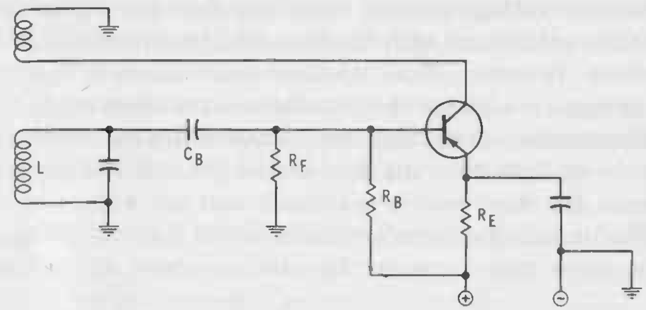


Fig. 2-7. Alternate tickler-coil oscillator circuit.

siderable mismatch occurs in this circuit due to the high tuned-circuit impedance to be coupled through C_B to the low-impedance base of the transistor. The mismatch causes severe instability; for this reason the circuit is seldom used.

Hartley Oscillator

For reasons of circuit complexity, the use of a link winding may be undesirable. Also the link may be required for coupling driving power to a succeeding stage. In such cases the tran-

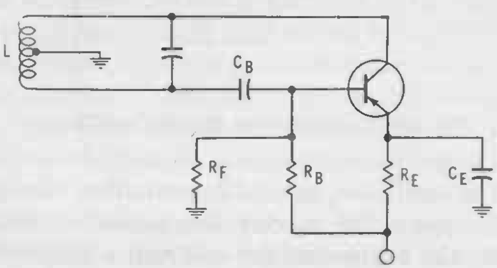


Fig. 2-8. Hartley-type oscillator.

sistor may be used in a Hartley-type oscillator (Fig. 2-8). In this configuration the tuned circuit is common to both the input and output circuits. The coil is tapped, and the tap point is at ground potential. Thus the RF at each end of the coil is 180° out of phase with respect to the other end. The transistor provides the remaining 180° of shift required to sustain oscillations. If the coil is center-tapped, capacitor C_B is made small in value to match the high-impedance tuned circuit to the low-base impedance. As mentioned earlier, this results in poor frequency stability. The reason for this is that C_B forms a capacitive voltage divider (and therefore an impedance divider) in conjunction with the base-emitter capacitance of the transistor. However, this junction capacitance is subject to wide changes in value as the transistor parameters are changed. For this reason a preferable impedance-matching method is to place the coil tap near the base end of the coil (where the impedances are more nearly matched) and use a large value of C_B. The impedance transformation would then be determined by the turns ratio between the winding above and below the tap.

The Hartley oscillator illustrates the point mentioned earlier, that ground or common is an arbitrary term. Observe the circuit in Fig. 2-9. Although this appears to be an entirely dif-

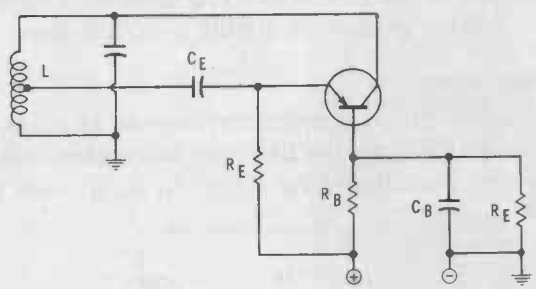


Fig. 2-9. Common-base Hartley oscillator.

ferent form of oscillator, careful examination reveals that the only actual changes are moving the ground symbol from the center tap to the bottom of the coil and a slight physical rearrangement of the position of the components. Note also that

moving the ground symbol has changed the Hartley from a common-emitter to a common-base configuration. However, the performance of either circuit is identical.

Colpitts Oscillator

Another oscillator circuit which generally operates in common base is the Colpitts. The Colpitts circuit may take on many similar forms with only minor differences tailored to suit the frequency of operation. The common-base Colpitts circuit seldom fails to oscillate. Because the impedances may be closely matched, the Colpitts has the best frequency stability of any oscillator circuit. The Colpitts has an additional advantage in that it does not require a tap on the coil.

Consider the circuit shown in Fig. 2-10 and assume that it is a common-base, radio-frequency amplifier. Signals applied to the emitter will appear greatly amplified but not phase-inverted

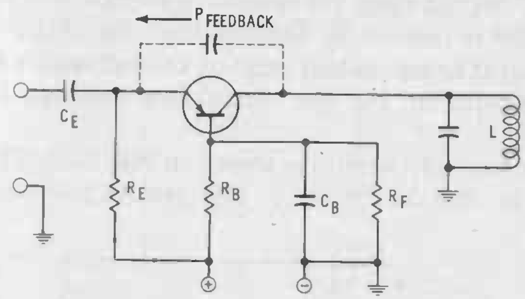


Fig. 2-10. Colpitts oscillator circuit.

in the collector circuit. If a capacitor is now connected between the collector and emitter, the in-phase energy is returned to the input of the amplifier and a feedback path is created. This configuration is particularly well suited for VHF applications where only a small amount of feedback capacitance is required. Many transistors have sufficient junction capacitance to oscillate without the aid of the external capacitor. This is particularly true if an RF choke is inserted in series with the emitter resistance.

Fig. 2-11 illustrates another form of Colpitts oscillator which is more suitable for the high-frequency portion of the radio spectrum. Here a capacitive voltage divider is connected di-

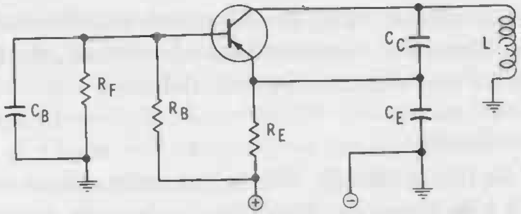


Fig. 2-11. Colpitts with capacitive-divider feedback.

rectly across the coil with the emitter connected to the junction of the two capacitors. As mentioned earlier, this capacitive divider makes an ideal system for matching the high impedance of the collector to the low impedance of the emitter. Because the emitter has a lower impedance than the collector, capacitor C_E is always much larger in value than C_C to provide a correct match. Capacitor C_E should be approximately 10 times the value of C_C. By changing the ratio of capacitance, the feedback can be varied to increase or decrease the vigor of the oscillator. A tap placed at an equivalent point on the coil results in exactly the same operation, but the circuit then becomes a Hartley oscillator.

Another form of Colpitts is shown in Fig. 2-12. This circuit is similar to that in Fig. 2-11. The ground has been moved

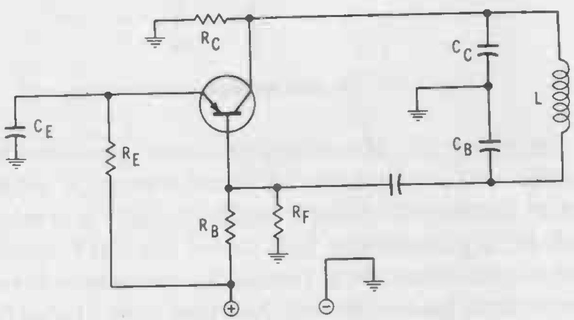


Fig. 2-12. Alternate Colpitts circuit.

from the bottom end of the tuned circuit to the junction of the capacitive divider, and a collector DC return resistor is employed. Although the action and performance of the oscillator are similar to those of the one shown in Fig. 2-11, the explana-

tion of operation is somewhat different. Since the dividers place an imaginary tap on the coil and since this point is at RF ground, the coil is capable of phase inversion as in the Hartley. Thus the signal fed back to the base through C_B is 180° out of phase with respect to the collector signal. With the additional 180° of phase shift supplied by the transistor, the stage is capable of sustaining oscillations. The impedance-matching functions of the capacitors in this circuit are the same as for Fig. 2-11.

One disadvantage of the Colpitts which may have been uncovered by the reader is the fact that the circuits shown so far require a variable inductance to change resonant frequency. If either of the two divider capacitors is varied, it will change the resonant frequency but will also upset the impedance match. Introducing this mismatch will degrade the oscillator performance. The problem can be avoided by using a dual-section capacitor; however, a simpler scheme is shown in Fig. 2-13. The divider capacitors are made 5 or 10% smaller than

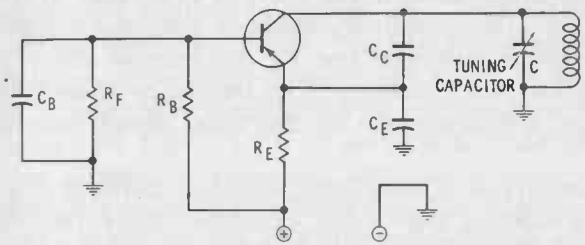


Fig. 2-13. Colpitts with tuning capacitor.

their normal value. A variable capacitor is then connected in parallel with the divider and used to adjust the resonant frequency of the circuit. This does not in any way upset the impedance-matching characteristics of the divider, and the theory of operation remains the same so long as the fixed capacitance (C_C) is at least 10 times the value of the variable capacitor. It should also be pointed out that for variable-frequency oscillators this scheme will provide straight-line frequency tuning of the oscillator. For example, 1 mmf of change at minimum capacitance will shift the frequency of oscillation the same amount as for a 1-mmf change at maximum capacitance.

Still another form of Colpitts is the Clapp oscillator illustrated in Fig. 2-14. The Clapp varies from the Colpitts in that a series-tuned resonant circuit (consisting of L and C) is substituted for the parallel tank. Capacitor C_C shunting the

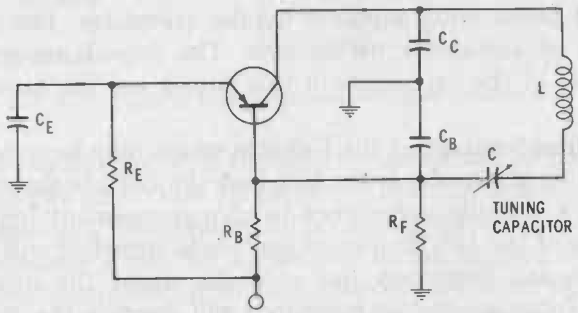


Fig. 2-14. Clapp oscillator configuration.

collector junction and C_B in parallel with the base junction are made quite large, being in the order of 1,000 mmf at 5 megacycles. This large capacitance tends to mask the variations in the junction capacitance and its detrimental effect on the oscillator frequency stability. The frequency of oscillation is determined almost entirely by the tuned circuit consisting of L and C.

Another version of the Clapp oscillator, shown in Fig. 2-15, permits grounding the tuning capacitor. Since the value of

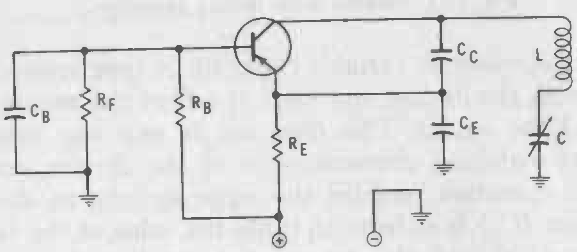


Fig. 2-15. Alternate Clapp oscillator circuit.

C_B is large with respect to C_E, the base is more nearly at RF ground and the transistor is considered to be operating in the common-base mode.

OSCILLATOR STABILITY

The oscillators which have been discussed so far are free running as contrasted to an oscillator which is stabilized by a quartz crystal. When a crystal is not employed, the circuit is generally referred to as a variable-frequency oscillator. To obtain the greatest frequency stability in any oscillator, three factors must be taken into consideration: (1) bias, (2) junction capacitance, and (3) temperature.

For best stability the oscillator should be biased near the center of the transistor I_E/I_C curve so that operation occurs over the linear portion of the curve. If excessive drive or improper bias carries the transistor violently into the nonlinear region, frequency instability will occur. For the same reason, the oscillator voltage source should be stabilized to prevent frequency instability.

The experimenter may note an interesting effect when working with transistor variable-frequency oscillators. The use of a common bias source for the collector and base-emitter junction will maintain a relatively constant ratio of the two voltages with varying supply potentials. The effect of these variations on frequency is partially canceled out as the change in one voltage is somewhat counteracted by the change in the other. The reason for this is that an increase in collector voltage causes an increase in frequency while an increase in emitter voltage causes a decrease in frequency. The experimenter will find that a particular value of emitter resistance and emitter current will cause variations in supply voltage to have a minimum effect on the oscillator frequency. The exact value can be found empirically by starting with an R_E value of 1,000 ohms. Experiment with the value of R_F while noting the exact amount of frequency shift for a given supply-voltage change. Increase R_E to 1,500 ohms and repeat the experiment. A *Micro Alloy Diffused Transistor* (*MADT*-Philco), for example, will show the best frequency stability with an R_E value of 3,300 ohms and approximately 1 to 2 ma of emitter current when operating from a 12-volt source.

The supply voltage changes the frequency of oscillation because of its effect on the transistor junction capacitance. It is logical to assume that by using high values of C across the inductance in the tuned circuit, small changes in junction capacitance will have less effect on the frequency of oscillation. Obviously a change of 1 mmf out of 1,000 mmf will have less effect on the frequency of the tuned circuit than will 1 mmf change out of 100 mmf. For this reason always use as much capacitance as possible in parallel with the tank coil, particularly in the Hartley circuit. Excessive capacitance will lower the tank impedance to the point where oscillations may cease. The exact amount of capacitance in the tank is a matter for experimentation. To get in the "ballpark," a good approximation is to divide the frequency of oscillation into the constant number 5,000. Thus at 5 mc, one uses 1,000 mmf of capacitance in the tank. The inductance value then must be juggled to make the circuit oscillate at 5 mc. If it does not oscillate with this much capacitance, the value can be reduced and the inductance increased. Conversely, if the circuit still oscillates strongly, try additional capacitance for even more stability. It might be pointed out that such high-C circuits tend to make the inductance more critical. This is quite true, but the overall stability with high-C circuits is better due to the masking of junction-capacitance variations caused by parameter changes.

For the same reasons, make the capacitive-divider values as large as possible (while maintaining a 10-to-1 ratio) in the Colpitts and Clapp circuits.

Temperature change causes some unique problems in transistor oscillator circuits. Changes in temperature have the same effect on components as in tube-type, variable-frequency oscillators. In tube equipment, however, the temperature of components is usually elevated above ambient because of the heat generated by the tubes. Thus, changes in ambient temperature have less effect on the tube oscillator. In transistor equipment the components usually operate at ambient temperature, and changes in ambient have an adverse effect on the frequency of oscillation. For this reason temperature compensation of tuned circuits may be required. For best stability always use good quality components, such as silver mica or dipped mylar capacitors and temperature-stable coil forms. Employ transistors which have a low value of C_c, such as mesa and *MADT* types. Silicon transistors are less affected by temperature than germanium and therefore are recommended for variable-frequency oscillator applications.

Crystal-Controlled Oscillators

In this type of frequency-stabilized oscillator a quartz crystal either replaces the tuned circuit or takes over its frequency-determining function. Quartz crystals are used because of their extremely high Q and excellent frequency stability over a given temperature range.

Fig. 2-16 shows the electrical equivalent circuit of the crystal. Capacitor C_S, inductance L, and resistance R in series

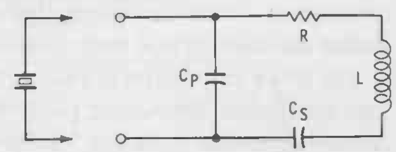


Fig. 2-16. Equivalent LCR circuit of a crystal.

represent the mechanical vibration characteristics of the crystal. These three components will exhibit a low impedance at one frequency, which is determined primarily by the physical dimensions of the quartz and which, in turn, determines its vibrating characteristics. This point is called the series-resonant frequency of the crystal.

The equivalent circuit also shows a parallel capacitance (C_P), representing the electrostatic capacitance between the crystal electrodes. These electrodes may be either pressure plates or plated silver discs in intimate contact with the quartz. The shunt or parallel capacitance C_P, in conjunction with the inductance of the crystal, forms what might be thought of as a second tuned circuit in the crystal. This is called the parallel-resonant frequency and is always slightly higher in frequency than series resonance. The difference between the two frequencies may be anywhere between a few hundred cycles and a few kilocycles, depending on the type of crystal. The difference is always very small compared to the actual frequency of either point. The impedance of the crystal at parallel resonance is extremely high.

Fig. 2-17 shows a curve of impedance versus frequency. It should be pointed out that the impedance of the crystal is also high at frequencies other than the resonant frequency. However, the *highest* impedance occurs exactly at parallel resonance while the *lowest* impedance occurs exactly at series

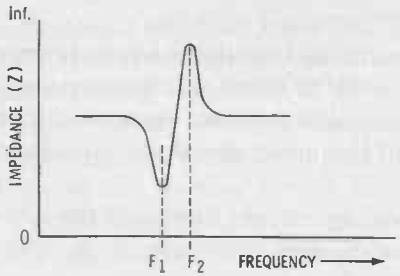


Fig. 2-17. Impedance-versus-frequency curve of a crystal.

resonance. Note the steep slope between the two points; this indicates that it is a very high-Q tuned circuit.

Fig. 2-18 shows the crystal-controlled version of the tickler-coil oscillator discussed previously. The crystal (Y) is connected in series with the feedback path and operates at series

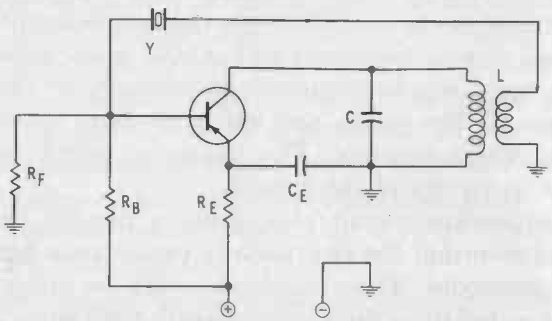


Fig. 2-18. Crystal-controlled version of a tickler-coil oscillator.

resonance. Since the crystal impedance is lowest at series resonance, it may be thought of as a switch. Above or below series resonance the impedance is so high that it attenuates the feedback energy below the point where the circuit can sustain oscillations.

Fig. 2-19 is the crystal-controlled equivalent of the Hartley. Capacitor C_B in Fig. 2-8 is replaced by the crystal in Fig. 2-19. Once again the crystal operates at series resonance as in the previous circuit.

Figs. 2-20 and 2-21 are the crystal-controlled equivalents of the Colpitts. As in the Hartley, the crystal replaces C_B and operates at series resonance.

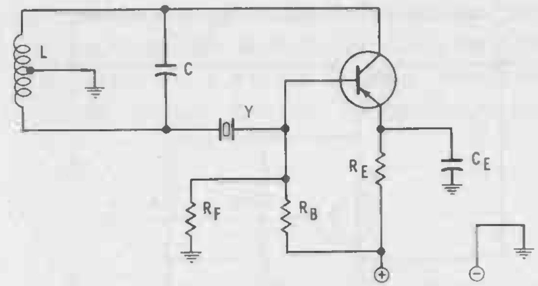


Fig. 2-19. Crystal-controlled equivalent of a Hartley oscillator.

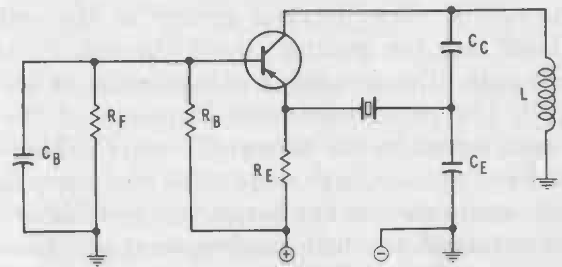


Fig. 2-20. Crystal-controlled version of the Colpitts oscillator.

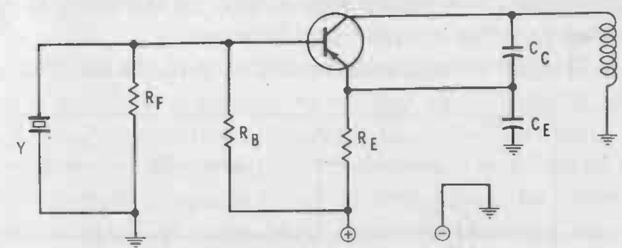


Fig. 2-21. Alternative version of Colpitts, crystal controlled.

Fig. 2-22 is somewhat different from the previous circuits in that the crystal replaces the tank circuit and operates at parallel resonance. Although this is another variation of the Colpitts, it is sometimes referred to as a Pierce oscillator. If the crystal were replaced by a parallel-tuned circuit, the configuration would be identical to the Colpitts discussed earlier.

Capacitors C_C and C_E form an impedance-matching capacitive divider in shunt with the equivalent parallel-tuned circuit

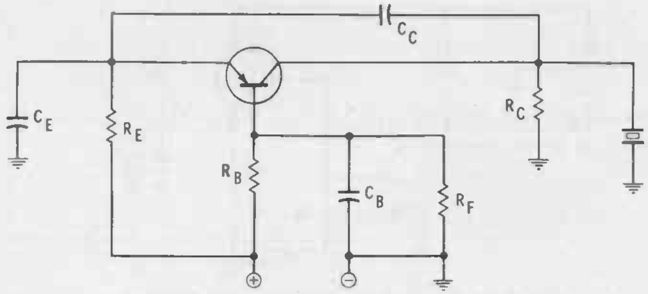


Fig. 2-22. Circuit of a Pierce oscillator.

inside the crystal. The in-phase energy at the collector is coupled back into the emitter circuit through C_C to create a feedback path. The frequency of oscillation is determined primarily by the parallel-resonant frequency of the crystal and to a lesser extent by the values of C_C and C_E. These capacitors should be made as large as possible to swamp the effect of junction capacitance on the frequency of oscillation. As the values are increased, however, the frequency of oscillation will be pulled away from parallel resonance lower in frequency. A point will be reached, as the frequency of oscillation approaches series resonance, where the phase shift in the crystal inductance is too great to support oscillations.

Fig. 2-23 is the common-emitter version of the oscillator just

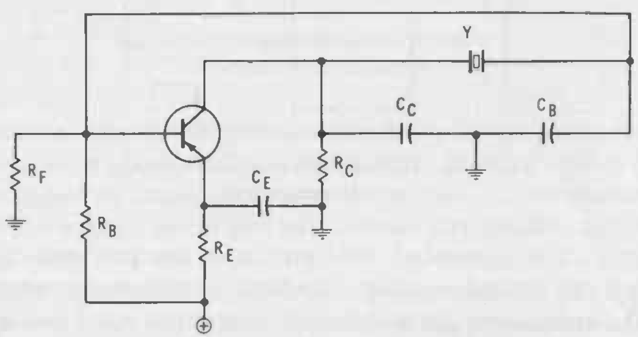


Fig. 2-23. Common-emitter Pierce oscillator.

described. In this circuit the crystal assumes a phase-inverting function the same as for a parallel-resonant L/C circuit. Radio-frequency energy coupled to the base is 180° out of phase

with respect to the collector signal. When the phase is again inverted by the transistor action, the stage oscillates.

Fig. 2-24 illustrates a rather unusual oscillator. At first glance it appears to lack most of the necessary requirements

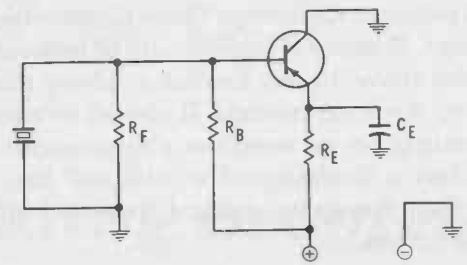


Fig. 2-24. Unusual transistor oscillator circuit.

for an oscillator circuit (tuned circuit, impedance-matching capacitors, etc.). If you ignore the ground connections for a moment, the theory of operation becomes somewhat easier to understand. The crystal operates at series resonance. Capacitor C_E is relatively small and somewhat critical in value (500 mmf at 5 mc). The base-emitter capacitance of the transistor, in conjunction with C_E, forms an impedance divider. Neglecting the grounds, C_E is in series with the crystal, and feedback occurs between emitter and base through the crystal. Although there is no phase inversion or voltage amplification, the transistor does have current gain. A transistor which has an f_t of 100 mc or more will oscillate quite vigorously. The effect of junction capacitance on frequency can be minimized by connecting a capacitor equal in value to C_E between the base and emitter connections of the transistor. Output RF may be obtained across C_E. A low-impedance output, quite independent of the oscillator section, may be obtained by lifting the collector from ground and connecting a 50- to 500-ohm resistor in series with this point. The output voltage will vary between a fraction of a volt and several volts, depending on the value selected.

Overtone Oscillators

Crystals have another characteristic which has not been mentioned. If the output tank circuit of Fig. 2-20 and 2-21, for example, is tuned to a harmonic of the crystal frequency, the circuit will oscillate at this frequency. If a 10-mc crystal were used in either circuit and the output tank tuned to 30 mc, the crystal would oscillate at the third harmonic. The crystal is excited to vibrate three times as fast. Most crystals are capable of establishing vibration modes at 3, 5, and 7 times their natural resonant frequency. These higher-order modes are called overtones. Because of the difficulty of fabricating crystals for frequencies above 15 mc, oscillators above this frequency usually employ overtone crystals. It should be mentioned that crystals operating on an overtone always oscillate at series resonance. Thus a fundamental crystal will have an output slightly less than the actual marked frequency multiplied by the order of overtone.

3

Building Oscillators

Now that the theory of transistor transmitters has been covered, let's take a look at the practical aspects and construct several circuits which illustrate these principles. Since the beginning of any transmitter is the oscillator, let's start there.

Several of the circuits which follow, such as the oscillators in this chapter, are built on *electronic pegboard*, manufactured by Vector and by Lafayette Radio. This method of construction allows simplified step-by-step assembly and results in the most reproduceable performance short of using an etched circuit board. Several of the projects in the following pages include circuit-board layouts, in case the experimenter prefers to use this method of construction.

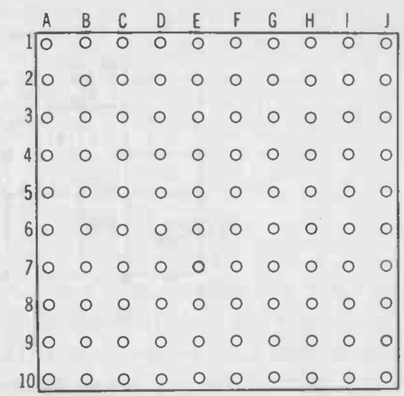


Fig. 3-1. Layout guide for construction projects.

The use of a pegboard chassis also provides an "electronic roadmap" which is standardized at 10 holes square. Exact positions on the "map" can be located by key letters and numbers, as shown in Fig. 3-1. For example, A-1, J-1, A-10, and J-10 are the four corner holes.

CHART 3-1. Recommended Transistor Types

Manufacturer	Type Numbers
Amperex	OC-170, 2N2672
Motorola	2N741
Philco	2N1727, 2N1745
RCA	2N370, 2N371, 2N372
	2N384
Texas Instruments	2N711

The circuits are designed so that the type of transistor used is not critical. Chart 3-1 shows several popular transistors that are used in many of the projects. Each circuit was optimized with the low-cost, high-gain Amperex 2N2672, then it was tested with the other types to verify operation.

BUILD A CRYSTAL CHECKER

Fig. 3-2 is the diagram of a crystal tester which not only can measure crystal activity but also can be used to check the frequency of oscillation. Any reasonably active crystal with a

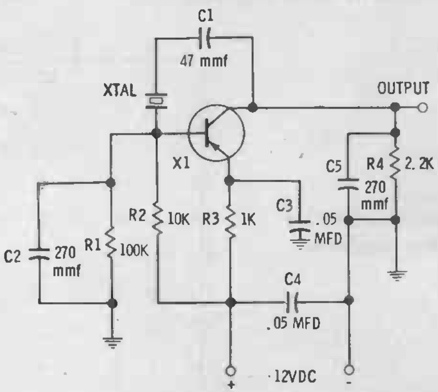


Fig. 3-2. Schematic of the crystal checker.

fundamental frequency between 100 kc and 15 mc will oscillate in the crystal checker. This includes those hard-to-get-going surplus crystals between 370 and 500 kilocycles used in World War II tank transmitters. The crystal checker can also be used as a 100-kc frequency standard for setting dial calibrations on communications receivers. It is only necessary to insert a suitable 100-kc crystal and replace C1 (47-mmf mica) with a 7- to 45-mmf rotary trimmer capacitor. This trimmer is used to zero-beat the crystal with WWV at 5 or 10 mc.

How It Works

The circuit is similar to the one described in Chapter 2, except for the addition of C1. This capacitor is added so that the crystal "sees" the standard calibration capacitance of 32 mmf. Thus the crystal oscillates at parallel resonance and the frequency can be determined by measurement with an accurate frequency meter or communications receiver.

Construction

Insert pegs at A-1 and J-1 and connect together with a piece of solid tinned wire (Fig. 3-3). This line becomes the positive bus. Insert pegs at A-10 and J-10 (Fig. 3-3) and connect to-

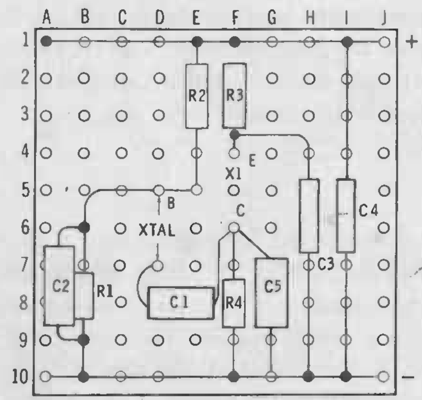


Fig. 3-3. Layout of the crystal checker.

gether as before. This is the negative bus. Install the pegs for the transistor at E-5 (base), F-6 (collector), and F-4 (emitter). The transistor can be soldered to these pegs or a socket can be used as shown in the accompanying photograph (Fig. 3-4).

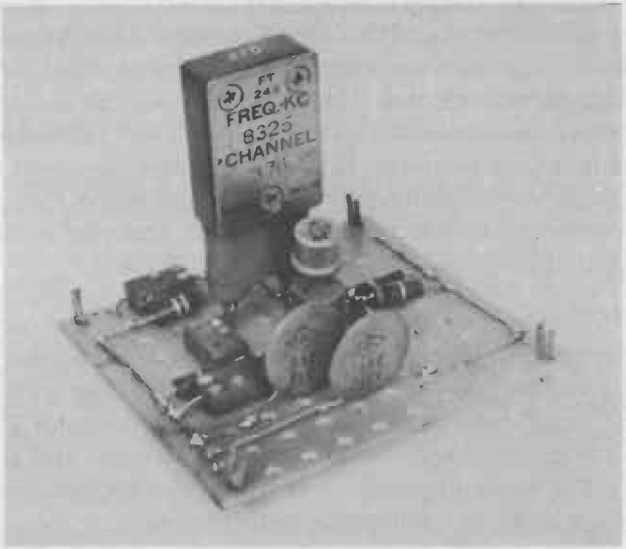


Fig. 3-4. Photo showing crystal socket.

The two remaining pegs are installed at D-5 and D-7 for the crystal socket. If the unit is mounted in a metal box, these points can be connected to a crystal socket on the front panel. A socket suitable for hermetically sealed crystals (HC-6 and CR-18/U style) can be connected in parallel with the large-pin socket for testing both types of crystals.

Testing

The circuit is energized by connecting a source of voltage (6 to 12 volts) in series with an SPST toggle switch and a 0 to 5-ma meter to terminals A-1 or J-1 (positive) and A-10 or J-10 (negative). With 12 volts applied, the meter will read approximately 2 ma with no crystal in the circuit. When the crystal is inserted, the current will increase. The activity of the crystal is roughly proportional to the increase in the meter reading. The meter is not required, of course, if the circuit is used as a signal generator or frequency standard.

If the oscillator is installed in a communications receiver, there is no need to use a battery power supply. A source of 6

to 12 volts, positive DC, is available at the cathode of the audio-output tube. The negative line should be connected to chassis ground. A lead from the collector of X1 should be wrapped around the receiver antenna terminal and coupled for proper signal level.

BUILD AN 80/40-METER PEANUT WHISTLE

It's truly amazing how far a fraction of a watt can be transmitted when it is radiated by an efficient antenna. The "peanut whistle" shown in Figs. 3-5 and 3-6 has been used to contact stations more than 100 miles away on the 40-meter band. Signal reports indicated that more distant contacts could have been made. The circuit can also be used for testing crystals on their fundamentals at frequencies between 2 and 15 mc.

How It Works

The peanut whistle is a practical use for the circuit shown in Fig. 2-24. However, in Fig. 3-5 a 330-ohm resistor has been

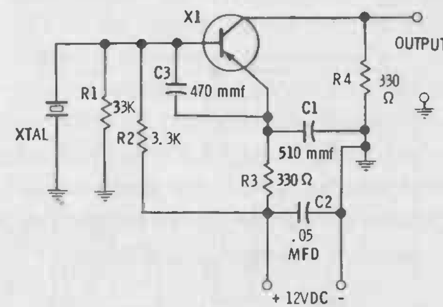


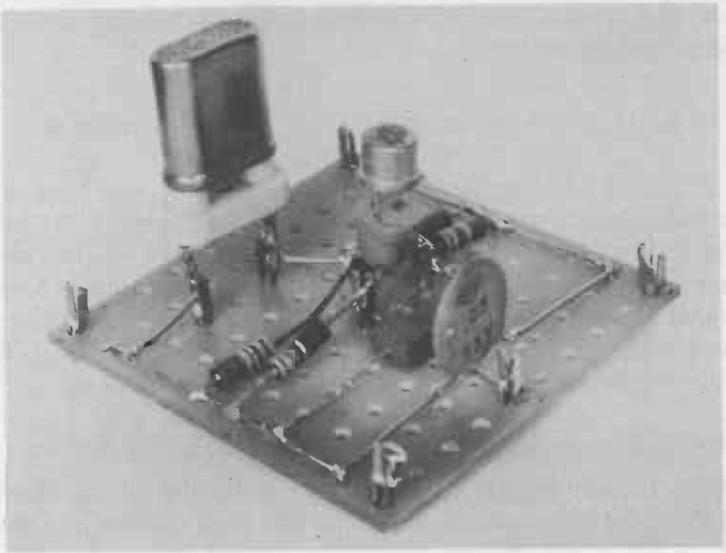
Fig. 3-5. The 80-40 meter "Peanut Whistle."

connected in series with the collector of X1. The signal when taken from this point is isolated from the oscillator section of the transistor.

Construction

The peanut whistle is built on a 2½" × 2½" board. The transistor is positioned the same as in Fig. 3-3, and pegs are placed in the four corner holes. The crystal is positioned at

C-6 and C-8. A peg for the output is inserted at J-6. The exact layout is shown in Fig. 3-6 and the accompanying photograph.



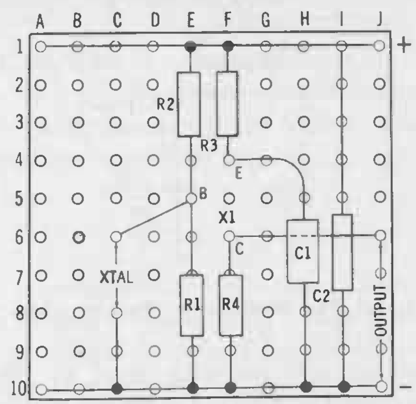


Fig. 3-6. Layout for the "Peanut Whistle."

Testing

The circuit can be used as a crystal oscillator, exactly as shown in Fig. 3-5. The output is quite strong, and there is no need to couple the signal into the receiver antenna terminal.

To use the peanut whistle as a transmitter, the output terminals are connected to the folded-dipole antenna (Fig. 3-7). The 330-ohm resistor (R4) should be disconnected to avoid absorbing transmitter power. The folded-dipole type of antenna, such as the one shown in Fig. 3-7, must be used to com-

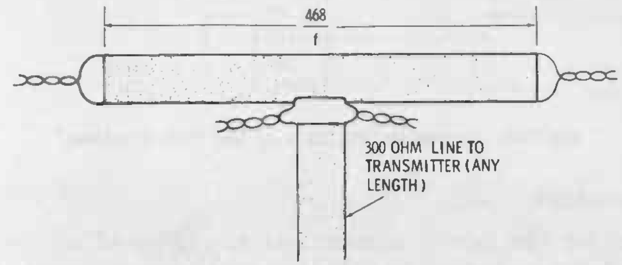


Fig. 3-7. Suggested antenna for the "Peanut Whistle."

plete the collector DC circuit. This antenna is constructed from 300-ohm transmission line of the type used for television reception. The length of the antenna section (in feet) can be accurately determined by dividing 468 by the frequency in megacycles. Thus, for the 40-meter novice band (7.2 mc) the antenna should be cut to 65 feet. The two ends of this piece are twisted together and soldered. One conductor is cut exactly at the center and attached to the transmission-line part of the antenna, as shown in Fig. 3-7. The other end of the transmission line is connected to the transmitter output terminals. A telegraph key in series with a 12-volt battery completes the circuit. Any of the transistors shown in Chart 3-1 can be used in this circuit.

BUILD THE CB CYCLONE

The one-transistor transmitter shown in Fig. 3-8 is capable of impressive performance. The power output is less than one-tenth watt, yet the transmitter is capable of transmitting signals a thousand miles or more when the skip is in on ten meters and the Citizens band. The unit may be used on the Citizens band without a license, provided the circuit is adjusted for no more than 100 milliwatts input (8 ma at 12 volts DC) and is connected to an antenna not more than 5′ long.

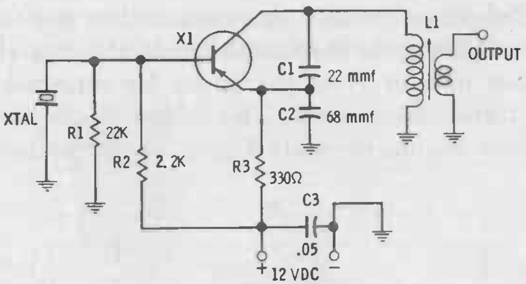


Fig. 3-8. Schematic diagram of the "CB Cyclone."

Construction

Pegs for the power connections are inserted at the four corner holes as before, but the transistor is positioned at E-4 (emitter), D-5 (base), and E-6 (collector). The crystal is inserted at C-6 and C-8. A guide to parts layout is shown in Fig. 3-9.

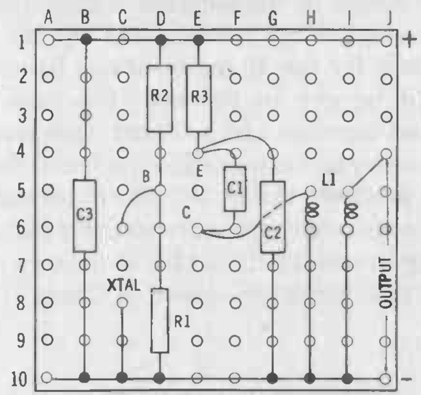


Fig. 3-9. Layout for the "CB Cyclone."

To insure duplicating the performance of the unit, the coil-winding details shown in Fig. 3-10 should be followed closely. The exact type of slug-tuned coil form is not important so long as the outside diameter is approximately 5/16″ in diameter. A printed-circuit type of coil form can be seen in Fig. 3-11. The circuit will work equally well on the Citizens band or 10 meters with the component values given and the coil shown. When

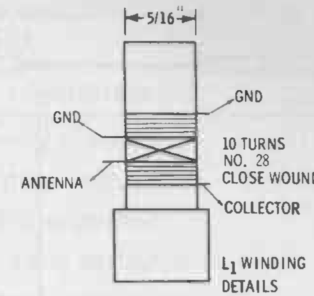


Fig. 3-10. Winding details for L1.

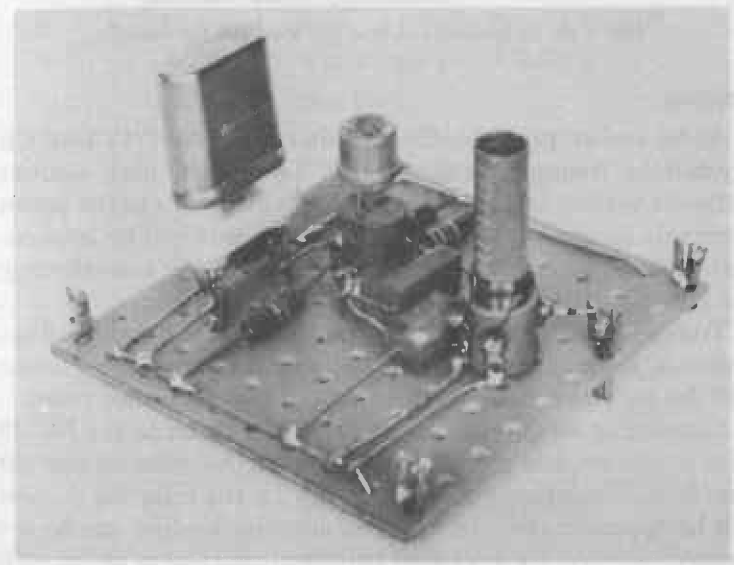


Fig. 3-11. Finished "CB Cyclone."

ordering a crystal for this unit, a third-overtone type for the frequency of operation should be specified.

A suitable antenna is shown in Fig. 3-12. The length depends on the frequency of operation and can be determined by dividing the frequency in megacycles into 468. The coaxial shield lead is connected to J-10, and the center conductor is connected to J-4.

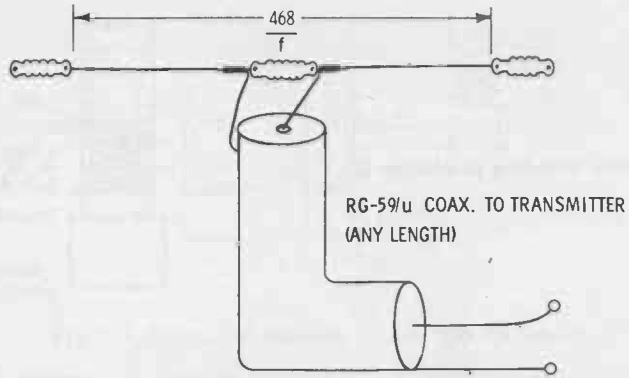


Fig. 3-12. Suggested antenna for the "CB Cyclone."

Testing

As in earlier projects, the circuit can be used to test CB crystals for frequency and activity. To test the unit, connect a 12-volt battery in series with a 0 to 10-ma meter to the power terminals on the board. The collector current will be approximately 3 ma with the crystal out. If the circuit is oscillating, the current will rise when the crystal is inserted.

Tuning is similar to that of a tube transmitter. As the coil approaches resonance by tuning the slug, the collector current will dip and then rise on either side of the resonance point.

Connecting an antenna, or dummy load (such as the No. 49 pilot lamp shown in the photo), will make the collector current rise. With the antenna shown in Fig. 3-12, the collector current will be approximately 10 ma. The antenna loading can be decreased by inserting a 50-mmf trimmer capacitor in series with the output lead and coaxial cable center conductor. The pilot lamp will load the transmitter more heavily than the antenna, causing it to draw more than 10 ma. When the transmitter is tuned properly, the bulb will glow brightly.

Although all the transistors shown in Chart 3-1 will oscillate in this circuit, only the Amperex 2N2672 is capable of the power output described. The 2N711 and 2N741, for example, put out only enough power to barely light the bulb. For this reason the 2N2672 is highly recommended.

A suitable modulator to permit voice transmission is described later in the book.

CRYSTAL CHECKER PARTS LIST

Quantity	Item No.	Description
1	C1	47-mmf mica capacitor.
2	C2, C5	270-mmf mica capacitor.
2	C3, C4	.05-mfd disc capacitor.
1	R1	100K, ½-watt, carbon resistor.
1	R2	10K, ½-watt, carbon resistor.
1	R3	1K, ½-watt, carbon resistor.
1	R4	2.2K, ½-watt, carbon resistor.
1	X1	Transistor (see Chart 3-1).
1	Xtal	See text.
1		Piece of "Vectorbord" (type A) 10 holes square.
9		Push-in terminals (Vector T-30 or equiv.).

PEANUT WHISTLE PARTS LIST

Quantity	Item No.	Description
1	C1	510-mmf mica capacitor.
1	C2	.05-mfd disc capacitor.
1	R1	33K, ½-watt, carbon resistor.
1	R2	3.3K, ½-watt, carbon resistor.
2	R3, R4	330-ohm, ½-watt, carbon resistor.
1	X1	Transistor (see text).
1	Xtal	See text.
1		Piece of "Vectorbord" (type A) 10 holes square.
9		Push-in terminals (Vector T-30 or equiv.).

CB CYCLONE PARTS LIST

Quantity	Item No.	Description
1	C1	22-mmf mica capacitor.
1	C2	68-mmf mica capacitor.
1	C3	.05-mfd disc capacitor.
1	L1	10 turns of No. 28 enameled wire closewound on 5/16"-diameter slug-tuned form. Link 3 turns same wire closewound over center of primary coil.
1	R1	22K, ½-watt carbon resistor.
1	R2	2.2K, ½-watt carbon resistor.
1	R3	330-ohm, ½-watt carbon resistor.
1	Xtal	See text.
1		Piece "Vectorbord" (type A) 10 holes square.
15		Push-in terminals (Vector T-30 or equiv.).

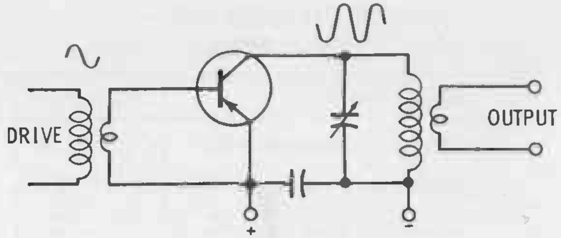
4

Power Amplifiers

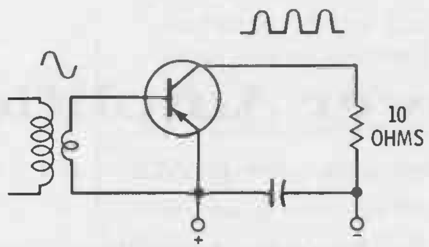
Although the low-power transmitters described in the previous chapter are capable of impressive range, it is considerably easier to make contacts with more power going into the antenna. This can be accomplished by adding a power-amplifier stage to the oscillator circuits described earlier.

FLYWHEEL ACTION

To work efficiently, the power amplifier must be properly matched in both the input and output circuits. The tuned-circuit Q must be high to insure a minimum of harmonic radiation. Fig. 4-1A is the circuit of a typical link-coupled, power-amplifier stage. The drive from the oscillator consists of a sine wave applied to the base-emitter junction of the power amplifier. Negative half cycles forward-bias the stage and cause a large pulse of collector current to flow. No current will flow during the positive half cycle, and the output, therefore, consists of a series of pulses resembling the output of a half-wave rectifier. This waveform is shown in Fig. 4-1B; it is obtained by replacing the tank circuit in Fig. 4-1A with a 10-ohm resistor. Alone, these pulses, although they are capable of conveying intelligence, are rather useless. They are rich in harmonics and should not be applied to an antenna where



(A) With inductance in output.



(B) With resistance in output.

Fig. 4-1. Typical RF power amplifier showing input and output RF waveforms.

the harmonics can be radiated or cause interference on unauthorized frequencies.

To minimize harmonic radiation, it is customary for the collector current to flow through a tank circuit (Fig. 4-1A), consisting of a high-Q coil and capacitor combination resonant at the signal frequency. Each time a pulse of current flows through the coil, it causes a ringing similar to a damped wave and has a duration of several cycles. Each pulse reinforces this damped wave, and the result is a continuous series of relatively pure sine waves that are free of harmonics. The action is similar to that which occurs in oscillator circuits (Chapter 2).

FREQUENCY MULTIPLICATION

The tank circuit can also be resonated at a multiple of the drive frequency to reinforce the production of harmonics. In this type of frequency multiplier, every second or third pulse (for second- or third-harmonic production) will cause the tank

circuit to *ring* at the harmonic frequency. This process can be repeated many times by cascading frequency multipliers to permit a low-frequency oscillator to control a much higher carrier frequency.

INTERSTAGE COUPLING

To be useful, the power output of one stage must be coupled to the input of a succeeding stage. The coupling system should introduce a minimum of loss since the power output is generally determined by the amount of driving power.

There are three principal ways in which two stages may be coupled together, through inductive or capacitive coupling, and through filter networks. Although inductive coupling is undoubtedly the most popular system, particularly for high-frequency equipment, the other two methods have advantages which the experimenter should be aware of.

Inductive coupling generally involves the use of a primary coil connected to the driving stage with a link coil wound around the primary and connected to the driven stage. Occasionally a secondary coil may be inductively coupled to the primary to minimize harmonics. In this case, the link should be wound over the secondary.

Link coupling is generally used for impedance matching. If the loaded Q of the tuned circuit is not important, the collector is usually attached to the primary, as in Fig. 4-1A. The impedance step-down ratio is easily computed from the formula:

$$N = \sqrt{\frac{Z_P}{Z_S}}$$

where,

N is the turns ratio,

Z_P is the primary impedance,

Z_S is the secondary impedance.

However, in transistor transmitters the impedances are sometimes difficult to determine, so the number of turns on the link winding and the degree of coupling are usually adjusted for maximum power output. A good starting point is 1 turn on the link winding for every 8 to 10 turns on the primary. Maximum coupling will occur when the link is near the center of the coil. Thus, by moving the link toward the center of the coil, while observing power output, it is easy to determine whether more or fewer turns are required on the link.

There is an amazing amount of power lost due to incomplete coupling in this type of tank circuit. Even though the link is adjusted for maximum power output, considerable driving power does not reach the driven stage due to coupling losses. For this reason, powdered-iron toroid coils are becoming very popular. A group of typical low-frequency toroids are shown in Fig. 4-2. This type of coil form is characterized by the fact

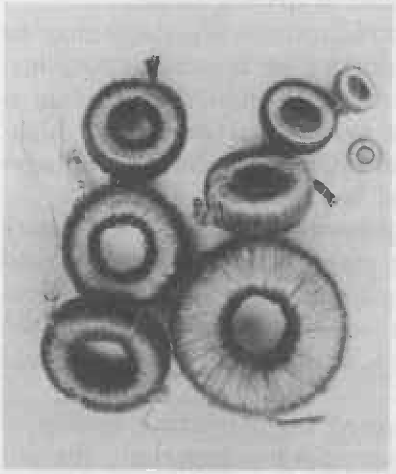
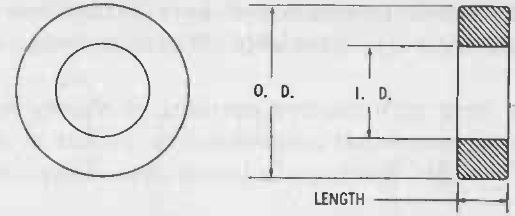


Fig. 4-2. Some toroid examples.

that virtually all the lines of force are contained in the doughnut core. Thus, a link wound around this core is likely to be cut by more lines of force than in an airwound coil. Further, it is possible to mount the coil very close to a metal chassis without adversely affecting the Q of the tuned circuit. Two companies selling powdered-iron toroid cores suitable for use in transmitters are Micrometals, 72 E. Montecito Avenue, Sierra Madre, California, and Radio Cores, Inc., 9540 Tulley Avenue, Oak Lawn, Illinois. A chart of popular toroids sold by Radio Cores is shown in Chart 4-1. The Radio Cores 57-1540 (1.2″ diameter) and 57-1541 (¾″ diameter) are particularly useful for frequencies between 500 kc and 10 mc. The material used in these toroids is called *Carbonyl E*. A variation of the same material, called *TH*, is excellent for 10 to 30 mc, while

Chart 4-1. Typical RF Toroids



DIMENSIONS			RADIO CORES, INC. MATERIALS AND PART NUMBERS				
O.D.	I.D.	LENGTH	10	11	12	21	22
.275 / .285	.152 / .162	.120 / .130	57-2336	57-2653	57-2651	57-2652	57-2486
.307 / .317	.152 / .162	.120 / .130	57-2655	57-2654	57-2656	57-2657	57-2658
.375 / .385	.200 / .210	.130 / .140	57-3372	57-3605	57-3596	57-3765	57-3766
.432 / .442	.245 / .255	.177 / .197	57-2002	57-1556	57-1542	57-1538	57-1553
.505 / .515	.307 / .317	.177 / .197	57-1540	57-1557	57-1543	57-1555	57-1554
.685 / .695	.370 / .380	.240 / .260	57-1753	57-1673	57-1677	57-1687	57-1686
.713 / .723	.280 / .290	.255 / .265	57-2339	57-2659	57-2427	57-2660	57-2661
.785 / .795	.495 / .505	.365 / .385	57-1541	57-1735	57-1736	57-1879	57-1878
1.052 / 1.072	.575 / .585	.430 / .450	57-2666	57-2667	57-2668	57-2669	57-2670
1.302 / 1.322	.775 / .785	.435 / .465	57-1826	57-2107	57-1516	57-1705	57-2662
1.990 / 2.010	1.240 / 1.260	.535 / .565	57-1817	57-2108	57-1465	57-2110	57-2663
For 2" Diameter Toroid	Relative Permeability		8.70	8.10	7.7	16.2	37.0
	L in mh for 1000T#30 PE		12.3	11.5	11.0	23.0	52.5
COMPARATIVE MATERIAL COST			MED	HIGH	HIGHEST	HIGH	LOWEST
COLOR CODE			Violet	Gray	Orange	Brown	None
FINISH			BLACK PHENOLIC RESIN (.005" Max. Increase Over Dimensions Shown in Table)				

Courtesy Radio Cores, Inc.

SF is optimum for 30 to 50 mc, and *J* is useful up to 140 mc or more. Another material, called *IRN-9*, is used between 50 and 200 mc and is somewhat less expensive than the various *Carbonyl* mixes. The forms are not expensive and usually cost from 10 to 25 cents, depending on size and material. There is, however, a minimum charge per type, so it is wise to standardize on a particular size and mix.

Because these toroid forms are generally unavailable, except direct from the manufacturer, and because they can seldom be found in wholesale radio supply stores, the author has used airwound coils exclusively in the projects to follow. However, the use of toroids is mentioned here as an experimenter's option, since their use invariably increases power output and efficiency.

Another form of inductive coupling is shown in Fig. 4-3. This circuit is somewhat more complex in that it requires an RF choke for a DC return path in the base of the driven stage.

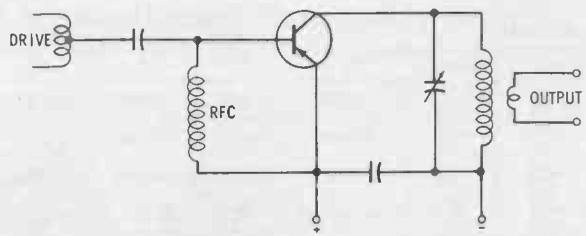


Fig. 4-3. Power amplifier with a tapped coil for input impedance matching.

This circuit has the advantage of coupling more energy to the driven stage than link coupling without the use of toroids. The formula for calculating the impedance ratio is the same as before; however, Z_S now comprises the impedance between the tap point and the bottom of the coil. This points up a major disadvantage of the circuit. The tap must be adjusted for maximum power output. With coils having closely spaced turns of insulated wire, the problems in attaching a tap are obvious. When airwound coils are used, such as *B & W* or *Air Dux*, turns can easily be shorted with solder, which drastically reduces the Q of the coil and the amount of power coupled to the following stage. Often a shorted turn can only be detected by testing the coil with a Q meter.

The coupling system favored by the author is the capacitive voltage divider shown in Fig. 4-4. The capacitive divider can be extremely useful, and the principle should be thoroughly understood by the experimenter. Capacitors C1 and C2 in Fig. 4-4 are in series and form the capacitive portion of the resonant circuit. However, they also act to form an artificial or imaginary tap on the coil. If, for example, the capacitors were equal in value, the effect would be exactly the same as placing a tap at the center of the coil. If C2 were 10 times the

capacitance of C1, the impedance transformation would be the same as for a coil tapped at one-tenth the turns up from the bottom of the coil. With either this tap, or a capacitive

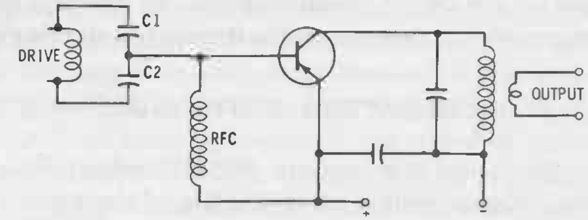


Fig. 4-4. Power amplifier with capacitive divider for input impedance matching.

divider with a 1-to-10 ratio, the circuit would exhibit a 100-to-1 stepdown in impedance. It is obvious that by juggling the values of the two capacitors, while maintaining resonance, any desired impedance ratio may be obtained. By making C1 and C2 variable, it is possible to obtain a continuously variable impedance system much the same as a roller coil with a sliding contact.

For some reason, the capacitive divider coupling system also results in a reduction of 10 to 20 db in the amplitude of harmonics transferred from the driver to the driven stage, or antenna in the case of a final tank circuit. Experimental evidence, in many applications, repeatedly indicates this to be the case. If an antenna is coupled to the junction of C1 and C2 in a final tank circuit, capacitor C1 acts as the tuning capacitor, while the action of C2 resembles that of a loading capacitor in a pi-network.

The low-pass filter or pi-coupling system shown in Fig. 4-5 can actually be evolved from the capacitive divider just de-

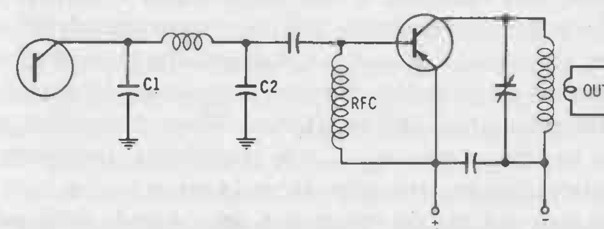


Fig. 4-5. Low-pass filter coupling system.

scribed. Here the junction of C1 and C2 is grounded rather than the junction of C2 and the coil. However, as mentioned earlier, ground is an arbitrary point and can be moved around to suit particular circuit considerations. As before, the value of the two capacitors determines the impedance stepdown ratio.

COLLECTOR MATCHING

Up to this point the various circuit configurations have shown the collector connected to the top of the tuned circuit. This may be satisfactory in low-power stages (e.g., less than 1 watt) which do not draw much collector current. However, for high-power stages the collector impedance is so low that it may represent only a few ohms shunting the tank. This reduces the Q of the coil to an unusable level, prevents proper flywheel action, and causes a drastic loss of coupling. The capacitive-to-inductive ratio can be increased to raise the Q, but this results in rather enormous and unwieldy capacitor values. In such cases it is customary to tap the transistor down on the tank circuit to some low-impedance point (Fig. 4-6)

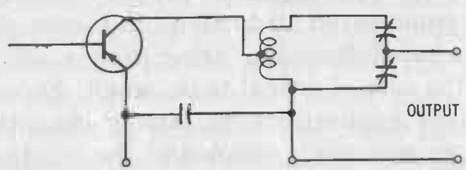


Fig. 4-6. Tapped coil for output impedance matching.

or employ some other impedance-matching scheme. For purposes of preliminary experimenting the collector impedance can be approximated by determining the DC resistance using Ohm's law. For example, if the stage draws 1 ampere at 12 volts, the collector resistance will be approximately 12 ohms. However, in class C this load is not across the tank at all times, and in actual practice the collector impedance is roughly two times the calculated DC resistance. When the collector impedance has been estimated, the impedance transformation can be calculated by using the formula given earlier.

If the taps are placed too far up on the coil, the loaded Q will be severely reduced and proper flywheel action will be

impossible. If the taps are too far down on the coil, the losses will increase due to the very high voltages induced at the top of the coil. Once again, the best position can be determined by adjusting the tap position for maximum power output. An approximation which is handy for determining the initial position of the taps is to assume an impedance across the coil of 10,000 ohms. The collector tap can then be placed at the correct number of turns to transform the collector impedance to this higher value. The link or capacitive divider can then be arranged to transform the 10,000-ohm impedance down to the input impedance of the driven stage.

While on the subject of power amplifiers and the problems of coupling them to an antenna, the "sloppy-tuned circuit" technique should be mentioned. Tapping the collector of a power amplifier down on the final tank coil results in certain problems which are not obvious without the aid of a high-speed oscilloscope. During cut-off periods between drive cycles, considerable voltage will appear on the collector due to the back emf caused by the collapsing field of the coil. This potential will generally be more than twice the supply voltage. A preferable circuit is shown in Fig. 4-7. Although it appears to have

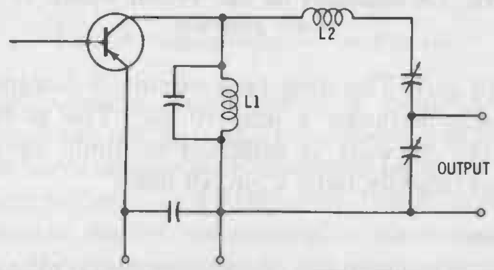


Fig. 4-7. "Sloppy-tuned circuit" for output impedance matching.

evolved from Fig. 4-6, the operating conditions are quite different. Coil L1 actually forms a resonant circuit in conjunction with the collector capacitance of the transistor. This circuit is parallel resonant at the signal frequency. Coil L2 forms another resonant circuit at the signal frequency in conjunction with the two output capacitors. These two tuned circuits provide more harmonic rejection than the one-coil circuit shown in Fig. 4-6.

For greater efficiency, which is obtained with higher Q, it is customary to use an external capacitor in parallel with L1. A typical circuit, for example a 20-watt marine transmitter at 2 mc, employs a .002 mica in parallel with a 15-turn coil close wound on a .05-inch diameter form.

WORK DX WITH THE PEANUT WHISTLE II

The two-transistor version of the Peanut Whistle (Fig. 4-8) uses the oscillator section described in Chapter 3. An Amperex 2N2672 is added to increase the power level about

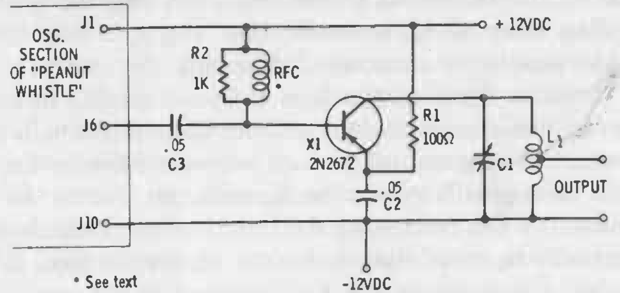


Fig. 4-8. Schematic for the "Peanut Whistle II" power amplifier.

10 times (10 db). The final tank circuit is designed to tune all frequencies between 3 and 10 mc. The power output, approximately .25 watt, is sufficient to dimly light a No. 47 pilot lamp or brightly light a No. 49 lamp.

Construction

The entire transmitter is constructed on an aluminum plate measuring 3" × 6". A 2½" cardboard piece is cemented to one end to insulate the pegs in the oscillator section from shorting to the chassis. Connections are made to the power amplifier stage by means of a three-lug terminal strip. The 365-mmf tuning capacitor is bolted to the end of the metal strip. Be sure to use short screws in the threaded tuning-capacitor mounting holes to prevent shorting the stator plates. The coil is mounted to the rear of the tuning capacitor. The form, measuring 1" in diameter, was salvaged from the junk box and was complete with threaded spade lugs for chassis mounting.

If you use paper or plastic tubing, without lugs, fabricate two L brackets to secure the coil to the chassis. Plastic sleeving, stripped from hook-up wire, is used to insulate the transistor leads. The full lead lengths are used. The collector lead is soldered to the lug on the tuning capacitor, while the base and emitter are connected to the terminal strip. A suggested layout is shown in Fig. 4-9.

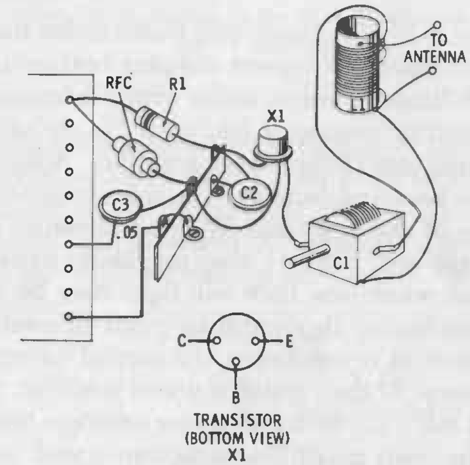
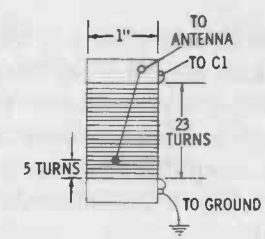


Fig. 4-9. Suggested layout for Fig. 4-8.

The radio-frequency choke (RFC) is "homebrew" since its characteristics are not critical. It is constructed by scramble-winding 50 turns of fine wire (any size between No. 30 and No. 40 is satisfactory) on a 1,000-ohm, 1-watt carbon resistor. The two ends of the coil are soldered to the resistor leads. Be sure to glue the turns so they do not come unscrambled.

Winding details of the final tank coil (L1) are shown in Fig. 4-10. Like the choke, this coil is not critical. It consists of 23

Fig. 4-10. Winding details for L1 in Fig. 4-8.



turns of No. 22 wire wound on a 1" form. The space between each turn is about the same as the wire diameter. Output is taken from a tap placed 5 turns up from the ground end. This tap is wired to a terminal lug on the coil form and connects to the center conductor of the coaxial transmission line. The shield connects to chassis ground.

Testing

Connect peg A-10 of the oscillator board to the minus terminal of a 12-volt battery. Connect the plus battery terminal to A-1 on the oscillator board in series with a telegraph key. To test the transmitter connect a No. 47 or No. 49 pilot lamp to the output terminals of the power amplifier. Apply power by depressing the key, and tune the variable capacitor for maximum brillance of the test bulb. With an 80-meter crystal installed, the bulb will light at near maximum capacitance. A second position where the bulb will light may be noted near minimum capacitance. Be careful to avoid this setting on 80 meters, however, as it represents the second harmonic of the crystal frequency. If the crystal is above 3,650 kc, the second harmonic will fall outside the 40-meter amateur band. Always remember, tune near maximum capacitance with an 80-meter crystal and near minimum capacitance with a 40-meter crystal.

A suitable antenna was described in Chapter 3. This antenna would be 126.5' long for the 3.7-mc novice band and 65.5' long for the 40-meter novice band. The transmitter will tune up at approximately the same position as for maximum bulb brilliance. However, to get maximum power output you should tune the final amplifier for a maximum reading on your station-receiver S meter or a field-strength meter.

BUILD THE CB CYCLONE II

The circuit shown in Fig. 4-11 is a power amplifier for the oscillator described in Chapter 3. It employs a low-cost silicon transistor and increases the power of the oscillator approximately six times to over .25-watt output.

The amplifier operates in the common-emitter configuration. The collector is connected to a tap on the coil which is series tuned.

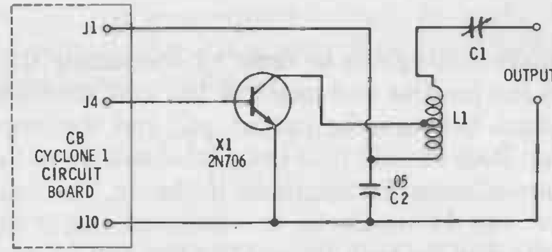


Fig. 4-11. Schematic for the "CB Cyclone II" power amplifier.

Construction

The method of construction is similar to previous projects. Pegs are mounted in the four corner holes and wired together for ground and power connections (Fig. 4-12). The transistor is mounted at C5 (collector), C7 (emitter), and D-6 (base). Note that the wire connecting the base of the transistor to

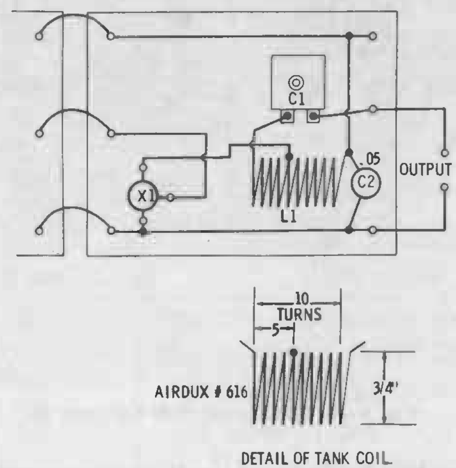


Fig. 4-12. Layout for the "CB Cyclone II."

the output of the oscillator board passes between the collector and emitter pins. A ¼" hole was drilled at G-3 to mount the trimmer capacitor used to resonate the final tuned circuit. A peg for the output connector was installed at J-5. The coil is attached between the tuning capacitor C1 and the bypass capacitor C2, and is supported by its own leads. The finished transmitter is shown in Fig. 4-13.

Testing

Preliminary testing can be done by connecting the 12-volt battery to the positive and negative bus and attaching a No. 47 pilot lamp between the output peg and the negative or ground line. Both the coil in the oscillator and C1 in the power amplifier are adjusted for maximum brilliance. The final amplifier current can be measured to determine power input, by breaking the wire between the positive bus and C2 and inserting a milliammeter. The final tunes just like a tube transmitter; that is, it will dip at resonance and the current in the dip will increase as the final is loaded heavier into an antenna. With

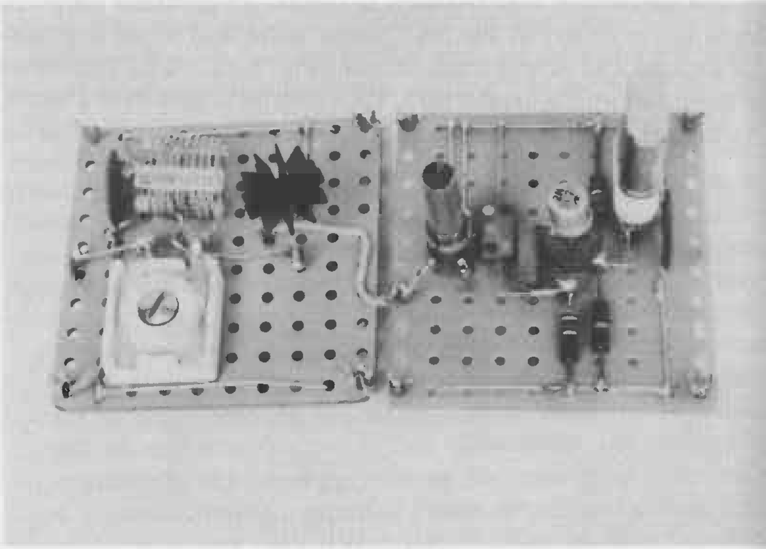


Fig. 4-13. Completed "CB Cyclone II."

normal tuning the final current will be approximately 50 ma. Loading on an antenna is the same as for a No. 47 pilot lamp. An antenna such as the one shown in Fig. 4-11 should be used. CW operation can be accomplished by using a telegraph key in series with the battery. In this manner, both the oscillator and the power amplifier are keyed. The CW note is extremely good.

As in the Cyclone I, the circuits are designed to tune both the 10-meter band and the 11-meter Citizens band. When

onnected to the recommended dipole antenna, this transmitter is capable of working more than 1,000 miles with the ght propagation conditions.

PEANUT WHISTLE II PARTS LIST

Quantity	Item No.	Description
1	C1	365-mmf variable capacitor (J. W. Miller No. 2111).
2	C2, C3	.05-mfd disc capacitor.
1	L1	23 turns No. 24 enameled wire wound on 1″ form. Spaced diameter of wire. Tap at 5 turns (see text and Fig. 4-10).
1	R1	100-ohm, ½ watt, carbon resistor.
1	R2	1K, 1 watt, carbon resistor (see
1	RFC	text).
1	X1	2N2672 *(Amperex)*
1		Piece of sheet aluminum, 3″ × 6″.

CB CYCLONE II PARTS LIST

Quantity	Item No.	Description
1	C1	10-70 mmf variable trimmer capacitor (*ARCO* 302).
1	C2	.05-mfd disc capacitor.
1	L1	10 turns of No. 24 (*Airdux* No. 616), spaced diameter of wire, ¾″ diameter, tap at 5 turns.
1	X1	2N706
1		Piece of "Vectorbord" (type A) 10 holes square.
10		Push-in terminals (Vector T-30 or equiv.).

5

The Novice Powerhouse

Most of the transmitters described so far have used very low power. The transmitter to be described in this chapter runs 5 watts input and is capable of the same performance as a novice transmitter employing a 6V6-type tube. It is capable of transmitting for hundreds or even thousands of miles when properly tuned to a correct antenna. Even if the antenna is not exactly the right length or if the transmitter is slightly mistuned, there is still sufficient reserve power to compensate for some of the loss. The transmitter operates on both the 80- and 40-meter novice bands and has a tuning range of 3 to 8 megacycles. Later a voice modulator will be described for this transmitter so that you can work AM.

HOW IT WORKS

The transmitter circuit is the type generally referred to as a master-oscillator power amplifier (MOPA). It consists of a crystal-controlled oscillator link-coupled to a power-amplifier stage. The output of the power amplifier, in turn, is connected to the transmitting antenna. A circuit is also included so that relative power output can be observed on a meter.

The circuit for the high-power novice transistor transmitter is shown in Fig. 5-1. Transistor X1, a *PADT*-50 (Amperex), is

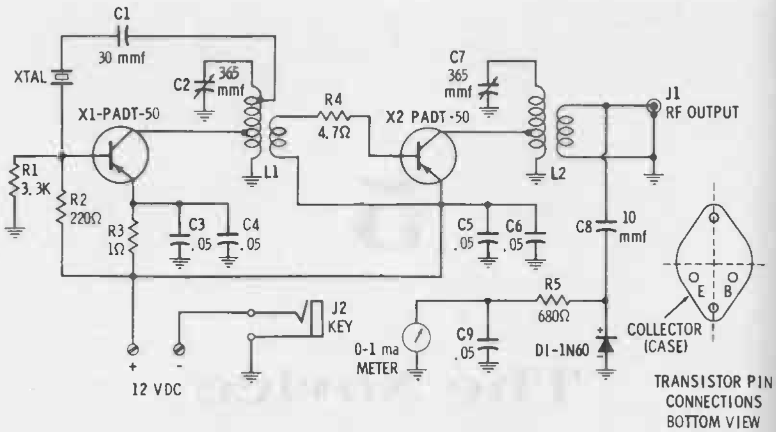


Fig. 5-1. Schematic of the "Novice Powerhouse" transmitter.

connected in the common-emitter configuration and serves as the crystal oscillator. The collector of the oscillator is connected to a low-impedance point on coil L1. This coil also serves as an impedance step-up transformer to provide feedback (through the crystal) to the base of the oscillator stage. Resistors R1 and R2 forward-bias the oscillator, while R3 provides DC degeneration and prevents thermal runaway. Capacitors C3 and C4 bypass this point so that it is not degenerative to the oscillator signal. Capacitor C1 is 30 mmf (±10%) and determines the exact frequency of oscillation. The value of C1 is chosen so that the crystal "sees" a 32-mmf load capacitance and the crystal oscillates at exact parallel resonance. Coil L1 is resonated with a 365-mmf variable capacitor; it tunes the circuit between 3 and 8 megacycles.

Output from the oscillator is link-coupled to power-amplifier stage X2, a second *PADT*-50. Resistor R4 (4.7 ohms) is included to limit the base current of the stage and prevent overdriving. No emitter stabilization is required in this stage since the transistor is operating in deep class C. In other words, if the crystal is removed or C2 is detuned (so that the oscillator no longer functions), X2 will draw no collector current. It only conducts on a portion of the negative half cycle. Capacitors C5 and C6 bypass the stage to prevent any RF feedback from reaching the oscillator circuit. The collector of this stage is also tapped at a low-impedance point on coil L2, which is

resonated at the signal frequency by means of C7, a second 365-mmf capacitor. RF power output from the amplifier is link-coupled to the transmission line through connector J1. Although an RCA-type phono connector is used, a coaxial connector may be substituted, if desired.

The RF voltmeter is a relative-output indicator and works in the following manner. Capacitor C8 (10 mmf) samples a small amount of the RF appearing across the link and transmission line. This radio-frequency energy is rectified by D1, a 1N60 general-purpose germanium diode. A filter consisting of R5 (680 ohms) and C9 (a .05-mfd disc capacitor) removes the RF component and leaves only DC. This DC is applied to a 0- to 1-ma meter which provides a relative indication of the amount of voltage appearing across the link and transmission line. Capacitor C8 determines the exact reading of the meter. With the values shown, the meter will read almost full scale on 40 meters. If it reads more or less than full scale, the value of C8 can be varied to suit the constructor. Capacitor C9 is not critical and may be any value between .001 mfd and .05 mfd.

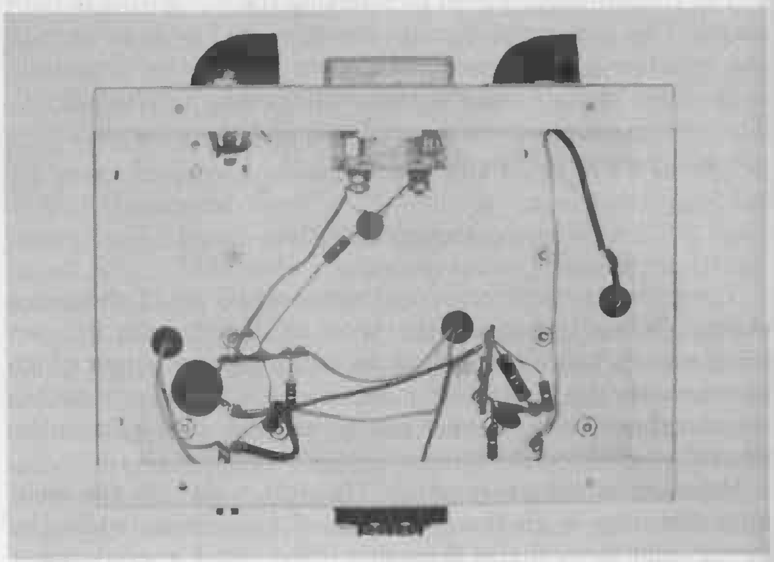
The transmitter is keyed by breaking the B+ circuit to both stages. This is accomplished by inserting the key in series with the negative battery terminal. When the key is depressed, both stages operate and the transmitter emits a carrier signal. The keying note is excellent and even detuning the oscillator will not cause excessive chirp.

CONSTRUCTION

The entire transmitter is constructed on a 5″ × 7″ aluminum chassis. When laying out the holes on the chassis, it is an excellent idea to make the drawing on the paper wrapper which accompanies the chassis. It minimizes scratching or marking the chassis and the wrapper can be retained to duplicate the project at a later date.

Referring to the photograph (Fig. 5-2), note that the oscillator circuitry is on the right side of the chassis while the power amplifier occupies the space at the left. A crystal socket is mounted on the front panel for quick changes of frequency. A meter found in the author's junk box was used for the RF voltmeter. This meter has a basic 0- to 1-ma movement and is

(A) Top-front view.



(B) Wiring view.

Fig. 5-2. Complete Novice Powerhouse transmitter.

not critical. Any similar meter with this range may be used. The key jack is mounted on the left side of the front apron directly below the power-amplifier tuning adjustment. Although an open-circuit key jack was used in this project, a closed-circuit variety can also be used. When testing and tuning up, it is only necessary to remove the key plug from the jack to energize the transmitter.

A layout for the transmitter is shown in Fig. 5-3. When laying out the holes for the transistor, the job can be made

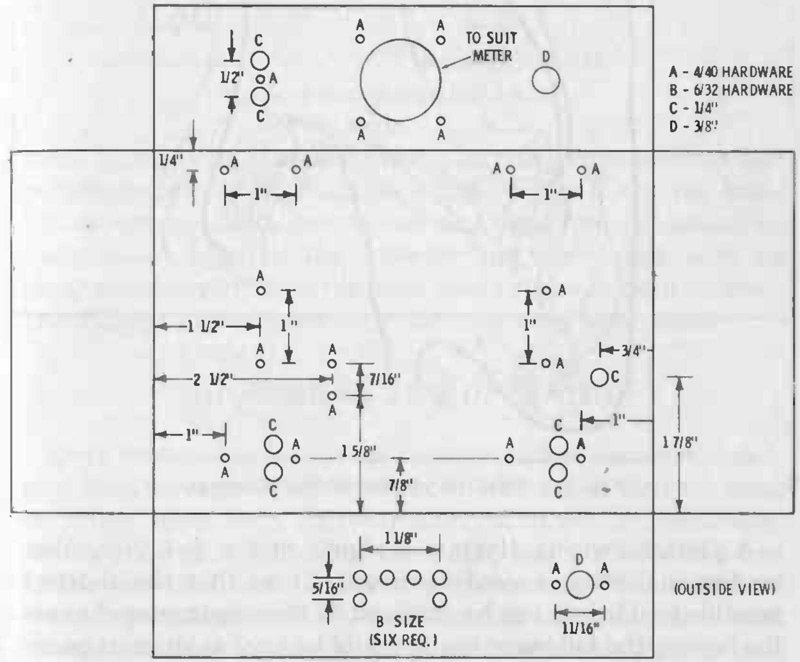


Fig. 5-3. Chassis layout for Novice Powerhouse transmitter.

easier by positioning the mica washer in the proper area and tracing the holes with a pencil. A transistor mounting kit may make mounting the *PADT*-50's easier. Solder lugs are used under the head of the transistor mounting screws to connect the collector to the coil in each stage. Although a terminal-strip type of power connector is used on the rear apron, any suitable battery connector may be used at the constructor's option. Below the chassis you will note that two terminal strips are used as tie points. These strips are positioned to permit the shortest possible lead length in critical circuits and the layout should be followed closely.

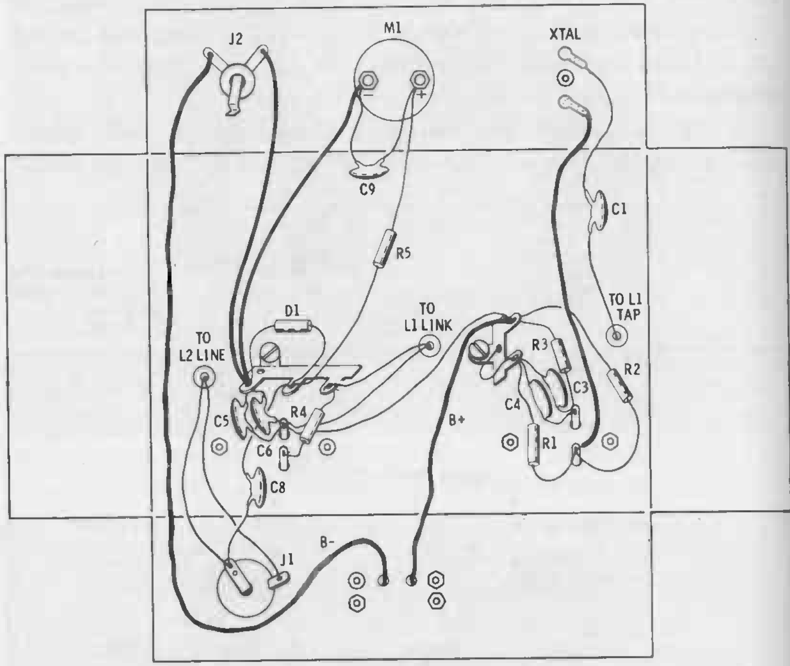


Fig. 5-4. Pictorial chassis-wiring diagram.

A pictorial wiring diagram is shown in Fig. 5-4. Note that no terminal strip is used to mount C1 so that the shortest possible lead length can be obtained. If the constructor changes the layout, the following leads should be kept as short as possible: the wire between C1 and the crystal, the length of leads associated with R4, the bypass capacitors, and the lead lengths between the tops of the coils and the associated tuning capacitors (C2 and C7).

Coils L1 and L2 were found in the junk box and consist of 25 turns of No. 22 bare wire spaced approximately the same diameter as the wire. A piece of *AIRDUX* 816 coil material may be substituted for these coils if desired. Taps are placed on the coils by heating the wire carefully to avoid loosening the turns and butt-soldering a short length of wire at the tap

point. The position of the links on L1 and L2 is somewhat critical; the drawing shown in Fig. 5-5 should be followed closely for initial settings. Note that both power RF transistors

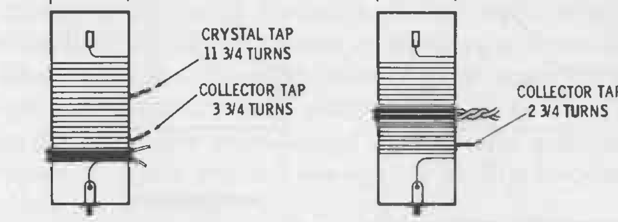


Fig. 5-5. Coil-winding detail.

must be insulated from the chassis by using mica washers and petroleum jelly or, if available, silicon grease. Once the transistors are mounted, but before any wires are connected to them, check between the collector and the chassis with an ohmmeter to verify that the mica washer has not been shorted. It is difficult to check this once the circuit has been wired.

ADJUSTMENT AND OPERATION

Once the construction of the transmitter has been completed, give it a thorough visual inspection to ascertain that no wires or points have been short-circuited. Connect an ohmmeter between the plus and minus terminals, insert a telegraph key in the key jack, and close the key contacts. The ohmmeter should read approximately 100 ohms in one direction. Now reverse the two ohmmeter leads and the meter should read about 2,000 ohms. If these approximate readings are obtained, they indicate that there are no short circuits in the transmitter that will discharge the battery or destroy the transistors.

Construct a dummy-load bank as shown in Fig. 5-6. This dummy load will represent a load of approximately 50 ohms to the transmitter, and the bulbs will light to about half brilliance when the connector is inserted in the transmitter antenna connector. Once the ohmmeter test has been made, you may connect a battery to the power connector on the rear apron. With no crystal inserted, there should be no reading on the

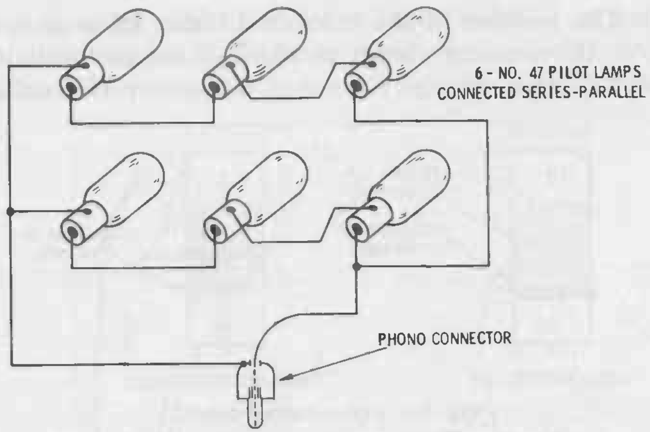


Fig. 5-6. Dummy-load bank for transmitter testing.

transmitter meter. Insert a crystal for the 40-meter CW band and set both capacitors near minimum capacitance (clockwise). Now turn the oscillator adjustment counterclockwise very slowly. If the circuit is working properly, the meter will now indicate slightly. Now slowly turn the power amplifier adjustment (C7) counterclockwise. The meter reading should increase as you turn the knob and the lamps should start to glow. Turn the oscillator adjustment for maximum brilliance of the bulbs or maximum reading on the meter. Once again turn the amplifier adjustment for maximum brilliance of the bulbs and the highest meter reading obtainable. Repeat these adjustments, back and forth, until absolute maximum power output has been obtained.

If the bulbs do not seem to be particularly bright, try moving the oscillator link up or down slightly and retuning the oscillator and amplifier for maximum power. If the power output decreases, move the link in the opposite direction. Each time the link is adjusted it will be necessary to recheck the setting of C2 and C7. Once the maximum power output has been obtained by varying the oscillator link, adjust the amplifier link for maximum power in the same manner. You may find that more power output can be obtained by reversing the two leads on the links. When you have completed these adjustments the bulbs should be about half brilliance and the meter should read near full scale. If it reads beyond full scale, reduce the

size of C8. You can listen to the CW note in your station receiver while keying the transmitter. It should sound clean with no chirps or clicks.

To finish testing the transmitter, insert an 80-meter, novice-band crystal in the socket. Retune both the oscillator and final power-amplifier adjustments near maximum capacitance position. The bulbs should again light to the same brilliance, although the meter may not read as high as before. This is because the capacitive reactance of C8 increases with decreasing frequency. Again recheck the keying on this frequency to verify that the note is pure.

Now connect your novice "powerhouse" to an antenna. For best results an antenna such as the one shown in Fig. 3-12 should be used. Insert a suitable crystal and peak both the oscillator and power-amplifier adjustments for maximum meter reading. You may note that the meter reads higher or lower when connected to the antenna. This is because the impedance of the antenna may be more or less than the bulbs. Once again it may be necessary to vary the value of C8 to provide a usable meter reading. Once a value has been selected for C8, it need not be changed unless a different antenna is employed. You will not need to adjust this transmitter with a field-strength meter since the RF-voltmeter circuit indicates maximum power output the same as a field-strength meter. You may find, however, that moving the link on the power-amplifier coil (L2) will produce slightly more output than it did on the dummy load.

That completes the adjustment, and you are now ready to work DX with the novice transmitter. As in earlier projects, you are cautioned not to tune the transmitter to its second harmonic when operating on 80 meters. Always remember that the variable capacitor should be set near maximum capacitance with an 80-meter crystal and near minimum capacitance with a 40-meter crystal.

Most of the parts for this transmitter are readily available; however, special mention should be made of the *PADT*-50 transistors. These are made by Amperex and there are no substitute types. These transistors are available from any radio store handling Amperex devices or may be obtained from various mail-order distributors. If you have difficulty obtaining the transistors, the name of the nearest source may be obtained by writing Amperex Semiconductor Products, 230 Duffy Avenue, Hicksville, L.I., N.Y. The constructor may have some difficulty, also, in locating resistors R3 and R4. Values smaller than 10 ohms usually are stocked only in the larger radio stores.

NOVICE POWERHOUSE
PARTS LIST

Quantity	Item No.	Description
1	C1	30-mmf, 10% dipped-mylar or silver-mica capacitor.
2	C2, C7	365-mmf maximum, variable capacitor (J. W. Miller No. 2111 or equiv.).
5	C3, C4, C5, C6, C9	.05-mfd disc capacitor.
1	C8	10-mmf, 10% dipped-mylar or silver-mica capacitor.
1	D1	IN60 germanium diode.
1	J1	RCA-style phono connector.
1	J2	Open- or closed-circuit key jack.
1	L1	25 turns No. 22 bare or enameled wire wound on 1" diameter form. Spacing between turns same as wire diameter. Collector tapped at 3¾ turns, crystal tapped at 11¾ turns. Link, two turns No. 24 plastic-covered hookup wire wound over bottom 3 turns of coil.
1	L2	Same as L1 except collector is tapped at 2¾ turns. Link is two turns wound over center of coil.

NOVICE POWERHOUSE
PARTS LIST (CONT.)

Quantity	Item No.	Description
1	M1	0 to 1-ma meter.
2	X1, X2	*PADT*-50, RF power transistor *(Amperex)*.
1	R1	3.3K, ½-watt, 10% carbon resistor.
1	R2	220-ohm, ½-watt, 10% carbon resistor.
1	R3	1-ohm, ½-watt, 10% carbon resistor (see text).
1	R4	4.7-ohm, ½-watt, 10% carbon resistor. (see text).
1	R5	680-ohm, ½-watt, 10% resistor.
1		Chassis 5" × 7" × 2" (aluminum).
1		crystal socket.
3		¼-inch rubber grommets.
1		Two-screw terminal strip.
2		Knobs.
2		Mica washers, and transistor mounting hardware.
		Hardware and wire as required.

6

Modulation

By definition, modulation is the process of adding intelligence to a carrier signal. Modulation can be simple, as in the case of continuous wave (CW), that is, making or breaking the carrier to form dots and dashes. It can also be extremely complex, such as for time-division multiplex or video-modulation schemes.

The carrier can be modified by voice modulation in two principal ways. One system, used for hi-fidelity broadcasting and mobile radiotelephones, is frequency or phase modulation. The other method, known as amplitude modulation, is the subject of this chapter.

FREQUENCY MODULATION

Although frequency-modulation projects are not included in this book, the reader should be familiar with the technique. The heart of an FM transmitter is the frequency-modulated oscillator stage. Fig. 6-1 shows a typical FM oscillator stage. Modulation is applied to the oscillator in such a manner that the frequency of oscillation varies in proportion to the applied signal. For example, as a positive audio peak is applied to the oscillator, the frequency might move higher. During negative peaks the frequency would shift lower than the average frequency of oscillation. As the modulation waveform crosses its zero axis, the frequency of oscillation is the same as if no modu-

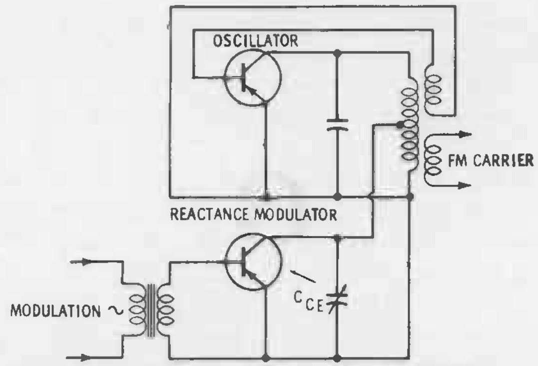


Fig. 6-1. A basic reactance-modulated oscillator circuit. Bias and voltage omitted for simplicity.

lation were applied. The "distance" the frequency changes is in proportion to the modulation amplitude, while the rate of change is determined by the modulation frequency.

The FM oscillator is usually followed by a limiter to remove any traces of amplitude modulation, as shown in the block diagram of Fig. 6-2. Although only one multiplier is shown, the

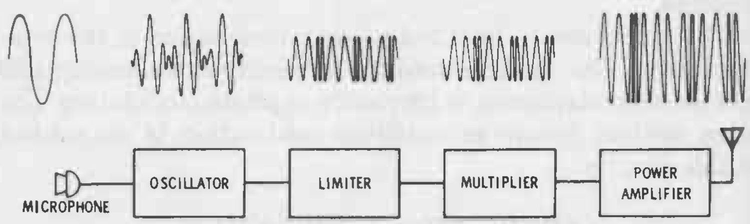


Fig. 6-2. Waveforms found in a typical FM transmitter.

limiter may be followed by several doublers or triplers to increase the oscillator output to the desired signal frequency. Each time the frequency is doubled, the modulation bandwidth will also increase by a factor of two.

In the receiving system the FM signal is again passed through a series of limiters, this time to remove impulse noise, static, ignition pulses, etc. Finally the signal is applied to a discriminator; this circuit has the characteristic of generating positive or negative voltages for signals above or below the average frequency of oscillation. Thus the output of the dis-

criminator will be the signal which originally modulated the transmitter.

AMPLITUDE MODULATION

As the name implies, this system varies the amplitude of the carrier wave rather than the frequency. This is generally accomplished in the final or power-amplifier stage in an amplitude-modulated transmitter. Amplitude modulation can be introduced by varying the gain of the amplifier as shown in Fig. 6-3. T_a is a modulation transformer inserting modulation

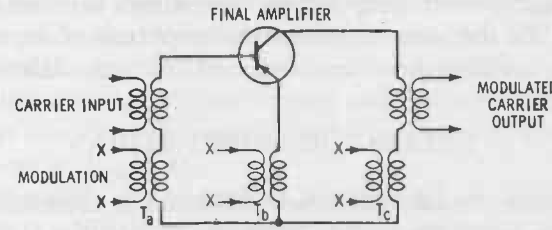
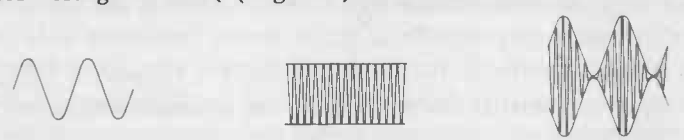


Fig. 6-3. Basic method of modulating an AM transistor transmitter. Only one of the three transformer connections would be used.

into the base circuit, while T_b injects the signal into the emitter. Either of these systems varies the gain of the transistor and causes the amplitude of the carrier to change in direct proportion to the modulation waveform.

Although either connection requires only a small amount of modulating power, both types of modulation are inefficient and make transmitter tuning (for best modulation) very critical. The most satisfactory modulation system is to vary the collector voltage as in T_c (Fig. 6-3).



(A) Modulation waveform. *(B) Carrier waveform.* *(C) Modulated carrier.*

Fig. 6-4. The amplitude of the modulation waveform increases or decreases the carrier amplitude.

Fig. 6-4 shows the result of modulating a carrier wave with a sine wave. These waveforms are what would be observed on an oscilloscope if measurements were made in a typical transmitter employing one of the three modulating methods just described. The drawing in Fig. 6-4A is a 1,000-cycle waveform that has been amplified by the modulator stage. Fig. 6-4B is the carrier signal applied to the final amplifier stage. Actually it also consists of a series of sine waves, as in Fig. 6-4A, but because of the high frequency, they show up as a solid bar on the screen of the average oscilloscope. Fig. 6-4C shows the modulated carrier wave. When the modulation and carrier are combined in correct proportions, the valleys produce zero RF output while the peaks are twice the amplitude of the unmodulated carrier. This condition is termed 100% modulation.

COLLECTOR MODULATION

Amplitude modulation is accomplished in transistor transmitters in much the same manner as in vacuum-tube equipment; however, it is easier with vacuum-tube equipment than with transistors, as you will soon see.

Since high-level collector modulation is by far the most popular system, let's examine its behavior in detail. Fig. 6-5

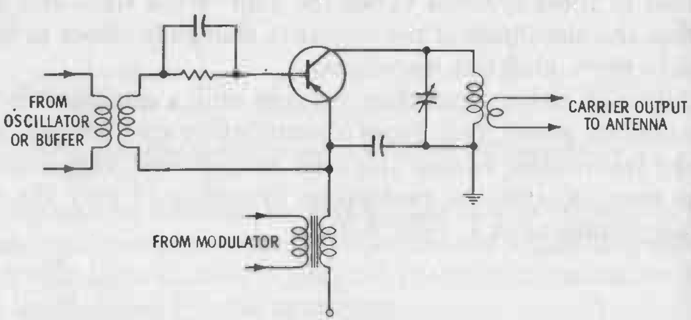


Fig. 6-5. Output circuit of a typical AM transistor transmitter.

is the simplified circuit of a typical collector-modulated transistor final amplifier. The point marked "AM carrier output" connects to the transmission line and radiating antenna. Negative 12 volts is applied to the collector of the power amplifier

in series with the secondary of the modulation transformer and the radio-frequency tuned circuit. The negative-going peaks add to the supply voltage while the positive-going half of the modulation waveform opposes the supply. It might be thought of as two batteries connected in series. When they are connected positive to negative, they add; but when they are connected positive to positive, they oppose each other.

Under conditions of 100% modulation the peak negative-modulation voltage will equal the supply, and the collector voltage will be −24 volts rather than −12 volts. By the same token, one-half cycle later the positive peak will exactly cancel the supply, and the instantaneous collector voltage will be zero. Doubling the collector voltage causes the RF peaks to reach twice the amplitude of the unmodulated carrier. One-half cycle later the transmitter output will be zero. This corresponds to the peaks and valleys in the modulated carrier waveform shown in Fig. 6-4C.

Unfortunately the preceding description, while technically accurate, is a highly utopian explanation as far as transistors are concerned. Although the theory of modulation is the same for tube and transistor transmitters, the results are not the same. One major problem is the built-in "flat-topping" of the modulated waveform usually found in transistor transmitters. The condition becomes much more noticeable as the power input is increased. Two additional and equally vexing problems are drive requirements and carrier leakthrough.

Flat-topping of the modulated waveform is a direct result of the inability of the transistor to handle high input powers. During the peak power-output period the instantaneous voltage is approximately doubled, as explained earlier. Ideally the collector current also doubles so that the peak power input is four times the average value. In a transistor transmitter, however, this increased power input is always accompanied by a decrease in h_{fe}, the gain of the transistor. At the same time the modulation waveform is trying to produce more power, the falling h_{fe} is opposing that power increase. The saturation resistance of the transistor is also causing more voltage to be lost across the junction, since it is in series with the supply and modulation. The result is a built-in compression of the modulation peaks, which prevents the stage from being modulated

100%. Fig. 6-6 shows some of the more realistic waveforms found in transistor transmitters. Fig. 6-6A shows the typical

(A) Typical modulation envelope of a vacuum-tube final amplifier.

(B) Modulation envelope of a transistor final amplifier.

Fig. 6-6. Comparison of modulation waveforms.

output of a vacuum-tube final amplifier, while Fig. 6-6B illustrates the likely waveform of a transistor final amplifier that is not capable of handling the peak powers involved.

The drive requirement of a tube and transistor final is also quite different. Unlike a tube final, the power output of a transistor final is largely determined by the drive available. Even though the stage is RF biased into class C, the increase in power input due to modulation may make it drop into class-B operation. This lack of drive during peak periods tends to compound the peak compression mentioned earlier. If the compression is severe, the oscilloscope may show what appears to be a 100%-modulated waveform (slightly distorted) and yet a dummy-load lamp will show downward modulation. Be-

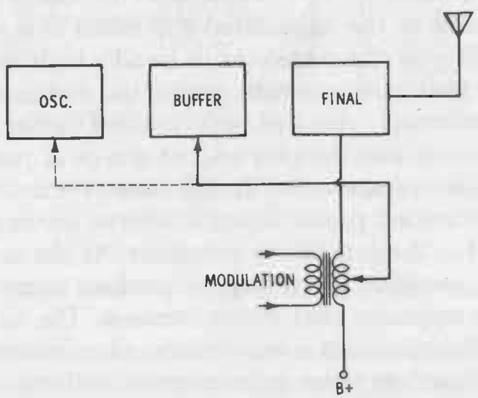


Fig. 6-7. Preferred method of modulating both driving and final stages.

cause of this problem, it is usually mandatory that the buffer stage driving the final, or sometimes even the oscillator, also be modulated. Fig. 6-7 shows one system of accomplishing this. The tap of the secondary of the modulation transformer, and therefore the amplitude fed to the buffer and/or oscillator, is determined by the power involved. Generally this tap is experimentally adjusted so that the driving stages are never quite cut off. If excessive modulation is applied to the driving stage or stages, negative speech clipping will occur and severe splatter will result. A "trick circuit" to minimize this effect is

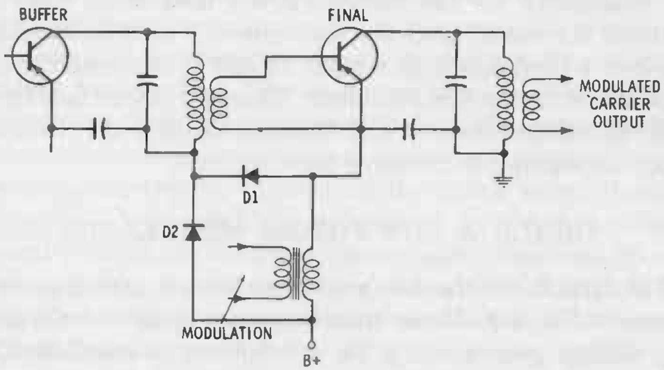


Fig. 6-8. "Trick circuit" for increasing positive peak modulation.

shown in Fig. 6-8. The final is collector-modulated with the modulating voltage connected in series with the positive power-supply lead. During positive peaks the modulation adds to the supply voltage. Positive peaks are also coupled to the buffer stage through diode D1 to increase the drive during peak periods; diode D2 is cut off during this period. During negative-peak periods, splatter is prevented since diode D1 no longer conducts audio to the buffer, and the stage is clamped to B+ through conducting diode D2. The result is a noticeable increase in upward modulation with no change in the negative or valley part of the waveform.

The problem of leakthrough can also be very annoying. During the period when the modulation supposedly cancels the supply voltage, there will still be a noticeable amount of power output. This is quite unlike vacuum tube RF power amplifiers.

The reason for this effect is the junction capacitance. With the base-emitter junction forward-biased and zero volts on the collector, the collector-base junction exhibits considerable capacitance. Thus the RF signal appearing on the base is capacitively coupled to the output tank circuit. The result is shown in Fig. 6-6B. Note that this waveform is the result of the problem just discussed and is typical of those found in transistor transmitters.

Now that some of the pitfalls have been examined, let's construct several modulators for use in conjunction with the transmitters described in this book. The low-power modulator forms the foundation for the medium-power modulator (500 milliwatts or 0.5 watts) and the high-power 5-watt modulator. It employs a Darlington-pair input circuit to increase the input impedance to over 400,000 ohms. Thus the circuit can be used with crystal and ceramic microphones which must "look into" a high impedance to preserve bass response.

BUILD A LOW-POWER MODULATOR

The circuit for the low-power modulator just described is shown in Fig. 6-9. Three transistors are used to build up the tiny voltage generated by the microphone to a sufficient level to be used as a modulator for the oscillator projects described earlier.

Transistors X1 and X2 form a Darlington pair to raise the input impedance and have no voltage gain. The driver tran-

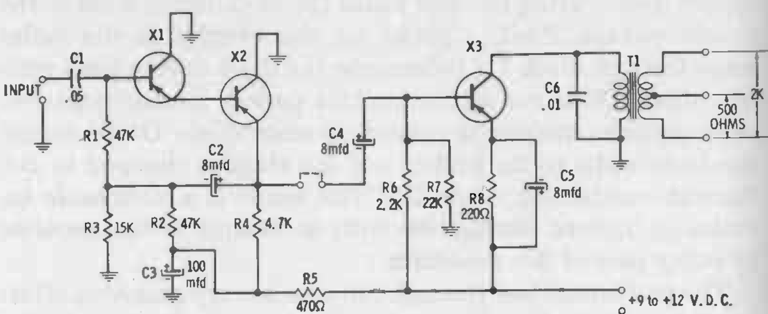


Fig. 6-9. Schematic diagram for the high-impedance microphone preamplifier and modulator.

sistor (X3) exhibits considerable power gain. Approximately 1 volt rms undistorted output will develop across a 470-ohm resistor connected to half of the T1 secondary.

The circuit works in the following manner: transistor X1 acts as an emitter follower. Forward-bias is developed at the junction of R2 and R3. Resistor R1 isolates the bias circuit from the input circuit. Transistor X2 is a second emitter follower; the input impedance of the emitter follower is determined by its output impedance (the emitter load resistor plus the input impedance of the driven stage) multiplied by the beta of the transistor. Thus when two emitter followers are cascaded, the normally low input impedance is raised to several hundred thousand ohms. In addition, capacitor C2 provides a small amount of degenerative current feedback to increase the impedance even further. The audio voltage developed across R4 is approximately the same amplitude as that generated by the microphone, but across a much lower impedance.

The audio voltage is coupled to the driver stage through capacitor C4. The driver (X3) operates in the common-emitter mode and exhibits considerable voltage gain. Resistors R6 and R7 form a voltage divider to provide forward-bias for X3, while R8 stabilizes the gain and current over wide temperature and voltage variations.

The amplified voltage is developed across the primary of T1 (10K impedance). Capacitor C6 tunes the primary to attenuate high audio frequencies. If a standard 10K to 2K center-tapped transistor interstage transformer is used, the output impedance is either 500 ohms or 2,000 ohms, as shown in Fig. 6-9. This transformer is suitable for modulating the oscillators described earlier, or it can be used to drive the half-watt modulator which follows.

Construction

The entire circuit is constructed on a circuit board measuring 2″ × 4″. This board was supplied by the W. H. Paulin Company, Box 122, Upland, California. However, if it is more convenient, electronic "peg board" can be used with the construction style employed earlier. There is no need to use transistor sockets, since the circuit is not critical as to transistor type. Virtually any PNP general-purpose audio transistor can be employed. The circuit can be installed in an LMB box or incorporated as part of the larger modulators which follow.

The constructor should carefully observe the electrolytic polarity. For example, if capacitor C4 is reversed, the error can damage output transistor X3.

If it is desired, a volume control can be included in the circuit. Resistance R4 would be replaced by the two end connections of a 5,000-ohm audio-taper volume control. The dotted connection in Fig. 6-9 would be broken and the negative end of C4 would be connected to the arm or center terminal of the volume control.

The circuit has been designed to amplify only the voice frequencies and produce communications quality. If higher fidelity performance is desired, capacitor C6 should be deleted to increase the high audio frequencies, while replacing C5 with a 100-mfd electrolytic will extend the low-frequency response. A slight amount of bass boost can be obtained by connecting the 100-mfd capacitor between the emitter end of C5 and ground.

If the circuit is wired as shown in Fig. 6-9, it should work immediately without any further experimentation. Connect a

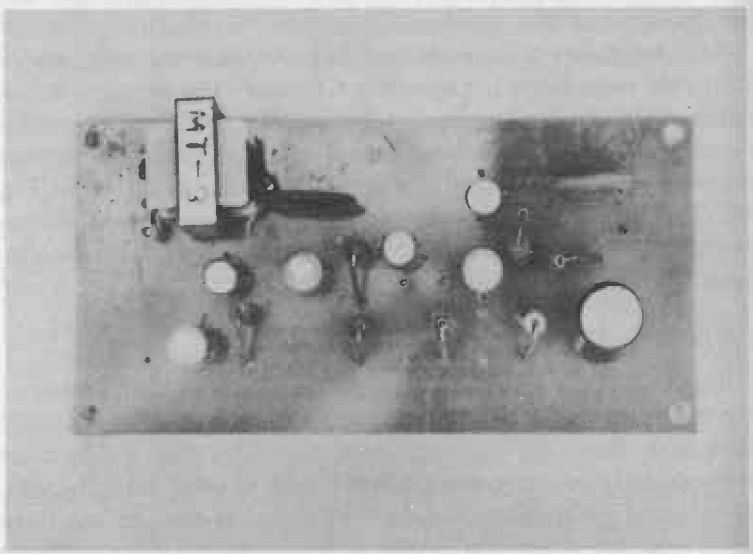


Fig. 6-10. Parts location on modulator board.

microphone (ceramic, high-impedance dynamic, or crystal) to the input, apply power, and connect a pair of headphones to the 2,000-ohm output connection. If the unit is working properly, you should hear your voice clearly with no distortion. Approximately 2 to 3 volts rms can be measured across the headphones or a 2.2K resistor connected to the amplifier output. Two views of the finished unit are shown in Figs. 6-10 and 6-11.

Fig. 6-11. Completed low-power modulator.

BUILD A HALF-WATT MODULATOR

The project previously described delivers only a few milliwatts of audio and therefore is not capable of 100% modulation of any circuit but a very low-power oscillator. It does, however, make an excellent driver for this circuit to be described.

How It Works

The circuit shown in Fig. 6-12 is known as a class-B push-pull modulator. Unlike the class-A driver (X3 in Fig. 6-9), the transistors in class-B circuits only amplify on alternate half cycles. Collector current pulses alternate in each half of the

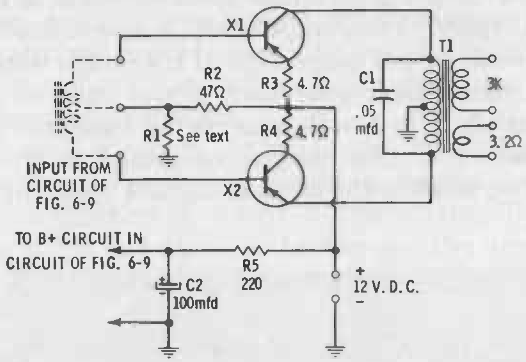


Fig. 6-12. Schematic of the half-watt modulator.

primary of T1. The half-cycle waveforms generated by each transistor are combined in the secondary to recreate an amplified replica of the driving signal.

The output of the driver signal becomes a push-pull signal by connecting the center tap of the driver transformer to the B supply. When the audio signal is negative to the base of X1, conduction occurs. At the same time the signal on the base of X2 is positive-going; thus X1 conducts while X2 is cut off. During the next half-cycle period the polarity reverses; then X2 conducts while X1 is cut off.

If the transistor characteristic were completely linear from cutoff to saturation, the waveform shown in Fig. 6-13A would

(A) Output waveform from properly biased stage.

(B) Output waveform with insufficient crosswave bias.

Fig. 6-13. Waveforms showing crossover-distortion effect on a normal sine wave.

represent both the input and output waveform of the modulator. Unfortunately this is not the case. All transistors have a "hook" in their curve at the low-current end. This "hook" must be overcome by applying forward bias to the base-emitter junction. If this is not observed or if the amount of forward

bias is insufficient, the output waveform will resemble that shown in Fig. 6-13B. This effect is known as crossover distortion and is very annoying to the ear. The sound is similar to that created by a speaker with a rubbing voice coil. Even a slight amount of distortion at this point on the waveform is immediately obvious to the ear, even though it can hardly be observed with an oscilloscope. Crossover distortion can be minimized by incorporating a forward-bias network. This is the purpose of R1 and R2 in Fig. 6-12. Resistor R1 will be approximately 6.8K for the components shown. It should be adjusted for a collector current of 3 ma or until the output signal has the least amount of audible distortion with minimum no-signal collector current. The transistors draw more current on signal peaks and the collector current will vary between 3 ma no signal and 30 ma on signal peaks.

Resistors R3 and R4 are included in the circuit to prevent thermal runaway. They also supply a small amount of degenerative feedback to increase the input impedance of the stage and reduce distortion. A filter network consisting of R5 and C2 is included to prevent feedback through the B+ circuit to the driver stage.

Transistor T1 is a special transistor type used in nine-transistor Citizens band walkie-talkies; it has three windings. The primary impedance is 500 ohms center tapped, while the modulation winding is 3,000 ohms and the speaker winding is 3.2 ohms.

Construction

The modulator is constructed on a small LMB chassis box and is shown in the accompanying photograph (Fig. 6-14). The circuit board containing the preamplifier and driver circuitry is mounted inside the box on the rear apron wall. The volume control, which must be used to prevent overdriving the modulator, is mounted at the front apron along with the microphone jack. The two class-B transistors are mounted by drilling holes in the top of the box which are the same size as the outside diameter of the transistor package. This is done so the transistor makes a tight press fit into the holes and permits the chassis to act as a heat sink. If other types of transistors are used, clips may be employed to ther-

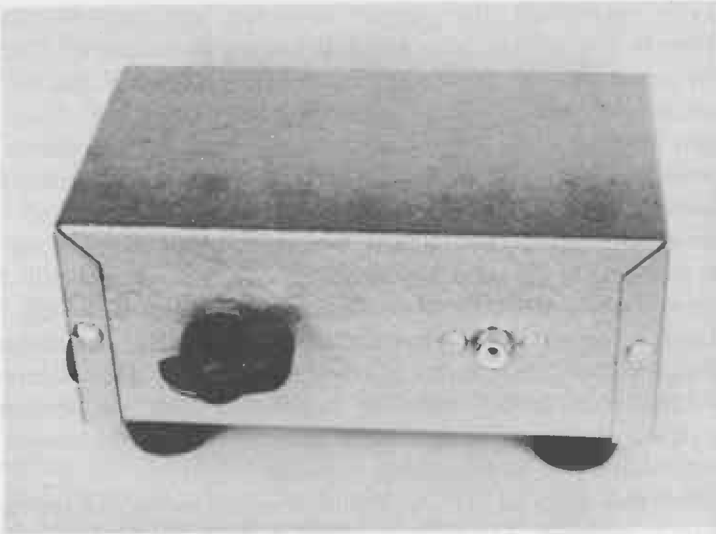


Fig 6-14. Top view of finished half-watt modulator.

mally connect the transistor case to the chassis. The transistor must be mounted to a heat sink to increase the dissipation and prevent destruction or thermal runaway at high audio signal levels.

Transformer T1 is mounted by drilling two small holes in the chassis and bending the mounting ears over. It is a good idea to solder these ears together with a short piece of wire to secure the transformer. The output wires are attached to a four-lug terminal strip mounted near one end of the box.

Testing

Once the construction of the half-watt modulator has been completed, it should be tested in the following manner. Connect a microphone and a battery power source to the circuit and temporarily connect a speaker to the 3.2-ohm winding on T1. Check the idling collector current by inserting a 0 to 10-ma meter in series with the grounded center-tap connection on the output transformer. Adjust the value of R1 until the current reads 3 ma. Disconnect the meter and have someone talk into the microphone while you listen to the quality of audio in the speaker. If it seems to be satisfactory, decrease the resistor value slightly and again check the current and audio

quality. Also try making the resistor slightly larger in value and note the increase in distortion. Select a value for R1 which produces the lowest idling collector current consistent with reasonable audio quality.

This modulator is used by connecting the 3,000-ohm winding in series with the supply voltage to the modulated stage. (This technique was described earlier in this chapter.)

BUILD A 5-WATT MODULATOR

Although the half-watt modulator just described will handle most of the projects in this book, the 5-watt,80/40 meter transmitter and the 5-watt Citizens band rig to be described later require additional modulating power.

The 5-watt modulator also employs class-B circuitry for high efficiency and low battery power consumption. The circuit shown in Fig. 6-15 was developed by the author to obtain maximum undistorted power with minimum power consumption. The circuitry is somewhat unconventional and is described in detail.

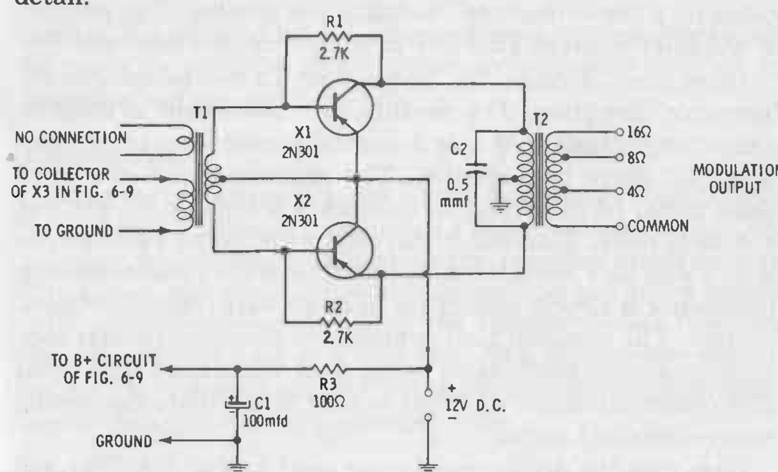


Fig. 6-15. Schematic diagram of the five-watt transmitter modulator.

How It Works

At first glance it appears that no forward bias is applied to the class-B transistors. Actually two forward-bias networks are used; they are independent of each other. Resistor R1, in conjunction with the winding resistance in the upper half of the secondary of T1, forward biases X1 while R2 and the transformer winding resistance biases X2. This circuit has the advantage that the operating point of each transistor is individually adjustable for minimum crossover distortion, although this is seldom necessary with the RCA 2N301 power transistors. An even more important benefit of this circuit is the large amount of degeneration supplied through the bias network. The large audio signal on the collector of each transistor is fed back to the base circuit through the 2.7K resistor and applied to the base in shunt with, and in opposition to, the driving signal. This negative feedback tends to smooth out any distortion created by the class-B action. It is so effective that only 25 ma of collector current is required to minimize the crossover distortion. This is half the value recommended by RCA for the 2N301's. The 2-to-1 reduction in battery consumption may be important in many applications. The pulsating current can cause audio feedback if the B supply to the preamplifier driver is not well filtered. Audio feedback is indicated by a "motorboating" sound in the speaker. The purpose of the filter network (R3-C1) is to prevent this feedback.

Capacitor C2 tunes the primary of T2 to reduce second-harmonic distortion. The modulation transformer is actually a matching transformer that is generally used to match a class-B output stage to a speaker. The primary impedance is 48 ohms center tapped while the secondary is capable of matching 4, 8, or 16 ohms. Thus the circuit can be used as an audio amplifier as well as a modulator. It so happens that the secondary impedance is almost perfect for 5- to 10-watt transistor transmitters. The common lead connects to B+, the 16-ohm connection delivers modulation to the final, while the 4-ohm point (the center tap of the winding) is used to modulate the driving stages described earlier.

Note that the driver transformer used in Fig. 6-9 (T1) will not match the low input impedance of the 2N301 transistors. In this application the transformer was replaced with the same type used as a modulation transformer in Fig. 6-12. However, the transformer is reversed so that the primary is used as a secondary and vice versa, to provide a correct impedance

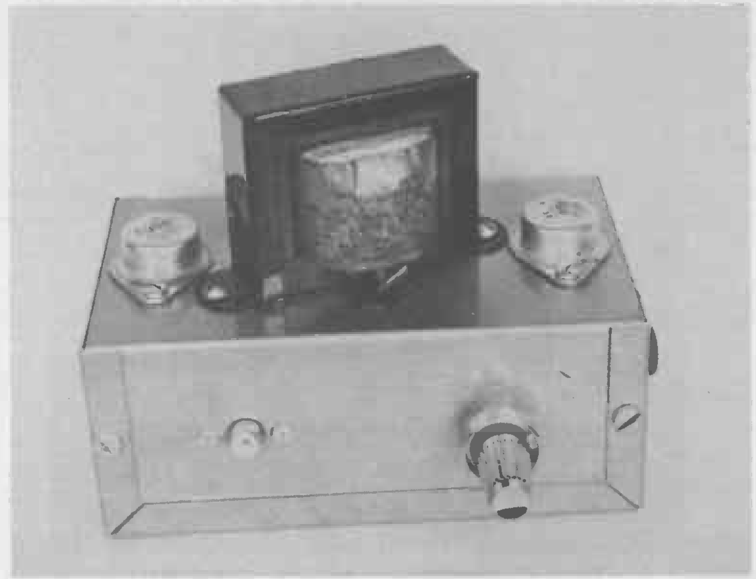


Fig. 6-16. Finished 5-watt modulator.

match. No substitution of the power audio transistors should be made as the RCA 2N301's were intentionally selected for their low leakage and high beta.

Construction

Like the earlier project, the 5-watt modulator is constructed in an LMB aluminum chassis box (Figs. 6-16 and 6-17). The preamplifier-driver circuit board is also mounted inside the box. The power audio transistors are mounted on top of the chassis with mica washers to insulate the transistor case from the chassis which is used as a heat sink. The transformer mounts next to the transistors, and connections to the secondary are made through a four-lug terminal strip.

Testing

It is an easy matter to connect the modulator to the 80/40 meter transmitter described in Chapter 5. The connections are shown in Fig. 6-19. Note that if both CW and AM capabilities are desired, a double-pole, double-throw toggle switch should be connected to the secondary of T2. The switch is used to

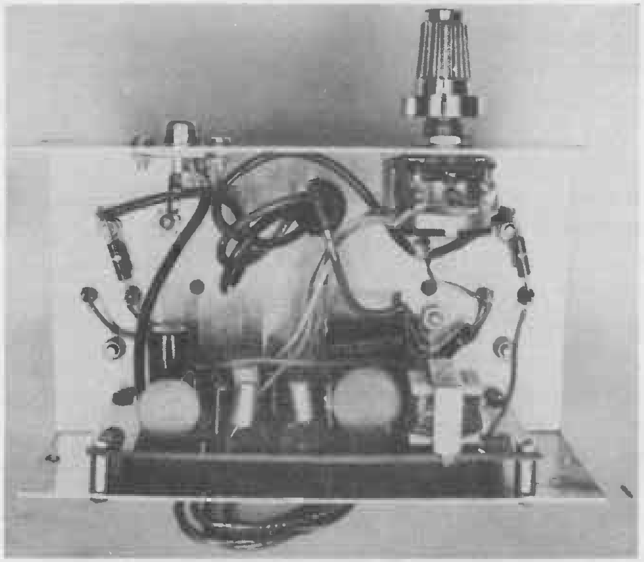


Fig. 6-17. Internal view of 5-watt modulator.

short the common bus to the 4- and 16-ohm connections, thus placing maximum voltage on the transmitter circuits and preventing chirp in the CW signal emission.

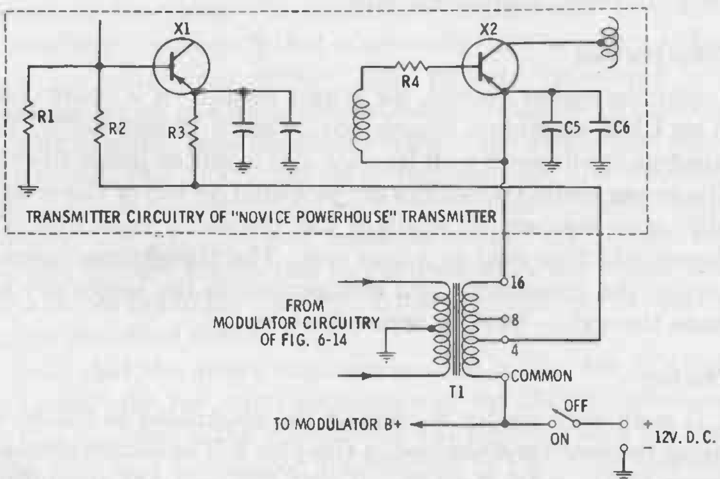


Fig. 6-18. Circuitry showing the method of connecting the modulator to the 80/40-meter transmitter project of Chapter 5.

HIGH-IMPEDANCE MICROPHONE PREAMPLIFIER AND MODULATOR

PARTS LIST

Quantity	Item No.	Description
1	C1	.05-mfd, 75 WVDC, disc capacitor.
3	C2, C4, C5	8-mfd, 15 WVDC, electrolytic capacitor.
1	C3	100-mfd, 15 WVDC, electrolytic capacitor.
1	C6	.01-mfd, 75 WVDC, disc capacitor.
3	X1, X2, X3	General purpose PNP Audio transistor such as 2N404 (RCA), 2N1274 (TI).
2	R1, R2	47K, ½-watt, carbon resistor.
1	R3	15K, ½-watt, carbon resistor.
1	R4	4.7K, ½-watt, carbon resistor.
1	R5	470-ohm, ½-watt, carbon resistor.
1	R6	2.2K, ½-watt, carbon resistor.
1	R7	22K, ½-watt, carbon resistor.
1	R8	220-ohm, ½-watt, carbon resistor.
1	T1	Transistor driver transformer, primary impedance 10K, secondary 2K center tapped (see text). (W. H. Paulin WP-3, Lafayette TR-98, or equiv.).

NOTE: The author has arranged to have a circuit board and kit of parts for this project made available through the W. H. Paulin Co., Box 122, Upland, California.

HALF-WATT MODULATOR
PARTS LIST

Quantity	Item No.	Description
1	C1	.05-mfd, 75 WVDC, disc ceramic.
1	C2	100-mfd, 15 WVDC, electrolytic capacitor.
2	X1, X2	2N1274 transistors (Texas Instruments).
1	R1	See text.
1	R2	47-ohm, ½-watt, carbon resistor.
2	R3, R4	4.7-ohm, ½-watt, carbon resistor.
1	R5	220-ohm, ½-watt, carbon resistor.
1	R6	5000-ohm, audio-taper potentiometer.
1	T1	Transistor modulation transformer, 500-ohm, center-tapped primary, 3K and 3.2-ohm secondaries (W. H. Paulin WP-6, Lafayette TR-119 or equiv.).
1		LMB chassis box (LMB No. 880 or equiv.).
1		Volume control.
1		Microphone jack.
1		Terminal strip.
1		Knob.

5-WATT TRANSMITTER MODULATOR
PARTS LIST

Quantity	Item No.	Description
1	C1	100-mfd, 15 WVDC, electrolytic capacitor.
1	C2	0.5-mfd, 100 V, paper capacitor.
2	X1, X2	Audio power transistors, type 2N301 (RCA).
2	R1, R2	2.7K, ½-watt, carbon resistor.
1	R3	100-ohm, ½-watt, carbon resistor.
1	T1	Modulation transformer used as a driver, primary 3,000 ohms, secondary 500 ohms center tapped, 3.2-ohm winding not used (see text).
1	T2	Class-B, transistors-to-speaker, matching transformer, primary impedance 32 ohms, center tapped, secondary 4, 8, or 16 ohms. (TRIAD TY-64X or equiv.).

7

Tunnel-Diode Transmitters

Contrary to what the name implies, there is no physical comparison between the tunnel diode and a passageway through obstacles.

Fig. 7-1. Examples of tunnel diodes.

The small size of the tunnel diode is deceiving. Unlike a vacuum tube or transistor, the tunnel diode has only two terminals. It is capable of oscillating at frequencies from the subaudio range up to several thousand megacycles. This subminiature semiconductor Samson is truly an amazing device. Some examples of typical tunnel diodes are shown in Fig. 7-1.

HOW IT WORKS

The theory of operation, when it is used in a relaxation oscillator circuit, is somewhat analogous to that of a neon bulb (Fig. 7-2). A common type of neon bulb, such as the NE-51,

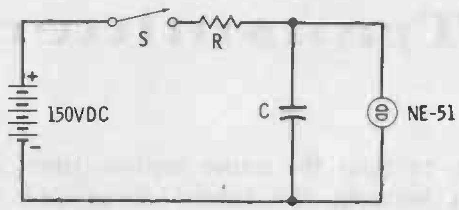


Fig. 7-2. A relaxation-type oscillator using a neon bulb as the negative-resistance device.

will produce oscillations when it is connected in this manner. When switch S is closed, voltage is applied to the RC network and the capacitor acquires a charge through resistor R, which has a value of several megohms. At some point on the exponential charging curve the voltage across the capacitor (and consequently the lamp) is sufficient to ionize the gas contained in the neon bulb. At this point the bulb fires, or strikes, and rapidly "switches" from an infinitely high impedance to a very low impedance. This discharges the capacitor, and since there is insufficient voltage to sustain the gas ionization, the bulb

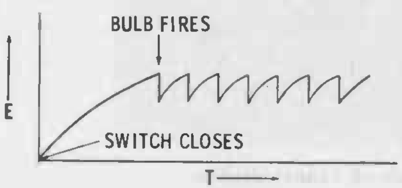


Fig. 7-3. Waveform generated by the circuit shown in Fig. 7-1.

is extinguished. Since the circuit is now reset to its original state, the capacitor once again charges and the cycle repeats. The circuit will continue to oscillate in this manner so long as voltage is applied to the input. The circuit generates a sawtooth waveform as shown in Fig. 7-3. The frequency of the oscillation is determined by the time constant of R and C. Incidentally, this circuit makes an excellent code-practice oscillator simply by substituting a telegraph key for S and connecting a pair of earphones in series with a .001-mfd capacitor across the neon bulb.

THE TUNNEL DIODE

Unlike the neon bulb which can ionize and deionize only a few hundred times a second, the tunnel diode does not depend on the firing of a gas. The tunnel diode is "cultured" to exhibit

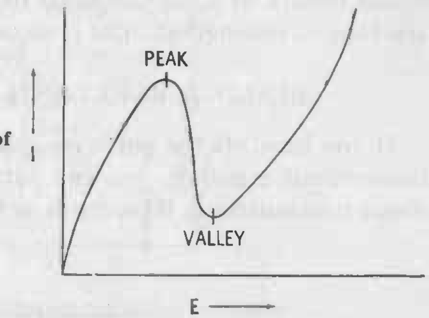


Fig. 7-4. EI characteristic curve of a typical tunnel diode.

a voltage-current curve such as the one shown in Fig. 7-4. As the voltage is increased, the current also increases until the peak point is reached. Increasing the voltage further *suddenly* causes the current flowing through the diode junction to decrease. This corresponds to the valley point on the characteristic curve. Decreasing the voltage will cause the operating point to rise back to the peak point. If a suitable energy-storage circuit is included in the circuit, the device will oscillate continuously between the peak and valley points. Some tunnel diodes are capable of switching between the peak or valley ten thousand million times in 1 second (10 kilomegacycles).

The theory of the tunnel diode is somewhat difficult to explain, because it operates on a principle entirely different from that of a transistor. Most readers are familiar with the barrier found in diode junctions. This depletion region, which forms the PN junction, prevents the recombination of electrons and holes. It is necessary to apply a potential to the diode junction to drive charge carriers over the barrier. The barrier has intangible dimensions and is usually described in terms of the quantity of voltage required to initiate conduction. In the case of germanium this *barrier potential* is 0.15 volt; it is 0.6 volt for silicon materials.

Unlike the usual diode junction described, the charge carriers in a tunnel diode seem to actually tunnel through the barrier. At the point the barrier is overcome the charge carriers appear almost instantaneously on the other side. Actually the curve in Fig. 7-4 is somewhat idealized in that the trace between the peak and valley occurs so fast that it is virtually invisible on all but the most expensive oscilloscopes. The transition occurs in 2 nanoseconds or less. The nanosecond is a fraction of one-millionth of a second.

BUILD A 60-SECOND TRANSMITTER

If you have all the parts ready to assemble and "tack" the connections together, you can actually assemble this tunnel-diode transmitter in 60 seconds or less.

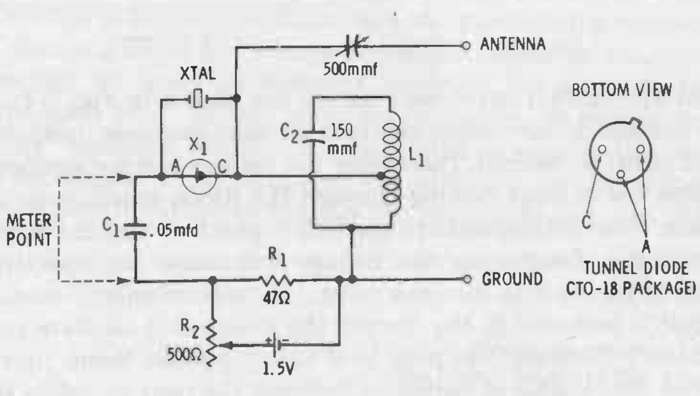


Fig. 7-5. The tunnel-diode oscillator.

Since there is no particular rush, however, let's assemble the transmitter on a piece of pegboard, as in an earlier project. Fig. 7-5 is the schematic diagram for an 80/40-meter tunnel-diode transmitter. A tuned circuit (L1), resonant at the crystal frequency, is connected in series with the tunnel diode (X1). A 47-ohm, ½-watt carbon resistor completes the circuit. Critical tunnel-diode current is adjusted by means of a potentiometer in series with the 1.5-volt battery.

Construction

The board used measures 9 holes by 7 holes. Insert pegs in the four corner holes and tie the upper and lower pegs together as in the pictorial diagram (Fig. 7-6). Mount the printed-circuit coil form in holes F-2, G-2, and G-3. The crystal socket is mounted in holes D-3 and D-5. The tunnel diode is connected

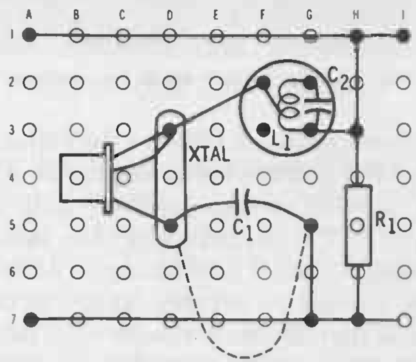


Fig. 7-6. Layout of the tunnel-diode oscillator.

in parallel with the crystal. The leads can be "tacked" across the crystal, or a transistor socket can be mounted in holes A-4, B-3, and B-5. Connections to the tunnel diode are also shown in Fig. 7-5. A Hoffman HT-8 type was used, but virtually any tunnel diode will work. The popular General Electric 1N2939 carried by most radio stores makes an excellent oscillator.

Adjustment

Disconnect the wire across the .05-mfd capacitor and replace it with a 0- to 10-ma meter. Connect a 500-ohm potentiometer and a dry cell in series (Fig. 7-7). The plus lead should connect to A-1 and the minus wire to A-7. Set the potentiometer to the

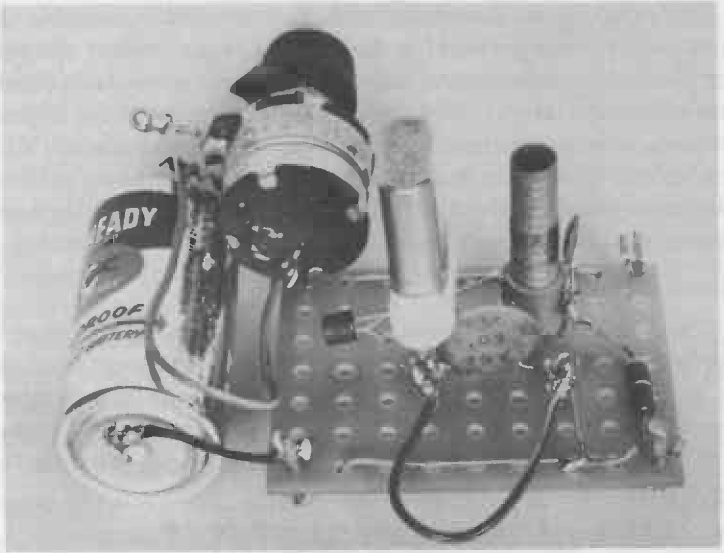


Fig. 7-7. The tunnel-diode transmitter.

maximum-resistance position. Observe the action of the tunnel-diode current as the potentiometer resistance is decreased.

The current should increase smoothly until it reaches approximately 4 ma. The current will then suddenly drop to a lower value (about 2 ma). Continuing to decrease the resistance causes the current to decrease even further to approximately 1 ma, and then it slowly rises to 4 or more ma. Do not decrease the potentiometer resistance beyond this point; the diode could be damaged because of excessive current flow. At some point between the 2-ma and 1-ma currents described previously, the diode becomes "unstable" and is capable of oscillation. Listen on a communications receiver to the frequency the crystal is cut for and adjust the coil slug slowly. The antenna should be disconnected from the receiver and a 3′ length of wire connected instead. This will prevent reception of strong signals which might mask signs of oscillation in the tunnel-diode transmitter. It may be necessary to juggle the adjustment of the coil and potentiometer to initiate the oscillation. You will have no trouble recognizing the correct signal once it starts. The note is very pure and stable. If the coil or potentiometer is misadjusted, oscillations will be heard on other

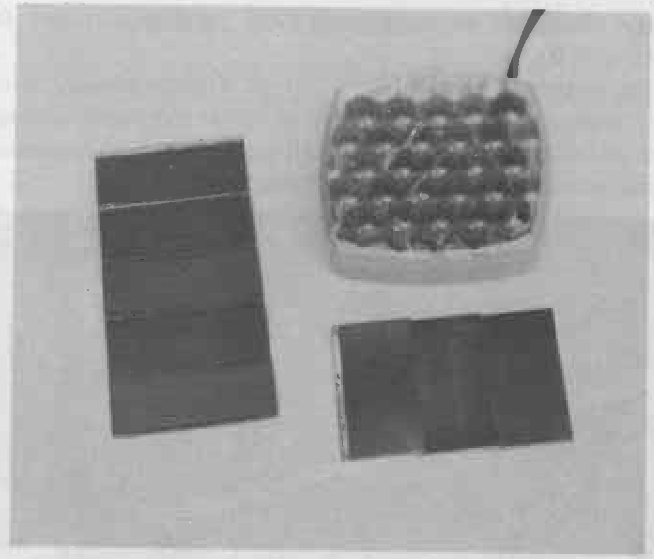


Fig. 7-8. Light-power solar cells.

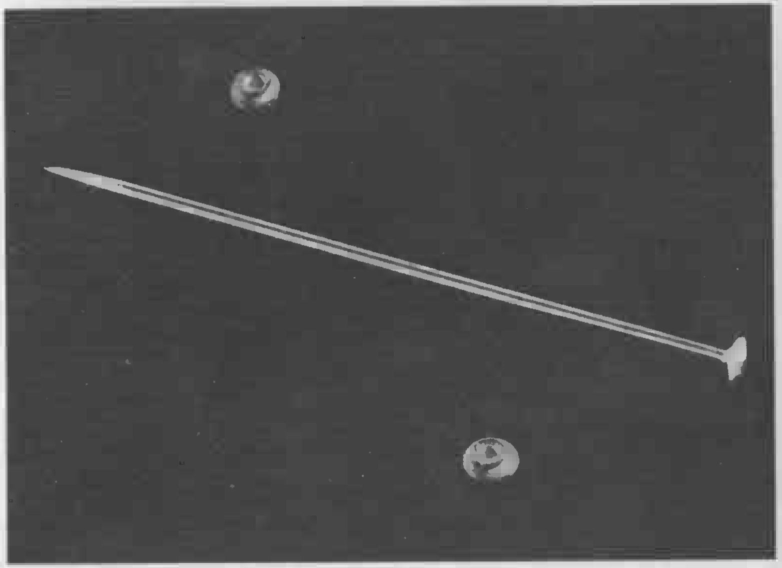
frequencies but they will be very unstable and obviously not crystal controlled. Once you get the "feel" of the circuit, connect an antenna to the transmitter in series with a 50- to 500-mmf trimmer capacitor. The value of this capacitor may require adjustment, depending on the length of the antenna.

The transmitter can be keyed by connecting the telegraph key in series with the antenna. Although the power output is measured in microwatts rather than milliwatts, this "pee-wee powerhouse" is capable of communications. In 1960, a New Zealand ham, Lester Earnshaw, ZL1AAX, contacted a station 160 miles distant on the 80-meter CW band using a similar tunnel-diode transmitter.

In describing his transmitter (*Radio Electronics*, March, 1961, page 34), Earnshaw points out that the silicon solar cell is an excellent current source for the tunnel diode. It is possible to make the transmitter operate on "light power" only by replacing the battery and potentiometer with a silicon solar cell of the type shown in Fig. 7-8.

A line of extremely tiny germanium tunnel diodes has recently been announced by Sylvania. Enclosed in the *Dot* package, the diode measures only .050" in diameter and .030" high (Fig. 7-9).

If the experimenter cares to invest a little money, one of these tunnel diodes in conjunction with the solar cells described before could be built into an entirely self-contained transmitter in a size smaller than an air-mail stamp.



Courtesy Sylvania Electric Products Inc.

Fig. 7-9. Tiny germanium diodes pictured next to a common pin.

The cost might be prohibitive for most amateurs as the *Dot* diodes are priced at from \$30 to \$130. They could, however, be used to construct a number of novice experimental circuits by the interested, well heeled amateur.

TUNNEL-DIODE TRANSMITTER

PARTS LIST

Quantity	Item No.	Description
1	C1	.05-mfd disc capacitor.
1	C2	150-mmf mica capacitor.
1	R1	47-ohm, ½-watt, carbon resistor.

TUNNEL-DIODE TRANSMITTER

PARTS LIST (CONT.)

Quantity	Item No.	Description
1	R2	500-ohm potentiometer.
1	L1	Printed-circuit type coil resonant at crystal frequency.
1	X1	Tunnel diode (General Electric 1N2939 or equiv.).
1		Crystal socket.
1	Xtal	Crystal for desired operating frequency.
1		Piece "Vectorbord" 9 holes × 7 holes (type A).
14		Push-in terminals (Vector T-30 or equiv.).

8

High Power for CB or 10 Meters

This interesting project uses only four silicon planar transistors (plus those in the modulator) and is capable of generating more than 3 watts RF power output on the Citizens band or 2.5 to 3 watts on the 10-meter amateur band. The transistors are designed for very high-frequency operation, and the transmitter should produce more than 2 watts output on six meters (by rewinding the coils), although this has not been verified.

HOW IT WORKS

The circuit for the high-power CB transmitter is shown in Fig. 8-1. The popular 2N706 is used as a common-base oscillator with feedback occurring through the transistor capacitance between the collector and emitter. The crystal (a third-overtone, 27-mc type) is connected between base and ground. It operates at series resonance and bypasses the base at its overtone, thus permitting oscillation only at that frequency. Resistors R1 and R2 provide forward bias, while R3 is used for emitter degeneration and stabilization. An RF choke is connected in series with R3 to prevent the low resistance from loading the feedback energy.

The 27-mc oscillator energy is developed across L1 and coupled to the buffer (X2) through a link winding. Resistor

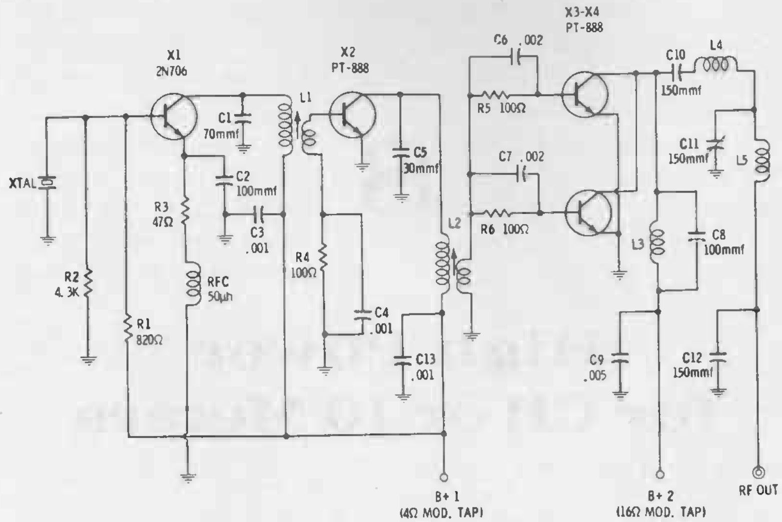


Fig. 8-1. Circuit of high-power CB and 10-meter transmitter.

R4 limits the buffer base current. Reverse bias is developed across the resistor which biases the stage into deep class-C operation. No emitter resistance is required since the stage operates only over a very narrow portion of the positive half cycle.

Slightly less than 1 watt of drive is applied to the final, which consists of two PT-888's connected in parallel. Drive-limiting resistors are used here also. Capacitors are used to bypass the signal around the resistors for maximum drive. Separate net-

Fig. 8-2. Parts layout for CB and 10-meter transmitter.

works are used in each transistor base circuit to equalize the drive level to unmatched transistors.

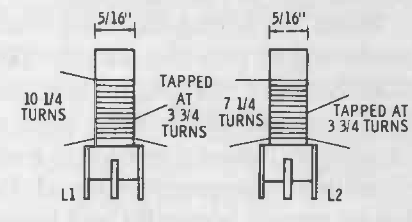
Capacitor C8, the collector capacitance, and inductance L3 form the "sloppy-tuned circuit," discussed in Chapter 4 and is broadly resonant at 27 mc. A double pi network consisting of C10, D11, C12, L4, and L5 couples the RF energy from the transistor collector circuit to the antenna. This somewhat complex circuit is required to provide some 40 db of attenuation to the second-harmonic energy, which could interfere with Channel-2 television or other services.

B-supply connections 1 and 2 can be tied together and connected to a source of 12 to 14 volts DC for CW operation. The high-power Citizen band transmitter can also be used for voice transmissions by connecting it to the 5-watt modulator described in Chapter 6.

Construction

The transmitter is constructed on a printed-circuit as shown in Figs. 8-2 and 8-4. This is necessary because of the low impedance and high circulation currents involved. The emitter

Fig. 8-3. Coil-winding detail.



leads of X2, X3, and X4 should be kept as short as possible or tied to the ground line through additional strips of tin to reduce the lead inductance. Printed-circuit coil forms are used for L1 and L2, while L3, L4, and L5 are airwound and are self supporting. Coil winding data are given in Fig. 8-3. Fig. 8-4 shows a view of the completed transmitter.

Testing

Once the construction has been completed, recheck the wiring carefully. The transistors are rather expensive, particularly when compared to the low-cost, general-purpose devices used elsewhere in the book. A collector-base short circuit or

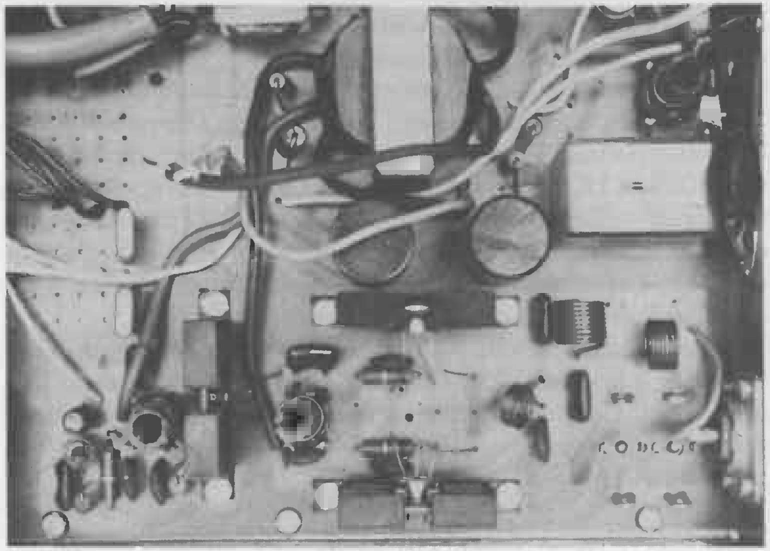


Fig. 8-4. View of complete transmitter.

accidental reversal of the battery polarity can instantly destroy one or more of the transistors.

When the wiring has been verified, connect a dummy load, consisting of four No. 44 pilot lamps connected in series parallel, between the output terminal and ground. Tie B+1 and B+2 together and connect them to plus 12 volts. If a receiver capable of tuning the crystal frequency is available, use it to monitor the transmitter signal. Connect a voltmeter (use the lowest range) across R4 and tune the slug in coil L1. The only voltage developed across R4 is due to signal rectification, and it provides an excellent oscillation-strength indicator. Tune L1 for maximum voltage, which should be about 1 to 2 volts. Next connect a 1K resistor to the base of X3 or X4 and measure the voltage between the free end of the resistor and ground; this should read approximately 3 volts. Peak coil L2 for a maximum reading on the voltmeter. At this point the dummy-load lamps should glow slightly. Peak variable capacitors C11 and C12 for maximum output. The constructor can also check the value of L3 by spreading or compressing turns for maximum output. After each movement of the coil, recheck the setting of C11 and C12. Also recheck the adjustment of coils L1 and L2, and

repeak the adjustment for maximum output. The bulbs should glow brightly.

This transmitter is connected to the 5-watt modulator constructed earlier in much the same way as the 80/40 meter transmitter. Both the oscillator/buffer sections and the final must be modulated. Connect B+1 (in Fig. 8-1) to the 4-ohm tap on the modulation transformer and the final B+ (B+2) to the 16-ohm connection. When the transmitter is modulated, the brilliance of the dummy-load bulbs should increase noticeably.

Obtaining Components

The transistor types specified (except the 2N706) have manufacturer's numbers and may be confusing. They are manufactured by Pacific Semiconductors (Aviation Blvd., Hawthorne, California) and are the type used in the Cadre Citizens band 5-watt transceiver. The transistors may be obtained from any large supply house which handles Cadre equipment; they cost $4.50 each. The remaining components should be readily available at radio wholesale stores.

HIGH-POWER CB AND 10-METER TRANSMITTER

PARTS LIST

Quantity	Item No.	Description
1	C1	70-mmf silver mica capacitor (DM-15 or NPO).
2	C2, C8	100-mmf silver mica capacitor (DM-15 or NPO).
2	C3, C4	.001-mfd disc ceramic capacitor.
1	C5	30-mmf silver mica capacitor (DM-15 or NPO).
2	C6, C7	.002-mfd disc ceramic capacitor.
1	C9	.005-mfd disc ceramic capacitor.
1	C10	150-mmf silver mica capacitor (DM-15 or NPO).
2	C11, C12	15- to 150-mmf trimmer capacitor (Arco or equiv.).

PARTS LIST (CONT.)

Quantity	Item No.	Description
1	L1	10¼ turns, No. 26 enameled wire closewound on a slug-tuned, printed-circuit coil form. Link 3¾ turns, interwound in bottom 4 turns of primary.
1	L2	7¼ turns, No. 18 enameled wire closewound on a slug-tuned, printed-circuit coil form. Link 3¾ turns interwound in bottom 4 turns of primary.
1	L3	7 turns, No. 26 enameled wire, ¼" inside diameter, closewound and air supported.
1	L4	10 turns, No. 18 enameled wire ⅜" inside diameter, closewound and air supported.
1	L5	7 turns, same as L4.
1	X1	2N706 transistor.
3	X2, X3, X4	PT-888 transistors (Pacific Semiconductors, see text).
1	R1	820-ohm, ½-watt, carbon resistor.
1	R2	4.3K, ½-watt, 5% carbon resistor.
1	R3	47-ohm, ½-watt, carbon resistor.
3	R4, R5, R6	100-ohm, ½-watt, carbon resistor.
1	Xtal	27-mc, third overtone crystal.

Index

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TRANSISTOR TRANSMITTERS for the AMATEUR

by Donald L. Stoner, W6TNS

ABOUT THE AUTHOR

Don Stoner, well known for his numerous articles on various phases of amateur radio in national magazines, has authored several other books. He is an instructor at Chaffey College in Ontario, California, and president of Stoner Electronics. Much of his time is spent as a consultant, designer, and builder of special communications equipment.

The rewards and enjoyment of a fascinating hobby are well known to all radio amateurs. The greatest satisfaction for many hams comes from constructing and operating their own rigs and thus keeping abreast of the latest developments in communications.

Transistor Transmitters for the Amateur is a book written for all amateurs, whether they want to build their own transmitter or merely acquaint themselves with transistor transmitter theory. Starting with a short history of communications transistors, the book progresses through transistor oscillators, RF amplifiers, modulators, and simple antennas. Each construction project contains photographs, schematic diagrams, chassis-layout drawings, tuning procedures, and operating instructions.

The projects described include: a crystal checker and calibrator; an 80/40 meter "Peanut Whistle"; the "CB Cyclone"; the "Peanut Whistle II"; the "CB Cyclone II"; the "Novice Powerhouse"; a high-impedance microphone amplifier and modulator; a half-watt modulator; a 5-watt modulator; and a high-power CB or 10-meter transmitter—useful equipment for all amateurs, CB'ers, students, and experimenters.

Presented in an informal style with extensive use of photos and drawings, *Transistor Transmitters for the Amateur* is a book that you'll want in your shack as a reference as well as a construction guide.

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