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A Second Look
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At one time or another I suppose most of us have wondered why our semiconductors are numbered as they are with 1N, 2N and 3N numbers. To the newcomer, it must look like a meaningless bunch of random numbers—and it isn't a great deal more to the oldtimer! Receiving-tube numbers follow certain guidelines, as do many transmitting tubes, so what happened to solid state? To say that it is a numbering system that grew on us isn't too far from wrong—but that's getting ahead of the story.

During the early days of radio, each vacuum tube manufacturer numbered his tubes as he saw fit. When there were only a few tube types available—with very little difference in their characteristics—this was no great problem. In fact, if you look in any of the old radio books, you'll see that none of the circuits specify what tube to use; it didn't make that much difference.

However, with the invention of the tetrode and pentode, and the introduction of tubes with vastly different characteristics, the problem became pretty sticky. The replacement problem was particularly bad, so in 1933 the industry voluntarily adopted a numbering system that called for a number, a letter and another number. The first number denoted the filament voltage range, the middle letter was a serial designation and the last number indicated the number of useful elements for which terminals were provided (including internal shield and shell connections). This is pretty much the system still in use today.

Transmitting tubes didn't run into the standardization bugbear until 1942. With the war, and the great number of tube types being manufactured, a standard number-letter-number system was adopted for transmitting and special-purpose tubes. Examples of devices assigned number from this system are the 1N21, 2C39, 2E26 and 3E29. The first number indicated the power rating of the heater: 1 for zero power, 2 for up to 10 watts, 3 for up to 20 watts and so on. The letter indicated the structure or function: B for diodes, C for triodes, D for tetrodes, E for pentodes, N for crystal diodes and rectifiers, etcetera. The final number was a serial designation, started at 21 to avoid conflict with the receiving-tube system.

Except for the use of 1N for semiconductor diodes, this system was scrapped in 1946 for a purely numerical system starting at 5500. Although several manufacturers wanted a numbering system for diodes (and later transistors) that told the user more about the device than a simple serial system, none was ever agreed upon. In the meantime, the system we use today kept growing. The first digit came to indicate the number of elements minus one—thus a 1N34A is a diode, a 2N706 is a triode and a 3N159 is a tetrode; the N indicates a solid-state device, and the last number is the order of registration.

This brings up another question—who decides precisely what number will be assigned to a particular transistor? The answer is an industry sponsored committee that registers all new transistors (and vacuum tubes), assigning the next open number in the system. If the system doesn't make much sense to you, don't feel too badly; a number of professional groups have tried to put some meaning into it, but without a great deal of success. The big argument against any new system at this point in time concerns the great number of solid-state devices already on the books—making any new numbering system impractical.

Jim Fisk, W1DTY
Editor
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The technically-minded amateur has known for some time that a vacuum tube becomes progressively less effective as the frequency of operation is increased. Amplifiers require greater driving power, output power drops off, and the manufacturer may be obliged to derate the tube at the high-frequency end of the operable range. If the frequency is raised high enough, the gain of the amplifier will drop to an unusable level. Upon further increase in frequency, the gain will drop to unity or less. At the same time this is happening, the input impedance of the amplifier drops as does the maximum impedance realizable in the plate circuit.

Numerous factors contributing to the reduction of amplifier output at vhf and uhf can be listed and divided roughly into three groups:

1. Circuit-reactance limitations
2. Circuit and tube loss limitations
circuit-reactance limitations

At the very-high and ultra-high frequencies there exists a situation which is quite different from that which exists at low frequencies. At low frequencies the electrical circuits and the tube are quite distinct. As frequency increases, this ceases to be true, and it is found that part of the resonant circuit exists within the tube (fig. 1). The electrode leads, while they are normally short and have large surface area in modern tubes, have a small but finite inductance. As the frequency of operation is raised, the reactance of the lead inductance will become quite appreciable, often reaching undesirable proportions in the vhf-uhf regions. The effect of lead inductance is to create a voltage drop such that the applied voltage across the external terminals of the tube will not entirely appear across the electrodes. Driving voltage is thus lost across the lead inductance.

In addition, while the interelectrode capacitances may be small, at ultra-high frequencies they approach a large fraction of the capacitance required to establish resonance in an external circuit. As such, the inter-electrode capacitances represent a limitation in terms of actual operation as the external tank circuit may “disappear” within the tube. The combination of the electrode-lead-inductance and the interelectrode capacitance may cause an internal resonance in the uhf region.

A typical internal resonance experienced in a large power tube is the circuit consisting of the control-grid cage and mounting cone, and the screen-grid and cone. These two assemblies can form a quarter wavelength long tuned-line circuit shorted at the tube envelope by the capacitor consisting of the two contact rings and the dielectric material used in the envelope. The internal resonant frequency could be in the range of 1400 MHz for a five- to ten-kilowatt tube. This could lead to a 1400-MHz parasitic oscillation.

The smaller the tube, the higher the resonant frequency. All tubes will have internal resonances and the designer must move them out of the normal usable frequency range or load the circuit in such a way as not to degrade the performance within the rated frequency range. Even if resonances do not occur, the combination of reactances within the tube may constitute an undesired network that creates a mismatch between the driver and the tube electrodes.

magnitude of lead inductance

The most important reactance encountered in a vacuum tube is that associated with the lead inductance. An estimate of the magnitude of lead inductance may be made using the following equation:

$$L = 1.00508L (2.303 \log_{10} \frac{4L}{d} - 1 + \frac{d}{2L})$$

where

- \(L\) = inductance (microhenries)
- \(L\) = wire length (inches)
- \(d\) = wire diameter (inches)

This equation assumes no mutual inductance with some other nearby lead or wire.

As an example of how great lead inductance can be, consider the case of a lead that is 0.1-inch in diameter and one-inch long. This lead will have an inductance of 0.015 \(\mu\)H and an inductive reactance of 47 ohms at 500 MHz. An inductive reactance of this magnitude, for example, between the screen element and the screen by-pass capacitor outside the tube can cause stability problems. A voltage drop exists across the
screen reactance caused by the current flowing through it which is the VHF current charging the output capacitance of the tube. This voltage is impressed upon the screen, which thereby is removed from ground potential, disturbing the grid-plate isolation normally afforded by this element.

This source of positive input resistance can be better understood by realizing that it results from a signal appearing across the cathode lead inductance, driving the cathode a small amount, as in a cathode-driven amplifier. (In a cathode-driven amplifier, the alternating current component of the plate current has to flow through the cathode lead inductance to reach the cathode. Since the driving signal is in series with the output load through the cathode-to-plate resistance of the tube, some of the power in the load is supplied by the driver.) Unwanted feed-through power is thus supplied by cathode lead inductance in a grid-driven stage.

The above equation shows that as the frequency is increased, the grid-loading due to cathode-lead inductance increases. For example, if the frequency of the amplifier is tripled from 144 to 432 MHz, the input resistance of a particular tube at 432 MHz would be one-ninth the resistance at 144 MHz. If it is desired to drive the 432 MHz tube to the same maximum grid voltage swing, then nine times the power is required just to cover the input loading due to cathode-lead inductance.

There will be other losses such as skin

cathode lead inductance and input loading

Tube gain will also be adversely affected due to a reduction in input resistance because of the cathode-lead inductance. A small amount of cathode-lead inductance in conjunction with the grid-to-cathode capacitance of a vacuum tube will cause a resistive load to appear across the input of the tube. The magnitude of this added shunt input resistance is:

\[ R = \frac{1}{\omega^2 L_k C_{gk} g_m} \]

- \( R \) = input resistance due to cathode-lead inductance
- \( \omega = 2\pi f \) where \( f \) is frequency (hertz)
- \( L_k \) = cathode lead inductance (henries)
- \( C_{gk} \) = Grid-to-cathode capacitance (farads)
- \( g_m \) = transconductance (mhos)
effect, dielectric, radiation, transit time and the circuit reactance effects shown in fig. 1. The amount of grid loading due to cathode-lead inductance may be reduced somewhat by separating the input and output circuit paths back to the cathode. Some miniature tubes and low power transmitting tubes, such as the 4CX250B, 6146 and others, have multiple cathode leads to minimize the cathode-lead inductance effects. Transistors also have this problem, only it is called emitter-lead inductance instead of cathode-lead inductance.

vhf/uhf cathode-lead inductance neutralization techniques

It is possible to neutralize the effects of the cathode-lead inductance by choosing the value of cathode by-pass capacitor so it will be approximately series resonant with the total lead inductance (tube, socket and circuit inductance). This technique is particularly effective in low noise stages of vhf/uhf receivers.

Fig. 2 illustrates stage gain as a function of cathode by-pass capacitance value in a high power vhf amplifier. This particular graph was obtained from an experimental 100-MHz, 5CX1500A amplifier running in class B. The 5CX1500A cathode lead, socket and circuit lead inductance was measured to be just over four nanohenries. A capacitance of 637 picofarads was calculated to be necessary for cathode lead inductance neutralization. The graph agrees fairly well with the calculated data.

A neutralization technique described in the October 1939 issue of Electronics is of interest (fig. 3). I have had no personal experience with this technique, but it does appear to have merit. The voltage drop $e_L$ across $L$, caused by the cathode current, is reversed in polarity with respect to $e_R$, in the sense that $C_{gg}$ and $L$ are in series between grid 1 and the cathode. Thus, the current flowing back to the grid through $C_{gg}$ is 180 degrees out of phase with the applied voltage, $e_{Input}$. For a certain value of $L$ the currents through $C_{kg}$ and $C_{gg}$ will be equal as well as opposite in phase; thus, the conductance between grid 1 and the cathode is zero. For neutralization of the cathode lead inductance the following ratio must be met:

$$\frac{L}{L_k} = \frac{C_{gg}}{C_{kg}}$$

For the 5CX1500A 100-MHz amplifier component values for neutralization of the cathode lead inductance could be:

- $L_k = 4$ nanohenries
- $C_{kg} = 35.8$ picofarads
- $C_{gg} = 10$ picofarads
- $L = 14$ nanohenries

The well known 4CX250B—rated to 500 MHz.
which will be discussed in the section on interelectrode capacitances.

**intermodulation distortion and input loading**

An appreciable amount of input loading can increase intermodulation distortion. Any tube plate-characteristic non-linearity will cause a variation in this input loading with signal level and thereby present a varying load to the driver, thus causing increased distortion in the drive voltage.

(fig. 4. Neutralizing cathode lead inductance with an inductor in the screen lead on the tube side of the screen bypass capacitor.)

$L_k$ and $C_{gg}$ are typical characteristics of the 5CX1500A and SK840 socket while $C_{gg}$ and $L$ are added components.

A very similar method of neutralizing that part of input resistance caused by the cathode lead inductance uses an inductance in the screen-grid lead and has been often used in 6146-type gear at 144 MHz (fig. 4). Basically, this circuit is the same as fig. 3. There are two differences: the point K (cathode pin) is now at ground potential and only screen current flows through the inductance $L_k$.

Other circuits may be used to minimize the effects of the cathode-lead inductance. The grounded-grid, cathode-driven (or probably more correctly called the "grid-separation" circuit), is often used (fig. 5). In this case the cathode is driven while the grid in a triode, or the grid and screen grid in a tetrode, are operated at some low rf potential. The grid structures then act like a shield between the input and output circuits. The main advantage, as far as cathode-lead inductance is concerned, is that this inductance is now just another inductor in series with, and therefore a part of, the input tuned circuit. There are other advantages that can be credited to the "grid separation" amplifier.

**screen lead inductance**

The screen lead inductance between the screen element and the screen by-pass capacitor may help or hinder the operation of an amplifier. Below the self-neutralizing frequency of the tube, the screen lead inductance is usually detrimental to the stability of the amplifier, as the rf current flowing through this inductance will cause an unwanted rf voltage to be developed. The point where the screen bypass is connected to the screen terminal may very well be at rf ground potential, but the potential of the screen itself may be varying above and below ground by the magnitude of voltage developed across the screen lead inductance, $e_L$ (fig. 4).

The magnitude of the developed voltage depends on the inductance of the screen lead and the frequency of operation. The higher the frequency, the greater the inductive reactance and the greater the rf current through this inductor. The current is greater.
because the capacitive reactance of the output capacitance of the tube will be smaller, and the capacitor will be charging to the same rf plate voltage swing each cycle.

At operating frequencies below the self-neutralizing frequency of a tetrode, screen inductance may be added in order to neutralize the amplifier. That is, the self-neutralizing frequency is lowered to the operating frequency. At operating frequencies above the self-neutralizing frequency of the tube, a series capacitor is sometimes added to move the self-neutralizing frequency up to the operating frequency (fig. 6). As was briefly touched upon previously, sometimes it is desirable to add a certain amount of screen lead inductance to neutralize the number of leads to use to provide the inductance he needs for his design.

In a cathode-driven (grounded-grid) amplifier, control-grid inductance is very important. Just as in the case of the screen-lead inductance in a grid-driven tetrode amplifier, the control-grid inductance in a cathode-driven amplifier may aid or hinder the designer. The control-grid inductance may cause instability, a loss in drive voltage due to the voltage divider effect (fig. 1) or it may be used to provide a method of neutralizing the amplifier.

**plate lead inductance**

The plate in modern tubes used for uhf operation is usually designed with a massive anode structure in order to dissipate the heat that is generated in this element of the tube. For this reason, plate lead inductance is usually low enough so it is not of any great concern. If normal good engineering practice is followed in designing uhf circuitry, the plate lead inductance becomes inconsequential.

**interelectrode capacitance**

In addition to lead inductance, interelectrode capacitance plays an important role in the operation of tubes in the uhf regions. Interelectrode capacitances due to active parts of the tube structure are incapable of reduction beyond a certain point. However, in many tubes the interelectrode capacitance results largely from capacitance between leads in areas of the tube where electrons do not flow. It is the job of the tube designer to reduce this unnecessary capacitance to a minimum.

---

fig. 5. Cathode-driven or grid-separation amplifier.
input capacitance

The input capacitance of a grid-driven tube is the sum of the grid-to-cathode and grid-to-screen capacitances. The larger the input capacitance, the greater the drive power must be. This can be explained by the very large increase in input charging current necessary to charge the input capacitance. As the frequency increases the reactance of the input circuit becomes smaller, and for the same peak grid voltage, the charging current may limit the magnitude of the peak grid voltage. Peak grid voltage can also be minimized by operating with less bias. Quite often in certain amplifiers the class-B mode is more desirable than class-C operation. Reducing the peak grid voltage and the charging current will reduce the amount of power that must be dissipated by the control grid. Radio frequency power dissipated by the control grid unfortunately cannot be measured by the dc meters on the front panel of the amplifier, so the operator has no means of knowing if charging currents cause excessive grid temperature. High input capacitance also limits the bandwidth of the input circuit. For those applications requiring large instantaneous bandwidth, great care must be taken in the design of the equipment.

As power is proportional to the square of the current, the power lost in the input circuit necessarily increases as the charging current rises. The driver must supply this extra power.

To reduce this loss, the circuit designer must keep the input capacitance down and

\[
\text{fig. 6. Screen neutralization at operating frequencies above self-neutralizing frequency of tube.}
\]

\[
\text{fig. 7. Chart for computing resistivity and depth of penetration for metallic conductors between 100 MHz and 100 GHz.}
\]
The grid-separation or cathode-driven amplifier offers quite an advantage as far as input capacitance is concerned. The input capacitance consists only of the cathode-to-grid capacitance. For the same tube in the cathode-driven configuration, the input capacitance will be roughly half that value in the grid-driven circuit. This is quite an advantage for applications requiring wide bandwidths.

**output capacitance**

The output capacitance of a power tube is an important factor in determining what plate-load resistance can be used. This in turn determines the stage gain and power output that is available. The equivalent shunt resistance (plate-load) of a parallel resonant circuit can be written as 

\[ R_L = \frac{Q}{2\pi f C} \]

or 

\[ R_L = \frac{Z_{rf}}{Q} \]

where \( R_L \) is the plate-load resistance, \( Q \) is the loaded \( Q \) of the resonant circuit and \( f \) is the resonant frequency. For operation at a given frequency, to increase the shunt resistance it is necessary to decrease the shunt capacitance. This can be done to a point by reducing the circuit capacitance and increasing the tank coil inductance, maintaining the same frequency of resonance. Eventually, this process is limited by the fact that the capacitance external to the tube has been reduced to zero; the shunt resistance is finally determined by the tube interelectrode capacitance. The larger the interelectrode capacitance, the smaller the shunt resistance that can be realized. Accordingly, power output tends to drop off as the load resistance, or as the square of frequency, as frequency increases.

There are other problems brought about by the effect of output capacitance. The output capacitance must be charged and discharged during each cycle of the radio frequency. Again, as the frequency increases, the reactance of the output capacitance decreases. Therefore, with the same value of peak rf plate voltage, the current flowing through the output capacitance must increase as the frequency increases. The output capacitance of a tetrode is made up of the screen grid structure, the plate structure and the tube envelope.

The charging currents must flow over the surface of these components of the tube, all of which have varying degrees of rf resistance. It is possible for the charging currents to exceed the dissipation rating on the screen grid even though the dc meters indicate all is within ratings. It is advisable in vhf/uhf circuits, therefore, to try to achieve the lowest usable value of peak rf plate voltage by using the lowest plate load resistance and drawing the highest plate current consistent with desired output power and efficiency. Running the tube in this manner will lighten the rf stress on the tube seals and reduce the rf current in the screen grid structure thus providing for a potentially longer tube life.

**feedback capacitance**

The feedback capacitance in a grid-driven amplifier is the capacitance from anode to the control grid. The higher the frequency of operation, the greater the chance for instability due to rf feedback from the output circuit through feedback capacitance to the input circuit. In the vhf/uhf region this capacitance and other tube capacitances and inductance must be adjusted to provide for neutralization by added circuit components.

The grid-separation amplifier helps minimize the effects of the feedback capacitance. The feedback capacitance in this configuration is very much less than the grid-driven case since it is the capacitance from anode to cathode with the grid, or grids, shielding the output from the input. In some applications no neutralization will be required. Other applications may require quite extensive neutralization.
Distributed-amplifier tube discussed in text. By choosing the number of control-grid leads, the designer can control lead inductance. The tube in the right-hand photo has a screen by-pass capacitor installed. Center pin is one of the heater pins; the square tab on the capacitor is the other heater connection and the cathode. The threaded pins are control-grid leads.

circuit and tube-loss limitations

The power losses associated with a tube and circuit all tend to increase with frequency. In the vhf/uhf region, almost all radio-frequency current flows in the surface layers of a conductor because of skin effect (fig. 7). The resistance and rf losses in a conductor increase with the square root of frequency because the layer in which the current flows decreases in thickness as the frequency increases. Insulating supports in the tube and external circuit have losses associated with the molecular movements produced by the electric fields. These dielectric losses will usually vary directly with frequency. Also, there will be additional losses due to the radiation of energy from the wires and leads carrying rf current. The power radiated from a short length of wire carrying current increases as the square of the frequency.

All these factors contribute to a general reduction in tube and circuit efficiency as operating frequency is increased. In the manufacture of power tubes the resistance losses are reduced by increasing conductor surface area and by proper choice of lead materials. Dielectric losses are reduced by
selection of envelope and insulating materials. Support insulators are positioned, when possible, out of high voltage fields. Radiation losses are reduced by constructing vhf and uhf tubes and circuits so as to be totally shielded. At times it is prudent to use concentric line construction techniques so that tube and circuit fields are entirely confined.

transit-time limitations

Electron transit-time effects can contribute to reduced tube output in many ways. Transit time is the finite time an electron takes in going from the cathode to the grid. If the transit time (a function of grid-to-cathode distance and grid-to-cathode voltage) is an appreciable fraction of one ultra-high-frequency cycle, then an electron in transit in the grid-to-cathode region might be still heading for what was once a more positive location than the cathode surface it just left, but now finds the grid may be less positive or perhaps even negative (fig. 8). As a result of transit-time effect in the cathode-grid region, there will be a dispersal of "out of step" electrons. Because of this dispersion of the electron stream, the plate-current pulses are not as sharp as the current pulses liberated from the cathode.

In addition, energy is required to accelerate the electron towards the anode, and this energy is supplied by the driver. As the operating frequency is raised, more energy is required because the grid-input resistance due to transit time varies inversely as the square of the frequency. That is, if the frequency is doubled, the input resistance due to transit time effects will be one-fourth that at the lower frequency. The extra power required to overcome transit-time loss due to grid-input resistance is supplied by the driver and appears as lost drive-power—required but put to no practical use other than to heat the tube seals and waste precious exciter output.

Paradoxically, transit-time loss and cathode-lead inductance loss (both of which cause input loading) are not all evil because they often tend to stabilize a "wild" stage. Cure of the trouble may lead to higher stage gain with the possibility of oscillation and instability!

Circuit loading can be used to achieve stability and this is often done at the lower frequencies by adding resistors across i-f transformers, for example. The circuit designer, therefore, finds that a tradeoff exists between loading and stage gain that will work to his advantage or disadvantage, depending upon his ability to analyze the circuit. There is often more than one successful path to a proper design and the good circuit engineer has several alternative paths (in his head, if not on paper) to choose from. A candid realization of uhf/vhf effects will help the circuit designer obtain maximum power, efficiency and reliability from his equipment.

references

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High-power linearity with Eimac power-grid tubes.
solid-state circuits
for
single sideband

Transistors haven't taken over ham equipment as quickly or as thoroughly as they have some electronic gear, but semiconductors of all kinds are steadily working their way in. You see more solid-state construction in homebrew rigs than in commercial ones. Nevertheless, when you look over the range of ssb equipment now available, you find there's hardly a circuit that hasn't gone solid-state in some brand or some model. No serious ham can help being interested.

At first, only small size and portability (battery operation) were the reasons for using transistors. But semiconductors have been around for 20 years now, and they are more reliable and stable than ever before. With few exceptions (high power, extremely high frequency) transistors can now perform most tube functions—and in many cases do a better job. Furthermore, the field-effect transistor (FET) has made high-impedance solid-state circuits possible.

Here are some places you'll see transistors in single-sideband transmitters and receivers.

speech or af amplifier

The two-stage amplifier of fig. 1 is typical of audio transistors used in sideband rigs. This circuit is from the Galaxy V Mark 2 transceiver. Q9 and Q10 are speech amplifi-
ers between the microphone and the balanced modulator. Power and frequency demands are modest, and hundreds of inexpensive transistor types can do this job.

The dc supply hookup is simple—positive voltage to the collectors, and emitters returned to ground through stabilizing resistors R105 and R109. R106 and R112 drop the supply voltage to the right collector potential, and C157 and C124 decouple the two stages from each other and from the rest of the transceiver.

Q9 gets base bias through R143 and R104, with C121 for decoupling. Q10 uses R107 to carry base bias; R107 is a voltage divider with R108.

The audio signal is brought from the microphone through blocking capacitor C120 and is applied to the base through a filter consisting of R103 and C122. This filter removes any stray rf, but has little effect on speech signal. Emitter resistor R105 is unby-passed; that adds a little degeneration to improve linearity in the stage. The audio output from Q9 develops across collector load R106.

This output signal is fed to the base of Q10 through C123. Q10 is a phase splitter, with collector and emitter resistors (R111 and R109) equal. The collector signal goes to a VOX adapter. The emitter signal is fed through C126 and microphone gain control R110 to the balanced modulator.

audio output stage

Audio power requirements aren't stringent in ssb receivers; a watt or two is usually sufficient to drive the speaker. Many sets use a pair of transistors in class-B push-pull service—like the one in fig. 2, from the Gonset 910A transceiver.

Since these transistors are pnp types, they require negative collector voltage (with respect to emitter). In this hybrid rig, however, supply voltage is positive with respect to chassis. The collectors of Q10 and Q11 are connected to ground through the low dc resistance of T5's split primary. The emitters go to the 12-volt line through R32, a stabilizing resistor. The collectors are thus negative with respect to the emitters. Base bias comes from a voltage divider (R30-R31).

The signal from the audio driver stage is coupled to Q10 and Q11 by the split secondary windings of input transformer T4. The collectors feed their output signals to the split primary windings of T5. The secondary matches a 3.2-ohm speaker.

crystal bipolar oscillator

Raytheon's Sideband Engineers SB-34 transceiver uses a germanium bipolar pnp transistor as a 456.38-kHz oscillator. On transmit, it generates the carrier frequency, and on receive it supplies the carrier-reinsertion signal. Output from this stage is fed to the balanced modulator, which is also the bal-
anced detector for receiving.

The circuit—shown in fig. 3—is a Pierce crystal oscillator. Feedback is from collector to base via the crystal.

As in the preceding circuit, the supply voltage is connected to the stage in what some people call an upside-down arrangement, which is used a lot with pnp transistors. The emitter and base are positive with respect to chassis, and the collector is grounded (through L11). The collector is thus negative with respect to the emitter.

Base bias is taken from divider R18-R19. The emitter stabilizing resistor (R20) is bypassed by C19, to prevent degeneration. Output voltage is developed by the collector across L11. Since the output is untuned, Q12 is operated class A, with a fairly clean sine-wave output.

**FET crystal oscillator**

One advantage of the field-effect transistor (over the bipolar) is that the FET's high impedance loads a crystal very lightly, making a more stable oscillator. FETs also have better temperature stability than bipolars.

These advantages are put to good use in the Drake R-4B receiver, which uses an n-channel junction FET as a calibration oscillator. The circuit is shown in fig. 4. It's another Pierce type, with feedback through the crystal from drain to gate. C119 is a trimmer to allow zero-beating the 100-kHz oscillator with WWV.

R73 returns the gate to ground for dc. Source bias is used, and source resistor R72 is bypassed for rf by C120. Output voltage is developed across drain load R155, decoupled by C189. C188 feeds this rf to a buffer amplifier. C138 lowers the drain-circuit impedance, thereby stabilizing the oscillator. The output is very high, so the shunting effect of C138 does no harm.

Q8 is operated class A and produces a clean sine wave at 100 kHz. Other stages divide this down to 50 and 25 kHz. The last stage produces lots of harmonics, putting harmonics every 25 kHz across the bands of interest. By rearranging jumpers in the divider stages, the user can get 25-, 50-, or 100-kHz markers to calibrate from.

**Voltage-tuned vfo**

In the past, vfo's have been LC-tuned, usually with an air-dielectric capacitor as the variable element. Some new transistor vfo circuits, however, use a voltage-variable capacitor (also called a varactor or varicap). The Gonset 910A transceiver uses a varactor to tune a vfo over the range of 5.5-6.5 MHz.

The circuit is shown in fig. 5A, and it's somewhat complex. Look at how it's redrawn in fig. 5B, though, and you'll recognize a
Colpitts oscillator; C32 and C34 form the feedback divider. L1 and all those capacitors are in series across the divider.

Q17's collector is negative with respect to its emitter, as befits a pnp bipolar. R33 is the load resistor, connected from collector to ground. R46 is the emitter stabilizing resistor, and connects the emitter to the positive 10-volt line. Base bias comes from a divider (R45-R34) between the 10-volt line and ground.

Still referring to fig. 5B, note that in the tank circuit there is only one variable element. This symbol represents six capacitors and a variable-capacitance diode. In fig. 5A you can see those parallel capacitors on the left side of the diagram. C26, C27, and C28 furnish the bulk of capacitance that resonates with L1. C29 and C30 are trimmers for aligning the vfo. C31 is a blocking capacitor, so the control voltage can be connected to D8 without affecting the base of Q17.

A voltage divider (R68, R47, and the vfo tuning control), connected from the 10-volt line to ground, supplies the control (tuning) voltage. As the slider of the tuning control is moved, it varies the bias applied to the diode. As the capacitance of D8 varies, it changes the frequency of the oscillator over a 1-MHz range. L2 and C98 decouple the tank circuit from the control-voltage line.

**i-f amplifier**

For an example of a simple i-f amplifier, look at fig. 6—the circuit used in Gonset's 910A transceiver; frequency is 9 MHz.

You should recognize the upside-down dc supply circuit used for Q5. Its collector is grounded for dc through T2's primary, and the emitter is returned to the 12-volt positive line through emitter stabilizer R16 (bypassed for rf by C12) and the rf gain control. Base bias is applied through R56, but is overridden during certain signal conditions by voltage from the agc line. Stage gain is varied two ways: by the manual rf gain control and by the agc line.

Transistor i-f transformers—such as T10 and T2—often use a lot of turns on the tuned winding (the primary in this case) to improve the Q of the tuned circuit. With many turns on the coil, the parallel capacitor can be small. Also, the slug-type tuning is more effective. But, because bipolar transistors are low-impedance devices, the primary of T2 is tapped at a low-impedance point for the collector connection. The low-impedance base of the next stage is fed from a low-impedance, few-turn secondary.

**two-way i-f amplifier**

The Raytheon SB-34 transceiver uses a novel i-f amplifier stage at 456 kHz. During transmission, the stage amplifies the dual-sideband signal from the balanced modulator and applies it to the mechanical sideband filter. During reception, the direction is reversed. Incoming rf is heterodyned down to the 456-kHz i-f and passed through the mechanical filter. The i-f stage amplifies these signals and passes them to the balanced modulator, which then functions in reverse as a product detector.

The circuit of this bilateral i-f amplifier is shown in fig. 7. The emitters of the two transistors share a common stabilizing resistor (R33) and by-pass capacitor (C29), and both emitters are tied to the 12-volt positive line. Q5's collector is returned to dc ground through the mechanical filter (not shown). Q6's collector is returned to dc ground through the secondary of T2.

As you may have guessed, one transistor is on during transmit, and the other is off. They switch during receive. The swapping is done by changing base bias. In the transmit
mode, line A is grounded. R31 and R32 become a voltage divider from 12 volts to ground, forward-biasing Q5 and turning it on. The sidebands of 456 kHz are fed from the balanced modulator through T2, and then through C28 and R30 to Q5, which amplifies them and sends them on to the mechanical filter.

Line B meanwhile is 12 volts positive. Divider R37-R38 has 12 volts at both ends. Q6 is therefore not forward-biased, and does not conduct.

In the receive mode, line A gets 12 volts, which turns off Q5. Line B is grounded, so now divider R37-R38 develops bias at the base of Q6. Rf coming from the mechanical filter goes through C30 to the base of Q6, where it is amplified and fed through T2 to the balanced-modulator/product-detector.

**rf amplifier**

In one commercial transceiver, a simple rf amplifier is used. Its circuit is shown in fig. 8.

The dc circuit is pretty much what you've seen before. No emitter resistor is used; the emitter is tied directly to 12 volts—although it is bypassed for rf by C106.

Base bias is developed by divider resistors R97 and R98; but notice: they're connected between the 12-volt line and the agc line. This keeps the rf stage from overload by applying agc action.

Rf from the antenna is fed through the T/R relay to tuned circuit L17-C129, which is broadly resonant across the transceiver's operating range of 49.975 to 54.025 MHz. The rf input is tapped down on L17 to match the 50-ohm antenna.

Another tap on L17, a little higher, matches the base impedance of Q21. From there, dc blocking capacitor C107 feeds the rf to Q21.

A tap on L16 matches the collector impedance of Q21. The resonance of tank circuit L16-C105 is broadened by R95. The rf output signal developed across the tank is fed through C104 to the first mixer.

**two-way mixer**

Earlier, you saw the two-way i-f amplifier used in the Raytheon SB-34 transceiver. The same principle is used in that unit's vfo mixer. The circuit is shown in fig. 9.

The dc hookup is about the same as in the i-f amplifier in fig. 7. The collectors are returned to ground through T3 and T6, and the transistors share a common emitter resistor (R45) and bypass (C43). Base bias is developed by dividers from the 12-volt line to control lines A and B.

In the transmit mode, Q7 is on; it amplifies the sideband signal, sending it on to the hf mixer and ultimately to the power amplifier and antenna. During the receive mode, Q8 is on; it amplifies incoming rf and sends it to the mechanical filter.
But the purpose of this stage is to add the vfo signal to the i-f signal. The vfo operates between 5457 and 5707 kHz. Transformer L4 couples its output to the emitter-base circuits of both Q7 and Q8.

For transmitting, the sideband signal passes from the mechanical filter through T3 to Q7. There, it is heterodyned with the vfo (5457-5707 kHz). The new difference frequency is between 3175 and 3425 kHz, and this signal is applied to T6. Of course, T6 accepts only the specific difference frequency produced by the vfo. It’s tuned by the A and B sections of C48, which are on the same shaft as the vfo’s tuning capacitor.

For receiving, the system works in reverse. The incoming rf is in the range of 3175 to 3425 kHz. It’s tuned by C48 and T6 and passed to Q8. There, it is heterodyned with the vfo signal, producing a difference frequency of 2289 kHz, which is fed to the next mixer stage.

**Product detector**

In the Gonset 910A, a transistor is the product detector for ssb (a diode is the a-m detector). The emitter-follower circuit is shown in fig. 10.

The rf sidebands around 9 MHz are brought through T3, whose secondary is a low-impedance link that matches the low base impedance of Q7. C16 keeps dc out of T3 while coupling the rf to Q7’s base. A steady signal from the 9-MHz crystal oscillator is also fed to the base of Q7, where it heterodynes with the sideband signal from transformer T3.

The collector of Q7 returns to ground...
through resistor R23, and is grounded directly for rf by C19. The emitter is fed dc from the 12-volt line through load resistor R24. C131 eliminates any rf that is "hanging on," and C20 couples the audio to the volume control. Base bias is developed by divider R20-R21.

Q8 rectifies the audio signal and develops a negative dc output across the R125, R126, and part of R122. C138 and C139 bypass what audio is left, and the negative dc voltage is used to bias the grids of the rf and i-f amplifiers.

Manual gain control for both rf and i-f stages is provided by R122. The arm taps off a portion of the voltage from the negative 110-volt line, and applies it to the grids of the rf and i-f amplifiers through R126. Only if the dc output of Q8 exceeds the amount set by R122 does the agc take effect.

fig. 8. Rf stage uses tapped coils to match transistor impedances.

fig. 9. Two-directional mixer stage with common oscillator for both receive and transmit.

agc amplifier

One transistor agc circuit, from the Galaxy V Mark 2, is shown in fig. 11. Audio signals are fed through C137 from the product detector to the base of Q8. This is an npn stage which requires positive collector voltage. But the collector must develop a negative voltage to apply to vacuum-tube grids (the set is a hybrid).

If the collector must be somewhat negative with respect to ground, the emitter and base must be even more negative. As you can see, a 100-volt negative line is tied through R124 to the base (through R121) and the emitter (through R123). The rf gain control (R122) forms the bottom part of the divider from the 100-volt negative line. All this makes the collector positive with respect to emitter.
bias regulator

National's NCL 2000 is a linear power amplifier in which a pair of power tubes in parallel furnish 2000 watts PEP when driven by an ssb exciter. For proper sideband operation, the PA grid bias must be held constant despite varying line voltage or load changes. Fig. 12 shows the circuit that accomplishes this task.

A dc supply furnishes 90 volts negative directly to the collector of the regulator transistor. This —90 volts is also tied, through R39, to the emitter of Q2, the error amplifier. Q2's emitter is held at —18 volts by zener diode D9, as a steady reference. C17 and C12 bypass any transient voltages that might be too quick for the zener diode.

The collector of Q2 is also tied to the 90-volt negative line, through R38. C6 and C8 bypass any transients that get into this branch. Q2's collector is more negative than its emitter or its base (which goes to ground through Q1, and through the sensing divider R40-R41-R42 to ground. This places Q1's emitter—and the PA grid-bias line—normally somewhere between —25 and —45 volts with respect to ground.

If the voltage on the grid-bias line changes, it affects voltage at Q2's base. Q2 amplifies this change and inverts it at the collector.
Thus, any error signal is applied to the base of Q1. The emitter-collector conductance of Q1 changes and restores the original voltage at Q1's emitter. The normal bias value is adjusted manually by R41, which sets the operating point for the base of Q2.

**future of transistors in ssb**

That's a real handful of transistor circuits. Most sideband functions, you have seen, can be performed by today's transistors or diodes. You may have noticed that no rf power stages were mentioned here. This lack will soon be filled. New silicon “overlay” transistors promise up to several hundred watts and can work beyond 50 MHz. Some are available now, but at high cost. In time, though, all but the very highest-powered ssb transmitters will be completely solid-state.

**drive indicator**

The Drake R-4B receiver is often used with a companion transmitter. When it is, the receiver vfo is used to excite the transmitter. The frequency is then identical for receiving and transmitting.

But if the transmitter were fired up without rf drive, the final tubes could be damaged. A neon lamp might be used to monitor rf drive directly, but the rf level is too low to fire a neon. So, a transistor switcher is used instead, as shown in fig. 13.

R150 and R151 form a divider from the 150-volt line to ground. Q10 is an npn grounded-emitter transistor, and is not normally conducting, so it has a high emitter-collector resistance. The lamp stays off.

When the vfo operates, rf from the first mixer stage is rectified by D15; a positive voltage develops across R153. This dc is applied through R77 to the base of Q10. (D17 is a protective diode which shunts any negative peaks to ground.) The transistor is biased into conduction and becomes a low resistance from its collector to ground. The neon fires.

---

**fig. 12. Two-transistor circuit regulates power-tube grid bias.**

**fig. 13. Transistor switch turns on neon lamp when rf is present.**
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You must still select these semiconductors for the higher frequencies—but it's worth it.

**Superior gain characteristics** of the metal-oxide semiconductor field-effect transistor (MOSFET) have been exploited in ham-band converters for six and two meters in *Ham Radio*'s June and August 1968 issues.\(^1\)\(^2\) Although its gain diminishes slightly between 200 MHz and 300 MHz, the MOSFET—particularly the dual-gate type—is still a fine performer at 220 MHz. The deluxe combination of low noise, high gain and low cross-modulation are packaged in the following solid-state converter, which is reasonably easy to construct and tune.

**Circuit considerations**

The 1¼-meter band is more demanding of semiconductors than are the lower frequencies. Only by selecting high-quality components was it possible to achieve the desired level of performance. Much to my surprise, the oscillator and multiplier transistors are equally as critical as those used in rf stages.

The first rf amplifier uses an RCA 3N159 dual gate MOSFET, which is a low-noise version of the RCA 3N140. In using the 3N159, you achieve the lowest possible noise figure, about 2.5 dB. Very low feedback capacitance...
—less than 0.02 pF—allows operation with no neutralization. The second stage, being less critical, uses the 3N140 operating at lower gain to prevent overload.

Two rf stages may be stagger tuned if wideband operation is desired. Unlike the two-meter converter, which had to be stagger tuned, this converter can have both stages peaked on the same frequency. This added utility is made possible by using dual-gate MOSFET's in both rf stages, with Teflon sockets, to minimize feedback capacitance.

Only one change is made in the 3N141 mixer circuit from that in the 6- and 2-meter converters. It is the addition of a neutralizing circuit from gate 2 to ground. The L and C resonate at 14 MHz, effectively bypassing gate 2 at the intermediate frequency. (Values must be changed if output other than 14 MHz is employed.) Incoming signals will be improved one S-unit or more with this circuit. The technique may also be used on the two-meter converters.

Birdies and "garbage" on 220 have not helped to make it a popular band. One way to improve the situation is to use a high-frequency local oscillator as was done here. A 2N3478 drives a 103-MHz fifth overtone crystal whose frequency is doubled in a second 2N3478. Cheaper transistors did not deliver enough injection voltage for good gain in the mixer.

I find that the use of back-to-back diodes in the antenna circuit is quite controversial. Some feel they are essential to prevent burnout of the front-end transistors, while others say this is not so. My friend Ed, W2DMR, uses a 3N140 preamplifier with no diodes on ten meters with a kilowatt. His only isolation is a DK60G relay. He operates within 1/4 mile of another kilowatt on the same band without problems. I have removed the diodes in all my converters without mishap, but 10 watts of rf is the greatest power I've generated on any band. Diodes are shown in the schematic for those who care to use them.

Sometimes rf burnout has been traced to slow decay of the transmitter's output, rather than the lack of isolation of the relay. In this case, large amounts of rf may load the converter during coaxial relay switching. A very good solution to this problem is found in delaying the relay, or by using two sequentially operated relays—one for the transmitter; one for the receiver.

construction

As with the 6- and 2-meter converters, all components were mounted on copper-clad printed-circuit board according to the layout shown. The board is fastened to a Bud CB1626 chassis. Satisfactory operation is once again possible without shields. For those who may have forgotten the rules for handling MOSFET's, the following should be observed:

1. Keep MOSFET leads shorted until ready to use. (These devices are shipped this way.)

2. When cutting leads, grasp lead and case with fingers to reduce possibility of electrical and mechanical shock.

3. Do not solder or change components with MOSFET's in their sockets. (Such a practice may be acceptable if you use a soldering iron with a grounding system.)

4. Never insert or remove transistors when power has been applied.

performance plus

A comparison of this converter was made with my Nuvistor converter whose noise

* Depends on the transistors. Quality control and inspection mean different things to different manufacturers. Better use the diodes and be safe. Editor.
L1 2½ turns number 20, spaced winding, ¼" diameter; tapped 1 turn from ground

L2, L3 2½ turns number 22, ¼" diameter

L4 0.68 μH (J. W. Miller 9310-08)  L6 1.5 μH (J. W. Miller 9310-16)

L5 3 μH, 17 turns number 22 on 3/8" diameter  L7 0.22 μH (J. W. Miller 9310-02)

L8 3½ turns number 22 on 3/8" diameter slug-tuned form, red core (J. W. Miller 4300-3)

103-MHz crystal is 5th overtone (International Crystal FM-1-GP)

All 10-pF trimmers are JFD type VAM-010 (Allied Radio 43F345)*

fig. 1. Schematic diagram of the 220-MHz converter. Capacitors marked BM are button micas; bypass capacitors are plugs or laminated ceramics.

*Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60660. Order catalog number 43F3475. $3.95 each plus postage: shipping weight, 2 ounces.
The noticeable quieter MOSFET's are music to my ears. I find that the gain is comparable, which is not an easy achievement with the best of the current JFET's.

While 220 is an orphan band in many parts of the U.S., the Mt. Airy VHF Society has kept it alive in the Philadelphia area. Only one station has managed to overload my MOSFET converter. (He is hard on the tubes too!) Admittedly, this is no test for comparison, but I feel the MOSFET is at least as good as the proven two-rf-stage 2-meter converter\textsuperscript{2} with respect to cross modulation and dynamic range.

**mosfet's and the future**

Only a few of the semiconductor manufacturers are manufacturing MOSFET's, and not all are interested in the depletion-mode types for vhf such as the RCA transistors used in this series of three articles. Stability problems of the oxide layer which once plagued the MOSFET have been largely solved, giving time for curing the ills of static burnout and frequency limitations.

MOSFET amplifiers at 500 MHz are practical now, but high-volume production is not in effect on these units. Some feel the

*Bud CB-1626 chassis available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog number 42E7812, $5.55 plus postage; shipping weight, 12 ounces.*
upper frequency limit of the MOSFET is 1000 MHz, but science may soon disprove that notion. Even now, Motorola has incorporated a layer of silicon nitride to eliminate static burnout of the oxide layer.

Fairchild has integrated additional circuitry in the MOSFET package to reduce the burn-out hazard, while other manufacturers are experimenting with still different techniques. Presently, there is no simple solution compatible with all MOSFET's, but the fact also may be changed in the near future.

The metal-oxide semiconductor field-effect transistor answers the challenges of high capacitance, high gain, low noise, inexpensive construction, wide dynamic range and low cross modulation. It is so distinctive that it invites comparison in all rf and many dc applications.

references

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- Built-in dual calibrators (25 and 100 KHz)
- Built-in Clarifier (off-set tuning)
- All crystals furnished 80 through the complete 10 meter band
- Provision for 4 crystal-controlled channels within the amateur bands
- Provision for 3 additional receive bands
- Break-in CW with sidetone
- Automatic dual acting noise limited
- and a sharp 2.3 KHz Crystal lattice filter with an optimum SSB shape factor of 1.66 to 1.

Design features include double conversion system for both transmit and receive functions resulting in, drift free operation, high sensitivity and image rejection.

- Switch selected metering
- The FT dx 400 utilizes 18 tubes and 42 silicon semi-conductors in hybrid circuits designed to optimize the natural advantages of both tubes and transistors.
- Planetary gear tuning dial cover 500 KHz in 1 KHz increments
- Glass-epoxy circuit boards
- Final amplifier uses the popular 6KD6 tubes.

This imported desk top transceiver is beautifully styled with non-specular chrome front panel, back lighted dials, and heavy steel cabinet finished in functional blue-gray. The low cost, matching SP-400 Speaker is all that is needed to complete that professional station look.

SPECIFICATIONS:
- Maximum input: 500 W PEP SSB, 440 W CW, 125 W AM.
- Sensitivity: 0.5 uv, S/N 20 db.
- Selectivity: 2.3 KHz (6 db down), 3.7 KHz (55 db down).
- Carrier suppression: more than 40 db down.
- Sideband suppression: more than 50 db down at 1 KHz.
- Frequency range: 3.5 to 4, 7 to 7.5, 14 to 14.5, 21 to 21.5, 28 to 30 (megahertz).
- Frequency stability: Less than 100 Hz drift in any 30 minute period after warm up.

CLARIFIER CONTROL — Does the work of an external VFO — allows operator to vary receive frequency 10KHZ from transmit frequency, or may be used as an extra VFO combining transmit and receive functions.

SELECT CONTROL — Offers option of internal or outboard VFO and crystal positions for convenient preset channel operation.

FUNCTION CONTROL — Selects crystal calibration marker frequency and desired transmit mode of operation.

FT dx 400 $599.95 — SP-400 $14.95
A simple multiband antenna system that features automatic stub matching as you change bands

Transmission lines can be used for various purposes besides transferring power to an antenna. Depending on the electrical length of the line, it can function as a frequency-dependent switch, an impedance-matching transformer or a reactive circuit element. You can devise a number of interesting antenna designs with the first two functions. Normally these characteristics are used separately, but there is no reason why they can't be combined.

Before you digest any designs, it's important to have a clear idea of how transmission-line sections work as impedance-matching transformers and frequency-dependent switches.

Transmission line characteristics

A lossless transmission line one-quarter wavelength long will transform the impedance at its input terminals to an impedance at its output terminals equal to

\[
\frac{(Z_0)^2}{Z_{in}}
\]

where \(Z_0\) is equal to the impedance of the line.

However, this is only true when the input...
or output impedance is a pure resistance. Although the whys of this impedance transformation can be shown mathematically, the proof is somewhat tedious. Perhaps the simplest way to visualize the action is to remember that the input voltage and current vectors undergo a 90° phase shift on the quarter-wave line. Therefore, their relative amplitude values are reversed. If the input impedance is lower than the line impedance, the output impedance is always higher. If the input impedance is greater than the line impedance, the output impedance is always lower.

Any number of quarter-wave sections can theoretically be used in series if one particular line does not provide the desired transformation. If the transmission line is one-half wavelength long, the output impedance is the same as the input impedance since this is the same as putting two quarter-wave sections back-to-back.

The limiting case occurs as shown in fig. 1 when the input impedance is either zero (short circuit) or infinite (open circuit). It can be seen from the impedance-transformation formula that the output impedance must be opposite from the input impedance. With this in mind, we can use the quarter-wave transmission line as an rf switch.

However, the switch is frequency dependent. If the input impedance is zero at a frequency where the transmission line is a quarter-wavelength long, the output impedance is infinite and looks like an open circuit. If line length is fixed and the frequency is doubled, the transmission line represents one-half wavelength and the output impedance is the same as the input impedance.

Therefore, by controlling the termination and length of the transmission line, it can be used as an impedance transformer, a 1:1 impedance transfer element or a frequency-dependent switch.

**dipole harmonic operation**

A simple resonant half-wave dipole presents a pure resistive impedance which matches 50- or 70-ohm cable fairly well. If the dipole is used at harmonic frequencies, its terminal resistance and reactance will vary as shown in fig. 2. At even multiples of a half-wavelength, the terminal impedance is highly resistive with practically no reactive component; at odd multiples, the terminal impedance is resistive at almost the same value as at the fundamental frequency with no appreciable reactive component. This is the reason you can use a 7-MHz dipole successfully on 15 meters with the same feedline.

To feed maximum power into a dipole at any multiple frequency of its fundamental, you have to match the antenna's terminal impedance. The usual way to do this is to use a resonant transmission line and an antenna coupler at the transmitter. By using combinations of quarter-wavelength transmission line sections at the antenna, however, it's...
possible to obtain multiband operation without an antenna coupler or any tuning or bandswitching circuits. The transmission line going to the transmitter will still operate at a low SWR on each band.

**Some basic antennas**

Once you understand how the transmission line sections can be used, you'll undoubtedly be able to come up with designs of your own. As a starter, here are some basic designs I have investigated.

A dipole which automatically bandswitches from 80 to 40 meters is shown in fig. 3. On 80 meters, the open-circuited quarter-wavelength sections (shown horizontally) between points X and Y effectively short out the vertical quarter-wavelength section. This effectively connects points X and Y together as in a normally fed dipole.

On 40 meters, the flat-top portion of the antenna is a full wavelength long with a rise in input impedance as shown in fig. 2. The horizontal transmission line section which was one-quarter wavelength long on 80 meters is one-half wavelength long on 40 and presents an open circuit between terminals X and Y.

The vertical quarter-wavelength section is effectively connected between terminals X and Y; this section acts as a transformer, and its characteristic impedance is chosen to match the transmission line to the transmitter.

Normally, if 50- or 70-ohm coaxial cable is used, the matching section can be made from 300-ohm twinlead. The characteristic impedance of the horizontal transmission line sections aren't particularly important since they only perform a switching function.

The antenna operates as a normal dipole on 40 meters; the main radiation remains broadside to the line of the flattop with about 1.8-dB gain.

Variations of the same antenna for different bands are shown in fig. 4. The antenna of fig. 4A is designed for 20 and 10 meters.
and is simply a scaled version of the antenna shown in fig. 3. The antenna of fig. 4B is for 15 and 10 meters but is a bit different. Since you can't cut a half-wave dipole for 15 meters and use it as a full-wave dipole on 10, a different approach is used. The dipole is cut one-half wavelength long on 10.5 MHz; this makes it a full-wave dipole on 15 and approximately three half waves long on 10.

If you analyze the switching section, you'll find it produces a short circuit on 10.5 MHz, an open circuit on 15 meters and a short circuit again on 10 meters. The antenna is matched on 10.5 MHz and while this feature has no amateur value, it may be useful for WWV reception. Because of the length of the flattop, the antenna has about 1.8-dB on 15 and 10 meters. However, on 10 meters the radiation pattern takes the form of a cloverleaf characteristic of a 3/2-wave dipole.

The antenna shown in fig. 4C is a little more conventional although it is basically the same as fig. 3. Operation is possible on three bands—40, 20 and 15 meters—because the flattop is 3/2-wave long on 15 and the transmission-line switch shorts out the matching section.

The use of transmission line sections doesn't have to be limited to dipoles; they can be used just as easily with unbalanced vertical antennas. The antenna in fig. 5A is basically one half the antenna shown in fig. 4C. Although it resonates on 40, 20 and 15 meters, it is mainly useful on 40 and 20 since the radiation angle is quite high on 15 meters. This is because the vertical radiator exceeds consideration (typically 70% for 300 ohm twinlead). The line sections shouldn't be formed into coils and placed at the center of the flattop since this will introduce spurious resonances. It's better to bring both ends of the sections to the center of the antenna in the form of a drooping “U” as shown in fig. 6 and then make all the interconnections.

The values shown in fig. 2 are typical for wire antennas made from number 12 or 14 wire. If the antenna is made from very thin wire and used on 80 meters, the impedance values at harmonic frequencies may be higher than those shown. In this case it's a good idea to put the antenna together temporarily without the switching sections and operate it on the harmonic frequency; then measure the SWR to determine if you've picked the right impedance for the matching section.

You should have no difficulty with the switching sections if they are cut to formula length. The power handling capability of the antenna is determined by the rating of the transmission line used for the line sections.

**construction**

You should observe a few precautions when building any antenna. For one thing, the physical length of the transmission line must take the velocity factor of the line into consideration.
solid-state
current-controlled
tuning

Coming for the most part as fallout from military research and development—where they have been used for almost ten years—are a fascinating line of variable inductors. These inductors are electrically variable, but unlike varicaps, they alter the inductance of a circuit as a control current is varied from 0 to 60 mA.

How to use a new electronic component that adds versatility to a number of circuits.

Cutaway of a typical Vari-L current-controllable inductor.
One of the circuits developed by the Vari-L Company* to demonstrate typical applications of the Mite series variable inductors is shown in fig. 1. As the control current is varied from 0-60 mA, this super-regenerative receiver's detector tunes from 20 to 50 MHz. The audio output can be fed to a simple transistor amplifier. While the inductor shown presently sells for $15.80, the manufacturer expects production costs to fall appreciably as time goes on.

(at zero control current). With the 100-mH units, frequency bands of 3 through 30 kHz can be covered; with the .05 µH inductor, a 250-300 MHz range can be realized.

When you use these inductors, it's recommended that the signal winding be connected to the tuned circuit the same way as any conventional high-Q coil. Next, it is resonated with a suitable tuning capacitor, and the control winding is connected to a source of continuously-variable current (such as a battery and a rheostat).

fig. 1. Superregenerative receiver using a Vari-L coil. Receiver tunes from 20 to 50 MHz as the control current is varied from zero to 60 mA.

**how the inductor works**

These tiny variable inductors are two-part assemblies. The control winding is wound on the support legs of a U-shaped ferrite or laminated core. The signal winding is wound on a ferrite core positioned over the open end of the U; this winding is used as a tunable coil in a Colpitts oscillator or other two-terminal networks. When ac or dc current is passed through the control winding, it causes the magnetic flux to vary—decreasing the over-all permeability of the signal-winding core.

These Mite-series units are available with inductances from 100-mH to .05 µH.


By varying the current, the operating frequency is smoothly controlled. A tracked chain of several units can be controlled equally well—tuned to the same or harmonically related frequencies. The characteristic curve of a typical Mite inductor is shown in fig. 2.

**applications**

Since the inductance can be varied slowly or at high speed rates—up to several million times per second—command can come from just about anywhere. It can be a manually-operated source, an electronic circuit or a transducer. They are useful for remote control or closed-loop systems, for sweeping oscillators or passive filters and...
for frequency switching.

These variable inductors have been used in receivers, both in local oscillators and gang-tuned rf amplifiers, sweep generators of all types from audio to vhf, frequency modulators for transmitters, ultrasonic test generators and spectrum analyzers. Since the unit is sealed, ruggedly built and without any moving parts, field failures are unknown. Early versions are still going strong after 12 years according to the manufacturer.

Additionally, Vari-L is now concentrating on the development of new uhf variable-inductor designs and resonant-cavity units which will extend the useful region of the devices into the gigahertz range. This should be of particular interest to uhf enthusiasts looking for better ways to remotely control their plumbing installations, such as preamps/converters that are employed at the feedpoints of parabolic dishes and corner reflectors.

One vhf application is shown in fig. 3.

![Fig. 3. Oscillator that tunes from 170 to 240 MHz.](image)

This oscillator tunes from 170 to 240 MHz. The frequency is varied by the 60-mA control current fed to the MP-1 inductor at terminal three. The output link is a conventional hairpin located close to the tank and adjusted for maximum output at the frequency desired. The 2.5-pF variable capacitor may be a fixed value if only 220-MHz operation is desired. Do not substitute other transistors for the 2N1493.

Another interesting application is shown in fig. 4. This circuit provides one milliwatt into a 50-ohm load in the 260- to 380-MHz range. A Fairchild 2N3563 is recommended by the manufacturer, although a 2N918 can be substituted with good results using the same circuit constants. Usual construction techniques are recommended to insure oscillation over the entire frequency range. Further, a shielding can is recommended for the Mite MP-1.
fig. 4. A 260- to 380-MHz vfo developed by Vari-L engineers to demonstrate the versatility of the MP-1 inductor.

Lastly, fig. 5 shows how a Vari-L MK-100 inductor can be used in a 5 to 75-kHz continuously variable oscillator. With the 2N2712 transistors shown, output will be on the order of 0.1-volt rms with three-percent distortion. Unlike previous examples, this circuit requires a somewhat lower control-current range—zero to 15 mA. Other transistor types can be used with good results, including the 2N4124, 2N4265, HEP-54 and GE-10.

In conclusion, it should be noted that Vari-L products are presently not widely available at electronic distributors. If you're interested, it's suggested that you contact the manufacturer directly for literature. Vari-L will respond to interested amateurs and will ship single-unit quantities by first-class mail.

references

fig. 5. Variable oscillator that tunes from 5 to 75 kHz.
some notes
on
cubical quad
measurements

Much has been written, and probably much remains to be written, about cubical quad antennas. A review of measurement methods for these popular antennas seems in order for those desiring to improve the performance of their quads and for those who wish to build one from scratch.

The basic beam consists of a driven element and reflector. Measurement data are given here for arrays with n-elements, however, for those interested in a larger antenna.

Test equipment includes:
1. Accurately calibrated receiver
2. Grid-dip oscillator
3. Antennascope (rf bridge)
4. Swr bridge
5. Signal source capable of providing a few watts at the frequencies of interest.

The accurately calibrated receiver is used to check the grid-dip oscillator to obtain a frequency reading within one-half percent. The signal source may be your transmitter or transceiver if power can be reduced to a few watts.

driven element resonant frequency

First, disable all elements except the driven element. Open these elements and fold the wire back over itself. If you don’t do this, you’ll get several readings on the GDO when it’s coupled to the driven element. Now loosely couple the GDO by using a small “gimmick” made from a male coax connector and a short loop of wire. Solder one end of the wire to the connector shell, and connect the other end to the center pin, making a loop about a half-inch in diameter. The gimmick replaces the coax on the driven element for the measurement. Next, raise the antenna as high as you can conveniently reach to make the measurement. The plane of the antenna should be perpendicular to the ground. Find the resonant point, using the GDO with very loose coupling, then check the GDO with your calibrated receiver. This should give an accurate measurement of the driven element resonant frequency. If there are other elements to be measured, proceed in the same manner. Keep a record of the data.

Cubical quad driven elements are cut according to the relationship

\[ L = \frac{1000}{f} \]

where

- \( L \) is the length of wire (feet)
- \( f \) is the resonant frequency (MHz)

Some manufacturers use a constant of 1005, but the difference, in terms of total percent length, is insignificant for amateur work.

On 20 meters, the wire length should be changed 5.7 inches for each frequency change.
of 100 kHz. In other words, if your driven element resonates at 14.2 MHz, and you want it to resonate at 14.3 MHz, subtract 5.7 inches from the total length, or about 1.5 inches from each side (4 X 1.5 = 6, which is close enough).

**reflector resonant frequency**

If coils are used in the reflector to lower its frequency, simply couple the GDO loosely to the coil, and measure \( f \); it should be 3 to 5 percent lower than the driven element at the point chosen for its operating frequency. If a stub is used to lower the frequency, couple the GDO to the stub. If you can't get a satisfactory reading by this method, then coil up the wire in the stub to make a small loop, and couple the GDO to the loop. The loop should be a single turn, or even a half turn, to get the correct resonant frequency. If \( f \) is obtained by increasing the wire length 3 to 5 percent, follow the procedure given above.

**director resonant frequency**

An identical procedure is used for checking director resonant frequency. These elements should resonate from 3 to 5 percent higher in frequency than that of the driven element. (I use 5 percent.) It's possible, if more than one director is used, to resonate the first director at 5 percent, the second at 4 percent, etc. No data have been obtained concerning shortening each director from the previous one by a fixed amount, but yagis are often constructed in this manner, so the same reasoning might apply to a quad.

**input impedance**

Couple the antennascope or rf bridge to the GDO, and couple the output of the rf bridge to the driven element. Set the GDO at the desired frequency, and tune for a minimum dip on the rf bridge. Next, read the input impedance on the rf bridge. A curve may be plotted, if desired, of frequency versus input impedance. It will give you a good idea on how flat the line really is. Remember, loose coupling is used in all measurements in order not to affect accuracy due to oscillator pull.

The coupling line between the rf bridge and the driven element should be as short as possible, not more than six inches to a foot, and should be coax cable of the same impedance used to feed the quad. Also the reflector and director must be in place, and connected, for these readings. The quad should be in exactly the same condition as it will be when used, except for height above the ground.

**front-to-back ratio**

Raise the quad in a vertical position above ground and as high as possible. You must still be able to reach the reflector coils, or the stub, whichever is used. Next, energize the quad at the resonant frequency. Point the quad away from where the measurements will be taken. This may be either a cooperating amateur a few blocks away, or your own receiver a few blocks away, or a field strength meter a few wavelengths away.

Now, adjust the stub or the coils (number of turns) until a minimum reading is obtained on the receiver or field strength meter. Measure the reflector frequency again, and if it isn't within 3 to 5 percent range there is something wrong with your measurements or your method of measuring.

**standing wave ratio**

In all cases, when using an swr bridge, first set the bridge "wide open," or at its most sensitive point. Then feed in just enough rf to get a full scale reading in the forward direction. Next, reverse the bridge, and read the swr. It's not necessary to hoist the quad in place to make these measurements. Quads have no open ends and thus are quite insensitive to nearby objects.

If the swr bridges were perfect, it wouldn't matter whether the bridge operated at the full-power output of the transmitter or at very much reduced power. Measuring instruments are not perfect of course, hence their limitations must be taken into account. Remember, too, that when the swr bridge is set to read reflected power, it also reads a small fraction of the incident power, and any stray coupling will affect the accuracy of the reading. Also note that the true swr will always be lower, but never higher, than the instrument indicates.

A theoretically perfect swr bridge would read the same regardless of the amount of
power fed to it. In a way, therefore, the difference between the full-power reading and the reduced-power reading is somewhat of a test for the quality of the bridge assuming that no power is leaking to the reverse diode.

Now, let's assume that the actual mismatch is 2:1 between the feed line and the antenna. This is the maximum you can read on the bridge regardless of where you insert the bridge into the line. If there is in fact a mismatch, you'll obtain a different reading everywhere along the line. If the true input impedance is 50 ohms, then if you measure one-half wavelength from the input point, taking the coax propagation factor into consideration, the measured impedance will be 50 ohms.

At one-quarter wavelength from the input point the measurement would be either 25 or 100 ohms, depending upon whether or not the quad input impedance was higher or lower than that of the cable. However, at exactly one-half wavelength from the feed point, it would be immaterial what cable impedance were used as long as the propagation factor is accounted for in measuring the one-half wavelength.

The reason for this is that the reflected voltage, traveling back along the line from load to source, meets the incident voltage. If these voltages meet at multiples of one-half electrical wavelength on the line, the voltages will add, because at that instant they will be exactly in phase. It is possible, with certain line lengths, to get a reading of 1:1 when the actual swr is considerably greater than this. The only way to be sure is to use several different line lengths of the same cable, and if the swr comes out approximately the same, then you're reasonably sure that the swr is being read correctly. With a perfect match the swr will be 1:1 regardless of where the meter is placed.

**optimum element spacing**

According to reference 1, the optimum gain occurs in a quad with 0.125-wavelength spacing between driven element and reflector. If the spacing is increased to 0.25 wavelength the gain falls off about one dB. There is an infinite number of spacings that may be used on any beam, but I found that 0.125 wavelength between reflector and driven element is indeed optimum. (The input impedance is about 50 ohms as well.) I also found that a 0.1-wavelength spacing between driven element and director, or directors, leaves little to be desired.

It's true that many successful quad users space the elements at considerably greater distances with possibly some advantages. But from a practical standpoint, spacings as indicated above have many advantages also, such as a reasonable size structure that can be rotated with a rotator of moderate strength, and one that can be put on a tower of the ordinary ham variety.

To obtain the excellent performance of a multielement quad (unless you want to use a boom of 25 to 50 feet), spacings of the order of 0.125 wavelength between driven element and reflector, and 0.1 wavelength between driven element and directors, are quite in order.²³

**references**

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The easiest way to operate on two meters is to buy a "Two'er," select a crystal for the emergency frequency, then hope someone in the nearest large city will hear you. Well, with a good beam you might be heard; then again, if band conditions are poor—forget it.

I tried to buy a linear for my Two'er that would up the output at least ten times, i.e., to 30 watts nominal. None was available that would preserve the audio quality Heath is known for and still provide a respectable signal in the nearest town (which is some 83 miles away) with anybody on two.

Obviously, to get the most from the audio in the Two'er, $\text{AB}_1$ operation is a must. So I built such a linear using a 7984 pentode with series-tuned input, ferrite choke loading instead of resistor swamping and a toroidal-tank output. It gave the needed punch, and now I'm heard every day even with 20-dB tropo fadeouts, if the guys on the other end are using at least one Nuvistor ahead of their receivers. The circuit described is a refinement of an earlier amplifier using a novel input system that puts the power where it's needed.
development

Some of the problems I encountered involved the old TVI hangup. Living in a deep vhf TV fringe area, it’s impossible to operate even class B on two meters. An exception might be a rig with a class-B single-balanced modulator with very little carrier—but even then some TVI would be a problem. In my case, however, uhf TV reception in Lexington is so popular that a dsb rig with a kW PEP final is feasible, but only because people hereabouts switch from vhf fringe channels to local uhf TV. Of course, anything like class C is “verboten.”

I had a surplus of 125-V, 50-mA power transformers and several E. F. Johnson variable “M” capacitors. Most of the hardware was scrounged from other projects that never got off the runway. This included two UG-1094/U BNC connectors, 1000 PIV rectifiers rated at 750 mA and a Bud CU-622 Convertabox for the chassis.

Having built the linear “à la haywire” six months before, I tossed out the phenolic Compactron socket for a new mica-filled one, doubled the number of transformers, and used a newer Convertabox. Everything at last fit well, and most of the hole centers were eyeballed-in using a steel scriber and fierce concentration on a 10c plastic pocket ruler. Everyone around Lexington said my old 7984 bomb was loud, but the newer version was even better, and no one at either time could hear and see me!

Except for the parallel-tuned toroid output, over-all design of the two-meter linear is
straightforward, and the series input circuit is conventional except as noted below. If desired, ceramic coil forms could be used and a physically "linear" quarter-wave line inserted in place of the IRN-9 toroids; but toroids keep the circuits small, and more power is coupled into the coaxial output and not into the wiring.

I tried a high-Q quarter-wave line and got less out. There is a limit, though, to the use of toroids on two meters. After trying unsuccessfully to get more than 25 watts measured output from a high-power PL-177WA amplifier, I have bowed a bit to the conservative by observing that a practical limit (tube dissipation basis) is 35 watts for ICAS operating conditions. This is all that seems worthwhile for AB, with the 7894. I did heat up some larger cores quite a bit, using the PL-177WA, but this tube is 40 watts "larger."

A new circuit feature is claimed here. As you read on you will notice that a grid-loading control is provided, saving lost watts in the output by peaking the control grid without a swamping resistor. This is really the most important part of what I have to say under "development," because by removing that swamping resistor and loading the grid only lightly you get very high efficiency in AB, and drive high enough to push the 7984 to its limit without an insertion amplifier.

In fact, before I accidently found that series tuning would give an rf voltage peak of the correct impedance at the control grid, I thought an amplifier would be necessary. Not so! But don't forget the trick is to center-tap the series inductance and load with a Ferroxcube VK-200-10/3B semiconductor low-z ferrite choke.*

**layout**

Looking at the photo, from left to right we see the input connector, 30M11 input capacitor and the slug-tuned grid-loading inductance coil form. Next is the blower and 7984. Just to the right is the output BNC connector, output tuning and bias pot. The rest of the controls are as labeled.

The underchassis view shows the input tuning control; rear, Ferroxcube choke and series inductance, shield partition and heavy-duty ceramic coupling capacitors (the only kind that would handle the circulating current) and trifilar output tank circuit wound on two Permacor 3/8-inch IRN-9 sample toroids.** The rf components occupy only a small fraction of the chassis space. The rest of the chassis is used for the voltage-doubling power supply and bias-multiplier filters.

A series-limited neon lamp is used as the pilot and is held with a large Tinnerman speednut. The on/off switch is special in that it mounts in a 3/8-inch hole and not in the usual 7/16-inch or larger hole, which saves buying a punch for just this size.

* Order from Mr. R. Worban, Ferroxcube Corporation of America, 360 E. North Avenue, Northlake, Illinois 60164. 11c each plus shipping.

**Permacor Division of Radio Cores, 9540 South Tulley Avenue, Oak Lawn, Illinois 60453; Number 57-6075 IRN-Q material.
Button bypassing is used on both screen leads and also for plate circuit decoupling. These 240-pF buttons were salvaged from a surplus ARC-3. The plate choke is 2.7 µH, chosen for a cutoff of 149 MHz, with preference for the higher reactance obtained using the Miller chokes as opposed to the Ferroxcube. Remember that the VK-200 series are lossy when used with vacuum tube circuits for other than this application in swamping the grid. Their use in transistor circuits is a notable exception, where impedances are very low. Twenty pF was the value chosen for plate coupling, since two of these will have a series-resonant mode near the two-meter band.

Incidentally, I tried to use "dog-bone" capacitors and disc ceramics with no luck. They all burned up. When I switched from these to the more expensive ceramics I used pins 3 and 4 of the 7984 because they carry the bulk of the rf load. Pin 5 is fine for dc coupling.

Fig. 3. Different input circuits discussed in the text: Inductive division in A, capacitive division in B.

The circuit

Fig. 1 shows the voltage-doubler power supply. Screen regulation is considered a myth for AB1 operation using a pentode of such excellent structure for rf. If the plate voltage should get out of hand, so will the screen voltage, tending to reduce the plate current; also there's a bleeder and voltage divider using a 20k/10W resistor from screen to ground.

The input circuit is capacity coupled using material) from center position as resonance is achieved. That is, inductive division, like capacitive division, (to quote RCA) feeds the right voltage and current at two-meter rf frequencies to match the input admittance and susceptance of the high-C 7984. Fig. 3 shows both tubes. Capacitive effects are represented by $C_{in}$ (fig. 3a), with bead choke resistance swamping.

If you'll also look further into the RCA Transmitting Tube Manual, you'll notice they use only center-tapped series-tuned circuits for vhf since strays have already limited that fraction of the input which can be capacitively divided, and varying the slug of the coil form is the best way to optimize an input circuit like this.

Other arrangements are shown for high-C input vhf tubes (fig. 3B). In general, use at least one and one-half times the input capacity of the tube for the series variable, and
always center tap the inductance. Also, use permeable cores in the form instead of the brass slug type, unless extreme bandwidth, lowered Q and extra power are available. Use of solenoid-wound rf chokes for class-C and -B operation is recommended, but use ferrite for linears that don’t draw grid current.

operation

I used this linear for over a year without a blower, and have never had difficulty with it. It has a measured output of 12 watts (with a new 7984) in AB1, and will go to AB2 if the bias is set too low. (You can tell by listening for distortion in the monitor.) For AB1 as opposed to class C, where the input power is specified, I chose plate dissipation as a more accurate gauge for rating, times the book value for theoretical efficiency, i.e., 33% of a maximum rating of 35 watts dissipation, which would be about 12 watts output.

On the air I say “50 watts,” and this can be measured as input in the usual manner. My goal is 2000 watts dc input to two 4CX1000’s in AB1. Though of questionable legality, this would give no more than 660 watts output, like the class-C boys, with virtually no TVI.

references


ham radio

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P. O. Box 1504 San Diego, Cal. 92112
more ohmmeter troubleshooting

Last month, I explained how to use your ohmmeter for safe troubleshooting. Before I ran out of space, I reviewed series and parallel paths and also told you how to get accurate readings on your ohmmeter.

I hope you learned to recognize the "odd" series-parallel paths I described. When you're tracking down a fault in some receiver or transmitter, how easy it is for you depends on how well you spot the resistance paths that aren't always obvious. This month, I'll start off by telling where you might find some of those paths.

hidden resistance paths

There are two small parts most likely to cause false ohmmeter readings: electrolytic capacitors and transistors. Electrolytics are part of all power supplies and are used for decoupling along dc supply lines. In either use, they often get in the way of ohmmeter tests. Transistors foul up ohmmeter readings because of their "diode" nature. To make matters worse in transistor equipment, many coupling capacitors are electrolytic. They mix their leakage currents with the transistor leakage currents, and you can drive yourself nuts sorting out the true paths.

Fortunately, these two paths follow known behavior patterns. If you know the patterns, you can figure out if you're being misled. Here are some of them.

1. In power-supply paths. The leakage resistance of most electrolytic capacitors is nearly 50k. Learn to allow for it when you're checking resistances along B-plus lines. The diagram in fig. 1 shows power-supply paths in one receiver schematic (simplified, of course).

Suppose you clipped the common lead of your ohmmeter to ground and the test lead to the plate of V4. Since there is no apparent path to ground, you might expect to read an open circuit, or infinity. Instead, you read around 600k. As you can guess, the ohmmeter is reading the resistance of R13 plus the leakage resistance of the power-supply capacitors.

2. Allow for diode action. Suppose the reading from the plate of V4 to ground is almost exactly 470k. What does that mean? Maybe a short in one of the electrolytics? To check, you move the test lead to the junction of R3-C4. A very low resistance there seems to mean there's a shorted electrolytic.

But before you jump to a false conclusion, reverse the test leads: put the common lead on the R3-C4 junction and the other (red) lead to ground. The low reading disappears!
It's confusing, isn't it?

What happens is that the voltage of the ohmmeter battery causes the rectifier diodes to conduct and give the low reading. Whenever the negative side of the ohmmeter battery (sometimes, but not always, the red test lead) is touched to the cathode end of the rectifier, the ohmmeter reads the diode’s forward resistance (which is low). An ohmmeter thus connected from R3-C4 to ground might read through R1-R2-R3 and the diode.

The pnp transistor is wired up as a mixer. Suppose you connect your ohmmeter across R1 to measure it. Instead of the expected 4.7k, you read about 1k and figure R1 is bad. However, if you reverse the ohmmeter leads, you measure 4.7k. How come?

The base-emitter junction of a pnp transistor is forward-biased when the base is negative and the emitter is positive. If you connect the test leads the wrong way, the ohmmeter reading includes the base-emitter resistance (100 ohms or so). It is in series with R3, and both are in parallel with R1. The result: a wrong measurement.

How about when you measure R2? There should be no diode action between base and collector. What would you think, then, if you tried measuring R2 and got a 40k reading? Another hidden path is probably the cause. Fig. 2B shows the possible cause: leakage through the power-supply electrolytic capacitor (which may not even be drawn at that location on the schematic). The path through R4, the capacitor’s leakage, and R1 lowers the resistance read by an ohmmeter connected across R2. You can thus get an

3. Consider transistor leakage. A form of diode action takes place in transistors. That is, the ohmmeter battery can forward-bias a transistor junction and make it draw current. The ohmmeter then reads the junction resistance as well as whatever circuit-path resistance you’re measuring. There’s an example in fig. 2.
erroneous reading, even if the transistor is normal.

A significant clue, though it might fool you, is the fact that reversing the ohmmeter leads in this case doesn't change the reading. The trouble is therefore not diode action.

4. Watch out for paths drawn elsewhere. The path just described is a common example. There are others harder to spot. One is shown in fig. 3.

The situation in fig. 3A is apt to be overlooked in tube circuits. The trouble is that schematics show only the common power-supply connections. If there happens to be a ground path somewhere else on the schematic, you can get a low ohmmeter reading you can't account for.

In fig. 3A, for example, if you measure from the plate of V1 to ground, the voltage-divider network in the screen circuit of V9—which may be drawn at the other end of the diagram—causes an incorrect reading. (It should be infinity.) Even if you disconnect the power-supply capacitor, the too-low reading persists. The equivalent circuit is redrawn in fig. 3B to show how the circuit looks to the ohmmeter. (R9 has no effect, since it does not go to ground.)

You'll run into this voltage-divider situation more often in transistor sets than in tube equipment. Base-bias and emitter-bias dividers are standard practice. They cause trouble only when they are in parallel with each other but are widely separated on the schematic.

5. Something else to watch for in transistor circuits is shown in fig. 4. This one can fool you for several reasons. First, there is the diode action of the base-emitter junction. Second, there are parallel paths created by the power-supply electrolytic (not shown, as usual). Third, the parallel path depends on where you connect the ohmmeter.

The four rearrangements of the circuit show what the ohmmeter "sees" as it is connected between ground and each of the four points: A, B, C, and D. Anytime you have trouble figuring out paths, redraw them the way I have these.

From A, the ohmmeter should measure either about 850 ohms or about 4.6k. The diode is the reason for the alternative. One way, the ohmmeter makes the diode conduct and measures its forward resistance plus R4; the other way, it reads only the resistance of R1, since the diode can't conduct. The resistances of R1, R2, and R3 are comparatively so high, they have little effect even though they are in parallel.

From B, the ohmmeter should measure about 35k. The diode shows little effect, because R1 and R4 are relatively small compared to R2. The ohmmeter therefore reads about the same either way. The 68k–72k is in parallel with the (approximate) 77k.

From C, there are several considerations. First of all, you don't have to worry about the effects of R2, R3, and the power-supply electrolytic. They are so much larger than R1 that the 4.7k is the only resistance that can affect that leg of the circuit. The diode action of the base-emitter junction is important, though. Connected one way, the ohmmeter
reads only the 820 ohms of R4. Connected
the other, it reads slightly less because diode
action puts R1 in parallel; around 750 ohms
seems about right.
From D, the ohmmeter mainly reads the
value of leakage in the power-supply elec-
trolytic—about 30k. The much higher values
of R2 and R3 isolate any action of the base-
emitter junction, or the effects of R4 or R1.
The condition of the electrolytic determines
exactly what the ohmmeter reads.

3. If you’re still looking for the reason for
a low reading, see if you can find any other
parallel paths drawn elsewhere on the
schematic diagram. If you don’t find any,
you are probably on the trail of the fault.

**what to do about parallel paths**
There are times when it’s faster and easier
to eliminate the effects of parallel paths or
to get around their effects some way. Once
you spot a wrong reading, track down the
guilty parallel component. You can’t always
figure it out from Ohm’s law, either.
One of my first suggestions is: Don’t make
resistance measurements with respect to
ground. Don’t clip the common lead of your
ohmmeter to ground. Make each measure-
ment between specific points in the circuit.

As an example, think about a circuit like
the one in fig. 1. You can clip the common
ohmmeter lead at the junction of R2-R3, and
measure the branches through R9, R11, R4,
etc., directly. That way, you eliminate any
effects from power-supply electrolytics. To
measure in the branch that starts through R5,
you can clip the common lead at the junction
of R5-R1 and eliminate any diode-action
problem with the rectifiers.

**fig. 3. Sometimes a parallel path is hid-
then by the way the schematic is drawn.**

**knowing the cause of wrong readings**
These general descriptions of where to
look for “hidden” parallel paths also sug-
gest ways to deal with them. When your
ohmmeter readings are different from what
the schematic leads you to expect, here are
some quick steps to run down the cause.
1. If a reading is low, reverse the ohm-
meter leads. If the reading increases, there’s
diode action in the circuit, which may be
causedit by a diode, a power-supply rectifier,
or a transistor. In unusual cases, slight dif-
fences may be traced to electrolytic capac-
itors that don’t “form” under reverse polarity.

2. If an ohmmeter reading starts low and
builds up slowly, an electrolytic capacitor is
causing it. This is the normal charging action
of the electrolytic. Reversing the ohmmeter
leads gives a lower reading, which builds back
up again to the higher one. The reading is
anywhere from 25k to 50k, depending on the
condition of the electrolyte in the capacitor.
The charging action is what you watch for.
And so on. By making point-to-point measurements, you don't have nearly so many parallel paths to consider. Also, as you move the main test lead on out a branch, it's easier to figure out the effects of added circuit resistances. Analysis is much simpler.

My second suggestion is to disconnect unwanted parallel paths when they get in your way. This happens seldom, if you plan your point-to-point measurements. But when it does happen, it's usually simple to unsolder a parallel component or branch.

On printed-circuit boards, you can often unsolder just one lead. Or, use a single-edge razor blade to slit across the foil and disconnect a circuit branch; that's easily restored afterward by a solder "bridge" across the tiny slit.

My third suggestion is to pull a tube or transistor out of its socket if you think it's fouling up your attempts to interpret resistance readings. Transistors that are soldered into their printed-circuit boards aren't that simple, yet can be handled without much trouble. Just unsolder the base connection (or slit the foil as already described).

In a circuit that contains a FET (field-effect transistor), unsolder the gate connection and the source connection. You can get low-resistance readings through some FET channels, even when the gate is disconnected.

My fourth suggestion for dealing with parallel paths is a last resort. When you can't analyze a circuit any other way, disconnect and test each part in the circuit—one at a time. It takes time, but if you've done your preliminary testing right, there aren't too many parts to test. At least you'll have narrowed the trouble down to one small section of the circuit.

in dangerous circuits ... the safe way

I brought up resistance measurements in the first place to explain how to trouble-shoot transmitters without exposing yourself to dangerous high voltages. Hunting a fault in the "live" plate circuit of an ssb linear can be a real pain in the neck. You need a special meter to measure high voltage, and usually you can't even reach test points without cheating a bunch of interlocks. There may be an overcurrent relay, too, or an undercurrent relay. In other words, the practical way to troubleshoot is with the transmitter or linear completely disconnected from power, and opened up. For that, you need ohmmeter measurements.

The plate circuit of a commercial linear
amplifier is shown in fig. 5. Mainly, you check for continuity. However, you can disconnect the power-supply lead and make measurements between ground and points in the circuit.

If an interlock switch isn't working, your measurements show it. The normally open (n.o.) switch is there to short out high voltage across the capacitors when you open the transmitter compartment. At the same time, the normally closed (n.c.) interlock switch removes applied voltage. Start your tests with power disconnected and the interlocks pushed in the way they would be with the compartment closed.

If everything's normal, you should get a low resistance reading from the plate cap of either tube to the clip that connected to the power-supply terminal. Check the opening and closing of the overcurrent relay with your finger. An open choke, tuning tank coil, relay contact, relay coil, n.c. interlock, or metering resistor (R1) spoils continuity, but each one is easy to track down. Just clip your common ohmmeter lead to the power-supply clip, and work your way to the tube plate-caps with the other test lead.

A shorted or leaky by-pass capacitor (C1, C2, C4) shows up when you connect your ohmmeter, set for its highest megohms range, between the plate-cap and ground. Be sure the PA current meter is not connected; it could be an unseen parallel path. Disconnect the power-supply clip, too, for the same reason. Set the ohmmeter to its lowest range to test the n.o. interlock switch; it should read zero ohms to ground when the cover is off the transmitter high-voltage compartment and infinity with the cover in place.

In high-voltage circuits, especially the ones that carry high-energy rf, certain troubles only show up when the circuit is energized. For example, capacitor C2 or C1 might break down under high-voltage stress, yet test okay by the ohmmeter. The only sure test is to remove it from the circuit and substitute another one. This is costly, though, when the capacitor is an expensive one, such as a bathtub type. For these unusual cases, there are some special tests I'll talk about in a future column. You can make these tests without working inside the "live" transmitter or linear amp.

**individual parts tests**

As a wrapup for this two-month coverage of ohmmeter tests, here are some parts tests. Usually, these tests are made with the part disconnected from its circuit. When other tests and resistance readings point to a particular part, the final test is of the part itself. The following tests are quick and dependable.

Test **resistors** and potentiometers by measuring their resistance. For potentiometers, also check how smoothly the resistance changes. Connect your ohmmeter between one end lug and the center lug, and turn the control shaft slowly; any unusual or sudden resistance change means you need a new control.

Test ordinary **capacitors** by measuring with the highest range on your ohmmeter. There should be absolutely no leakage (infinite resistance). Touch the two capacitor leads together for a moment before applying the ohmmeter. Capacitors above 0.01-μF show a slight charging "kick" of the ohmmeter pointer. Those that don't are probably open. A shorted capacitor reads zero ohms between the two leads.

Test **electrolytic capacitors** by applying the ohmmeter leads in first one direction and then the other. Settle for the connection that gives the highest reading. It should exceed 30k, but not go beyond 60k or so. If it's too low, the capacitor is leaky. If it's too high, the capacitor is becoming ineffective. Notice how long it takes to "build up" to the highest ohmmeter reading. If it takes more than 3 or 4 seconds, the capacitor needs replacing.

There are three tests for **coils and transformers.** Test them first for continuity of windings. The schematic diagram may list the dc resistance of each winding; if some turns are shorted, the reading will be much too low. If a winding is open, the ohmmeter shows nothing.

Test coils and transformers second for leakage between windings. Use the highest ohmmeter range. Even several megohms—an awfully high resistance—is no good between transformer windings; they should be totally isolated (infinite resistance).

Test coils and transformers third for leakage from windings to ground. With the transformer mounted as usual but with all its leads disconnected, clip one ohmmeter lead to
ground and touch the other to each winding lead. There should be infinite resistance, since each winding should be totally isolated from the frame of the coil or transformer. If you think maybe the coil frame isn't grounded, clip one ohmmeter lead to the frame and touch the other to each winding. There should still be no reading. A reading other than infinity indicates leakage.

Test a relay by checking its coil the same way you test a transformer winding. Check the coil for continuity; if you know the dc resistance, measure it. If the ohmmeter reads infinity, the relay coil is open. Then check the relay contacts, with the Rx1 ohmmeter range. Normally closed contacts should measure zero ohms when the relay is at rest; at the same time, normally open contacts should measure infinity. Push the armature in with your finger. The normally open contacts should measure zero ohms that way, and the normally closed contacts should measure open (infinite resistance).

Test speakers and headsets by measuring them with the Rx1 ohmmeter range. Listen as you make the ohmmeter connection; the voltage of the ohmmeter battery usually makes speakers and headsets “pop.” The ohmmeter pointer should show very low resistance for speakers and a few hundred ohms for headsets.

Test microphones in one of three ways, depending on type. For carbon microphones, connect the ohmmeter across the element leads and whistle into the mike. The reading should vary with each whistle. If it stays steady, the granules are packed and the element needs replacing. If the ohmmeter reads infinity, the mike cord is probably open, or the element is making poor contact.

For dynamic microphones, the exact resistance depends on the impedance of the mike. However, there should be some sort of reading, even if an input transformer is used inside the mike. Also, if you listen, there is usually a “pop” in a dynamic mike when you connect or disconnect the ohmmeter.

For ceramic microphones, the easiest test is to connect the ohmmeter and listen for a slight “click.” The ohmmeter shows a resistance reading only if the mike is defective; a good ceramic or crystal mike reads infinity resistance.
Finally, connect the ohmmeter across the push-to-talk leads, and operate the button on the mike; a zero-ohms reading when the button is held means the p-t-t contacts and wires are okay.

Test diodes by measuring them first one way and then the other. A diode should measure low resistance in one direction and high resistance in the other; if the two readings are closer together than 100-to-1, the diode is probably bad. Power-supply rectifiers usually measure 100–200 ohms in the low-resistance direction; other types measure much lower.

A conventional bipolar transistor should show high resistance between emitter and collector, no matter which way the ohmmeter is connected. Between the base and emitter, however, resistance should be low in one direction and high in the other—just as in a double. If the base-emitter resistance is high in both directions, the transistor is open; if it's low in both directions, it's shorted. If resistance between emitter and collector is lower than several thousand ohms in either direction, the transistor is probably useless.

A field-effect transistor normally shows low resistance between drain and source—usually no matter which way the ohmmeter is connected. It should show extremely high resistance in one direction between gate and drain, and fairly high in the other direction. A MOSFET, sometimes known as an IGFET, should show infinite resistance between its gate and any other element, no matter which way the ohmmeter is connected.

next month

Everyone has his pet way to troubleshoot. In past repair bench columns, I've shown you several ways. You've read about signal tracing, signal injection, and now resistance measurement.

The best commercial technicians use several methods in various combinations that fit particular servicing problems. In my next column, I'll talk about one that is probably the most common: voltage testing. When a part goes bad, it usually upsets the dc operating voltages in its circuit. Those voltage changes, if you know how to interpret them, are a valuable clue to what part is bad. Next month, you'll find out how.
Late reports that reached me in October indicated that ionospheric critical frequencies were not much higher (about 10%) than predicted by ESSA. Peak values at Raratonga in the Cook Islands (latitude 21° S, longitude 160° W) rose to almost 15 MHz, compared to about 20 MHz during the peak of previous cycles. Local midnight critical frequencies have been running as high as 12 MHz, however. On the West Coast, peak values of critical frequencies have been running as high as 11 MHz, again about 10% higher than predicted.

We have reports of 50 MHz F2-layer openings during the first half of October between Hawaii and the Marshall Islands, South America, Mexico, and even (weakly) the West Coast of the United States; also between Japan and the Marshall and Cook Islands. Transequatorial scatter propagation between Hawaii and the Cook Islands was an almost nightly occurrence. However, the big, strong signal F2-layer openings into temperate latitudes hoped for this fall have so far been absent.

Quite a few six-meter operators have been missing a good bet by not running more nighttime schedules with and listening for beacon stations in the southern hemisphere. I was pleasantly surprised during the latter part of September to hear Z1AA’s beacon station on 51.022 MHz three nights in a row between 11 PM and 1 AM local daylight time. He was weak, both here and in Southern California at W6NLZ, but you’d never have guessed that 51 MHz was open to the South Pacific by monitoring the low edge of the band.

The active sun

Let’s continue our discussion of the sun, which last month stressed its stability. The sun rotates in the same direction as the earth. An observer at the north star would see both the sun and earth rotating in a counterclockwise direction as the earth revolves counterclockwise around the sun. Thus the solar surface rotates from our east to west. Since the axis of the sun is inclined 7° to the ecliptic, we are able to see the sun’s North pole only from June to December.

The sun does not rotate as a solid body—the equatorial regions rotate faster (25-day period) than the polar regions (30-day period). It is this equatorial acceleration and the presence of a general magnetic field that is the cause of the solar cycle. The sun’s general field penetrates the surface only in polar regions greater than 55° latitude and is called a poloidal field. The intensity of the poloidal field is about 1 gauss, compared to fields of 0.24 to 0.61 gauss observed on the surface of the earth. At solar latitudes less than 30°, the general magnetic field is confined to a sheath of about 5% of the solar radius. At the equator, the general magnetic field is believed to have an intensity of about 5 gauss.

The motions of ionized media (plasma)
are closely coupled with the magnetic and electric fields. Relative motion between a plasma and a magnetic field is inhibited since such motion would produce an electric field which would result in large current flow (since the plasma is an excellent conductor). Therefore, the plasma is forced to move along magnetic field lines; otherwise the magnetic field is dragged along with the plasma.

The differential rotation wraps the subphotospheric field lines around the sun to form a toroidal field. The toroidal field strength continually rises as the field lines are stretched and crowded together like a rubber band that has been wound up tightly. When the magnetic field strength reaches a critical value of about 250 gauss, tubes of magnetic flux may "bubble up" through the photosphere to form bipolar magnetic regions. It is in these "centers of activity" that sunspots, solar flares and filaments occur.

**sunspots**

Sunspots are visible as small dark areas on the solar disk. They appear dark because their temperature is only about 3000° K compared to 6000° K for most of the photosphere. The low temperature is due to the presence of high magnetic fields (about 3000 gauss) which inhibit convection.

Sunspots are believed to develop in pores; a pore is a small (1000- to 2000-mile diameter) moderately dark area which may occur anywhere on the sun and has a normal lifetime of a few hours. Only pores within centers of activity develop into sunspots. The surrounding hotter plasma tends to flow into a pore, carrying its magnetic flux along and concentrating it.

As the magnetic field strength grows, convection from lower hotter levels is decreased, the surface temperature drops, and the spot grows. At full development, the spot may be 20,000 miles or more across and consists of a dark umbra surrounded by a penumbra with very fine radial filamentary structure. It may last from as short as a few days to as long as a hundred.

The appearance and motion of sunspots, bipolar magnetic regions and other solar activity follow an eleven-year cycle, or, more properly, a twenty-two year cycle, since the magnetic fields reverse every eleven years.

Sunspots tend to occur at high solar latitudes early in the 11-year cycle and at low latitudes late in the cycle. The distribution of spot latitude vs age of the cycle follows a curve of

\[
\sin \varphi = \pm \frac{1.5}{n + 3}
\]
where \( \varphi \) is the latitude of maximum occurrence and \( n \) is the age of the cycle in years. This theoretical curve is plotted on fig. 1 along with the observed Zurich smoothed sunspot numbers from April 1954 to December 1967 and ESSA predicted smoothed sunspot numbers through 1970.

Over 84% of the sunspots occur in bipolar pairs. The preceding and following (with respect to solar rotation) spots have opposite magnetic polarity, and the preceding spot is closer to the equator. In the opposite hemisphere, the magnetic polarity of preceding and following spots is reversed. The preceding spot has the opposite polarity as the polar field in the same hemisphere at the beginning of the cycle. This is consistent with the theory of spot development from the twisting of the polar field due to equatorial acceleration.

The preceding spot is generally larger and has more magnetic flux (by a ratio of 3:1) than the following spot. Sunspots rotate about their axes (which are tilted in the forward direction) in the direction of the Coriolis force* (counterclockwise in the Northern Hemisphere, clockwise in the Southern Hemisphere). In general, however,

* Usually defined as a deflecting force acting on a body in motion (as an airplane or missile) due to the earth’s rotation. In this case it refers to the force on a sunspot due to the sun’s rotation.

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FIG. 2. Time chart of maximum usable frequencies for January, 1969 centered on 90° West longitude.

FIG. 3. Maximum range to the north due to absorption.

the sunspots are calm magnetic lands in a sea of great turbulence that forms the rest of the bipolar magnetic region and the whole of the solar disk by comparison.

Sunspots are a secondary phenomenon since the bipolar magnetic region forms first; they themselves do not radiate increased ultraviolet and x-ray emission that affects the earth’s ionosphere. In fact, the
fig. 4. Maximum range to the northeast (top time scale) and to the northwest (lower time scale) due to absorption.

fig. 5. Maximum range to the east (top time scale) and to the west (lower time scale) due to absorption.

fig. 6. Maximum range to the southeast (top time scale) and to the southwest (lower time scale) due to absorption.

fig. 7. Maximum range to the south due to absorption.
total radiant energy output per unit area in a sunspot may be only a tenth of normal for the solar disc.

Sunspot numbers** are by no means as accurate an index of solar activity as you might wish, since they are somewhat artificial in nature. For example, a large sunspot pair somewhere on the solar disk would result in a Wolf number 12; add an insignificant small spot somewhere else on the surface, and the Wolf number becomes 23, a jump of almost a factor of two. This artificial nature of the sunspot numbers is one of the reasons for the poor correlation between monthly sunspot numbers and monthly average ionospheric critical frequencies. The answer is a better index of solar activity, such as the combined area of all the sunspots, or of bipolar magnetic regions, or (now that we have space probes) of the magnitude of the ionizing radiations that affect our ionosphere.

metamorphosis

The magnetic flux lines joining the two poles of a bipolar magnetic region may expand outwards to great distances in the corona—and may break away with the flow of plasma away from the sun. As the solar cycle progresses, the preceding regions of flux migrate towards the equator and neutralize each other; the following regions of flux migrate towards the poles where they first neutralize the old poloidal field and then from a new poloidal field of reversed polarity. The energy for maintaining and concentrating the magnetic flux comes from the kinetic energy of differential rotation.

The reversal of the poloidal field takes place near the maximum of the solar cycle, but strangely enough, both poles need not reverse simultaneously. At the peak of sunspot activity in the southern solar hemisphere in the middle of 1957, the southern poloidal field reversed, but the peak of sunspot activity and reversal of the poloidal field in the Northern Hemisphere did not occur until the end of 1958. These reversals left the poloidal field of the sun parallel to that of the earth.

**Sunspot numbers were discussed in detail in the November, 1968 issue of ham radio, page 72.

One or both solar poloidal fields may have already reversed this year.

propagation for january

A time chart of predicted 4000-km F2-layer muf for January 1969 is given in fig. 2. Maximum range versus time of day charts for 160 to 20 meters are shown in fig. 3 to 7. The first half of January will be a repeat of December with conditions becoming more like those of November as the month progresses. It will be a great month for high-frequency propagation in the Northern Hemisphere from frequencies as low as 1.8 MHz at night to as high as 40 MHz during the day.

The 50-MHz operator aspiring to WAC is now wishing that (in addition to a kilowatt, a big beam and hilltop location) the location was south of 30° latitude and that he had enough time to keep the band under constant surveillance. With that, we offer a salute to those dedicated chaps who keep beacon stations running in Rhodesia, Mexico, Hawaii, the Cook Islands, Australia, and Japan.

references


ham radio
State of the Art in '69? or...

THE GOOD OLD DAYS?

It wouldn't transceive... and it was big and heavy... But... K6CTV's 1959 station was designed for DXing and contesting... and it was a joy to operate!

REMEMBER WHEN...

YOUR DREAM RECEIVER

- offered a choice of really steep skirted selectivities?
- let you slice off adjacent-channel QRM or set the audio passband to your taste with a single control?
- tuned anywhere in the band independent of the transmitter?

AND THE TRANSMITTER

- didn't overheat if you held the key down for five minutes?
- had effective speech clipping built in?
- keyed silently and cleanly for instant CW break-in?
- could be “spotted” exactly where you wanted it?
- band-changed as easily as the receiver?
- used rugged, high linearity output tubes for SSB?

AND BOTH OF THEM

- completely covered all ham bands from 160 through 10 meters in convenient, full mega’cycle” ranges?
- read out frequency simply and unambiguously to a kilo’cycle”?

Of course, hardly anyone who’s enjoyed the convenience of transceiving on sideband is likely to give it up, and none of us really want to wrestle ninety-pound desk-crushers. But is it necessary to sacrifice performance and convenience features that were readily available a decade ago?

TODAY at SIGNAL/ONE we believe that “state-of-the-art” equipment ought to be at least as convenient and effective in every respect as were the “separates” of the “good old days”. We’re demonstrating that conviction with a smooth blend of sophisticated modern engineering plus traditional ham ingenuity—bringing you a superb new line of NO COMPROMISE equipment—a new criterion for the next decade in amateur radio.

Performance... Convenience... Quality...

“It Speaks for Itself”

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2200 Anvil Street N. • St. Petersburg, Florida 33710
antenna and rotator
preventive maintenance

Here's graphic proof that Ben Franklin's Advice about a pound of prevention is still true.

Most amateurs agree that the antenna system is the heart of any station. Much has been written on the emphasis that should be put on erecting the best possible radiating system to compete on the crowded bands. I'd like to add some footnotes to the antenna literature in the area of preventive maintenance.

The initial investment in an antenna and rotator is substantial. How many times have you heard fellows say on the air, "Sorry, my antenna's stuck on Europe. Big storm passed through, and I haven't had a chance to go up the tower and fix the rotator." With a little care and forethought, you can ensure reliable operation during the worst weather, even if you use the more inexpensive equipment.

The following notes explain how to avoid the disappointment and down time that can occur with stuck rotators and vibrating antenna elements during the winter months.

the antenna

It's pretty discouraging to come home, anticipating an evening of hamming, and find your pride and joy in bits and pieces all over the back yard. Beam antenna elements are usually made of aluminum alloy. Unless they're given the proper treatment, they just won't withstand the rigors of weather. Airframe manufacturers know this—aluminum
alloy is used extensively in this application. Where in the aircraft the aluminum is used determines the amount and type of treatment it's given before installation.

Most amateurs don't have the facilities for giving aluminum the correct treatment for withstanding environmental effects. The treatment is quite involved and includes chemical and electro-chemical techniques. It is quite expensive, also, if used for small quantities of metal.

One thing you can do, before hoisting your antenna to the top of your tower or mast, is give it several coats of zinc chromate paint. Put on as many coats as you can, and work it into every nook and cranny. For each coat of zinc chromate properly applied, you can count on at least six months of corrosion-free operation.

Now let's look at metal fatigue caused by vibration.

Tie a small weight on the end of the twine and drop it through the element.

vibration proofing

All that's needed to dampen antenna element vibration is some sisal twine and something with which to plug the ends. Measure the length of all your elements plus a little extra, say a foot or so. Drop the twine through each element, and secure the twine with rubber caps over the element ends. A tennis ball is good. It may look a little strange, but who cares what the neighbors think? It's solid enough to make a good seal and will hold the twine securely. Make sure there's no moisture in the elements. The twine will dampen the vibration by changing the bending-mode frequency to some harmless low value.

Another method is to pack the elements loosely with spun glass. Wear gloves when handling this stuff, though, because it can be pretty treacherous. It can penetrate your skin and cause all sorts of agony. Your boom can be sealed with a sponge rubber ball a little larger in diameter than the boom. Use the twine treatment here, also, and pull it up tightly. It will help keep your beam on a straight course and ease your jangled nerves...
when your hear the wind wailing at 3 a.m. I'd like to emphasize that the spun glass must be packed into the tubing loosely. If it's packed too tightly you will zap the element.

Another way of damping element vibrations is to fasten several rubber or plastic strips to the end of each element. Since the mechanical impedance of any rod-like element is high at the free end, an energy absorber at the end of an element will reduce the amplitude of vibration and minimize breakage. The energy absorber can be made from a five- or six-inch length of garden hose, split lengthwise every 90° and held at the end of the element with a hose clamp. These simple energy absorbers are most effective if they are clamped at the very end of each element, but to prevent changes in element electrical length and impedance, they should be placed so the free ends are about two inches in from the ends of the elements.

the rotator

Lots of stories have made the rounds about whether TV antenna rotators are strong enough to support amateur beams. The photos show my AR22 rotator, which was purchased, used, in 1963. At that time it had been in use for six months, and I've been using it constantly since. The reason it's down now is because I'm installing a new tower and two stacked five-element Cushcraft beams.

Given half a chance, I find that the AR22 rotator is very satisfactory. The secret is to eliminate the tremendous moment arm caused by a large mass of metal on a long length of mast above the rotator. Add to this the simple fact, sometimes overlooked, that any rotator must be perfectly centered with respect to its mast. Unless the correct diameter mast is bolted to the rotator, the eccentric rotation will load the rotator bearings, and you've got problems.

The AR22 manufacturer say that 1 1/2-inch diameter pipe is correct. Maybe so, but I found this isn't true for my rotator. In my case, a diameter of two and one-sixteenth
inches is correct. I had a piece of aluminum stock turned down to this size, then had it stepped down above the rotator clamps to slip inside a 1½-inch mast.

It may be necessary to file down the edges of the shoulders where the U-bolts come through to properly seat the aluminum reducer. I didn’t have to, though.

The picture with the ruler has been purposely double exposed to show that the reducer is, in fact, centered in the rotator. I made the first exposure with the rotator pointed in one direction, then I turned it exactly 180 degrees and made the second exposure. Note that there’s no visible deviation at the end of the reducer. If there were, two distinct centers would show, and two outer edges would also show.

**the thrust bearing**

A thrust bearing located somewhere between the top of the rotator and the top of the tower takes all the weight off the rotator. The only load on the rotator will then be turning torque, which is as it should be.

If the thrust bearing is exposed to the weather, it’s going to get loaded with snow and ice. (California amateurs can skip this part. However, does anyone know what smog does to aluminum?) I made a copper umbrella to cover the thrust bearing, as shown in the photo. It slips onto the mast and is fastened with an ordinary automobile radiator hose clamp. The same idea can be used at the top of the tower where the mast enters the tubing.

Instead of a copper umbrella, you could wrap a piece of soft rubber around the mast and secure it with the same type of hose clamp. This will form a skirt to keep out the weather. Be sure the skirt flares at the bottom, else it might get loaded with ice and stick to the tower. It might help to lubricate the mast-skirt interface with low-temperature silicone to avoid this minor problem.

*ham radio*

The hood or umbrella clamped on to the mast directly above the thrust bearing, which is out of sight and out of the weather; it’s mounted on the plate welded to the tower cross-sections.

*those days before spark...*

“Yuh can’t get thru a pileup like that... yuh gotta better chance out here on the edge.”

*January 1969*
SWAN 508
FULL COVERAGE EXTERNAL VFO

The Model 508 Frequency Control Unit is designed for full coverage of 80, 40, 20, 15, and 10 meters. It provides for transmitting and receiving on separate frequencies, and plugs directly into the back of the 500C. A separate Dual-VFO adaptor is no longer required, since the relay control circuitry is built into the 508. A panel control permits selection of VFOs so that operation may be transceive mode with the 500C VFO, transceive with the 508 VFO, or transmit on the 500C and receive on the 508. The Model 508 features eight ranges of 500 kc each, with 5 kc calibration. It may also be used with the 350C transceiver.

$125

SWAN 500C
FIVE BAND TRANSCEIVER

80 through 10 meters • 520 watts • Home station, mobile, portable operation • SSB-CW-AM.

The new model 500C is the latest evolutionary development of a basic well proven design philosophy. It offers greater power and additional features for even more operator enjoyment. Using a pair of the new heavy duty RCA 6L6G tetrodes, the final amplifier operates with increased efficiency and power output on all bands. PEP input rating of the 500C is conservatively 520 watts. Actually an average pair of 6L6G's reach a peak input of over 570 watts before flattopping!

The 500C retains the same superior selectivity for which Swan transceivers are noted. The filter is made especially for us by C-F Networks, and with a shape factor of 1.7 and ultimate rejection of more than 100 db, it is the finest filter being offered in any transceiver today.

For the CW operator the 500C includes a built-in sidetone monitor, and by installing the Swan VOX Accessory (VX-2) you will have break in CW operation.

Voice quality, performance and reliability are in the Swan tradition of being second to none.

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SWAN 117XC
MATCHING AC POWER SUPPLY

Complete A.C. supply for 117 volts, 50-60 cycles, in a matching cabinet with speaker, phone jack, and indicator light. Includes power cable with plug for transceiver, and A.C. line cord. Ready to plug in and operate.

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SWAN 14-117
12 VOLT DC SUPPLY

Complete D.C. supply for 12 volt mobile or portable operation. Includes cables, plugs, and fuses. Will also operate from 117 volt A.C. by detaching the D.C. module & plugging in 117 volt line cord. Negative ground standard. Positive ground available on special order.

$130

PHONE PATCH, Model FP-1 ...........$48

ASK THE HAM ... WHO OWNS ONE
We are proud to announce production of the deluxe model 250C, new contender for top honors on the 6 meter band. The 250C is built to performance standards of the highest order, including a receiver front end employing dual Nuvistors in cascode with a noise figure of less than 3 db. Deluxe features include a built in 250 KC crystal calibrator; 'S' meter; selectable sideband; vernier control of the megacycle tuning range; and accessory sockets on the rear apron to receive the model 210 external VFO, the new Swan Noise Silencer and the VX-2.

The famous model 250 is a consistent winner in VHF competition -The 250C is designed to set new records! We are confident that the 250C will satisfy the most critical requirements of the serious VHF operator.

$420

SPECIFICATIONS:
- Frequency Range: 50-54 MC
- Power Rating: 240 watts PEP Input in SSB mode, 180 watts CW Input, 75 watts AM input
- Two 6146 B Power output tubes
- Distortion Products: down approx. 30 db
- Unwanted Sideband: down more than 40 db
- Carrier Suppression: better than 50 db
- Receiver Noise Figure: Better than 3 db, with two 6CW4 nu
vistors in Cascode
- Selectivity: 2.8KC at 6 db down, with crystal lattice filter at 10.9 MC.
- Antenna Matching: Wide range Pi network.
- Metering circuits: S-meter on Receive mode, P.A. Cathode Current and relative output in transmit mode.
- 250 KC Crystal calibrator
- Selectable upper and lower sideband

ACCESSORIES:

- Receiver Mode switch provides for AM reception
- Accessory sockets for noise silencer, external VFO and VOX unit

MATCHING AC
POWER SUPPLY
Model 117XC .............. $105
12 VOLT DC SUPPLY
Model 14-117 ............. $130
SWAN NOISE SILENCER
Model NS-1 ............... $36
EXTERNAL VFO
Model 210 ................ $120
PLUG-IN VOX UNIT
Model VX-2 ............... $35
PHONE PATCH
Model FP-1 ............... $48
2KW LINEAR AMPLIFIER
Model 6B. With power supply .......... $660

AND FOR 2 METERS
SWAN TV2 TRANSVERTER

A receiving and transmitting converter, which may be used with the 250C or 500C to operate in the 144-148mc band. Provides 240 watts P.E.P. transmit power, and low noise receiver front end with Nuvistors in Cascode.

$295

One of the main reasons Swan maintains its position as the leading manufacturer of amateur radio equipment is our dedication to the principles of Value Analysis, and Value Engineering.

We continually examine our products to find ways of improving their performance, reducing costs, and making them better values. As new components are developed in the electronic field, we analyze them to see how they can be used to improve the performance of our transceivers.

The most recent results of our Value Analysis and Value Engineering are shown on these pages: The 500C 5 band SSB Transceiver, our new External VFO, the 510X Mars Oscillator, and introducing our new 6 Meter Transceiver, the 250C.

Yet with all the improvements, the increased reliability, and proven performance, Swan's Value Engineering results in substantially lower prices than competitive equipment. And every piece of Swan equipment is backed up by service second to none. Value Analysis, Value Engineering keep Swan in the lead. Visit your Swan dealer soon.

SWAN ELECTRONICS
OCEANSIDE, CALIFORNIA
A Subsidiary of Cubic Corp
Tompkins Radio Products has just announced a new squelch control for noise-free monitoring on hf and vhf Tunaversers. This makes it possible to have receiver squelch performance with an economical converter. The solid-state squelch circuit doesn't need any connections to the a-m broadcast radio or the converter other than the coaxial feedline.

This new unit features complete electronic control without relays and has a fully adjustable squelch setting. It is available in two models—the model ST that fastens to the bottom of a Tunaverter or the model SU which has its own mounting bracket. Furnished complete with connecting cable for $17.50 (model SU $1 extra). For more information, write to Herbert Salch and Company, Marketing Division of Tompkins Radio Products, Woodboro CTI, Texas 78393.

Allied Radio has recently announced a new solid-state receiver for amateurs and swl’s, the A-2515. This new five-band receiver tunes all the amateur bands from 80 to 10 meters, international shortwave, aircraft, marine and the standard a-m broadcast band. The bands covered are 150-400 kHz, 550-1600 kHz, 1.6-4.8 MHz, 4.8-14.5 MHz and 10.5-30 MHz.

A total of 24 semiconductors are used in the circuit with two field-effect transistors in the front end for maximum sensitivity and low noise. Four mechanical filters provide sharp station separation. The noise limiter and agc reduce noise, blasting and fading. The built-in bfo and product detector give good reception of CW and ssb. The illuminated slide-rule dial has calibrated bandspread for the amateur bands; an S-meter is included on the front panel. Other features include a push-pull audio stage with a thermistor for low distortion, provisions for receiver muting and a headphone jack.

The receiver is equipped with dual power supplies, one for 117 Vac and the other for 12 Vdc, so the receiver may be operated from house current, autos, boats or camp sites. $99.95 from Allied Radio Corporation, 100 North Western Avenue, Chicago, Illinois 60680.
TALK POWER

The AutoLevel is the ultimate in volume compressors. This unique device provides all the talk power your transmitter can use. The AutoLevel was designed for use with SSB or AM transmitters, with or without ALC capabilities.

The AutoLevel is not an audio or RF clipper — all compression is attained by a photo-optical regulator which provides 14 dB's of compression with a minimum of wave form distortion.

The AutoLevel is easily installed in the mike line, and it contains its own power supply; (there's no need to bother with batteries). It can also be used with your phone patch for the utmost in ease of operation.

When you're ready for the finest, ask your local dealer for the AutoLevel.

SPECIFICATIONS

<table>
<thead>
<tr>
<th>dB's compression</th>
<th>14 dB minimum</th>
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<tr>
<td>Wave form distortion</td>
<td>negligible</td>
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<tr>
<td>Input impedance</td>
<td>suitable for dynamic or crystal microphone</td>
</tr>
<tr>
<td>Output impedance</td>
<td>50K (nominal)</td>
</tr>
<tr>
<td>Power supply</td>
<td>115 volts AC</td>
</tr>
</tbody>
</table>

AUTHORIZED DEALERS

(listed alphabetically)

HENRY RADIO STORE
931 N. Euclid
Anaheim, California 92801
Tel: 714-772-9200

HENRY RADIO STORE
11240 West Olympic Blvd.
Los Angeles, California 90064
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L. A. AMATEUR RADIO SUPPLY
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Redondo Beach, California 90278

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Box 506
De Witt, Arkansas
Tel: 501-984-2830

PIONEER-STANDARD ELECTRONICS, INC.
5403 Prospect Avenue
Cleveland, Ohio 44103
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PIONEER-STANDARD ELECTRONICS, INC.
SREPCO ELECTRONICS DIVISION
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Tel: 513-224-9871

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114 North Third Street
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Tel: 614-221-2335

VALLEY HAM SHACK
4109 N. 39th Street
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Tel: 602-953-4850

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January 1969
new allied catalog

Allied’s new 1969 catalog “Electronics for Everyone,” is now available. This new book presents the latest in electronic components, vacuum tubes, semiconductors, wire and cable, test instruments, microphones and speakers, antennas and towers and amateur equipment as well as a full line of high-fidelity equipment and tape recorders, transistor radios, TV sets and video tape recorders. Specialty lines include a big selection of electronics kits, cameras and accessories, weather instruments and a wide selection of electrical accessories. The 1969 catalog, number 280, is available on request from Allied Radio Corporation, Post Office Box 4398, Chicago, Illinois 60680.

motorola semiconductor catalog

Motorola has just announced a new catalog of 175 semiconductors, books and accessories in the HEP line of devices for the amateur, experimenter and electronic hobbyist. The HEP line includes semiconductors at reasonable prices for almost any experimental or replacement use.

There are many new devices listed in this catalog: high-voltage germanium power transistors, complementary silicon npn/pnp power transistors, a unijunction transistor, a bilateral trigger diode, a sensitive low-power SCR, digital RTL and linear integrated circuits, an IC kit, new IC sockets and a TO-5 heat sink.

Other semiconductors in the HEP line include small-signal and power silicon and germanium transistors, field-effect transistors, silicon rectifiers and bridges, zener diodes, SCR’s, power transistor mounting kits, heat sinks and books.

Copies of this handy new catalog, the Motorola HEP Catalog, MHA27-4, are available at your local Motorola HEP distributor or by writing to the Technical Information Center, Motorola Semiconductor Products, Inc., Post Office Box 20924, Phoenix, Arizona 85036.
Jim Fisk  W1DTY
editor of Ham Radio
reviews
The new

Radio Communication Handbook
published by

"One of the most exciting handbooks I have seen in a long time"

"This new book covers every aspect of amateur radio communications and is filled with solid state projects and circuits"

"If you're troubled by the lack of semiconductor circuits and modern technology, the Radio Communication Handbook is the answer; it is the most up to date radio handbook that has crossed my desk"

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Comtec also stocks most other R.S.G.B. publications
There's an error in one of the specs for the new Hammarlund HQ-200 receiver advertised on page 27 of the December issue—the a-m sensitivity should read 1 $\mu$V for 10 to 1 signal-to-noise ratio.

The code wheels shown with the article on the digital wind-direction indicator in the September issue are in error. Fig. 3, shown as a Grey-coded wheel, is actually a binary wheel. Fig. 2 is complete gibberish—a result of the photographic process used to make the illustrations for the magazine. A correct Grey-coded wheel is shown below.

In the schematic for the five-watt modulator on page 10 of the October issue, there should be a 270-ohm, 2-watt resistor in the line that runs from the center tap of the Stancor TA38 transformer (red lead) to the voltage supply. In addition, the line going to pin 9 of the integrated circuit should be connected to the two lines that cross it—one between pins 8 and 11 and the other from pin to the transformer.

In the decade counter shown in fig. 3 on page 44 of the November issue, pin Q on flip-flop C should be connected to pin J of flip-flop D.
THE **ICE-1**

**2 METER FM TRANSCEIVER**

**FULLY SOLID STATE - NO TUBES**

- Operates on 117 VAC — 12 VDC — or optional internal NI-CAD battery
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- Light weight — Less than 4 1/2 lbs.
- Built in 117 VAC power supply
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- Transmitter and receiver channels individually switchable
- 3 channels transmit — 3 channels receive
- Push-to-talk operation
- Transmitter output — 4 watts minimum

**COMPLETE WITH ONE SET OF CRYSTALS ON 146.94, 117 VAC AND 12 VDC**

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<tr>
<td>POWER CABLES, LESS MICROPHONE AND ANTENNA</td>
<td>$285.00</td>
</tr>
<tr>
<td>MICROPHONE</td>
<td>16.00</td>
</tr>
<tr>
<td>EXTRA CRYSTALS (TRANSMIT OR RECEIVE)</td>
<td>7.00</td>
</tr>
<tr>
<td>BUILT IN NI-CAD BATTERY AND CHARGER</td>
<td>47.00</td>
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**ICE**

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**January 1969**
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5 band, 520 Watt
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The new Swan 500C offers you higher power, improved styling, and many new deluxe features, yet maintains the same high standards of performance, rugged reliability and craftsmanship that have become the trademark of the Swan Line. Backed by a full year warranty and a service policy second to none, we feel the Swan 500C will establish a new standard of value for the industry.

$520

SWAN 350-C $420.00
SWAN 250-C (NEW MODEL) $420.00

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14C DC MODULE $65.00
VX-11 VOX $35.00
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SWAN MARK II LINEAR AMPLIFIER
2000 WATT P. E. P.

The new Mark II provides full frequency coverage of the amateur bands from 10 through 80 meters, and also MARS frequencies. Two Eimac 3-400Z Triodes easily provide the full legal power input: 2000 Watts P.E.P. in SSB mode or 1000 Watts AM, CW, or RTTY. It may be driven by any transceiver or exciter having between 100 and 300 watts output.

Planetary vernier drives on both plate and loading controls provide precise and velvet smooth tuning of the amplifier. Greatly reduced blower noise is provided by a low RPM, high volume fan.

Complete with Tubes: $395

MATCHING POWER SUPPLY

The power supply is a separate matching unit which may be placed beside the Mark II amplifier, or with its 4½ foot connecting cable, may be placed on the floor. Component quality is of the highest caliber. Silicon rectifiers deliver 2500 volts D.C. in excess of 1.2 ampere. Computer grade electrolytic filters provide 40 mfd capacity for excellent dynamic regulation. A quiet running fan allows continuous operating with minimum temperature rise, extending the life and reliability of all components.

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**CLEANING HOUSE.** 6-meter linear amplifier, Knight T-175. 300 watts PEP ssb, 150 watts eff. less than 10 watts drive; may be used on a.m., built-in power supply. Almost new. 6-meter draft-6 meter 6-element beam. $15. Johnson 275-watt Matchbox with swr bridge, mint, $50. cavity oscillator with tubes tunes from 300 to 3000 MHz nationally advertised at $80; excellent, $20. Waters model 384 dummy load, like new, $50. WD1TY, Box 25, Ridgé, N. H. 03246. (603-899-2860).

**GONSET GSB-6** SIDEWINDER, brand new, sealed factory carton, $285; ac supply, $60; dc supply, $50. 500-watt 913A linear, $245; all new equipment. Van WD2LZD, 607-785-5662.

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**THE LAKE COUNTY AMATEUR RADIO CLUB, INC.** announces its 16th Annual Banquet to be held at Teherl's Restaurant, U.S. 30 and 41 (near Tealville, Ind.), at 6:30 p.m., CST, February 8. Chicken dinner, entertainment, speeches. Please attend with your wife or girl friend. Tickets $4.00 each. Contact S. Brier, W9EQG, 385 Johnson Street, Gary, Indiana 46302.


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<td>B&amp;W 5100B</td>
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