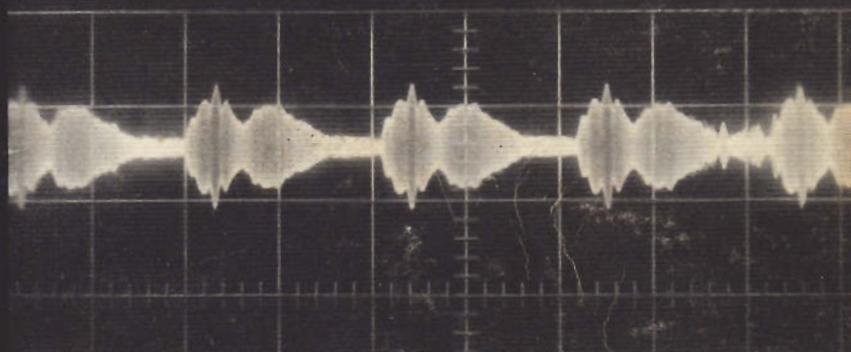


SINGLE SIDEBAND

*for the
Radio Amateur*

\$3.00

A DIGEST
OF
AUTHORITATIVE
ARTICLES
ON
AMATEUR
RADIO
SINGLE
SIDEBAND



PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

**SINGLE
SIDEBAND
for
the
RADIO
AMATEUR**

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Foreword

Many amateurs think of single sideband as a relatively new method of phone communication, but its origins actually go back to the early part of this century. Fifty years ago it was demonstrated mathematically that two sidebands accompany the "carrier" wave as a result of the modulation process. The two are identical, except for their frequency relationship to the carrier, and each one carries the complete intelligence in the signal. This immediately suggested that only *one* actually was necessary. The other, along with the carrier, could be eliminated before transmission, with a considerable saving in power and spectrum space.

In a short time the system was applied to telephone communication and, somewhat later, to long-wave transoceanic telephony. When short waves came into use the transatlantic circuits were extended to the higher frequencies. The technical problems were difficult, a principal one being the attainment of the necessary frequency stability. However, by the end of World War II much progress had been made in frequency stabilization, and shortly after the war's end a small group of amateur experimenters took up the art. The success of their efforts was eye-opening, not only to the amateur world but also to commercial and government hf communication services—which up to that time had made almost no use of the system for general multifrequency communication. Today, "ssb" has almost supplanted the long-standard amplitude-modulation method in the lower-frequency amateur bands, and is making rapid inroads into vhf communication.

The pages of *QST* have carried the complete story of this transition from a-m to ssb, covering the whole field of amateur ssb technique in the process. Building a book around the significant articles published in *QST* has been, and continues to be, a logical way to present the constantly-changing ssb picture. This new edition continues the plan of selecting major articles describing principles, practice, and current circuit methods. Where necessary, the material has been edited to be most useful as of the time this edition is published, but in other respects the author's original version is retained.

"Sideband," through elimination of carrier heterodynes and reduction of channel width, has remarkably improved the communication capacity of our phone bands. Nevertheless, Utopia is still not here. The number of amateur stations has increased greatly in the past few years and, unfortunately, not all operators are sufficiently acquainted with the principles of proper adjustment of ssb equipment. In this edition, special emphasis has been placed on those principles and methods that an amateur *must* observe in adjusting and using his transmitter in order to avoid causing unnecessary interference. In addition, there is a large selection of well-tried equipment designs which the experimentally-inclined constructor can modify, combine, and select from to his heart's content.

This book, both in its present and earlier forms, could not exist had it not been for the enthusiasm with which its many authors took up single sideband and made known their findings through the pages of *QST*. The editors count it a privilege to preserve their works in a book such as this.

Newington, Conn.

JOHN HUNTOON
General Manager, A.R.R.L.

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A History of Amateur Single Sideband

The current interest in single sideband was triggered off in 1947 when, on September 21st of that year, O. G. Villard, jr., W6QYT, put W6YX on single sideband in the 75-meter band and worked W6VQD. The 20-meter band was "opened" by W6YX on October 9th, when WØNWF was worked.

Exactly one week after hearing W6YX, on sideband, Art Nichols, WØTQK, had literally thrown together a 20-meter rig and was working W6YX, while scores of interested amateurs around the country were listening and finding, to their amazement, that a sideband signal could be copied on a normal communications receiver.

However, this was not the first amateur sideband operation. Back in 1933 Robert Moore, W6DEI, built and operated a sideband transmitter. It was described by him in *R9* magazine, and there were perhaps a half dozen sideband stations on the air back in 1934.

The basic sideband techniques are almost as old as radiotelephony itself, and communications companies have used sideband in commercial point-to-point service since the early 30s. Why were the amateurs so slow in utilizing this superior radiotelephone technique?

There are several contributing factors. Back in the early 30s there was not as much interest in phone as there is now. The usual receiver was a regenerative one, with or without r.f. stage, and superheterodynes were rather rare. The commercial point-to-point sideband stations used crystal-controlled transmitters and receivers always held as close to one frequency as possible, and reports in technical journals said that a tuning error of 20 or 30 cycles was the limit. This discouraged most amateurs who thought about trying sideband, because amateur radio is a "band" rather than a "channel" affair, and changing frequency is a large part of our operating.

But receiver stability sneaked up on amateur radio without any great fanfare, and by 1947 there were enough good receivers in use to copy the signals of W6YX, WØTQK and the others and establish the practicability of amateur sideband. It was also found that the tuning error could be on the order of 100 or 200 cycles and still permit acceptable copy, if one could forget concepts of "high fidelity" amateur phone. And, finally, a complexity that was frightening in the 30s is so commonplace now that it is no longer a consideration.

An Introduction to Single Sideband

There are several different methods of transmitting speech by radio; "single sideband" is one of them. "Amplitude modulation" (a.m.), is the method used in the regular broadcast band, all of the international short-wave broadcasting, and many of the mobile services (land, marine and air). Another method is "frequency modulation" (f.m.), which is used by the "hi-fi" f.m. broadcasting stations, TV sound channels, and many mobile services. Single sideband is the third method for transmitting intelligence; it is used by commercial services for many of their long-distance circuits. All of these methods are used in amateur radio.

The nature of a.m. and f.m. signals is such that they are very easy to tune in on receivers that are designed to receive them. These receivers are relatively simple and inexpensive to build, and even a young child can tune in a signal acceptably on them. Frequency stability isn't much of a problem. On the other hand, it

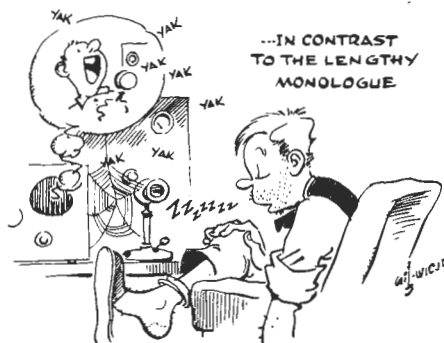


requires a certain amount of skill and understanding to tune in a single-sideband transmission, because a slight error in tuning will turn the signal into "monkey chatter." Also, if the signal is allowed to overload the receiver, it will become distorted and impossible to tune in with any setting of the dial.

The reason for the popularity of single sideband, despite its complexity when compared with a.m. or f.m., is that it offers practical operating advantages that can be obtained by no other known method. As is explained in articles to be found in this book, more of the power generated by the transmitter is used to produce speech at the receiver than when a.m. or f.m. is used. This means that for a given amount of available power the transmitting station can enjoy greater range and longer useful operating time. Through the use of "voice-controlled break-in" it is possible for several stations to operate on the same frequency in a rapid-fire "round table" that closely approaches a normal conversation, in contrast to the lengthy monologue-type contact customary on a.m. and f.m.

Transmitters

Sections of this book describe the two prin-



cipal methods of generating a single-sideband signal. These methods are called the "filter system" and the "phasing system." Single-sideband signals are usually generated at some fixed frequency and then "heterodyned" to the operating frequency and amplified. A single-sideband signal cannot be frequency-multiplied the way an f.m. signal can, and it is poor economy to generate the signal in the output stage as is customary with a.m. After heterodyning to the operating frequency, a single-sideband signal is brought to maximum power level in a "linear amplifier." A linear amplifier is an amplifier that amplifies the input signal with very little distortion; in technical terms it means that the amplifier has low intermodulation.

The generation of a single-sideband (abbreviated "s.s.b.") signal starts with the production of a "double-sideband suppressed-carrier" signal in a device called a "balanced modulator." A double-sideband suppressed-carrier signal (abbreviated "d.s.b.") is the same as an a.m. signal except that the carrier frequency has been suppressed. In the filter system of single-sideband generation, the double-sideband output from the balanced modulator is fed to a sharp filter, where one sideband is passed but the other is rejected. In the phasing system, two balanced modulators are used, to generate two double-sideband signals that differ only in phase. When the phase (and amplitude) relations between the two signals are proper, combination of the two signals will result in cancellation of one sideband and reinforcement of the other, to yield a single-sideband signal.

Filters of sufficient selectivity for s.s.b. generation can currently be built as high as 9 Mc. or so, and this sets the upper frequency limit of initial s.s.b. generation by the filter method. The phasing system has no such restriction, although usual practice is to generate

the signal in the 2- to 10-Mc. range. Both systems have their faithful followers; the phasing method is admired for its economy and its electrical subtlety, and the "brute-force" filter system for its dependable long-term stability.

Receivers

In many ways the problems of receiving s.s.b. signals are the same as those encountered in receiving code signals, and any good c.w. receiver is a good s.s.b. receiver, up to the point where the selectivity of the c.w. receiver is too great to pass all of the s.s.b. signal. But the same general features are required: a slow tuning rate, freedom from backlash, good signal-handling ability, and good frequency stability. Older communications receivers lacking in selectivity and signal-handling ability can often be made adequate by the addition of a "single-sideband adapter" that provides additional selectivity and a better detector. These adapters will use either the filter or the phasing principle to obtain the necessary selectivity. To avoid distortion, linearity must be maintained throughout the receiver, so it is necessary to exercise good judgment in setting the r.f. gain control when automatic gain control is not available, and this and s-l-o-w tuning are the two basic secrets of success in receiver tuning for s.s.b. reception.

Transceivers

It is universal operating practice for s.s.b. contacts to be made with all stations involved transmitting on the same frequency. The desirability of this can easily be appreciated; no retuning of the receiver is needed when successive stations take their turns in a round-table. If a station calls CQ, an operator wishing to answer will first "zero" his transmitter on the same frequency—i.e., will zero-beat his transmitting frequency with the incoming signal—since the CQer will be listening on that frequency. Other stations that may break in later will follow the same practice.

This has led to the development of "transceiver"-type equipment. A transceiver is a combined transmitter-receiver in which the same frequency-controlling element is used for

both transmitting and receiving, so that the transmitting and receiving frequencies are *automatically* the same. Usually, the s.s.b. generating circuits also provide the selective elements for receiving. This effects a considerable saving in components and bulk, as well as cost.

Many commercially-available transceivers provide for transmission and reception on only one of the two possible sidebands, as including circuits for sideband selection would add appreciably to the cost of the equipment. For this and other reasons there has been a species of standardization on the particular sideband used in the various amateur bands. Nearly all operation in the 3.5- and 7-Mc. phone sub-allocations is on *lower* sideband, while the *upper* sideband is used on 14, 21 and 28 Mc.

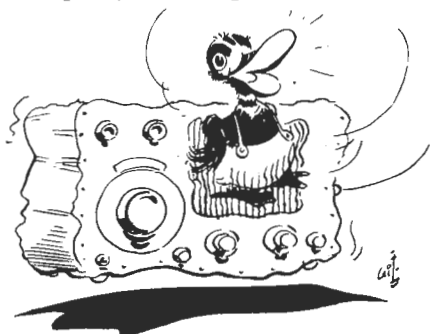
What Bands?

Technically, s.s.b. communication becomes more difficult at the higher frequencies. Setting and maintaining a receiving (or transmitting) system to within 50 or 100 cycles of a given frequency is relatively much easier at 3.9 Mc. than it is at 28.9 or 144.9 Mc. This is one of the reasons why the s.s.b. activity is greater on the lower-frequency bands than it is in the higher portions of the spectrum. However, the rewards of s.s.b. techniques are greatest on those bands where selective fading and/or weak signals are encountered, as on 14, 21 and 28 Mc. and v.h.f.

Equipment

Although this book contains many articles on the construction of s.s.b. equipment, it is realized that not everyone has the facilities or time to construct everything in his s.s.b. station. Fortunately, there are a number of good s.s.b. exciters and linear amplifiers on the market. A new, or second-hand, s.s.b. exciter offers a time-saving way of "getting your feet wet" in single sideband. Some older exciters deliver only a few watts of power, but even this is enough to try "barefoot" (directly into the antenna) for a few QSOs. Other exciters (and transceivers) will deliver well over 100 watts; in this power level they can certainly be classed as "transmitters." Some provide for optional a.m. or d.s.b. operation, while some of the latest designs restrict the operation to s.s.b. and c.w. or s.s.b. only. The optional a.m. operation is always at a lower power level than the s.s.b. or c.w., because a large percentage of the total power capability is used up by the carrier.

The power level of the station can be increased by adding a linear amplifier between the exciter and the antenna, and a number of such amplifiers are currently on the market. In selecting one, first make certain that your present exciter has sufficient output to adequately drive the amplifier. More-than-enough drive can always be dissipated in a supplementary load, but insufficient drive will limit the useful output from the linear amplifier.



TO AVOID DISTORTION IN THE RECEIVER, LINEARITY MUST BE MAINTAINED

» Here are the reasons why single-sideband is superior to other forms of modulated signal in radiotelephony. The soundness of the conclusions is confirmed by everyday amateur experience with s.s.b.—experience reflected in the fact that “sideband” is now the dominant mode on the popular h.f. bands.

Why Single Sideband?

DONALD E. NORGAARD, W6VMH

How and why can single sideband “buy” us better communications? First of all, a single-sideband signal uses up *less than half the space in the band* than that occupied by *properly-operated* a.m. or n.f.m. transmitters, regardless of power. Next, it doesn’t “waste any steam blowing the whistle”! By that is meant the relatively tremendous amount of power devoted to transmission of the carrier compared to intelligence-bearing sidebands. There just isn’t any “whistle blowing” to blot out the other fellow and rob yourself of “steam.” These things are mentioned first because they should be obvious and we want to start out agreeing with one another in this discussion.

Carrier and Sideband Relationships in A.M.

To keep things on a simple basis at first, assume that an ideal a.m. transmitter has a carrier *output* of 100 watts. We know that when this carrier is modulated, sidebands are generated in proportion to the strength of the modulating signal (until we reach 100% modulation), and that the carrier strength itself is not affected at *all* by modulation. A plot of the frequency spectrum (voltage *versus* frequency) of the simple case of steady 100% modulation of the carrier by a single tone (sine wave) of 1000 cycles would look like Fig. 1. The envelope (a plot of voltage *versus* time) would,

of course, have the appearance of Fig. 2. All right, so far? Our *Handbook* tells us that in a resistive circuit where the resistance stays constant the power is proportional to the square of

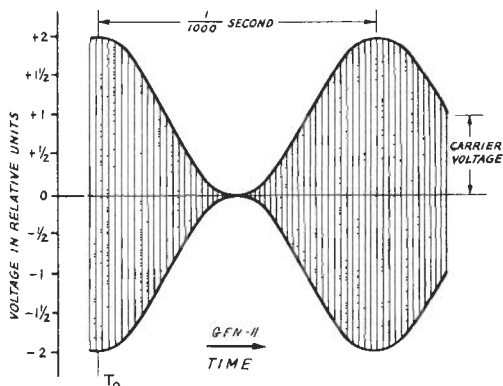


FIG. 2—Envelope of carrier 100% modulated by a 1000-cycle sine wave.

the voltage applied. In the case we are talking about, three voltages are applied; one is the carrier, and the other two are the upper and lower sidebands, respectively, in accordance with Fig. 1.

The voltage of each of the sidebands is half that of the carrier. Therefore, the power in each sideband is $(\frac{1}{2})^2$ times that of the carrier. Since it was assumed that the carrier output was 100 watts, the power in each sideband is 25 watts, and the *total* sideband power is 50 watts. This, incidentally, is the maximum single-tone sideband power that can be generated by amplitude modulation of a carrier of 100 watts. No one has ever been able to do better, because it just isn’t possible to do so. (It doesn’t help to overmodulate! This *cuts down* the desired sideband power and generates spurious sidebands called splatter.)

We can represent the information in Figs. 1 and 2 by means of a vector diagram and make some more calculations. In Fig. 3 the carrier voltage is given one unit length. Therefore, the upper and lower sideband voltages have one-half unit length, and are so indicated. Now, watch out for this one: In Fig. 3 the carrier vector is assumed to be standing still, though actually it makes one revolution per

From “What About Single Sideband?” *QST*, May, 1948.

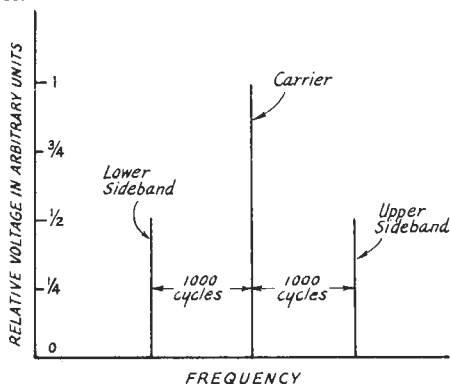


FIG. 1—Example of 100% modulation of a carrier by a single tone of 1000 cycles per second.

cycle of carrier frequency. Imagine you are standing at the origin of the carrier vector and are spinning around with it at carrier frequency. What you would see are the upper- and lower-sideband vectors rotating in *opposite* directions at the modulation frequency in such a way that the terminus of the last vector in the chain of three lies along the line of the carrier, bobbing up and down at 1000 cycles per second. As far as you could tell, the carrier vector does not move or change at all, and that is the impression Fig. 3 is intended to convey. At the instant of time (T_0 , Fig. 2) chosen for Fig. 3 the three vectors are all in line and add up to two voltage units. One two-thousandth of a second later the sideband vectors have rotated one-half turn each, and the three vectors add to zero, since $1 - \frac{1}{2} - \frac{1}{2} = 0$. This should make it easier to understand the relationship between Figs. 1 and 2 without too much trouble.

Now, here is the point of all this: The carrier vector is one voltage unit long—corresponding to a power of 100 watts. At the instant of time shown in Fig. 3, the total voltage is two units—corresponding to $(2)^2$ times 100, or 400 watts. One two-thousandth of a second later, the answer is easy—the voltage and power are zero. Therefore, the transmitter *must* be capable of delivering 400 watts on peaks to have a carrier rating of 100 watts. Stated differently, the excitation, plate voltage, and plate current must be such that the output stage can deliver this peak power. What about this? We are already up to 400 watts on a 100-watt transmitter! Yes, we are, and if the transmitter won't deliver that power we are certain to develop sideband splatter and distortion.

Under the very best conditions that can be imagined we need a transmitter which can deliver 400 watts of power on peaks to transmit a carrier power of 100 watts and a total maximum sideband power of 50 watts. What does this 100-watt carrier do for the transmission? The answer is it does nothing—for the simple reason that it does not change at all when modulation is applied. The carrier is just like

a hat rack—something to hang sidebands on. It seems silly to carry a hat rack around with us just so that we can say that we have brought two hats along. Yet, that is just exactly what we do when we hang two sidebands just so on a carrier and go out with the whole thing into our crowded phone bands to be jostled about. Far better to put on a hat and leave the hat rack home where it belongs. One hat? Certainly. It is ridiculous to go around trying to wear two dinky hats at the same time—especially in the rain!

Leaving the Carrier at Home

Sure, take a look at Figs. 1 and 3. Suppose we leave the carrier home and double the amplitude of each of our sidebands. This will still run our transmitter at its peak output capacity of 400 watts, all it can do. Well, the sideband power goes up all right. The sideband voltages are doubled, so our sideband power is four times what it used to be. That means each sideband is 100 watts, and our transmitter is not overloaded on peaks. The total sideband power is, of course, 200 watts. But this sideband power doesn't do much for us if it can't all be put to work. That is the situation with two sidebands and no carrier; the sideband power is in such a form that it doesn't lend itself readily to full utilization.

What if we leave one of the sidebands home, too? If we do, we can increase the voltage on the remaining one to two units and run our transmitter at its maximum peak power output of 400 watts. This time it is *all* sideband power. It so happens that sideband energy in this form is usable. Yes sir, all of it can be used, for it is just like c.w.! It is indeed, and we receive it in just the same way. All that is necessary is to set the b.f.o. in our receiver so that it is at the same frequency as the carrier we left home. Good. We don't have to carry our own hat rack around, and we don't have to go out with two little pint-size hats on either. Your host will let you hang your hat on his hat rack, and your hat won't know the difference, either, because the hat racks we are talking about are *identical*. What a fine thing that is. We put out 400 usable watts with a transmitter that could put out only 50 usable watts in the form of amplitude modulation.

Expressed in decibels, the ratio of 400 watts to 50 watts (8:1) is 9 db. But this isn't the complete story. The transmission covers only half the spectrum of the a.m. transmission and isn't blowing a loud whistle in the middle of it all. This kind of 8-db. gain doesn't bother the other fellow as much as if it were obtained with antenna gain on a.m. transmission.

Before climbing down from the ivory tower of theory we ought to see what hanging our hat on our host's hat rack really means. First of all, his hat rack has not been dragged through the mud and rain of propagation. It has our wet hat hanging on it and the hat

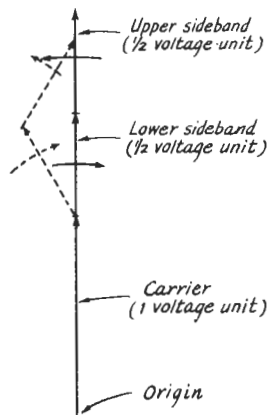


FIG. 3—Vector diagram of 100% modulation of an a.m. carrier at the instant (corresponding to T_0 in Fig. 2) when peak conditions exist. The broken vectors show the relationships at an instant when the modulating signal is somewhat below its peak.

won't fall off unless the hat rack is unsteady—it won't, provided we are not careless about how the hat is put there. The point is this: The sideband must be based on a good clean carrier of immaculate frequency stability, and our host's carrier must be stable, too. A good crystal-controlled oscillator or a really stable v.f.o. is a necessary part of a present-day transmitter, anyway, so there is no worry on this point. Receiver stability has become increasingly important through the years and it is quite likely that our host is today in possession of a fairly good receiver. At least, to hear him tell about it over the air or at the club, there never was a better one! But even if he doesn't



have the very best that can be constructed, he might be willing to steady it a little bit by hand or to do some tinkering with it in the free time between rag-chews and schedules (or CQs) so that he doesn't have to coax it along constantly. There is no denying that it can be done.

Transmitter Ratings

Back to earth again, we might worry about the little 100-watt transmitter straining itself to put out 400 watts, for that is what we said we wanted it to do for a short percentage of the time, but it probably would burn up if we kept that one sideband generated by the 1000-cycle tone pumping through it steadily. Fortunately, speech waveforms have a high ratio of peak to average power. It is average dissipated power that burns up tubes, so there is nothing to worry about on this score until we can learn how to talk with waveforms having a much lower ratio of peak to average power. Actually, the steady 100-watt carrier of an a.m. signal causes most of the dissipation in the 100-watt transmitter, but it was built to stand up under that kind of treatment.

While shrouded in theory, we were talking about *output* power, and managed to show that we could get 400 watts of sideband power output with single sideband at the same peak power that gave only 50 watts of sideband power in the case of a.m. That's fine for comparison purposes on a theoretical basis, but there is the practical matter of efficiency to consider. Let's lean over backward and say that a *good* Class C plate-modulated amplifier such as the one in our ideal 100-watt a.m.

transmitter runs with an efficiency of 80%. Neglecting the fact that the total input under modulation with speech is somewhat higher than the carrier input (which is $100/0.80=125$ watts), the dissipation in the output stage is 25 watts. Let us say, however, that the modulation still drives the transmitter to its peak output power of 400 watts, but has very low average power. Therefore, the peak sideband power output is 50 watts, with very low average power. Here is a strange way of rating things, but it means something: The peak *useful* sideband power is 50 watts obtained with a final-stage dissipation of slightly over 25 watts in the a.m. transmitter. The peak input power is, of course, $400/0.80=500$ watts, since the efficiency of 80% is pretty nearly constant with this type of operation. You have already guessed what the next thing is. The peak useful efficiency is

$$\frac{\text{peak useful power output}}{\text{peak input}}$$

or $50/500=10\%$. Who says high efficiency? This figure is not the true efficiency of the output stage—that's the assumed 80%—but it is the "communication" efficiency. The transmitter, of course, cannot tell the difference between carrier and sideband signals it deals with, so we must be satisfied with 10% "communication" efficiency as we have defined it.

Now let's look at the single-sideband situation. The output stage must be a linear amplifier. This linear amplifier will have characteristics quite similar to Class B modulators used, for instance, in the little 100-watt plate-modulated a.m. transmitter. Suppose we put into this transmitter the same speech waveform we used in the example above. This wave had a high peak-to-average power ratio, if you recall, and we were concerned only with conditions during the peak period. Things are adjusted so that the peak *output* is 400 watts in order to fall into our theoretical pattern. The theoretical maximum peak efficiency of a linear amplifier is 78.5%, but nobody ever got that much out of such an amplifier. However, with modern tubes we can get 70% peak efficiency quite comfortably, so let's use that figure in our calculations. All right, the peak power input is $400/0.70=572$ watts, which, if sustained, would get some tubes mighty hot at 70% efficiency, if they could dissipate only 25 watts. This signal isn't sustained, however, for we assumed a speech input wave having a high peak-to-average power ratio, and it is average power that makes plates incandescent. Well, all of this 400-watt peak output is useful "communication" power, and it is obtained at 70% efficiency. Thus we can say that the "communication" efficiency of the final stage of this single-sideband transmitter is 70%.

All this does sound wonderful. What about plate dissipation in the final stage? If we neglect the average dissipation during modula-

tion with our speech wave, the one might say that the total dissipation is close to zero. It certainly would be if we had vacuum tubes with linear I_p -vs.- E_p curves right down to cut-off. But there are plenty of tubes that make good linear amplifiers, and they do not have linear I_p - E_p curves at all. This generally means that the linear amplifier is operated in such a way that there is d.c. input even though there is no signal input. This d.c. input power, of course, heats the tubes when no signal is there, and represents most of the dissipation that the tubes are called upon to stand under conditions of speech modulation. In most cases good linearity is obtained when the no-signal input plate current is about 5% of the maximum-signal plate current. This means that the no-signal dissipation is about 5% of the maximum input power, since the d.c. input voltage is held constant. Therefore, the total dissipation would be something close to $572 \times 0.05 = 28.6$ watts.

That's within gunshot of the 25 watts which our a.m. transmitter burned up in the plates of its tubes. You have guessed it again; the output stage of the single-sideband transmitter delivering 400 watts peak communication output can use the same tubes that are necessary in the 100-watt-carrier-output a.m. transmitter which delivers 50 watts peak communication output.

The foregoing comparison isn't absolutely accurate, since the actual waveform of speech input is unknown. But it is a fair comparison, and experience and tests support the argument. That is what really proves the point.

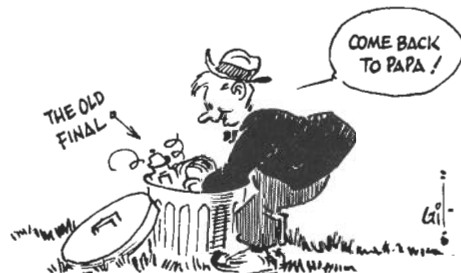
Signal-to-Noise Ratio

The business of receiving a single-sideband signal probably needs a little clarification. Let us examine the characteristics of receivers and find out what happens when a signal is received.

Theory says (and experience bears this out) that noise power is proportional to the effective bandwidth employed in a system. The noise we are considering now is "thermal noise," frequently called "receiver hiss." This is not to be confused with man-made noises of the impulse type such as automobile ignition, commutation noises, or even an interfering radio transmission. However, it is no figment of the imagination, since it can be measured, and, equally important, heard in our receivers. The single-sideband signal requires only half as much i.f. bandwidth as the a.m. signal requires to provide a given audio bandwidth. Therefore, we should not use more receiver bandwidth than the type of transmission requires us to use, since we do want to deal with pertinent facts in comparing one system with another. Reducing the effective receiver bandwidth by a factor of two cuts down the noise power output of the receiver by the same factor, when only thermal noise is considered. But this

reduction in bandwidth does not affect the ability of the receiver to respond to all of the sideband power it receives from a single-sideband transmitter. This begins to look as though we receive all of the single-sideband power available at the receiving location and hear only half the noise power that would be heard when receiving an equivalent a.m. transmission with the same receiver gain. This is absolutely true, so in haste we might put in another 2:1 factor of improvement in signal-to-noise ratio simply because we measure half the noise power when the bandwidth is cut in half. Apparently, this would then give the single-sideband system a 12-db. (16-to-1 power ratio) signal-to-noise ratio gain over the idealized a.m. system. In one sense this is true when considering power relationships alone, but before we reach any conclusions we should see how a detector responds to signals furnished to it by an i.f. amplifier.

We see from Figs. 1, 2 and 3 that the two sidebands in our idealized a.m. system each have 25% of the carrier power, but 50% of the carrier voltage. In an idealized a.m. receiver the detector is a linear or envelope detector, and linear detectors respond to voltage—definitely not to power as such. Therefore, the detector output corresponds to the envelope voltage, giving a demodulated signal voltage having a peak value equivalent to one voltage



unit if we assume that each sideband is $\frac{1}{2}$ voltage unit at the detector. The demodulated signal in this case is our modulating signal, a 1000-cycle sine wave. This may be expressed as one unit of 1000-cycle audio power at the detector output. The characteristics of thermal noise, however, are such that this same detector produces noise power output in proportion to the i.f. bandwidth, which, of course, is necessarily twice as great for a.m. reception as it is for single-sideband reception. So we can say that the a.m. receiver detector output (or audio output) has one signal power unit and two noise power units when two sidebands totaling one-half a power unit are applied to the detector. (These units are not necessarily the same, but are in the same classification. Obviously, this depends on the relative strengths of the signal and the noise.)

In order to produce the same detector output when only one sideband is applied to the

detector (along with a sufficient amount of locally-generated carrier at the correct frequency) its voltage must be the same as the combined voltage of the two sidebands that were applied in the case of a.m. reception. The power in this one sideband is twice the combined power of the two sidebands which produce the same voltage output from the detector. This is the same thing we saw when comparing total sideband power of two sidebands with the power of one sideband having the same voltage as the combined voltage of the two sidebands, when we discussed the transmitters. At the receiver we can say that we get one signal-power-unit audio power output from the detector with one unit of sideband power input applied to the detector, and one unit of noise power, since we can slice the i.f. bandwidth in half to reduce the noise power output by half.

It doesn't take much figuring to see that if it requires twice as much single-sideband power as it does double-sideband power, to get the same signal output power from a receiver with the noise power output half as much for single-sideband operation as for double-sideband operation, nothing has been gained in *signal-to-noise ratio*. But nothing has been lost, either. Since measurements confirm the reasoning we have just been through, we should give back that 3 db. we thought at first we had earned by reducing the bandwidth by two to one. Therefore, on an idealized theoretical basis we must conclude that *single-sideband operation can give 9-db. signal-to-noise ratio improvement over amplitude modulation operating at the same peak power output*.

Back again from the ivory tower we begin to wonder what significance this 9-db. system gain has, since we arrived at this figure on an *idealized basis*. This idealized condition included consideration of only the necessary facts in order to avoid confusion. But to the amateur, confusion in the form of QRM is not avoidable except under idealized conditions, which seldom, if ever, occur in the ham bands. In fact, commonplace man-made disturbances so completely mask out thermal noise in a good receiver operated on our low- and medium-frequency bands that we should try to evaluate the performance of single sideband working under the conditions we know we do have.

Impulse noise—the clicks and pops we hear—produces detector output voltage more or less proportional to bandwidth. Immediately we can say that single-sideband reception at half bandwidth will give us almost 3 db. receiver s/n gain with this kind of noise, provided we cut down the bandwidth in the right way. That's fine, because we can get a practical gain of almost 12 db. over this type of noise when we use single-sideband transmission. That's the kind of noise we want to beat!

QRM in A.M. and S.S.B. Reception

Another type of QRM is the usual one—interfering radio transmissions. These fall into several classifications which deserve individual consideration. The first case is that of interference which has a signal strength definitely lower than that of the desired transmission. With conventional receiver conditions (a.m. reception), all of the interfering energy that reaches the detector heterodynes with the carrier of the a.m. signal being received and produces a beat note between the two carriers, along with “monkey chatter” caused by the voice sidebands of the undesired transmission beating with the relatively strong desired carrier. A rejection filter may be used to put a



notch in the i.f. passband so that the carrier heterodyne is practically eliminated, but most of the monkey chatter remains. (This depends, however, on the shape of the i.f. passband when the notch filter is switched in.) In almost every case of this kind the heterodyne between carriers is much more bothersome than the monkey chatter, so it pays to notch out the interfering carrier. With single-sideband reception, the exposure to interference is cut down to half, but any interfering signals (carriers or sidebands) that lie within the band occupied by the desired transmission will cause heterodynes and monkey chatter in proportion to their strengths. The notch may be used to eliminate one carrier heterodyne, but that is about all it can do. The advantage of single-sideband reception in this case is principally that, on the average, only half the number of heterodynes will be heard, where interference is the only disturbance to otherwise flawless reception. Well, that helps.

The case of an interfering signal of about the same strength as the desired signal is next. If nothing is done to eliminate the interfering carrier before it reaches the detector, all of the sidebands that are passed by the i.f. amplifier are demodulated against each carrier, and there is as much monkey chatter caused by the desired sidebands beating with the interfering carrier as there is from the undesired sidebands beating with the desired carrier. In addition, there are usually equal amounts of halfway-intelligible speech outputs from each trans-

mission. Of course, the heterodyne of the carriers is by far the loudest signal heard, and it consists of a fundamental heterodyne note and a series of fairly strong harmonics throughout the audio band. Add a little QSB on both signals to this picture and not much is left of either signal. When the carrier of the interfering signal is put in the rejection notch a lot of the curse is removed. The remaining monkey chatter is, of course, more bothersome than in the case where the interfering signal was not so strong. With single-sideband reception under the same conditions, an interfering carrier produces a single-tone heterodyne, and the interfering sidebands produce monkey chatter, but nothing intelligible. Use of the notch filter can eliminate the carrier heterodyne, leaving only monkey chatter. Here again, the exposure to QRM is cut in half, since the receiver bandwidth can be cut in half without sacrifice of audio bandwidth, so the situation is similar to the first case (interference weaker than the desired signal) but, of course, worse. When the desired transmission is besieged by more than one interfering signal of equivalent strength only one of the carriers can be put in the notch, and the others have to be tolerated along with monkey chatter. The remaining heterodynes, however, are definitely less disturbing since they are not distorted in the detector. What is left is then purely a fight on the basis of strength and intelligibility. Single-sideband intelligibility is definitely of a superior nature.

When the interfering signal is stronger than the desired one, the stronger is the only intelligible one in a.m. reception, since the situation is the reverse of the first case. This is true until at least the undesired carrier is notched down so that it does not reach the detector. But all the troubles are not so easily disposed of. The low-level speech sidebands of the interfering transmission appear as monkey chatter, while the stronger ones which exceed the level of the desired carrier serve as virtual carriers against which the desired carrier and its sidebands are demodulated to produce whistles, groans, and monkey chatter of a kind that is horrible. It's all a weird mess in spite

of anything that can be done with the very best conventional receiver. With single-sideband reception of the desired weaker signal, all of the undesired noises are, of course, louder than in the previous cases, but that is the only difference. Notching out the chief offender—the interfering carrier—frequently wins the battle, but it is not certain to do so. After all, there are limits, but you have a fighting chance, because somewhere there in the background is perfectly clean intelligible speech without distortion. The only trouble is that the monkey chatter may be louder, but not funnier. Of course, two strong interfering transmissions partly or wholly within the receiver passband make just that much more trouble. Here again, the fact that the receiver bandwidth can be cut in half cuts down the average probability of trouble by a factor of two to one.

It has been assumed in the discussion of the QRM problem that the receiver is not overloaded by signals, and that the interfering signals are of good quality and frequency stability. The difficulties are greatly compounded when "rotten" signals are involved. The rotten signal not only does more damage than necessary to others using the band, but is out of luck when it is the recipient of QRM from other transmissions.

When single-sideband signals are in the rôle of interfering signals, the principal effect is monkey chatter unless the sideband strength is sufficient to put the interference in the class of a signal which exceeds the carrier strength (of an a.m. signal). Single-sideband reception clears up this difficulty, but does not eliminate *all* interference. Single-sideband reception of standard a.m. and n.f.m. signals with exalted carrier is possible and feasible. Such a receiving method improves the situation tremendously, but the full advantages cannot be exploited until single-sideband transmissions are the only ones involved.

Laboratory tests and on-the-air experience with single-sideband transmitting and receiving equipment indicate that single-sideband signals are the most QRM-proof signals that are known, as well as the least troublesome in creating QRM.

» Explaining the principles of the "balancing" or "phasing" method of generating a single-sideband signal, by an author who has done outstanding work in this field.

The "Phasing" Method of Generating Single Sideband

DONALD E. NORGAARD, W6VMH

Fundamentally, the "phasing" method of generating a single-sideband signal consists of removing one of the sidebands by means of a balancing process rather than by filtering.

The principle employed may be explained by reference to Figs. 1A and 1B, which are vector diagrams showing the relationship between carrier and sidebands produced in amplitude modulation. In Fig. 1A a carrier is shown in

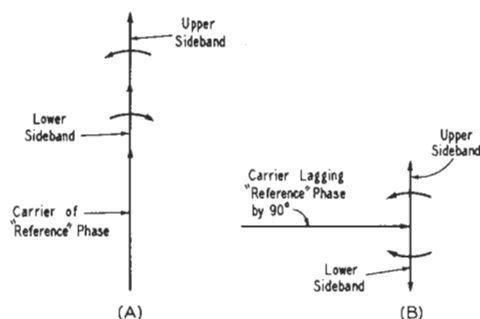


FIG. 1—The carrier and sideband relationship required to generate a single-sideband signal by the "phasing" or "balancing" method. The modulating signal in B leads the modulating signal in A by 90° . When the two signals represented by A and B are combined, the upper sidebands add and the lower side bands cancel out, resulting in a single-sideband signal.

"reference" phase, and the positions of the sideband vectors indicate that peak-envelope conditions exist at the instant shown. In Fig. 1B a carrier of the same frequency but 90° away from that of Fig. 1A is shown. The two sideband vectors in Fig. 1B indicate that the envelope has a value (at the instant shown) equal to the carrier; that is, the modulating signal is 90° away from that which gave the conditions shown in Fig. 1A.

If the conditions shown in Fig. 1A exist at the output of one modulating device at the same instant that the conditions indicated in Fig. 1B exist at the output of another modulating device, and if the sideband frequencies and magnitudes are the same, the simple sum

of Figs. 1A and 1B will consist of carrier and upper sideband only. It can be seen that the lower-sideband vectors are equal in magnitude and opposite in direction, and hence would cancel one another. How can this result be obtained in practice?

The vector diagram of Fig. 1A might be said to represent the output of a modulated amplifier where a carrier of reference phase is modulated by a tone of reference phase. Thus, Fig. 1B would represent the output of a second modulated amplifier where a carrier of the same frequency but 90° displaced from reference phase is modulated by a tone that is also 90° displaced from its reference phase. To make the whole thing work, the frequencies of all corresponding signals represented in the two vector diagrams must be exactly the same. This would suggest an arrangement such as Fig. 2, which would operate satisfactorily if the 90° phase-shift devices held amplitudes and phases of the respective signals to agree with the requirements indicated in Figs. 1A and 1B. The carrier phase-shifter is easy to

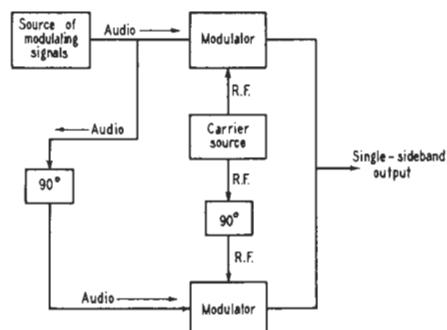


FIG. 2—A block diagram showing the circuits required to generate a single-sideband by the method of Fig. 1. This is an impractical method because there is no known means for obtaining the 90° audio shift over a wide range of frequencies.

build, since the carrier frequency is constant, but the modulating signal phase-shifter might not be, since it must work over a wide range of frequencies. The arrangement of Fig. 2 works in principle but not in practice, for any wide range of modulating frequencies.

From "A New Approach to Single Sideband," QST, June, 1948.

It so happens that two phase-shift networks having a *differential* phase shift of 90° can be inserted between the source of modulating signals and the modulating devices to generate sets of sidebands which can be combined to cancel one of the sidebands as indicated earlier. This leads to an arrangement such as that shown in Fig. 3, where the symbols " α " and " β " indicate the two networks that have a *difference* in phase shift of 90° over any de-

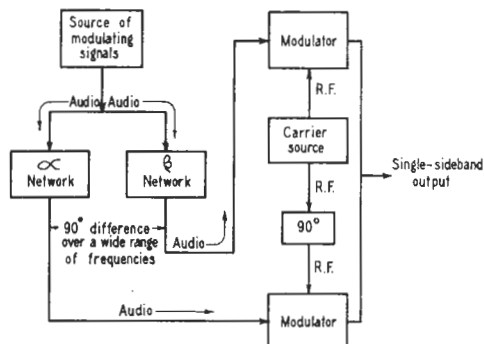


FIG. 3—The system outlined in Fig. 2 becomes practical by using two audio channels (α and β networks) with a constant phase difference of 90° .

sired range of modulating signal frequency. The principle of Fig. 3 has been found to be practical for several important reasons:

1) A carrier of any desired frequency can be used. This means that heterodyning the output to a higher frequency is not at all necessary as is the case when a filter is used to eliminate one sideband.

2) Conventional parts may be used in any and all of the circuits. There is no "problem of the filter." The cost, therefore, is low.

3) Any desired range of modulating frequencies may be employed. There is no theoretical limit to how low or how high these frequencies may be but, of course, there are practical limits. The phase-shift networks can be designed to cover a frequency range of 7 octaves, far more than is necessary for speech.

4) Modulation may be accomplished at any chosen power level. In the interest of efficiency, it is generally wise to carry out this

portion of the process at receiver-tube level, using linear amplifiers to build up the power.

5) Simple switching may be provided so that amplitude-modulation, phase-modulation or single-sideband signals may be generated.

The characteristics of typical wide-band phase-shift networks are shown in Fig. 4. It can be seen that the differential phase shift averages 90° over a frequency range of at least 7 octaves. Of course, the ideal differential phase shift is exactly 90° , and the excursions of the actual phase-shift curve are $\pm 2^\circ$ from this value. The ratio of undesired sideband to desired sideband is dependent upon this deviation, the most unfavorable points being at the peaks and valleys of the differential-phase-shift curve. The ratio

$$\frac{\text{undesired sideband}}{\text{desired sideband}} = \tan\left(\frac{\delta}{2}\right),$$

and for $\delta = 2^\circ$,

$$= \tan\left(\frac{2^\circ}{2}\right) = 0.0174, \text{ or } -35 \text{ db.}$$

The symbol δ represents the deviation of the actual performance from the ideal 90° , and, in the above example, δ was taken at its maximum value. The average attenuation of the undesired sideband is more than 40 db. over the band of modulating frequencies between 60 and 7000 c.p.s. There is little to be gained by improvement of this ratio, since subsequent amplifier distortions can introduce spurious components in sufficient amounts to mask any improvement gained by idealizing the phase-shift network characteristics.

A Practical Exciter Layout

While the block diagram of Fig. 3 is useful in explaining the principle of generating single-sideband signals, it does not represent a complete single-sideband exciter with enough gadgets to satisfy a person with a practical turn of mind. There is little to be gained by using single sideband unless the carrier is attenuated, but Figs. 1A, 1B, and 3 do not indicate this. Therefore, Fig. 5 is offered as a workable system that provides for carrier attenuation, amplitude modulation, phase modulation, single sideband, operation on 75- or 20-meter phone,

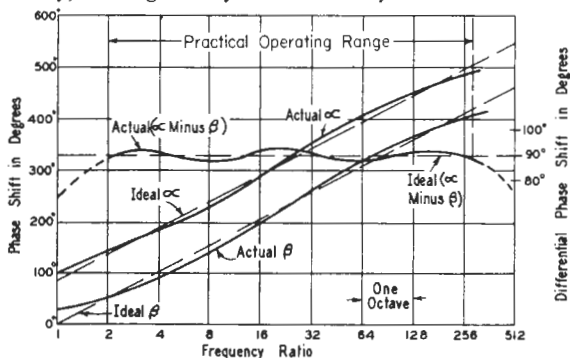
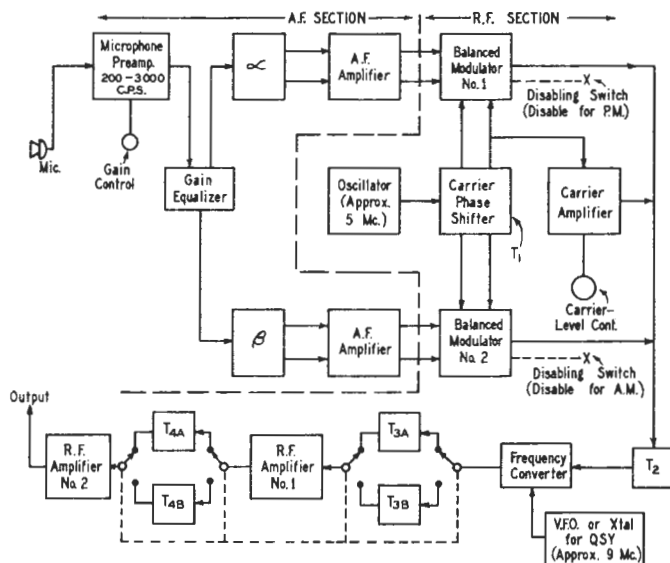


FIG. 4—This plot shows how the 90° difference between the α and β networks is maintained over a wide frequency range. The scale for the phase difference is given on the right-hand side of the graph.

FIG. 5—Block diagram of an exciter capable of generating s.s.b., a.m. and p.m. signals on either the 75- or 20-meter bands. Table I gives a description of the various components.



and QSY within these bands. If multiband operation and QSY are not desired, modulation can be accomplished at the operating frequency by appropriate simplification of the arrangement of Fig. 5.

It is not the purpose of this article to give specific circuit-design data for a complete single-sideband exciter; rather the purpose is to put out the over-all features that *must* be observed in order to satisfy the requirements of this system of generating single-sideband signals. For instance, the design of the bandpass circuits indicated in Fig. 5 is beyond the scope of this article. The advantage of using such an arrangement designed to cover the amateur band in use is that no tuning adjustments whatsoever need be made when it is desired to QSY. With ordinary circuits, best operation usually demands retuning when large percentage changes in frequency are made. However, ordinary tuned circuits can be substituted for the bandpass transformers, as in any transmitter.

A conservative output rating for an 807 output stage would be 30 watts peak, under drive conditions where the grid takes no current (Class AB₁). If suitable bias and drive are supplied to the 807, a conservative 50 watt peak output may be obtained. In either case, the output power is sufficient to drive additional amplifiers of fairly-sizeable ratings or to use directly as a low-power single-sideband phone transmitter.

The functional block diagram (Fig. 5) might appear formidable at first glance, but the whole arrangement lends itself to rather simple circuit design. Separate consideration of the two portions of Fig. 5 should not be taken to indicate independence of one from the other. It is well to keep in mind that in this system the audio-frequency circuits and the radio-fre-

quency circuits must work hand-in-hand in order to generate single-sideband signals of superior quality.

Notes on the Audio System

The audio-amplifier and phase-shift circuits are straightforward. The important consideration is that the phase-shift and amplitude relationships determined by the phase-shift circuits must be preserved over the entire voice range in succeeding parts of the system. Fortunately, there is nothing difficult about it, once the objectives are clearly in mind. These are:

- 1) Low harmonic distortion and noise.
- 2) Vanishingly small discrepancies in phase-shift and amplitude response.
- 3) Ease of control and adjustment.
- 4) Simplicity and low cost.
- 5) Stability of characteristics.

Most microphones in current amateur use require low-level amplification (the usual microphone preamplifier) to bring their output signals up to, say, a level of one or two volts. This is the job required of the audio amplifier ahead of the α and β phase-shift networks.

This is as good a time as any to mention the desirability of including in the "preamp" a bandpass or low-pass audio filter to pass the important speech band out to 3000 cycles or so, to conserve space on the bands. The operation of the rest of the circuits in the system in no way requires this, but good sportsmanship in the use of our bands does. It is good practice to eliminate unnecessary low frequencies, too, concentrating on the portion of the audio spectrum between 200 and 300 c.p.s. for maximum intelligibility. Why do anything about it at all, if the system as such does not require it? The answer has two important aspects—important to *you* as an occupant of the bands we share:

1) Intelligible speech does not require transmission of frequencies higher than 300 c.p.s. To do so adds practically nothing to intelligibility but does increase the space in the band required for transmission. It boils down to the fact that we want the "other fellow" to use as little of our bands as possible, and the Golden Rule certainly does apply in this matter. In addition, regardless of how "high fidelity"-minded one may be, crowded bands force the operator who listens to the transmission to restrict his receiver band-width so much that he receives only what is necessary, if even that much.

2) Elimination of frequencies below 200 c.p.s. removes a large percentage of the high-energy speech components that do not contribute to intelligibility. Such elimination permits the transmitter to concentrate its efforts on only the *essential* portions of speech power. In practice, this means something like 3 to 6 db. in system effectiveness. Two or three dollars spent on a suitable audio filter (and that's all one should cost) can give a transmitter a communication effectiveness equivalent to doubling or quadrupling its output power.

Notes on the R.F. System

Considerable flexibility is possible in the design of the radio-frequency portion of the block diagram in Fig. 5. The objectives in this portion of the single-sideband system are:

- 1) Very high order of frequency stability.
- 2) Provision for 90° r.f. phase shift in the excitation for the two balanced modulators.
- 3) Ease and stability of adjustment.
- 4) Absence of r.f. feedback.
- 5) Low distortion in modulation and subsequent amplification.
- 6) Provision for adjustable carrier level; generation of a.m., p.m., and single-sideband signals; output-level control.
- 7) (optional features) Operation on 75- or 20-meter bands; easy QSY within each band; choice of sideband transmitted.

Obviously, a number of methods exist for accomplishing these objectives. Many of the possible methods that may occur to the designer will satisfy the requirements quite well; some will not. Others, while technically adequate, may be difficult to adjust or may be impractical in some other way. Since the handling of radio frequencies is concerned in this portion, good mechanical layout and construction is of considerable importance. Also, since stability of adjustment is one of the principal objectives, it is a good idea to provide some sort of locking arrangement for the balance controls to prevent accidental shifting of their positions.

Balanced Modulators

Fig. 5 indicates the use of two balanced modulators. A little explanation might be helpful in understanding why and how balanced modulators are used.

In amplitude modulation the maximum strength of any sideband that can be produced is one-half the strength of the carrier. Since the carrier must be present in order to be modulated, but is not needed afterward (in single-sideband transmission, that is) it can be balanced out. This, then, is one job that the balanced modulator is called upon to do—namely, to permit sidebands to be generated, but to balance out the carrier after it has served its purpose. There are many forms of balanced modulators; some balance out one or the other of the two signals supplied; others can balance out both input signals. But none of them can balance out *one* sideband and not the other. Nature itself seems to be quite positive about that.

Since the signal that is to be balanced out is an alternating-current wave, it is necessary in the process to take account of phase relationships as well as magnitudes. Unless the two signals which are to be balanced have a phase difference of exactly 180°, perfect balance cannot be obtained by any amount of adjustment of amplitudes alone. This, incidentally, may explain why trouble is sometimes encountered in neutralizing an amplifier, since the same principle is involved. In the case of the balanced modulator, the perfection of balance required is usually quite high, and some means

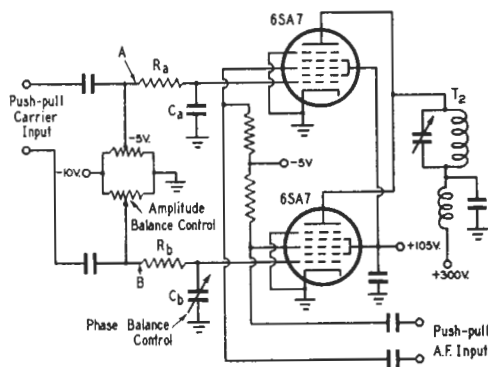


FIG. 6—A typical balanced modulator, using 6SA7 tubes. Provision is included for obtaining amplitude and phase balance of the r.f. (carrier) input.

for satisfying the conditions necessary for balance must be provided. Very few arrangements automatically provide the conditions necessary for perfect balance and frequently those that do are limited to operation at low frequencies, where circuit strays have negligible effect. It has been found practical to "grab the bull by the horns" and use some arrangement where separate phase- and amplitude-balance adjustments are provided, rather than to hope for a fortuitous set of conditions that might permit balance.

The circuit shown in Fig. 6 illustrates this philosophy. Fundamentally, only one of the tubes need be supplied with modulating sig-

nal, two tubes being necessary only to allow balance of the undesired component (the carrier) in the output. If, however, each tube is made to generate sidebands as well as to balance the carrier from the other, the ratio of residual unbalanced carrier signal to desired output is made smaller at low cost. Likewise, even small amounts of the modulation defect known as carrier shift are effectively reduced. The carrier signals at points A and B in Fig. 6 are made as nearly equal in magnitude and opposite in phase as is feasible using circuit components of ordinary commercial tolerances. The RC circuit between point A and grid No. 1 of the first modulator tube (a 6SA7 converter tube in this example) may be designed to provide about 20° phase shift at the operating frequency, by suitable choice of R_a and C_a . The RC circuit in the other grid can be designed to produce variable phase shift from 10° to 30°, by adjustment of the trimmer capacity, C_b . This permits a phase correction of $\pm 10^\circ$ —usually sufficient to insure perfect phase balance of the signals applied to the tubes. No attempt is made to equalize the magnitudes of the signals in the grid circuits because it is almost too much to expect that a perfectly-balanced pair of tubes could be found in order to take advantage of balanced amplitudes. Instead, the function of amplitude balance is accomplished by means of a bias adjustment on one of the tubes of the pair, so that the carrier signals are balanced out in the plate circuit of the tubes. That, incidentally, is what must happen anyway, regardless of the method used. The picture is completed by applying push-pull modulating signals to the No. 3 grids so that the sidebands produced by the separate modulation processes in each tube add together in the common plate circuit. The audio-frequency component balances out in the plate and screen circuits, this being a case of a balanced modulator that balances against each of the input signals. However, slight unbalance of the audio-frequency signals does absolutely no harm in the particular application of this circuit, so no provision is made for balance adjustment at low audio frequencies.

In any balanced modulator the efficiency is necessarily low, since at least one of the input signals is dissipated in the modulating elements or associated circuits. In the case of a balanced modulator that suppresses the carrier, the efficiency cannot possibly be greater than 50%. The efficiency obtained in practice is more like 5% to 10%. Where two balanced modulators are used (as in Fig. 5) the efficiency is still lower, since the unwanted sideband is dissipated. This situation leads to the choice of generating a single-sideband signal at very low power level where the inescapably low efficiency in the generation of the signal wastes no large amounts of power.

Good operating characteristics are obtained with 6SA7 tubes in this application when the

Table I
Explanation of Fig. 5

Microphone preamplifier	Sufficient gain to bring microphone output to a voltage level of approx. 2 volts, peak-to-peak.
α, β	Phase-shift networks.
A.F. amplifier	Push-pull self-balancing amplifier with good phase and amplitude characteristics. Maximum output required approx. 2 volts, peak-to-peak.
Balanced Modulators 1 and 2	Two 6SA7 tubes (in each). See Fig. 6 for details.
Carrier phase-shifter, T_1	5-Mc. double-tuned transformer with push-pull output from each winding at low impedance. Output on each line 2 volts, peak-to-peak.
Carrier amplifier T_2	6SJ7 tube. 5-Mc. double-tuned transformer.
Disabling switches	Bias controls for No. 3 grids of modulators. Can be ganged to permit s.s.b.-a.m.-p.m.
Carrier-level control	Bias control on grid No. 1 of carrier amplifier. Minus 10 volts to minus 3 volts range.
T_{3a}, T_{4a}	Bandpass double-tuned transformers to cover 75-meter phone band.
T_{3b}, T_{4b}	Bandpass double-tuned transformers to cover 20-meter phone band.
Frequency converter R.F. amplifier No. 1	6SA7 converter tube. 6AK6 beam tube. Operates as Class A amplifier.
R. F. amplifier No. 2	807 beam-power output tube. Can be operated as Class A or B amplifier.

No. 1 and No. 3 grids are supplied with maximum signals of about 1 to 2 volts peak-to-peak, at a bias of about 5 volts, negative. Other voltages are the same as recommended for converter service.

As in the case of the audio system, the radio-frequency circuits can employ receiving tubes of extremely modest ratings up to the point in the system where the signal levels reach the power-tube class. For instance, the r.f. portion of Fig. 5 up to the grid circuit of the output stage would somewhat resemble in over-all magnitude and construction the i.f. portion of an average communication receiver. The versatility of Fig. 5 should make it attractive, although some of this versatility is obtained at the expense of circuit complication not fundamentally a part of single-sideband operation. This is apparent when comparing Fig. 5 with Fig. 3.

» A brief theoretical discussion of lattice crystal filters, and some pointers on building and adjusting a practical filter.

Crystal Lattice Filters

C. E. WEAVER, W2AZW, AND J. N. BROWN, W3SHY

The ability of receivers to attenuate the undesired adjacent channel signals is termed "skirt selectivity." The filters to be described achieve high attenuation outside the passband through the very high "Qs" of the crystals themselves. In some cases, the crystals yield Qs of well over 10,000, which are certainly not obtainable in coil- and capacitor-tuned circuits.

Theory

The equivalent electrical circuit of a piezoelectric crystal is shown at A in Fig. 1. The circuit has both a series-resonant frequency and a parallel-resonant frequency. This is shown graphically in B, where the reactance of the equivalent circuit is plotted for all frequencies between zero and infinity. The series-resonant frequency, f_r , occurs first, where the curve crosses the zero-reactance line, and the parallel-resonant (antiresonant) point, f_a , occurs where the curve rises to high values of inductive reactance (+) and then breaks sharply through

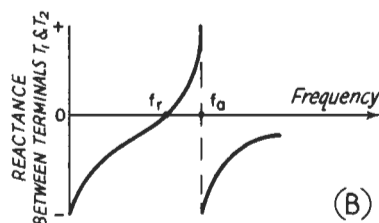
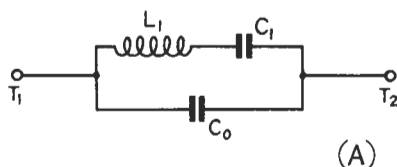


FIG. 1—The equivalent electrical circuit of a piezoelectric crystal (A). The reactance varies with frequency as in (B).

FIG. 3—Schematic diagram of a two-section crystal lattice filter.

T₁, T₂, T₃—Replacement-type i.f. transformers.

X_A, X_B—Matched pairs of FT-241 crystals (see text).

C₁—1- to 5-pf ceramic capacitor (see text).

From "Crystal Lattice Filters for Transmitting and Receiving," QST, June, 1951.

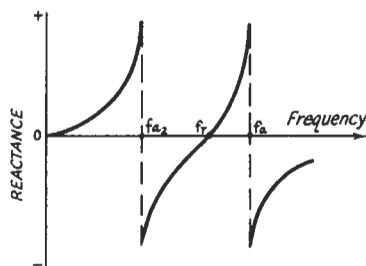
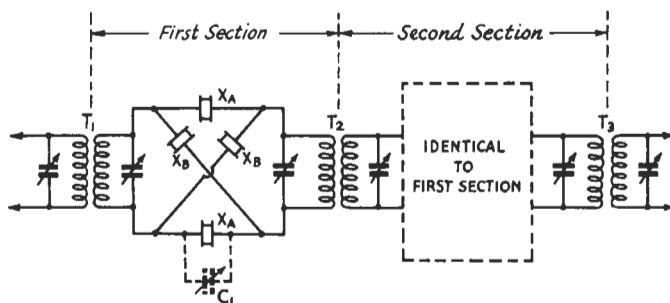


FIG. 2—Reactance plotted against frequency for crystal shunted by an inductance.

zero to a high capacitive (−) reactance. For most crystals, the two resonant frequencies occur within a few hundred cycles of each other. It is these two resonance points and what we can do with them that will occupy our attention for a moment. The problem is to spread these two resonant frequencies so that the crystals can be used as elements in a filter network. This "spreading" can be done by using either a series or a shunt inductance with the crystal. Fig. 2 shows the result of shunting a crystal with a coil. You will note that not only have we f_r and f_a but we have created a new parallel-resonant point, f_{a2} , which will be of use to us later.

Now, let's choose two pairs of identical crystals and connect them as shown in Fig. 3. You will notice that the shunt coils mentioned above have been moved to the input and output of the lattice network. This is accomplished by a mathematical transformation beyond the scope of this article. Suffice to say, the coils have the same effect as if they were connected directly across the crystals. This, of course, suggests the use of radio-frequency transformers (ordinary i.f. transformers) as input and output devices as well as spreading coils for f_r and f_a . It might be well to mention at this point



that when f_r and f_a are spread, f_r remains fixed in frequency and only f_a is moved higher.

Let us briefly consider what happens inside the lattice filter. Assume that the pair of crystals connected in shunt (\times connected) are of identical frequency and are about 2 or 3 kc. higher in frequency than the pair of identical crystals connected in series (horizontally connected). Also assume that the coils used have spread the f_r and f_a of each crystal. Any over-spreading can be corrected by the i.f. transformer tuning capacitors, provided the crystals are exactly paired. (See later section on filter alignment.) A of Fig. 4 shows the reactance plot for both sets of crystals, the shunt pair being represented by the dashed curve. Careful alignment is necessary to make the series-resonant frequency of the series crystals (solid curve) correspond to the parallel-resonant frequency of the shunt-connected crystals (dashed curve) and vice versa. The attenuation curve, B in Fig. 4, shows the resulting bandpass characteristic. We have points of very high attenu-

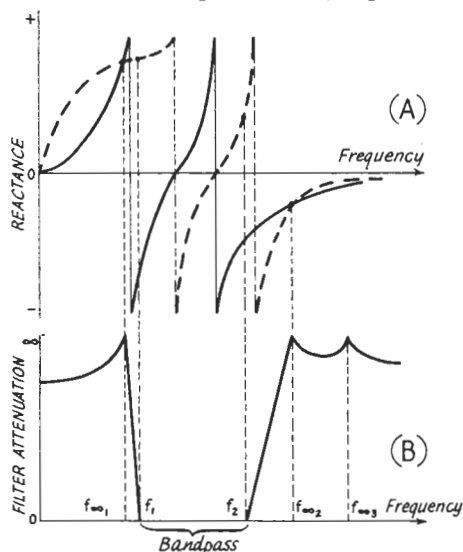


FIG. 4—The reactance-vs.-frequency characteristic for the two pairs of crystals in a lattice filter section is shown at (A). The resultant attenuation characteristic is shown in (B).

ation ($f_{\infty 1}$, $f_{\infty 2}$, and $f_{\infty 3}$) whereas the reactance values are equal and have the same sign (either + or -). We have a bandpass for those frequencies where the reactances of the two filter arms are opposite in sign.

Practical Filters

A workable filter can be constructed at the cost of only a very few dollars. The FT-241-A low-frequency surplus crystals were tried with very good success. Very inexpensive coupling devices were used—ordinary replacement i.f. transformers (Meissner No. 16-5712). There is one sacrifice made that was probably due either

to an improper choice of transformers or an impedance mismatch between crystals and transformers. This was an insertion loss of approximately 12 to 15 db. in the middle of the passband. However, the authors felt that this did not handicap the system too greatly, as this was less than could be gained in a single stage of ordinary i.f. amplification.

Now for the choice of crystals for a given bandpass. For a 5- or 6-kc. bandpass the crystals should be chosen from the FT-241-A series with the two groups of four crystals being separated in channel designation number by two channels; for example, four crystals on Channel 40 and four on Channel 42. For a bandpass of 2.5 or 3 kc., the channel numbers should be consecutive; that is, Channels 40 and 41, for example. Each pair of these crystals for each filter section must be carefully matched so that they are on the same frequency or as close to the same frequency as possible. The pairs should be *within ten or twenty cycles*. If you have several crystals available, a careful selection might be made to match crystals. A signal generator and a vacuum tube voltmeter can be used to do this. Connect the crystal in series with the "hot" lead of the signal generator and the probe of the v.t.v.m. Now sweep the signal generator slowly through the frequency of the crystal, and you will discover that there will be a small indication for any randomly chosen frequency. As the generator frequency is increased through the crystal's fundamental frequency, the v.t.v.m. indication will increase sharply to a very high value and then will break sharply to a very low value, perhaps unreadable on the instrument. The high indication was the series-resonant frequency, f_r , and the null was the parallel or antiresonant frequency, f_a . With a lot of patience and a little cussing, it will be possible to match pairs of crystals using this method. Edge grinding of the lower one of a pair of crystals will fix this matching problem. But be careful—only one or two very light swipes on the fine-grain side of a new flat Carborundum stone. And take heart, because it sounds worse than it actually is. What happens if these crystals are not closely matched? There will be very narrow attenuation slots in the edges of the passband of the filter. The commercial companies get around this problem by putting two sets of silver plating on a crystal and attaching four terminals, making the one crystal serve as two identical crystals. It's a very nice trick but not too practical for a ham to try, and it wouldn't work with this type of crystal.

Now, assuming that you have eight crystals chosen, four crystals per section, each section requiring two pairs of identical crystals, we will proceed. Mount them as shown in the photograph of the sample filter, or in any convenient manner. The physical layout shown is almost identical to the electrical layout. One word of caution: Capacitive leakage around the filter

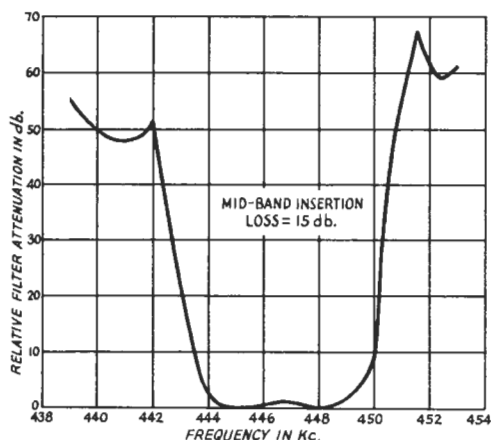


FIG. 5—Attenuation characteristic of an experimental crystal lattice filter (two sections) suitable for receiver use. The crystals were Channels 40 and 42 of the FT-241 series.

sections must be avoided because the high attenuations cannot be realized if there are alternate signal paths other than through the filter elements. Use of shielding is recommended.

Alignment

To align either of the two filters, the following equipment or combinations of equipment will be needed: a BC-221 frequency meter or equivalent calibrated source of r.f. energy covering the range of 400 to 500 kc., and a low-frequency receiver such as the BC-348, BC-453, or a panoramic adapter whose input covers the frequency range we are concerned with. In lieu of the receiver or panoramic adapter, a simple crystal-controlled converter could be built to heterodyne the low-frequency in ques-

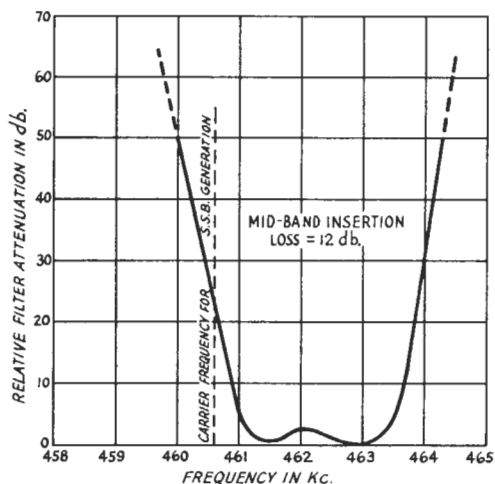
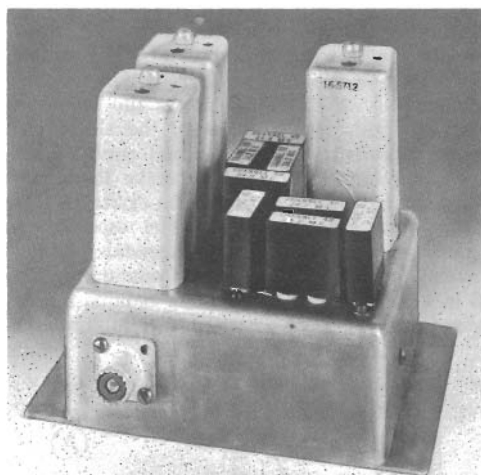


FIG. 6—Attenuation characteristic of an experimental crystal filter (two sections) suitable for a s.s.b. receiver or transmitter. The crystals were Channels 49 and 50 of the FT-241 series.

tion up to a range covered by an existing high-frequency receiver. Use of the receiver and S-meter as a tunable vacuum tube voltmeter indicator is suggested if the approximate "db. per S unit" value is known.

Specific step-by-step adjustments for alignment of these filters will not be given in this article. They would be long and space-consuming and rather pointless in an article of this



An experimental crystal lattice filter for receivers using surplus crystals. Its attenuation characteristic is shown in Fig. 5.

general nature. Instead, a few pointers will be given, and we have faith that the old ham ingenuity will fill in the rest. The first step is to peak the i.f. transformers for the midband frequency of the filter. It may be necessary to align each roughly with the signal source and indicating instrument coupled loosely to each separate transformer in turn in order to get sufficient signal through the whole filter for further alignment. Once this is done, the various sharp peaks and valleys in the passband characteristic must be ironed out to give a smooth shape. If you have been careful in the matching of the crystals, the passband will be fairly well defined. Mismatch of these pairs of crystals will cause the passband to be bumpy and attenuation outside the passband will not be as high as possible. A little cut-and-try is in order here. Place a small "gimmick" capacitor¹ across one of the higher-frequency crystals and run the signal generator through the frequency of the filter again. You will have to judge whether you are doing any good; if not, try another value for the little gimmick capacitor. Usually only one or two pf. will be sufficient to align a typical off-frequency crystal. The trimmer adjustments on the i.f. transformers may be used to equalize the passband characteristic and make it flat.

¹ A gimmick capacitor is a low-capacity affair made by twisting two No. 22 enam. wires together for an inch or so. The capacity is reduced by cutting the wires.

» Here's a two-in-one special—enough information on h.f. crystal filters to design one for most any frequency, bandwidth and shape factor, and also a ready-to-build 5.5-Mc. filter for s.s.b. transceiving.

High-Frequency Crystal Filters for S S B

D. J. HEALEY, W3HEC

Many articles have appeared in *QST* describing crystal filters for s.s.b. operation.¹⁻⁵ However, none of these supplied a design procedure and also gave the precise performance of the resulting filters. This article describes a particular type of filter that was built for a homemade transceiver.

The theoretical shape of the selectivity characteristic attainable with simple crystal filter arrangements was calculated first and found to be inadequate for good sideband suppression. The effect of mismatch when filter sections are cascaded without vacuum-tube isolation improved the steepness of the selectivity characteristic but at the expense of ripple in the pass band. By inserting a small resistance between two sections of a three-section filter, the ripple was reduced without greatly affecting the shape factor (ratio of the bandwidth at some high attenuation to the bandwidth at low attenuation) of the selectivity curve. A filter constructed according to this design from FT-243 surplus crystals performed as predicted.

In filters such as the one used in the transceiver described by W3TLN⁶ it is not unusual to obtain spurious responses as close as 15 kc. to the pass band which are suppressed by only about 20 db. In the filter described in this article the spurious responses are attenuated more than 50 db. even with a crystal whose principal spurious frequency was only 7 db. down from the main response.

Simple Filter Sections

Fig. 1 shows the equivalent circuit of a crystal neglecting its spurious modes. This circuit has the reactance vs. frequency curve shown in Fig. 2. L and C are series resonant at f_r and f_a is the antiresonant frequency of C_0 and the LC combination. By utilizing crystals in a lattice structure as shown in Fig. 3, a

selective filter is obtained. The lattice is a bridge, and it is obvious that maximum un-

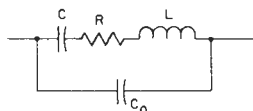


FIG. 1—Equivalent circuit of a crystal. C and L are the motional capacitance and inductance of the crystal, and R represents the frictional loss. C_0 is the electrode and holder capacitance shunting the crystal.

balance of the bridge will occur when one arm has an impedance which is capacitive while the other arm is inductive. When the impedances are equal, the bridge will be balanced. The reactance vs. frequency curves of the crystals can then be used to indicate the regions of the pass band and the stop band. Fig. 4 shows what happens when the antiresonant frequency of one pair of crystals is made equal to the resonant frequency of the other pair. It is observed that the pass band of a simple lattice is limited to the region between the antiresonant frequency of the higher-frequency crystals and the resonant frequency of the lower-frequency crystals. For the case where the reactance curves in the stop band are equal only at zero frequency and infinite frequency, analysis of the circuit shows that the frequency difference $f_2 - f_1$ corresponds to the bandwidth at which the attenuation is approximately 7 db.

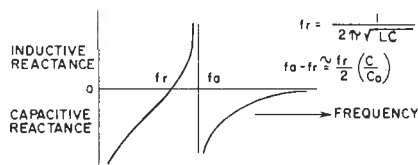


FIG. 2—Reactance vs. frequency characteristics of a crystal. The series-resonant frequency, f_r , is that of C and L . The anti-(parallel) resonant frequency, f_a , is that of the circuit formed by C and L in one branch and C_0 in the other.

The resistive component of the crystal may be transformed to an equivalent parallel resist-

From October, 1960, *QST*.

¹ Weaver and Brown, "Crystal Lattice Filters," page 38.

² Good, "A Crystal Filter for Phone Reception," *QST*, October, 1951.

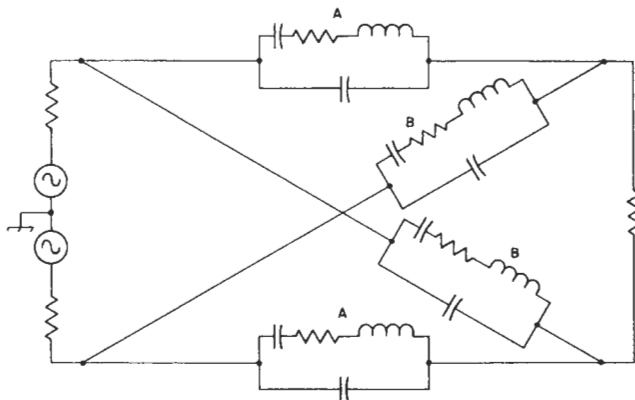
³ Burns, "Sideband Filters Using Crystals," *QST*, November, 1954.

⁴ Morrison, "Cascaded Half-Lattice Crystal Filters for Phone and C.W. Reception," *QST*, May, 1954.

⁵ Vester, "Surplus-Crystal High-Frequency Filters," page 48.

⁶ Vester, "Mobile S.S.B. Transceiver," *QST*, June, 1959.

FIG. 3—Equivalent circuit of a full lattice crystal filter. The series crystals, A, are the same frequency, as are the shunt crystals, B. Note that the lattice could be redrawn as a bridge circuit.



ance shunting the crystal which is essentially constant for the small frequency range of concern. If all crystals have the same Q and inductance, then the equivalent resistance shunting the crystals is the same, and if one considers the resistive bridge thus formed, it is balanced under these conditions. The loss resistance can then be neglected except as it modifies the termination of the filter and affects insertion loss. The point is that the filter behaves essentially as though its elements had infinite Q .

Identical results can be obtained with half as many crystals in a half-lattice circuit. As shown in Fig. 5, this is the equivalent of a full lattice in which the impedances of the elements are one half those of the half lattice. The basic circuit, Fig. 5B, shows two crystals and an ideal transformer having unity coupling. A practical transformer which does not have unity coupling can be represented by the circuit of Fig. 5C. Leakage reactance appears in series with the crystals of the lattice and will lower their resonant frequencies. In narrow-band filters, this can be prevented by connecting a capacitor which resonates with the leakage inductance at the center frequency of the filter in series with the center tap⁷ as in Fig. 5D. This allows a simple center-tapped coil to be used for the ideal transformer.

Designing An S.S.B. Filter

In building a filter for a transceiver, an intermediate frequency of 5500 kc. was selected. This choice is satisfactory for 100-db. suppression of spurious signals in the receiver except on 15 meters. There a 5th-order intermodulation product falls in the pass band so that with a simple mixer the spurious response is only attenuated 75 db. However, by using a simple balanced mixer, the desired suppression of 100 db. is realized. The balanced mixer also reduces the preselection requirements on other bands.

Since the pass band is on the order of twice $f_a - f_r$, this frequency difference must be about

1500 c.p.s. for a filter capable of passing the voice frequencies. The second formula in Fig. 2 shows that $f_a - f_r$ depends on the ratio of the capacitance shunting the crystal to the motional capacitance of the crystal. For AT-cut crystals (the ones you get for \$2.95), this ratio is about 250 minimum. Therefore, AT-cut crystals above 750 kc. meet the requirements, and capacitive terminations are feasible. For example, the shunt capacitance, C_o , of a typical AT-cut crystal is about 3 pf. If it is desired to terminate a half-lattice filter with a circuit capacitance of, say, 15 pf., this will reflect as a total shunt C of $15/2 + 3 = 10.5$ pf. across each crystal. Under such conditions, AT-cut crystals can be used for amateur s.s.b. applications at any frequency above 2625 kc.

Surplus FT-243 crystals (BT-cut) are available for 5500 kc., and their use is economically attractive. These crystals, however, have C_o/C

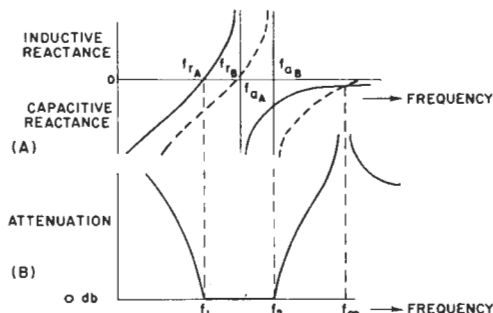


FIG. 4—(A) Reactance characteristics of crystals A and B in the lattice of Fig. 3. In the type of filter described in this article, the series-resonant frequency of the higher-frequency crystal, f_{rB} , is made equal to the antiresonant frequency of the lower-frequency crystal, f_{rA} . (B) Attenuation curve of a filter using the crystals of (A). In the pass band between f_1 and f_2 the series and shunt reactances are opposite, the bridge is unbalanced, and nearly all of the input signal appears at the output. At other frequencies the reactances are similar, and the bridge approaches balance and shows little output.

⁷ Kosowsky, Patent No. 2,913,682.

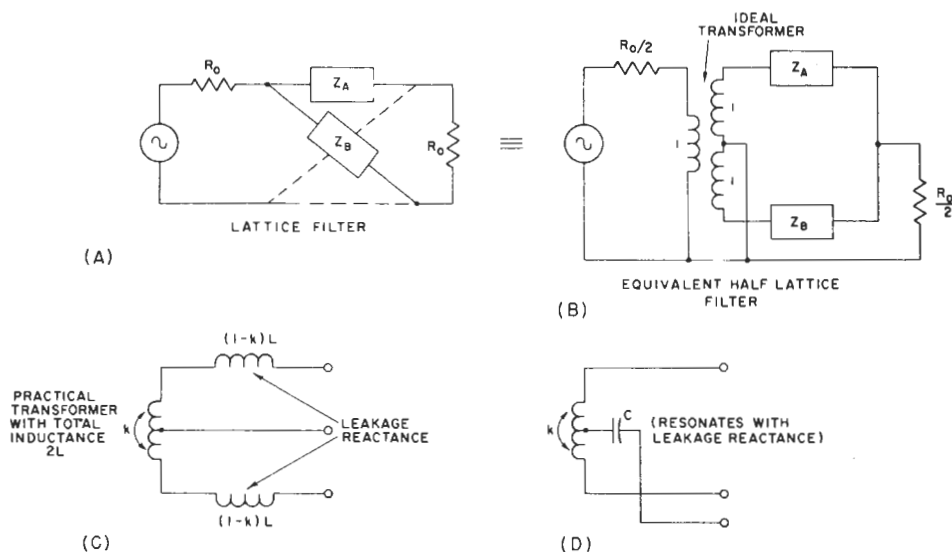


FIG. 5—(A) Basic circuit of a lattice filter and (B) an equivalent half-lattice filter. The transformer is an ideal one with unity coupling between windings. (C) Diagram of a practical transformer showing the leakage inductance which results from a coupling coefficient, k , less than unity. (D) How the leakage reactance may be tuned out by inserting capacitor, C , in the center-tap lead.

ratios around 4000 minimum,⁸ which is too high to allow the desired pass band to be obtained. The pass band can be widened by paralleling inductance with the crystals. This will raise their antiresonant frequencies and leave their resonant frequencies unchanged. One must be cautious in doing this since the inductance will also be antiresonant with the total effective capacitance of the crystal at

⁸ Herman Shall, W3BWK, has pointed out that the high ratios measured on FT-243 surplus crystals result from the nature of their mounting. Actually, BT-cut crystals can exhibit ratios as low as 400. Some of the FT-243s on the surplus market may be AT cut, and may have ratios around 2000. AT-cut crystals which are plated can have ratios as low as 250. If the ratio is other than indicated in the article (around 4000) the termination tuning will be slightly different. However, the range of tuning available with the specified coils should be adequate for any surplus crystals available.—Ed.

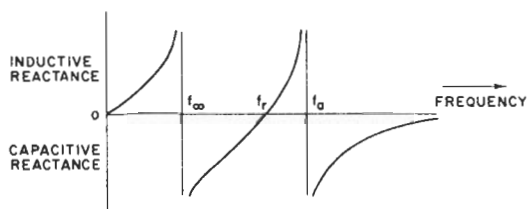


FIG. 6—Reactance vs. frequency characteristics of a crystal and inductance, L_0 , in parallel. f_r , the resonant frequency of C and L , the motional capacitance and inductance of the crystal, is the same as in Fig. 2. f_a , however, is higher than the antiresonant frequency of the crystal alone, since L_0 decreases the effective inductance across shunt capacitance C_0 at frequencies above f_r . L_0 is also antiresonant with the total effective capacitance of the crystal at some lower frequency, f_∞ .

some lower frequency, f_∞ in Fig. 6. If this new frequency of infinite attenuation is too close to the center frequency of the filter, the pass-band characteristics may be distorted.

Measurements made on an FT-243 crystal resonant at 5502.195 kc. showed a motional capacitance, C , of 0.0038 pf. and a shunt capacitance, C_0 , of 14.7 pf. The spacing between the resonant and antiresonant frequen-

cies is therefore $\frac{5502195}{2} \left(\frac{0.0038}{14.7} \right)$ or 711 c.p.s.

To provide a desired spacing of, say, 1422 c.p.s., the effective shunt capacitance must be reduced to one half the value of C_0 . This would require the addition of a *negative* capacitance of $14.7/2 = 7.35$ pf. If an inductance, L_0 , in parallel with capacitance C_0 is considered the susceptance of the parallel combination is given by

$$B = 2\pi f \left(C_0 - \frac{1}{4\pi^2 f^2 L_0} \right) \\ = 2\pi f C_0 \left(1 - \frac{f_\infty^2}{f^2} \right)$$

where f_∞ is the antiresonant frequency of L_0 and C_0 . (The capacitive contribution of the motional capacitance and inductance below their resonant frequency is small enough to be neglected.) For the case above, it is desired to reduce the susceptance of C_0 alone, $2\pi f C_0$, by

a factor of one half. This means that $1 - \frac{f_\infty^2}{f^2}$ must equal $\frac{1}{2}$, so L_0 and C_0 must be antiresonant at 0.707 times the filter frequency, or

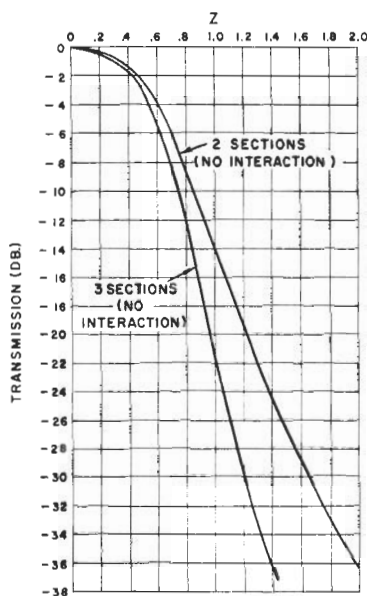


FIG. 7—Theoretical selectivity characteristics of two and three cascaded identical filter sections when interaction due to mismatch between the sections is ignored. Z is the normalized frequency variable defined in the text. It is proportional to the frequency difference from the center of the pass band.

about 3889 kc. for the 5500-kc. filter. Higher values of f_{∞} will also work as long as they are not too close to the center frequency.

For convenience, the filter was built with input and output coils which resonate with 56 pf. plus about 10 pf. tube and circuit capacitance. This is equivalent to about 33 pf. across each crystal, or a total effective C_o of 47.7 pf. Since the effective C_o desired is only

$$7.35 \text{ pf.}, 1 - \frac{f_{\infty}^2}{f^2} = 7.35/47.7 = 0.154, \text{ and } f_{\infty}$$

is 0.92 times the filter frequency, or 5060 kc. This new frequency of infinite attenuation, due to resonance between the coil paralleling the crystal and the total capacitance, is sufficiently removed from the center frequency so that it has negligible effect on the pass-band shape. Design equations for narrow-band capacitor-only filter circuits should therefore be sufficiently accurate for a filter built around FT-243 crystals at 5500 kc.

For narrow-band filters of this type it is convenient to describe the selectivity characteristic in terms of a normalized frequency variable, Z .

$$\text{Let } f = f_o + Z \left(\frac{f_2 - f_1}{2} \right)$$

where f_1 is the resonant frequency of the

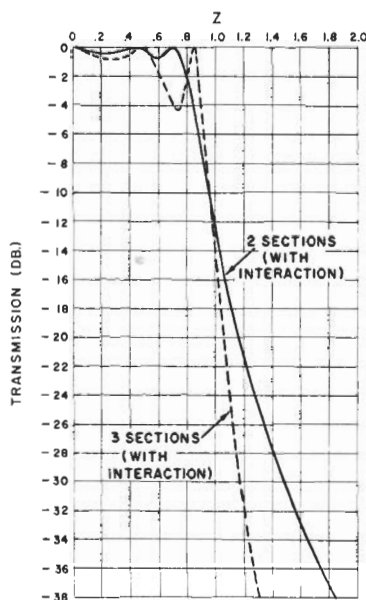


FIG. 8—Theoretical selectivity characteristics of two and three cascaded identical filter sections when interaction is taken into account. These curves are steeper than those of Fig. 7, but there is considerable ripple in the pass band.

lower-frequency crystal, f_2 is the antiresonant frequency of the higher-frequency crystal as modified by the circuit, and $f_o = \frac{1}{2}(f_1 + f_2)$, the center frequency of the filter.

Fig. 7 shows the selectivity characteristics of two- and three-section filters (each section being two crystals in a half lattice) when the effects of mismatch between sections⁹ are neglected. Note that these are normalized curves, plotted in terms of the variable Z .

If we consider only two sections and a 6-db. bandwidth of 2400 c.p.s.; the upper curve of Fig. 7 tells us that $Z = 0.7$ corresponds to 1200 c.p.s. In these plots, Z is proportional to the separation from the center of the pass band. If we assume a low audio cutoff of 300 c.p.s., the carrier must be $1200 + 300 = 1500$ c.p.s.

⁹ The classical method of filter design uses the notion of a characteristic impedance for a filter section. When several filter sections having the same characteristic impedance are cascaded, the over-all selectivity characteristic should be the product of the characteristics of the individual sections (or the sum of their responses in decibels). The difficulty is that the image impedance required to terminate the filter in its characteristic impedance is not realizable with ordinary resistive terminations. As a result, there is reflection at the termination which is a function of frequency, and the filter section does not provide the correct image impedance for an identical section which may precede it. In practical filters a match is obtained only on the average over the frequency range of interest. The input impedance, therefore, varies from the image impedance value, and it is this variation which causes a practical multisection filter to have a response which is different from that which would be expected from the characteristics of the individual sections.

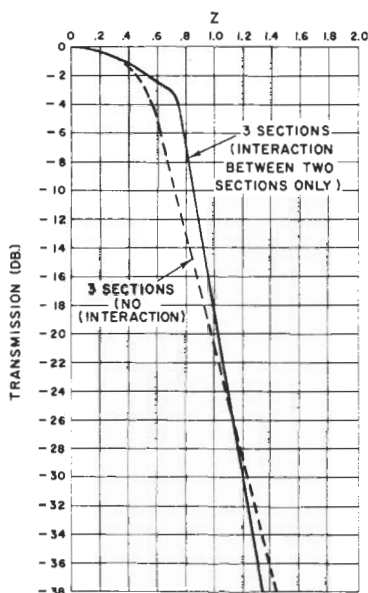


FIG. 9—The solid curve shows the selectivity of a three-section filter with interaction between two sections only. This design is an excellent compromise between those of Figs. 7 and 8. The dashed curve drawn for comparison is for a three-section filter without interaction.

from the center frequency. Therefore, the carrier must lie at

$$Z = \frac{1500}{1200} 0.7 = 0.875.$$

At this frequency the carrier attenuation would only be about 10.5 db. The undesired

sideband would extend from $Z = 0.7 \frac{1800}{1200} =$

1.05 to $Z = \frac{4200}{1200} 0.7 = 2.45$. The corresponding

sideband suppression would vary between 15.4 and 43.2 db.

This is not adequate for our purpose, so three sections must be considered. Here, the 6-db. down frequencies correspond to $Z = 0.617$. Carrier rejection would be 11.6 db. and sideband suppression varies between 17.9 and 58.4 db. This still does not meet our requirement, which is 30- to 40-db. suppression of the undesired sideband with a low audio cutoff of 300 c.p.s., and about 20-db. carrier rejection due to filter selectivity.

Fortunately, the interaction which occurs due to mismatch between cascaded identical filter sections will help us achieve this goal. Fig. 8 shows the selectivity characteristics of two- and three-section filters when effects of mismatch are taken into account. As can be

seen, these curves are much steeper than those of Fig. 7.

Consider as before a 6-db. bandwidth of 2400 c.p.s. For three sections, 1200 c.p.s. corresponds to $Z = 0.91$, and at the carrier frequency

$$Z = 0.91 \times \frac{1500}{1200} = 1.138. \text{ At this value}$$

of Z , the rejection is 26.5 db. At the low audio end of the undesired sideband, $Z = 0.91 \times \frac{1800}{1200} = 1.36$ and the attenuation is 40.5 db.

Thus the three identical crystal filter sections satisfy our requirements.

If we build three sections and reduce the interaction of one section, the attenuation characteristic becomes that shown in Fig. 9. The large pass-band ripple of Fig. 8 is reduced, and by allowing some interaction but not the full interaction of one section, the pass band can be made nearly flat. In Fig. 9, $Z = 0.8$ when the response is 6 db. down. If we again consider a 2400-c.p.s. bandwidth and 300-c.p.s. audio cut-off, $Z = 0.8 \times 1500/1200 = 1.0$ at the carrier frequency; the corresponding attenuation from Fig. 9 is 19 db. The undesired sideband extends from $Z = 1.2$ to $Z = 2.8$, so sideband attenuation will range from 30.3 to greater than 60 db. Such a filter will meet the requirements as well as provide a flat pass band. The shape factor for 30 db./6 db. is 1.49 and for 60 db./6 db., about 2.5.

Since $Z = 0.8$, 1200 c.p.s. from the center frequency, these values can be substituted into the equation which defined Z . Solving for $f_2 - f_1$ gives a value of $2 \times 1200/0.8$, or 3000 c.p.s. The resonant and effective antiresonant frequencies of each crystal should therefore be separated by half this amount or 1500 c.p.s. The resonant frequencies of the two sets of crystals should also differ by 1500 c.p.s. Table I lists the measured characteristics¹⁰ of the crystals actually used. Crystals 1-3 are resonant near 5502.2 kc.; crystals 4-6 are near 5503.9 kc., giving a separation of 1700 c.p.s.

¹⁰ Shunt capacitance C_0 can be found by connecting a small capacitance, C_1 , in series with the crystal and measuring the shift in resonant frequency. The resonant frequency of the crystal alone is

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

If f_{r1} is the resonant frequency of the crystal and C_1 combined, then

$$f_{r1}^2 = f_r^2 \left(1 + \frac{C}{C_1 + C_0} \right).$$

Since $\frac{C}{C_1 + C_0}$ is much less than 1,

$$f_{r1} - f_r \approx \frac{f_r}{2} \left(\frac{C}{C_1 + C_0} \right)$$

or

$$C = 2 \frac{\Delta f}{f_r} (C_1 + C_0)$$

where Δf is the difference ($f_{r1} - f_r$) between the resonant frequencies.

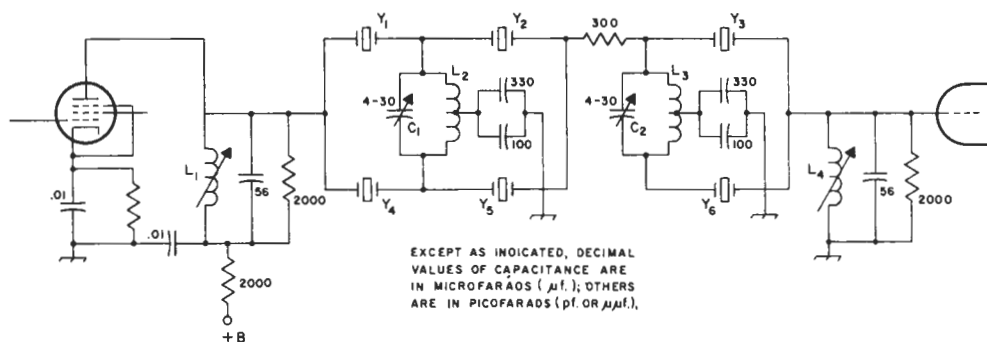


FIG. 10—Circuit diagram of a filter designed according to the methods of this article. Resistances are in ohms, and resistors are $\frac{1}{2}$ -watt composition; capacitors are disk ceramic except as noted.

C_1, C_2 —4–30-pf. mica trimmer.

L_1, L_4 —50 turns No. 38 enamel, close-wound on 17/64-inch diam. ceramic slug-tuned form (CTC LS-6, National XR-81 or similar).

L_2, L_3 —60 turns No. 38 enamel, close-wound on 17/64-inch ceramic form (CTC LS-6, National

XR-81 or similar with powdered-iron core removed), center tapped.

Y_1, Y_2, Y_3 —All same frequency (near 5500 kc.).

Y_4, Y_5, Y_6 —All same frequency and 1500 to 1700 c.p.s. different from Y_1, Y_2, Y_3 .

The impedance of a filter such as this is given by the expression.¹¹

$$R_o = \frac{1}{2\pi C_o'} \left(\frac{f_1}{f_1 f_2 - f_o^2} \right).$$

C_o' is the shunt capacitance of the crystal plus the reflected circuit and tuning capacitance or about 47.7 pf. as shown earlier. f_1 is the resonant frequency of the lower-frequency crystals, 5502.2 kc. f_2 is the antiresonant frequency of the higher-frequency crystals, which is 5503.9 kc. plus about 1500 c.p.s. or 5505.4 kc. f_o is 5060 kc. as calculated above. Putting these values into the equation gives a value of 3920

ohms for R_o . In a half lattice the termination should be $R_o/2$ or 1960 ohms.

2000-ohm terminations are used with the filter that was built. Slight variations in the terminations from these values will affect the selectivity only a small amount and can be used to get almost flat pass-band response.

The effective parallel resistance of the coils is about ten times the filter impedance and has negligible effect on the filter characteristics.¹²

¹¹ A simpler expression can actually be used for the filter under consideration. $R_o = \frac{1}{2\pi f_o C_o''}$ where f_o is

the center frequency of the filter, and C_o'' is the effective capacitance required (7.35 pf. in this case) so that the correct $f_a - f_r$ is exhibited by the crystal.

¹² The input and output coils will have little effect on the impedance of the filter if $2\pi Q f_o L$ (or $Q/2f_o C$) is large compared to $R_o/2$. Q and L are the Q and inductance of the coil, C is the capacitance across the coil, f_o is the center frequency and $R_o/2$ is the terminating resistance. This requirement is usually met in h.f. filters such as the one described. At lower frequencies such as 450 kc., the required $R_o/2$ is higher. Then the terminating resistor must be chosen so that it and the effective resistance of the coil in parallel will give the desired termination.

Table I

Measured characteristics of the crystals used for the filter described in this article.
 f_{rs1} and f_{rs2} are spurious frequencies.

No.	f_r (cps)	C (pf.)	C_o (pf.)	f_{rs1} (cps)	atten. f_{rs1} (db.)	f_{rs2} (cps)	atten. f_{rs2} (db.)
1	5502195	0.00380	14.7	5516800	> 9	5559800	12
2	5502227	0.00356	12.3	5519900	13.5	5552000	15.5
3	5502212	0.00290	12.7	1	> 15 ²	1	> 15 ²
4	5503960	0.00334	12.3	5523000	9	5547200	9
5	5503927	0.00348	14.0	5536200	7	5570200	7.5
6	5503860	0.00311	13.8	1	> 20 ²	1	> 20 ²

¹ These frequencies were not recorded.

² Attenuation greater than figures shown.

Fig. 10 is a diagram of the complete filter. The first two sections are connected "back-to-back," and full interaction takes place. The 300-ohm resistor between sections two and three reduces interaction and smooths the pass-band response as shown above. The leakage reactance between the two halves of L_2 and L_3 is tuned out by the capacitors connected in series with the center taps of these coils. L_1 and L_4 , the input and output coils, resonate with the calculated value of terminating capacitance at 5060 kc. and effectively reflect the needed inductance across the crystals. The 2000-ohm resistors complete the termination called for by the design equations.

All the crystals were purchased as 5500-kc. FT-243s and etched to the desired frequencies with hydrofluoric acid. It is best to wash each crystal with soap and water and measure its frequency before etching. The crystals in each set of three should be as close to each other in frequency as possible, and the separation between the two groups should be about 1500 c.p.s. A simple comparator circuit¹³ will allow two crystals to be checked simultaneously and compared, using an oscilloscope and audio oscillator to measure the frequency separation.

Tuning the filter is quite simple since all four adjustments can be peaked for maximum output at a fixed alignment frequency. This frequency should be on the high side of the pass band and can be the carrier frequency used for lower sideband transmission (5505.5 kc. in the case of the filter described). Using the carrier frequency it is only necessary to unbalance the balanced modulator to obtain a c.w. alignment signal. Of course, a signal generator and r.f. probe-equipped v.t.v.m. can also be used. C_1 , C_2 , L_1 and L_4 are simply adjusted for maximum output.

A slightly better shape factor can be had by detuning the carrier oscillator to a lower alignment frequency corresponding to about the 4-db.-down point on the high-frequency side of the pass band. Fig. 11 shows the measured performance of the filter when aligned at 5505.2 kc. The 6-db. bandwidth is 2750 c.p.s., somewhat greater than the 2400-c.p.s. design figure because the average spacing of the crystal pairs used was 1700 c.p.s. instead of 1500 c.p.s. At 30 db. down, the bandwidth is 3950 kc., so the 30 db./6 db.-shape factor is 1.44. This agrees well with the theoretical value of 1.49 from Fig. 9.

The spurious crystal responses occur as indicated in Table I, but the over-all filter exhibited more than 52-db. attenuation at the nearest spurious frequency (5516.800 kc.). The others could not be measured since they were attenuated more than 60 db., which attenuation

level was beyond the capability of the measuring setup used.

It should be noted that this filter is better used to pass the lower sideband than the upper one. When aligned at the 5505.5-kc. carrier frequency, the filter provides 20 db. of carrier attenuation with a 6-db. down audio pass band which extends from 300 to 2800 c.p.s. The undesired upper sideband is attenuated more than 40 db. for all audio frequencies above 350 c.p.s.

If the filter is aligned at 5505.2 kc. and the carrier set at 5505.6 kc., carrier suppression is 19 db. for a 6-db. audio pass band of 300-3050 c.p.s. Upper sideband suppression is better than 40 db. for audio above 300 c.p.s.

The cutoff on the low-frequency side of the band is somewhat less steep than on the high side. This is believed to be due to the use of less than ideal coupling coils between the filter sections. Using an alignment frequency of 5505.5 kc. and a carrier frequency of 5502.2 kc., the upper sideband audio pass band is 450-2950 c.p.s. for 20-db. carrier suppression. Undesired sideband attenuation is greater than 30 db. for audio above 350 c.p.s. and greater than 40 db. for audio above 750 c.p.s.

When the filter is aligned at 5505.2 kc. and the carrier is placed at 5502.1 kc., the audio pass band is 400-3200 c.p.s. for 20-db. carrier attenuation. Lower-sideband suppression will be more than 30 and 40 db. for audio frequencies above 300 and 700 c.p.s., respectively.

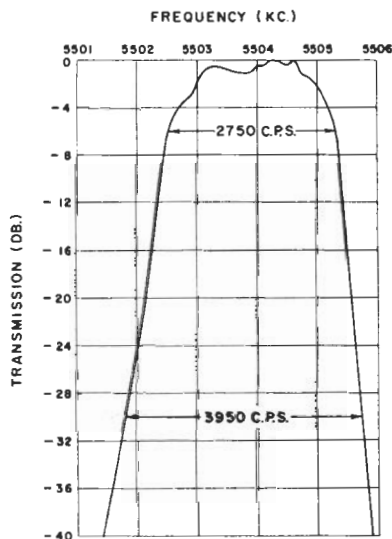


FIG. 11—Measured selectivity characteristic of the filter described in this article when aligned at 5505.2 kc. The 6-db. bandwidth is 2750 c.p.s., and the 30 db./6 db. shape factor is 1.44.

¹³ Clark, "Hints & Kinks," *QST*, December, 1959.

» Using the methods and circuits outlined here, the problem of making a usable high-frequency (i.e., in the 4- to 7-Mc. range) crystal filter doesn't sound too tough, even with limited test equipment. If you've been interested in some of the newer transmitting and receiving techniques using filters in this range, here's a way to give them a whirl without a large investment.

Surplus-Crystal High-Frequency Filters

BENJAMIN H. VESTER, W3TLN

After all the recent QST articles on uses for high-frequency crystal filters, I've really been coveting one for a mobile s.s.b. transceiver I'm planning. The commercial price tags on filters being what they are, I decided it would have to be built from surplus crystals, or not at all. Having, during the earlier days of s.s.b., suffered with a low-frequency crystal filter (typical report was, "Gee, your voice sounds funny"), I decided to do a little reading before dragging out the soldering iron this time.

An article by Kosowsky¹ boils a lot of "long-

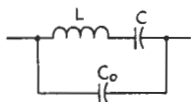


FIG. 1—The equivalent circuit of a crystal. L and C are the electrical equivalents of mechanical constants of the crystal, while C_0 is the shunting capacitance of the electrodes and holder.

hair" literature on crystal-lattice theory into a fairly simple and understandable form. One of the most interesting points to me was the fact that the crystal filter designer considers the narrow-band high-frequency crystal filter for s.s.b. to be the "easy" design—the problem getting much more exotic for the wide-band high-frequency filter. Since my buddy, W3HEC, was already tackling the tough problem of making a good low-frequency filter with the FT-241 crystals, I took the easy way out and tried my hand with the high-frequency unit.

Some Background

If you're planning to try your hand at it, it will help if you grab a few fundamental concepts on crystal lattice filters first. The properties of the crystal itself are pretty well known, the approximate equivalent circuit being shown in Fig. 1 and the change of reactance or impedance being shown in Fig. 2. The crystal has two resonances very close together, L and C being in series resonance at f_z , and L , C and C_0 being parallel resonant at f_p . These reso-

nances have been given names by the network theory boys, the series resonance being called a "zero" of impedance (for obvious reasons) and the parallel resonance being called a "pole" of impedance. The symbols used for these are shown on Fig. 2. These poles and zeros are mighty convenient little symbols for handling networks, the most convenient part being the fact that if you have a circuit with several poles and zeros, you can often manipulate the circuit values so as to get some of the zeros each to cancel out a pole. Hence, a circuit with a multitude of resonances (or poles and zeros) can be arranged to have its response equivalent to only a few resonances.

The universal crystal filter is a lattice circuit. The lattice is usually developed in full "four-arm" form (i.e., as a bridge circuit) and then the equivalence of the half-lattice is proved. The reader is referred to Kosowsky's article and its bibliography for the full treatment on this. We will settle for a few statements on crystal lattice filters which have been adequately proven by others. Consider the simple one-section half lattice shown in Fig. 3. The first important point to consider is that the only way in which the lattice can give a high insertion loss between input and output is for the impedances of A and B to be about equal, so that the voltage at their common connection (point O) is equal to the voltage at the coil center tap. Our crystals will meet the require-

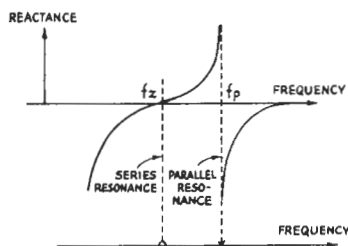


FIG. 2—Reactance characteristics of a crystal. The series-resonant frequency, f_z , is that of L and C (Fig. 1) in series; the parallel-resonant frequency, f_p , is the resonant frequency of the parallel circuit formed by L and C in one branch and C_0 in the other.

¹ From January, 1959, QST.

² Kosowsky, "High Frequency Crystal Filter Design Techniques and Applications," *Proceedings of the IRE*, Feb., 1958.

ment pretty well if they have the same holder capacitance, so the primary problem is to build the coil so that the voltage from Terminals 1 to 2 is exactly the same as the voltage from 3 and 4. The method for realizing this will be discussed a little later.

Crystals *A* and *B* are chosen to be different in frequency for the half lattice. Thus it is obvious that if we are at a zero (series resonant) frequency of, say, crystal *A*, the impedance balance of *A* and *B* is spoiled and there is a voltage showing up between point *O* and the center of the coil. This will also occur at the pole (parallel resonant) frequency of crystal *A*. The same can be said for crystal *B*, only the unbalance is in the opposite direction. This leads us directly into the statement that the pass band of the crystal filter will be as wide as the spacing of all the poles and zeros. This says nothing about the ripple or variation in transmission in the pass band, however, and if *A* and *B* are far apart the ripple or dip may be tremendous. Here's where the network theory boys' trick of pairing off poles and zeros comes in handy. A little study with Fig. 2 of the way in which the impedance change around a zero differs from that around a pole will give an idea how the lattice crystals can be arranged to give a flat pass band. Fig. 4 shows the desired arrangement. The series-resonant frequency of crystal *B* is arranged to coincide with the parallel-resonant frequency of crystal *A*. This will theoretically give a perfectly flat pass band from the zero of crystal *A* to the pole of crystal *B*.

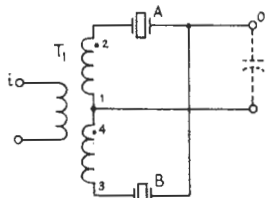


FIG. 3—The half-lattice crystal filter. Crystals *A* and *B* should be chosen so that the parallel-resonant frequency of one is the same as the series-resonant frequency of the other. Very tight coupling between the two halves of the secondary of T_1 , is required for optimum results.

Our problem is now resolved down to determining the pole-zero spacing for the available crystals. The surplus FT-243 crystals in the 5-Mc. range (this choice of frequency was obviously based on the excellent results being obtained with the popular HT-32 transmitter) have a measured spacing of about 2.2 kc. between their series- and parallel-resonant fre-

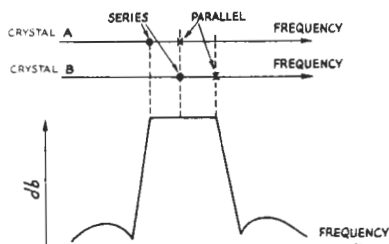


FIG. 4—The theoretical attenuation-vs.-frequency curve of a half-lattice filter shows a flat pass band between the lower series-resonant frequency and higher parallel-resonant frequency of the pair of crystals.

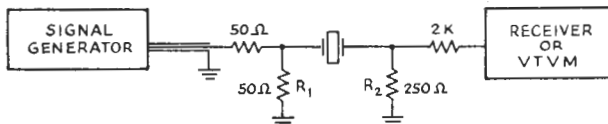
quencies. Thus, two of them spaced 2.2 kc. apart in frequency are theoretically capable of giving a 4.4-kc. bandwidth. Practically, it is very difficult to get quite this much bandwidth.

If we examine the effects that the external coupling circuitry has on the pole-zero spacing, it can be shown that both an increase and a decrease in the spacing can be accomplished, by shunting inductance or capacitance, respectively, across the crystal. The most familiar example of this to most of us is in pulling a crystal oscillator's frequency by shunting a capacitor across the crystal. This technique, you will remember, only works where the crystal is being used in its parallel-resonant mode. Referring back to Fig. 1, it is easily seen that a parallel capacitor makes C_0 larger and lowers the parallel-resonant frequency (pole). It will not affect the series-resonant frequency (zero), so the effect of the parallel capacitor is to move the pole closer to the zero. Similarly, it can be shown that an inductance shunted around the crystal will push the pole away from the zero; unfortunately, however, this also introduces a second parallel-resonant frequency. Even the network theory boys begin to sweat a little when they begin to manipulate this many poles and zeros in a lattice circuit, so we hams had better avoid the complications, and shy away from trying to add tuned inductors on the input and output of the filter. If we are forced to use an inductor, we will make its inductance large enough to avoid its resonating with C_0 anywhere near the desired pass band.

Preliminary Measurements

Now that we have some ideas as to how crystal filters work, we will get more specific and look at the procedure by which one may be evolved. To measure the spacing between the series- and parallel-resonant frequencies,

FIG. 5—Setup for measuring the series- and parallel-resonant frequencies (or pole-zero spacing) of a crystal.



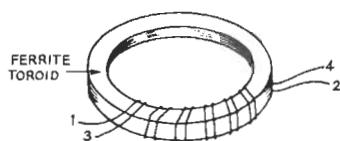


FIG. 6—Bifilar winding on a toroidal core.

we must be careful to avoid having the test circuit put shunt capacitance across the crystal and give erroneous results. The circuit in Fig. 5 was used by the writer. To eliminate the extra shunt capacitance that a socket would add, the crystal holders were soldered directly into the circuit. The signal generator can be most any kind, so long as it has a slow tuning rate—I used one of the Command transmitters. The measurement detector can be a scope, a v.t.m. (with r.f. probe), or the station receiver. The low resistance, R_2 , across it should swamp out any small amount of input capacitance it might have. If a receiver is used, a 1K or 2K resistor should probably be put in series with its input to isolate the crystal from the receiver front-end tuned circuits. The series- and parallel-resonant frequencies are, of course, at the peak and null of the signal across R_2 . Any decent communications receiver will measure the frequency difference; best accuracy is obtained by measuring the harmonics of the generator with the receiver in the sharp crystal-filter position.²

Initial measurements of the two 5645-kc. crystals I had showed a pole-to-zero spacing of 2.2 kc. on one and 2.4 kc. on the other.

²I.e., after adjusting the generator to the series-resonant frequency, let the generator alone and shift the receiver to some higher range where a generator harmonic can be heard and its frequency measured. Then shift back to the fundamental frequency, adjust the generator to the parallel-resonant frequency, shift the receiver again and then measure the generator harmonic adjacent to the first one. The frequency separation between the crystals is of course equal to the frequency difference between the harmonics divided by the order of the harmonic. This method usually will give improved accuracy only if the receiver calibration can be read to the same accuracy—e.g., 1 kc. per dial division—on the harmonic range as on the fundamental. —Editor.

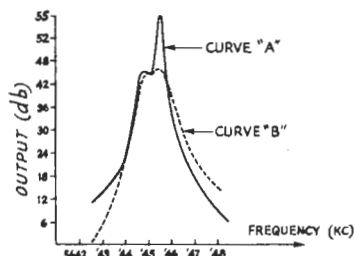


FIG. 7—Measured attenuation curves of a half-lattice filter using two nominal 5645-kc. crystals having series-resonant frequencies separated by 560 cycles. A—without resistance termination; B—with 10,000-ohm terminating resistor. In taking the data for these curves and those shown in Figs. 8, 9 and 11, the attenuation was based on the manufacturer's calibration of the receiver used in the tests.

Their series-resonant frequencies were about 560 cycles apart. I decided to try these out first to get a bearing on the problem.

As indicated earlier, the push-pull coil must have very tight coupling between its two secondaries and should be chosen with a high enough inductance to avoid resonance with the crystal shunt capacitance near the pass band. I used a $\frac{1}{4}$ -inch ferrite toroid (origin and properties unknown) with the secondaries wound bifilar. The bifilar winding is illustrated in Fig. 6. The enclosed LS series coils made by CTC probably would work just as well. (It would probably be very difficult to get tight enough coupling with air-wound coils, however.) I arbitrarily made each half of the secondary coil with an inductance of 50 microhenrys; this required 25 bifilar turns, or 50 turns total. The exact inductance is not at all crucial—the important thing is the tight coupling.

Experimental Results

A filter was constructed with the circuit shown in Fig. 3. It was fed from a low impedance and its output was fed into a 6AK5 mixer grid, the mixer grid effectively shunting

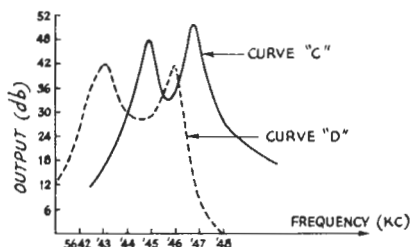


FIG. 8—Attenuation curves of half-lattice filter with crystals of the same nominal frequency as in Fig. 7, but with 1.5-kc. separation. C—with 0.5-megohm terminating resistor; D—shunt coil added across the output to resonate with capacitance present at that point.

some capacitance across the crystals. This mixer was used to beat the filter output signal into a range which was covered by my receiver (a 75A-3) so the receiver could be used for both db. and frequency measurements. The initial response was as shown by curve "A" in Fig. 7. A 10K resistor was then added to terminate the filter and the response squared up

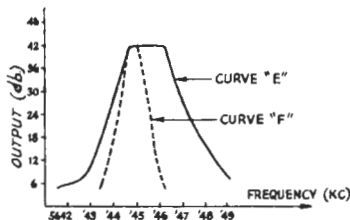


FIG. 9—E—half-lattice filter using same crystals as in Fig. 8, but with 1500-ohm terminating resistor. F—using two nominal 5644-kc. crystals separated 300 cycles, with 3900-ohm terminating resistor.

FIG. 10—Half-lattice filters cascaded in a back-to-back arrangement. The theoretical curve of such a filter has increased skirt selectivity and fewer spurious responses, as compared with a simple half lattice, but the same pass band as the simple circuit.

(as shown by curve "B") to give a passable 1-kc. high-frequency filter.

This was sufficiently encouraging, so I dug out the ammonium bifluoride³ etching bath from its hiding place and moved the upper-frequency crystals to a frequency 1500 cycles above the lower frequency (W2IHW's technique for etching crystals is really simple).

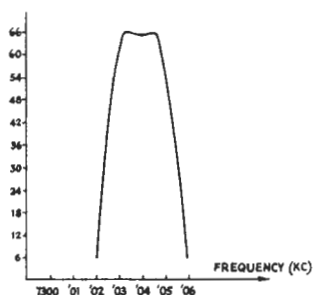
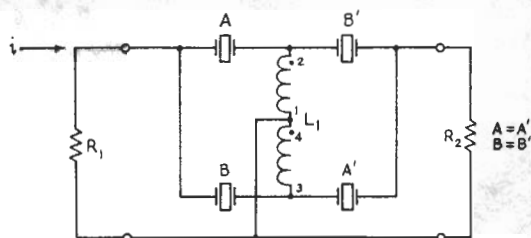


FIG. 11—Attenuation curve of filter using four nominal 7300-kc. crystals, pairs separated 1.5 kc., in the circuit of Fig. 10.

The initial results with this were anything but encouraging. Curve "C" in Fig. 8 illustrates the results. It was obvious that the capacitance across the lattice output had shoved the poles too close to the zeros, or else the 0.5-meg. terminating resistor was improper. I tried tuning the capacitance out with a slug-tuned coil and got all kinds of interesting results (curve "D" in Fig. 8 is typical), none of them usable. When I terminated the filter with lower values of resistance, however, the results improved markedly. With just the right resistor, 1.5K in this case, the pass band was flat over a reasonable width. Curve "E" in Fig. 9 shows the final results. The bandwidth is just barely great enough for phone use.

Since I had one other 5645-kc. crystal which was 300 cycles from one of the original crystals, I substituted it in and got curve "F" in Fig. 9. This time a 3.9K terminating resistor gave the flattest pass band.

If greater rejection off the skirts is required, there are several ways in which these sections can be cascaded. A simple technique is to connect them back-to-back as shown in Fig. 10. This method of connection will minimize spurious off-frequency response since the probability of getting the spurious responses of crystals



A and B to line up with those of crystals A' and B' is pretty small. The coil, L_1 , is again wound bifilar and R_1 and R_2 are chosen experimentally for the best pass band. The crystals should be matched as closely as you can read their frequency—this is pretty easy with the etching technique. Fig 11 shows the response I got from four 7300-kc. crystals, connected like Fig. 10 (crystals A and A' were 1.5 kc. higher than B and B'). The same bifilar coil was used.

I measured the spacing between series and parallel resonance of a few of the other surplus crystals that were lying around and got the following results:-

Crystal Frequency	Type	Pole-Zero Spacing
8725 kc.	FT-243	2.7 kc.
7250 kc.	FT-243	2.3 kc.
7380 kc.	Plated-surplus	5 kc.
7010 kc.	Plated-surplus	6 kc.
8900 kc.	Plated-harmonic cut	20 kc.

The plated crystals will give wider-band filters.

If you're interested in an asymmetrical filter which has a gradual fall-off on one side, then the circuit shown in Fig. 12 can be used. Here both the crystals are on exactly the same frequency. The coils are again bifilar and C is tuned to give the desired pass band. The potential bandwidth here is only half that obtained with the half-lattice. It should work nicely with the plated crystals, however.

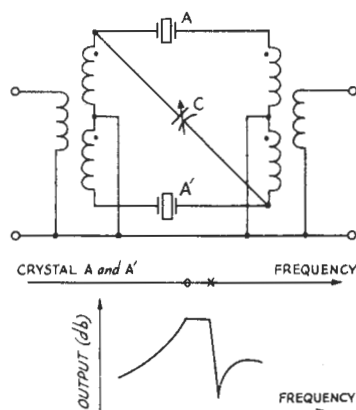
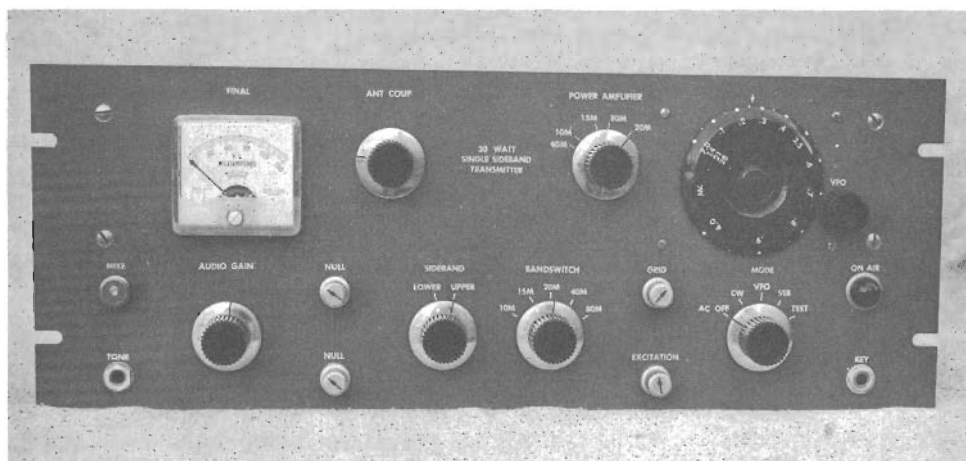


FIG. 12—An asymmetrical filter and theoretical attenuation curve.

³ Newland, "A Safe Method for Etching Crystals," page 52.

» By taking advantage of readily available surplus units, this s.s.b. exciter can be built for less than \$150. It contains all of the conveniences and features found in most advanced units.



A phasing-type s.s.b. exciter. All adjustments can be made from the 7-inch rack panel. Controls along the top, from left to right, are for antenna coupling, the multiband tuner in the final, and the v.f.o. Along the bottom are connectors for microphone input and test-tone input (for alignment purposes), audio gain control, carrier null controls, sideband selector, band switch, excitation and final grid-tuning controls, mode switch and key jack. (Photos by Rogers H. Connell, WFLA-TV News-Photo.)

A Phasing-Type Sidebander

ADELBERT KELLEY, K4EEU

Shortly after being bitten by the sideband bug, the author constructed a single-band phasing-type exciter. Several months later the excellent design by W6TEU¹ appeared, featuring all h.f. bands and voice control—in short, all the desirable features in one package. The project was immediately undertaken to rebuild along the general outlines of the “Sideband Package,” but as a phasing exciter rather than a filter rig.

This design differs from the “Package” in the following ways:

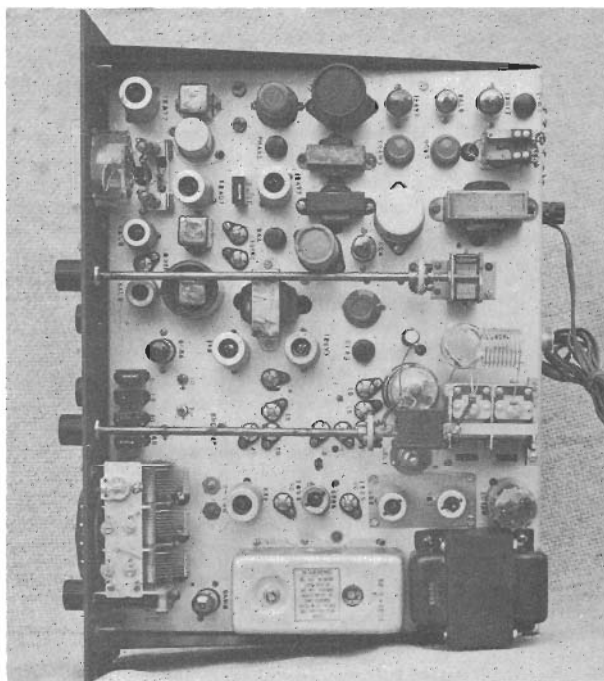
- 1) Sideband generation by the phasing method.
- 2) More elaborate voice-control system, to include receiver control.
- 3) Use of the voice-control system on c.w., to provide break-in.
- 4) Provision for side tone on c.w.
- 5) Modification of the power supply to increase the power output.

¹From November, 1959, *QST*.

¹Bigler, “A Sideband Package,” page 59.

K4EEU's s.s.b. exciter is assembled on a 17×13×3-inch chassis. On the portion toward the panel, the audio section is at the top, converter section for higher-frequency bands at the center, and the v.f.o. at the bottom. Along the rear, from top to bottom are the high-voltage filter choke, antenna-tuning capacitor, the final r.f. stage with its multiband tuner and the power transformer.

- 6) Use of additional surplus material to lower the cost.



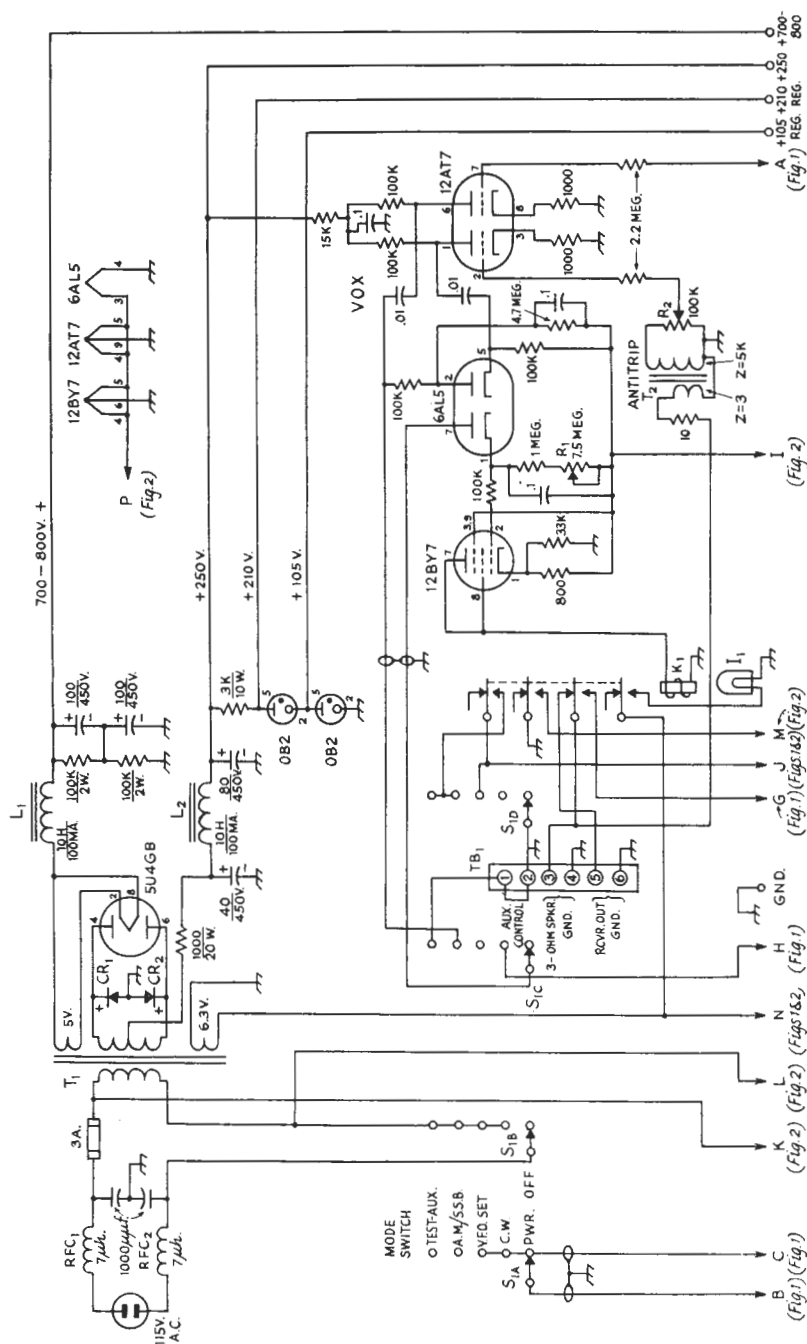


FIG. 3—VOX, power-supply and control circuits of K4EEU's s.s.b. exciter.

Unless otherwise indicated, capacitances are in μf . Capacitors marked with polarity are electrolytic. Other fixed capacitors may be ceramic, paper, or mica. Resistances are in ohms and resistors $\frac{1}{2}$ watt unless indicated otherwise.

CR₁, CR₂—Silicon rectifier, 840 volts r.m.s., 425 ma.
(Sarkes Tarzian 120SM).

I_1 —6.3-volt dial lamp.

K₁—4 p.d.t. 5000-ohm relay (Potter & Brumfield GPD coil and GP17 contact assembly).

L₁, L₂—10.5-hy. 110-ma. filter choke (Stancor-C1001).

R₁, R₂—Potentiometer, carbon.

S₁—4-pole 5-position rotary switch (Centralab PA-1013 or PA-2011).

T₁—Power transformer: 700 v.c.t., 160 ma.; 5 volts, 3 amp.; 6.3 volts, 5 amp. (Triad R-16A*).

T₂—5-watt output transformer, 5000 ohms to 3.2-ohm voice coil.

RFC₁, RFC₂—7- μ h. r.f. choke (Ohmite Z-50).

* Care should be used in substituting for this item if an increase in panel and cabinet height is to be avoided.

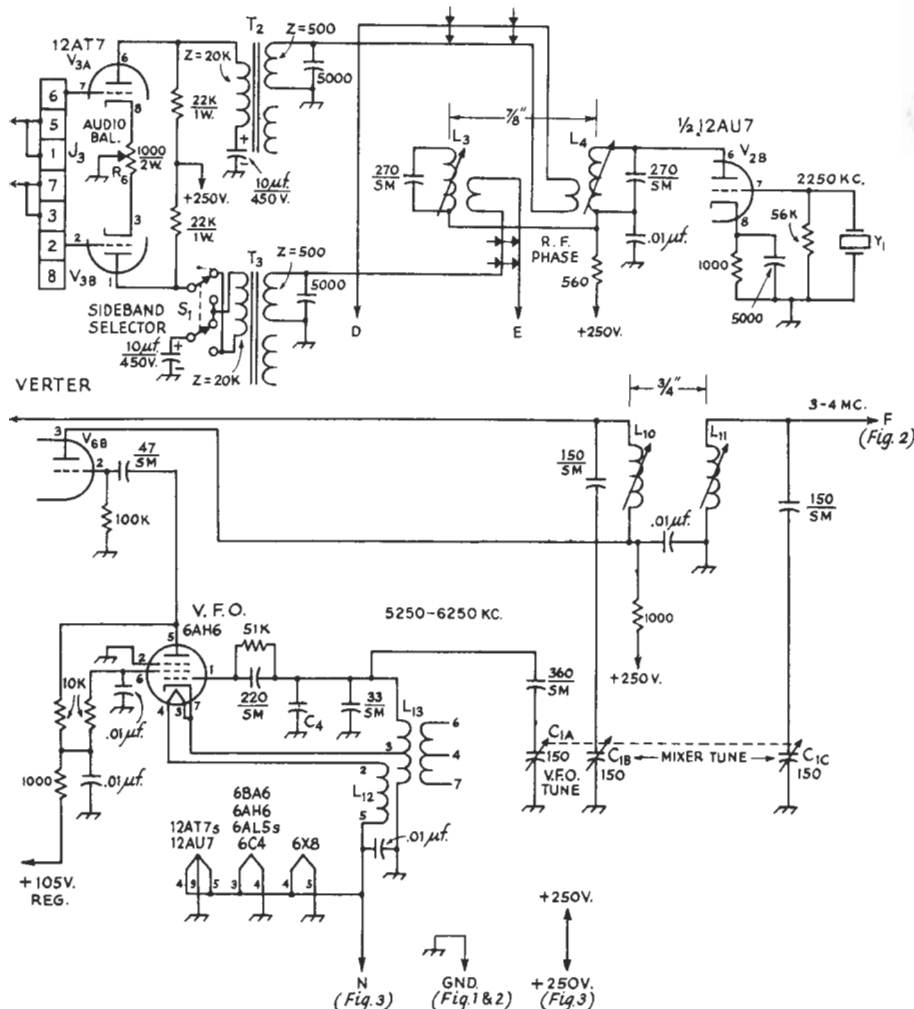


FIG. 1—Sideband-generator, 4-Mc.-output, and c.w.-keying sections of the 30-watt s.s.b. transmitter. Unless otherwise marked, capacitances are in pf. Capacitors with polarity markings are electrolytic. Other fixed capacitors not marked SM may be mica or ceramic. Unless indicated otherwise, resistors are 1/2 watt.

C₁—Three-gang capacitor, see text.

C₂, C₃—Disk ceramic (Centralab DDA104).

C₄—Fixed air padder in BC-458.

J₁—Microphone connector.

J₂—Closed-circuit jack.

J₃—Octal tube socket (for B & W 350 phaseshift net.)

J₄—Open-circuit jack.

K₁—2500-ohm relay, s.p.s.t. contacts (Potter & Brumfield LB-5).

L₁, L₂—26 mh. TV-width coil, (Thordarson WC-19).

L₃,* L₄,* L₈,* L₉*—45 turns No. 28 enam., adjusted to resonate at 2250 kc. (2-turn links on L₃, L₄. Arrows indicate twisted pair).

L₅*—16 turns No. 22 enam. double-spaced, c. t., adjusted to resonate at 2250 kc.

L₆—2 turns over center of L₅.

L7—6 turns at ground end of L8.

L₁₀, L₁₁—40 turns No. 34 enam., close-wound on 1/2-inch diam. iron-slug form (National XR-50).

L₁₂, L₁₃—Oscillator coil unit from BC-458.

R₁, R₂, R₅, R₆—Linear-taper potentiometer.

R₃, R₄, R₇—Audio-taper potentiometer.

RFC₁—2.5-mh. r.f. choke (National R-100S or equiv.)

RFC₂—0.5-mh. r.f. choke (National R-50).

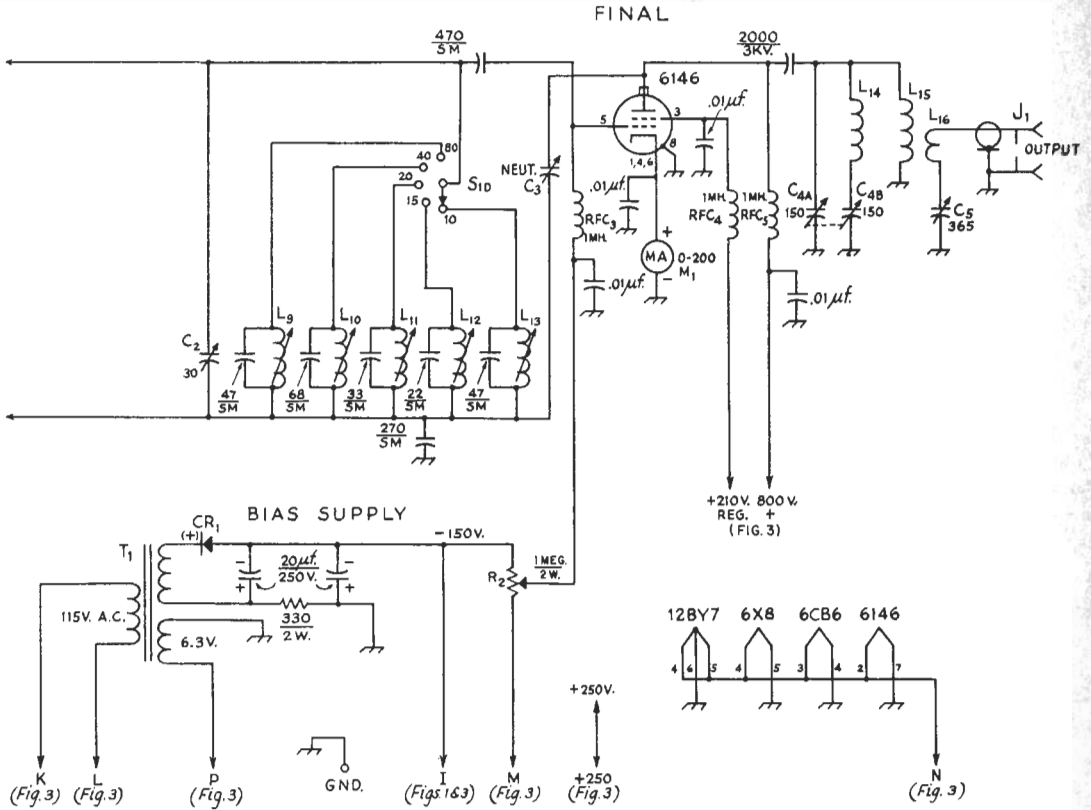
S₁—D.p.d.t. rotary switch (Centrolab 1462).

T₁, T₂, T₃—Miniature 20,000-to-500-ohm transformer (surplus or see QST Ham-Ads). Lafayette AR-151 (20,000 to 800 ohms c.t.) also is suitable; use one-half secondary winding.

T₄—5-watt tube-to-voice coil (3.2 ohms) output transformer (Thordarson 24S50 or equivalent).

Y₁—2250-kc. crystal.

* Wound on $\frac{1}{16}$ -inch diam. iron-slug form from surplus unit. If $\frac{3}{8}$ -inch forms are used, turns should be decreased by about 25 per cent. If $\frac{1}{4}$ -inch forms are used, turns should be increased by about 25 per cent. Half-inch forms will require a reduction of about 50 per cent in the number of turns.



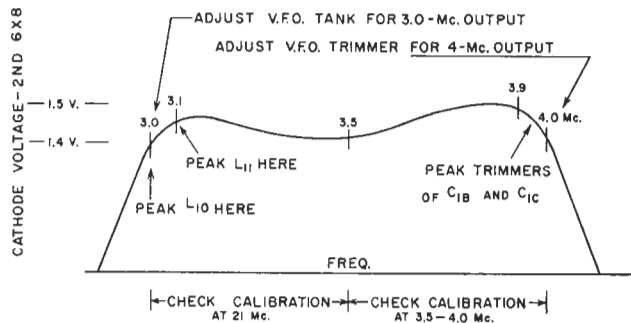
To get the v.f.o. to tune over the correct range it is necessary to insert a 360-pf. silver mica in series with the oscillator tuning capacitor and shunt 33 pf. across the coil. No other changes are necessary. The resulting v.f.o. does not have the linear dial of the W6TEU rig but the bandspread is adequate and it is satisfactory in all other respects.

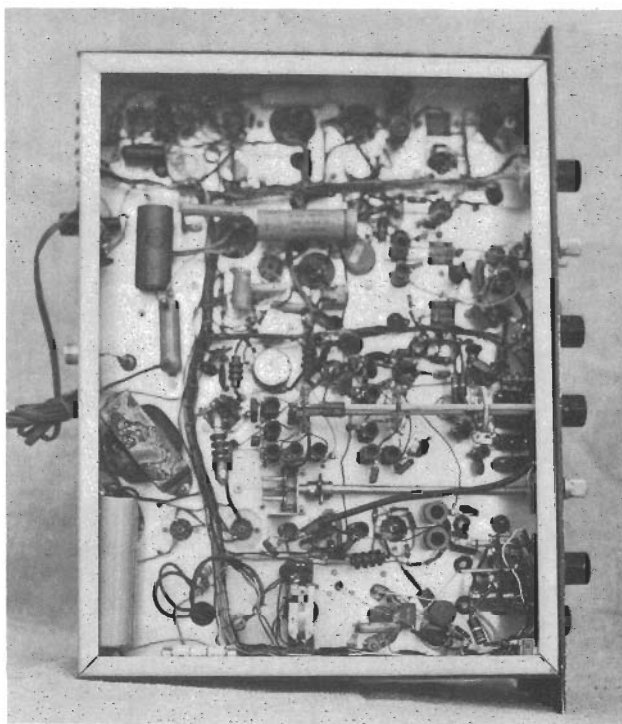
A few construction pointers might be in order. Be sure the v.f.o. shield makes a good all-around contact with the chassis, and provide a ground connection from the frame of the v.f.o. tuning capacitor through a grommited hole to the under side of the chassis. A small baffle shield was placed over the mode switch to eliminate a small change in frequency when the switch was operated. If carefully constructed, this v.f.o. will compare, in stability, with the best of them.

VOX System

The voice-control system uses a multicontact relay to turn the transmitter on, silence the receiver, provide adjustable time delay, and operate an indicator light. The mode switch controls this circuit and on c.w. provides automatic station control. When the key is pressed both relays close, the receiver goes off, the speaker is switched to side tone, and the transmitter final is activated. Subsequent keying holds the first relay shut until a pause in the transmission causes the VOX relay to open, activating the receiver. While it is not true break-in, this system approaches it. An entire QSO can be had without operating send-receive switches. To aid in operating, a simple

FIG. 4—Response curve of the band-pass coupler L₁₀ L₁₁ (Fig. 1) when correctly adjusted. Adjustment is discussed in the text.





Bottom view of the K4EEU s.s.b. exciter.

neon-tube oscillator provides keying side tone.

When the mode switch is turned to v.f.o., the voice-control system is inactive and there is sufficient signal from the transmitter to zero in on a signal in the receiver. To prevent audio feedback, the audio system is shorted out on c.w. or when setting the v.f.o. When the mode switch is turned to the A.M.-s.s.b. position, regular voice-control operation with anti-trip is available. The anti-trip levels are set by fixed pots on top of the chassis and need not be adjusted in day-to-day operation. The TEST position of the mode switch locks the transmitter on (when terminals on the back terminal board are shorted) for testing or to disable the VOX manually.

Power Supply

The power supply is of the "Economy" type and uses two silicon rectifiers to replace the usual tubes and filament transformers. These rectifiers are real space savers and their price is favorable when compared with the cost of the parts they replace. By using two connections to the bridge system, about 700 volts is available for the 6146 plate and 250 volts for the rest of the chassis without using a large power-wasting resistor. The high-voltage filter capacitors are standard cardboard-sleeve electrolytics and are the answer to a large capacitance in small space.

Exciter Coils

To allow better separation between the

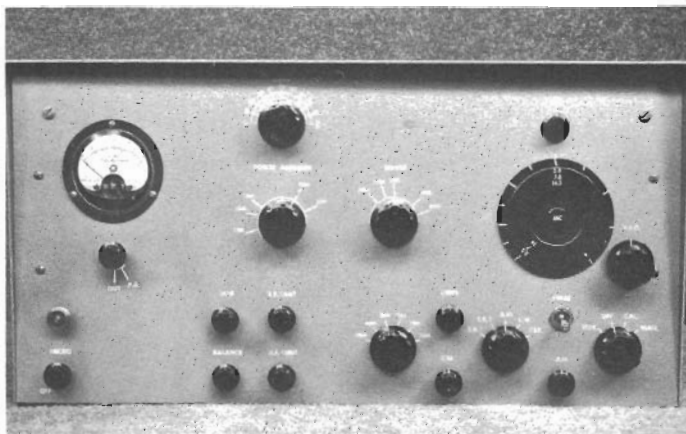
various coils in the bandswitch assembly, a shaft extension is used on the Centralab PA-302 6-inch shaft to lengthen it another inch. Surplus slug-tuned coils were used here, and all were adjusted to frequency with a grid-dip meter. There are eighteen of them used in this transmitter! It is helpful, but not essential, if they all are of the same kind. Other coils that will work are Miller No. 4400, CTC, North Hills, and even those removed from old TV sets. The use of slug-tuned forms helps in the final adjustment of the transmitter and, of course, eliminates the need for trimming capacitors.

Fig. 4 shows the response curve of the band-pass coupler (L_{10} , L_{11} , Fig. 1) when correctly adjusted, and with the band switch in the 80-meter position. The curve is formed by taking readings with a v.t.v.m. across the cathode of the second 6X8 mixer as the output frequency of the first converter is varied. Frequency points for the low-frequency half of the range can be checked by listening for the 21-Mc. harmonics on the receiver. The slugs of L_{10} and L_{11} are peaked for the two frequencies indicated at the low-frequency end of the range.

The author has built two transmitters of the above design and is convinced that it is bug-free if the illustrated chassis layout is used and good wiring practices are followed.

"The power output and multiband operation of this transmitter put it in the "most-desired" classification for the home constructor in search of circuit information. As added attractions, it includes provision for c.w., a.m., and f.s.k."

Panel layout. The microphone connector and gain control are to the extreme left. The meter switch is below the meter. Grouped to the left of lower center are controls for VOX sensitivity and r.f. limiter (above), mixer balance and a.f. limiter (below). At the lower right are the bandswitch, excitation and v.f.o. calibrate controls, mode switch, final-amplifier power switch and a.m. drive control, and the function switch (S1). On the upper portion of the panel, near center, are the loading control (C25) above, and controls for the



final multiband tuner and driver tuning. The small knob above the v.f.o. is the carrier-balance control.

Filter-Type 100-Watt-Output Sidebander

JOHN ISAACS, W6PZV

The hobby of amateur radio is many things to many people. The author is one of those who derive enjoyment from the construction of equipment. This includes new equipment plus the modification (improvement?) of existing commercial and surplus equipment. If one places a monetary value on his spare time, then it is not difficult to prove that the purchase of commercial gear will "pay off" in the end. The advocates of this philosophy are obviously in the majority and the author has no wish to convert anybody. The information presented here, it is hoped, will be of some interest to those who still like to "roll their own."

The design and construction of a multiband exciter requires a lot of time. There are bound to be mistakes. It is best to make as many of these as possible on paper before the first hole is cut. After all, you aren't going to construct several prototypes before making the final unit. A good approach is to benefit by the experience of others. An idea here, another there. Everything helps.

A set of objectives is always necessary for any worthwhile project. The author had these in mind for his new exciter:

1) Multiband operation with no plug-in coils.

- 2) Provision for c.w., a.m., s.s.b. and f.s.k.
- 3) Voice control and antitrip on s.s.b.
- 4) Built-in stable v.f.o.
- 5) About 100 watts peak output.
- 6) Some provision for r.f. or a.f. limiting.
- 7) Good carrier and sideband suppression on s.s.b.

The author's previous experience with s.s.b. had been limited to the phasing type of exciter. Results were not always satisfactory because of a continuous need for adjustment to maintain reasonable carrier suppression and a low order of sideband suppression. There are several successful commercial designs which employ the phasing method and many staunch advocates of same. Just for a change then, if for no other good reason, it was decided that the new exciter would employ the filter method. The new McCoy 9-Mc. crystal filter looked particularly promising. Also, the relatively new circuits using the 7360 tube appeared to offer advantages. A search of the literature revealed numerous good designs, including those found in some well-known commercial units. A design by W6TEU¹ and an adaptation by K4EEU,² looked especially interesting. Although the basic signal-generating

¹ Bigler, "A Sideband Package," page 59.

² Kelley, "A Phasing-Type Sidebander," page 80.

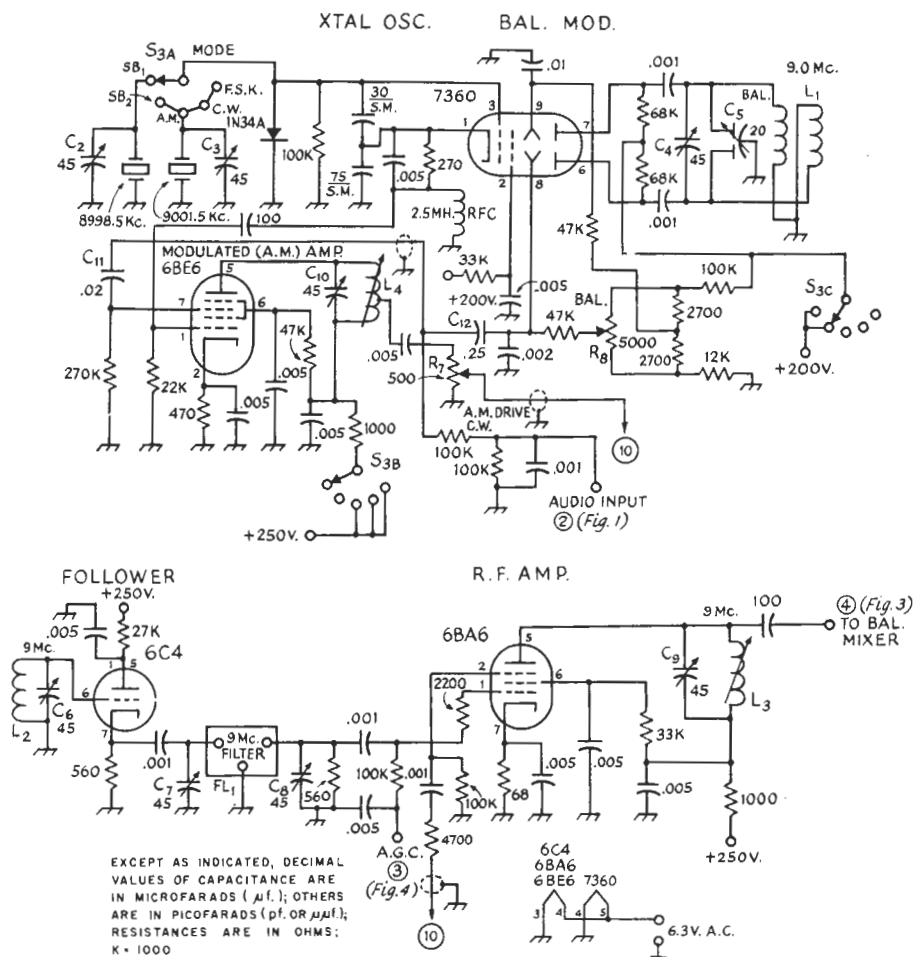


FIG. 2—Sideband and a.m. generator circuits. Resistances are in ohms and fixed resistors used by the author are rated (conservatively in most cases) at 1 watt unless indicated otherwise. Fixed capacitors of less than 0.001 μ f. are mica or silver mica (SM); others are disk ceramic, except as listed below.

C₂, C₃, C₄, C₆, C₇, C₈, C₉, C₁₀—7—45-pf. ceramic trimmer (Centralab 822-BN or equivalent).

C₅—Differential capacitor (Johnsan 19MA11/160-311).

C₁₁, C₁₂—Paper.

FL₁—Sideband filter (McCoy Electronics 32 B1).

L₁—32 turns No. 26 enam., bifilar-wound on $\frac{3}{8}$ -inch polystyrene rod.

L₂—38 turns No. 26 enam., close-wound on $\frac{3}{8}$ -inch polystyrene rod. Form is placed parallel to

form of L₁, forms spaced $\frac{3}{4}$ inch center to center.

L₃—30 turns No. 26 enam., close-wound on $\frac{3}{8}$ -inch ceramic iron-slug form (Miller 4400 form).

L₄—Same as L₃, tap at 7 turns from ground end.

R₇—Carbon control, linear taper.

R₈—Linear-taper control.

S₃—3-pole 5-position ceramic rotary switch (CRL P-272 index, 2 type RRD wafers, one pole of rear wafer [crystal switch] not used).

negative d.c. is applied to another grid of the 6BN6. This voltage acts to prevent the operation of the VOX relay on signals from the receiver speaker.

The s.s.b. signal is generated at 9 Mc. in a 7360 (see Fig. 2). This tube performs the functions of a crystal oscillator and a balanced modulator. Actually, two crystals are used. These are supplied with the McCoy 32B1 s.s.b. filter. They are at 8998.5 kc. and 9001.5 kc.

The passband of the filter is centered on 9000 kc. and is symmetrical. Sideband selection is made by connecting one or the other of the crystals into the circuit. The filter cuts off the unwanted sideband and also provides about 10 db. of carrier suppression.

At 9 Mc. it is necessary to provide the 7360 with both a resistance and a capacitance balance. Also, the plate coil is bifilar wound. After the initial adjustment of the capacitors,

DRIVER

2ND MIXER

BAL. MIXER

V.F.O.

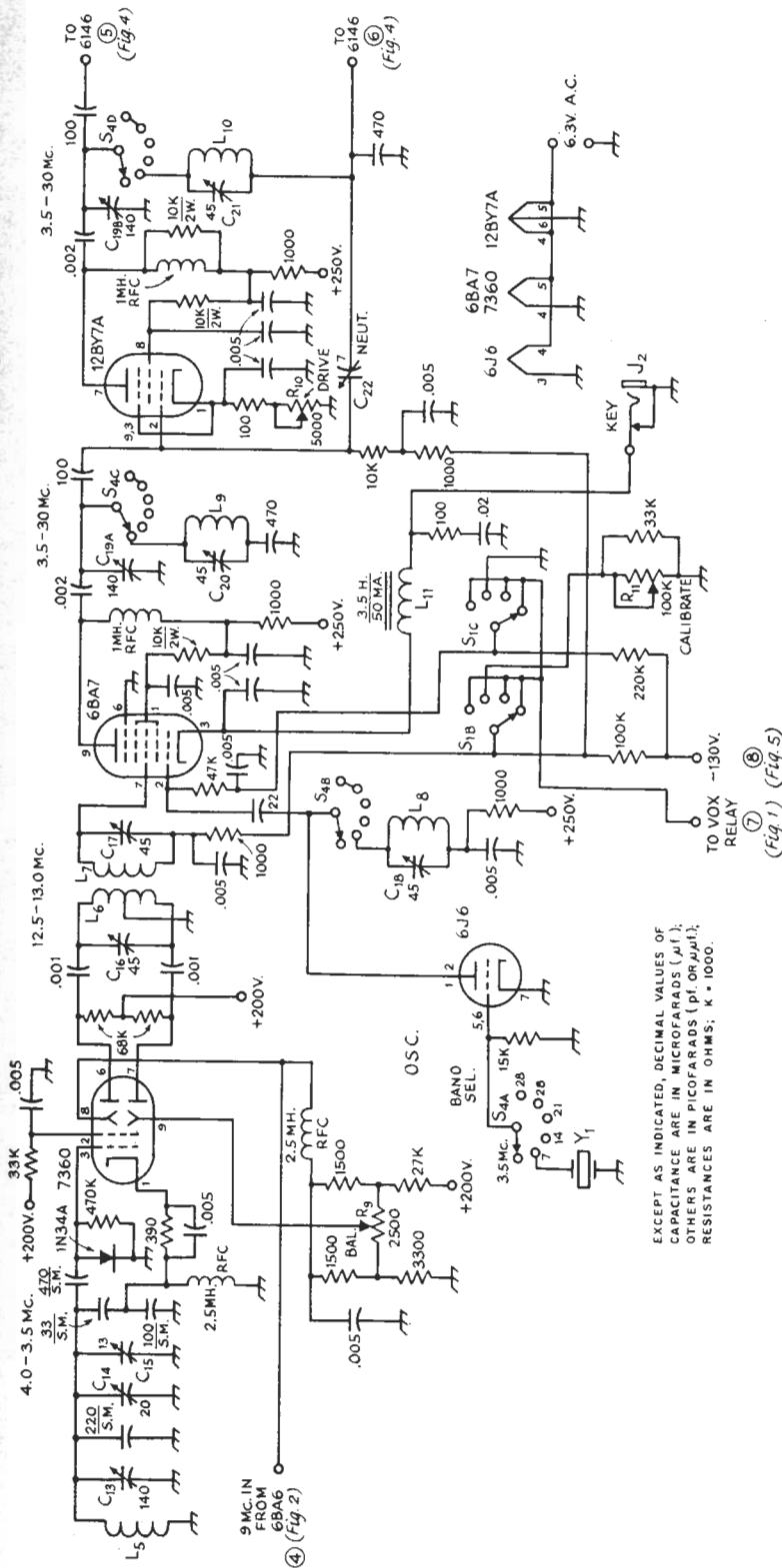


FIG. 3—V.f.o., band-heterodyning and driver circuits. Resistances are in ohms, and fixed resistors used by the author are rated (conservatively in most cases) at 1 watt. Fixed capacitors of less than 0.001 μf . are mica or silver mica (SM); others are disk ceramic. Triode sections of 6J6 are connected in parallel.

- C13—Midget two-bearing 140-pf. air variable.
 C14—N300 5–20-pf. ceramic trimmer (Erie TS-D).
 C15—NPO 2.5–13-pf. ceramic trimmer (CRL 822-BZ).
 C16, C17, C18, C20, C21—7–45-pf. ceramic trimmer (CRL 822-BN). In bandswitching circuits, a similar capacitor is connected across each coil.
 C19—Dual 140-pf. air variable.
 C22—NPO 1.5–7-pf. ceramic trimmer (CRL 822-EZ).
 J2—Closed-circuit jack.
- L5—26 turns No. 22 enam., close-wound on $\frac{1}{2}$ -inch low-loss bakelite form (National XR-50 form with slug removed).
 L6, L7—26 turns No. 26 enam., close-wound on $\frac{3}{8}$ -inch polystyrene rod, forms mounted as described for L12.
 L8, L9, L10—See coil table.
 L11—Filter choke (Stancor C-1080).
 R9, R10, R11—Linear control.
- S1—See Fig. 1.
 S2—4-pole 6-position ceramic rotary switch (CRL P-272 index head, 4 type XD wafers).
 Y1—16.5 Mc. for 2.5-Mc. output.
 Y2—20.0 Mc. for 7-Mc. output.
 Y3—27.0 Mc. for 14-Mc. output.
 Y4—34.0 Mc. for 21-Mc. output.
 Y5—41.0 Mc. for 28–28.5-Mc. output.
 Y6—41.5 Mc. for 28.5–29-Mc. output.

Chart of Harmonic Frequencies

Osc. Freq. (Mc.)	2.0	2.1	2.2	2.3	2.4	2.5	2.6	2.7	2.8	2.9	3.0	3.1	3.2	3.3	3.4	3.5	3.6	3.7	3.8	3.9	4.0	4.1	4.2	4.3	4.4
I.F. (Osc. + 9 Mc.)	11.0	11.1	11.2	11.3	11.4	11.5	11.6	11.7	11.8	11.9	12.0	12.1	12.2	12.3	12.4	12.5	12.6	12.7	12.8	12.9	13.0	13.1	13.2	13.3	13.4
3rd Har.	6.0	6.3	6.6	6.9	7.2	7.5	7.8	8.1	8.4	8.7	9.0	9.3	9.6	9.9	10.2	10.5	10.8	11.1	11.4	11.7	12.0	12.3	12.6	12.9	13.2
4th Har.	8.0	8.4	8.8	9.2	9.6	10.0	10.4	10.8	11.2	11.6	12.0	12.4	12.8	13.2	13.6	14.0	14.4	14.8	15.2	15.6	16.0	16.4	16.8	17.2	17.6
5th Har.	10.0	10.5	11.0	11.5	12.0	12.5	13.0	13.5	14.0	14.5	15.0	15.5	16.0	16.5	17.0	17.5	18.0	18.5	19.0	19.5	20.0	20.5	21.0	21.5	22.0

only an occasional adjustment of the potentiometer is required to maintain the carrier suppression. Following the 7360 is a 6C4 which is used as a cathode follower to provide the necessary match to the crystal filter. The output of the filter is amplified by a 6BA6 to get the signal up to a level for mixing.

Most commercial exciters make some provision for a.m. operation. This is done usually by carrier insertion or by unbalancing the modulator. In either case, the results leave a great deal to be desired. The proper ratio between carrier and sideband(s) is difficult to maintain. Also, a signal consisting of a carrier plus only one sideband produces some distortion in receivers equipped with a diode detector and set for normal a.m. operation.

In this exciter, some of the output of the 9-Mc. crystal oscillator is fed to grid No. 1 of a 6BE6 r.f. amplifier. Audio is fed to grid No. 3. The plate is tuned to 9 Mc. and the output is a standard a.m. signal. Proper adjustment of the cathode resistor and the audio input is necessary to obtain the proper degree of modulation. For a.m. operation, the B+ is removed from the plates of the 7360 and the output of the 6BE6 is fed to the suppressor grid of the 6BA6.

V.F.O. and Balanced Mixer

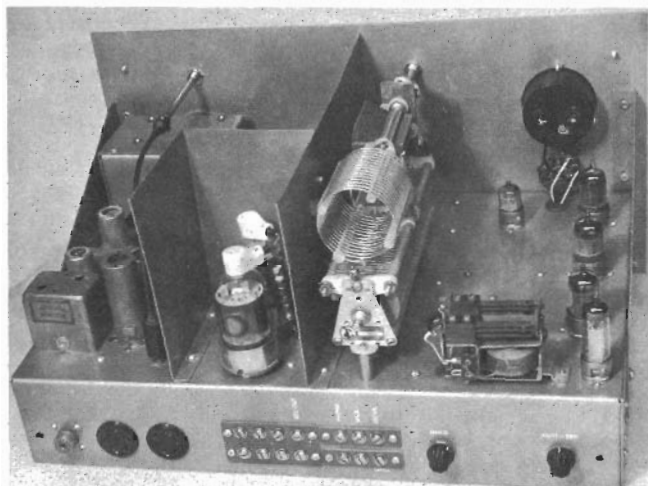
Mixing the 9-Mc. signal with a 5-Mc. signal produces output on either 75 meters or 20 meters, but further conversion is necessary to obtain a signal on the other bands. Because

of this, the idea of a 5-Mc. v.f.o. was dropped. A scheme used in one of the commercial exciters appeared to be the most promising. Mix the low-level signal with the v.f.o. signal and convert to a higher frequency. This frequency should be high enough so that the output circuit can be broad-banded to eliminate the need for tuning the mixer plate along with the v.f.o. A chart, illustrated here, was prepared to determine the best frequency range for the v.f.o., which was to cover a segment 500 kc. wide. Surprisingly enough, the range of 3.5 to 4.0 Mc. appeared to be the best. Just above and below this range, some of the oscillator harmonics fall within the i.f. range. Theoretically, it was not necessary to be quite so careful with oscillator harmonics. As shown in Fig. 3, the first mixer uses a 7360 (which also serves as the v.f.o.). The 9-Mc. signal falls well outside the i.f. passband and no trouble was to be expected from this source. Because of subsequent mixing, the oscillator is on 3.5 Mc. when the exciter output is 4.0 Mc.

Second Mixer and Amplifier

After the 9-Mc. signal is converted to the 12.5- to 13.0-Mc. range, it undergoes one more conversion to get to the desired band. Referring again to Fig. 3, mixing is done in a 6BA7, and a 6J6 crystal oscillator is used. In every case, the crystal oscillator is on the high side, so no trouble is experienced from harmonics of this oscillator. All crystals are third-overtone type, and the crystal oscillator includes a

This rear view shows the multiband tank assembly more clearly than the plan view. It also shows terminals for the power supply and other external connections. The VOX hold and antitrip controls are to the right.



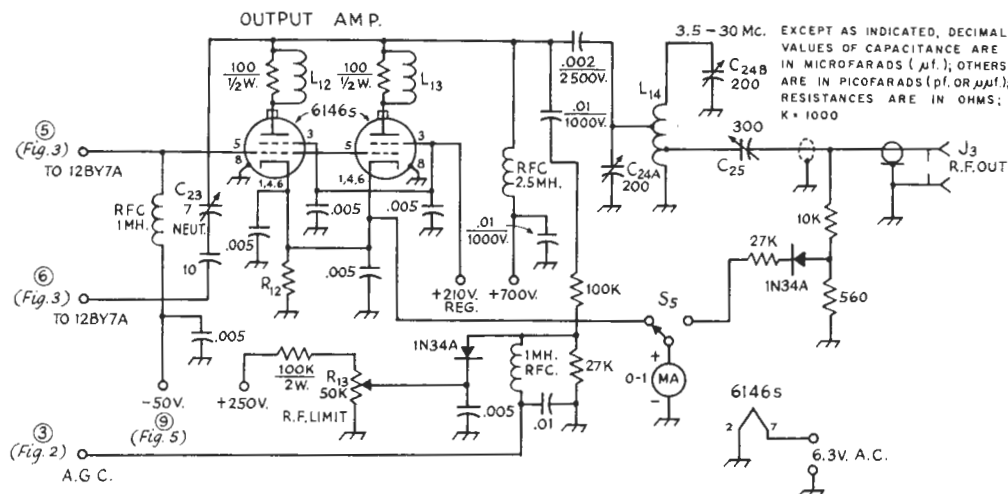


FIG. 4—Circuit of the output amplifiers. Fixed capacitors of less than 0.005 μf . are mica; others are disk ceramic. Resistances are in ohms and resistors are 1 watt unless indicated otherwise.

C23—Same as C22.

C24—Dual 200-pf. 2000-volt variable (Johnson 200FD20/155-505).

C25—Midget air variable (Hammarlund MC-325-M).

J3—Chassis-mounting coax receptacle.

L12, L13—5 turns No. 20 close-wound on associated resistor.

L14—15 turns No. 14, 2-inch diam., 6 turns per inch, tapped at 3 and 8 turns from ground and (Air Dux 1606T stock).

R12—300-times meter shunt.

R13—Linear control.

S5—S.p.d.t. rotary.

tuned plate circuit for each crystal. The output of the 6BA7 is on the desired amateur band and its plate circuit is tuned by a variable capacitor, C19A, adjustable from the front panel.

Following the 6BA7 is a 12BY7 which is tuned by a second section (C19B) of the same variable capacitor used for the plate of the 6BA7. (The 470-pf. capacitor from L9 to ground is a tracking corrector.) The 12BY7 is bridge-neutralized and a 5000-ohm control is connected in the cathode circuit to provide a means of adjusting the excitation to the power amplifier.

Power Amplifier

Two 6146 tubes are connected in parallel and used as the linear power amplifier. See Fig. 4. These tubes are also bridge-neutralized. To further stabilize things, a 10,000-ohm resistor is connected across the plate load of the 12BY7. A portion of the r.f. output is rectified by a crystal diode and fed back to the grid of the 6BA6. A control, R13, is provided so that the crystal will not rectify until some preset level is reached. This operates just like an a.g.c. circuit and minimizes the possibility of over-driving the power amplifiers and any subsequent linear. Another crystal diode is used to rectify a portion of the output so that it can be monitored by a 0-1-ma. meter. A 300-ma. shunt, R12, is provided in the cathode circuit of the 6146s and this is also

connected to the same meter through a selector switch.

The most interesting part of the power amplifier is the tank circuit. It is the multiband type and uses only one coil and a split-stator capacitor. The circuits used by W6TEU and K4EEU employed a similar tank which had two coils. The design of both types of multiband tank circuits is well covered by W6MUR.⁵

Power Supply

A total of six d.c. voltages is required for the operation of the exciter. Referring to Fig. 5, a bridge rectifier is used to provide the high voltage (700 v.). Eight silicon diodes plus one rectifier tube are used. This supply has a choke-input filter with an effective capacitance of 80 μf . The regulated screen voltage for the power amplifier is obtained from the 700-volt supply by using two 0C3 voltage-regulator tubes in series. Two more d.c. supplies are used. Both are directly connected to the 120-volt line. A polarized plug on the line cord takes care of the proper ground connection. One supply uses a single silicon diode and provides -130 volts and -50 volts for the bias circuits. The other supply uses two silicon diodes in a voltage-doubler circuit. It provides +200 volts and +250 volts for all of the exciter except the power amplifier.

⁵ Johnson, "Multiband Tuning Circuits," *QST*, July 1954.

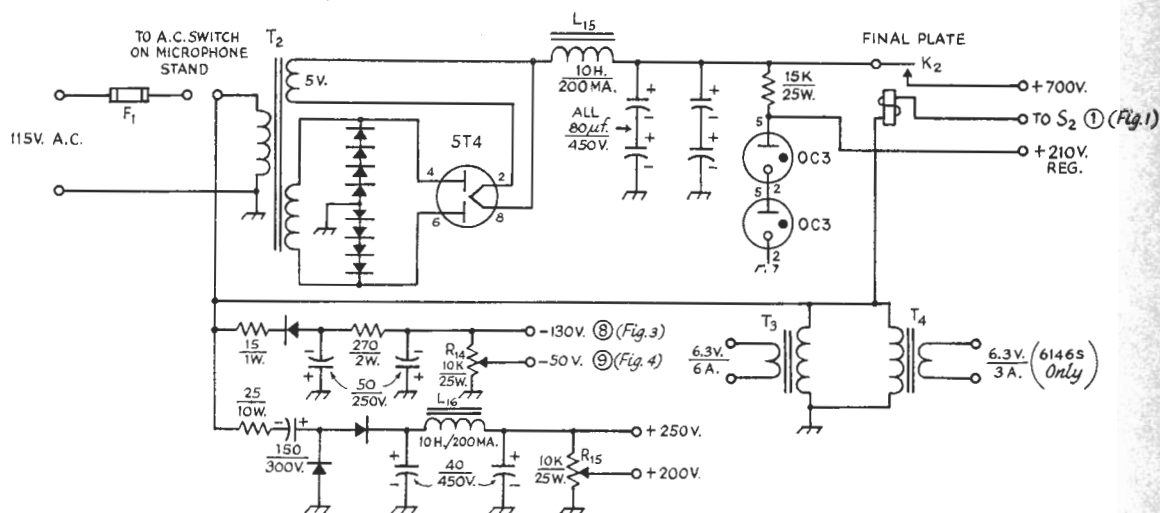


FIG. 5.—Power-supply circuits. Capacitances are in $\mu\text{f.}$ and capacitors are electrolytic. Resistances are in ohms. All solid-state rectifiers are 130-volt a.c., 500-ma. d.c. silicon units (Sarkes-Tarzian M-500). See text regarding use of polarized a.c. line plug.

F₁—Fuse, 5 amp.

K₂—S.p.s.t. 115-v. a.c. relay (Potter & Brumfield KT11A or similar).

L15, L16—Filter choke (Stancor C-2705 or similar).

R₁₄, R₁₅—With adjustable slider.

T₂—Power transformer: 800 volts r.m.s., c.t., 400 ma.; 5 volts, 3 amp. (Stancor PC-8412 or similar, 6.3-volt winding not used).

T₃—6.3-volt 6-amp. filament transformer.

T₄—6.3-volt 3-amp. filament transformer.

Control Circuits

Under stand-by conditions (see S_1), the exciter is producing no output and the receiver is connected to the speaker. The crystal oscillators and the v.f.o. operate continuously. A bias of -130 volts is applied to both the 6BA7 and the 12BY7. Consequently, there is no drive to the 6146s, which are biased with -50 volts on the control grids. A relay is provided in the 700-volt line so that the high voltage can be disconnected from the 6146s.

For transmitting, the -130 -volt bias must be removed, and the receiver speaker shorted. This function is performed by the VOX relay. Talking into the microphone will cause the relay to close. The relay will also be energized if the function switch is placed in the manual position. One set of contacts on the relay closes across the speaker voice coil. A second set grounds the grid returns of the 6BA7 and 12BY7 and effectively removes the -130 -volt blocking bias. These tubes then have a normal bias arrangement and so they amplify the signal produced in the low-level section and drive the 6146s.

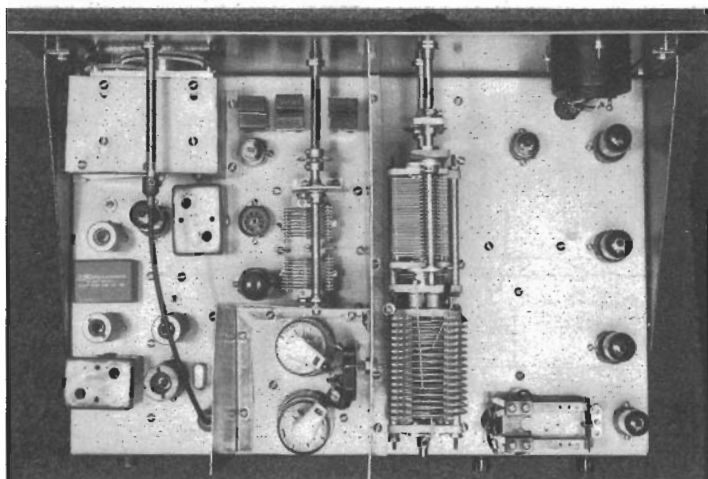
Provision is included for talking yourself on frequency. Remove the 700 volts from the plates of the 6146s, and then place the function switch S_1 in the calibrate position. The VOX relay remains de-energized, but a 100,000-ohm potentiometer, R_{11} , is connected into the -130-volt bias circuit. Advancing this control reduces the bias on the 6BA7 and the 12BY7. It is adjusted so that the modulated

output of the exciter can be heard in the receiver at about the same level as a regular signal. Talking is continued and the v.f.o. tuned until your voice sounds normal. The exciter is then within a few cycles of the desired frequency. This bias and calibrate circuit is very similar to the one used in the Central Electronics s.s.b. exciters.

Construction

The mechanical details of the exciter can be seen in the photographs. The original intention was to include the power supply on the same chassis as the exciter. However, it was decided to use components which were on hand, and these were heavy and required a lot of space. Also, RCA warns that magnetic fields will adversely affect the balance of the 7360 tube. Consequently, the power supply was constructed on a separate chassis and cables used to make the necessary connections to the exciter chassis.

Much time was spent in laying out the main components and arranging the controls so that the front panel would present a reasonably pleasing appearance. After the holes were cut and the construction started, it became apparent that insufficient room had been allowed for the low-level r.f. portion. The isolation between the input and output of the 9-Mc. filter is good, but the 9-Mc. crystal oscillator is not shielded as it should have been. As a result of this, and because of stray coupling to various leads, the carrier suppression is not



Chassis plan view. The tubes and relay to the right are in the audio and VOX circuits. Just to the right of center are the loading capacitor and components of the final-amplifier multiband tuner. To the left of the center shielding partition is the dual driver tuning capacitor with the six band crystals above it and to the right of the v.f.o. compartment. The 9-Mc. coils L_1 and L_2 are in the shield can at lower left, next to the carrier-oscillator tube and crystals, one of which is hidden by the 6146 shield; the other similar can contains L_6 and L_7 . The black box contains the sideband filter. The long flexible shaft extension operates the carrier balance control mounted below deck.

as great as expected. Measurements with available test gear show a carrier suppression of about 45 db. Theoretically, the 7360 can produce a suppression of 60 db. and the filter should add about 10 db. more.

The exciter is built on an 11×17×3-inch chassis. This just fits the LMB W-1D cabinet. This cabinet is 18 inches wide, 11 inches deep and 9 inches high. A pair of Bud MB-458 chassis-mounting brackets is used to brace the panel and chassis.

The method of constructing the bandswitching coil assemblies is described and shown by W6TEU in his article.¹ This arrangement works out very well and is less expensive than using slug-tuned coil forms.

Two r.f. transformers are specially constructed. One is L_1L_2 . The two coils are wound on $\frac{3}{8}$ -inch polystyrene rod as are the band-switching coils. The number of turns and the center spacing of the coils are given under Fig. 2. The coils are mounted vertically on a piece of Micarta plastic $\frac{3}{8}$ -inch thick. The shield is made from an old i.f. transformer can which was cut down. The r.f. transformer, L_6L_7 , is constructed in exactly the same manner. The tuning capacitors are mounted external to the cans.

Alignment

The initial tune-up of the exciter is no more complicated than the alignment of a multiband receiver. However, no signal generator is needed as this is already built in. No one should consider the construction of an exciter of this type without having at least two pieces of test equipment on hand. The first is a

vacuum-tube voltmeter with an r.f. probe attachment. The second is a grid-dip meter with reasonably accurate calibration. A frequency meter such as a BC-221 or LM is also useful to set the final calibration of the v.f.o. However, a receiver can be used for accuracy corresponding to the calibration of the receiver. For s.s.b. operation, accurate calibration is not usually necessary, as it is very convenient to talk yourself on frequency whether this be with a round-table or to a clear spot in the band. Of course, band-edge operation will require some kind of frequency standard.

The first thing to adjust is the balanced modulator. Turn the balance control, R_8 , to either end. Set the differential capacitor, C_5 , to mid-position. Connect the r.f. probe of the v.t.v.m. to the grid of the 6C4. With the 7360 tube operating, adjust the ceramic trimmers C_4 and C_6 for maximum output as indicated on the v.t.v.m. If this occurs with either capacitor at its maximum setting, either L_1 or L_2 needs more turns. Now, alternately adjust R_8 and C_5 so as to produce a minimum output. After this is done, a small readjustment of C_6 will usually result in a slightly lower minimum. More exact adjustments can be made after the whole exciter is operating, and with the signal tuned in on a receiver. In this case, the receiver S meter will serve as the output indicator.

Next, with the selector switch S_3 in the a.m. position, adjust C_{10} for maximum output with the r.f. probe connected across the 500-ohm a.m. control R_7 . With the r.f. probe connected to Pin 8 of the 7360 balanced mixer, adjust

C_9 for maximum output. It is assumed that the slugs of L_3 and L_4 are set so that these maximum adjustments occur within the tuning range of the ceramic capacitors. This completes the adjustment of the low-level section.

The v.f.o. section of the 7360 balanced mixer is adjusted so that it covers the range of 4.0 of 3.5 Mc. The r.f. probe is connected to Pin 7 of the 6BA7. The v.f.o. is set at 3875 kc. and C_{16} is adjusted for maximum output. With the v.f.o. set at 3625 kc., the ceramic trimmer C_{17} is adjusted for maximum output. The capacitance of the r.f. probe will have some effect on the tuning of L_7 , as well as L_2 and L_3 , so these adjustments can be checked again later when the output of the exciter is tuned in on the receiver. The object is to tune L_6 and L_7 so as to produce a passband from 12.5 to 13.0 Mc. The spacing between L_6 and L_7 will affect this also. The spacing of $\frac{3}{8}$ inch (as stated in the coil list) is not necessarily the best, but seemed to produce acceptable results with the shield used. It will be necessary to adjust the resonant frequencies of L_6 and L_7 experimentally to get the desired results. The output at 12.5 and 13.0 Mc. should be down to about half of the maximum so as to keep the passband reasonably narrow. This variation is easily compensated for by varying the drive control to obtain the required output at any desired frequency.

The crystal oscillator is adjusted next. Connect the r.f. probe to Pin 2 of the 6BA7. Disconnect the B+ from the 7360 balanced mixer and the plate of the 6BA7. Adjust each of the plate-circuit tank coils of the 6J6 so as to obtain maximum output for each band position. If the g.d.o. is first used to insure that each of the tank circuits will tune to the required frequency, no difficulty should be experienced.

Restore the 7360 balanced mixer and the 6BA7 to normal operation. Adjust the a.m. control R_7 to obtain output at Pin 7 of the

6BA7 as was done when L_6 and L_7 were being adjusted. Set the v.f.o. to 3750 kc. Set the band switch to 80 meters. Connect the r.f. probe to Pin 2 of the 12BY7. Set the ceramic trimmer (C_{20}) across the L_9 coil so that it is at about half of maximum. Adjust C_{19} for maximum output. Repeat this procedure for each of the other bands, recording the setting of C_{19} for each band.

The band-switching circuits of the 12BY7 are adjusted by shifting the probe to the grids of the 6146s, setting C_{19} to the recorded points in succession, and adjusting the C_{21} trimmers for maximum readings.

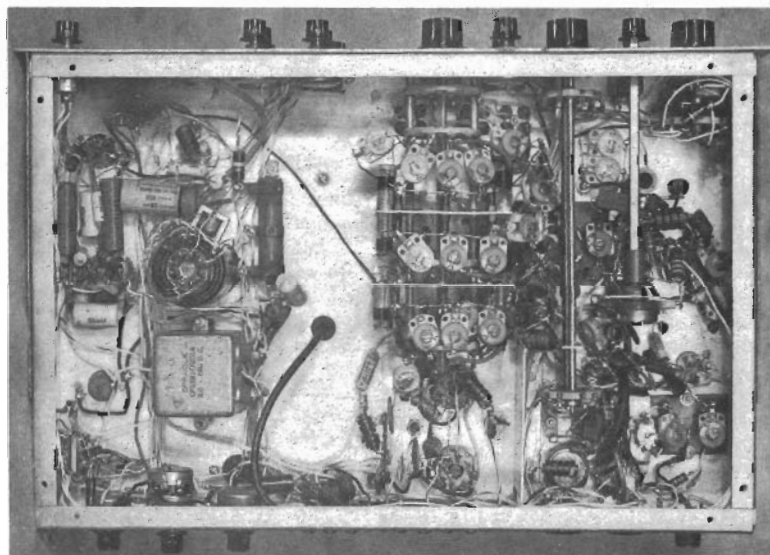
Neutralization of the 12BY7 is performed by disconnecting the B+ from the 12BY7 and adjusting the neutralizing capacitor C_{22} for minimum indication on the r.f. probe. This should be done on the 10-meter band.

The only remaining adjustment is the neutralizing of the 6146s. Connect a dummy load to the output and resonate the tank circuit to the 10-meter band (using the g.d.o.). Disconnect the B+ from the 6146s and adjust the drive control to obtain an indication on the r.f. probe connected to the tank circuit. Adjust the neutralizing capacitor C_{23} for minimum indication.

The preceding alignment information is necessarily brief and hits only the high spots. The previously-cited articles give additional information which should be useful. It is assumed that anyone with enough experience to build such an exciter would have no difficulty with its alignment.

The over-all performance of the exciter has proved to be very satisfactory. Judging from reports, the unwanted sidebands are down at least 40 db. The carrier suppression is apparently adequate as no adverse reports have been received on this score. The a.m. signal has good quality and cannot be distinguished from a standard plate-modulated signal.

Bottom view of the W6PZV s.s.b. unit, showing the band-switching assembly.



A L C Circuits

Automatic level control—or automatic load control, as it is called alternatively—is a form of delayed automatic gain control applied to a transmitter. Its purpose is to prevent modulation peaks from exceeding the linear range of operation. The principle is quite similar to that of a.g.c. as used in receivers. That is, some of the output of the last stage is rectified to develop a d.c. voltage that can be used to control the gain of an earlier low-level stage in such a way that the final output level will not rise above a predetermined value.

In the single-sideband transmitter the a.l.c. circuit is designed to allow modulation peaks to reach the linear peak-envelope level, but not to exceed it. To achieve this, the circuit is adjusted so that it comes into operation only when the amplitude is close to the peak-envelope value; that is, the gain control is delayed until the point of maximum output is almost reached, but then comes into action rapidly so the amplitude cannot reach the "flattening" point.

Rectification of Plate Output

Typical circuits are shown in Fig. 1. The circuit at A can be applied to amplifiers using any type of tube or circuit—i.e., triode or tetrode, grid-driven or cathode-driven. It works directly from the plate of the amplifier, taking a relatively-small sample of the r.f. voltage through the capacitive voltage divider C_1C_2 . This is rectified by the diode of CR_1 to develop a control voltage, negative with respect to ground, across the 1-megohm load resistor. The diode is back biased from a positive voltage source, the bias voltage being adjustable by means of the "level-set" potentiometer R_1 . CR_1 will be unable to rectify until the r.f. voltage exceeds the bias voltage, and by setting R_1 properly no gain-control voltage will develop until the r.f. amplitude is close to the peak-envelope point.

The d.c. control voltage is used to increase the negative bias on a low-level amplifier or mixer, preferably the former, as shown at C. The controlled tube should be of the variable- μ type. The time constant of the control-voltage circuit should be such that the control voltage will rise rapidly when rectification begins, but will hold down the gain during syllables of speech. The time constant can be adjusted by shunting additional capacitance, C_3 , across the 1-megohm resistor, R_2 , in Fig. 1A (the 0.01- μ f. capacitor is simply an r.f. bypass). A value of about 0.1 μ f. is representative.

The capacitive divider C_1C_2 should be designed to apply about 20 volts peak to CR_1 when the amplifier is delivering peak-envelope output. The total capacitance of C_1 and C_2 in series should not exceed 5 to 10 p.f.—i.e., should

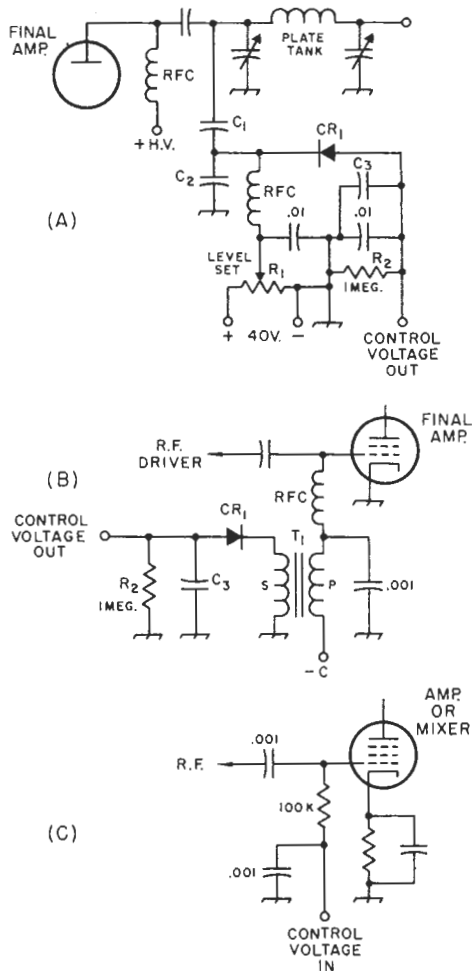


FIG. 1—Automatic Level control circuits.

(A) Control voltage obtained by sampling the r.f. output voltage of the final amplifier. The diode back bias, 40 volts or so maximum, may be taken from any convenient positive voltage source in the transmitter. R_1 may be a linear control having a maximum resistance of the order of 50,000 ohms. CR_1 may be a 1N34A or similar germanium diode. Other values are discussed in the text.

(B) Control voltage obtained from grid circuit of a Class AB₁ tetrode amplifier. T_1 is an interstage audio transformer having a turns ratio, secondary to primary, of 2 or 3 to 1. An inexpensive transformer may be used since the primary and secondary currents are negligible. CR_1 may be a 1N34A or similar; time constant of R_2C_3 is discussed in the text.

(C) Applying control voltage to the controlled amplifier or mixer.

be small in comparison with the tank tuning capacitance so tuning will not be seriously affected. For estimating values, the amplifier peak output r.f. voltage can be assumed to be equal to 75 per cent of the d.c. plate voltage. For example, if the amplifier d.c. plate voltage is 1500, the peak r.f. voltage will be of the order of $0.75 \times 1500 = 1100$ volts, approximately. Since about 20 volts is required, the divider ratio would be $1100/20$, or 55 to 1. This is also (approximately) the ratio of the capacitance of C_2 to that of C_1 . Thus if C_1 is 5 pf., C_2 should be $5 \times 55 = 270$ pf.

Tetrode Grid Rectification

The circuit of Fig. 1B is less flexible and can be used only with grid-driven tetrodes operated Class AB₁. It makes use of the fact that a small amount of rectification occurs in the grid-cathode circuit of a tetrode AB₁ amplifier before the driving voltage actually causes the net grid voltage to be zero and the grid current becomes large enough to cause flattening. This rectification causes a small audio-frequency current to flow in the grid circuit. In the circuit shown, the current causes an a.f. voltage to be developed in the secondary of transformer T_1 ; this voltage is rectified by CR_1 and filtered to negative d.c. by R_2 and C_3 . The resultant d.c. voltage is

used to control an amplifier or mixer as in Fig. 1C. The time constant of R_2C_3 should be chosen as described above. Resistance-capacitance coupling can be substituted for the transformer, although when this is done a voltage-doubling rectifier is generally used so the control voltage will be stepped up. Alternatively, an audio amplifier can be inserted between the grid circuit and the rectifier.

Controlled Stage

The circuits shown here can be modified as necessary to suit individual amplifier and exciter circuits. The details will vary with the actual equipment, but should not be difficult to work out if the principles of the system are understood. Either circuit is capable of developing the few volts of control voltage necessary to prevent the amplifier from being driven into the nonlinear region. The greater the gain between the control amplifier and the stage at which the control voltage is taken off (usually the final amplifier) the less control voltage required. That is, the control voltage should be applied to an early stage in the exciter. Preferably, too, the stage should be one operating on a frequency different from that of the final stage, to reduce the possibility of unwanted feedback.

Temperature Compensation of Oscillators

Finding the right values and coefficients of temperature-compensating capacitors for an oscillator circuit can be a long and tedious task. The following method is used to compensate an oscillator in the Hallicrafters HT-32 s.s.b. transmitter; the principle is applicable to any amateur rig.

In the HT-32 v.f.o. a series-tuned Colpitts (Clapp) circuit is used and, as is necessary in any good oscillator, everything is built like the proverbial battleship. Two capacitors of different temperature coefficients are used with a variable differential capacitor, as shown in Fig. 1. The oscillator is tested by recording the frequency change with temperature. The direction of the drift then indicates which way the differential capacitor must be moved to minimize the deviation.

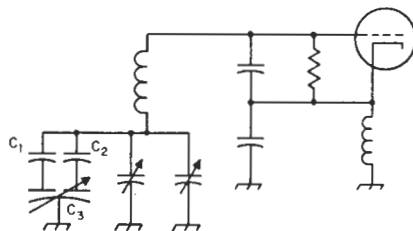


FIG. 1—The v.f.o. in the HT-32 can be set to the best condition of temperature compensation through the use of a differential capacitor of N1500 and NPO coefficients. Changing the rotor position of C_3 permits effective adjustment of the coefficient from an NPO characteristic to N1500.

» The ssb transmitting output of this little handful is 12 peak-to-peak volts at 9 MHz in receiving, the audio output is 2.5 watts for a 50- μ V 9-MHz input signal. Diode switching is used throughout, and there is a novel rf clipper which doesn't require an extra sideband filter.

Transistor Module for SSB Transceivers

A Complete I-F Audio System

GUY M. GILLET, ON5FE

Because of the many advantages to be gained with a transceiver, separate receiver-transmitter configurations are less used these days. Economy may be the most important reason; obviously, one circuit is cheaper than two, and there is needless duplication in separate units. Some other advantages are weight and size reduction, especially for mobile use, and ease of operation. However, a transceiver must often be more sophisticated, because of the problems presented by the switching.

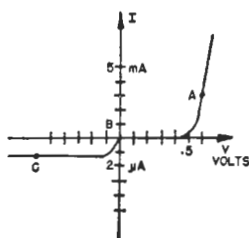


FIG. 1—Typical current-vs.-voltage characteristic of a silicon diode. Note change in current scale below horizontal axis.

These switching problems are not to be minimized, particularly when switching a "hot" lead from receive to transmit. Spurious signals can be induced in the wiring going to relay contacts. Also, energizing the relay may change such circuit parameters as stray capacitance and lead coupling. For example, the load capacitance on a crystal filter may change appreciably when switching from receive to transmit. These troubles can be avoided by diode switching. Let us explain briefly how it works:

A diode forward-biased by a few milliamperes dc (point A on Fig. 1) loses its rectifying property for small alternating currents. That is, a small reduction Δi in the polarization current may be interpreted as a current Δi passing through the diode in the reverse direction. For this Δi , the diode acts like a dynamic resistance of about 50 ohms.

In order to make the diode look like an open circuit, it is usually only necessary to reduce the bias voltage to zero, which results in a dy-

namic resistance of about 100 kilohms (operating point B). However, the diode is usually back-biased to point C to avoid eventual rectification problems and to reduce the effective junction capacitance. This results in a more positive switching action.

Fig. 2A shows a conventional spdt switch, while Fig. 2B is the diode equivalent. If X is connected to $-V$, D_1 conducts and D_2 is off. The ac signal passes through D_1 from point O to M as in the conventional switch. Switching is accomplished by changing X from $-V$ to $+V$ (V is the power-supply voltage) which turns D_1 off and D_2 on. The signal current must be small compared to the dc bias current to avoid nonlinearity.

Overall System

A block diagram of the complete transceiver is given in Fig. 3. This article covers only the blocks within the dotted outline. Design problems with the VFO, driver and power amplifier will not be treated here. Note that there are separate channels for transmitting and receiving; only the filter is switched.

It is obvious that this approach is different from the one often taken in commercial transceivers where, for instance, the i-f amplifier is used for transmit as well as for receive. It may seem that components are being duplicated, and it is true that this circuit arrangement does not offer any advantage if only a simple transceiver is being designed. However, if features such as VOX and rf clipping are desired, the circuit

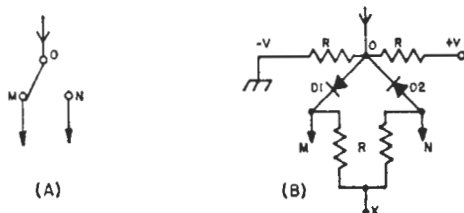
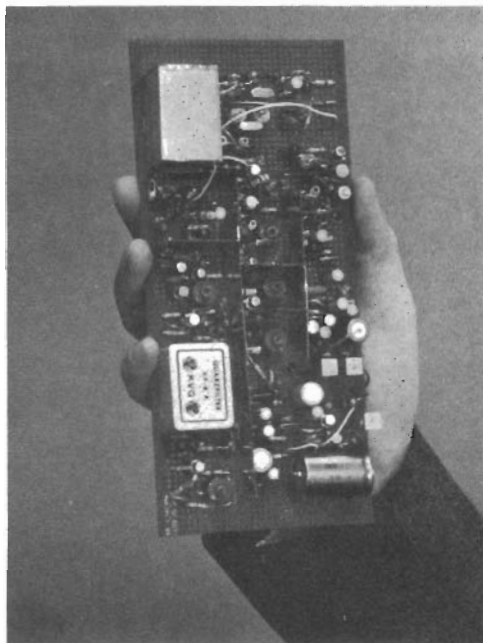


FIG. 2—Conventional switch (A) and corresponding diode switch (B). V is the source of voltage required for biasing the diodes, forward for conduction, reverse for nonconduction. Resistors R are needed for isolating the switched circuit from the dc source.



The small size of the module built by ON5FE is evident from this photograph. It contains a complete transceiver i-f system, including speech amplifier/balanced modulator for transmitting and product detector/audio amplifier for receiving, along with VOX and clipping circuits.

results in a simpler overall design because only one pair of switching points is needed.

Now let us turn to the individual circuits.

Microphone Amplifier and VOX System

The gain of the microphone amplifier in Fig. 4 is 45 dB, and its input impedance is about 150 kilohms. This high impedance can be obtained with a transistor if it is operated at very low collector current (30 μ A) providing it retains a high h_{fe} at this current. The approxi-

mate formula is $Z_{in} = \frac{26}{I_c} h_{fe}$, where i_c is the collector current in mA and h_{fe} is the small-signal current gain. Since noise has an effect at this first stage a low-noise transistor is used and the emitter-follower configuration must be avoided.

There is no microphone gain control, as it was found to be useless because of the rf clipping in later stages. The output is from the emitter and collector of Q_3 ; the balanced modulator gets the emitter signal while the collector output is rectified in a voltage doubler to operate VOX. The amplifier supplies 4 V peak-to-peak output before clipping. Heavy clipping, by talking too loud, must be avoided, but light clipping on higher voice peaks will cause no trouble. The clipping level can be varied from linear (no clipping) to 30 dB by adjustment of potentiometer R_1 in the emitter output circuit. Complete rf decoupling of the microphone amplifier is indispensable.

The VOX circuit is straightforward. Its sensitivity is adjusted by potentiometer R_3 , which sets the forward-bias level at the base of Q_4 when the microphone amplifier is operating. During reception Q_4 is cut off, which in turn cuts off Q_5 . Under these conditions Q_6 is saturated since it is biased for heavy conduction through R_5 . The point labeled R is at nearly +12 V when Q_6 is on; this voltage operates the diode switching circuit for receive, as mentioned earlier. When the rectified VOX signal applies positive bias to the base of Q_4 , this transistor saturates, turning on Q_5 , which then cuts off Q_6 because the voltage at Q_6 's base rises to nearly +12 V. Q_6 and Q_7 form a bistable flip-flop, and with Q_6 off Q_7 goes on, supplying 12 V for the transmit circuits at point T. C_1 and R_4 hold Q_5 on during momentary pauses in speaking.

Feedback from the speaker to the microphone is avoided by means of the antitrip circuit, which also uses a rectifying voltage doubler but with the diodes connected for negative dc output voltage. Adjustment is by R_2 . The three

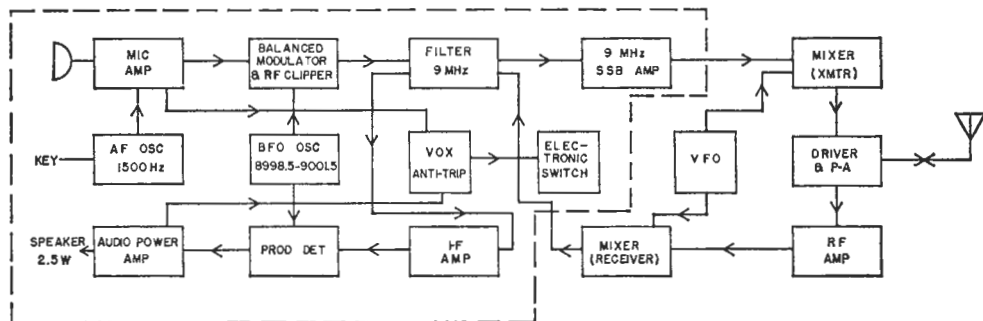


FIG. 3—Block diagram of the ssb module (part enclosed in dashed lines) together with the remaining essentials for a complete transceiver. The module includes i-f, balanced modulator, and audio circuits, and provides 9-MHz ssb output for transmitting.

adjustments, VOX, antitrip and delay, are nearly independent.

Balanced Modulator and RF Clipper

The balanced modulator, Fig. 5, uses two varactor diodes, back-biased at 4.5 volts through potentiometer R_7 , which is adjusted to cancel any mismatch in the diodes' characteristics. The 9-MHz carrier is applied to the arm of R_8 , and due to symmetry, no voltage will be developed across the 9-MHz resonant circuit, L_1C_2 . The audio voltage is applied to the arm of R_7 , causing the capacitance of the varactors to vary in opposite directions and thus aiding each other in the unbalancing. An interesting feature of this circuit is that the input is asymmetric—i.e., single-ended—for both the high-frequency and audio signals. Also, the input impedances are much greater than those of the classical four-diode modulator.

Inductor L_1 is constructed on a toroidal core because the coupling must be purely magnetic if a high carrier-rejection ratio is to be obtained. Note that the modulator can be easily unbalanced by connecting a resistance in parallel with R_8 (point K grounded). In this way, cw or a-m signals can be produced. The a-m signal could be taken from the open secondary of L_1 with point K at ground, although this possibility of producing a-m is not used in the present circuit.

European coil forms were used throughout this transceiver so information on the number of turns would be nearly useless if different coil forms are used. The turns ratio will provide sufficient information. One can easily construct suitable coils with his own material with the help of a grid-dip meter. The requirements are:

Audio: 4 V pk-pk max.

Carrier: 1 rms.

Carrier rejection: Better than 50 dB.

Turns ratio on L_1 : $n_p/n_s = 14/4$.

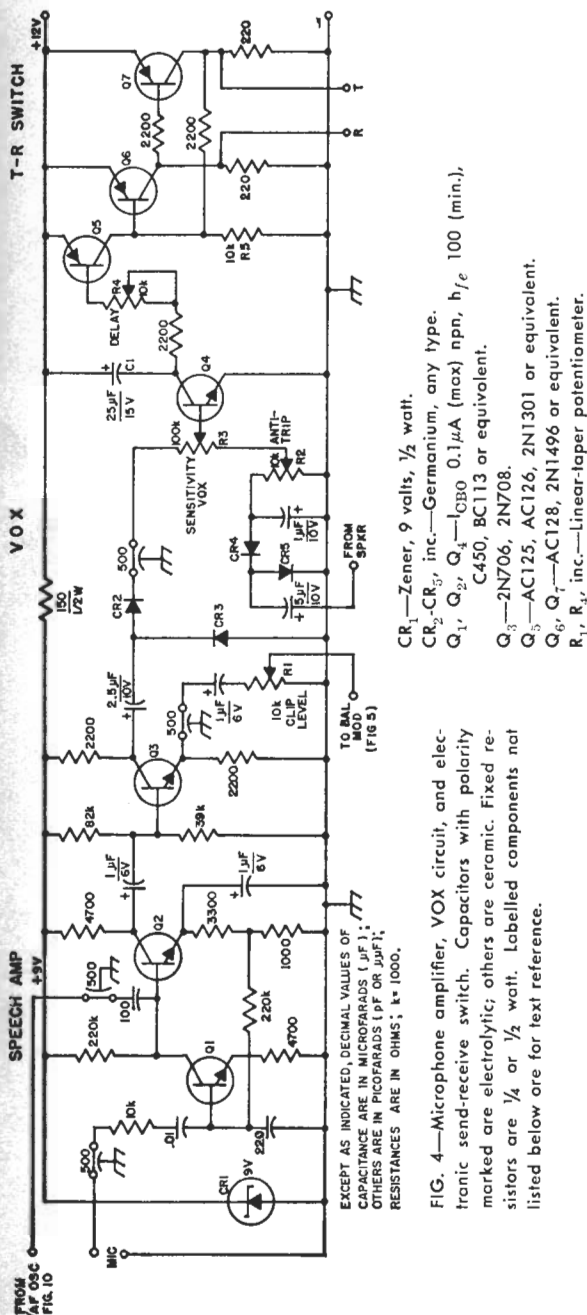
Varactor diodes: $C = 20$ pF at -4 V (not critical).

Output: 200 mV pk-pk rf for 2.8 V pk-pk audio input.

Distortion seems to be less than with the four-diode modulator. It is not necessary to have well-matched varactors; their differences may be cancelled by adjustment of R_7 .

RF Clipper

It is well known that speech remains intelligible even with infinite clipping; in other words, the amplitude contains relatively little information as far as intelligibility is concerned. However, keep in mind that in the absence of noise at the receiver a clipped signal will always be less intelligible than a normal signal without clipping. It is only under marginal conditions (high noise level) that clipping presents an appreciable improvement. Consequently, the operator must maintain control of the clipping level. Using the peak power as reference (and



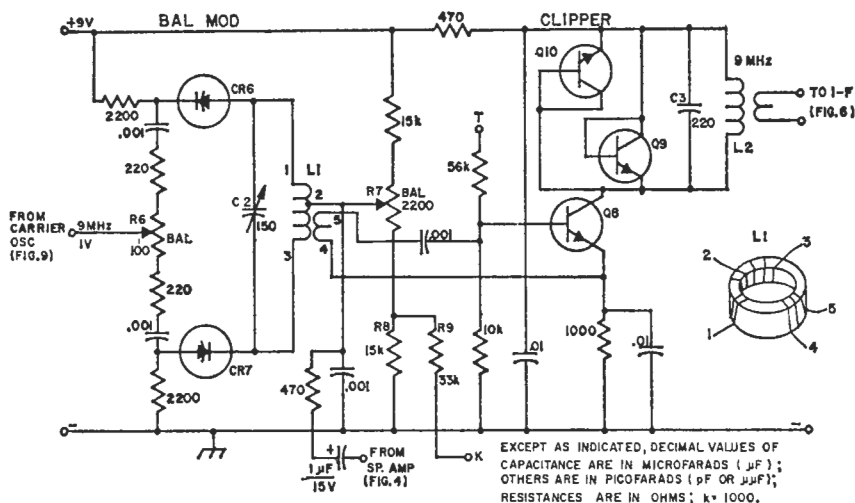


FIG. 5—Balanced modulator and r.f. clipper circuit. Inset shows construction of L_1 toroid with terminal numbers. Capacitor with polarity marked is electrolytic; other fixed capacitors are ceramic. Fixed resistors are $\frac{1}{4}$ or $\frac{1}{2}$ watt. Labelled components not listed below are for text reference.

C_2 —150-pF miniature variable.

C_3 —Mica.

CR_6 , CR_7 —Varactor diode; see text.

Q_8 , Q_9 , Q_{10} —2N706, 2N708.

R_6 , R_7 —Linear-taper potentiometer.

R_9 —App. 33,000 ohms; see text.

L_1 , L_2 —Toroid, pri./sec. turns ratio 14/4; primary to resonate at 9 MHz with C_2 and C_3 , respectively (app. 3 μH and 1.5 μH , respectively). Terminal 2 on primary is a center top.

this is a good approach to the problem, because ssb transmitters are limited by their peak power rather than their average power dissipation) it has been found that the following degrees of improvement can be obtained:

	Amount of Clipping	Improvement (in dB)
Audio clipping	15 dB	4 dB
	25 dB	5.5 dB
Rf ssb clipping	10 dB	4 dB
	20 dB	8 dB

The table indicates, for example, that during marginal conditions an rf ssb clipping level set at 20 dB can have a peak power 8 dB under the peak power of an unclipped signal and maintain the same intelligibility. The table also shows that rf clipping has greater efficiency than audio clipping. This is because in rf clipping most of the spurious signals created fall outside the pass band. However, a crystal filter must always follow rf clipping in order to eliminate spurious signals near the edge of the band and to avoid transmitting an excessively broad spectrum. Unfortunately, the crystal filter is presently the most expensive item in a single-side-band transmitter. To economize, we decided to clip the dsb signal and use only one filter to extract ssb and to follow the clipping stage. But "you get what you pay for," and dsb clipping creates more intermodulation products inside the transmitted bandwidth than ssb clipping does.

The rf clipper is a standard clipping circuit

except that two transistors in diode configuration are used instead of two diodes. The clipping action is slightly better. Two examples of rf clipping at different levels can be seen in the photographs taken at the collector of Q_8 . The third picture, taken at the output of Q_{12} , Fig. 6, is two-tone test modulation with strong clipping, after passing through the crystal filter. It shows some distortion, but at an acceptable level. When using a high clipping level, note that the plate of the power amplifier must dissipate a significantly higher average power.

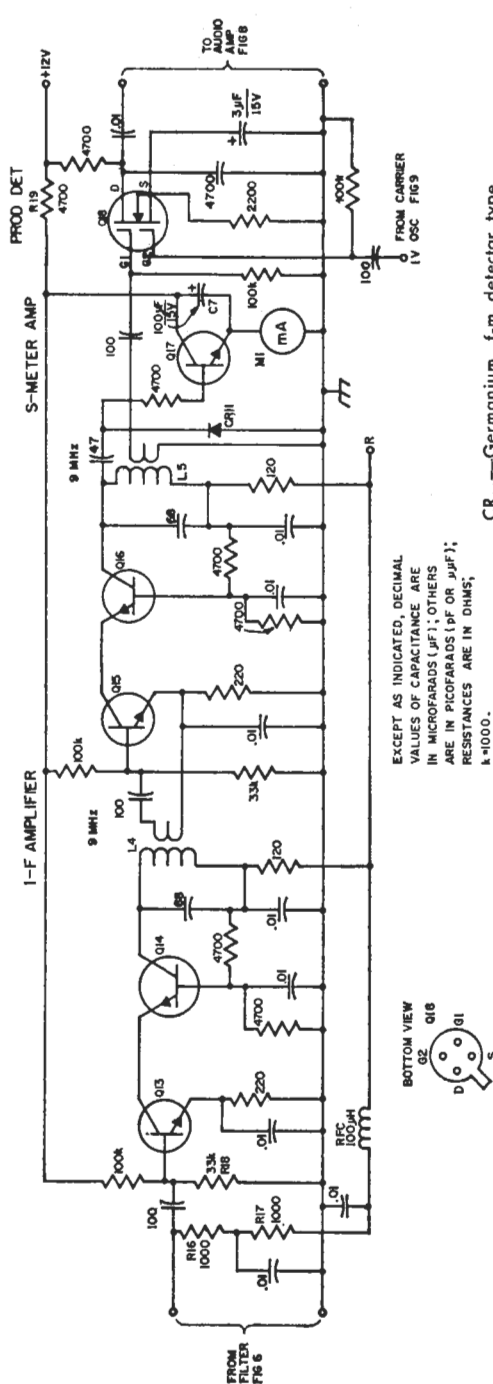
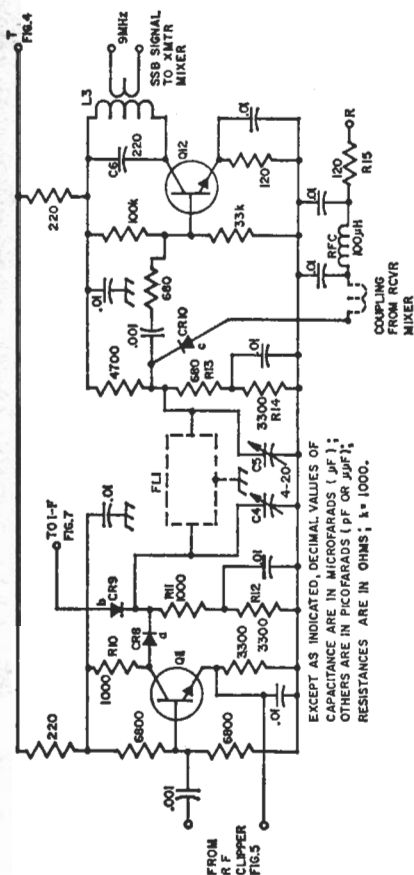
Crystal Filter

A commercial European filter (KVG type XF9A) was used. Its characteristics are:
 Bandwidth: 2.5 kHz at -6 dB.
 Pass-band irregularities: Less than 1 dB.
 Insertion loss: Less than 3 dB.
 Spurious response: Less than 45 dB.
 Shape factor: 1/1.7, 6 dB/50 dB.
 Input and output impedances: $R = 500$ ohms, $C = 30$ pF.

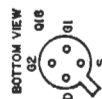
If this low-cost German filter is not available,¹ any filter can be used with minor changes in R_{10} , R_{11} and R_{13} .

Q_{11} provides the correct input impedance for the filter. Q_{12} is a buffer stage providing 12 volts peak-to-peak on its collector. This signal is large enough to feed a vacuum-tube mixer. If a transistorized mixer is desired, suitable

¹ KVG filters are available in the U.S. from Spectrum International, Topsfield, Mass. 01983. —Editor



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



$CR_{1,1}$ —Germanium, f-m detector type.

L_1, L_2 —Toroid pri./sec. turns ratio 10/1; primary to resonate at 9 MHz. with 68 pF (app. 4.5 μH).

M_1 —0-1 dc milliammeter.

Q_{11}, Q_{12} —inc.—2N706, 2N708.

Q_{13} —Same as Q_{11} , Fig. 4.

Q_{14} —MOSFET, 3N141.

R_{19} —App. 4700 ohms; see text.

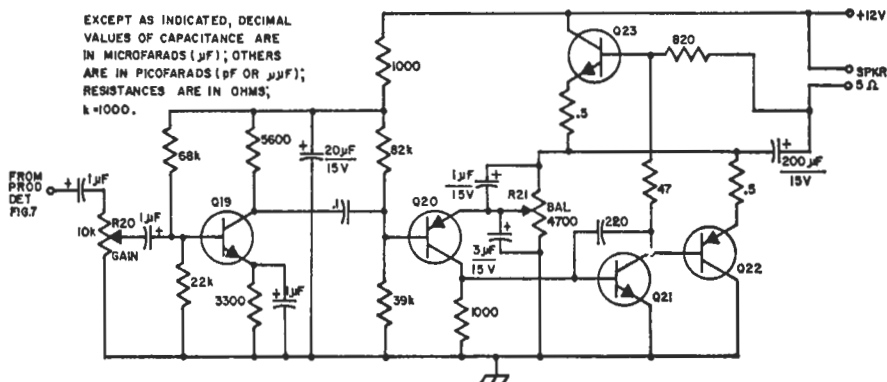


FIG. 8—Audio amplifier circuit. Capacitors with polarity marked are electrolytic; others are ceramic. Fixed resistors are $\frac{1}{4}$ or $\frac{1}{2}$ watt.

Q₁₉, Q₂₁—Same as Q₁, Fig. 4.

Q₂₀—AC125, AC126, 2N1301.

Q₂₂—Pnp, 800 mW, h_{fe} 200, I_{CBO} 100 μ A, I_C 1A
BV_{CBO} 25 V.

Q₂₃—Some characteristics as Q₂₂, but npn (complementary pair).

R₂₀—Logarithmic-taper potentiometer.

R₂₁—Linear-taper potentiometer.

matching is needed and it is recommended that a balanced mixer be used.

Filter Switching

During transmit, point T is at 12 volts and point R is at about 1 volt. CR₈ then conducts through R₁₀, R₁₁, and R₁₂. CR₉ and CR₁₀ are back-biased.

During receive, T is at about 1 volt and R is at 12 volts. CR₁₀ now is forward-biased through R₁₃, R₁₄, and R₁₅. The signal from the mixer is applied to the filter through CR₁₀. CR₈ is back-biased; CR₉ conducts through R₁₆ and R₁₇ (see i-f schematic, Fig. 7) passing the filtered ssb to the i-f amplifier.

I-F Amplifier and Product Detector

The traditional transistorized hf amplifier (common-emitter circuit) does not respond to agc as well as its vacuum-tube counterpart. The LC circuit is detuned and the Q varies when the agc acts. Circuit stability may even be compromised. In Fig. 7 these troubles are avoided by using two transistors per stage in a cascode arrangement. The first transistor is stable because its load is the very low input impedance of the next transistor, which is in a common-base configuration. The second transistor is stable in this common-base arrangement, so the LC circuit is practically independent of agc action. The gain of a cascode amplifier is slightly greater than that of a single transistor in the common-emitter mode. At 9 MHz the gain may be 50 dB or more per stage. Unfortunately, 80 dB cannot be obtained with one cascode stage and two stages provide too much gain. In order to reduce the gain, R₁₈ must be decreased until the gain is reduced to the optimum 80-dB figure.

The agc amplifier is Q₁₇, which also controls

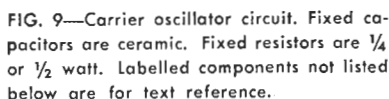
the S meter. The agc dynamic range is 0 to -70 dB without distortion. When no signal is present, the collector current of Q₁₇ is zero. When a signal is received, the transistor begins to conduct, causing a voltage drop in R₁₉ and reducing bias current in the cascade stages. The agc has a fast action and slow decay due to C₇. Q₁₇ can discharge C₇ very rapidly, but the capacitor must recharge through R₁₉. The time constant is $\frac{1}{2}$ second. Note that C₇ is not grounded; this prevents the S meter from responding to momentary peak currents.

The RCA 3N141 dual-gate MOSFET is a very good product detector. The high insulation between gates aided in design. As usual, the carrier must be injected at a higher level (about 20 dB) than the ssb signal to minimize the intermodulation products. For one volt rms carrier, a 100-mV rms signal gives 700 mV peak-to-peak audio, and a 10-mV rms signal gives 70 mV peak-to-peak audio.

Another I-F Circuit

An alternative i-f amplifier circuit is given in Fig. 11. Usually, successive amplifier stages are fed in parallel, but at high frequencies a series arrangement may also be used, when the load is not resistive. Only the i-f stages are series connected in Fig. 11, but an rf stage, mixer stage and product detector may be added easily.

The first stage uses an FET because our experimental model had an FET in the rf amplifier. Using the FET at the front end of the i-f amplifier resulted in a single agc circuit for i-f and rf. L₁, C₁ and C₂ (and similarly, L₂, C₃ and C₄) may be regarded as a pi circuit matching the high output impedance of Q₁ (about 50 kilohms) to the low input impedance of Q₂ (about 10 ohms). Generally, in an i-f amplifier the selectivity is of paramount importance, but



Y₁—8998.5 kHz.
Y₂—9001.5 kHz.
(Crystal frequencies are for a 2.5-kHz.-bw filter having a center frequency of 9000 kHz.)

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μ F); OTHERS ARE IN PICOFARADS (pF OR μ F²); RESISTANCES ARE IN OHMS; k = 1000.

TO
COLL.
OF Q1
FIG. 4

6Q25

AMPLITUDE
4700

1 μ F

KEY

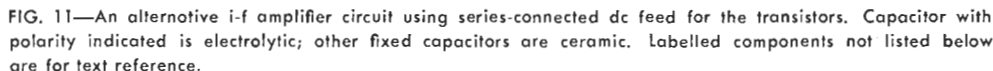
10k

33k

C12 .0012

C11 .0012

C10 .00012



L_3 —Toroid, pri./sec. turns ratio 10/1; primary inductance 4 μ H.
 M_1 —0-1 dc. milliammeter.
 Q_1 —MPF-102 or equivalent.
 Q_2, Q_3 —2N706, 2N708.

here the crystal filter bandwidth has already done the job of selection.

A gain greater than 100 dB can be provided by a circuit such as this without sacrificing stability. Q_1 remains stable because of the low (500 ohms) input impedance from the filter.

The agc needs a polarization voltage or bias of not less than 30 volts. A convenient source is the rf power-amplifier negative bias. The

value of R_1 is determined by $R_1 = \frac{V \text{ bias}}{5.0 \times 10^{-4}}$.

The Audio Amplifier

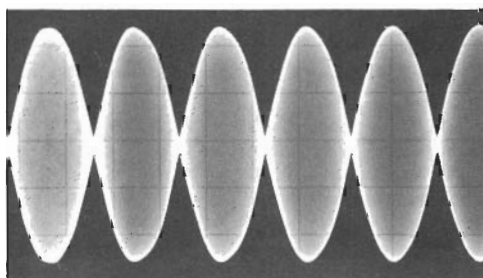
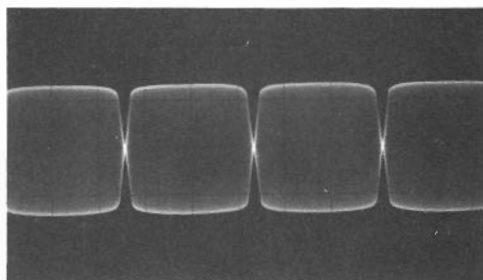
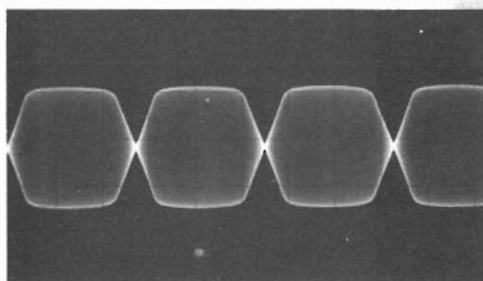
Any audio circuit which can provide the desired output from the 100-mV peak-to-peak input is, of course, suitable. A series push-pull transistor circuit such as is shown in Fig. 8 must have a power supply with very low output impedance. The point cannot be overstressed, for neither the audio output nor the other circuits can work properly otherwise. Regulation is especially important in the case of high power and low supply voltage. In the circuit presented, the feedback is applied to the emitter of Q_{20} , and not to the base as is often the case. With this configuration Q_{20} has a high input impedance.

Carrier Oscillator

In the carrier oscillator circuit, Fig. 9, the crystals for lsb and usb are used in a parallel-resonant mode and may be pulled slightly by a parallel capacitance. Again, diode switching is used to avoid long signal leads. When the sideband selector, S_1 , is at position 1, a dc current flows through CR_{12} and R_{20} to ground. Crystal Y_1 (usb) is then at ac ground through CR_{12} and C_{10} . At selector position 2, the dc passes from the +12 V input through R_{21} , CR_{14} , CR_{13} and R_{22} to ground. Crystal Y_2 is then at ground via CR_{13} and C_{10} . The asymmetric diode CR_{14} is necessary to allow the lsb crystal Y_1 to function. Without CR_{14} , Y_1 is shunted by the two resistors, R_{20} and R_{21} , and crystal Y_2 , which is near the series-resonant mode of Y_1 . R_{23} is necessary to prevent heavy distortion in the output of Q_{24} . The oscillator provides an output at a level of 1 volt rms.

Tone Oscillator

CW operation can easily be obtained by unbalancing the modulator, as mentioned earlier. Another method of cw operation is keying an audio oscillator in order to keep the VOX operational; this same audio oscillator is also useful as a tuning aid before ssb operation. The circuit of Fig. 10 is a classical phase-shift oscillator. The sine-wave output may be adjusted in amplitude by R_{24} . Output is taken from the collector and sent to the microphone amplifier through a 100-pF capacitor, Fig. 4. The 500-pF feedthrough capacitor in Fig. 4 will change the frequency, as do the values of phase-shift capacitors C_{11} , C_{12} and C_{13} . Keying does not



Top: Single-tone signal with moderate clipping, measured at the input to the 9-MHz. sideband filter. Center: Same with heavy clipping. Bottom: Heavily-clipped two-tone signal after passing through the sideband filter. Post-clipping filtering restores the original modulation envelope with little distortion.

alter any dc current, thus helping to avoid key clicks.

It was found interesting and helpful to set the tone to 1500 Hz and tune the carrier oscillator. In this way 8998.5 kHz and 9001.5 kHz signals can be adjusted with a 9-MHz standard (BC-221, for example).

Acknowledgments

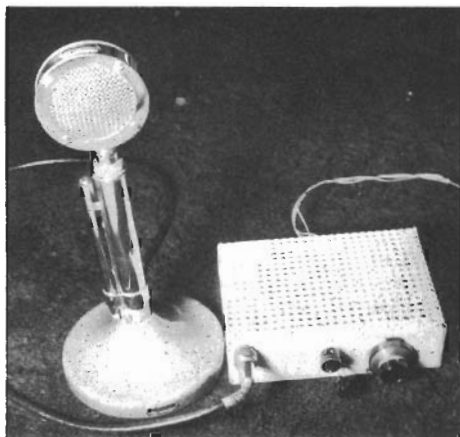
Thanks to my friend ON4JN, whose suggestions and aid during this project were most helpful. And thanks to K9QGL and VE3BHT who helped rewrite my translation.

References

- 1—Craiglow, Getzin and Swanson, "Power Requirements for Speech Communication Systems," *IRE Transactions on Audio*, November-December, 1961.
- 2—Pappenfus, Bruene and Schoenlike, *Single Sideband Principles and Circuits*, McGraw-Hill Book Company.
- 3—Perkins, "Transistor Cascode Circuit Improves Automatic Gain Control in Amplifiers," *Electronics*, June 2, 1961.

» The ultra-ultra in compactness demands going to transistors. Here is an example of an all-semiconductor transceiver that is practically pocket-size. The circuit techniques will be of interest to those who want to explore the solid-state field.

This complete 14-Mc. transistor transceiver is contained in an enclosure measuring 5 by 7 by 2 inches. The vernier tuning dial controls the VXO frequency by adjustment of C_6 . Of the two smaller knobs to the left, the lower one operates the transmit-receive switch, while the upper one is the r.f. gain control.



A Solid-State S.S.B. Transceiver

BENJAMIN H. VESTER, W3TLN

For some years, I've had a hankering to try my hand at a transistorized s.s.b. transceiver. Being somewhat prejudiced toward the single-conversion approach with a relatively high i.f., I've had to wait until the transistor art boiled out some good units for use in the h.f. region. At the same time, miniature low-voltage capacitors and other components have been developed and are now readily available at low prices. After surveying a recent wholesale flier, I decided the prices were now reasonable enough to start building. For reasons which are somewhat fuzzy now, I settled on 20 meters as the best band, although the design is suited to other bands as suggested later.

The basic arrangement of this transceiver is almost identical to that of the tube model described earlier,¹ the key features being (a) use of a high-frequency crystal filter to allow single conversion and (b) use of a VXO for the tunable oscillator. The transmit-receive switching is accomplished manually with a miniature wafer switch which interrupts the B+ to stages which are inactive for the mode in use.

With an eye toward future installation in my Volkswagen, I restricted myself to a single 6-volt power supply. As will be noted, this somewhat limits the amount of d.c. stabilization one can use, and also limits the power output obtainable.

Receiver Front End

The schematic starts in Fig. 1. The r.f. amplifier, Q_1 , is in a standard neutralized ground-

ed-emitter circuit with double-tuned input. With the poor intermodulation characteristics of transistors, as much selectivity as practical should be inserted "up front." L_1 and L_4 are wound on separate link-coupled powdered-iron toroids with an electrostatic shield between them. The whole r.f. stage is mounted in one of the Command-set i.f. cans with the two capacitors therein being used to tune L_1 and L_4 . L_2 and L_3 are each a single turn which is slid around the toroid until proper coupling is obtained; i.e., until a passband of about 500 kc. is obtained. The electrostatic shield is the same shielding disk found in the i.f. cans.

The collector coil, L_5 , is wound on a CTC LS-9 coil form. The LS-9 is a completely-shielded, ferrite-loaded form which is quite small. Having a group of these forms salvaged from a surplus military receiver, I used them throughout the unit. A small, tunable coil like this is, of course, a key factor in achieving miniature design.

The receiver mixer, Q_2 , is conventional, capacitor C_4 being chosen empirically to give the maximum mixer efficiency. L_7 and L_8 are wound on another LS-9 form with the appropriate impedance step-down for the crystal filter which follows in Fig. 2.

VXO

The VXO with transistors is slightly different from the tube type. The crystal operates in its series-resonant mode instead of parallel resonance. The VXO crystal, Y_1 , was one of several given to me by W3BWK; its fundamental

¹ Vester, "Mobile S.S.B. Transceiver," *QST*, June, 1959.

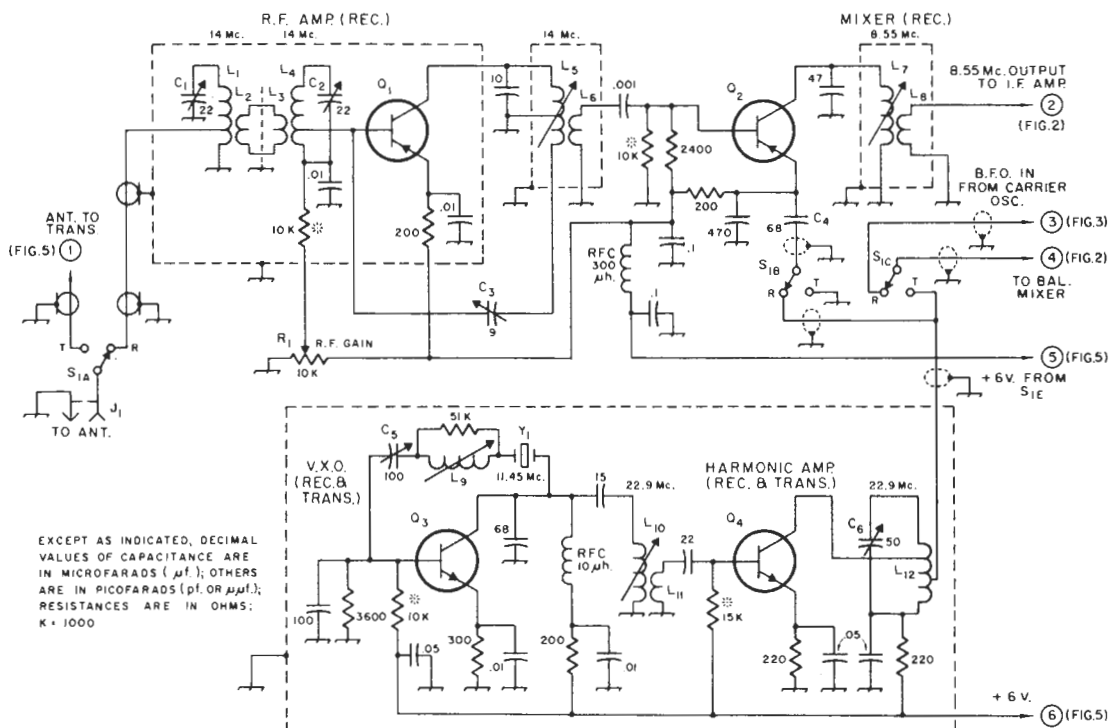


FIG. 1—VFO and high-frequency receiver circuits. The 22.9-Mc. VFO signal mixes with a 14.35-Mc. incoming signal to produce an 8.55-Mc. i.f. signal. On transmit, the 22.9-Mc. signal is transferred to the balanced diode mixer of Fig. 2. Fixed capacitors of decimal value are miniature ceramic or paper with a minimum rating of 6 volts. Others are NP0 ceramic or dipped silver mica. Resistors are $\frac{1}{4}$ watt. Resistors marked with asterisks are bias resistors (see text).

- C1, C2—Air trimmer from Command-set i.f. cans.
- C3—0.1–9-pf. trimmer, or "gimmick."
- C4—Nominal value; see text.
- C5—Air variable (Hammarlund APC-100B).
- C6—Ceramic trimmer.
- J1—Phono jack or chassis-mounting coax receptacle.
- L1—40 turns, tapped at 2 turns from ground end, on powdered-iron toroid (Stackpole D-1 iron) $\frac{1}{2}$ -inch outside diameter, $\frac{1}{4}$ -inch inside diameter, circular cross section (Henry L. Crowley Co., West Orange, N. J., part No. C-2776).
- L2—Single turn on L1; see text.
- L3—Same as L2, wound on L1.
- L4—Same as L1, tapped at 4 turns.
- L5—21 turns of double-strand No. 34 enameled (bifilar-wound) on CTC LS-9-5S shielded ferrite-slug form. Finishing end of one strand is connected to starting end of other strand to form center tap; two remaining ends connected to circuit as shown.
- L6—3 turns over L5.
- L7—25 turns on CTC LS-9-4S shielded ferrite-slug coil form.

- L8—6 turns over ground end of L7.
- L9—48 turns close-wound on 1-inch ceramic iron-slug form (National XR-60 form).
- L10—Inductance 3.5 μ h., scramble-wound on CTC PLST-2C4L/N iron-slug form.
- L11—4 turns over ground end of L10.

Note: Above coils are close-wound with No. 34 enameled wire.

- L12—12 turns No. 24, $\frac{1}{2}$ -inch diam., 32 turns per inch (B & W 3004 Miniductor), tapped at 4 turns and 7 turns from ground end.

Q1, Q2—2N700, or similar u.h.f. p.n.p. transistor (see text).

Q3, Q4—2N706, or similar u.h.f. n.p.n. silicon transistor (see text).

R1—Linear-taper control.

S1—Subminiature ceramic rotary switch, 2 sections, 5 poles, 2 positions (Centralab PS-117, 1 pole and 1 position not used). See Figs. 2 and 5 for remaining poles.

Y1—11.45-Mc. crystal.

frequency (11.450 Mc.) was half of the desired frequency (22.9 Mc.), so some harmonic selection and amplification was necessary. This was not quite so easy as with pentode tubes, and another transistor, Q4, was required. Prior to

putting in this stage, with its associated tuned circuits, the 11.45-Mc. signal leaking into the receiver mixer was enough to allow high-power teletype signals just below 20 Mc. ($11.45 + 8.55 = 20$ Mc.) to be heard in the receiver.

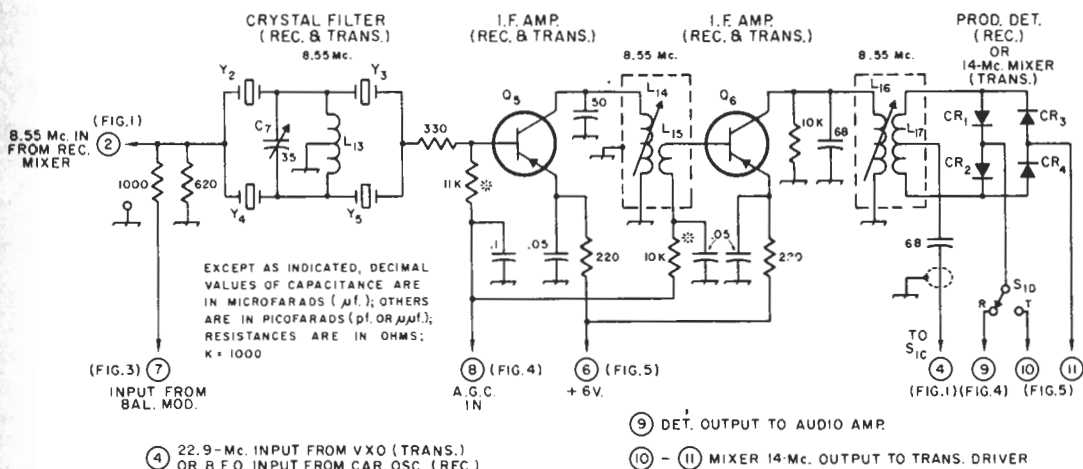


FIG. 2—8.55-Mc. i.f. circuit. On receive, the diodes in the output circuit operate as a product detector, the carrier oscillator (Fig. 3) serving as the b.f.o. On transmit, the 8.55-Mc. suppressed-carrier signal from the balanced modulator (also Fig. 3) passes through the crystal filter which strips off the unwanted sideband. The remaining sideband passes through the i.f. amplifier to the diode network, which now operates as a balanced mixer, where it mixes with the 22.9-Mc. V.X.O. signal to produce a 14.35-Mc. signal for the transmitter (Fig. 5). Fixed capacitors of decimal value are miniature ceramic or paper with a minimum rating of 6 volts. Others are NP0 ceramic or dipped silver mica. Resistors are $\frac{1}{4}$ watt. Asterisks identify biasing resistors (see text).

C7—Ceramic trimmer (Centralab 827D).

CR1, CR2, CR3, CR4—Germanium diode (CK706, IN34A or equivalent.)

L13—20 μ h., center-tapped, bifilar-wound on $\frac{3}{4}$ -inch ferrite toroid core, and connected as described for L5. Cores available from same source as L1. See references 1, 2.

L14—12 turns on CTC LS-9-4S, shielded ferrite-slug coil form.

L15—4 turns wound over ground end of L14.

L16—24 turns on CTC LS-9-5S iron-slug form.

L17—8 bifilar turns, wound over L16 and connected as described for L5.

Note: Above coils are close-wound with No. 34 enameled wire.

Q5, Q6—Same as Q1.

S1—See Fig. 1.

Y2, Y3—8550.3-ke. crystal.

Y4, Y5—8551.5-ke. crystal.

If you can get a crystal whose fundamental is at 22.9 Mc., you can avoid the extra stage.

Of course, the tuning range of the V.X.O. depends on the inductance of L9. I put on just enough turns to make the V.X.O. cover the most active part of the s.s.b. portion of the band with the slug all the way out. With the slug advanced to a preset stop, the V.X.O. tunes all the way down to the bottom end of 20. There is some loss in v.f.o. stability at this setting, but with the "cool" transistor circuits, the stability is still as good as that of a number of commercial receivers.

I.F. Filter and Amplifier

The crystal filter (Fig. 2) has been covered before^{1,2}; capacitor C7 and coil L13 are chosen to resonate approximately (disconnected from the circuit) at the passband frequency of the filter. Adjustment of C7 and the slug of L7 (Fig. 1) can then be made to optimize the filter passband.

The i.f. amplifiers, Q5 and Q6, are conventional, with coils wound on LS-9 forms being

used for interstage coupling. These stages were not neutralized, and some intentional interstage impedance mismatch was used to keep the circuits noncritical.

The diodes fed by Q6 serve both as a product detector for receiving and as the transmitter mixer, where the 8.55-Mc. i.f. signal and the 22.9-Mc. V.X.O. signal are mixed to produce 14-Mc. output. The diodes are in a balanced arrangement so that both the V.X.O. and the 8.55-Mc. signals are suppressed when transmitting. The diodes are garden-variety germanium with no particularly good balance requirements on them.

Carrier Oscillator and Balanced Modulator

The carrier oscillator and balanced modulator (Fig. 3) are conventional, and are both stuffed into the same Command-set i.f. can to contain the carrier leak-through. Both the fixed and variable capacitors already mounted in the i.f. can are used. L18 and L19 are wound on another miniature powdered-iron toroid which is supported by plastic tape wrapped around two of the posts in the i.f. can. Crystal Y6 is similarly supported on the other two posts.

² Arnold and Allen, "Some Ideas in a Ham-Band Receiver," page 181.

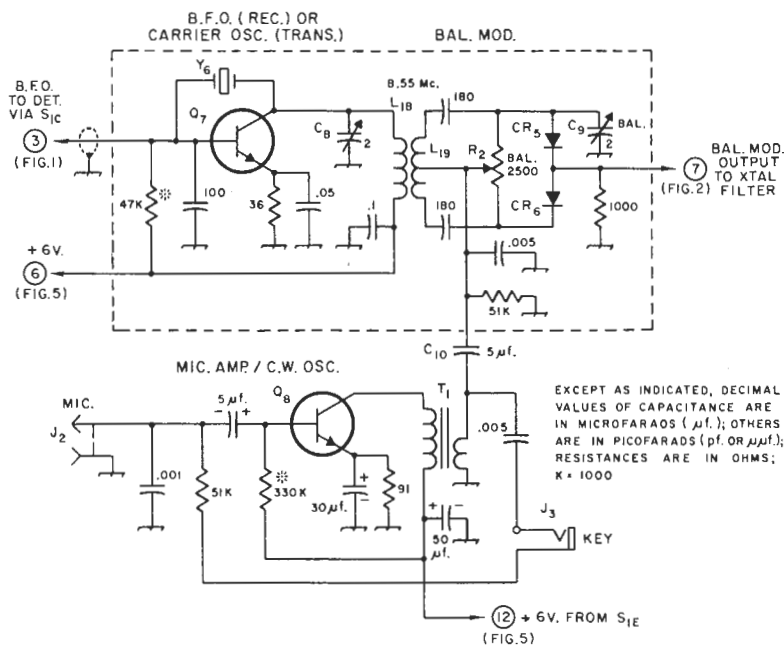


FIG. 3—Carrier-oscillator (b.f.o.), balanced-modulator, and transmitter audio circuits. A feedback circuit is provided to cause the microphone amplifier to oscillate for c.w. operation. Fixed capacitors of decimal value are miniature ceramic or paper. Except as listed below, others are NPO ceramic or dipped silver mica, except where polarity indicates electrolytic. All capacitors have a minimum rating of 6 volts. Fixed resistors are $\frac{1}{4}$ watt. Biasing resistors are identified by asterisks; see text.

C8, C9—22-pf. air trimmer from Command-set i.f. can.

C10—Ceramic or paper.

CR5, CR6—Same as CR1.

J2—Microphone connector.

J3—Miniature open-circuit phone jack; both sides must be insulated.

L18—60 turns No. 34 enameled, close-wound on toroid form same as described for L1.

L19—10 bifilar turns over L18, wound and connected as described for L5.

Q7—Same as Q3.

Q8—2N170 or similar.

R2—Linear control.

T1—Subminiature interstage audio transformer, 4:1 turns ratio, low-impedance winding in output.

Y6—8553.0-kc. crystal.

C9 was tried on both ends of the balance pot to obtain the best carrier suppression.

The audio amplifier used in the receiver could have been switched into use as a microphone amplifier, of course, with some small saving in parts. The additional switch contacts required didn't justify it with the parts and space I had available. As shown in Fig. 3, the addition of a feedback path around the microphone amplifier is a handy technique for generating a tone for both tune-up and c.w. operation.

Receiver Audio and A.G.C.

The audio amplifier (Fig. 4) was built around a couple of transformers I salvaged from a hearing aid, and transistors from the junk box. Anyone considering building a unit like this would do well to copy the audio circuits from Priebe's excellent receiver article,³ or buy one of the packaged units available from Lafayette Radio.

The a.g.c. rectifier and amplifier feed directly off the output transformer. As can be seen, this will provide a.g.c. to maintain the same audio level at all times. Having only enough panel space for a single gain control, I chose to make it an r.f. gain control. Of course, the audio level that the a.g.c. tries to hold could be adjusted by running R_3 to a potentiometer similar to the r.f. gain control. The "hang" action of this a.g.c. is not as good as with similar tube circuits, but it seems to be a reasonable compromise with miniaturization since it uses only four tiny parts.

Transmitter Output Stages

The transmitting amplifier, Q13, in Fig. 5, is a straightforward Class A stage. The "final," Q14, is a high-frequency silicon switching transistor which is run Class B, with the emitter grounded directly. The bias resistor, R_4 , must be empirically chosen for any particular transistor to give a static collector current of 3 to 5 ma. Since the switching transistor has a

³ Priebe, "All-Transistor Communications Receiver," QST, February, 1959.

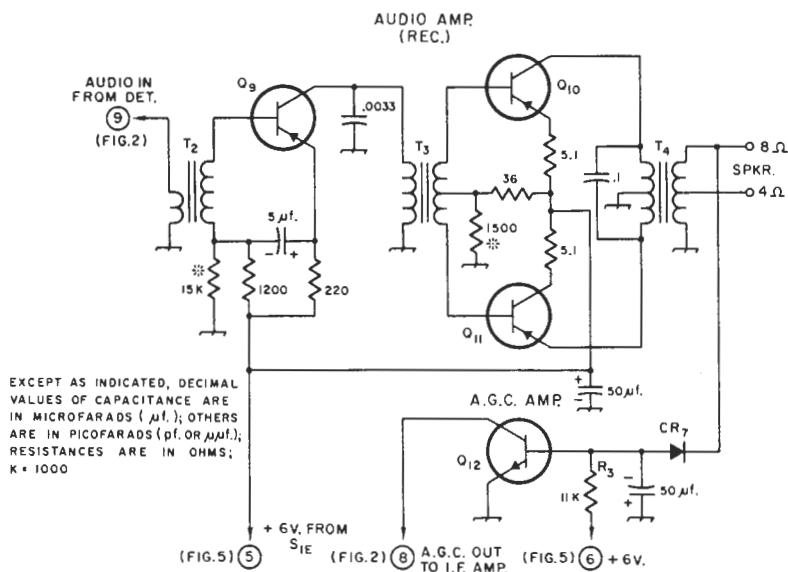


FIG. 4—Receiving audio circuit. An a.g.c. signal is obtained by rectifying and amplifying a signal taken from the audio output. Capacitors of decimal value are miniature ceramic or paper. Others are electrolytic. Both types have a minimum rating of 6 volts. Resistors are $\frac{1}{4}$ watt. Bias resistors are identified by asterisks; see text.

CR₇—Silicon junction diode, 50 p.i.v., 1N599 or equivalent.

Q₉—2N653 or similar.

Q₁₀, Q₁₁—2N586 or similar.

Q₁₂—Same as Q₈.

R₃—See text.

T₂—Same as T₁; low-impedance winding in output circuit.

T₃—Subminiature interstage transformer, 4:1 turns ratio, secondary center-tapped.

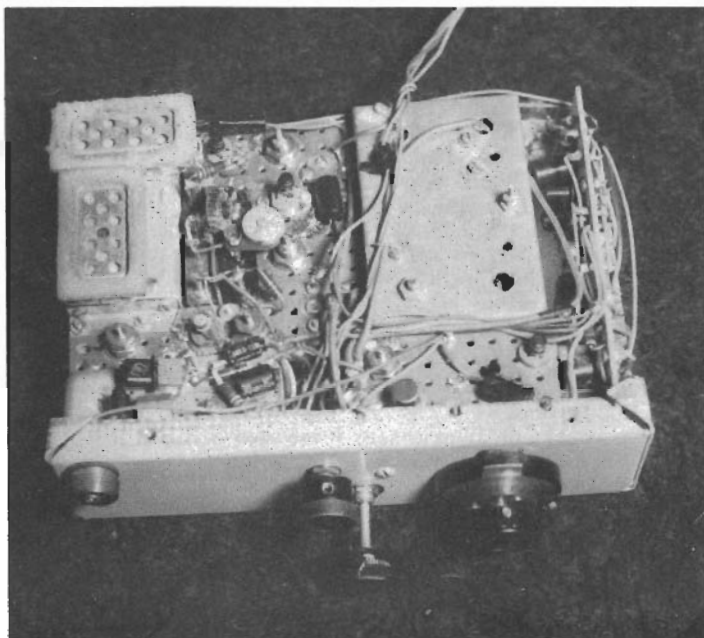
T₄—Transistor output transformer, 400 ohms, c.t., to 8.4 ohms, tapped at 4 ohms (Thordarson TR-22).

very low collector-saturation resistance, it has considerable peak-current capability and makes an excellent s.s.b. linear amplifier.

Constructional Details

The general layout of components is shown in the photographs and the sketch of Fig. 6.

Top view of the transceiver. The two rectangular speakers in the upper left-hand corner are flat dynamic microphones taken from junked hearing aids. The microphone-amplifier board is immediately to the rear of the microphone connector. It is mounted on top of the can containing the carrier oscillator. The basic chassis is a standard 5 \times 7 \times 2-inch unit with back apron sawed off.



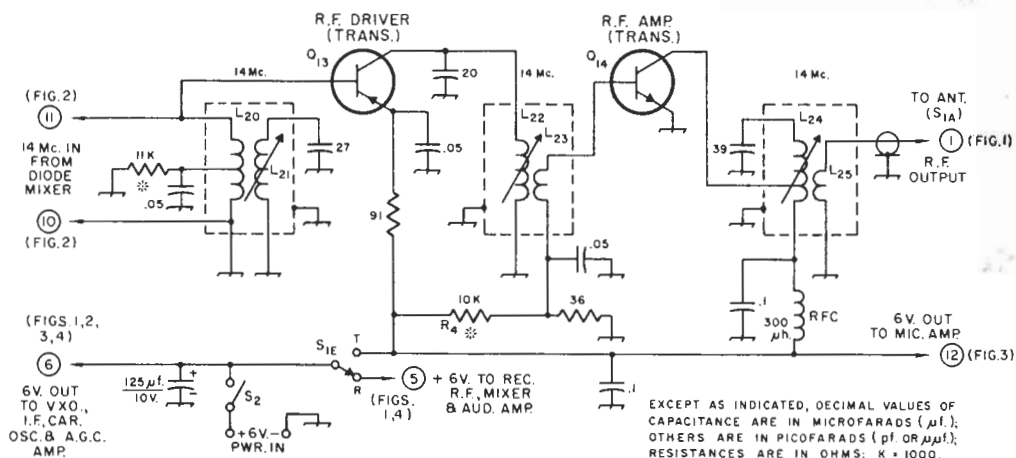


FIG. 5—Transmitter output circuits. This section receives 14-Mc. drive from the diode balanced mixer of Fig. 2. Capacitors of decimal value are miniature ceramic; others are NPO ceramic or dipped silver mica, except polarity indicates electrolytic. Capacitors have a minimum rating of 6 volts. Resistors are $\frac{1}{4}$ watt; asterisk indicates bias resistor; see text.

L20—4 bifilar turns, center-tapped, wound over L21 and connected as described for L5.

L21—21 turns on CTC LS-9-5S iron-slug form.

L22—21 turns on CTC LS-9-4S iron-slug form.

L23—6 turns wound over ground end of L22.

L24—16 turns on CTC LS-9-5S iron-slug form, tapped at 4 turns from low-potential end.

L25—6 turns wound over low-potential end of L24.

Note: Above coils are close-wound with No. 34 enameled wire.

Q13—Same as Q1.

Q14—Same as Q3.

R4—Nominal value, see text.

S1—See Fig. 1.

S2—S.p.s.t. slide or toggle switch.

As already noted, the miniature LS-9 coil forms are used wherever practical, with fixed miniature mica capacitors added for resonant tanks. Additional shielding is provided by using the two Command-set i.f. cans for critical circuits, and by enclosing the complete VXO in the smallest-size Minibox. The remainder of the r.f. circuits are mounted on subchassis made of copper-clad perforated boards. Since many of the components connect to ground, they can be soldered directly to the board, providing a good

low-inductance path. These boards are very easy to work with and simplify construction and assembly considerably.

The filter crystals squeeze in between the r.f. amplifier can and an under-chassis shield, and are held in place with a drop of glue. C7 is a Centralab type 827 ceramic trimmer capacitor. When Y2 and Y3 are placed end to end, the mounting-hole spacing of the capacitor matches the spacing of adjacent pins of the two crystals. The capacitor is slipped over the crystal pins for support. L13 is glued in place close to the capacitor. Other parts which are too heavy to be supported by their leads are glued in place.

To make assembly and disassembly possible with the crowded chassis, a number of captive nuts were used, fastened to the chassis and mounting brackets with epoxy (a two-tube mixture is now available in most hardware stores). The cover was made from perforated aluminum sheet with the corners folded over and epoxied together. After filing the corners smooth, several coats of spray paint were added to give a fairly professional-looking cover.

Components

Up to now, we have ignored the types of transistors used. The audio-transistor choices were made straight from my particular junk box. If your junk box is emptier, the Japanese units with matching transformers are an excellent choice. For Q1, Q2, Q5, Q6 and Q13, I

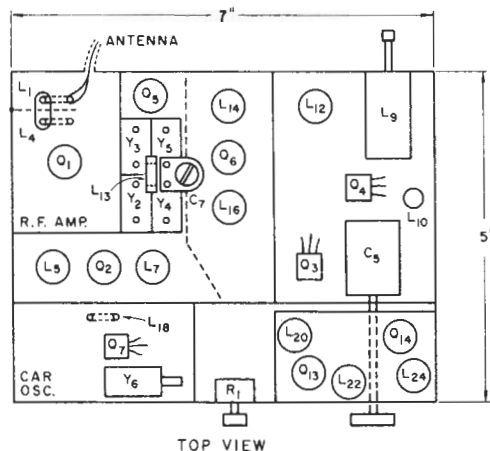
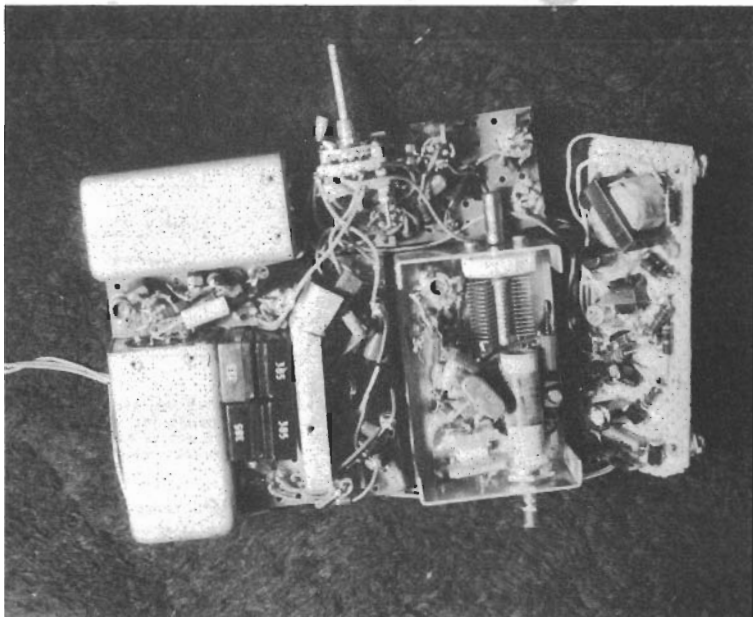


FIG. 6—Sketch showing layout of principal components.



Bottom view of the transceiver. The VXO is in the Minibox to the right of center. The carrier oscillator and balanced modulator are in one of the shielding cans to the left. The other, adjacent to the four crystals of the i.f. filter, contains the receiver r.f. stage. The receiver audio section is assembled on the perforated board to the right.

used some available 2N700s. The 2N1742 will serve in these circuits with essentially the same performance, and is somewhat cheaper. For Q_3 , Q_4 , Q_7 and Q_{14} , I used 2N706s. Actually, Q_{14} is the only stage that requires a silicon transistor of this quality. Any of the u.h.f. transistors will serve for Q_3 , Q_4 , and Q_7 . In fact, if these circuits are adapted for p.n.p. transistors, the 2N1742 will work fine.

Regardless of what transistor is used for each stage, it is wise to adapt each stage's bias resistor (all bias resistors are identified with asterisks) to the particular unit to give a collector current equal to that recommended for the transistor used. The 2N1742, for example, will require a bias resistor of considerably higher value.

Actually, there is little to be gained by neutralizing the receiving r.f. amplifier if a 2N700 is used. However, if a lower-frequency transistor is used, neutralizing may yield a sizable increase in gain.

The electrolytic capacitors used throughout were obtained from the C-923 assortment and the ceramic bypasses from the AS-510 assortment, both from Olson Radio. These are good-quality Japanese parts and quite cheap.

The transmit-receive switch is the latest Centralab subminiature wafer switch.

Other Bands

It is pretty obvious that by rewinding a few coils and using different VXO crystals, you can adapt the unit for operation on other bands. The carrier oscillator, i.f. circuits, and audio

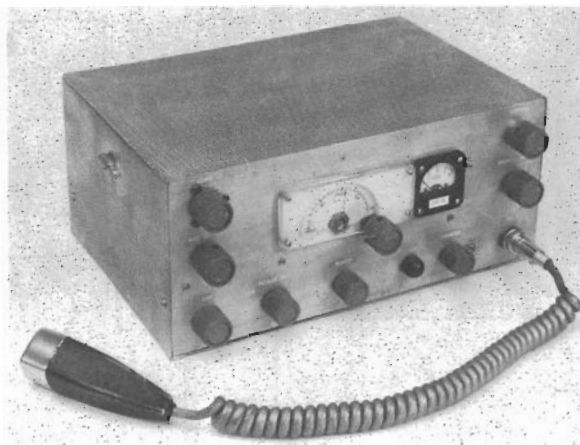
circuits remain as is. The VXO frequency should be chosen to be 8.55 Mc. above the highest frequency to be covered so as to get maximum tuning range. Of course, L_{10} and L_{12} must be rewound to resonate at the new VXO frequency. Similarly, L_1 , L_4 , L_5 , L_{21} , L_{22} , and L_{24} must be rewound to resonate on the new band. The same approximate turns ratios should be used for each transformer.

Results

With all such rigs, some mention of results is in order. First, the receiver is stable and selective, but is not very good as far as intermodulation is concerned. This is one respect in which transistors are inferior to tubes. The audio quality is limited by the small speaker. (I always demonstrate it with an external speaker.) On transmit, the carrier suppression is 45-50 db., and the other sideband is about 40 db. down. Local reports indicate that the signal is clear and clean. Since the "final" runs a puny 250 mw. p.e.p., I haven't worked much DX (Florida, Louisiana, Nebraska, and similar), but it is a dandy "local" rig. It's been running off a 6-volt lantern battery for several months now. It has made several trips cross country in my briefcase, and provided an excellent way to keep in touch with what the 20-meter s.s.b. gang is up to.

Of course, I intend to add a linear amplifier to bring the transmitter up to a reasonable output, but haven't definitely decided yet whether to succumb to tubes or wait for the price break on high-power h.f. transistors.

» *The 75-meter transceiver described here is relatively simple and inexpensive to build and adjust. It has been duplicated by several hams in the Delta area with marked success.*



This model of W5TAB's 75-meter transceiver was constructed by W5RQJ. The dial is homemade, but may be replaced by a conventional type. At the left-hand end of the panel are the loading and tuning controls of the pi-network output circuit; at the right-hand end are controls for receiver and transmitter audio. Along the bottom, from left to right are receiver r.f. trimmer, mobile power relay switch, buffer tuning control, and modulator balance control.

A 50-Watt PEP Output Transceiver for 75

KENNER E. DAY, W5TAB

The increasing popularity of transceiver operation, and the cost of commercial gear now on the market, prompted the author to design a single-band unit (75 meters) that could be built by the average ham with *some* building experience, yet would not require intricate alignment procedures, or elaborate test equipment not usually available to amateurs. Other design targets were the use of readily-available components, and lowest possible cost commensurate with adequate performance in both fixed-station and mobile operation.

Although some of the circuitry to be discussed is unorthodox, on-the-air testing of several of these transceivers has proved that the design is thoroughly practical. Commonly-available tube types are used, and a simple sideband filter using surplus crystals provides good suppression in the transit mode, and a fairly narrow passband for reception. Only four crystals are needed in all.

Circuit Principles

The complete wiring diagram of the transmitter (save for the heater wiring shown in Fig. 2) appears in Fig. 1. Several features were taken from an earlier *QST* article by K5BUQ,¹ so any similarity is *not* coincidental. When receiving, the incoming 4-Mc. signal is amplified in the 6AU6A r.f. stage, and then combined in the 6BE6 receiver mixer with a 3545-kc. signal from the v.f.o. to produce a signal at 455 kc. in the output of the mixer. This signal is fed through the selective crystal filter (Y_2Y_3) to a single i.f. stage using another 6AU6A. The amplified 455-kc. signal is coupled into the 1N34A diode detector where it is combined with the signal from the 455-kc. crystal-controlled 6C4 b.f.o. to produce audio output. The audio signal is amplified in the triode section of the 6EB8, and brought up to speaker level in the pentode section of the same tube. The r.f. gain control, R_1 , which is applied to the r.f. and i.f. stages, provides smooth control of audio output, so a separate audio gain control was not deemed necessary.

¹ From June, 1967, *QST*.

² Taylor, "A 75-Meter S.S.B. Transceiver," *QST*, April, 1961.

FIG. 1—Circuit of the 75-meter transceiver. Fixed capacitors of decimal value, unless listed below or indicated otherwise in the diagram, are disk ceramic and, unless indicated otherwise, are 500-volt. Others are silver mica or NPO ceramic 500-volt, except where polarity indicates electrolytic. Unless indicated otherwise, resistors are $\frac{1}{2}$ -watt.

- C_1 —"Gimmick" capacitor made by twisting two short lengths of insulated wire together.
 C_2 —Air trimmer (Hammarlund APC-50-B).
 C_3 —Mylar capacitor.
 C_4 —Midrange air variable (Johnson 167-11, or similar).
 C_5 —Miniature air trimmer (Johnson 160-110).
 C_6 —Air trimmer (Hammarlund APC-25-B, or similar).
 C_7 —Air variable (Millen 19325, or similar; see text).
 C_8 —Three-section broadcast-band t.r.f. variable capacitor, sections in parallel.
 C_9 —7-45-pf. ceramic trimmer.
 J_1 —Three-circuit microphone connector.
 J_2 —Chassis-mounting coaxial receptacle.
 J_3 —8-contact chassis-mounting male connector (Cinch-Jones).
 K_1 —Four-pole double-throw relay, 110 volts, d.c. (Potler & Brumfield KL17D, or similar).
 L_1 —10 turns No. 30 enameled, wound over ground end of L_2 .
 L_2 , L_3 , L_4 , L_5 , L_6 —35 turns No. 30 enameled, wound on $\frac{3}{8}$ -inch ceramic iron-slug form.
 L_7 —28 turns No. 26 enameled on $\frac{3}{8}$ -inch ceramic iron-slug form, wound tightly and doped.
 L_8 —24 turns No. 22 enameled on $\frac{7}{8}$ -inch ceramic form (surplus form).
 R_1 —Audio-taper control.
 R_2 , R_3 , R_4 —Linear control.
 S_1 —D.p.s.t. rotary switch (see Fig. 2 for second section).
 T_1 —Interstage audio transformer, single plate to p.p. grids (Stancor A-63-C).
 T_2 , T_3 —Miniature 455-kc. i.f. input transformer (Miller 12-C1).
 T_4 —Miniature 455-kc. i.f. output transformer (Miller 12-C2).
 T_5 —Audio output transformer, 5000 ohms to voice coil.
 Y_1 , Y_2 , Y_3 —455-kc. crystal (see text).
 Y_4 —453.148-kc. crystal (see text).

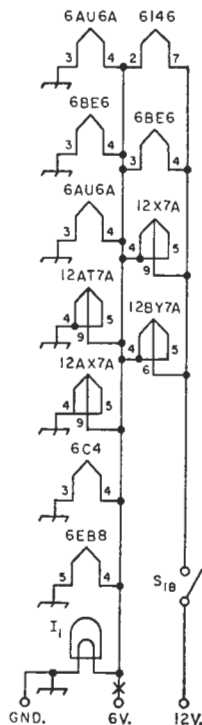
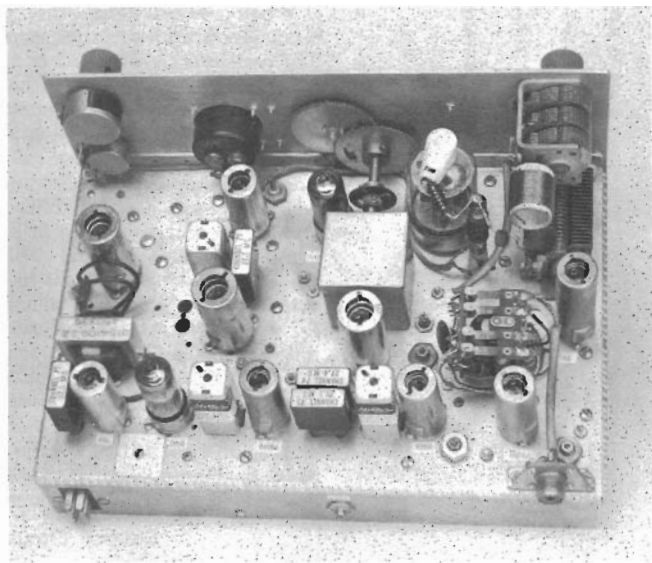


FIG. 2—Heater wiring diagram for either 6- or 12-volt operation. I_1 is a No. 47 6.3-volt 0.15-ampere pilot bulb. For 12-volt operation, the 12-volt terminal should be connected to Pin 4 of the plug for J_3 , Fig. 1, Gnd. to Pin 6, no connection to the 6-volt terminal. For 6-volt operation, S_{1B} should be transferred to the 6-volt line at X, the 6-volt terminal should be connected to Pin 5, the 12-volt terminal and Gnd. to Pin 6.

When transmitting, the crystal-controlled b.f.o. serves as the carrier generator at 455 kc. The oscillator signal is fed to a balanced modulator using a 12AT7. When the output circuit is adjusted for balance by potentiometer, R_2 , the carrier is suppressed. The application of audio from the speech amplifier results in a double-sideband suppressed-carrier signal at 455 kc. which is fed to a crystal filter consisting of T_5 , Y_4 , T_2 , Y_2 and Y_3 . (The 6BE6 receiver mixer is not active on transmit.) The filter attenuates the upper sideband by 20 to 30 db. The remaining lower-sideband signal is amplified in the i.f. stage, and passed along to the transmitter mixer, a 6CS6. Here it is combined with the 3545-kc. signal from the v.f.o. to produce mixer output at 4 Mc.—the same frequency as the receiving section. The 4-Mc. l.s.b. signal is amplified in the 12BY7A stage which drives the 6146 final amplifier. A pi-section output circuit provides a match to a low-impedance load.

Control Circuit

Reviewing the foregoing, it will be seen that three stages are common to the receiving and transmitting sections. These are the v.f.o., the b.f.o./carrier oscillator, and the i.f. amplifier with its crystal filter. Other stages are switched in and out, as necessary, by the four-pole double-throw relay, K_1 , which also switches the antenna. On receive, 250 volts is applied to the r.f. amplifier, r.f. gain control, receiving mixer, detector, and receiving audio. (The mixer is switched in the cathode circuit by a separate relay pole to avoid diode mixing in the



Lined up along the rear edge of the chassis, from right to left, are the 6AU6 receiver r.f.-amplifier tube, 6BE6 receiver mixer tube, T_2 , Y_2 and Y_3 , the 6AU6A i.f. tube, T_3 , the 6EB8 receiving-audio tube, 6C4 carried/b.f.o., and Y_1 . L_1/L_2 is to the right of output connector, and L_3 to the left. The 12AT7 v.f.o. tube is immediately to the rear of the box shielding the v.f.o. tuning capacitor, C_4 . The adjusting screw of L_4 and the screwdriver shaft of trimmer C_5 are discernible to the right of the 12AT7. The tube above and to the left of the box is the 12BY7A driver. L_6 is mounted between the box and the panel, and cannot be seen. Grouped at the upper left are the 12AX7 transmitting audio tube (above T_4), the 6CS6 transmitting mixer (below the meter), T_5 and Y_4 , and the 12AT7 balanced-modulator tube. L_5 is to the right of the meter. At the upper right are the 6146 and components of the pi network. L_7 is mounted on the output capacitor C_6 . The tube to the right of the change-over relay is the 0B2 regulator. On the rear apron are the power connector, J_3 , and the shaft of the bias control, R_4 .

receiver mixer while transmitting.) In addition, another pole of the relay disconnects the two 8- μ f. bypass capacitors in the speech amplifier. This was found to be necessary to avoid audio oscillation in the speech amplifier which occurred as the capacitor discharged after removal of voltage from the amplifier when switching from transmit back to receive.

When transmitting, voltage is removed from the stages mentioned above, and applied to the balanced modulator, the speech amplifier, transmitter mixer, and driver stage. The cathode resistor of the i.f. amplifier is switched to ground to remove it from the influence of the r.f. gain control and place it at full gain on transmit. (This switching also grounds the cathode resistor of the r.f. stage, of course, but since plate voltage has been removed from this stage, complications that might arise from this source are avoided.) Power to the final is not switched.

One side of the relay coil is connected to the 250-volt line through a 10,000-ohm series resistor. The coil circuit is completed to ground through the p.t.t. switch at the microphone.

The V.F.O.

A variation of the Vackar circuit, first noted in *QST* several years ago, is used in this important part of the transceiver. This circuit is easily adjusted, and provides constant output and adequate drive through very small coupling capaci-

tances, with a plate voltage of only 108 volts. This voltage is regulated by an 0B2 fed from the 250-volt supply through a 7000-ohm resistor. One section of a 12AT7 (V_{3A}) is used in the oscillator, while the other section (V_{3B}) is in a cathode follower driven by the oscillator. The latter serves to isolate the v.f.o. from the two mixers which it feeds. With this configuration, frequency shift is a matter of only a few cycles, comparing very favorably in this respect to commercial gear. No v.f.o. temperature compensation is included; drift is nominal after a thorough warm-up.

Crystals

By now, the reader may be wondering about the crystals. The author used surplus crystals in the 455-kc. range. The low-numbered FT-241 crystals, from Channel 38 to about Channel 75, are in a range that can be tuned to with ordinary $\frac{3}{8}$ -inch 455-kc. i.f. transformers. Two Channel 45 crystals (Y_2 and Y_4), and one Channel 44 crystal (Y_3) are used in the filter. These crystals are fairly close to 455 Kc. and Y_3 differs from the other two by about 1852 cycles. Using a Channel 45 crystal at Y_1 in the b.f.o./carrier oscillator, and tuning as described presently, lower-sideband output will be produced. For those unable to obtain the surplus crystals, Texas Crystals, Fort Meyer, Florida, advertises crystals in the 455-kc. range, 25-

cycle tolerance, in FT-241 holders for \$1.75 each. Three crystals of the same frequency are needed—two for the filter and one for the b.f.o. The additional crystal for the filter should be approximately 1800 to 2000 cycles lower in frequency.

Construction and Adjustment

The author has constructed several of these transceivers, and they have taken various physical forms. However, an $8 \times 12 \times 2$ -inch chassis provides enough space to avoid crowding of components if the layout shown in the photos is followed reasonably closely. A panel $5\frac{1}{2}$ inches high will provide clearance for the 6146 without submounting the socket. The dial is home-brew. The gearing was salvaged from old Command-set mechanisms. However, a National 5-to-1-ratio planetary-drive dial, or any similar conventional dial, may be used.

Careful orientation of the tube sockets will furnish convenient tie points for resistors and bypass capacitors and hold wiring between stages to a minimum. Low-potential wiring can be run around the edges of the chassis in bends and corners for neater appearance. As indicated in the diagram, shielded wire should be used for the connections to the microphone jack and gain control in the speech amplifier, for the balanced-modulator output connection, and in the coupling line between the i.f. amplifier and the transmitter mixer. Shielded wire is also preferable for heater circuits and other low-potential wiring.

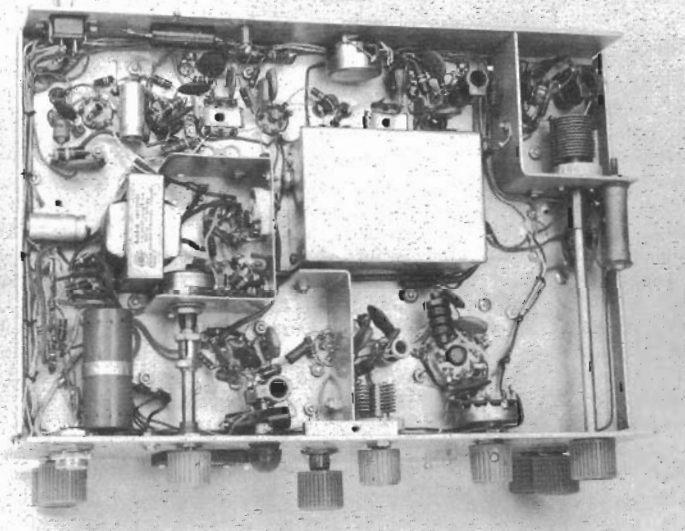
The transceiver can be built a stage or section at a time, testing each as it is completed. It is suggested that the v.f.o. be constructed first, using short leads. The tuning capacitor, C_4 , is placed above the chassis in a shielding box, with a connecting wire running through a small hole to the coil, which is enclosed in a second shielding box on the underside of the

chassis. Coil turns may have to be pruned, and capacitance juggled, to achieve the proper 200-kc. tuning range for the v.f.o. Assuming that the carrier-oscillator crystal is for Channel 45 (about 455 kc.), the upper limit of the v.f.o. range would be 3545 kc. to tune the transceiver to 4000 kc. The lower end of the range would be 3345 kc., to tune the transceiver to 3800 kc. Keeping the v.f.o. frequency on the lower side of the incoming signal seems to result in less drift than when the v.f.o. is tuned to the upper side. Listening on a receiver while adjusting the v.f.o. will assist the builder in getting the circuit into the proper tuning range.

After the v.f.o. is working, the receiver section can be constructed. To align the i.f. amplifier stage, couple output from a modulated signal generator to the receiver mixer stage with all four crystals in place. Tune the signal generator exactly to the frequency of the b.f.o. crystal. Remove this crystal, and peak i.f. transformers T_2 and T_3 for maximum audio output. Replace the b.f.o. crystal. Final alignment of the crystal-filter and i.f. stage can be done after construction of the transmitter stages.

Now peak the receiver mixer cell, L_3 , at 3900 kc. (A grid-dip oscillator will be helpful in rough tuning of circuits in the transmitter as well as in the receiver section.) The r.f. stage is rough-tuned by the slug of L_1L_2 , and the circuit is peaked by the 50-pf. trimmer, C_2 , which should be mounted on the panel.

After the receiving section is working, the transmitter section should be checked out. Peak the transmitter mixer coil, L_5 , at 3900 kc. In operation, the output of the mixer will fall off some at either end of the band, but should still be adequate for full drive to the final. Peak L_6 at 3900 kc. with C_6 set at mid capacitance. It will be noticed that part of the tuning capacitance in this stage is fixed to confine the tuning range to the vicinity of 4 Mc., thus



Grouped at left center are T_1 , R_2 and C_9 , the latter two mounted on a shielding bracket. L_5 is below the bracket. To the right is the box shielding the coil and other components of the v.f.o./cathode follower. Below the box are L_6 and driver tuning capacitor, C_6 . L_1/L_2 and C_2 are in the shielding compartment in the upper right-hand corner, and L_3 is to the immediate left.

avoiding the possibility of tuning to some other response in the output of the mixer. Those more mechanically able could gang-tune the mixer and driver stages by adding a small variable capacitor across the mixer coil, and coupling its shaft to that of the driver tuning capacitor, C_6 , to obtain full output across the band.

No special constructional precautions are necessary in the driver and final stages, except that a shield should be placed across the 12BY7A socket. Pins 3 and 9 of this tube are grounded, and the shield can be placed across these two pins when the socket is properly oriented on the chassis. The relay should be mounted on the chassis reasonably close to the pi-network components, since one pole of the relay switches the antenna.

The biasing control, R_4 should be set for a final-amplifier idling current of 25 to 30 ma.

If the transceiver has been constructed in sections, as suggested, proper alignment of the filter system, consisting of the three filter crystals and three i.f. transformers, can now best be done by feeding a sine-wave audio signal at low level, 1000 to 2000 cycles, into the microphone input, and observing the output wave form on a scope. This test set-up is covered under, "Testing an S.S.B. Exciter," in the A.R.R.L. *Radio Amateur's Handbook*. A little careful twisting on the i.f. transformer slugs will produce the proper pattern on the scope, indicating when the pass band of the filter is adjusted for maximum suppression of the unwanted sideband, and the carrier. Additional

information on filter alignment will be found in chapter 6.

In actual operation, transmitter adjustment is very simple. Press the push-to-talk switch. Set the v.f.o. to frequency, turn the carrier-balance control to one side, tune the final for maximum output, then adjust R_2 and C_9 for minimum final-amplifier idling current. If C_9 has no effect when connected to one plate of the 12AT7, it should be transferred to the other plate. That's all there is to it. I use a field-strength meter when tuning the final, but the plate-current dip is a fairly satisfactory indicator. During adjustment with the scope, the proper setting of the gain control to prevent overdrive and splatter should be determined.

Power Supply

For home-station operation, the author uses a small bench power supply delivering 600 volts at 150 ma., 250 volts at 75 ma., and 100 volts of bias. The Heath HP-10 supply is used for mobile work. The heater wiring diagram of Fig. 2 provides for either 6- or 12-volt operation.

Careful placement of small parts and wiring should enable the builder to construct a transceiver that will look neat, and perform satisfactorily in both fixed and mobile service. In mobile operation, good communication over distances from 100 to 300 miles is not uncommon. For points not covered here, it is suggested that prospective builders refer to the article by K5BUQ.

MEASURING SIDEBAND SUPPRESSION

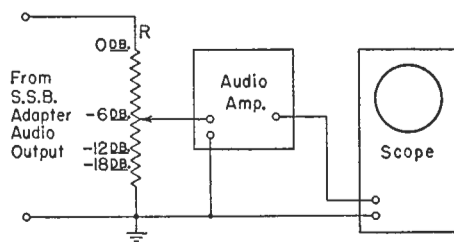
By calibrating the volume control (R) of a selectable-sideband adapter, it is easy to get direct readings of sideband attenuation.

Howard Wright, W1PNB, suggests this simple stunt for measuring sideband suppression of your own or the other fellow's signal. It requires that you have a selectable-sideband receiver or adapter of some kind and an oscilloscope, and the only other requirement is a calibrated volume control.

The volume control is calibrated with an ohmmeter. Assuming a 1-megohm volume control, the -6 db. point will be at half resistance or 0.5 megohm. The -12 db. point will be at half of this, or 0.25 megohm. The -18 db. point is half of this (0.125 megohm), and so on down the line in 6-db. steps.

The scope can be connected at any point in the audio amplifier following the calibrated volume control, and the sweep speed should be set low enough to make noise peaks appear as individual "spikes."

To measure sideband suppression, set the

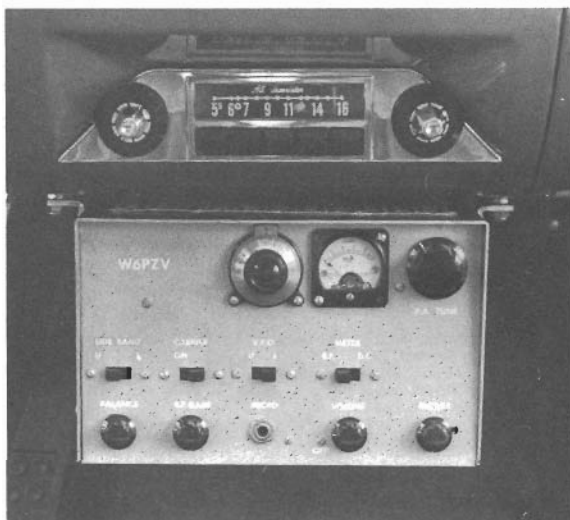


control at 0 db. and advance the r.f. gain control of the receiver to a point where the unwanted sideband gives a definite amount of scope deflection on peaks. Reduce the calibrated volume control setting (to save the loudspeaker) and switch to the desired sideband. Adjust the calibrated control until exactly the same amplitude peaks appear, and read the suppression ratio directly from control.

The accuracy of the system is limited, of course, by the care taken in calibrating the control, errors in reading the scope, and by the maximum possible amount of sideband suppression the selectable-sideband receiver is capable of. However, the limit of the receiver rejection can readily be found by occasionally tuning across an unmodulated carrier and measuring the point of receiver failure. Any reports given below this ratio will be accurate.

"The one-band transceiver has been a popular type of mobile station, since many—if not most—mobileers are one-band operators. The 7-Mc. transceiver described here has many circuit features that could be incorporated in equipment for other bands.

The transceiver installed in the author's car. The VXO and final-amplifier tuning controls flank the meter near the top of the panel. The slide switches, left to right, are for upper/lower sideband, carrier on/off, VXO frequency range, and meter. Controls along the bottom are for modulator balance, receive r.f. gain, receive audio gain, and final-amplifier drive. The microphone jack is at the center in this row. Some idea of the size of the unit can be gained by comparing it with the car broadcast receiver above. The output is 60 watts p.e.p.



A 7-MHz Mobile SSB Transceiver

JOHN ISAACS, W6PZV

By now there doesn't seem to be much doubt that s.s.b. is *the* way to go for mobile operation. The contemplated purchase of a new car finally triggered the decision to get going on a new mobile sideband rig. At that time, the only manufactured rigs that I felt were reasonably priced were also larger than the available space would accommodate. With size the main consideration, a number of other things were automatically resolved. The new rig would have to be a transceiver and it would have to operate on one band only. Also, it might be necessary to compromise on carrier suppression, unwanted-sideband suppression and power output. However, there could be no compromise on frequency stability.

After trying mobile s.s.b. operation with a separate transmitter and receiver, anything but a transceiver was out. Also, after operating mobile for over 15 years, it was found that although multiband operation has been available at all times, my operation has been confined to one band about 99 per cent of the time. It wasn't difficult to settle for operation on 40 meters. Daytime operation on 75 meters is somewhat of a problem, and 20 meters is in and out. I had just completed a multiband

exciter¹ using the McCoy filter, and a VXO looked like a good way to get needed stability in a small space. Every consideration was given to keeping the number of tubes to a minimum. Multipurpose tubes helped, but 10 tubes plus two VR tubes were finally needed to do the job.

Circuit Functions

Before going into detail, it might be well to describe the over-all scheme of the transceiver. Referring first to Fig. 1, a 6AR8 is used as a combination 9-Mc. crystal-controlled carrier oscillator and balanced modulator. The 6AR8 is similar to the 7360 but is less expensive. The 9-Mc. signal is fed to the No. 1 grid, while audio is fed to one of the deflectors. On transmit, the 9-Mc. d.s.b. signal from the balanced modulator is fed to the crystal filter of Fig. 2, which suppresses the unwanted sideband and also contributes further suppression of the carrier. The s.s.b. signal from the filter is amplified in the first of two 9-Mc. amplifiers and then fed to the 6BA7 transmitter mixer. Here the 9-Mc. signal is mixed with the 16.2-Mc. signal from the VXO to produce 7.2-Mc.

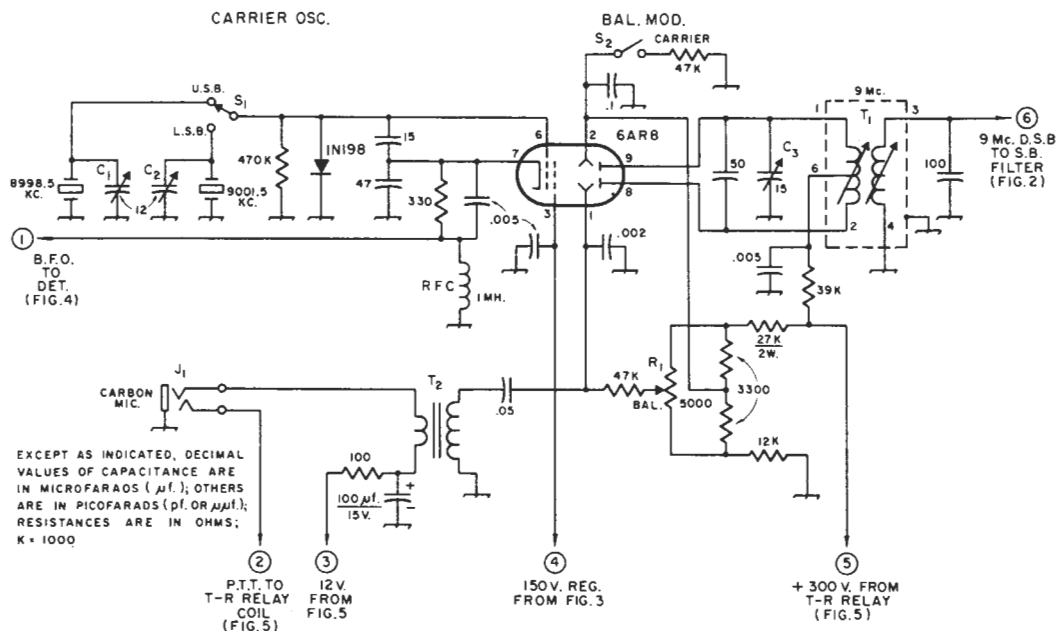


FIG. 1—Carrier-oscillator and balanced-modulator circuits. Fixed capacitors of decimal value are ceramic; others are silver mica, except polarity marking indicates electrolytic. Unless otherwise specified, resistors are $\frac{1}{2}$ -watt composition. See Fig. 5 for heater connections.

C₁, C₂—3–12-pf. ceramic, zero temp. coefficient (Centralab 822-FZ).

C₃—Air trimmer (Johnson 160–107, or similar).

J₁—Three-circuit jack.

R₁—Linear control.

S₁—S.p.d.t. slide switch (Carling S60B or equivalent).

S₂—S.p.s.t., or same as S₁.

T₁—10.7-Mc. f.m. discriminator transformer; see text (Meissner 17–3494).

T₂—Microphone transformer, 200 to 500K (UTC "Ouncer" 0–14).

output. This 7.2-Mc. output is fed to the 12BY7 driver of Fig. 3 and thence to the 6DQ5 final amplifier.

On receive, an incoming 7.2-Mc. signal is amplified in the pentode section of a 6AX8 (Fig. 2), and then mixed in the triode section of the same tube with the 16.2-Mc. signal from the VXO. The resulting 9-Mc. output signal from the receiver mixer is fed through the crystal filter and the following two 9-Mc. amplifiers to the detector and audio system of Fig. 4. The balanced modulator is disabled on receive, but the carrier oscillator remains in operation to furnish a b.f.o. signal to the detector. In either mode, inactive stages are disabled by removing plate voltage. The transmit-receive switch takes care of this.

Carrier Oscillator and Balanced Modulator

Returning to Fig. 1, two crystals are used in the crystal oscillator so that the carrier may be shifted to place either upper or lower sideband in the passband of the filter. These crystals are normally supplied with the McCoy filter as a package. The two frequencies may be trimmed to the proper spot in relation to the slope of the filter characteristic by means

of the trimmer capacitors (C₁ and C₂) shunting the crystals.

The balancing and output-coupling circuits of the balanced modulator are slightly different, and the values used also deviate somewhat from those normally shown for the 6AR8 or 7360 in this application. More trouble was encountered in this portion of the circuit than in any other. Transformer T₁ was originally a standard 10.7-Mc. i.f. transformer. The plates of the 6AR8 were shunt-fed through 68K resistors, and the plates coupled to the primary of T₁ through 0.001- μ f. capacitors. With this arrangement, the output from the crystal filter was very low, and the carrier suppression was poor. (The transformer was never designed for this application, so it can't be blamed.) A 10.7-Mc. discriminator transformer was substituted. The secondary is bifilar-wound, which is one point in its favor. All internal capacitors were removed, and the secondary was used as the center-tapped primary. With this revision, the results were much improved. The carrier suppression was better, and the output from the crystal filter was more than adequate. An r.f. probe, connected to the output of the 12BY7 driver, was used to measure the carrier

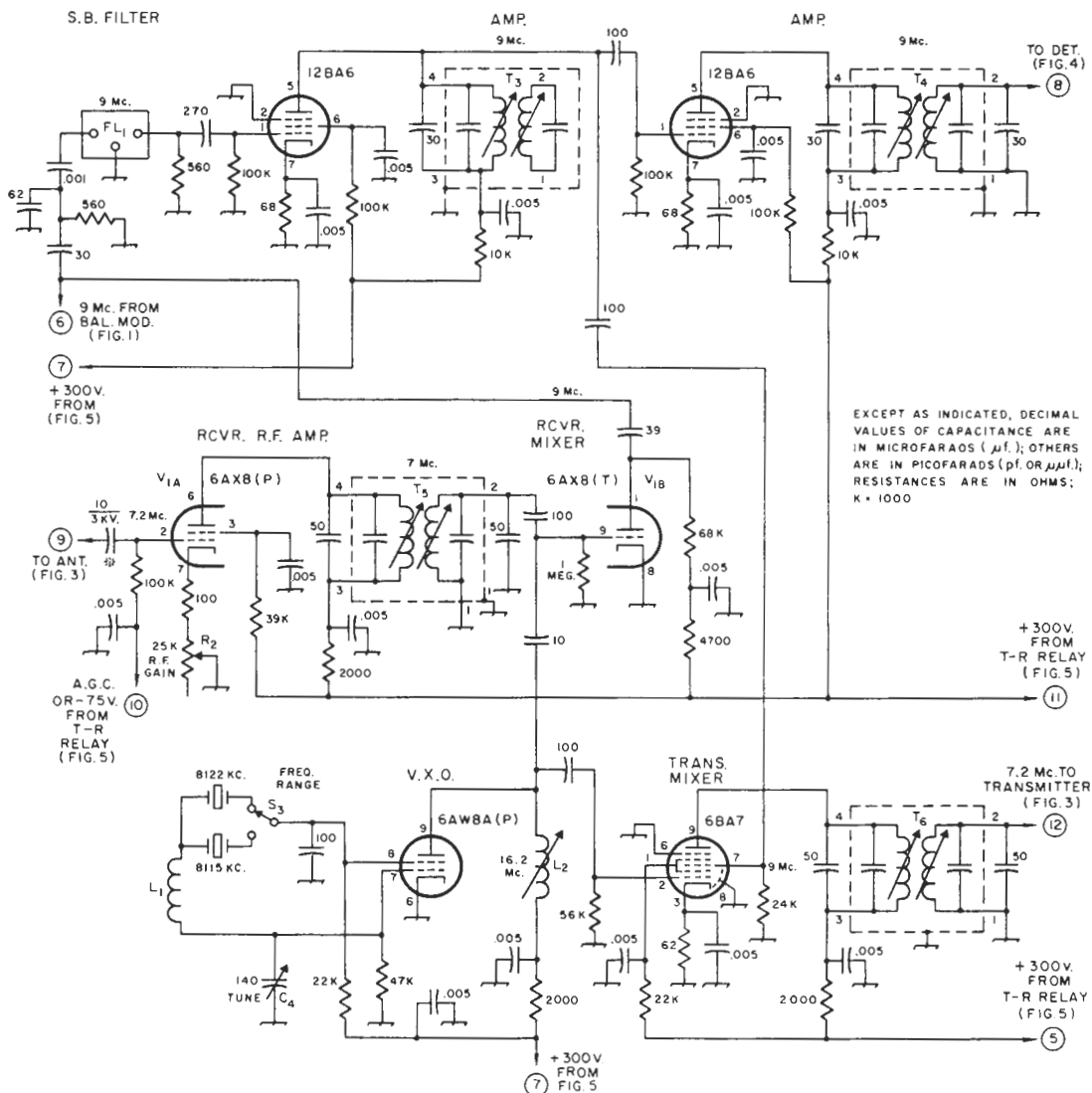


FIG. 2—i.f. and frequency-conversion circuits. Fixed capacitors of decimal value are ceramic; others are silver mica, except * indicates ceramic. Resistors are $\frac{1}{2}$ -watt composition. See Fig. 5 for heater connections.

C₄—Air-trimmer variable (Hammarlund APC-140-B).
FL₁—9 Mc. crystal filter (McCoy 32B1).

L₁—20 μh .—Approx. 45 turns No. 30 enam., close-wound on form taken from National R-33
100 μh . r.f. choke. See text.

L₂—18 turns No. 22 enam., close-wound on $\frac{3}{8}$ -inch iron-slug form.

R₂—Linear control.

S₃—Same as S₁.

T₃, T₄, T₅, T₆—10.7-Mc. i.f. transformer (Miller 1463).

suppression. The measured suppression is about 40 db. below the peak output. The minimum reading apparently is limited by the various beats produced in the 6BA7 mixer. At any rate, the effective suppression is better than 40 db. judging from the ratio of maximum to minimum readings on a field-strength meter. I am inclined to believe that the formula sometimes used to calculate carrier suppression of

sideband rigs is the same one frequently employed in determining gas mileage.

Modulation is accomplished by applying audio voltage to one of the 6AR8 deflectors. The required a.c. voltage is something less than 10 volts, and this is easily obtained by using a carbon microphone and an input transformer. The voltage for the microphone is taken from the 12-volt circuit through an RC filter. I have

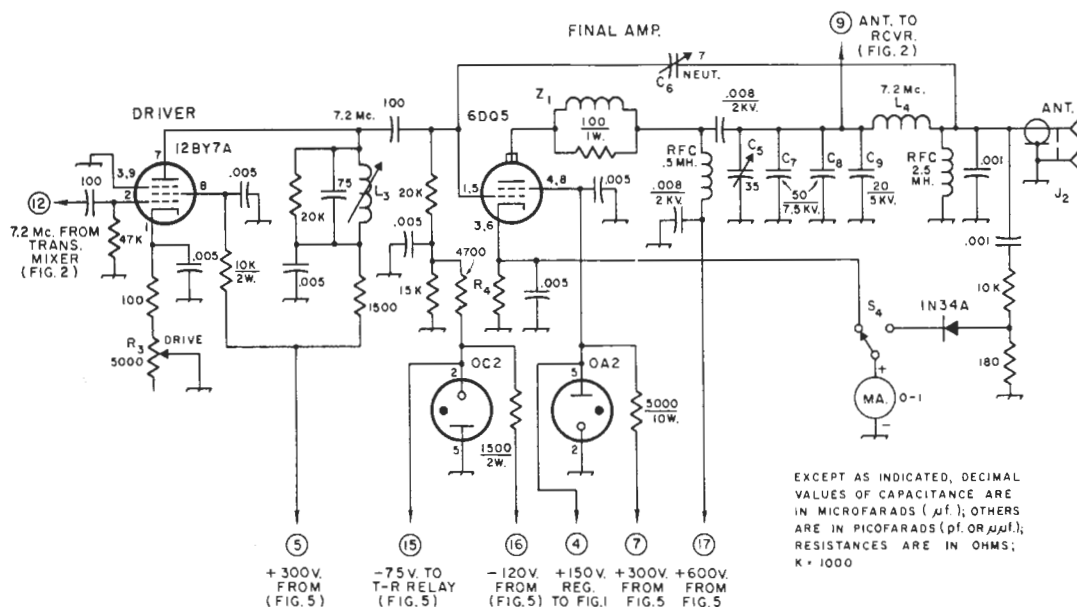


FIG. 3—Transmitting driver and final-amplifier circuits. Antenna connection to receiving r.f. amplifier is taken from the input side of the transmitting pi network. On transmit, the t.r. relay (K1 Fig. 5) applies -75 volts bias to the receiving r.f. amplifier to protect it from the transmitter signal. Capacitors of decimal value are ceramic; others are silver mica.

C5—Double-spaced midget variable (Hammarlund MC-35-SX or similar).

C6—NPO ceramic trimmer, 1.5–7 pf. (Centralab 822-EZ).

C7, C8—7500-volt ceramic (Centralab 850S-50Z).

C9—5000-volt ceramic (Centralab 853-20Z).

J2—Coaxial plug-in connector (UG1051/U).

L3—24 turns No. 26 enam. on $\frac{3}{8}$ -inch iron-slug form.
 L4—18 turns No. 20, $1\frac{1}{2}$ inches long on ceramic form $1\frac{1}{8}$ inches in diam.

R3—Wire-wound control.

R4—200 times meter shunt.

S1—Same as S1.

Z1—5 turns No. 18 wound on associated 100-ohm resistor.

always favored the carbon microphone for mobile operation. The cost is low and output is high enough to eliminate a preamplifier. There is never any r.f. feedback as is frequently encountered with low-output microphones and high-gain preamplifiers. In addition, the frequency response is designed for voice operation, and it doesn't make much sense to use a microphone which is flat from 100 cycles to over 10,000 cycles and then pass the output through a filter which limits the response to 300–3000 cycles.

Crystal Filter

The circuit diagram supplied with the filter shows variable capacitors connected across the input and output. These are to be adjusted for maximum output with the modulator unbalanced and using the 9001.5-ke. crystal. However, it was found that any capacitance added to either the input or the output of the filter caused the output signal to decrease. This is probably caused by the fact that the 62-pf. capacitor of the filter input impedance-matching network is connected across the input of the filter.

The VXO

The tuning of the transceiver is accomplished through the use of a VXO, or variable crystal oscillator. The circuit used (Fig. 2) has appeared in numerous sideband applications. The original circuit used a split-stator tuning capacitor, but experiment showed that by making one section fixed, about 80 per cent of the frequency shift could still be obtained with a single-section variable capacitor in place of the dual-section capacitor. Crystals at 8 Mc. are used, with the shift limited to about 9 kc. The output of the pentode section of the 6AW8 is tuned to 16.2 Mc. At this frequency, the shift is double, or about 18 kc.

Two crystals are used in overlapping ranges to cover from 7200 kc. to 7235 kc. The crystals used are regular ham-type units mounted in the small CR6U metal cans. The value of L_1 must be determined experimentally, and may vary depending on the crystal being used, although both of the crystals that I used provide about the same shift with the same coil. Connecting the inductance L_1 in series with the crystal causes its series-resonant frequency to

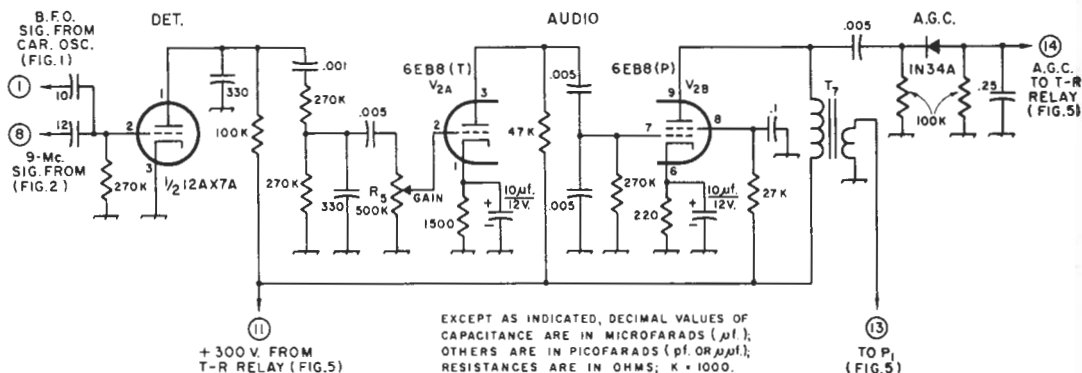


FIG. 4—Detector and receiving audio circuits. The b.f.o. signal is obtained from the carrier oscillator. Rectified audio supplies a.g.c. for the receiver r.f. stage. Capacitors of decimal value are ceramic or paper; others are mica, except polarity markings indicate electrolytic. Fixed resistors are $\frac{1}{2}$ -watt composition. See Fig. 5 for heater connections.

R₅—Audio taper control.

T₇—Output transformer, 10,000 ohms to 7 ohms (Thordorson 24S52).

be lowered. This means that the marked frequency of a crystal must be higher than the required highest output frequency. It was determined experimentally, using some crystals near 8 Mc., that the crystals should be ordered with a frequency about 6 to 7 kc. on the high side. Two crystals were then ordered, one at 8122 kc. and a second at 8115 kc. Using the circuit shown, these crystals actually cover 7202 to 7221 kc. and 7217 to 7235 kc. after mixing with the 9.0-Mc. signal. Nothing is guaranteed with respect to the crystals which you order. You're on your own.

L₁ is made by taking a 100- μ h. r.f. choke and first removing all of the wire. The core is then rewound with approximately 45 turns of No. 30 enameled wire. The actual number of turns is adjusted until the desired frequency shift is obtained. Using a grid-dip meter and a standard capacitor, the inductance was found to be approximately 20 μ h.

The frequency stability of the transceiver is excellent and the bandspread is very good, thus making tuning slow and easy.

Final Amplifier

A 6DQ5 is used in the final amplifier (Fig. 3). A pi network is used in the output and the tube is neutralized. The neutralizing circuit is a little unusual and is employed because of the physical problem of connecting a capacitor from the plate of the 6DQ5 to the bottom of the tank circuit of the 12BY7. The value of the output capacitor of the pi network is determined experimentally, and then the capacitor is soldered permanently into the circuit. This saves a separate loading capacitor and makes it possible to use the bridge neutralization circuit shown. This circuit would not work very well with a multiband amplifier, or one in which the output capacitor is variable.

The combination of C₅, C₇, C₈ and C₉ was chosen on the basis of compactness and availability. If space permits, the combination may be replaced by a single 150-pf. variable, or by other combinations of fixed and variable that will give an equivalent total. Any fixed capacitors used should be capable of carrying 3 or 4 amperes of r.f., such as high-voltage mica types or the 850S ceramic types mentioned.

Drive may be adjusted by means of R₃ in the cathode circuit of the 12BY7 driver.

The 6DQ5 screen voltage is regulated by an 0A2. This regulator also controls the No. 2 grid voltage of the carrier oscillator. The 0C2, operating from a -120-volt source, provides regulated -75 volts. A voltage divider in the grid-biasing circuit of the 6DQ5 reduces this to about -56 volts.

The metering system consists of a single 0-1-ma. meter which may be switched either across a multiplier resistor (R₄) to read final-amplifier cathode current, or to an output-indicating circuit sampling the r.f. output from the pi network.

Receiver Details

To avoid the need for an antenna-transfer relay or switch, input to the receiving r.f. stage (Fig. 2) is coupled through a small high-voltage capacitor permanently connected to the hot end of the transmitter output pi network. To offset the effect of the rather high r.f. voltage that appears at the signal grid of the 6AX8 pentode when transmitting, the t.r. switch applies a negative bias of 75 volts to this grid. This negative voltage is taken from the 0C2 in Fig. 3.

A 6BA7 and 6AR8 were tried as product detectors, but both proved to be very microphonic. However, the 12AX7 grid-leak detector shown in Fig. 4 works fine. Since the b.f.o. frequency from the carrier oscillator is always

30 amp., or about 400 watts. This is plenty for a transceiver rated at a peak input of about 100 watts.

Automobile alternators have three-phase output. They are Y-connected, but the neutral is not used, although it is brought to a terminal on some makes. The alternator contains six silicon diodes which convert the three-phase a.c. to d.c. When the alternator is operating at normal output, the a.c. voltage, line to line, is close to 10 volts. Measuring from line to neutral, it would be close to 6 volts. I opened my alternator and soldered three No. 10 wires to the a.c. output, ahead of the rectifiers. The rectifier connections were not disturbed, so the alternator functions in a normal manner as far as the d.c. output is concerned.

The next step was to raise the low a.c. output to a higher a.c. voltage. I found three surplus filament transformers with 10-volt, 10-ampere secondaries. The primary of each is tapped at 200, 220 and 240 volts. The 10-volt secondaries were connected in the delta configuration and then connected to the three wires from the alternator. The 220-volt taps are used and the primaries are connected in Y. Using the neutral, this stepped-up a.c. is applied to a three-phase rectifier consisting of 12 silicon diodes, 2 diodes per leg. The resulting d.c. output is approximately 600 volts and 300 volts. Because the ripple is low with the three-phase connection, very little filter is needed. A small step-down transformer is connected across one of the larger transformers and produces 110 volts a.c. which is rectified and filtered to produce the bias voltage.

The surplus transformers work very well. Actually, a rating of 10 volts at 5 amp. would be sufficient. Transformers rated at 6.3 volts, 10 amp. with 220-volt primaries could also be used by bringing out a neutral connection from the alternator and connecting the 6.3-volt windings from line to neutral. The 220-volt primaries would be connected in Y. However, probably neither 6.3- or 10-volt transformers with 220-volt primaries are a drug on the market. Control transformers rated at 12 volts 8 amp. are regular catalog items. These may be substituted for the 10-volt transformers mentioned at a sacrifice of 15 to 20% in output voltage.

The efficiency of this arrangement, with reasonably good transformers, is better than that of a good transistor power supply. To gain some idea as to the performance in this installation, look at the accompanying chart. The worst condition is with the engine idling and the headlights turned on. Here, at maximum output, the plate voltage is 490 volts and the plate current is 135 ma. for an input of 65 watts. The best condition is with the engine operating at driving speed and the headlights off. Here, at maximum output, the plate voltage is 630 volts and the plate current is 160 ma. for an input of 101 watts. Using

TABLE I

Engine	Mode	Lights	Low	High	Final
Idle	Rec.	Off	265 v.		
Idle	Rec.	On	255 v.		
Idle	Trans.	Off		500 v.	135 ma.
Idle	Trans.	On		490 v.	135 ma.
Driving	Rec.	Off	320 v.		
Driving	Rec.	On	310 v.		
Driving	Trans.	Off		630 v.	160 ma.
Driving	Trans.	On		600 v.	155 ma.

Driving speed equivalent to 40 m.p.h.

Engine idle speed: 580 r.p.m.

Generator regulator set at 14.2 volts.

Final plate current values shown were taken with maximum sustained modulation.

Minimum final plate current: 35 to 40 ma.

transformers designed for operation over a wide frequency range, the regulation and efficiency would be even better.

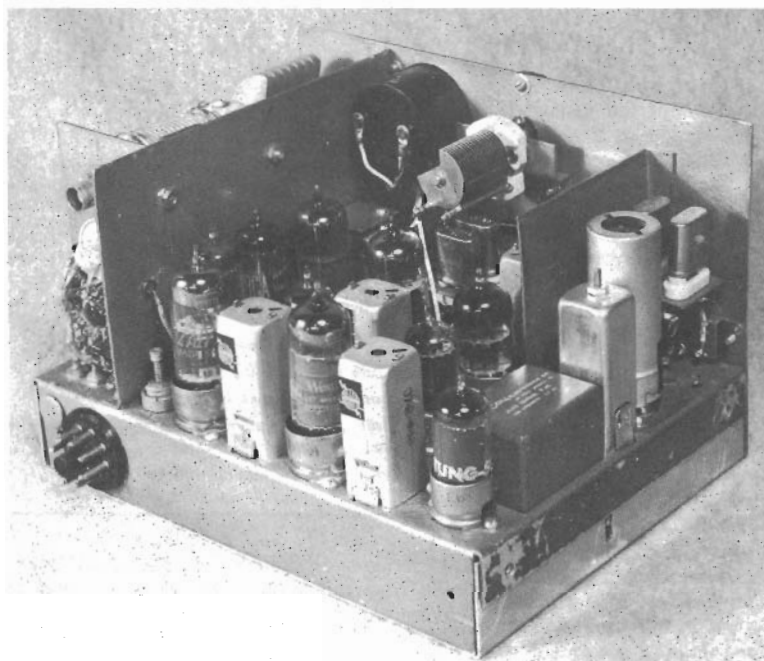
Control Circuit

Referring to Fig. 5, S_6 is the main power switch. This switch turns on all heaters, sets up the 12-volt circuits for the relays and microphone, and operates K_2 which turns on the high-voltage and bias supplies. In this condition, those transceiver stages which are used for both transmit and receive are supplied with 300 volts directly, while those used on receive only are similarly supplied through a back contact of K_1 . At the same time, the a.g.c. circuit from the audio amplifier is connected to the receiver input stage.

Switching from receive to transmit is controlled by a d.p.s.t. toggle switch (or p.t.t. switch) at the microphone. This switch closes the microphone circuit and simultaneously operates K_1 . K_1 shifts the 300-volt line from stages which operate on receive only to stages which operate on transmit only, and simultaneously operates K_3 which switches the 600-volt supply to the final amplifier. At the same time, K_1 switches the grid return of the receiver r.f. amplifier from a.g.c. to a fixed bias of -75 volts to protect the amplifier tube as described earlier. Since K_1 cuts off the plate supply to the audio amplifier, the speaker is muted.

Construction

The transceiver is small in size. In fact, it is too small. There is not enough room to work in easily and not enough metal surface and ventilation to adequately dissipate the heat generated by the tubes. The cabinet should have been a few inches deeper. I solved the heat problem by installing a small 12-volt blower directly behind the transceiver. This blows on the low-level part of the chassis.



Proceeding from right to left around the outer edge of the chassis, the sideband crystals are close to the panel, followed by the 6AR8, T_1 , the sideband filter, the first 12BA6, T_3 , the 6BA7 transmitter mixer, T_6 , the 12BY7A driver, and L_3 . To the left of the sideband filter are the second 12BA6 and the 6AX8, with T_4 still farther to the left, and T_5 just visible to the right. The VXO range crystals are below the VXO tuning capacitor. The detector and receive audio section occupies the space to the rear of the meter.

The cabinet is an LMB type W-1C. It measures 8½ inches long by 6½ inches wide by 6 inches high. Large holes were cut in the top and sides, and then pieces of perforated metal were riveted over the holes. This helps considerably with the ventilation. The photographs show the parts layout. It took a lot of planning to get all of the necessary parts into the small space. The balanced modulator and crystal oscillator are shielded from the rest of the circuit. The receiver front end is also shielded from the 9-Mc. amplifier strip. The 6DQ5 is shielded from the balance of the circuit. No difficulty was experienced which was traced to a lack of isolation or shielding, so the precautions were probably worth the effort.

There isn't much space available for the final amplifier. The 6DQ5 is mounted horizontally with the tank coil and capacitor above it. The tank coil is wound on a surplus ceramic form and is not affected by the heat. The fixed capacitors consist of three high-voltage ceramic capacitors mounted inside the coil form. These also are not bothered by the heat. Air-wound coils, using plastic strips for insulation, would not be suitable for this rig, although they work fine where there is adequate ventilation.

The cabinet, which included the chassis, was purchased in the unpainted condition. All of the holes were cut and the ventilating grilles riveted on, and then all of the metal pieces were painted with hammer-tone paint from an aerosol can. It is not difficult to get a good-looking paint job this way, and there is no worry about scratching the paint while the holes are being cut.

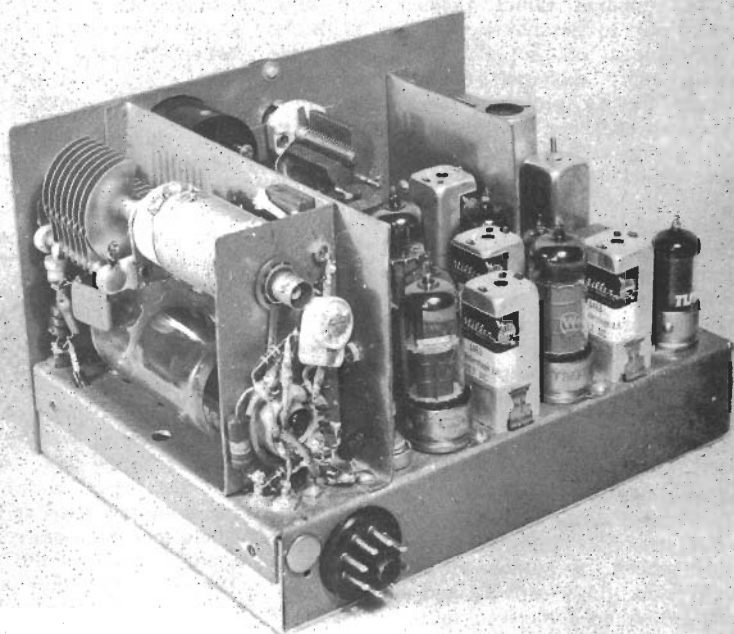
Alignment

A signal generator is not required for tune-up since there is already a built-in oscillator. Only two pieces of test equipment are required. The first is a vacuum-tube voltmeter with an r.f. probe attachment. The second is a grid-dip meter with reasonably accurate calibration.

The use of the commercially-made r.f. transformers saves a lot of time and trouble. They are small, well-shielded, and do not take up much space below the chassis. Using the additional external capacitors shown in the circuit diagram, they will tune to the required frequency with no time-consuming cutting and trying.

The first thing to adjust is the balanced modulator. The 9.0-Mc. oscillator must be working, and this can be checked by measuring the r.f. voltage at the cathode of the 6AR8. No voltage, no oscillation. The VXO should be disabled to avoid any possible stray pickup from it during the null adjustments to follow. Initially, capacitor C_3 is disconnected. The carrier switch, S_2 , is closed. This produces just enough carrier to facilitate tuning the transmitter with the aid of a field-strength meter, and makes it possible to tune the rig without danger of exceeding the dissipation rating of the 6DQ5. Connect the r.f. probe across the secondary of T_1 , and peak the tuning slugs for maximum output. Move the r.f. probe to the grid of the 6BA7 transmitter mixer, and peak the primary of T_3 (the secondary of T_3 is not used.) Now repeat the primary and secondary of T_1 . Set the trimmer (C_1) across the 8998.5-kc. crystal to about half capacitance and

This view shows the final-amplifier assembly. The 6DQ5 and pi-network coil are mounted horizontally from a bracket. The trimmer capacitor is for neutralizing.



switch this crystal into the circuit. The precise setting of C_1 requires a frequency meter like an LM or BC-221. It is possible to get additional carrier suppression by increasing the capacitance of the trimmer. There will be some sacrifice in audio quality if the frequency is set more than 10 db. down on the slope of the crystal-filter passband. This same adjustment will be required in C_2 with the 9001.5-kc. crystal switched in, but in this case the capacitance is increased to move the frequency up the slope of the filter and thereby provide less attenuation of the carrier. The 8998.5-kc. crystal will be used most of the time, since this is the one required to generate the lower sideband. After the transceiver is in operation, a check with a local station will be helpful in setting the crystal trimmers. Move the carrier frequencies as far down the slope of the filter as possible while still retaining acceptable quality. The local station will be able to tell you when you have gone too far.

Now open S_2 . Adjust the balance control, R_1 , for minimum output. Before making further adjustments, whistle directly into the microphone and note the reading on the v.t.v.m. All subsequent adjustments of the balanced modulator should not affect this maximum reading more than about 10 per cent. Now connect trimmer C_3 to one of the plates of the 6AR8. Start with a minimum setting and increase the capacitance. If the output does not decrease, try connecting the capacitor to the other plate. In all of these final adjustments of the balanced modulator, remember that a null is desired, not just reduced output; that is,

adjusting C_3 , R_1 , or the coil slugs in one direction should cause the output first to decrease and then to increase. The null is the minimum reading. Don't make the mistake of assuming that any decrease in output is an improvement in the null reading. Readjustment of the primary and secondary slugs of T_1 may improve the null. Just be sure that you are getting a null and not just decreasing the output by detuning the circuit. All of the null adjustments interact to some degree, so keep making adjustments until there is no further improvement.

Now get the VXO operating. Assuming that the circuit oscillates, and the value of L_1 is set as described earlier, the only adjustment to make is in L_2 . With the power off, use the grid-dip meter to set the frequency of L_2 to approximately 16 Mc. Turn on the VXO, connect the r.f. probe to Pin 2 of the 6BA7 and adjust L_2 for maximum output. Connect the r.f. probe to the No. 2 grid (Pin 8) of the 12BY7 driver. Close S_2 and peak the primary and secondary of T_6 for maximum indication. Be careful that these circuits are not being tuned to 8.2 Mc. instead of 7.2 Mc. Check by turning off the carrier switch. The meter reading should fall to a very low value. Re-adjust L_2 for a maximum reading also.

Move the r.f. probe to the input grid of the 6DQ5. During all of these initial adjustments, remove the plate and screen voltage from this tube. Adjust L_3 for maximum indication and readjust the primary and secondary of T_6 .

The final plate tank circuit is next. Check the tuning range of this circuit with the grid-

dip oscillator. Set the neutralizing capacitor, C_6 , at minimum value. Connect a 52-ohm dummy load to the antenna terminal. With the 6DQ5 filament on, but all other voltages removed, apply excitation to its input grid. Connect the r.f. probe to the plate. Adjust the tuning capacitor C_5 for maximum reading. If this occurs with C_5 set at maximum or minimum capacitance, the output circuit is not tuned to the correct frequency. Now adjust the neutralizing capacitor for a minimum reading. After the antenna is connected and the rig is on the air, some additional adjustment of the neutralizing capacitor may be necessary since a change in the output capacitance will unbalance the bridge neutralizing circuit. Some adjustment of the output capacitor may be required to suit a particular installation. Try 820, 1000 and 1200 pf., and see how much difference they make. The author found that a 0.6- μ h. coil connected from the bottom of the whip antenna to ground made a noticeable improvement in the output.

The remaining adjustments are for the receiver section. A modulated signal generator would help. Just peak T_4 and T_5 for maximum output on a received signal. No further adjustment of the receiver is necessary. All of the tuned circuits operate at 7 Mc. or higher, so there is no need to adjust them in normal operation since the VXO covers a range of only 35 kc.

Without going into excruciating detail, that's about it as far as the alignment and adjustments are concerned. The assumption is made that all bugs have been removed before the final alignment is attempted. Anyone who has

enough experience to start the construction of a single-sideband transceiver should know what to do when he runs into trouble. If not, he'd better not start.

Afterthoughts

My transceiver is installed in the car and is being used daily. Results have been very good, although not any better than commercial equipment having the same power rating, of course. It is a very pleasant surprise to find that you can regularly contact stations who would normally be beyond your range for a.m. It also works the other way. The sideband stations will come through local noise which would ordinarily make a.m. reception difficult, if not impossible.

I had the usual amount of troubles before the transceiver was operating in a satisfactory manner. Some design changes had to be made and the circuit shown in this article is not the same one used at the start of the project. There may be some concern about the unused triode sections of the 6AW8 and the 12AX7. The fact is that they just aren't used. Original plans included the full use of all tubes. It's better to have room for tubes you don't need than to have need for tubes for which you don't have room. Or something like that.

A drawing giving the chassis layout has been purposely omitted. Benefit by my experience. Use a larger chassis and cabinet. Use parts on hand where you can. You aren't going to be able to find some of the exact parts which I used. I have a larger junk box than you do.

Transceiver with Transistors [Almost]

VAROUJAN KARENTZ, W1YLB

This project started off innocently and unintentionally, as most projects do when a choice part or component has been acquired. In this case, it was the donation of a 455-kc. mechanical filter by WIHTK, along with his "maybe you can use this someplace" comment. Its subsequent incorporation into a transistorized (almost) transceiver evolved from some preliminary circuit experiments and then into a system concept which included the following objectives:

1. Compactness and portability for either fixed or mobile use.
2. Built-in a.c. or d.c. power supply.
3. Minimum battery drain when only receiving.
4. Operation on c.w., a.m., and s.s.b. (selectable sideband) with moderate output.
5. All-band (80 through 10) full frequency coverage.
6. Offset receiver tuning, audio-derived a.g.c., r.f. gain control, signal-strength and output-power indicator.
7. Construction with commercially available components wherever possible.
8. Stability adequate for s.s.b. and mobile use.
9. One-knob band switching.

The overall design, however, was compromised because the selection of mixing frequencies was determined by the availability of crystals already on hand. These crystals were borrowed from the home station SB-300 receiver and resulted in using higher oscillator and mixing frequencies than preferred. An inspection of the transistorized SBE-34 transceiver also indicated many desirable circuit features, which were utilized wherever adaptable.

Early in the design, serious consideration was given to a 100-percent solid-state unit. After some experimentation which indicated substantial drive requirements in order to obtain a minimum respectable power output (15 watts), tubes were chosen for the final amplifier and driver. R.f. power transistors do exhibit excellent efficiency—(up to 70 percent) but the low power gain, 15 db. or less, requires relatively high r.f. input power. In addition, the problems of band switching the extra stages and their associated input/output coil taps did not look inviting. Neither did the price of 30-Mc. r.f. power transistors.

As the design developed and stages were bread-boarded, a despairing observation became evident. Specifically, the conventional "well-stocked junk box" was almost useless.



"Almost" transistorized, in this case, means semiconductors everywhere but in the last two transmitter stages, where the r.f. power can be obtained more economically with tubes. The overall size, 11¾ by 5 by 10½ inches, and combination d.c./a.c. power supply make the transceiver useful either in the home station or the car.

The transition of construction techniques from tubes to transistors required the use of components and parts which were not ready to hand, particularly miniature low-voltage and low-wattage common components such as resistors and capacitors. Many of these items were either purchased new or removed from surplus transistorized equipment and printed circuit boards, in addition to an active advertising campaign among some W1 acquaintances.

With any new construction effort, various sizes and values of components are needed where substitution and experimentation are necessary. This added considerably to the total cost of the transceiver, since many of these components did not end up in the final unit.

Test equipment utilized included a v.o.m., v.t.v.m., audio signal generator, grid dip meter, regulated variable d.c. power supply, and a general coverage receiver. During the final alignment and performance checks, a high-frequency wide-band oscilloscope, frequency meter, and r.f. signal generator were used.

General Principles

The simplified block diagram, Fig. 1, indicates signal flow and the various stages comprising the unit. The sideband-generator concept used was originally described by W6TEU¹ as a vacuum tube exciter, and later a transistorized version was incorporated in the SBE-34.

¹ Bigler, "A Sideband Package", *QST*, June, 1958. Also in *Single Sideband for the Radio Amateur*.

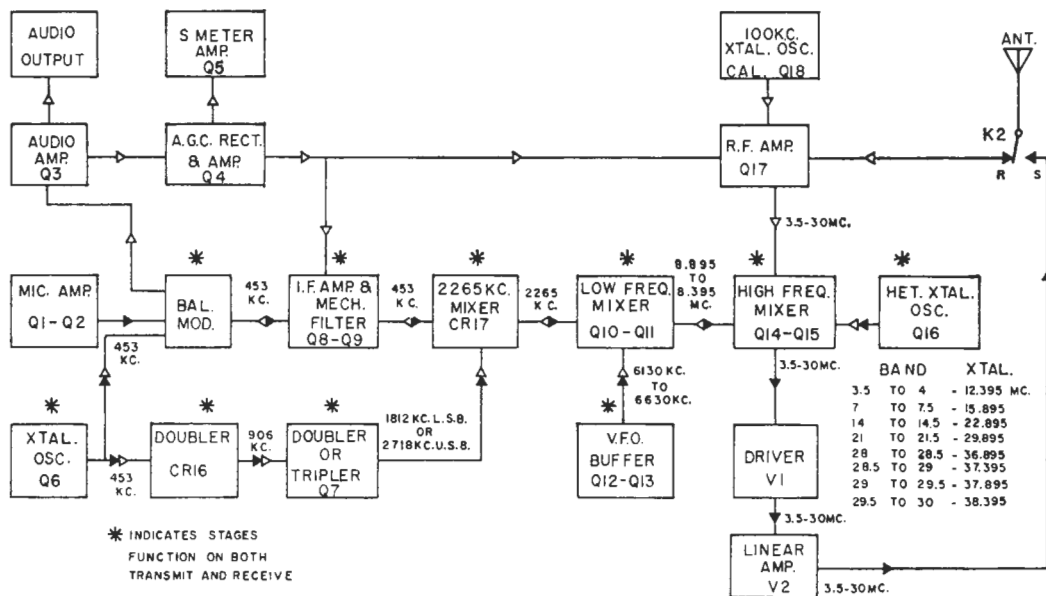


FIG. 1—Block diagram of the transceiver. Open arrowheads indicate direction of signal flow in receiving; solid arrowheads indicate direction in transmitting.

W6TEU's article provides an excellent description and alignment procedure. Basically, the 453-kc. carrier signal from Q_6 is fed into the balanced modulator, where the carrier is nulled out, and the sidebands are then fed through an amplifier and the 455-kc. mechanical filter, which strips off the lower sideband. Sideband selection is accomplished by doubling the carrier frequency and then either doubling or tripling again (in Q_7) to arrive at 1812 kc. for lower sideband or 2718 kc. for upper sideband. The selected frequency is mixed in CR_{17} , with the 453-kc. u.s.b. signal, resulting in a sum or difference suppressed-carrier frequency at all times of 2265 kc., upper or lower sideband. The s.s.b. 2265-kc. output of the sideband generator is fed into the low-frequency mixer stage, Q_{10} , which also receives the output of the v.f.o. The v.f.o. tunes a 500-kc. band from 6130 to 6630 kc. The resultant sum output of the low-frequency mixer is tunable from 8.395 to 8.895 Mc. This signal is then converted to the desired operating band in the following high-frequency mixer stage, Q_{14} , by the associated heterodyne crystal oscillator, Q_{16} . Since the heterodyne oscillator frequency is always on the high side of the h.f. mixer input signal, a single v.f.o. dial calibration will suffice for all bands when the proper heterodyne-oscillator crystals are selected. With the v.f.o. dial calibrated from 0 to 500 kc. the operating frequency is directly read on the dial by adding the lowest frequency in megacycles, for the band in use, to the dial reading. The 10-meter band requires four 500-kc. segments to cover 28 to 29.6 Mc.

In the TRANSMIT mode the output from the high-frequency mixer, Q_{14} , is fed to the 12BY7

Class A driver and from there to the 6JB6 Class AB₁ final linear amplifier. In the RECEIVE mode the signal from the antenna is coupled into the r.f. amplifier stage, Q_{17} , and thence to the high-frequency mixer, after which it follows a reverse path back through the mixers to the diode balanced modulator, which acts as a detector. The detected signal is then amplified by the audio amplifier, Q_3 , and the audio output stages. At the same time, the audio output is gain-controlled by the a.g.c. amplifier, Q_4 , which controls the gain of the r.f. amplifier, Q_{17} , and the 453-kc. amplifier, Q_9 .

Fig. 1 also indicates those stages which operate in either the receive or transmit modes. The amplifier/mechanical-filter, low-frequency mixer and high-frequency mixer perform bilaterally, and can be considered unidirectional in the selected mode, allowing signals to be passed in the desired direction. The injection oscillators Q_6 , Q_7 , Q_{12} , Q_{13} , and Q_{16} operate continuously. Other stages are biased off as required.

Microphone Amplifier

The mike preamplifier, Q_1 , and amplifier, Q_2 , are conventional common-emitter amplifiers. Q_1 is designed for low-impedance output, isolated and bypassed for r.f. by the RC combination of the 4700-ohm resistor and 470-pf. capacitor Fig. 2. The audio stages were built on a $3\frac{1}{2} \times 1\frac{1}{2}$ -inch epoxy board, allowing ample room for addition of a speech compressor at a later date. The two stages of audio provide ample audio gain for this use. These two stages provide sufficient gain (in excess of one volt output) even when a high-impedance -54-dbm.-output microphone is used. With

this mismatch the amplifier dynamic gain is reduced, the microphone is heavily loaded, and some low frequencies are attenuated. However, audio response is adequate since the microphone in use has a roll-off characteristic in the opposite direction. A matching input transformer (100,000 to 2000 ohms) could be used for a better match. The output would then be proportionally increased, and possibly Q_2 would not be necessary since only 0.2 to 0.3 volt of audio is needed to drive the balanced modulator. It should be noted that to reduce hum and feedback, Q_1 and Q_2 are decoupled from the 11.5-volt bus through a 150-ohm resistor and a 100- μ f. bypass capacitor.

Balanced Modulator

The diode ring-type balanced modulator, Fig. 2, provides approximately 35 db. of carrier suppression as measured with an r.f. probe and v.t.v.m. for TUNE/c.w. operation a small d.c. voltage is allowed to upset the balanced modulator through the CARRIER INSERT control, R_{12} , when the function switch, S_6 , (Fig. 4), is in the TUNE or c.w. position. The amount of voltage or carrier insertion is adjusted by this rear-panel 10K control pot. For c.w. operation a key is inserted into the normally-closed jack, J_2 , interrupting the d.c. path except in the key-down position. The c.w. note is remarkably smooth. This is partially attributable to the filter network composed of the 56K resistor and the two 0.1- μ f. capacitors. During c.w. operation the mike gain control should, of course, be fully counterclockwise. Amplitude modulation is possible by setting the amount of carrier insertion to the safe AB₁ plate-dissipation operating point of the 6JB6

final amplifier and adjusting the mike gain for proper modulation.

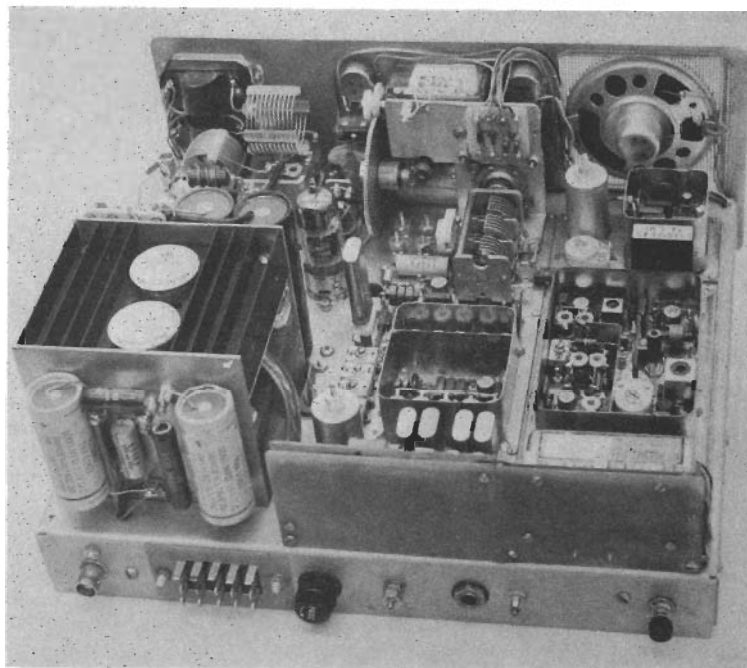
As in most balanced modulators, some interaction exists between the carrier null pot, the tuning of transformer T_{11} , and the 7-5-pf. trimmer capacitor, C_3 . Adjustments to each alternately are necessary to obtain maximum carrier null.

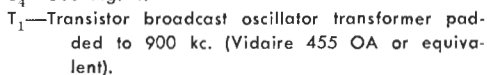
Amplifier, Filter and Low-Frequency Mixer

The 453-kc. common-emitter amplifiers, Q_8 and Q_6 , are controlled by the 11.5-volt d.c. or ground bus as selected by control relay, K_1 , (Fig. 4), as are also the low-frequency mixers, Q_{10} and Q_{13} , and high-frequency mixers, Q_{14} and Q_{15} (Fig. 3). In the TRANSMIT mode the bias resistors for Q_8 , Q_{10} and Q_{14} are grounded, completing the bias voltage-divider network path and allowing these transistors to conduct. In the RECEIVE mode the same bias resistors receive a positive voltage (base and emitter at same potential), effectively cutting these transistors off. In either case, the exact reverse biasing method is used for Q_9 , Q_{11} and Q_{15} , enabling conduction of the stages in the desired transmit or receive direction. The two capacitors in series across the input to the mechanical filter resonate the filter to 455 kc. and provide a better impedance match to the base of Q_9 .

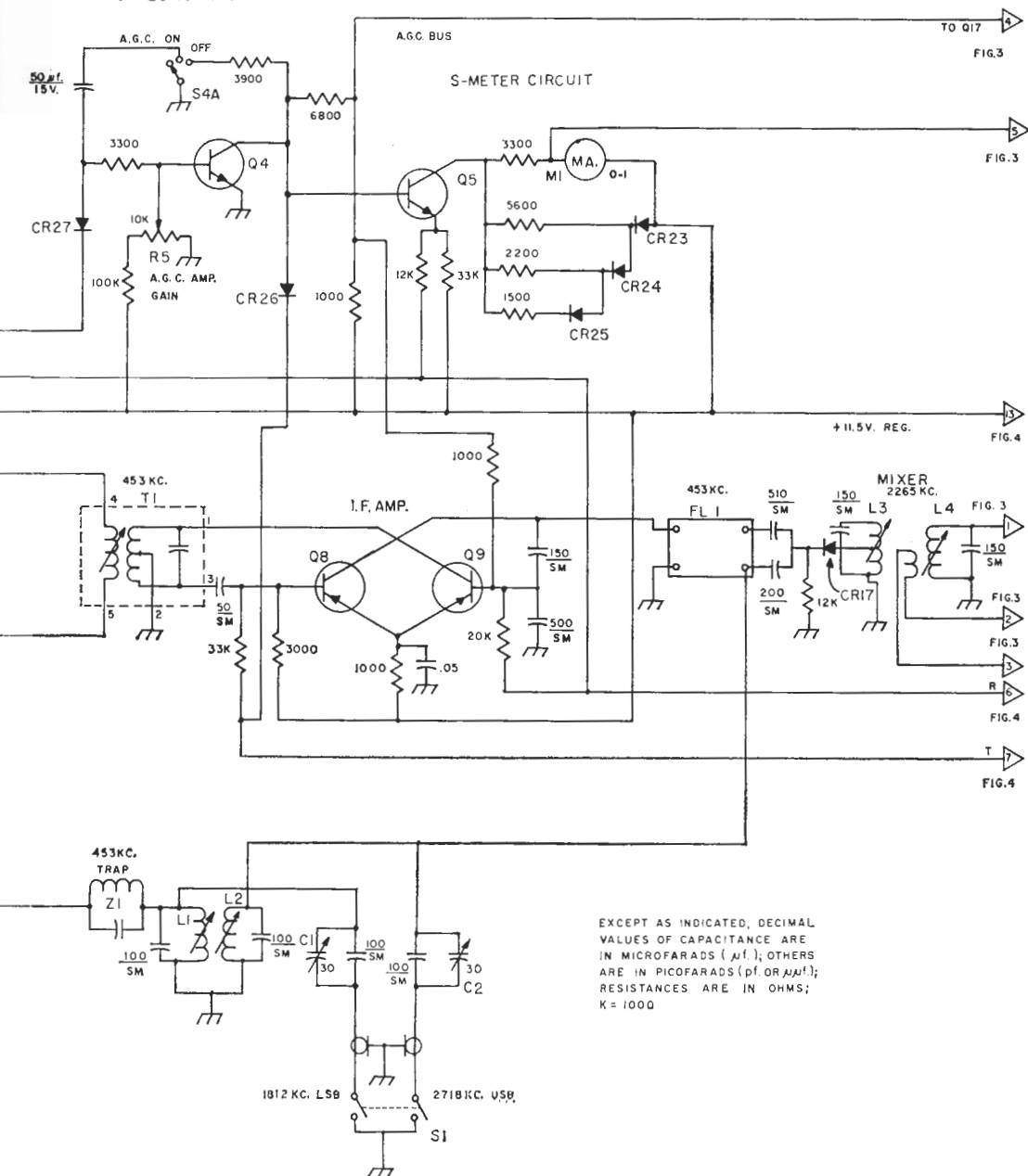
The 3.1-kc. bandwidth filter has a substantial insertion loss—almost 15 db. With 2.5 volts of r.f. at the collector of Q_8 , centered in the middle of the pass band, the output from the filter at the terminals is 0.5 volt. Limited information was available on the actual slope and attenuation characteristics of the filter in use.

In the top-of-chassis layout the transmitting driver and final amplifier occupy the left-hand section between the power supply and panel. Audio, i.f., and v.f.o. circuits are along the right-hand edge; the mechanical filter is visible beyond the upper edge of the circuit board mounted vertically along the rear chassis edge. In the center section, the receiving r.f. amplifier and mixer components are alongside the three-gang tuning capacitor; the heterodyne oscillator and its crystals are in the foreground.





A.G.C. RECT. & AMP.



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μ f); OTHERS ARE IN PICOFARADS (pf. OR μ pf.); RESISTANCES ARE IN OHMS; K = 1000

T₃—Transistor interstage audio transformer, 10,000 to 2000 ohms (Lafayette TR-96, center tap not used).

Y₁—453 kc. (Surplus FT-241A, Channel 45).

Z₁—Miniature 455-kc. i.f. transformer (see text).

DIODES AND TRANSISTORS

CR₁₂—CR₁₅, inc. CR₂₇—Germanium; 1N34A, 1N67A, 1N68, or similar, matched for forward resistance.

CR₁₆, CR₁₇—Germanium, see text.

CR₂₃—CR₂₆, inc.—Silicon; 1N914, 1N484, 1N645, or similar.

Q₁, Q₃—2N508 (p-n-p).

Q₂—2N396 (p-n-p).

Q₄—2N697, 2N440A, 2N1893, 2N1613, HEP-50 (n-p-n).

Q₅—2N1613, 2N697, HEP-50 (n-p-n).

Q₆—2N396A, 2N425, 2N1305 (p-n-p).

Q₇, Q₈, Q₉—2N396A, 2N425, HEP-51, 2N1305 (p-n-p).

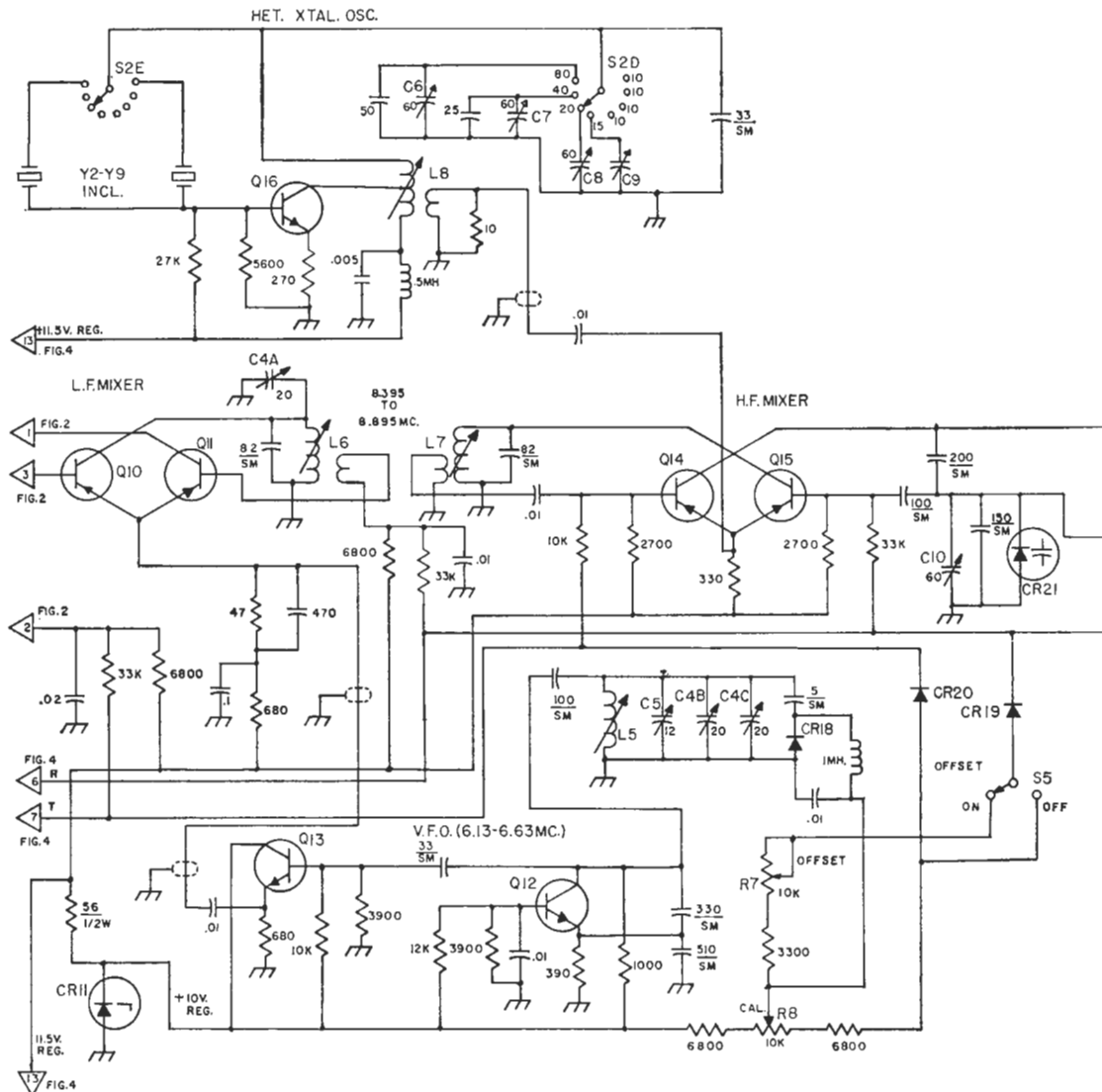


FIG. 3—Circuit of the low- and high-frequency mixers, heterodyne oscillator, v.f.o., driver and final amplifier, receiving r.f. amplifier and 100-kc. calibrator. Fixed capacitors marked SM are silver mica; others are ceramic. Unless otherwise indicated, fixed resistors are $\frac{1}{4}$ -watt composition.

C_3 —3-section variable; 6-20 pf. per section (Miller 1460).

C_5 —NPO ceramic trimmer, 3-12 pf. (Centralab 822-FZ or equivalent).

C_6, C_7, C_8, C_{10} —8-60 pf. mica trimmer (Arco 404 or equivalent).

C_9 —2-20 pf. mica trimmer (Arco 402 or equivalent).

C_{11} —5-45 pf. air padder with rear shaft extension, ganged with R_1 .

C_{12} —2-section superhet-type variable, 365 and 135 pf. (Lafayette 32G1101 or equivalent).

C_{13} —65-340 pf. mica trimmer (Arco 303 or equivalent) modified by adding $\frac{1}{8}$ -inch shaft for panel control.

C_{14} —1-8 pf. piston trimmer, plastic (Erie 532-000-BR or equivalent).

L_3 — L_{23} inc.—See Table I.

L_{24} —11 turns No. 16, air-wound, 1 $\frac{1}{4}$ in. dia., 8 turns per inch (B & W 3018 or equivalent).

L_{25} —14 turns No. 20, air-wound, 1 in. dia., 16 turns per inch, tapped 4th turn from ground end (B & W 3015 or equivalent).

M_2 —0.1 milliammeter (Calrad EW-2 or equivalent); indicates 2000 ma. full-scale in circuit shown.

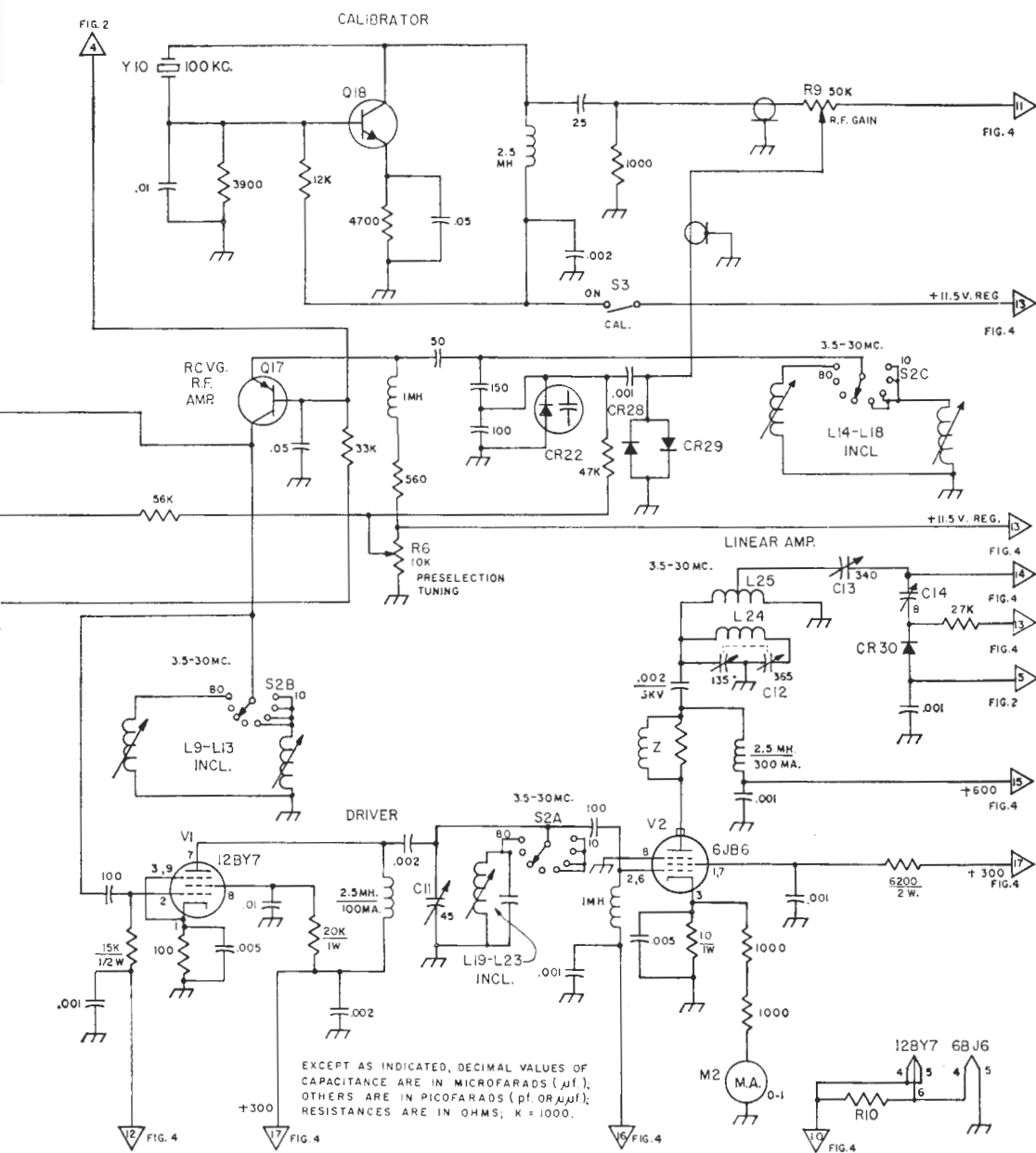
R_6 — R_{11} inc.—Linear-taper control, $\frac{1}{4}$ or $\frac{1}{2}$ watt.

R_{10} —7.5 ohms, 10 watts (TV ballast type, Hamilton-Hall FR-7.5).

S_2 —Ceramic rotary, 5 sections, 1 pole per section, 11 positions (8 used) (Centralab PS-21 sections with indexes; see text).

S_3 —S.p.s.t. slide switch.

S_4 —S.p.d.t. slide switch.



Y_2 - Y_9 , inc.—See Fig. 1 for frequencies.

 $Y_{10} = 100 \text{ kc.}$

Z—4 turns No. 16 spaced to occupy length of 100-ohm 2-watt composition resistor.

DIODES AND TRANSISTORS

CR₁₁—10-volt zener, 1/2 watt (1N758 or similar).

CR₁₈, CR₂₈-CR₃₀, inc.—Silicon (1N484, 1N645, or similar).

CR₁₀, CR₂₀—Silicon, matched forward resistances (1N434B, 1N484⁸, 1N645 or similar).

CR₂₁, CR₂₂—Voltage-variable capacitor (1N955, TRW V47 or V947, or similar).

Q₁₀, Q₁₁, Q₁₄, Q₁₅, Q₁₇—PNP r.f. type (2N2905A, 2N2672, 2N1132, 2N711, HEP-51* or similar).

Q₁₂, Q₁₃—NPH, r.f. type (2N706, 2N708, 2N918 or similar).

Q₁₆, Q₁₈—NPN, r.f. type (2N708, 2N918, HEP-50 or similar).

* HEP-51 not optimum for Q_{17} .

TABLE I

All coils listed below are close-wound on slug-tuned forms using enameled wire. Taps, when required, are counted off from the ground end of the coil. Shunt capacitors should be silver mica.

Coil	Form Dia. in.	Wire Size	No. of Turns	Tap Turns	Shunt Cap. pf.
L_1	$\frac{3}{16}$	33	75		
L_2	$\frac{3}{16}$	33	65		
L_3	$\frac{3}{16}$	33	80	45	
L_4	$\frac{3}{16}$	33	80	10-turn link	
L_5	$\frac{3}{8}$	26	25		
L_6, L_7	$\frac{3}{16}$	20	24	4-turn link	
L_8	$\frac{1}{4}$	24	10	7 2-turn link	
L_9	$\frac{3}{16}$	33	50		
L_{10}, L_{15}	$\frac{3}{16}$	26	26		
L_{11}	$\frac{3}{16}$	24	13		
L_{12}, L_{17}	$\frac{3}{16}$	24	8		
L_{13}, L_{18}	$\frac{3}{16}$	24	5		
L_{14}	$\frac{3}{16}$	33	55		
L_{16}	$\frac{3}{16}$	24	14		
L_{19}	$\frac{1}{4}$	26	40		150
L_{20}	$\frac{1}{4}$	26	22		100
L_{21}	$\frac{1}{4}$	22	14		75
L_{22}	$\frac{1}{4}$	22	12		50
L_{23}	$\frac{1}{4}$	22	17		22

If the newer type 2.1-kc. Collins experimenters' filter is used the insertion loss should not be as severe, on the assumption that the newer filters have improved characteristics. A different carrier-oscillator crystal frequency would have to be used to place the carrier at the proper point on the filter slope.

A 453-kc. tuned trap (miniature 455-kc. transistor i.f. transformer), Z_1 , is in series with the collector of Q_7 and coil L_1 . Apparently a small amount of fundamental signal (453 kc.) from the crystal oscillator was not adequately rejected in the frequency-doubler transformer T_2 or by the high-frequency tuned circuits L_1 and L_2 . Prior to inserting the trap this 453-kc. leak fed back through the filter out of phase with the suppressed-carrier signal and caused difficulty in balancing out the carrier on upper sideband. A double-tuned circuit substituted for T_2 , or possibly a different physical layout, would eliminate the need for this series trap.

It is necessary that double-tuned circuits be used wherever indicated in the schematic, to provide the selectivity necessary for rejecting harmonics and unwanted mixing frequencies and provide a clean signal for the following stages. The doubler diode, CR_{16} , and mixer diode, CR_{17} , where selected for optimum signal output, as were their associated bias resistors. A number of different diodes tried worked, but it was noticed that because of various characteristics a particular diode performed better. Both types finally used were unmarked germanium surplus.

The output (or input as may be the case) coil, L_6 (Fig. 3), of the low-frequency mixer stage is tuned and tracks with one section of the three-gang v.f.o. tuning capacitor. This provides uniform frequency response, along with rejection of unwanted frequencies, to its associated coil, L_7 . The mica trimmer of the variable-capacitor section is adjusted to obtain the padding necessary to tune L_6 through a 500-kc. range.

Audio Circuits

Audio amplifier Q_3 , Fig. 2, obtains the received signal from the ring modulator, now acting as a detector. Its output is fed into the audio-derived a.g.c. amplifier, Q_4 , and the pre-packaged 100-mw. audio output amplifier. The 0.02- μ f. capacitor and 12K resistor across the primary of T_3 improves the frequency response and provides a more constant load for Q_3 . The output audio amplifier is designed for a common positive battery supply, and therefore the amplifier printed circuit board has to be insulated from the receiver chassis and decoupled from the 11.5-volt supply. Correspondingly, the voice coil of the speaker must be returned to the proper terminals on the p.c. board. Q_3 is not biased off in TRANSMIT since the audio output amplifier is cut off completely. One hundred milliwatts of audio output with a miniature 3-inch speaker certainly does not appear very convincing to the high-fi-minded, but the result is gratifying. A 4-inch speaker with a large magnet gave significantly im-

proved output and response, but space limitations dictated the use of the smaller speaker.

A.g.c./Meter

The audio signal to the a.g.c. amplifier, Q_4 , is rectified by CR_{27} and applied as a negative-going voltage to Q_4 's base. A fast attack and slow release characteristic is obtained by the combination of the base bias resistors and the 50- μ f. capacitor. A.g.c. action reduces the gain of the r.f. amplifier, Q_{17} , and the 453-kc. amplifier, Q_9 , by decreasing their base-to-emitter voltage, which in turn reduces collector current. With no incoming signal Q_4 is conducting heavily and the potential at the junction of the 1000-ohm a.g.c. load resistor and the a.g.c. bus is less than 11.5 volts. As an incoming signal is applied to its base, Q_4 conducts proportionally less and the a.g.c. bus potential increases, thereby raising the base voltage of Q_{17} and Q_9 . The 6800-ohm resistor in series with the collector of Q_4 limits the a.g.c. action until an incoming signal exceeds the audio level where output variations are detectable. The gain of the a.g.c. amplifier is adjustable by a rear-panel control, R_5 . A.g.c. can be defeated completely by switch S_{4A} , which effectively shunts Q_4 and places the a.g.c. bus potential at approximately 9 volts.

The S-meter circuitry is unusual in that it provides approximately logarithmic compression by nonlinear action. As the a.g.c. voltage applied to the base of Q_5 increases, the collector current decreases and the potential across each silicon diode (CR_{23} , CR_{24} , CR_{25}) rises, exceeding the conduction point (0.5 volt average) of each diode successively as determined by its series resistor. Current is now shunted through each diode, limiting the current through the meter, M_1 . The 1500-ohm resistor and CR_{25} establish full scale or 30 db. as indicated on the meter. As the incoming a.g.c. voltage decreases, Q_5 conducts more heavily, the potential across CR_{25} becomes less and it stops conducting, followed by CR_{24} and CR_{23} in that order, thus reducing the compression. With the 3300-ohm resistor in series with the meter, compression does not begin until a reading of S9 is indicated, hence approximately 30 db. of logarithmic compression is achieved. This action is dependent, of course, upon the non-linear a.g.c. characteristics and r.f./i.f. gain variations from band to band—the primary downfall of all S-meter circuitry. The values of the voltage divider resistors in the emitter circuit of Q_5 are selected experimentally so that when R_5 is properly adjusted M_1 will be zeroed. A separate pot in the emitter circuit of Q_5 could be substituted and would provide ease of adjustment. CR_{26} acts as a diode switch to cut off Q_5 in TRANSMIT. This zeroes M_1 and allows it to function as a relative-output meter from the circuitry associated with the final amplifier tank.

Variable-Frequency Oscillator

The v.f.o. construction departs from the tried and true philosophy of rigid and heavy con-

struction, yet retains good thermal and mechanical stability. The entire v.f.o., with the exception of C_4 and the calibrate and offset circuitry, was mounted on a copper-clad 1½ by 3-inch epoxy board. Q_{12} , Q_{13} and L_5 are contained in a shielded enclosure. The oscillator, Q_{12} , is in a common-emitter Colpitts configuration, with an associated emitter follower, Q_{13} , used for isolation. The collector voltage for Q_{12} and Q_{13} is regulated by a Zener diode, CR_{11} . One volt c^d r.f. output is available at the emitter of Q_{13} . Two sections of the variable capacitor, C_4 , are paralleled in order to make the v.f.o. cover from 6130 to 6630 kc.; the adjustments available by the slug in L_5 and trimmer capacitor C_5 enable the frequency range and tracking to be set.

The v.f.o. circuit incorporates dial-calibration and receiver-offset features. CR_{18} is a silicon diode which exhibits a slight capacitance variation when reversed biased, and is placed in series with a 5-pf. capacitor across the v.f.o. coil. By varying the bias voltage the frequency of the oscillator can be changed independently of C_4 by about 15 kc. The calibrating pot, R_8 , initially sets the v.f.o. frequency to correspond with the dial (digital counter) reading. R_8 always functions in TRANSMIT, but in RECEIVE it functions only when the offset switch, S_5 , is in the off position. If it is desired to change the v.f.o. frequency while in the RECEIVE mode, the offset pot, R_7 , is switched into the circuit. This control will vary the received frequency approximately 4 kc. either side of the dial reading while the transmitting frequency always remains where it was set by the dial. The diode switch, CR_{19} , and the position of the offset switch determine when the offset control is in the circuit. When S_7 is in its off position, CR_{20} does not conduct during RECEIVE but CR_{19} does conduct, keeping the calibrate pot in the circuit. On TRANSMIT, CR_{20} always conducts regardless of the switch position. CR_{19} and CR_{20} must be evenly matched in forward resistance since unequal voltage drops would change the voltage on CR_{18} when switching from TRANSMIT to RECEIVE, thereby causing a frequency shift.

V.f.o. stability was achieved by an effective, but not yet well recognized, simple method. Very briefly, transistor junction heating, from whatever source, varies the transistor characteristics—significantly, its capacitance—resulting in frequency drift. This junction heating in an oscillator is also a function of the feedback voltage, which determines to some extent the collector current. By using a high-Q tuned circuit (as in any oscillator) and selecting the correct amount of feedback voltage or collector current, a set of operating conditions can be established which will minimize oscillator drift (other than that caused by external temperature changes). In this case, a fixed regulated voltage (10-volt Zener diode regulator) was selected and various values of feedback capacitance were tried experimentally until the drift of the oscillator was recognized as doing posi-

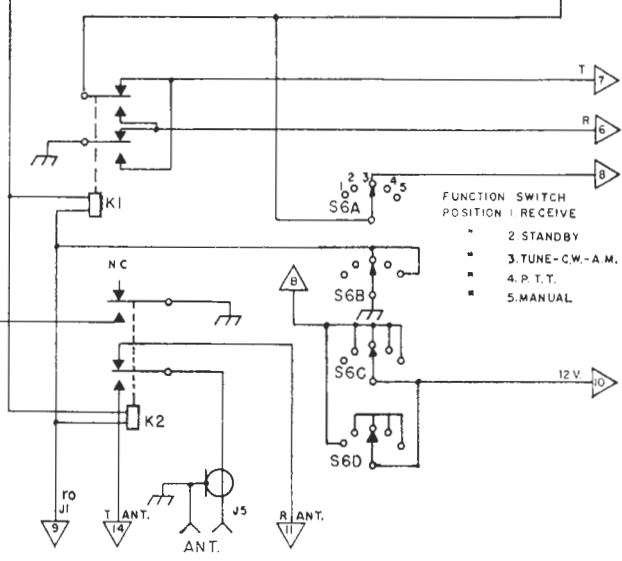


FIG. 4—Power and bias supplies, and control circuits. Unless otherwise specified, fixed resistors are $\frac{1}{2}$ watt. Capacitors with polarity indicated are electrolytic; others are paper.

- CR₁, CR₂—Silicon, 800 p.i.v., 500 ma.
CR₃—Silicon, 800 p.i.v., 100 ma.
CR₄—CR₇, inc.—Silicon, 50 p.i.v., 5 amp.
CR₈—Zener, 6.5 volts, 0.75 watt.*
CR₉—Zener, 5.0 volts, 0.75 watt.*
CR₁₀—Silicon, 25 p.i.v., 500 ma.
I₁₋₄, inc.—Pilot lamp, 6–8 volts, for meter
and dial illumination (Lafayette 99
C 6262).
J₃—Chassis-mounting 15-contact male conn.
(Cinch Jones P-315-AB).
P₁, P₂—Cable-mounting 15-contact female
connector (Cinch-Jones S-315-CCT).
K₁, K₂—D.p.d.t., 12-v. d.c., 10–20 ma.
Q₁₉—2N697, 2N1613, 2N1893, or equiva-
lent (n-p-n).
Q₂₀, Q₂₁—2N441, 2N442, HEP231, or equivalent
(p-n-p).
R₁₁—50,000-ohm linear control, ½ watt.
S₄—2 pole, 3-position rotary, 5-amp. contact rating
(one section, S_{4a}, Fig. 2, is used for a.g.c.
switching; S_{4b} is used for power switching.
These functions are combined to conserve
panel space).
S₆—Miniature ceramic rotary, 2 sections, 2 poles per

section, 5 positions, non-shorting (Centralab PS-111).

S₇—S.p.s.t. toggle (on R₂).

T₄—Transistor converter/a.c.; primaries, 12.6 volts d.c. and 115 volts a.c.; secondaries 280 volts d.c., at 150 ma., 12.6 volts a.c., 3 amp. (Thardarson TR-294 or Stancor P-8195).

Note: Heat sink is Motorola type MS-15 modified to take two TO-36 transistors.

* CR₈ and CR₉ series for 11.5-volt output.

tive; then the values were changed to find the point where drift was going negative. The capacitance values indicated in Fig. 3 are those which fell in between. The alternative method would be to select the optimum value of the feedback capacitor to maintain maximum Q and then adjust the collector voltage in varying increments (noting voltage values) until the drift rate changes from negative to positive. At the zero-drift point a Zener diode (or combination of them) can be substituted to maintain the collector voltage at that point. It should be noted that this is not temperature compensation in the normal sense—i.e., it is not applicable to thermal changes in external components.

The v.f.o. drive uses a Jackson dual-ratio vernier control to allow either fast or slow tuning. The digital counter and associated gears were obtained from various surplus sources, including some local WI's who dug real deep to the very bottoms of their junk boxes. A lucky combination of ratios was made up to obtain exactly the required 0 to 500 counter reading from minimum mesh to full mesh of C_1 . A circular direct-driven dial is much simpler and of course would not require any gearing. For information, with the gearing available the last gear ended up at the digital counter with a one-to-one ratio. This was necessary in order to have the digital counter read correctly by turning in the reverse direction to the tuning capacitor; with the heterodyne crystal oscillator on the high side of the mixer frequency, the v.f.o. frequency decreases as the signal frequency increases.

Heterodyne Oscillator

Link coupling is used from the heterodyne crystal oscillator, Q_{16} , to the emitters of the high-frequency mixers, Q_{14} and Q_{15} . Although a different crystal is used to cover each of the four segments of the 10-meter band, L_8 with the parallel 3-pf. capacitor allows oscillation to take place with any one of the four. Trimmer capacitors resonate the coils for each of the lower bands. On 80 and 40 meters, an additional fixed capacitance is shunted across the trimmer.

Varicap Tuning

Among the problems of tunable circuit design are those of matching to the input of transistors and the extra switching required to connect each tuned circuit's low-impedance tap to the transistor. A compromise was reached by eliminating the extra switching in the r.f. amplifier and h.f. mixer stages while still retaining an acceptable impedance match. Both the r.f. amplifier, Q_{17} , and high-frequency mixer, Q_{15} , utilize a voltage-variable capacitor diode (CR_{21} and CR_{22}) for tuning the band in use. These diodes (Varicaps), specifically designed for relatively high- Q r.f. applications, are used in a series-parallel combination with fixed voltage-divider capacitors for impedance matching. The two Varicaps are remotely con-

trolled by a common front-panel pot, R_6 . R_6 is ganged to the 12BY7 driver tank-circuit capacitor, C_{11} , and thus is used for single-control preselector tuning in RECEIVE and driver output tuning in TRANSMIT. Trimmer capacitor C_{10} in the base of Q_{15} is a padding adjustment for CR_{21} to keep the capacitance range consistent with the frequency band it covers. In circuits of this type where r.f. voltage is applied, the d.c. bias across the Varicap must be greater than the developed r.f. voltage since it is possible that the capacitance of the Varicap can be changed by the r.f. voltage if it exceeds the d.c. bias level. This normally undesirable situation is put to good use, when Q_{14} is conducting, to provide some degree of low-level a.l.c. action. The d.c. bias is reduced slightly, with the L/C ratios adjusted to maintain resonance at the desired frequency. When the r.f. voltage amplitude increases with speech and exceeds the threshold d.c. bias, the change in Varicap capacitance detunes the circuit and the output proportionally levels off.

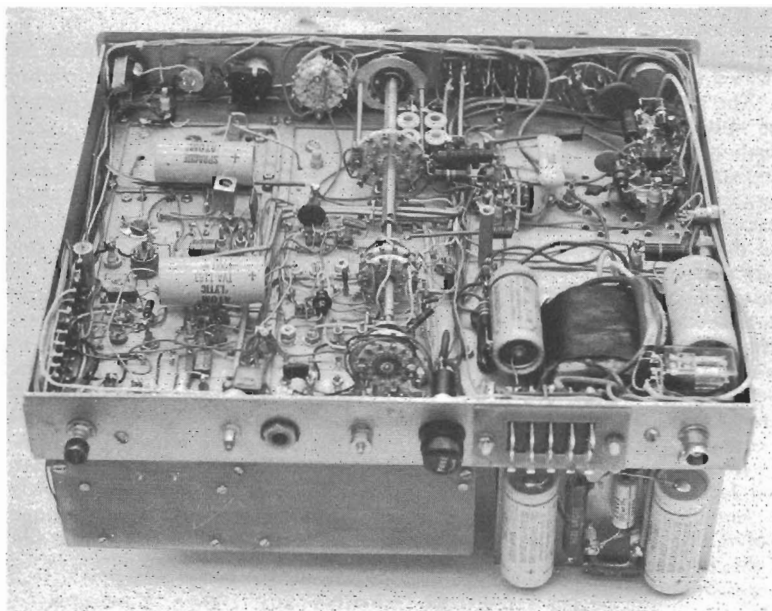
Rf. Amplifier and 100-kc. Calibrator

Q_{17} is a common-base amplifier for maximum voltage gain and high-impedance output; the latter is desirable for minimizing loading of the high-frequency mixer and driver input stage. Protection is provided from transmitted r.f. by two silicon diodes, CR_{28} and CR_{29} , which conduct to ground when the r.f. voltage is greater than 0.5 volt at the front end. The antenna is tapped down for impedance matching by the capacitor voltage divider mentioned previously, and the circuit is tuned by CR_{22} . R.f. gain is controlled right at the receiver front end, ahead of the amplifier, and a strong signal at the antenna that could cause overloading can be effectively attenuated by this control. As Q_{17} is always operating at maximum gain, no compromise is made on a.g.c. characteristics, as usually is necessary in normal r.f. gain control circuits.

By tying one end of the r.f. gain-control pot, R_9 , to the output of the 100-kc. crystal calibrator, variable-amplitude calibration injection is available. When the calibrator is turned on by S_3 and R_9 is rotated toward the calibrator end, signals coming from the antenna are attenuated. Eliminating incoming signals and atmospheric background noise makes the 100-kc. markers easily identifiable across any band.

Driver and Final Amplifier

The 12BY7 class A driver is completely cut off in RECEIVE by applying -80 volts through K_2 (Fig. 4) and the grid resistor. In TRANSMIT the grid resistor is allowed to complete its normal path to ground. Up to 3.5 volts peak r.f. is available at the grid of this stage on 10 meters. All the driver output coils, L_{19} to L_{23} , are swamped with resistors (not shown in the schematic) to provide a constant load and to prevent self-oscillation. The values of these resistors were not critical, and they were ex-



The band switch extends along the center of the underside of the chassis. Wiring here is principally between the circuit boards that make up individual sections of the transceiver.

perimentally selected to allow sufficient drive to the final amplifier, yet maintain stability. There is more than enough drive on all bands and heavy swamping was necessary, particularly across the 80- and 40-meter coils, to prevent grid current flow in the final amplifier. As information, the values used were: L_{19} and L_{20} , 4700 ohms; L_{21} , 6800 ohms; L_{22} and L_{23} , 10,000 ohms; all $\frac{1}{2}$ watt. Optimum values should be determined experimentally.

A multiband tuner is used in the final tank circuit. It resonates in 80-, 20-, 40-, 15-, 10-meter sequence from maximum to minimum capacitance.

C_{13} , the output loading capacitor, is adjusted conventionally for loading the amplifier into the antenna.

Relative power-output indication is obtained by rectifying a portion of the r.f. output by CR_{40} and applying it to the combination output/S-meter. Relative reading on the meter can be adjusted by C_{14} . Since the meter circuitry is at a positive potential the diode rectifier path for this circuit must be returned to the 11.5-volt bus rather than to ground. The 0-1 milliammeter, M_2 , is connected as a voltmeter to indicate 200 ma. full scale. Final-amplifier resting plate current is adjusted to 25 ma. on TRANSMIT by the bias pot, R_{11} (Fig. 4); on RECEIVE the resting current decreases to 15 ma. because the 100K resistor in the grid circuit of the 12BY7 is lifted from the ground and increases the bias voltage. Plate current in the TUNE/C.W. position with the amplifier loaded is 150 ma.

Power Supply and Regulator

High voltage is obtained from a dual-purpose power supply. The power supply transformer, T_4 , is a readily available item and is especially

designed for either 115 volts a.c. or 12 volts d.c. input. The secondary, which is rated at 280 volts, 150 ma., is used with a voltage-doubling rectifier-filter which raises the B+ voltage to 600 volts, and also supplies a 300-volt output for the driver B+ and the screen of the final. The supply has been loaded to a full 200 ma. continuously without any evidence of excessive heat. Negative voltage for the final amplifier and the driver grids is obtained by a shunt rectifier directly off the secondary a.c. winding.

In a.c. operation full-wave bridge rectification is used from the 12-volt a.c. winding to provide d.c. voltage for the transistors. This voltage is filtered and then regulated to 11.5 volts by Q_{19} . CR_8 and CR_9 are Zener diodes which establish the reference level for the 11.5 volt regulated bus. This figure was chosen in order to allow for possible voltage drop in supply leads from a 12-volt battery when d.c. power supply is used. A 10- or 11-volt Zener probably could be substituted with no change in overall operation, but in that case the optimum values for bias resistors for the various stages might be different from those given and should be determined experimentally.

With 12 volts d.c. input, CR_{10} acts as a one way current valve, preventing 12 volts d.c. from being applied back through the d.c.-to-d.c. converter. This is necessary in order to allow the receiver to be turned on by switch S_7 , on R_7 , and yet not allow the filaments or other circuits to draw current when the "receiver only" mode is selected. Total current drain in the "receiver only" mode is 140 ma. Half of this current is used by the four illuminating lamps for the dial and meters.

The d.c.-to-d.c. converter portion of the power supply circuitry is that recommended

by the transformer manufacturer, with the exception that higher-power transistors (Q_{20} and Q_{21}) are used. Both transistors are mounted on a finned heat sink attached to the top of T_1 . The end bells of T_1 were removed to save space and enable T_4 to be mounted horizontally.

Construction Notes

The balanced modulator, its associated crystal oscillator and double/tripler, the mechanical filter/453-k.c. amplifiers, and low-frequency mixers were constructed on a 6½ by 3-inch copper-clad epoxy board. Another copper-clad board, 8½ by 3½ inches, was utilized for the heterodyne oscillator, r.f. amplifier, high-frequency mixer, and the v.f.o. tuning capacitor. The a.g.c. amplifier, S-meter amplifier, d.c. regulator, and 100-ke. calibrator were located on the two boards where it was found convenient, since their associated circuitry was not critical with placement. Each stage was constructed and tested individually before going on to the next stage. The v.f.o. used the same construction, and likewise was tested and corrected for stability as previously described.

Most of the components were mounted above the boards and their leads interconnected either by direct wiring or through terminals beneath the board. For most components the copper-clad board was drilled to accept the wire size of each lead, and then countersunk by hand with a larger size drill just deep enough to remove the copper foil around the hole, to prevent shorting. Ground connections were soldered directly to the copper surface. The boards were mounted to the 11- by 9-inch cut-out chassis after most of the individual stage construction and testing had been finished.

The metal boxes and shields (other than for the v.f.o.) visible in the photographs were used as a precautionary measure rather than from necessity. However, it was considered good practice. With the close spacing of components and wiring, care was taken in placement of the various r.f. stages to minimize undesired coupling.

The band switch, S_{22} , is actually three separate ceramic rotary assemblies ganged together. The first section, using a single wafer, is mounted on a bracket placing the wafer 2 inches behind the panel. A two-wafer assembly, for the presclector, is similarly mounted in line with the first so that its front wafer is 4 inches away from the single wafer; the shafts of the two switches are ganged with metal tubing and set screws. The third assembly also has two wafers, separated 3 inches from the second section; this assembly (in the heterodyne oscillator circuit) is similarly ganged to the second switch.

Only those transistor types that were available for use and were either directly substituted or found suitable for operation, after appropriate base-bias adjustments were made, are indicated in the diagrams. The variety of transistors used indicates that many other types

of small-signal high-frequency transistors can be effectively used. Whatever the types chosen, the base-bias resistors should be adjusted individually for best performance, even for transistors nominally of the same type, since the operating characteristics do vary somewhat from one unit to another. The utilization of transistor sockets greatly simplified circuit testing. For mobile operation, soldered-in transistors would probably be desirable, but good-quality sockets have proven reliable.

Conventional precautions were taken concerning transistor handling, heat, applying voltages, polarity, and so on, during the construction and testing. Even with these normal precautions 8 transistors were destroyed by sheer negligence, because of a variety of circumstances—including accidentally applying the full r.f. output of the linear directly to the emitter of the receiving r.f. amplifier.

As a side note, after all of the transistor circuitry was completed and working it was noticed that the copper-clad boards had begun to tarnish heavily. An attempt was made to remove the tarnish with alcohol and a detergent. This proved catastrophic. Whatever the chemical reaction that took place, within days corrosion crept over the boards and under components until it appeared that the entire transceiver would have to be scrapped. As a last resort, the entire transceiver was immersed in a tub of soap and water, carefully washed, scrubbed and then rinsed. Then the chassis was placed in a 250-degree oven to bake out. The copper-clad boards with the components were later sprayed with clear Krylon. Corrosion is no longer evident.

Overall performance of the transceiver has been very good. It has been operated on all bands and modes, with gratifying reports. Single-tone power output into a Byrd wattmeter indicates 52 watts minimum output on 80 through 15 meters and 45 watts on 10 meters. Receiver sensitivity, while not accurately measured, compares favorably with that of the home-station receiver, and the set has been used as a "second receiver" for DX chasing. A few birdies are evident in the receiver, but only two of these are bothersome, falling in the phone portion of the 15- and 10-meter bands. All others are of very low amplitude and barely discernible. An exceptionally strong adjacent-channel local station will produce cross-modulation, but this can be controlled to a certain degree by the r.f. gain control, and the effect is not serious unless the desired station is very weak. No doubt an FET r.f. amplifier would solve this situation, and it is planned eventually to replace the existing r.f. amplifier. The low current drain in the "receive only" mode is a decided advantage, since automobile battery drain can be considered negligible. There was no need for any special noise suppression for mobile operation, thanks to the substantial amount of filtering used in the d.c. regulator input circuit.

» Getting set up for operation on the 160-meter band is not difficult to do if one already has a s.s.b./c.w. transceiver for the 75-meter band. This article describes an easy-to-build transceiving converter that will work with most commercial transceivers. It has a peak power output of approximately 25 watts and can be used on c.w. and s.s.b. to cover the range from 1.8 to 2.0 MHz.

A Transceiving Converter for "160"

DOUG DEMAW, WICER

There is a good chance that the reader has been wanting to try his hand at "top-band" operation now that the privileges on 1.8 MHz. have been expanded. Also, since s.s.b. operation has become legal on the 160-meter band, existing equipment can be placed in operation to enable the user to visit this interesting band.

For quite some time it has been practical to generate s.s.b. signals in the v.h.f. and u.h.f. regions of the spectrum by using transmitting converters in combination with an existing 14- or 28-MHz. s.s.b. transmitter. The low-band transmitter signal is taken at low power (usually under 5 watts) and mixed with a crystal-controlled oscillator signal to produce the desired *sum* frequency, e.g., a 14-MHz. s.s.b. signal is beat with a crystal-controlled 130-MHz. signal to produce 144-MHz. s.s.b. energy. Getting from the 75-meter band to 1.8 MHz. can be done in a like manner by using the *difference* frequency of a 5800-kHz. crystal-controlled oscillator and that of a 3.8-MHz. s.s.b. transceiver. This combination results in a frequency of 2000 kHz. Moving the transceiver's frequency to 4.0 MHz. results in a difference frequency of 1.8 MHz., the low end of the 160-meter band. This method is used with the simple 3-tube circuit described here. Receiving is handled in the same manner, beating the incoming 1.8-MHz. signal with the 5800-kHz. energy to produce an i.f. of 4 MHz., thus utilizing the 75-meter transceiver's receiver section for listening to the 160-meter signals.

Circuit Data

Looking at the circuit of Fig. 1, V_{1A} operates as a crystal-controlled oscillator to produce a 5800-kHz. local-oscillator signal for transmitting and receiving. This stage operates continuously. Output from V_{1A} is fed to the transmitting mixer, V_{1B} , and to the receiving mixer, V_3 . V_{1B} is turned off by means of K_{1C} , the changeover relay, during receive. During transmit, 3.5-MHz. s.s.b. or c.w. energy is supplied to the cathode of V_{1B} , across a 470-ohm resistor. This is mixed with the 5800-kHz. local-oscillator output at V_{1B} and results in a 160-meter signal at the output of V_{1B} . A high- Q tuned circuit couples the mixer output to the

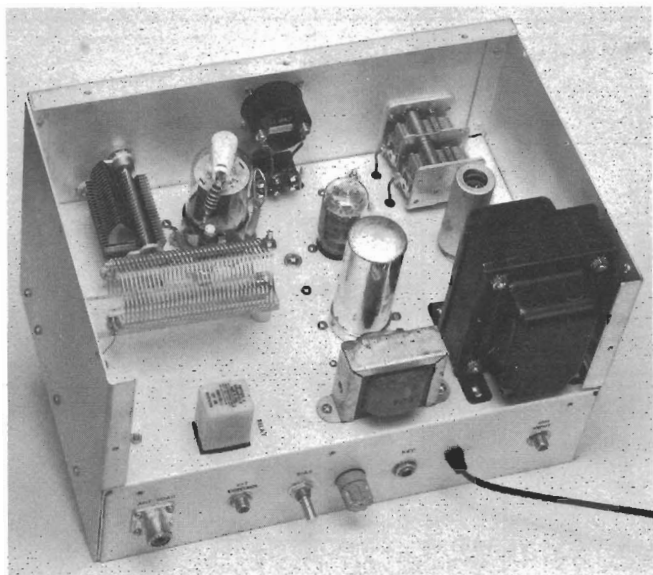


The transceiving converter is housed in a homemade aluminum cabinet which measures $8 \times 8 \times 12$ inches. Perforated aluminum is used for the top and back sides of the cabinet to assure good ventilation.

grid of the power amplifier, V_2 . The 6146B p.a. stage amplifies the 1.8-MHz. signal input power is approximately 35 watts p.e.p.

During receive the local-oscillator energy is fed to the receiving mixer grid (V_3) and beats with the incoming 160-meter signal to produce a receiving i.f. of 3.5 to 4 MHz., depending upon the dial setting of the 75-meter transceiver. Output from the mixer is routed to the transceiver through K_{1A} and J_1 . During transmit, V_3 is turned off by K_{1C} . A double-tuned high- Q input circuit is used at V_3 to reduce images, and to lessen the chances of front-end overload from strong local b.c. stations. A hand-pass tuned circuit is used at the output of V_3 to assure that only the desired i.f. signal reaches the input of the 75-meter transceiver.

The metering circuit reads plate current—200 ma. full scale—by measuring the voltage drop across a 10-ohm 5-percent resistor, R_2 . The 2000-ohm 5-percent metering resistor, R_3 , provides a full-scale meter reading of 2 volts, corresponding to 200 ma. of current flow through R_1 . M_1 is a 0 to 1-ma. instrument. It reads relative r.f. output voltage when S_1 is switched to R.F. A resistive divider is connected to the output line of the p.a. stage and CR_1 rectifies



Looking into the top of the transceiving converter, the power supply is at the lower right. Directly ahead of the power transformer is the receiving mixer, V_3 , and its tuning capacitor, C_6 . V_1 is to the left of V_3 , just ahead of the filter capacitor. The p.a. section of the unit is at the upper left. C_5 is below the chassis, directly under C_4 . C_3 , the neutralizing wire, is encased in spaghetti tubing and is visible adjacent to the 6146B tube. Relay K_1 is at the lower left.

the r.f. which appears at the junction of the two resistors. A 22,000-ohm "linearizing" resistor helps to make the meter respond more uniformly to the changes in r.f. voltage. If greater accuracy is desired for the plate-metering circuit, 1-percent resistors can be used at R_2 and R_3 , though the 5-percent resistors should be suitable for this application.

A probe-type neutralizing circuit is used at V_2 . C_3 is actually a stiff piece of bus wire, three inches in length, which is fed through the chassis by means of an insulating bushing. The wire is placed adjacent to the tube's anode, and is in the same plane as the anode. It is moved to and from the tube envelope to vary the capacitance between it and the tube plate. Adjustment of C_3 is discussed later.

Construction

An aluminum chassis which measures $12 \times 8 \times 2\frac{1}{2}$ inches is used as the base for this equipment. A home-made panel and cabinet is used to enclose the unit. The panel is 8 inches high and is 12 inches wide. The top is fashioned from perforated aluminum material which was obtained from the hardware store (Reynolds aluminum).

The layout should be apparent from the accompanying photographs. All long runs of r.f. wiring should be made with subminiature coax cable (RG-174/U), grounding the shield braid at each end of the cable.

Checkout and Tune Up

Some provision should be made to reduce the power output of the 75-meter transceiver to be used with this equipment. No more than 5 watts of drive should be necessary; too much drive can damage V_{1B} . Approximately 30 r.f. volts will appear between the transmitting mixer cathode and ground when normal 3.8-MHz. drive is applied. Some transceivers are capable

of supplying sufficient output on 3.8 MHz. by removing the screen voltage from the p.a. stage. Or, it may be practical to disable the p.a. and borrow some output from the driver stage by means of link coupling. The stout-of-heart may wish to merely turn down the speech gain of the transceiver until the desired power level is reached. This method was used in the ARRL lab while working with a KWM-2, but could lead to disaster if the audio level was inadvertently turned up beyond the desired point.

Before testing the 160-meter unit, make sure that the changeover relay, K_1 , is connected to the remote keying terminals of the 75-meter equipment by means of J_1 . Then, connect a 160-meter antenna to J_2 and listen for 160-meter signals, atmospheric noise, or Loran pulses. Peak the incoming signals by means of C_6 . For reception on the low end of the 160-meter band, C_6 should be almost fully meshed. The slugs of L_4 and L_5 should then be adjusted for best signal response. When receiving near the high end of the band, C_6 should be near midrange. Coils L_6 and L_7 form a bandpass circuit and should be stagger-tuned to give uniform response across any desired segment of the 160-meter band, e.g., 1800 to 1900 kHz., or 1900 to 2000 kHz. If the receiving section is performing properly, one should be able to copy a 0.3- μ v. c.w. signal without difficulty in areas where minimum atmospheric and man-made noise levels prevail. Ordinarily, however, noise levels prevent such weak-signal reception. If no signals can be heard, check V_{1A} to make certain it is working properly. The 5800-kHz. signal can be monitored on a general-coverage receiver to determine if the oscillator is operating.

Attach a 50-ohm dummy load to J_2 before testing the transmitter section of the equipment. Set R_1 for a resting plate current of 25 ma. for V_2 . This adjustment should be made without drive applied at J_1 , but with K_1 energized. Next,

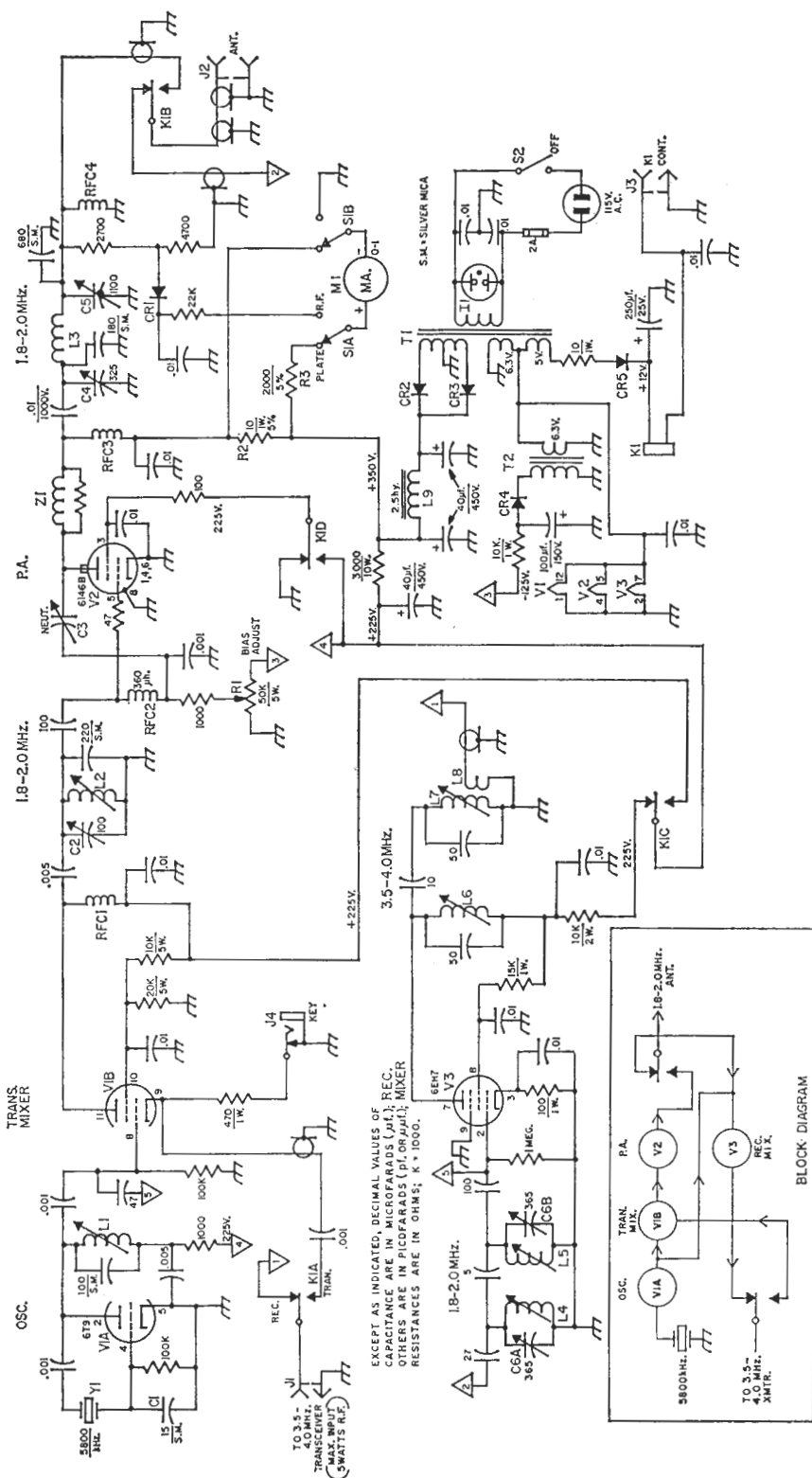


FIG. 1—Schematic diagram of the 160-meter equipment. Fixed decimal-value capacitors are disk ceramic unless otherwise noted. Polarized capacitors are electrolytic. Fixed-value resistors are $\frac{1}{2}$ -watt composition unless indicated otherwise. A block diagram is included to show the direction of signal flow.

- C₁—Feedback capacitor. (May require slightly more or less capacitance, experimentally, for best oscillator starting.)
- C₂—100-pf. variable (Hammarlund HFA-100A).
- C₃—See text.
- C₄—325-pf. variable (Hammarlund MC-325M).
- C₅—3-section broadcast-type variable, all sections in parallel (J. W. Miller 2113).
- C₆—Two section broadcast-type variable (J. W. Miller 2112).
- CR₁—1N34A germanium diode.
- CR₂, CR₃—1000 p.r.v., 1-ampere silicon diode.
- CR₄—600 p.r.v., 750 ma. silicon diode.
- CR₅—50 p.r.v., 2-ampere silicon diode.
- J₁—115-v.a.c. neon indicator (part of S₂).
- J₂, J₃—RCA phono connector.
- J₄—SO-239 style coax connector.
- J₅—Closed-circuit phone jack.
- K₁—4-pole double-throw 12-volt d.c. relay (Polter & Brumfield KHP17D11).
- L₁—5- to 8- μ h. adjustable inductor (J. W. Miller 21A-686RB1).
- L₂—12.9- to 27.5- μ h. adjustable inductor (J. W. Miller 42A225CB1).
- L₃—20- μ h. inductor; 35 turns No. 18 wire; spaced one wire diameter between turns, $1\frac{1}{2}$ inch diameter. Use 35 turns of Polycoid No. 1759 inductor.
- L₄, L₅—12.9- to 17.5- μ h. variable inductor (J. W. Miller 42A225CB1).
- L₆, L₇—23.8- to 39.6- μ h. adjustable inductor (J. W. Miller 21A335RB1). J. W. Miller Co., 5917 S. Main St., Los Angeles, Calif. 90003.
- L₈—6 turns small-diameter insulated wire wound over ground end of L₇.
- L₉—2.5-hy. 100-ma. filter choke.
- M₁—0 to 1-ma. d.c. panel meter.
- R₁—50,000-ohm, linear-taper, 5-watt control.
- R₂, R₃—See text.
- RFC₁—1-mh., 75-ma. r.f. choke (National R-50 or equiv.).
- RFC₂—360- μ h. r.f. choke (Millen J300-360 suitable).
- RFC₃, RFC₄—2.5-mh., 250-ma. r.f. choke (Millen 34102).
- S₁—D.p.d.t. toggle.
- S₂—S.p.s.t. rocket switch with built-in pilot lamp (Carling Electric Co. Type LT1LA50). Carling Electric Co., 505 New Park Ave., West Hartford, Conn. 06110 (catalog available).
- T₁—Power transformer. 540 volts c.t. at 120 ma., 5 volts at 3 amps., 6.3 volts at 3.5 amps. (Allied-Knight 54C1466 or equivalent).
- T₂—6.3-volt, 1-amp. filament transformer, reverse connected.
- Y₁—5800-kHz. fundamental-type crystal (International Crystal Co.).
- Z₁—Parasitic suppressor; 5 turns No. 18 wire over body of 47 ohm, 1-watt resistor.

apply approximately 2 watts of 3.8-MHz. (carrier) drive at J₁. Switch S₁ to read r.f. voltage, then tune C₂, C₄, and C₅ for maximum meter reading. Next, L₁ can be peaked for maximum oscillator output, while still observing the meter. After the foregoing adjustments are made monitor the plate current and tune for a dip the p.a. plate current by adjusting C₄. C₅ is the loading control, and it should be adjusted so that the dip in plate current is rather broad to assure tight coupling to the antenna—necessary if a good-quality signal is to be had. When the p.a. is properly adjusted the plate current should be approximately 100 ma.

If the 6146B stage is stable there will be no changes in plate current, other than the normal dip, as C₄ is tuned through its range. If additional peaks or dips occur, adjust the spacing between the neutralizing wire and the tube's anode until no instability is noted. With the drive disconnected from J₁, tune C₄ through its range and observe the plate current. Only the resting plate current should be registered if the amplifier is stable. By coupling a sensitive wavemeter to L₃ during the latter test,¹ self-oscillation will be apparent as r.f. output when C₄ is tuned. Fine adjustments to C₃ can then be made until no spurious output is noted.

When operating c.w., insert sufficient carrier to bring the p.a. plate current up to 100 ma. at dip. The key can be plugged into the exciter's key jack, or into J₄. Since K₁ is not designed for highspeed keying, it might be best to use J₄ as the keying terminal.

Final Comments

It should go without saying that the true measure of any ham station's performance can be taken from its antenna system. This is as true for 160-meter operations as it is for any other band. A random-length wire will usually give random results; a good antenna will give good results when used with good equipment. A quarter-wavelength vertical antenna, worked against a good ground system (even if the vertical element is physically short and uses lumped inductance to achieve resonance) will give good results. If space permits, a half-wave dipole, as high in the air as possible, will do an excellent job. Good results can sometimes be obtained by using an end-fed horizontal quarter-wavelength wire, as high in the air as possible. The latter should be worked against a good earth ground, and the more of the wire that is vertically oriented (current end) the better.

¹"Are You Putting Out On The Correct Band?" QST, March 1967, p. 25.

» If you are stuck with a monoband transceiver for the 80-meter band, this article shows how you can get on 10 and 15 with relatively little trouble and expense.

The transverter is built on a 7 × 11 × 2-inch chassis with bottom cover. A 6 × 4 × 5-inch utility box (LMB) houses the final amplifier. Along the chassis, from left to right, are the controls for the band switch (S_1), drive control (R_1), meter switch, and the receiving tuning capacitor (C_1). Of the two log-gle switches to the right, the upper one controls the external transverter power supply, while the lower one controls the two change-over relays. The pi-network control and band switch are on the front of the amplifier compartment. The screened ventilation hole in the top cover of the box is located immediately above the 6146.



A 21/28-MHz Transverter for 3.5-MHz Transceivers

DENNIS M. PETRICH, KØE00

Many owners of 80-meter transceivers are interested in operating on the higher-frequency bands from time to time, but do not feel that the cost of a new five-band rig is justified. Finding myself in this position, the problem was solved by building the transverter shown on these pages. With this unit added to the monobander, I can cover both 10 and 15. With a good junk box at hand, the cost was only about thirty dollars.

Fig. 2 shows a block diagram of the transverter. On receiver it operates as a conventional converter. A 21-MHz. signal, for instance, is fed to a 6AK5 r.f. amplifier, and then to a 6J6 mixer, where it is combined with a 25-MHz. signal from the oscillator to produce a 4-MHz. signal at the input of the transceiver. For a 28-MHz. signal, the oscillator frequency is 32 MHz., which again results in 4-MHz. mixer output. On transmit, the 4-MHz. signal from the transceiver is fed to a 12AU7 mixer/Class A amplifier where it is combined with the 25-MHz. (or 32-MHz.) signal from the oscillator to produce a 21-MHz. (or 28-MHz.) signal at the input of the 6CL6 driver. The final is 6146 AB₁ linear operating at about 70 watts input. The 10-meter band is covered in steps of 500

kHz., the oscillator frequency being changed appropriately.

It should be pointed out here that the tuning of the transceiver when working into the transverter will be in a direction opposite to normal. When the transceiver is tuned to 4 MHz., the received and transmitted signals will be at the low edge of whatever 500-kHz. segment has been selected; when the transceiver is tuned to 3.5 MHz., the received and transmitted signals will be at the high-frequency end of the chosen segment. Similarly, lower-sideband transceiver operation results in upper-sideband output from the transverter, a fortunate circumstance for those having 80-meter transceivers providing only lower-sideband output.

The complete r.f. circuit is shown in Fig. 1. The input and output circuits of the receiver r.f. amplifier are gang-tuned by a dual-section capacitor, C_1 . A double-tuned bandpass circuit couples the mixer to the input of the transceiver. The oscillator is crystal-controlled, the proper crystal being selected by S_1 . The output circuit of the oscillator is tuned to twice the crystal frequency in all cases.

On transmitting, the 3.5-MHz. signal from the transceiver is fed to the mixer cathode through an attenuating network that reduces the drive to a suitable level. The 50-ohm resistor

provides a proper load for the transceiver. The mixer circuit may be a bit unorthodox, but it proved to be the best of several tried.

The output circuits of the transmitter mixer and driver are individually tuned by C_3 and C_4 . The capacitors in these circuits, as well as those in the receiver r.f. amplifier stage, have sufficient range to cover both 21 MHz. and 28 MHz., so coil switching is required only in the oscillator. The final-amplifier output circuit is a pi network suitable for matching 50-ohm antenna systems. The 6146 is neutralized by the capacitive bridge method. C_5 is the neutralizing capacitor. Drive to the final amplifier can be adjusted by means of R_2 in the cathode circuit of the driver.

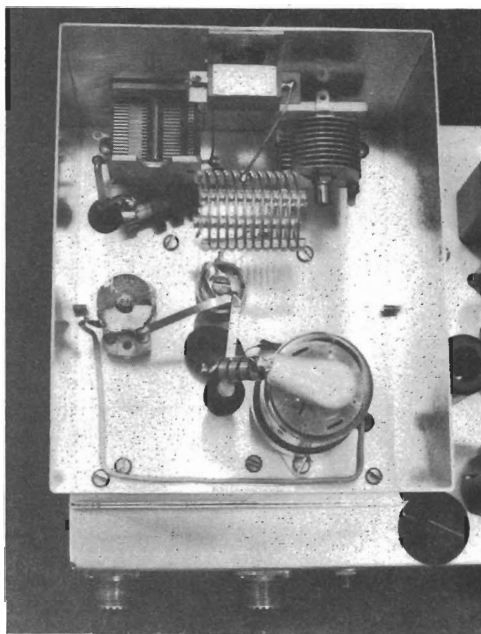
Final-amplifier grid, screen, or plate current can be monitored by meter M_1 . The multiplying resistances shown result in full-scale readings of 5, 25 and 200 ma., respectively, with 1000 ohms in series with the meter.

The transmit-receive change-over functions are performed by two relays, K_1 and K_2 . K_1 shifts the input and output r.f. connections. K_2 applies 180 volts regulated to the receiving mixer and r.f. amplifier, and cutoff bias to the grid of the final amplifier on receive. On transmit, the relay applies 325 volts to the transmitter mixer/amplifier and driver, and cutoff bias to the receiver r.f. amplifier and mixer.

The power-supply diagram is shown in Fig. 3. A silicon-diode bridge rectifier with center tap provides 700 volts for the final-amplifier plate, and 325 volts for the transmitter mixer and driver. A pair of VR tubes in series, operating from the center tap, supply 180 volts, regulated, for the oscillator and receiving tubes, and the screen of the final. A filament transformer in reverse, T_2 , driven by the filament winding on T_1 , provides -120 volts for biasing purposes.

Construction

Most of the necessary constructional details will be found in the photographs and their captions. However, one or two special precautions were found to be necessary. The only difficulty encountered in obtaining satisfactory results

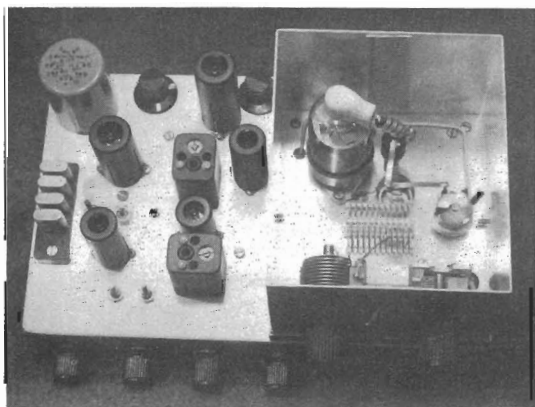


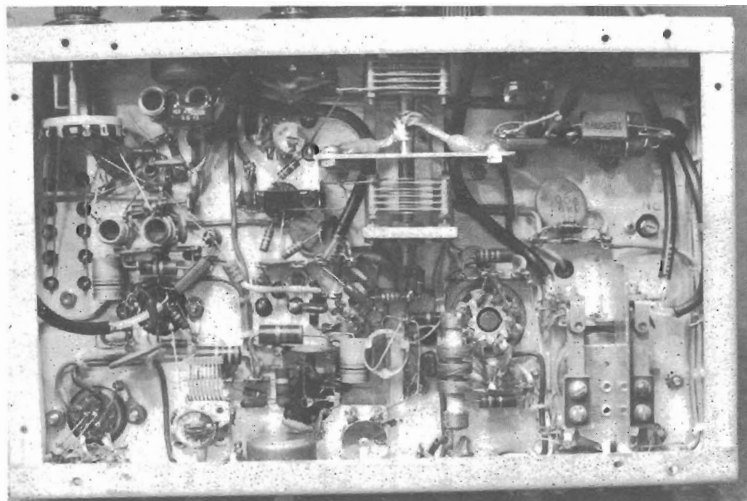
Components of the output pi network are mounted on the front wall of the amplifier compartment, C_8 to the left, C_6 to the right, with the coil and its band switch in between. The neutralizing capacitor is at left center.

from the unit when it was first put into operation was in stabilizing the receiving r.f. amplifier stage. The trouble was finally eliminated by adding pigtailed of flexible wire to ground the center of the rotor shaft, between the two sections of the dual tuning capacitor. It seems probable that this would be unnecessary if two separate capacitors, each with its front- and rear-bearing wiper contacts, were ganged.

The box that I used for the amplifier compartment had a haminertone finish. If this type of box is used, the finish should be scraped away in the appropriate areas to allow good electrical contact between the box walls and the covers, and between the bottom cover and the chassis. R.f. input and output connections were made using RG-58/U to provide a shielded conductor. Aside from these precautions, only the usual

The receiving r.f. amplifier input and output coils (L_5/L_6 and L_4) are in shielding cans mounted directly above their tuning capacitors, which are below chassis. The 6AK5 is between the two coils. The tube to the left of the forward shield can is the receiving mixer. The slugs of the bandpass coils (L_1/L_2 and L_3) are in front of this tube. To the rear of this tube are the tuning slugs of the crystal-oscillator coils, and the oscillator tube. The tube to the left of the rear shield can is the transmitting mixer. Along the rear edge of the chassis are the power-control relay (K_2), and the driver tube (between the input and output tuning controls of this stage).





Under-chassis view. The dual receiving-tuning capacitor is near upper center, with a shield running between the two sections. At the upper left is the crystal-oscillator switch. The pair of coils below the switch are the oscillator coils. Those above the switch are in the receiving output bandpass coupler. The driver input and output coils are below and to the left of the dual capacitor, at right angles to each other. The antenna change-over relay is in the lower right-hand corner.

practice of keeping r.f. connecting leads as short as possible, and keeping low-potential leads down flat against the chassis was followed.

Receiving Adjustment

The wiring should be checked thoroughly before applying power to the transverter. The crystal oscillator should be checked first with power applied, and S_1 in the receiving position. Turn S_1 to the Y_1 position (12.5-MHz. crystal). Using the g.d.o. as an r.f. indicator, adjust the slug of L_7 for maximum response at 25 MHz. Then turn S_1 to the Y_3 position (16.25-MHz. crystal), and adjust L_8 for maximum response at 32.5 MHz. This single adjustment should be satisfactory for all 10-meter crystals.

Next, connect the transverter to your transceiver or receiver, and feed the signal generator to the transverter antenna jack. Tune the generator to 21 MHz., and adjust the receiver to 4 MHz. Turn S_1 to the Y_1 position, and C_1 to maximum capacitance. Adjust L_4 and L_5 to peak the generator signal on the receiver S meter. Be sure to reduce the signal-generator output as the coils are adjusted closer to resonance to avoid overloading the receiver input circuits. Then adjust L_3 for maximum response.



Now move the receiver tuning down to 3.5 MHz., and change the signal-generator frequency to 21.5 MHz. Retune C_1 for maximum signal, and peak L_2 . Repeat the adjustments of L_2 and L_3 once again to make sure that the bandpass coupler is working properly. With an input signal of constant level, there should be no appreciable change in response as the generator is swept across the band.

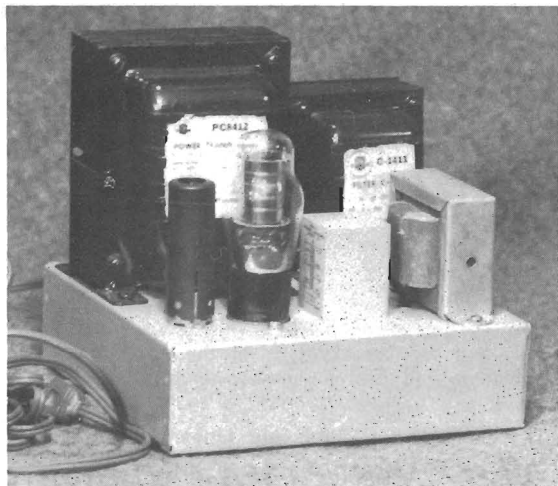
Now retune the signal generator to 29.5 MHz., and adjust your transceiver or receiver to 3.5 MHz. Adjust C_1 of the transverter for maximum receiver signal. Some slight readjustment of L_4 may improve the response at 29.5 MHz. without significant change in the response at 21 MHz.

At this point, the receiver section should be completely aligned, and connecting an antenna to the transverter should yield a barrage of 15- or 10-meter signals when the bands are open.

Transmitter Alignment

To align the transmitting section, place S_1 in the Y_1 position and connect the signal generator, through a 0.005- μ f. capacitor to the cathode (Pin 8) of the 12AU7 mixer. Set the signal generator to 4 MHz., and switch the transverter to transmit. With the signal generator turned off, adjust the final-amplifier bias voltage (by means of R_3) to -51 volts, or for an idling current of 16 ma. Turn the signal generator on, and adjust it for its fullest output at 4 MHz. Advance the drive control to maximum (R_2 shorted out), and peak C_3 for maximum

Along the rear apron of the chassis, from left to right, are input and output r.f. connectors, a phone jack for external power-control connections, the bias-adjusting control, ground terminal and power-supply connector. The phone jack has two insulated terminals, requiring a 3-circuit plug. (For safety, the power connector should be a male type.—Editor)



The power supply is assembled on a $7 \times 7 \times 2$ -inch chassis. The bias transformer is between the OA3 voltage regulator and the low-voltage filter choke. Diodes, filter capacitors, and resistors are inside the chassis.

amplifier plate current. This should occur with C_3 near maximum capacitance (with C_2 adjusted appropriately to make this occur). Next, dip the plate meter with the amplifier tuning capacitor, C_6 . Peaking C_4 should now result in a marked increase in plate current.

Next, shut the transverter off, and disconnect the 700-volt and 180-volt leads from the final. Connect your receiver input to the transverter's antenna connector, and tune the receiver to 21 MHz. At this time, turn the transverter back on, and peak the receiver on the signal being generated by the signal generator, and amplified by the driver. With C_3 and C_4 peaked, and the final-amplifier output circuit tuned to resonance, adjust the neutralizing capacitor, C_5 , for minimum signal on your receiver S meter. The gain of the receiver should be set quite low for this adjustment. The final is now neutralized, and no further adjustment of C_5 should be necessary.

Care should be taken, when peaking C_3 and C_4 to make sure that you aren't tuning up on the crystal-oscillator signal rather than the generator signal. A quick check can be made by turning the signal-generator's output level down. If no change in plate current results, you are tuning up on the wrong signal. Readjusting C_3 and C_4 will solve the problem.

Disconnect the receiver from the antenna terminals of the transverter, and replace it with a 50-ohm dummy load. Reconnect the 700-volt and 180-volt leads to the final amplifier. With

the signal generator still feeding the transmitting mixer, as before, you should now be able to load the final to a plate current of 115 ma., representing an input of 70 watts. Check the screen current to make sure that it's running under 15 ma. The meter should show no grid current for AB_1 operation.

Next, turn S_1 to the Y_4 position (16.5-MHz. crystal), and turn S_3 to the 10-meter position. Adjust the signal generator to 3.5 MHz., and peak C_3 , C_4 and the pi-network tuning capacitor as before. Maximum amplifier plate current should result with C_3 and C_4 near minimum capacitance. Now adjust the pi-network controls for proper loading of the final amplifier. You should now be ready to drive the unit with your transceiver, and the signal generator can be disconnected.

To use the transverter with a transceiver, connect the output of the transceiver to the input of the transverter through a length of coax line, and connect the antenna to the output of the transverter. The transverter requires very little driving power. Most of the transceiver power will be absorbed by the 50-ohm resistor at the input of the transverter. Since this resistor has a rating of only 20 watts, the input to the driving transceiver should be limited by keeping the microphone gain down. If full input to the transceiver is necessary for tuning up, the transceiver should be adjusted first while working into a dummy load, or 80-meter antenna, before applying the signal to the transverter. I use an NCX-3 transceiver, and I have found it satisfactory simply to tune the transceiver by peaking it up on a 21- or 28-MHz. signal on receive, since this operation also tunes the transmitter stages. For the low power required to drive the transverter, accurate tuning isn't necessary.

Now, apply power to the transverter (S_1 in the transmit position). Switch the transceiver to c.w. and back off the carrier insertion. Adjust C_3 and C_4 for a peak in final-amplifier grid current, increasing the carrier insertion if necessary to obtain a small reading. As soon as the driver circuits have been peaked, back off on the carrier insertion until the grid current just disappears. Then adjust the transverter pi network for an input of 70 watts. Then switch to s.s.b., and adjust the microphone gain control until the meter shows a slight flicker on the loudest voice peaks.

When loaded to a plate current of 115 ma., measurement with a Bird wattmeter showed an output of 38 watts. On the first try, several JAs and UAQs were worked with reports of excellent quality on both bands. The transverter is now being used to drive a Thunderbolt linear, and no problems have arisen.

"Did you know that an "ideal" vacuum tube would be a comparatively low-performance linear amplifier? This article goes to the roots of intermodulation distortion and summarizes the present state of the art of tube design.

Intermodulation Distortion in Vacuum Tubes

WILLIAM I. ORR, W6SAI

An ideal linear amplifier is one in which the output envelope amplitude is at all times directly proportional to the input envelope amplitude. Amplitude distortion results when the magnitude of the output signal is not strictly proportional to that of the driving signal. This class of distortion (which is the principal type encountered in linear amplifiers) includes *intermodulation distortion*, a particularly interesting type of amplitude distortion encountered in single-sideband service. In passing, it should be noted that intermodulation distortion (abbreviated IMD) occurs only in a nonlinear device driven by a complex signal having more than one frequency. As speech is made up of multiple tones (or frequencies) and as the perfect linear amplifier has yet to be built, the situation leading to IM distortion exists in most s.s.b. amplifiers. Once the intelligence-bearing signal has been generated, the amplitude relationships existing in the intelligence must be faithfully retained or the s.s.b. signal will blossom into a broad, fuzzy caricature of

itself, and the unlucky user of the nonlinear equipment will find his on-the-air popularity waning.

The Vacuum Tube and Linearity

The vacuum tube is the heart of the linear amplifier, and the amplifier is designed about it.¹ The passive circuit elements—resistors, capacitors, inductors, etc.—are entirely linear and they affect circuit operation only insofar as they determine the operating parameters of the tube. The linearity of the tube is open to question. The more linear the tube, the less stringent the demand placed upon the circuitry to achieve a desired degree of over-all linearity. The results obtained are a balance between excellence and economy.

The vacuum tube utilizes electrons emitted from a hot cathode by impressing upon them an electric field which varies with time. During the passage of the electrons from cathode to plate, the field is manipulated in such a way as to alter the number of electrons arriving at the plate of the tube. The electron flow (or cathode current) is a $3/2$ power function of the applied electrode voltages. This so-called " $3/2$ -power law" of Child and Langmuir is theoretically valid for uniform tube geometry and holds true for any space-charge-limited electron flow under the influence of an external field (Fig. 1). The $3/2$ -power law is not a linear function, and in practical tubes the cathode current is not a straight-line function of grid voltage. Further, practical tubes depart from the $3/2$ -power law to some extent, depending upon tube geometry, space charge, electron interception by grids, and emission limitations.

The relationship between the electric field and cathode-current flow within the tube described by this natural law plays an important role in the establishment of tube linearity. In practical amplifiers, for example, the magnitude relationship between input and output signals is not perfectly constant at all signal levels within a given range. The relationship defining amplifier linearity is termed the *envelope transfer function*, and ideal and typical transfer functions are shown in Fig. 2. The fundamental cause of a non-ideal, nonlinear amplifier transfer function may be traced directly to the

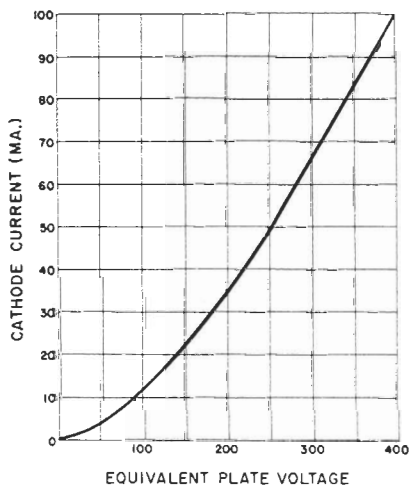


FIG. 1—The electron flow (cathode current) in a vacuum tube is a nonlinear function of the equivalent plate (or plate and screen) voltage and is described by the $3/2$ -power law. This curve illustrates typical electron flow, which plays an important part in establishment of tube linearity.

¹ This discussion applies to vacuum tubes. Similar conclusions may be drawn about transistors, but such conclusions are not within the scope of this article.

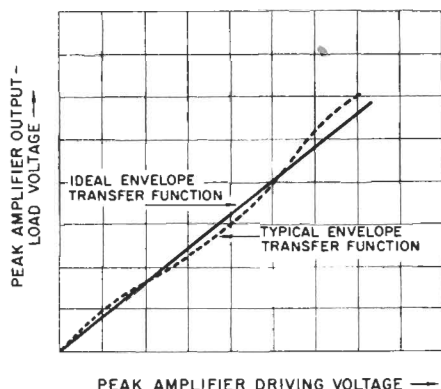


FIG. 2—Amplifier linearity is defined by the envelope transfer function. Departure from linearity is illustrated by curvature of the function (dotted curve) and may be traced directly to the nonlinear relationship between cathode current and electrode voltage shown in Fig. 1.

nonlinear relationship between the plate current and grid voltage of the tube employed in the amplifier. This relationship approximates the 3/2-power law throughout the operating region above cutoff.² An examination of intermodulation distortion reveals the importance of significant cathode-current departure from this fundamental law as regards amplifier linearity.

Measurements made on a wide variety of power tubes, from small to large, filamentary types and oxide cathode, triodes and tetrodes, in grid- and cathode-driven service, have shown conclusively that the magnitudes of the intermodulation distortion products are significantly affected by almost everything: changing heater

²Cutoff may be thought of as that amount of grid bias required to reduce the idling plate current of a vacuum tube to virtually zero.

³"Approximate Intermodulation Distortion Analyses." Report CTR-173 by R. E. Cleary, Collins Radio Co., Cedar Rapids, Iowa; "Linear Power Amplifier Design," W. B. Bruene, *Electronics*, August, 1955; "Linearity Testing Techniques for SSB Equipment," Icenbice and Tellhaver *Proc. I.R.E.*, December, 1956, pages 1775-1782. "Intermodulation Distortion in High Powered Tuned Amplifiers," R. C. Cummings, Consultant, Eitel-McCullough, Inc., San Carlos, California.

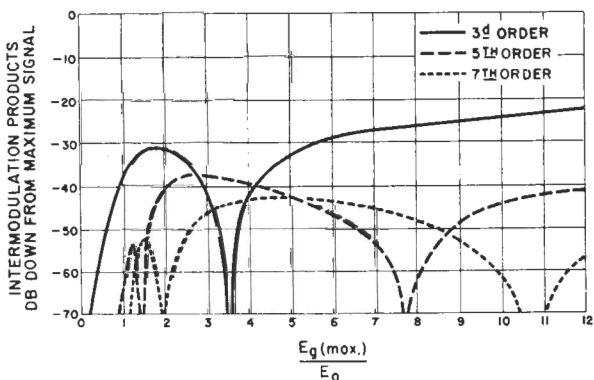
FIG. 3—Intermodulation distortion products may be predicted mathematically. This universal family of IMD curves applies to all perfect tubes obeying the 3/2-power law. The curves are plots of IMD level (Y axis) referred to the driving signal expressed as a ratio of drive to operating bias. As the drive is increased, the various IMD products pass through maxima and minima. Misleading conclusions of amplifier performance may be drawn if the equipment happens to be tested near a cusp on the IMD curve, where a particular product drops to an extremely low level. The whole operating range of the equipment must be examined to draw a true picture of IMD performance.

or filament voltage by only a few per cent; slight shifts in bias voltage, idling current, screen voltage, plate or grid tuning; neutralization, loading—all these factors and others even more obscure enter into the determination of intermodulation distortion.

Mathematical Analysis

IMD products may be calculated by several methods.³ The results of different valid mathematical techniques are in good agreement with each other, and also agree in general with data obtained from two-tone tests conducted with the IMD analyzer. A theoretical family of IMD curves of a perfect tube obeying the 3/2-power law is shown in Fig. 3. This universal family of curves applies to all tubes, regardless of operating parameters or tube type. Changes in electrode potentials and circuit values (and even changes in tube type) will produce characteristic curves of this general configuration, but of course the signal level at which particular value of distortion occurs will be different in each case.

In Fig. 3 intermodulation distortion products, expressed in decibels below the output level of the tube, are plotted along the Y axis. The ratio of a two-tone driving signal $E_{g(max)}$ to operating bias, E_o (relative to cutoff voltage) is plotted along the X axis. When E_o is zero, the tube is biased at cutoff (class B). Ratios of $E_{g(max)}/E_o$ greater than one, but less than infinity, represent the possible range of class AB operation. Starting on the curve at the no-signal point ($E_{g(max)}=0$), the IMD products are nonexistent. As $E_{g(max)}$ is increased, the IM products increase throughout the range of class-A operation and into the class AB region, until a maximum IM distortion figure for the 3rd-order products of about -30.7 decibels is reached at an $E_{g(max)}/E_o$ ratio of about 1.7. The 3rd-order product then drops to zero (minus infinity) again for a ratio of $E_{g(max)}/E_o$ of about 3.5, after which the IM product again increases, gradually rising to a level near -20 decibels for class-B operation. Fifth-order and 7th-order (and higher-order)



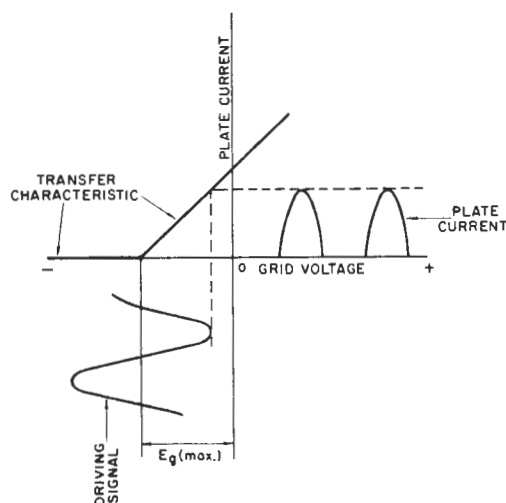


FIG. 4—An ideal tube transfer characteristic departs from the 3/2-power law. The ideal characteristic shown here consists of two linear portions, with the operating point set at the intersection. Half-wave plate current pulses are converted to sine waves by the flywheel effect of the plate tank circuit. Poor tank circuit Q , therefore, will have adverse effect on over-all linearity.

products follow this same general behavior, compressed along the X-axis, and are shown in dotted lines on the graph.

The results of this theoretical study indicate that the perfect 3/2-power tube will provide 3rd-order IM products no better than -20 to -30 decibels below maximum power output, and that the IM product varies markedly with drive level, dropping to zero at various points in the dynamic operating range. Thus, the perfect tube, obeying a fundamental law of physics, is a mediocre performer from a linearity point of view. As far as IM distortion goes, it is a poor device to use in equipment designed for linear amplification of intelligence-bearing signals.

Practical Linear Amplifier Tubes

Does this theoretical study actually mean that all tubes are poor linear amplifiers or that it is impossible to achieve IM distortion products of a better order than -20 decibels? Not at all. The study concerns itself with a *perfect* tube that implicitly follows the 3/2-power law. Of course, there is no such device, and *practical* tubes (i.e.: tubes that can be manufactured) depart from this law to a greater or lesser extent. The practical tube, in general, shows an improvement in over-all linearity as a result of departure from the 3/2-power law. The practical tube, in addition, does not have a definite value of cutoff grid voltage, it does not have constant amplification at all points within the structure, and current deviations and amplification variations occur with changes in

plate voltage. Current intercepted by the screen and control grids modifies the plate characteristic, and the "constants" that express the 3/2-power law vary with actual operating conditions. Theoretically, IM distortion as a result of this law should be independent of tube type. We know from experimental data that such is really not the case, as practical tubes exhibit transfer characteristics departing markedly from the 3/2-power law. In many instances, an improvement in linearity occurs when the tube departs from this law. For example, an ideal transfer characteristic for a tuned amplifier is shown in Fig. 4, consisting of two linear portions with the operating point set at the intersection. The resulting plate current consists of rectified and amplified half sine waves, the plate tank circuit converting this misshapen wave into an equivalent sine wave by virtue of the fly-wheel effect. The equivalent sine wave is directly proportional to the input signal at all amplitude levels from zero to the maximum value shown.

Alternatively, distortionless linear amplification may be achieved from another transfer characteristic having, instead of the discontinuity exhibited in the first example, a smooth curve of the form shown in Fig. 5. The operating point of the tube is chosen at projected cutoff. Ideally, the curved portion of the transfer characteristic should be a portion of a so-called "second-order" curve (a half-parabola, to be exact). A characteristic such as this is

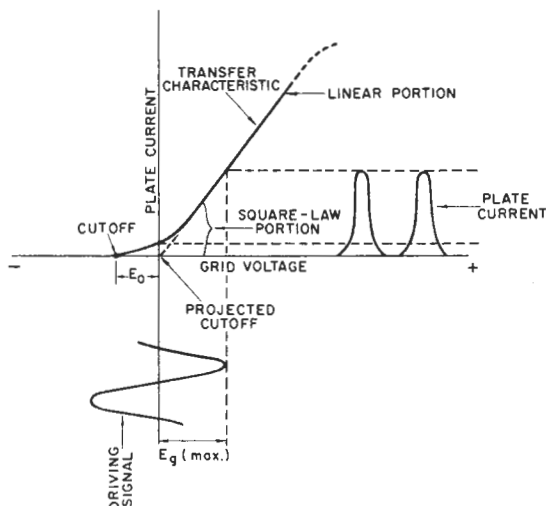


FIG. 5—Another ideal transfer characteristic for a linear tube consists of this form of curve, where the central portion is straight and the lower portion resembles a parabola. Practical tubes exhibit transfer characteristics of this general class, the upper portion of the curve showing additional curvature resulting from saturation of the electron stream in the grid-plate area of the tube. Plate current pulses are converted to sine waves by flywheel action of plate tank circuit.

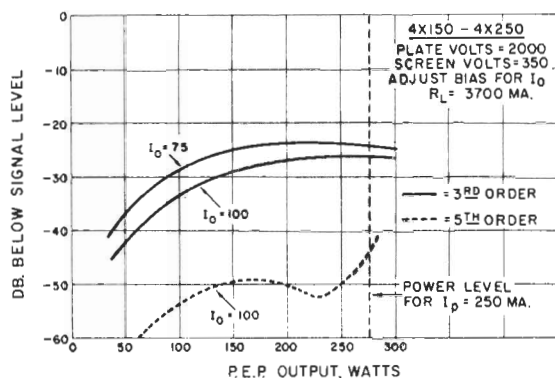


FIG. 6—A family of IMD curves for the 4X150-4X250 external-anode tubes. These curves are representative of this type of tube, and are typical for tubes made by different manufacturers. Intermodulation distortion products average about -25 decibels below peak signal, for 3rd-order products, while 5th-order products average -43 decibels below peak signal. These curves are representative of most small transmitting tubes of this type. Changes in loading or circuit parameters will alter shape and position of the curves.

termed *square law*. Distortion products added to the exciting signal by such a curvature can be filtered out of the output signal by the tuned plate tank circuit *because all of these products fall in the harmonic regions of the exciting signal*. A distortionless replica of the input signal is thus available at the output circuit of the amplifier. Other transfer characteristics exist which also will provide lower-distortion output.

A practical tube may have a transfer-characteristic exponent falling somewhere between 1.5 (3/2-power law) and 2 (square law); its transfer characteristic would approximate the curve of Fig. 5, wherein the central portion is fairly linear and the lower portion resembles a parabola. The upper portion of the characteristic may show additional curvature resulting from saturation of the electron stream in the grid-plate area of the tube. That is to say, the grid or screen "robs" the plate of the greater portion of the available electrons and causes a corresponding drop in plate current.

Intermodulation tests run on tubes having this general transfer characteristic show distortion products generally in agreement with the 3/2-power law. Shown in Fig. 6 are IM curves based upon typical measurements made on the 4X150-4CX250 family of external-anode tubes. With fixed values of plate and screen potential and plate load impedance, measurements were made at two levels of resting plate current over the operating range of the tube. At the recommended value of resting plate current, the 3rd-order IM products rise gradually and smoothly as power is increased to the maximum value of 500 watts (referred to a single-tone plate current of 250 ma.) until at this value the products reach a level of -26 db. below the p.e.p. signal. Decreasing the resting plate current to 75 ma. will degrade the IM curve by several decibels, as shown. Fifth-order products at the recommended value of plate current are below -43 db. at maximum plate current level. The addition of 10 decibels of negative feedback to a circuit employing this style of tube will reduce the IM products below the values shown by

approximately 10 db., so equipment with feedback designed around this tube (other factors being equal) should be able to reach the region of -35 db. IM distortion at full power. Individual tubes (and similar tubes made by different manufacturers) will vary from these curves by two to three decibels. Fig. 7 shows the variation in IM products between three tubes under fixed operating conditions. Changes in loading or other parameters will alter the shape and position of these curves.

Referring back to Fig. 3, tubes of this type are operated under conditions corresponding to a ratio of $E_{g(max)}/E_0$ in the range of 2 to 3 at maximum signal, and therefore distortion must pass through the third-order product maximum of about -31 db. within the operating range. Actually, maximum distortion appears near the 70% to 100% power level and is of the order of -25 db. or so. These curves are quite representative of most power tubes employed in amateur equipment, common varieties of transmitting tubes falling in the -20 to -30 decibel intermodulation range. Judicious use of feedback with these tubes will allow IM distortion products to fall in the -30 to -40 decibel range.

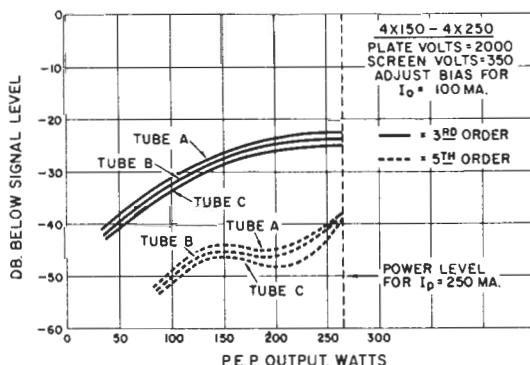


FIG. 7—Intermodulation distortion products vary from tube to tube of the same type, and also vary tube to tube as operating conditions are changed. Small "receiving-type" transmitting tubes are usually poorer than these curves by five to ten decibels.

» Putting TV sweep tubes to work in linear amplifiers takes a bit of doing if maximum efficiency and reasonable life are to be realized. This article focuses on some techniques that can be applied to all TV sweep tubes suitable for use in linear amplifiers—the guidelines for smoother operation and fewer tube failures.

Sweep-Tube Linear-Amplifier Design

DOUG DeMAW, W1CER

It wasn't too many years ago that some of us who were s.s.b. experimenters learned that a buck could be saved by "operating" on certain brands of 1625-type tubes, thus making them suitable for use in grounded-grid amplifiers. It was necessary to saw a window in the tube base, locate and separate the cathode and beam-forming plate leads, and bring them out to individual base pins. At roughly 25 cents apiece, surplus prices, it was no strain on the pin-money fund to put four of them to work in parallel and run them at approximately 500 watts d.c. input. In fact, many a "pair of shoes" was built along such lines and driven by exciters capable of 20 to 50 watts peak output. Little thought, if any, was given to the matter of proper balancing of the tubes in order to assure equal load sharing. Similarly, little attention was given to tube cooling. After all, at 25 cents per copy, it didn't matter too much is one had to replace a set of tubes every three or four months.

Things have changed considerably since the early 1950s, at least where the matter of tubes for use in linears is concerned. The emphasis has been on the use of big, somewhat-expensive tubes of the 4-1000A, 3-400Z, and 4CX1000A

species. These tubes can provide the legal maximum-power levels for amateur operation, while performing with minimum distortion, but require costly hardware and expensive power supplies to reach the 100-watt d.c. input-power level. TV horizontal-sweep tubes, especially those intended for color receivers,¹ offer the builder a low-cost approach to linear-amplifier design while still permitting reasonable signal quality and power output.² This article is based on tests performed on an experimental 800-watt-d.c.-input linear amplifier. Four GE 6KD6 tubes are used in the grounded-grid, parallel-connected arrangement of Fig. 1.

Sweep-Tube Ratings

Although there are many ways to rate a vacuum tube's performance capabilities, we will confine our discussion here to the manufacturers' *design-maximum ratings*. Under this heading we can find the limiting values of operating and environmental conditions which apply to what the manufacturers call a "bogey" tube—a tube of a given type, with average characteristics. These ratings are set to assure reasonable tube life in television receivers. Any departure from them is at the designer's risk, and will depend upon the life he expects from the tube in his particular circuit.

Table I lists the design-maximum ratings of the 6KD6 tube as given by General Electric Co. (A check of the tube manual showed that RCA rates their 6KD6s the same way.)

Since 990 volts is listed as the recommended maximum, a 900-volt d.c. supply was chosen to power the amplifier. This voltage permits the

From July, 1968, QST.



¹ A number of TV sweep tubes are used in commercial s.s.b. ham equipment. Some of them are: 6DQ5, 6GB5, 6GE5, 6HF5, 6JS6, 6JE6, 6KG6, 6KD6, and 6LQ6. They have different maximum ratings and require different operating voltages, but they are all candidates for use in linear amplifiers.

² A rundown on intermodulation (IMD) distortion products, and power ratings for TV sweep tubes as linear amplifiers, was given by Bill Orr, W6SAI, and Bob Sutherland, W6UOV, in *Ham Radio*, April 1968, page 9.

This experimental sweep-tube linear amplifier uses four Shurite edgewise plate meters to allow each of the four 6KD6 tubes to be independently monitored as discussed in the text.

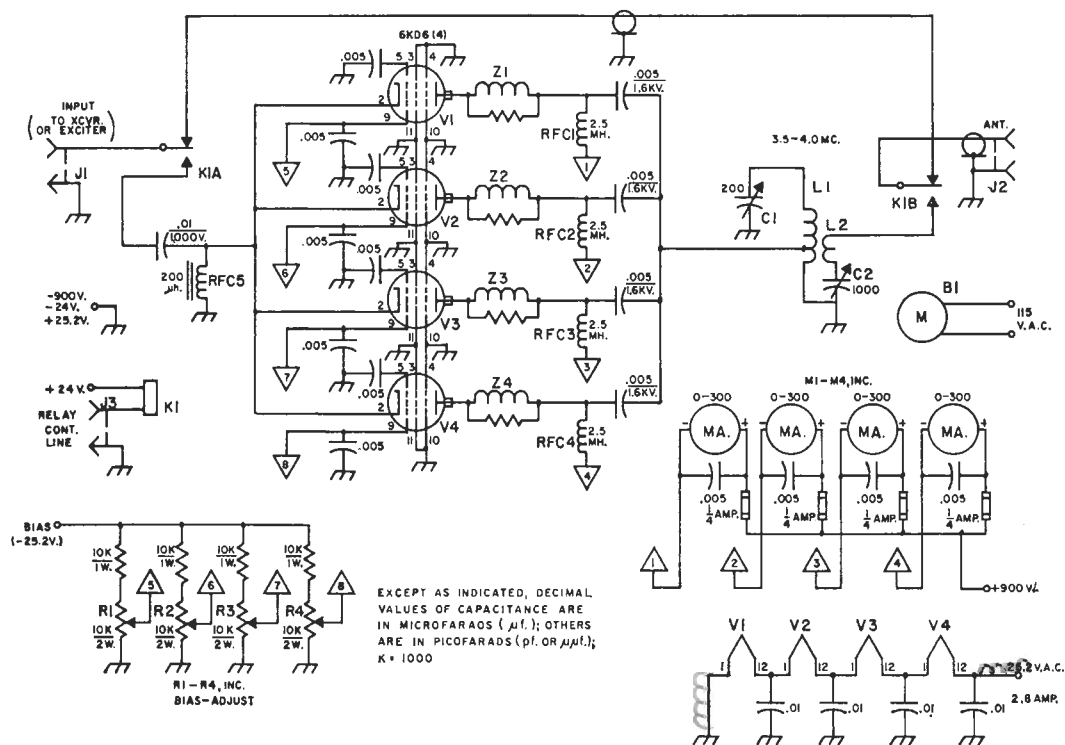


FIG. 1—Schematic diagram of the experimental 800-watt sweep-tube amplifier. Capacitors of decimal value are disk ceramic. Fixed-value resistors are composition. K_1 permits the exciter or transceiver to be switched through the amplifier for low-power operation, and so that the operator can receive through the amplifier.

B_1 —High-speed cooling fan (see text).

C_1 —Transmitting variable (from Command Set transmitter).

C_2 —Three-section broadcast-type variable, all sections in parallel.

J_1, J_3 —Phono jack.

J_2 —Coax fitting, chassis-mounting (SO-239).

K_1 —D.p.d.t. 24-volt d.c. relay, 10-ampere contacts.

L_2 —18 turns No. 12 wire, $2\frac{1}{2}$ -inch dia., 3 inches long (made from B & W 3900-1 coil stock, or Polycoids 1774). Tap at approximate mid-point.

L_1 —5 turns No. 14 wire, 3-inch diameter, approximately $\frac{3}{4}$ inch long (or 5 turns from Polycoids 1779 stock). Mount over ground end of L_1 .

M_1-M_4 , inc.—0-300 ma. d.c. meter (Shurite edge-wise meters used in this model).

R_1-R_4 , inc.—10,000-ohm linear-taper control (Ohmite 2-watt carbon type, or equal).

RFC_1-RFC_4 , inc.—2.5 mh., 250 ma. (Millen 34102).

RFC_5 —200 μ h. (see text for particulars).

Z_1-Z_4 , inc.—Parasitic suppressor; 8 turns No. 22 enam. wire wound over the body of a 56-ohm 1-watt carbon resistor. Use pigtailed of resistor as anchor points for the coil.

maximum desired input power (800 watts) to be reached while operating at a lower plate current than would be needed if a 500- or 600-volt plate supply were used. The lower voltage would require that more tubes be employed to obtain the same d.c. input power without exceeding the 33-watt plate dissipation rating of each tube. Also, the higher plate voltage and lower total plate current result in a plate-load impedance (higher) that is less difficult to match. In this instance the impedance is of the order of 500 ohms, using 900 volts at 890 ma. to obtain the desired 800 watts c.w. input. If a 500-volt supply were used, with six tubes in parallel and 1.6 amperes of plate current, the load impedance would be roughly 150 ohms—a difficult value to deal with when using a conventional tank circuit.

The filament requirements are 6.3 volts at 2.8 amperes per tube. By operating the filaments in parallel one can use a 6.3-volt transformer but the current rating must be 11 amperes or better. To get around this high-current requirement the four 6KD6s were connected for series-filament operation. Happily, the transformer ratings called for 25.2 volts at 2.8 amperes—a standard transformer type available from Allied-Knight or Stancor. Of course a series-parallel arrangement could also be used with a 12-volt transformer, if desired. This would call for a current rating of at least 5.6 amperes.

Since the circuit of Fig. 1 uses the tubes in a grounded-grid hookup, the screen grids and beam-forming plates are grounded and do not have a d.c. operating voltage applied to them.

TABLE I

6KD6 Maximum Ratings

D.C. Plate-Supply Voltage.....	990 Volts
Peak Positive Pulse Plate Voltage.....	7000 Volts
Positive D.C. Beam Plate Voltage.....	70 Volts
Screen Voltage.....	200 Volts
Peak Negative Grid-Number 1 Voltage.....	250 Volts
Plate Dissipation (Absolute-Maximum Value).....	33 Watts
Screen Dissipation.....	5.0 Watts
D.C. Cathode Current.....	400 Ma.
Peak Cathode Current.....	1400 Ma.
Heater-Cathode Voltage	
Heater Positive with Respect to Cathode	
D.C. Component.....	100 Volts
Total D.C. and Peak.....	200 Volts
Heater Negative with Respect to Cathode	
Total D.C. and Peak.....	200 Volts
Grid-Number 1 Circuit Resistance.....	2.2 Megohms
Bulb Temperature.....	225 C.

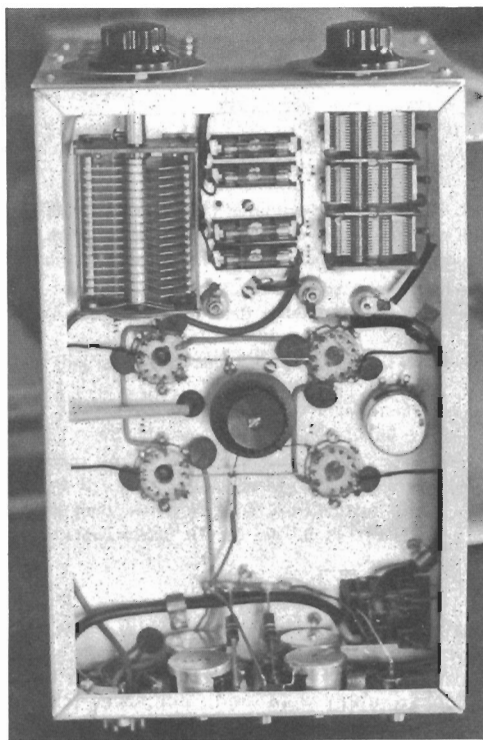
The current they draw varies with the level of the driving signal voltages at a given instant. Tubes operated this way perform as high- μ triodes, and the drive required is below the level that would cause the grids and beam-forming plates to exceed their safe dissipation ratings. The tubes in this amplifier have been in service for several months with no noticeable change in their performance—a good indication that the grids have not been damaged by excessive current.

A negative bias voltage, approximately 7 volts, is applied to the control grids of the tubes to set their resting plate currents, for class AB operation, to 25 ma. per tube with the plate voltage specified here. With 500 volts on the plates it is possible to operate the tubes at zero bias, since the static plate current will be low enough to keep the plate dissipation at a safe level (roughly 28 watts) with no signal applied.

A safe bulb temperature must be maintained at all times if one does not want the tube's envelope to melt or crack. For maximum life the tubes should be operated at temperatures well below the 225 degrees C. that the manufacturer specifies. An example of the catastrophic results of improper cooling can be seen in an accompanying photo. Generally, some type of forced-air cooling should be employed to insure against damage from heat.

What About Matched Tubes?

Whenever power-handling tubes are operated in parallel some provision should be made to assure equal- or near-equal load sharing. The d.c. plate current should be closely matched during both the no-signal and applied-signal periods. Some manufacturers of s.s.b. equipment have solved this problem by careful grading of the tubes, using only those that have similar dynamic characteristics. Handpicked tubes do not come easy for the home experimenter, however, because it might be necessary to purchase as many as twenty in order to get



Individual fuses are used to protect each plate line of the amplifier. They can be seen between the two variable capacitors near the front of the chassis. A toroidal-wound cathode r.f. choke is mounted at the center of tube sockets. Four bias-adjust controls are located on the rear apron of the chassis, permitting the resting plate currents of the tubes to be equalized. Each tube is set for 25 ma. resting current. This requires approximately -7 volts on the control grids. Tubes other than 6KD6s would require different bias values. The unused control at the right center of the chassis is for experimental purposes.

six whose characteristics are closely matched. At \$3.85 per 6KD6 (\$3.47 each in ten lots or more) one might have to spend \$70 to obtain six matched tubes worth only \$23—an impractical solution indeed!

By experimenting with the bias voltages of the four tubes in the amplifier of Fig. 1, it was learned that the no-signal plate currents could be matched when using any four tubes from a set of ten 6KD6s which were on hand. Separate bias-adjust controls permit each tube to be set for the class of operation desired. Fortunately, the plate-current readings tracked within 10 percent over the entire range from no signal to maximum signal when this method was used. (Initial tests were made with a single 300-ma. d.c. meter to avoid errors which could have resulted if four separate meters of unknown tolerance had been employed.)

If reasonable balance is not effected between paralleled tubes—especially if a single meter is used to monitor the total plate current, there is no way of knowing whether one tube is drawing *all* of the no-signal plate current while the remaining tubes are just loafing along. This could mean that the safe plate-dissipation ratings of one or two tubes in a group of several could be exceeded while *idling*! For this reason it is prudent to monitor the plate current of each tube in the amplifier, either by using a single meter and a suitable switching arrangement, Fig. 2, or by using four separate meters as was done here. The latter permits simultaneous monitoring of plate currents, permitting instant comparison of operating conditions. Four low-cost meters run approximately the same as one top-quality meter, so it's pretty much an even-stepsen proposition where cost is concerned. The less-costly meters (Shurite in this instance) have a 5-percent accuracy rating as opposed to the usual 2-percent accuracy of the more expensive instruments. This could mean a difference of as much as 30 ma., at full scale between any two 300-ma. meters, but since 30 ma. represents only a 10-percent error in operating currents it is well within the ball park according to the safe operating limits determined in these experiments.

Parallel Operation

With the circuit of Fig. 1, the total plate current is 890 ma. (800 watts c.w. input). The smaller r.f. chokes cannot handle such high current. Chokes that can take high values of d.c. are expensive. In this circuit four 250-ma. r.f. chokes are used, each tube having separate d.c. feed. This keeps the cost down and makes possible the independent metering of each tube's plate circuit. Initially, four 1-mh. chokes were used and were promptly destroyed when the power output of the amplifier reached upwards of 200 watts. The four chokes in parallel had a series resonance in the 80-meter band—sure death to r.f. chokes in power circuits. Watch out for series resonances when selecting r.f. chokes for a given band of operating frequen-

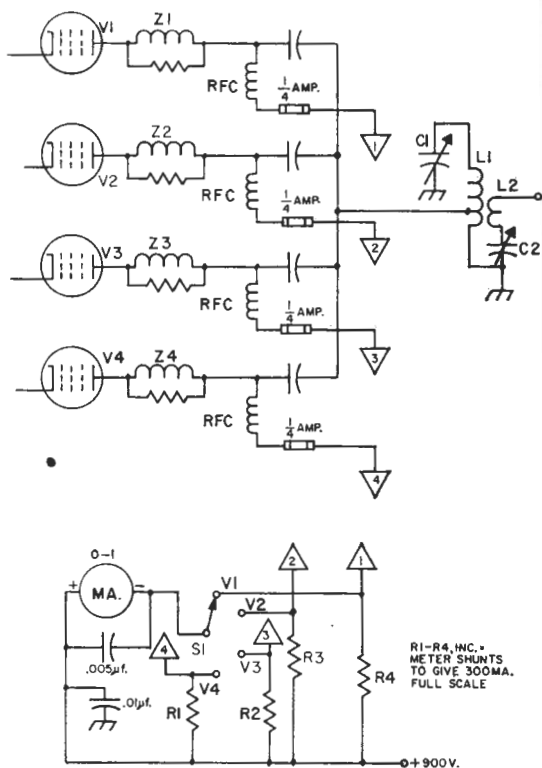


FIG. 2—Alternate method for metering the plate currents of the tubes. This system allows the use of a single 0-1 ma. meter. Meter shunts R_1 through R_4 should be calculated to provide a full-scale meter reading of 300 ma.

cies. A grid-dip meter can be used to locate the series-resonant frequencies of an r.f. choke by checking with the choke terminals shorted. This should be done with all operating voltages removed, but with the circuit connected for otherwise-normal operations.

To offer some protection to the plate meters, a 250-ma. fuse (don't use slo-blo fuses) is installed in each plate lead. Should one of the tubes short out during normal operation, the fuse will open the circuit and protect the meter. As a bonus feature, the defective tube will be turned off and the operator can continue his transmission by reducing the drive to a safe level. Repairs can be made later, when it's more convenient. Each meter is shunted by a 0.01- μ f. disk ceramic capacitor to prevent damage by r.f.

All four cathodes are returned to d.c. ground through a single homemade 200- μ h. choke. Its value is not especially critical provided its impedance is high compared with 50 ohms. An Indiana General CF-117-Q2 toroid core was wound with 90 turns of No. 24 enameled wire to provide the inductance used here.³ Alternatively, a cathode choke could be made by close-winding No. 24 enameled wire to a coil length

³ "Toroidal-Wound Inductors," *QST*, January 1968, page 11.

of approximately 4 inches on a $\frac{3}{8}$ -inch-diameter, $4\frac{1}{2}$ -inch-long piece of ferrite rod.⁴

It is customary to keep the heaters of indirectly-heated tubes above r.f. ground in cathode-driven linear amplifiers, the usual procedure being to install a bifilar-wound inductor in series with the heater feed to the tubes. Since a peak driving power of only 40 to 50 watts is needed to fully excite this amplifier, and the exciter used in these tests was capable of producing up to 100 watts peak output, the filament chokes were omitted since some extra loss could be tolerated. This also worked well on 7 Mc. However, if the amplifier were to be used on 14, 21, and 28 Mc., it would be wise to include the heater chokes.

An eight-inch long, close-wound layer of No. 20 enameled or Formvar-insulated wire on a 10-inch length of $\frac{3}{8}$ -inch diameter ferrite rod would be suitable. Two such windings would be used, one in series with each the filament leads to the 6KD6s, as close to the tube sockets as practical. For even better coupling efficiency between the exciter and the amplifier, a tuned input circuit can be used.⁵ Such a circuit offers an improvement in the linearity of the output waveform of the amplifier.⁶

The Matter of Cooling

If we were to adopt an appropriate slang expression regarding the operating temperatures of sweep tubes, it would no doubt be "don't lose your cool." Heat is the worst enemy of sweep-tube r.f. amplifiers. It not only reduces tube life but is the primary factor that limits the safe input. For these reasons a great deal of emphasis was placed on the matter of tube operating temperature in this circuit.

Some earlier work was done on this subject by L. F. Jeffrey, WA4KFO, an applications and test engineer at the G. E. tube plant in Owensboro, Ky. "Jeff" wrote an interesting and informative paper, jointly with P. E. Hatfield, W9GFS, in which 6KD6 operating temperature versus life was discussed.⁷ They established some test conditions to simulate typical ham-style s.s.b. operation and measured the bulb and plate temperatures of the tubes with special laboratory equipment. During the tests an input power of 448 watts p.e.p. (two tubes) was used, without forced-air cooling. The bulb temperature varied between 135 and 160 degrees C., and the plate temperature varied between 290 and 460 degrees C. They indicated that plate temperatures in excess of 500 degrees C. will definitely shorten tube life. The tempera-

ture can easily rise above the safe 500-degree maximum if the tubes are confined in a small space and do not have forced-air cooling (such was the case with the three melted tubes in the lower left photo).

Jeffrey and Hatfield made additional tests with the same setup, but with forced-air cooling on the two tubes. The resultant bulb temperature fluctuated between 35 and 55 degrees C. while the actual plate temperature ranged between 265 and 450 degrees C. These reduced temperatures offer a worthwhile safety factor.

A high-speed cooling fan with a blade length of $1\frac{1}{2}$ inches—3 inches overall diameter—is used in the amplifier shown in Fig. 3. The fan blades are located approximately two inches away from the tubes. Air is directed against the tubes even when the amplifier is in standby. Additional cooling effectiveness is assured because of the spacious cabinet in which the amplifier is housed (Fig. 4). Large-diameter air vents are located on the top and sides of the cabinet. The rear of the cabinet is enclosed (for TVI purposes) by a sheet of perforated aluminum to enhance the "breathing" capabilities of the assembly.

There is a trend these days toward compactness in transmitting equipment, and this can lead to unhappy results if proper attention is not given to adequate ventilation. To be on the safe side when designing a piece of power-handling equipment allow plenty of cabinet space so air can circulate freely. Locate the exhaust vents near the top of the enclosure; the intake vents should be near the bottom, just above chassis level.

Although heat-dissipating anode connectors were not used in this amplifier, they are worth considering in any permanent design. With air blowing across them, the plate temperature should drop even lower than with the present arrangement.

If a more compact amplifier is desired, the builder might consider using chimneys around each tube and a squirrel-cage blower to direct a heavy air stream up through the bottom of the chassis and around each tube socket. A ring of $\frac{1}{4}$ -diameter holes could be drilled in the chassis around each tube socket to allow the air to flow up through the chimneys. Of course the underside of the chassis would have to be airtight to insure against reduced air flow around the tubes. The chimneys could be fashioned from ordinary food containers such as canned vegetables come in. Each chimney should be painted inside and out with flat black paint to prevent the heat from being reflected back into the tubes. Heat-resistant dull black Bar-B-Q paint is available from most hardware stores in spray cans and is quite suitable for this.

Stability Considerations

Parasitic suppressors are used in the plate leads of all four tubes. They should be installed as a matter of course when building any power-

⁴ $\frac{3}{8}$ and $\frac{1}{2}$ -inch diameter ferrite rods, 12 inches long, are available from Lafayette Radio Electronics, 111 Jericho Tpk., Syosset, L.I. New York 11791.

⁵ "Gimmicks & Gadgets," QST, May 1968.

⁶ Orr, Rinaudo, and Sutherland, "The Grounded-Grid Linear Amplifier," QST, August 1961.

⁷ "Sweep Tubes In SSB Service," L. F. Jeffrey, WA4KFO, and P. E. Hatfield, W9GFS. (An independent paper written and printed by these amateurs.)

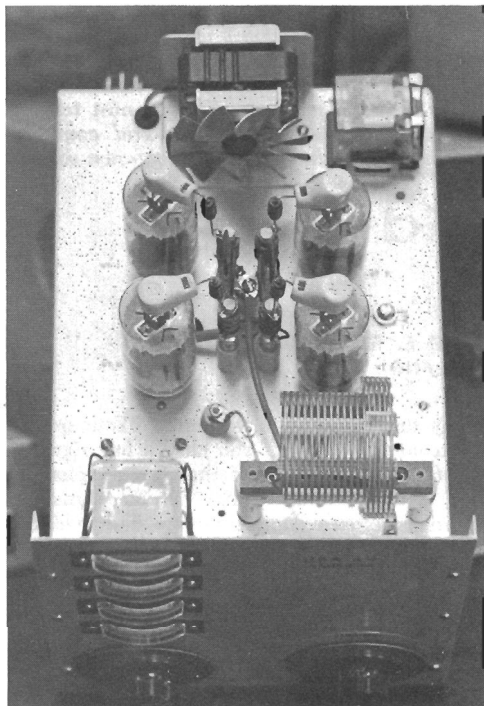


FIG. 3—A high-speed cooling fan is mounted on the chassis just behind the four 6KD6 tubes to provide plenty of air flow across the envelopes. The plate r.f. chokes and blocking capacitors are grouped between the tubes. Plate coil L_1 is mounted on a Millen jack-bar assembly, permitting greater flexibility when experimenting with L/C ratios.

type r.f. amplifier. Each screen-grid terminal on the tube sockets (pins 3 and 11), and each beam-forming plate pin (4 and 10), is grounded directly to the chassis with heavy buss wire, using the shortest lead length possible. This minimizes stray inductance—an aid to stability. Pins 5 and 9 connect to the control grids and each pin on all four sockets is bypassed to the chassis with a 0.005- μ f. disk ceramic capacitor. Good r.f. bypassing is imperative here if the amplifier is to be "tame." Random oscillation can cause excessive currents to melt a sweep tube faster than a blow torch can dispatch an ice cube!

Performance Notes

We do not attempt to rate this experimental amplifier at more than 800 watts c.w. input. If the builder of such an amplifier wants to risk shortened tube life he can raise the power to 1000 watts. This amplifier has been tried on c.w. at 1000 plate volts with a loaded plate current of 1 ampere; no bad effects were observed after a few hours of such operation. The grids and beam-forming plates of the tubes no doubt would be operating near or at their critical dissipation ratings with this kind of power input. A 6-tube amplifier would be a more satisfactory

approach to the higher-power level. During the foregoing tests a power *output* of 600 watts was measured into a 50-ohm dummy load on both 7 and 3.5 Mc. No evidence of harmonic TVI was observed on an RCA color TV set operating in the same room when the amplifier was connected to an 80-meter antenna. The TV antenna was 6 feet away from one end of the 80-meter half-wave antenna.

At 800 watts input (c.w. conditions) an output of 500 watts was easily obtained, indicating an overall efficiency of 62.5 percent. This compares favorably with the performance of the WA4KFO-W9GFS four-tube amplifier described in their technical paper. They used a pi-section output tank. This amplifier, Fig. 1, uses a tapped plate tank to obtain the right load resistance. Either method works well provided the right constants are chosen.

During tests with a two-tone signal a p.e.p. input of 1200 watts was established. The peak-envelope output power was approximately 700 watts (WA4KFO obtained slightly more), resulting in an efficiency of roughly 60 percent. It can be seen from this that a sweep-tube amplifier of this kind is capable of giving a worthwhile boost to signals from medium-power transmitters. A "barefoot" exciter with 100 to 200 watts p.e.p. input might produce a marginal signal under some conditions. By adding a 1200-watt sweep-tube amplifier (p.e.p. input), the signal level should rise 8 to 11 db, a worthwhile increase when the going gets rough.

The circuit of Fig. 1 is wired for use with transceivers of the Heath HW-12 variety. That is to say, a relay is used to provide a switch-through feature so the amplifier can remain in standby when not needed, yet the transceiver can be operated through it.

A final word of warning! Sweep tubes are not designed for continuous operating at these power levels. Do not attempt to use them as a.m. linear amplifiers except at *extremely* low levels. In tuning up, dip the plate current *quickly*. Tune-up should always be done at very low power—50 to 100 watts output—limiting the key-down time to no more than 20 seconds. After initial tuneup the power can be increased to its peak value and the tuning *quickly* touched up. The forward-power reading on an s.w.r. indicator is helpful for optimizing the final adjustments.

The hints should help prospective builders of sweep tube amplifiers to avoid some common pitfalls. The rules apply to all TV sweep tubes that are suitable for use in linear amplifiers, and there are many of them that work well. An inexpensive amplifier can be built by following the basic design described. It is worthwhile to consider a husky TV-set power transformer and bridge-rectifier setup for building a low-cost power supply. It should be possible to get 700 or 800 volts d.c.

» Grounded-grid linear amplifiers enjoy a widespread popularity and not without justification. This article spells out the reasons for the popularity and tells why certain grounded-grid amplifier designs are better than others. This is must reading for any sidebander contemplating a grounded-grid amplifier, whether it be a commercial product or the home-grown variety.

The Grounded-Grid Linear Amplifier

WILLIAM I. ORR, W6SAI, RAYMOND F. RINAUDO, W6KEY,
and ROBERT I. SUTHERLAND, W6UOV

To the hi-fi enthusiast, the linear amplifier is a high-fidelity music amplifier. To the s.s.b. enthusiast, the linear-amplifier package, when placed on the end of a sideband exciter, will make the exciter sound bigger, louder, and more commanding to other amateurs. The fact of the matter is that the s.s.b. linear amplifier is a high-fidelity amplifier in the true sense of the word. Although the hi-fi man thinks in terms of fidelity, and the sidebander thinks in terms of linearity, they are both talking about the same thing.

It is interesting to note that a good hi-fi audio amplifier can theoretically be converted into a low-distortion linear amplifier for sideband service by replacing the audio circuits with suitable r.f. tank circuits. Indeed, for r.f. work, push-pull circuitry is not even required as it is in audio service, because the flywheel action of the r.f. tank circuits will supply the missing half cycle. Finally, the operating parameters for a particular tube—plate, screen, and grid voltages, driving voltage, load resistance—are easily calculated for audio work, and apply equally well for r.f. service. For example, the 811-A tube is rated for Class-B audio service as a high- μ triode. The correct operating conditions (bias, load resistance, plate dissipation, etc.) are exactly the same when the tube is used in Class-B r.f. (grounded-cathode) amplifier service.

Why Linearity?

For sideband service the r.f. power amplifier must be truly linear. It must be capable of high-fidelity reproduction. *That is, the envelope of the signal existing in the plate circuit must be an exact replica of the envelope of the exciting signal.* This statement is a good definition of a linear amplifier. It implies that the power gain of the stage must be constant regardless of the signal level. Any deviation from this happy state creates distortion products that appear in the signal passband and adjacent to it.

Unfortunately, many amateurs judge a sideband signal by its "quality"—that is, the "pleasing" aspect of the voice being transmitted.

Many times one hears the report "Your quality is excellent, old man. You have a fine signal"—yet the listener observes that the recipient of this flattering observation has a signal as broad as a barn door, complete with whiskers and splatter that obliterate half the phone band! Obviously, the criterion of quality of a sideband system is what you *don't* hear, not what you *do*! The place to examine a sideband signal for linearity and quality is in an adjacent channel, not in the frequency band of the signal itself.

How Good Is "Good Quality"?

The excellence of a sideband signal is judged by the amount of (or lack of) sideband splatter in nearby channels. Theoretically, a sideband signal should be about three or four kilocycles wide—just as wide as the voice passband of the equipment. However, the poor sideband operator's ear has been brutally deafened by so many rotten signals that he often accepts any s.s.b. signal as "good quality" as long as it does not blanket the dial of his receiver.

Over the years a nice, easy, vague figure of "30 decibels down" for distortion products has become a password for good-quality, low-distortion, amateur sideband equipment. Since the measurement technique is usually undefined, and practically no amateurs have equipment sufficiently sophisticated to measure the intermodulation products of a sideband signal, this figure has become a byword for most commercial and homemade amateur equipment on the air. Valid or not, this magic number seems to be the socially correct distortion figure applied in all cases to all equipment!

Distortion—What It Means

If the output signal of a linear-amplifier stage is a replica of the exciting signal, there will be no distortion products. However, as vacuum tubes and circuit components are not perfect, this situation is as yet unreachable. As shown in Fig. 1, the transfer characteristic of a typical tube is approximately linear. This tube suffers no pain when amplifying a single signal (such as a carrier or a single tone), but has the interesting property of *mixing* when a

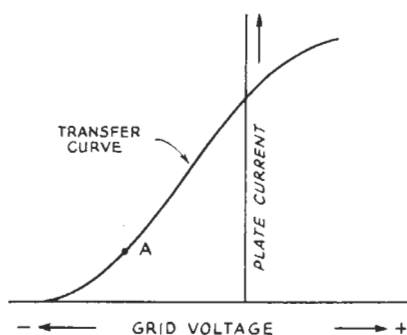


FIG. 1—Plate current vs. grid voltage curve (dynamic characteristic) of a vacuum tube. This curve is linear in the center portion and exhibits deviations at either extremity. The shape of the curve and the choice of the zero-signal operating point (A) will determine the distortion produced by the tube. Mixing action caused by nonlinearity produces distortion products which cannot be eliminated by the tuned circuits of the amplifier.

multiple-signal source is applied to it. This means that a voice signal (made up of a multiplicity of tones) will become distorted and blurred by the inherent mixing action.

A standard test to determine the degree of mixing for a given circuit or tube is the *two-tone* test, in which two radio frequencies of equal amplitude are applied to the amplifier and the output signal is examined for spurious products (Fig. 2). These products, or "garbage," fall in the fundamental signal region and atop the various harmonics. The tuned circuits of the amplifier filter out the spurious signals falling in the harmonic regions, which are termed "even-order" products. The "odd-order" products, unfortunately, fall close to the fundamental output frequency of the amplifier,

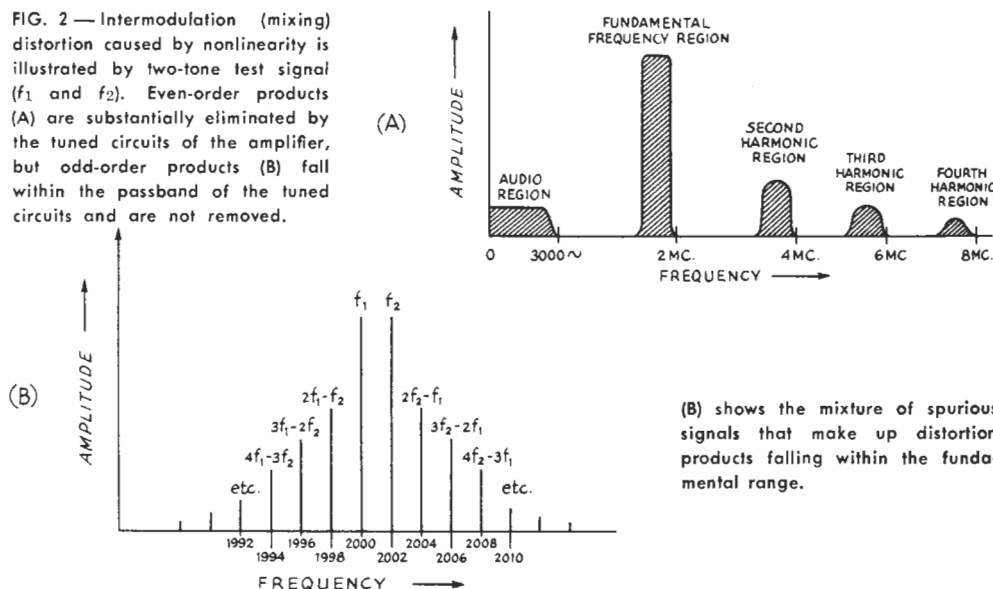
and cannot be removed by simple tuned circuits. These are the spurious frequencies that cause a poorly designed or incorrectly adjusted linear amplifier to cover the dial with splatter.

Shown in the illustration are two frequencies that make up a typical two-tone test signal. In this example, they are 2000 kc. and 2002 kc. Now, if the amplifier is perfect, these two signals will be the only ones appearing in the output circuit. An imperfect (but practical!) amplifier will have various combinations of sums and differences of the signals and the harmonics generated by the nonlinear transfer characteristic of the tube. Some of these unwanted products fall within the passband of the tuned circuits of the amplifier and are radiated along with the two test tones.

If the odd-order products are sufficiently attenuated, they will be of minor importance and can be ignored. The sixty-four-dollar question is: Of what magnitude can these spurious products be without becoming annoying? How much "garbage" can be permitted before the signal becomes intolerable to the operator trying to maintain a QSO in an adjacent channel?

The answer to these questions depends upon the type of information being transmitted and the degree of interference that can be tolerated in the adjacent channel. Certain forms of information (not voice) require an extremely low value of spurious products within and adjacent to the passband, otherwise the information will be seriously degraded. Odd-order products greater than 0.001 per cent of the wanted signal may be damaging to the intelligence. Translated into terms of decibels, this means the unwanted odd-order products must be 50 decibels below the wanted signal. This takes some doing, and is orders of magnitude more

FIG. 2—Intermodulation (mixing) distortion caused by nonlinearity is illustrated by two-tone test signal (f_1 and f_2). Even-order products (A) are substantially eliminated by the tuned circuits of the amplifier, but odd-order products (B) fall within the passband of the tuned circuits and are not removed.



(B) shows the mixture of spurious signals that make up distortion products falling within the fundamental range.

strict than is necessary in amateur voice communications.

In actual practice, it would seem that if the odd-order products are less than 0.1 per cent of the peak signal power level, the adjacent-channel QRM will be tolerable in everyday amateur communications. This indicates a distortion-products magnitude 30 decibels below the peak output power level of the transmitter. Such a state of affairs can be attained by modern techniques and tubes without too much trouble, provided attention is given to circuit design and operating parameters of the equipment. Of course, if distortion levels less than this can be reached, so much the better. Unfortunately, some equipments presently operating in the amateur bands and masquerading as "linear" amplifiers exhibit distortion levels of 20 decibels or less below peak power output. Use of equipment of this dubious quality quickly reduces the popularity of the operator to zero, and will probably lead to a brick through the shack window if continued!

The Grounded-Grid Linear Amplifier

For amateur service, the grounded-grid circuit professes to be the answer to many of the ills besetting the linear amplifier. It generally requires a level of drive that is compatible with the great majority of sideband exciters (70 to 100 watts). With proper choice of tubes, it may be operated in a zero-bias condition, eliminating the need for expensive and heavy grid (and screen) power supplies. Neutralization is not usually required. In addition, claims are made that the inherent feedback of the grounded-grid amplifier improves the stage linearity and drops the magnitude of the distortion products. This all sounds too good to be true, and an examination of the grounded-grid amplifier may be in order to see if it is the answer to the sidebander's prayers.

The classic grounded-grid amplifier is shown in Fig. 3. The control grid is at r.f. ground potential and the driving signal is applied to the cathode via a tuned circuit. The control grid serves as a shield between the cathode and the plate, making neutralization unnecessary at medium and high frequencies.

The input and output circuits of the grounded-grid amplifier may be considered to be in series, and a certain portion of the input power appears in the output circuit. This feed-through power helps somewhat to stabilize the load the amplifier presents to the exciter, and also provides the user with some "free" output power he would not otherwise obtain from a more conventional circuit. The driver stage for the grounded-grid amplifier must be capable of supplying the normal level of excitation power required by the amplifier plus the feed-through power. Stage power gains of 5 to 25 can be achieved in a grounded-grid amplifier.

Measurements made on tubes in the Power

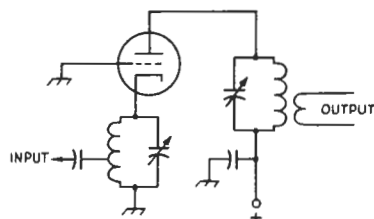


FIG. 3—The grounded-grid amplifier has the input circuit between cathode and ground. The control grid acts as a screen between the plate and the cathode, making neutralization unnecessary in most circuits. The input and output circuits are in series and a portion of the input power appears in the output circuit. The driver stage for the grounded-grid amplifier must be capable of supplying normal excitation power plus the required feed-through power. High-C cathode tank preserves waveform of input signal and prevents distortion.

Grid Tube Laboratory of Eitel-McCullough, Inc., showed that an improvement of 5 to 10 decibels in odd-order distortion products may be gained by operating various tubes in the grounded-grid configuration of Fig. 3, in contrast to the same tubes in the grid-driven mode. The improvement in distortion figure varied from tube type to tube type, but all tubes tested showed some order of improvement when cathode driven.

The tuned cathode circuit consisted of a bifilar coil, which carried the filament current, and a large-value variable capacitor. The circuit was high-C, with the excitation tap placed to provide a low value of s.w.r. on the coaxial cable to the exciter.

The Untuned Cathode Circuit

After sufficient measurements had been made with the circuit of Fig. 3, the apparatus was modified to simulate the popular untuned cathode input circuit of Fig. 4. It was imme-

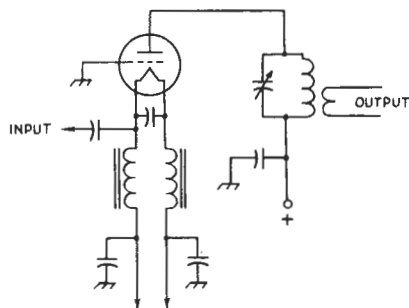


FIG. 4—Popular amateur-style grounded-grid amplifier used untuned filament choke in place of cathode tuned circuit. Laboratory tests showed that this simplified configuration produced higher intermodulation distortion products and had less power output than the "classic" circuit of Fig. 3, regardless of the type of tube used. In addition, the untuned input circuit proved hard to match and drive with pi-network-type sideband exciter.

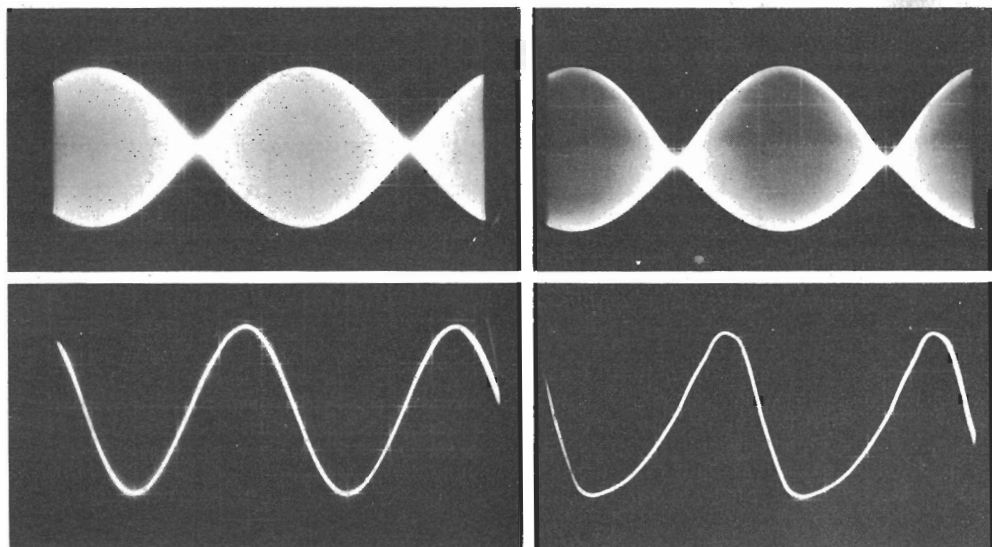


FIG. 5—Waveform distortion caused by half-cycle loading at cathode of grounded-grid amplifier can be observed in oscilloscope studies. Upper left: Two-tone test signal when tuned cathode circuit is used. Lower left: 3.5-Mc. waveform (single tone) from sideband exciter as seen at cathode tank. Upper right: Two-tone test signal when untuned cathode circuit is used. Lower right: 3.5-Mc. waveform (single tone) from sideband exciter, showing severe distortion of waveform when untuned cathode circuit is used.

diately found that all the tubes tested in the previous circuit gave noticeably poorer results when used with an untuned cathode circuit. Power output dropped by 5 per cent or so, greater grid driving power was required, and linearity suffered to a degree. Specifically, the third-order products rose approximately 3 to 4 decibels over the values produced by the circuit of Fig. 3, and the fifth-order products rose 5 to 6 decibels over those figures recorded with the tuned cathode circuit. The higher-order distortion products also rose accordingly. These results were consistent regardless of the *type* of tube under test.

Observing the input waveform at the cathode of the grounded-grid amplifier revealed a pronounced distortion of the r.f. waveform, caused by the loading effect over one-half cycle of a single-ended Class-B amplifier (Fig. 5). Plate and grid currents drawn over this portion of the cycle loaded the input circuit. Unless the output regulation of the exciter is very good, the portion of the wave on the loaded part of the cycle will be seriously degraded, as shown. The exciter used for these tests was operating Class A and was well swamped to improve regulation. Under normal circumstances using an amateur-type exciter, degradation of the input wave form may reach a more serious degree. Obviously, the circuit *Q* of the exciter output tank at the end of a random length of interconnecting coaxial line is not sufficient to prevent this form of wave distortion.

The solution to this problem is to employ either a high-*C* tuned circuit of the form shown in Fig. 6A, or untuned filament chokes

in conjunction with a simple pi-network or tuned circuit as shown in Figs. 6B and 6C. Either arrangement will supply the necessary flywheel effect to retain good r.f. waveform at the cathode of the stage.

Adjustment of the Tuned Cathode Circuit

The cathode circuit is resonated to the operating frequency by means of the variable capacitor. Resonance is indicated by maximum grid current in the amplifier. A low value of s.w.r. on the coaxial line to the exciter is established by adjusting the tap on the tuned circuit, or by varying the input capacitor of the pi network. S.w.r. correction should be made with the amplifier running at maximum input. When the tap is correctly set, maximum grid current and minimum s.w.r. will coincide at one setting of the capacitor. No cutting and trimming of the coaxial line is required, and the exciter will be properly loaded. This is a boon, indeed, to the owners of s.s.b. exciters that have a fixed pi network.

Grid-Current Measurement

Measuring the grid current of a cathode-driven amplifier can be exasperating, as it is a ticklish job to "unground" the grid sufficiently to permit a metering circuit to be used yet still hold the grid at r.f. ground potential. The inherent inductance of most bypass capacitors permits the grid circuit to float above the ground at some high frequency, and as a result, the amplifier exhibits instability and parasitics. This problem can be avoided by using the measuring circuit of Fig. 7A. The control grid is grounded through a 1-ohm composition re-

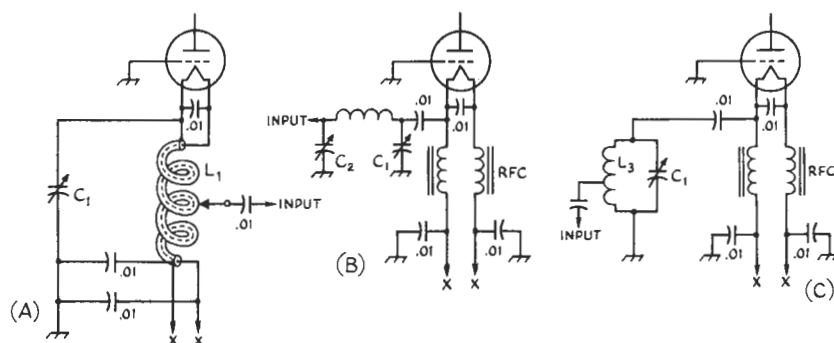


FIG. 6—Tuned cathode network for zero-bias tube may take the form of bifilar circuit (A), pi network (B), or a shunt LC circuit (C). A Q of 5 is recommended for optimum results. However, as this leads to rather bulky circuits at the lower frequencies, the Q may be decreased to 2 or 3 without serious effects. Capacitor C_1 is a 3-gang broadcast-type unit. Coils L_1 , L_2 and L_3 are adjusted to resonate to the operating frequency with C_1 set to about 13 pf. per meter of wavelength. Capacitor C_2 is approximately 1.5 times the value of C_1 . The input tap on coils L_1 and L_3 , or the capacitance of C_2 , are adjusted for minimum s.w.r. on the coaxial line to the exciter.

sistor, bypassed by a 0.01- μ f. disk capacitor. The voltage drop generated by the flow of grid current across the resistor can easily be measured by a millivoltmeter calibrated to read in terms of grid current. Individual grid currents for each of a parallel pair of tubes may be measured by the circuit of Fig. 7B.

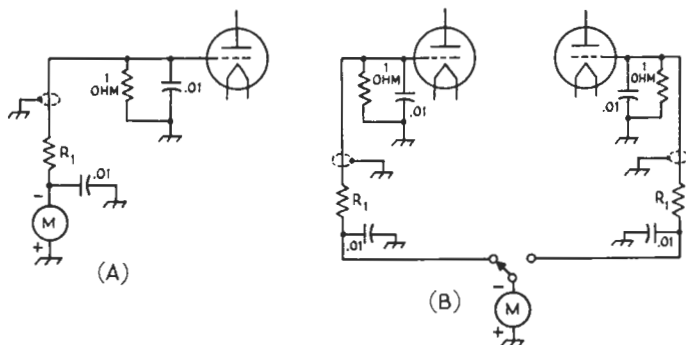
The internal resistance of the 0-1 d.c. milliammeter plus the series resistor R_1 determines the maximum current that can be measured. Suppose it is desired to read grid current of the order of 60 milliamperes. It would be convenient, therefore, to have the meter read 0-100 milliamperes, as the reading of the meter scale can easily be multiplied by 100 to obtain the actual value of current. Now, when a current of 100 milliamperes flows through 1 ohm, there exists a potential of 0.1 volt across the resistor. Therefore, the meter should read 0.1 volt full scale to correspond to a grid current of 100 ma. Assume the meter has an internal resistance of 55 ohms (such as the Triplett No. 22-T). The voltage drop across the meter itself is 0.055 volts when one milliamperes flows through it, but at one milliamperes the resistance must be 100 ohms

for a voltage drop of 0.1 volt. The difference between 100 ohms and 55 ohms, or 45 ohms, must therefore be added in series with the meter to convert it to read 0.1 volt full scale. On the other hand, the meter by itself across the 1-ohm resistor would indicate 0.055 volts full scale, corresponding to a grid current of 55 milliamperes. If the grid current is below this latter figure no series resistor will be required for the meter. Conversely, higher values of grid current would call for greater series resistance.

Summary

The use of the tuned cathode circuit in a grounded-grid linear-amplifier stage improves linearity, increases the power output, makes the stage easier to drive, and reduces the burden placed on the sideband exciter. It is the firm belief of the authors that the advantages of this circuit are well worth the added cost of parts and the extra controls. It is, of course, possible to dispense with the tuned cathode circuit provided the user understands the handicaps he must assume by omission of this important circuit element.

FIG. 7—Grid current in a grounded-grid amplifier may be measured across a low impedance without upsetting the stability of the amplifier. (A) Grid is grounded by a 1-ohm composition resistor in parallel with a 0.01- μ f. ceramic disk capacitor. Resistor and capacitor leads are cut very short, and lead to metering circuit is shielded. (B) A single meter may be used to measure individual grid currents of two tubes.



» Neutralization and control of grid isolation within the cathode-driven amplifier permit the designer to adapt the basic circuit to the particular operating conditions at hand. Power gain and feedthrough power may be varied, and the amplifier can be stabilized for proper operation over a wide frequency range.

The Cathode-Driven Linear Amplifier

WILLIAM I. ORR, W6SAI AND WILLIAM H. SAYER, WA6BAN

The cathode-driven, or grounded-grid, amplifier¹ is ideally suited to amateur s.s.b. or c.w. service and seems to be gradually relegating the grid-driven amplifier to the junk box. The attributes of the cathode-driven amplifier are impressive: it has reasonable power gain, it usually requires no auxiliary neutralization below 30 megacycles or so, it offers lower residual circuit capacitance, and parasitic suppression is not difficult. Under certain conditions, moreover inherent negative feedback exists in this configuration, to the benefit of amplifier linearity. Finally, a portion of the cathode r.f. drive power shows up in the output circuit, thus providing a

degree of "free" output power not otherwise available from a conventional grid-down circuit.

Strictly speaking, the extra output power is not "free," as r.f. power is expensive compared to d.c. plate power and may only be "free" if it is unavoidably available. It is generally referred to as *feed-through* power, but the implication in this term may be misleading, as this portion of the drive power does not appear in the load circuit of the cathode-driven stage until after it is converted to a varying d.c. plate potential effectively in series with the main amplifier power supply. This *converted drive power* performs a useful function in Class AB₁ and Class B linear service by swamping out the undesirable effects of nonlinear grid loading and presenting a reasonably constant load to the exciter.²

The purpose of this article is to examine certain aspects of the cathode-driven amplifier, not widely recognized, that afford additional flexibility and versatility under particular operating conditions, and which permit accurate and complete neutralization to be achieved when needed.

The Basic Cathode-Driven Circuit

First discussed in *QST* in September, 1933,³ the cathode-driven circuit has generated a considerable body of literature over the past few decades (see bibliography). The circuit is believed to have first been conceived circa 1920 by Ernst Alexanderson of alternator fame. Used about 1938 in European short-wave broadcast and TV service, this unique amplifier configuration became popular in U.S. post-war low-channel TV transmitters about 1944 or so.

The basic cathode-driven circuit is shown in Fig. 1. It may be operated either as an oscillator or as an amplifier by proper choice of components and potentials. The grid of the tube is nominally at r.f. ground potential and the exciting signal is applied to the cathode, or filament. For amplifier service, it is assumed that the cathode is instantaneously driven positive with respect to ground (the grid), the plate

From June, 1967, *QST*.

¹ The term "cathode-driven," or "grid-isolation" is preferred over "grounded-grid," as the latter implies that the grid is at r.f. and d.c. ground. This is often not the case.

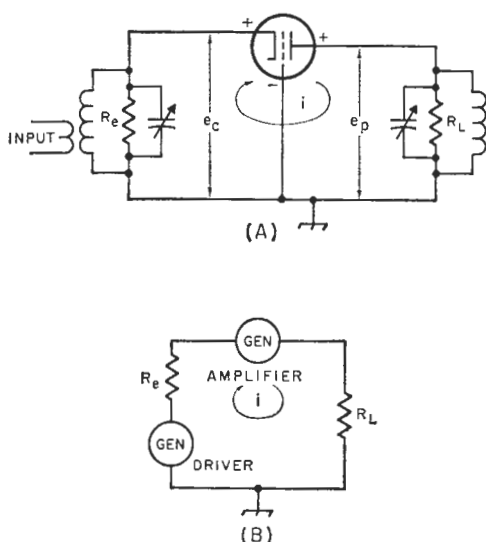


FIG. 1-A—The cathode-driven circuit. Driving voltage (e_c) is applied to the cathode of the amplifier and the output voltage (e_p) appears across the plate load impedance, R_L , in phase with e_c . The grid of the tube is at nominal ground potential. B—The driver and cathode-driven amplifier are in series with respect to the amplifier r.f. voltages. Amplifier cathode current (i) flows through the load resistance of the driver, contributing a degree of r.f. feedback.

² Pappenfus, Bruene and Schoenike, *Single Sideband Principles and Practice*, McGraw-Hill Book Co., N.Y. (1964).

³ Romander, "The Inverted Ultra-audio Amplifier," *QST*, September, 1933.

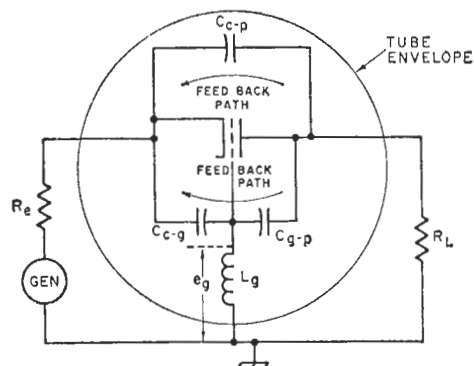


FIG. 2—Distributed constants of cathode-driven tube. Cathode-to-plate (C_{c-p}), cathode-to-grid (C_{c-g}) and grille-to-plate (C_{g-p}) capacitances, together with grid-leads inductance (L_g) make up feedback paths that must be neutralized for proper operation of the cathode-driven amplifier. Two feedback paths enter the picture: the direct path from plate to cathode via C_{c-p} , and a more devious path via series capacitors C_{c-g} and C_{g-p} .

will become more positive with respect to the cathode, and also with respect to ground. The instantaneous plate voltage, in effect, is developed in series and in phase with the exciting voltage, and the driver and amplifier stages may be thought of as operating in series to deliver power to the load, R_L . The delivered power is the sum of converted drive power and amplifier power, less any power from the driver required by the amplifier grid circuit. A parallel-tuned circuit is used in the cathode of the amplifier to enhance the regulation of the driver stage, to complete the plate circuit r.f. return path to the cathode, and to provide proper driver termination over the operating cycle.⁴

As the cathode-driven amplifier is effectively in series with the driver stage, the output current passes through the load resistance of the driver (R_e), causing a voltage drop across that resistance which opposes the original driving voltage. This indicates that inverse feedback is inherent in the cathode-driven amplifier to some degree if the driver has appreciable load resistance.⁵

Neutralization

The familiar cathode-driven amplifier used in h.f. amateur service is usually not neutralized. That is to say, no external neutralizing circuit is built into the amplifier. This omission has led to the general belief that the "grounded grid acts as a shield" and neutralization is not necessary in any and all cathode-driven amplifiers. The accepted proof of this belief is the fact that most h.f. amplifiers, in most instances, will not oscillate in use. Operation of an unneutralized

cathode-driven amplifier in the upper portion of the h.f. spectrum, however, may provide unpleasant surprises. Many amateurs have found to their chagrin that such an amplifier is often a tricky "beast" to tame at 10 and 6 meters.

The reason for the unwanted instability is simple. Wires and leads represent finite inductances, and their position relative to each other and to other circuit components represents capacitance; both these quantities may have an effect upon amplifier performance. Vacuum tubes have these distributed constants within their envelopes in the form of interelectrode capacitances and lead inductance.

Voltage feedback from output to input through the distributed constants of the tube has a deleterious effect on amplifier performance. The magnitude, phase and rate of change with respect to frequency of this feedback determine the dynamic stability of the amplifier, and control of feedback is termed *neutralization*. The purpose of neutralization of any amplifier, regardless of circuitry, is to make the input and output circuits independent of each other with respect to voltage feedback and the resulting reactive currents.⁶ When a cathode-

⁶ In fact, the cathode voltage is dependent to a degree upon the output voltage, as the input and output circuits are in series.

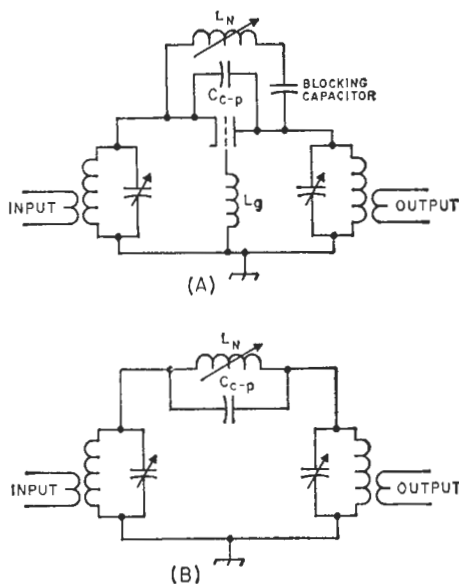


FIG. 3—A—Cathode-plate inductive neutralization. Capacitive feedback path between cathode and plate via C_{c-p} may be neutralized by making the capacitance part of a parallel-resonant circuit tuned to the operating frequency by the addition of L_N . A blocking capacitor is used to remove the d.c. plate voltage from the coil. Neutralization is frequency sensitive. B—Equivalent circuit; high-impedance parallel-resonant circuit nullifies feedback path between input and output circuits via plate-to-cathode capacitance.

⁴ C. E. Strong, "The Inverted Amplifier," *Electrical Communication* (England), Volume 19, No. 3, 1941.

⁵ J. J. Muller, "Cathode Excited Linear Amplifiers," *Electrical Communication* (England), Volume 23, September, 1946.

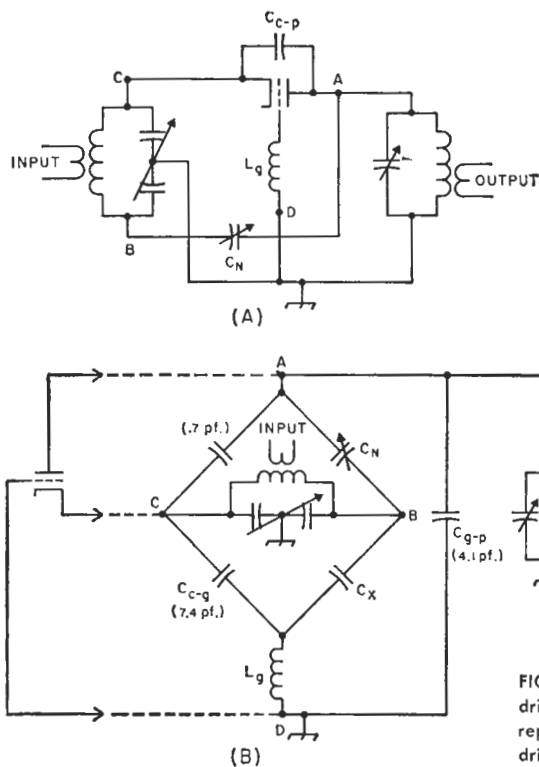


FIG. 4-A—Cathode-plate bridge neutralization. Balanced input provides equal out-of-phase voltages at points B and C. When C_N is equal to C_{c-p} , equal out-of-phase voltages will cancel each other at point A and feedback path via C_{c-p} is neutralized. B—Neutralization circuit redrawn in bridge form, with typical capacitance values for 3-400Z triode shown in parentheses. Bridge is balanced except for capacitance C_x , representing residual capacitance to ground at point B. If the balanced input circuit is high-C in comparison to interelectrode capacitances of tube, capacitances C_{c-g} and C_x are swamped out and bridge may be considered to be balanced.

driven amplifier is operated at the higher frequencies, the internal capacitances and the inductance of the grid structure of the tube contribute to the degree of feedback (Fig. 2). To achieve stability, the various feedback paths through the distributed constants inherent in the tube structure must be balanced out, or nulled, in some fashion by neutralization techniques. Proper neutralization may be defined as *the state in which, when plate and cathode tank circuits are resonant, maximum cathode voltage, minimum plate current, and maximum power output occur*. This definition implies that the input and output circuits are independent of each other with respect to common reactive currents, and that tuning of the circuits reveals no interaction.

As the grid of the tube is at nominal ground potential in a cathode-driven amplifier, it appears that this element may act as a screen, or

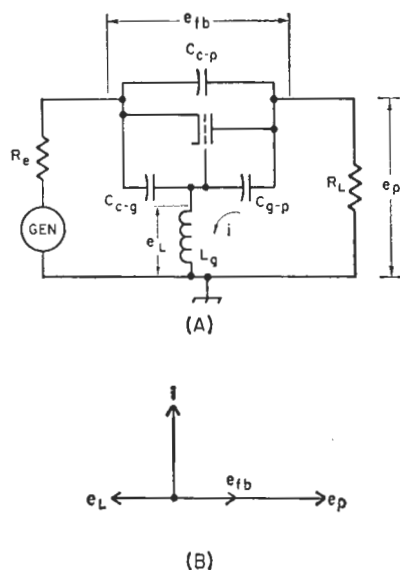


FIG. 5-A—Three-terminal representation of cathode-driven tube. See text for explanation. B—Vector representation of feedback voltages in cathode-driven tube.

shield, between the output and input circuits and that instability or oscillation due to feedback paths through the interelectrode capacitances of the tube may be avoided, or reduced to negligible values. At the lower frequencies, particularly with respect to well-shielded, low-gain tubes, this belief may be true. However, in the higher-frequency region the practical tube (i.e., the tube that can be built) departs to an important degree from this simplified concept.

Neutralizing the Cathode-Driven Amplifier

Suitable operation of the cathode-driven amplifier often requires some form of neutralization when the frequency of operation approaches the upper reaches of the h.f. spectrum. Complete circuit stability requires neutralization of *two* feedback paths, for which separate techniques are required.

The first feedback path involves the cathode-to-plate capacitance, C_{c-p} . Although the capacitance involved is small, the path is critical and requires neutralization. Neutralization may be accomplished either by a shunt inductance (Fig. 3) or by a balanced capacitance bridge circuit (Fig. 4). The first technique consists of connecting a reactance from plate to cathode of such magnitude as to transmit back to the cathode circuit a current equal in value but opposite in phase to the current passing through the cathode-to-plate capacitance. The bridge technique is a version of the well-known capacitance neutralizing circuit used in conventional

grid-driven amplifiers to balance out the effects of grid-plate capacitance. The balanced input circuit provides equal out-of-phase voltages to which the cathode of the tube and the neutralizing capacitor are coupled. As the value of the neutralizing capacitor is equal to the cathode-to-plate capacitance of the tube, the voltages are balanced at the junction of the two capacitances, which is the plate termination of the cathode-driven tube. Both capacitances are usually quite small, and the effect of series lead inductance in the bridge circuit is relatively unimportant. Consequently a reasonable bridge balance over a wide frequency range may be obtained with a single setting of the neutralizing capacitance.

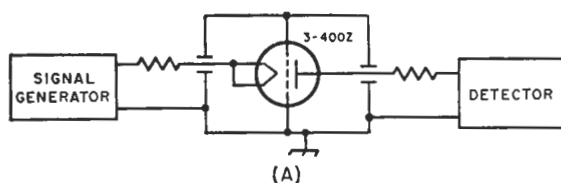
The shunt-inductance neutralizing circuit of Fig. 3, on the other hand, has the disadvantage of requiring adjustment for each working frequency, as the external inductance and cathode-to-plate feed-through capacitance form a frequency-sensitive parallel-resonant circuit at the operating frequency.

Either neutralizing circuit may be properly balanced⁷ even though the grid of the tube may not be at actual ground potential because of internal grid inductance, L_g . Intrastage feedback resulting from this inductance requires a separate, unique solution, apart from the neutralizing technique just discussed.

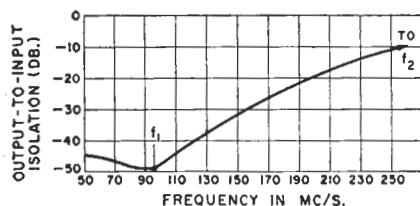
Grid-Inductance Neutralization

The second feedback path in the cathode-driven stage includes the grid-to-plate capacitance, the cathode-to-grid capacitance and the

⁷ With physically-large tubes having appreciable series input inductance, in-phase neutralization is often required. This may be achieved by adding external cathode-to-plate capacitance, or by detuning the shunt inductor from the condition of parallel resonance.



(A)



(B)

FIG. 6-A—The self-neutralizing frequency of a cathode-driven triode may be measured by observing the transmission properties of the cold tube when treated as a three-terminal network. B—Typical plot of intrastage isolation of 3-400Z triode mounted in test fixture. Self-neutralizing frequency of tube is about 100 megacycles.

series grid inductance, L_g , as shown in Fig. 2. The grid inductance represents the sum of all possible feedback paths *through* the grid structure, plus the actual series inductance of the grid structure. In practical tubes, there is no possibility of avoiding all inductance in the path between the active grid element of the tube and ground. This path exists because the grid is not a solid, intercepting structure. After all, openings must exist to permit electrons to pass from the cathode to the plate! Capacitance leakage can exist between the cathode and the plate through these openings. In addition, Maxwell's equations state that changing electric and magnetic fields propagate each other through space. In the vicinity of the real grid structure, the electric field about the "input" side of the structure gives rise to currents flowing in the structure which, in turn, cause an electric field to exist about the "output" side of the structure. In addition, electromagnetic coupling through the interleaved grid structure is also observed.⁸

These spurious coupling paths result in an apparent r.f. leakage through the cathode-to-grid and grid-to-plate capacitances that is often many times greater than that predicted by actual measurement of the internal capacitances. A simplified picture of this complex path may be seen as an inductance in series with the grid-to-ground path, common to both input and output circuits (Fig. 2). If this path is not neutralized, a voltage e_g appears on the grid of the tube which either increases or decreases the driving voltage, depending upon the value of internal capacitances and grid inductance. With sufficient spurious grid voltage, the cathode-driven stage may oscillate, or be unstable, even though the cathode-to-plate feedback path discussed earlier is completely neutralized.

The voltage e_g on the so-called "grounded grid" is determined by a complex action between the total cathode-to-plate capacitance and a separate low- Q circuit composed of a capacitive voltage divider (C_{g-u} and C_{u-p} in series) together with the grid inductance, L_g . A certain frequency at which these two feedback paths nullify each other is termed the *self-neutralizing frequency* (f_1) of the tube. This frequency usually occurs in the lower portion of the v.h.f. spectrum with small transmitting tubes. All the elements comprising the neutralizing circuit are *within the tube*. However, connecting the tube into the circuit by wiring or socketing will alter this frequency.

The self-neutralizing phenomenon comes about because of a frequency-sensitive voltage balance that takes place within this network, Fig. 5A, and which may be explained by a simple vector diagram, Fig. 5B. The r.f. plate voltage (e_p) causes a current (i) to flow through

⁸ Feedback admittance also is enhanced by the self-inductance of the grid wires, which provides common coupling between input and output circuits. The inductive coupling may partially compensate for the feedback through the cathode-to-plate capacitance. (See Bibliography, item 3.)

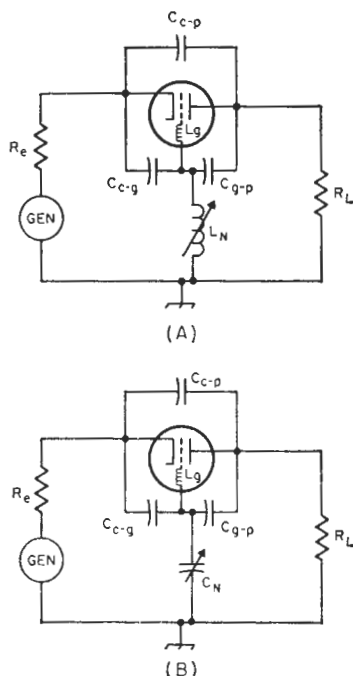


FIG. 7-A—The point of self-neutralization may be shifted lower in frequency by the addition of an inductance (L_N) in series with the grid-to-ground termination of the tube. B—The point of self-neutralization may be shifted higher in frequency by the addition of a capacitor (C_N) instead of an inductor.

C_{g-p} and L_g . If the reactance of L_g is small in comparison with the reactance of C_{g-p} (as would be the case below the self-neutralizing frequency), the current i will lead the plate voltage e_p by 90 degrees. In flowing through L_g this current will develop a grid voltage (e_L) which is 180 degrees out of phase with e_p and with the voltage e_{fb} fed back to the cathode via C_{c-p} and series-connected C_{c-g} and C_{g-p} .

At some frequency the voltage e_L developed across L_g will just equal the voltage fed back through the interelectrode capacitances (e_{fb}). The frequency at which e_L is equal to e_{fb} is the self-neutralizing frequency. At this frequency a cancellation of feedback voltages occurs and the complex feedback path is nullified, or "neutralized." (A second, somewhat higher, frequency at which the complex grid configuration is in a series-resonant state with respect to intrastage isolation is called the *grid series-resonant frequency* (f_2) of the tube.)⁹

The Self-Neutralizing Characteristic Curve

The self-neutralizing characteristic of a cathode-driven triode may be determined by

treating the tube as a passive three-terminal network and measuring transmission as a function of frequency. The tube is placed in a test fixture which is contrived to insure that the frequency measured is dependent on the tube and socket only (Fig. 6). A signal is applied to the "cold" tube through an appropriate attenuator and a detector is used to measure the transmission voltage through the tube. Investigation over a range of frequencies will produce a typical plot such as shown in Fig. 6B. The point of maximum isolation is the self-neutralizing frequency, f_1 . Measurements are not quantitative, as nothing is known about the impedance of the input or output circuits. The relative isolation with respect to frequency, however, is the interesting parameter.

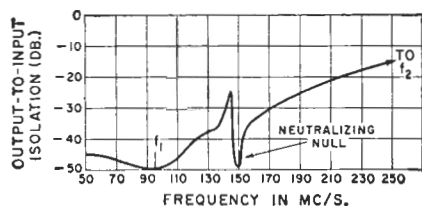
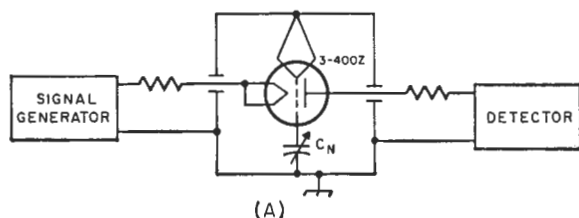
The self-neutralizing frequency (a broad null of several hundred kilocycles) may be moved about by manipulation of the external grid-to-ground circuitry of the tube, or by changing the capacitive feedback path. Or, if desired, a secondary point of neutralization may be created, as described later. If the desired frequency of operation is above the self-neutralizing frequency the voltage developed on the "grounded grid" will be too great and the series grid inductance, L_g , must be reduced, or the feedback path adjusted to establish self-neutralization. If the operating frequency lies below the self-neutralizing frequency, the voltage on the "grounded grid" will be insufficient to cancel the feedback voltage and the series grid inductance must be increased.

The portion of the plot around the point f_1 has been experimentally verified by observing the intrastage leakage (transmission) properties of a 3-400Z zero-bias triode mounted in an SK-510 socket and fixed in a partition in an r.f.-tight enclosure. Observation was over the range of 50 to 250 megacycles, and the self-neutralizing frequency was seen to be in the neighborhood of 100 megacycles (Fig. 6B). Above this frequency, the intrastage isolation gradually deteriorated as the series-resonance frequency, f_2 , of the grid element was approached. Near the latter frequency, tube operation is impractical, being further complicated by transit-time effects and other v.h.f. phenomena.

The Self-Neutralizing Frequency

The self-neutralizing frequency of a cathode-driven triode depends to a large degree upon the size of the tube, the interelectrode capacitances, the physical configuration of the grid structure and the inductance of the grid leads and terminals. Below this frequency, the tube can be neutralized by the addition of a small inductor (L_N , Fig. 7) in the grid-to-ground path. Above this frequency, neutralization may be achieved by reducing the reactance of the path by the addition of a suitable series capacitance, C_N . To demonstrate this a variable capacitor was placed in series with one grid terminal of the 3-400Z mounted in the test fixture. At any frequency between f_1 and 250 megacycles the shape of the plot could be

⁹ "Care and Feeding of Power Grid Tubes", application bulletin No. 13, EIMAC, a Division of Varian, San Carlos, Calif.



(B)

altered by adjustment of the capacitor, providing a neutralizing "null," Fig. 8, in the curve of about the same amplitude as observed at the lower frequencies. The Q of the neutralizing circuit (one grid lead plus the capacitor) was considerably higher than the Q of the grid system, and the neutralizing adjustment proved to be rather frequency-sensitive. The original self-neutralizing frequency (f_1) was little altered by the addition of the auxiliary circuit.

A second test conducted on a larger tube (the 3X2500A3, a 2.5-kw. low- μ triode) showed that it could be neutralized on the lower-frequency side of the self-neutralizing frequency f_1 by the addition of a suitable inductor between the grid terminal and ground. Both techniques are shown in Fig. 7.

It should be noted that intrastage self-neutralization and cathode-plate neutralization are interlocked. In the lower portion of the v.h.f. spectrum only one technique may be necessary to achieve a satisfactory degree of neutralization, at least as far as amplifier stability goes. At 6 meters, for example, either system will completely stabilize many amplifiers in most situations. At higher frequencies such is not the case, and both feedback circuits may require attention and manipulation to allow the amplifier in question to be properly neutralized.

General Remarks

Conclusions to be drawn as to the degree of intrastage isolation, or as to the requirement for neutralization in a cathode-driven amplifier, tend to be clouded unless backed by measurements made on the equipment, just as is the case with grid-driven amplifiers. In the latter instance, neutralization of the circuit is almost taken for granted. Not so with cathode-driven amplifiers, as adequate isolation and stability have often been achieved at the lower frequencies even with tubes that were not designed for this purpose. It is unwise to jump to the gen-

FIG. 8-A—The 3-400Z may achieve neutralization over a wide v.h.f. range by the addition of a series capacitor in one grid lead. Neutralization adjustment is frequency sensitive and must be peaked for maximum intrastage isolation of the operating frequency. B—Plot of intrastage isolation of 3-400Z, showing neutralizing null added by the series grid capacitor. Null may be moved about between f_1 and f_2 . A similar neutralizing effect may be obtained at frequencies lower than f_1 by the circuit shown in Fig. 7-A.

eral conclusion that this special situation exists in all cases.

At the lower frequencies, particularly with well-shielded, low-capacitance tubes, neutralization may not be necessary, and this permits the circuit designer to make use of circuit techniques and practices that afford variation of power gain, converted drive power, and degree of inverse feedback to the cathode driven amplifier. Specifically, these parameters may be varied to meet the demands of the system or to adjust the converted drive power requirement of the amplifier to match the available drive power of the exciter. These circuit schemes, however, should not be confused with the separate problems of amplifier neutralization, discussed in this article.

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AN 811-A 200-WATT GROUNDED-GRID LINEAR AMPLIFIER

The amplifier shown in Fig. 1, 3 and 4 requires about 15 watts of excitation power to drive it to full peak input (200 watts) on 3.5 through 30 Mc. For convenience and compactness, the amplifier is completely self-contained; silicon-diode rectifiers in the plate and bias supplies contribute materially to the small size.

Referring to the circuit diagram in Fig. 2, the input impedance of the grounded-grid 811-A amplifier (about 300 ohms) is stepped down through an "L" network to offer approximately 50 ohms as a load for the driver. The network makes for little or no complication, since the circuits are fixed-tuned and, once adjusted, need not be touched again. It will be noted that on the 15- and 10-meter bands no lumped capacitance is used in the network; this is because the capacitance of the length of RG-58/U running from S_{1B} is sufficient.

The filament choke, RFC_1 , is an inexpensive homemade one (described later). Since the filament winding of the power transformer has no center tap, two 22-ohm resistors are used to provide a center tap for the filament circuit. In the band-switched plate circuit, a commercial inductor (with two winding pitches) is used, and because the output capacitor is not large enough on 80 meters, on that band an additional 500 pf. is switched in by S_2 .

To meter grid or plate current, a 0-1 milliammeter is used as a 0-1 voltmeter to measure the drop across 10 ohms in the grid circuit or

2.5 ohms in the plate circuit, giving 0-100 and 0-400 ma. full-scale readings respectively.

A panel operate-standby switch, S_4 , removes the fixed grid bias during operate periods. If an external control is available, as in a VOX-controlled s.s.b. exciter, S_4 is left open and the external circuit connected through J_2 .

All of the power is derived from a single husky TV power transformer. The plate power is derived from a voltage-doubling circuit using inexpensive silicon diodes and 450-volt electrolytic capacitors. The filament voltage for 811-A is obtained from one transformer secondary, and another 6.3-volt secondary is utilized in a voltage-doubling circuit to provide cut-off bias for the 811-A, to avoid diode-noise problems if an electronic t.r. switch is used. High-voltage filtering is furnished by four 40- μ f. capacitors connected in series.

Construction

The amplifier is built on a 10 \times 12 \times 3-inch aluminum chassis, with a panel and back panel of 0.063-inch aluminum measuring 9 \times 12 inches. One-inch aluminum angle stock is used to make side and top lips that take the perforated-aluminum cover. The cover, not shown in the photographs, is a single piece 10 inches wide bent in a broad "U" shape; it is held to the lips by sheet-metal screws.

Capacitors C_1 and C_2 are fastened to the top of the chassis by 6-32 hardware; C_1 is located

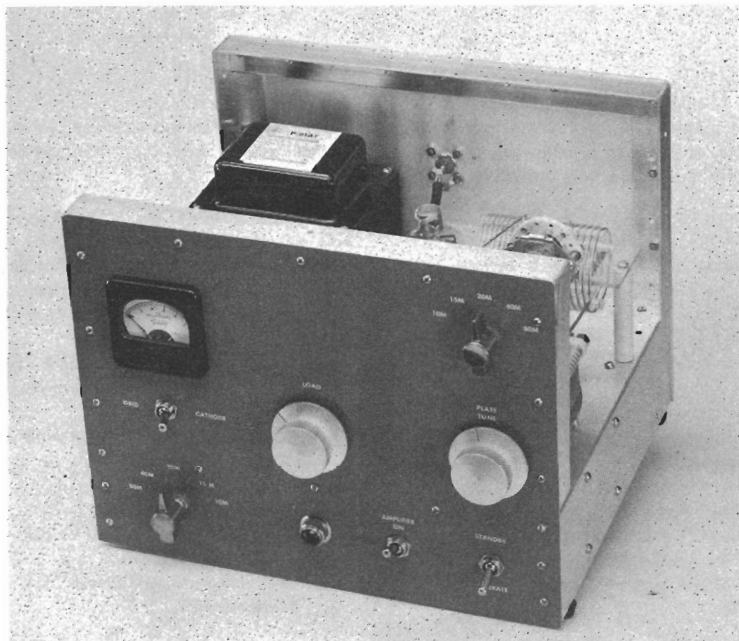


FIG. 1—The 200-watt grounded-grid amplifier with its perforated-metal cover removed. This compact amplifier uses an 811-A and a simple 1300-volt power supply. To simplify construction, two bandswitches are used (input at lower left, plate at upper right). The single meter can be switched to read either grid or plate current.

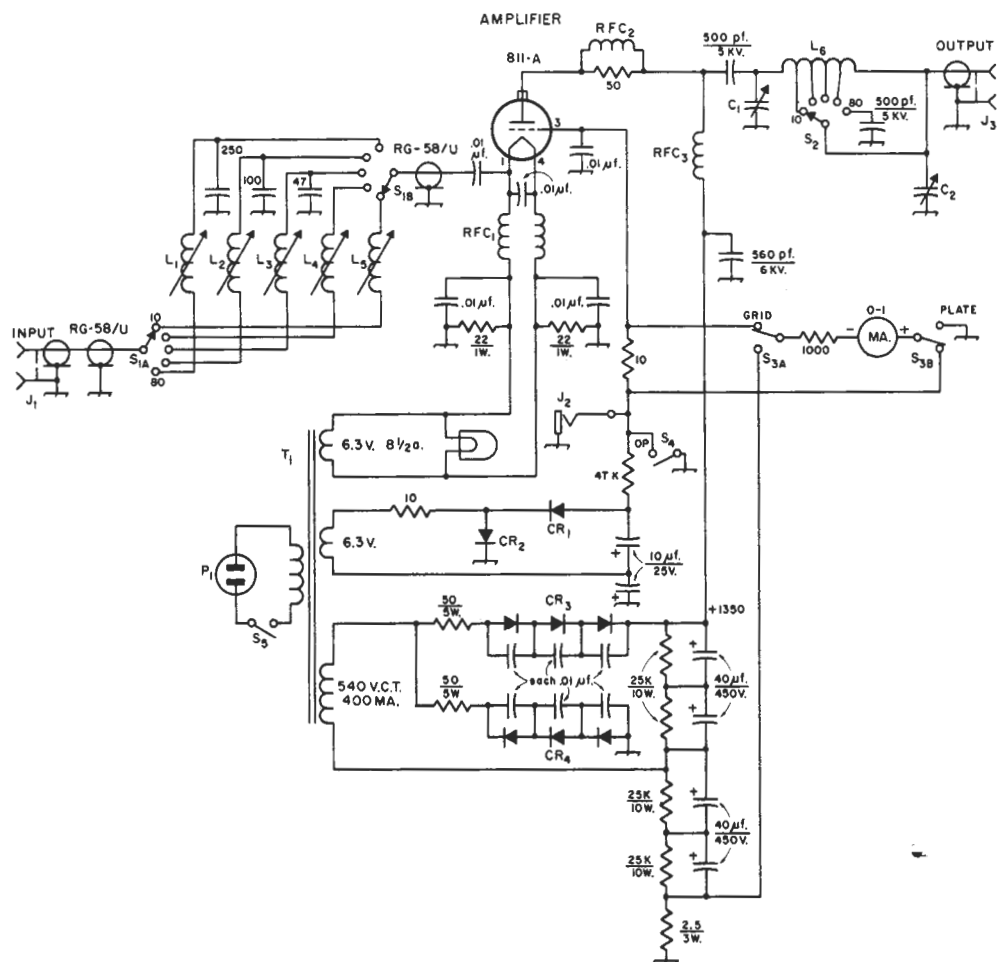


FIG. 2.—Circuit diagram of the 200-watt grounded-grid linear amplifier. Unless specified, all capacitances are in picofarads (pf. or $\mu\text{mf.}$), all resistors are $\frac{1}{2}$ watt, all resistances are in ohms. Capacitors marked with polarity are electrolytic; 0.01- $\mu\text{f.}$ capacitors are 1200-volt disk ceramic.

- C₁—250-pf. variable, 0.045-inch spacing (Johnson type 154-1).
C₂—3-gang capacitor, 365 pf. each section (Allied Radio 60 L 726).
CR₁, CR₂—200 p.i.v. 750 ma. silicon (RCA 1N3253 or equiv.).
CR₃, CR₄—Each three 600-p.i.v. 500-ma. silicon diodes in series (RCA 1N3195 or equiv.).
J₁, J₃—Coaxial receptacle, chassis type (SO-239).
J₂—Open-circuit jack.
L₁—5-9- μ h., adjustable (Miller 4505).
L₂—3-5- μ h., adjustable (Miller 4504).
L₃, L₄—1-1.6- μ h., adjustable (Miller 4502).
L₅—0.4-0.8- μ h., adjustable (Miller 4501).
L₆—22 turns No. 12, 2-inch diameter, tapped at 2, 3, 5, and 10 turns from C₁ end (Air-Dux PI 1608D6).
P₁—Mounting plate a.c. plug (Amphenol 61-M1).
RFC₁—Dual winding, 29 turns No. 14 Formvar or Nylclad, spacewound on ferrite rod. See text.
RFC₂—4 turns No. 14, 5/8-inch diam., 1 1/4 inch long, wound outside two 100-ohm 1-watt resistors in parallel.
RFC₃—1-mh. r.f. choke (National R-154U).
S₁—2-pole 6-position rotary ceramic (Centralab PA-2003).
S₂—1-pole 6-position rotary ceramic (Centralab PA-2001).
S₃—D.p.d.t. toggle.
S₄, S₅—S.p.s.t. toggle.
T₁—560 v.c.t. 400 ma.; 6.3 v. 8.5 a.; 6.3 v. 4.5 a. (Stancor P-8167).
Knobs are Barker & Williamson 901; bar knobs are National HRB.

far enough in from the edge so that its stator will clear the cane-metal side by $\frac{1}{4}$ inch or better. The plate bandswitch, S_2 , is supported by an aluminum bracket that is fastened to the rear of C_1 . The 500-pf. plate-blocking capacitor and the RFC_2 assembly are supported by the top of RFC_3 , and the 500-pf. 80-meter output padding capacitor is bolted to the chassis below S_2 . Plate coil L_6 is supported by two $2\frac{1}{2}$ -inch ceramic pillars. To reduce the height taken by the 811-A above the chassis, the 811-A socket is supported below the chassis by mounting it (Amphenol 49RSS4) in a recessed shell (Amphenol 61-61).

Underneath the chassis, the two toggle switches, the 6.3-volt pilot lamp, and the band-switch S_1 , are mounted on the front lip of the chassis. The input inductors, L_1 through L_5 , are clustered around the bandswitch, as are the several capacitors associated with this circuit. Lengths of RG-58/U run from the arms of S_1 to the input jack, J_1 , and the 811-A socket. The unused socket pin (No. 2) is used as a tie point

for the coaxial line and the 0.01- μ f. coupling capacitor.

The filament choke, RFC_1 , is made by winding No. 14 Formvar or Nylclad wire on a $7\frac{1}{2}$ -inch length of $\frac{1}{2}$ -inch diameter ferrite antenna core (Lafayette Radio, N.Y.C., MS-333). To obtain a high-Q coil, the two windings are wound parallel but spaced by lacing twine to give 29 turns in each coil. The coil is wound by securing the two ends and the length of spacing twine in a vise, securing the other wire ends to a 2-terminal strip held in place by a $\frac{1}{2}$ -inch diameter nylon cable clamp, and then winding the coils as the wires are stretched taut. Each turn of the core winds two turns of wire and one of twine. The twine is left on the coil, and no insulation is required between wires and core when the recommended surface covering (Formvar or Nylclad) is used. The choke assembly is supported below the chassis by 1-inch ceramic posts and the nylon cable clamps.

The bias-supply rectifiers, resistors and capacitors are mounted on a multiple tie-point

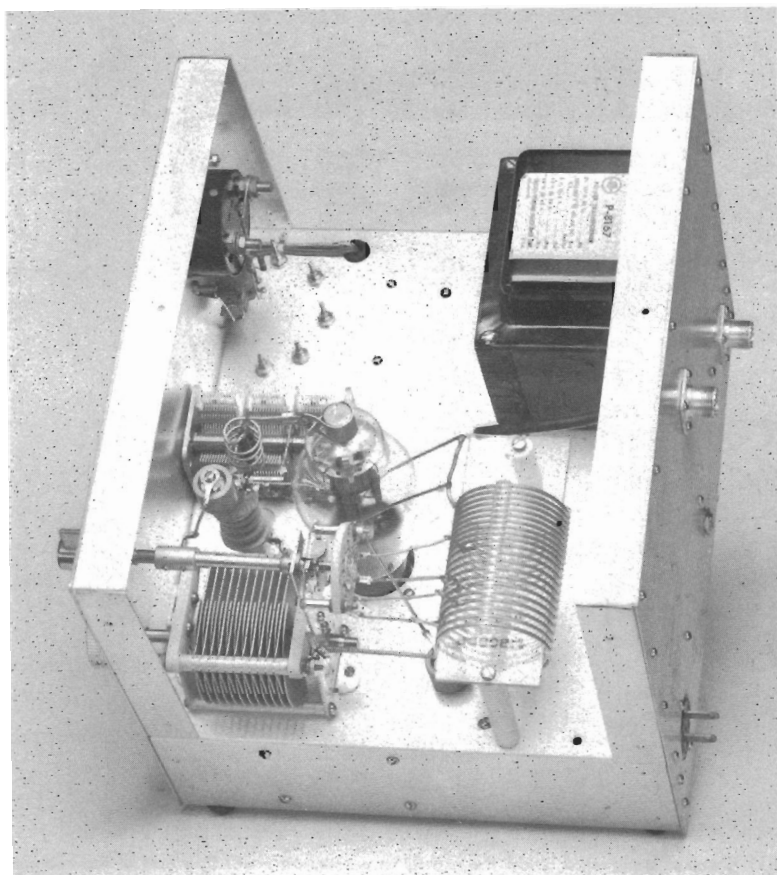


FIG. 3—A top view of the 811-A amplifier. The adjusting screws for the five switched input circuits project through the chassis under the meter. A bracket fastened to the back plate of the plate tuning capacitor (lower left) supports the plate bandswitch.

strip. In the high-voltage supply, the diodes and capacitors are mounted on a 4×7 -inch piece of $\frac{1}{16}$ -inch thick prepunched phenolic terminal board (Vector 85G24EP) with push-in terminals (Vector T-28). The resistors, both 50-ohm 5-watt and 25,000-ohm 10-watt, are mounted on tie points or narrow strips of terminal board located several inches from the diode and capacitor board. The reason for this is simple: the resistors become hot and might damage the diodes if mounted too close to them. The 2.5-ohm 3-watt resistor consists of three 7.5-ohm 1-watt resistors connected in parallel.

Tuning

When the wiring has been completed and the power supply checked (+1500 volts no-load, about 1450 with the 811-A drawing idling current of 30 ma.), the amplifier can be checked on a band with a driver capable of delivering a

peak signal of 15 watts or so. A dummy load should be used during initial tests, and an output indicator (r.f. ammeter or voltmeter) is very useful. Using a c.w. signal to drive the amplifier, it should be found possible to load the amplifier so that at plate-circuit resonance the plate current is 160 ma. and the grid current is about 27 ma. As the drive is reduced the grid and plate currents should drop back at roughly the same rate. If the amplifier is not loaded heavily enough, the grid current will run proportionately higher than the plate current. There is, of course, no real substitute for a two-tone linearity test, but the above figures will serve as a rough guide. When the amplifier has been loaded to the figures above with a c.w. driving source, an s.s.b. signal driving it to peak output will kick the plate meter to about 80 ma. (0.2 on meter) or the grid meter to 15 ma. (0.15 on meter).

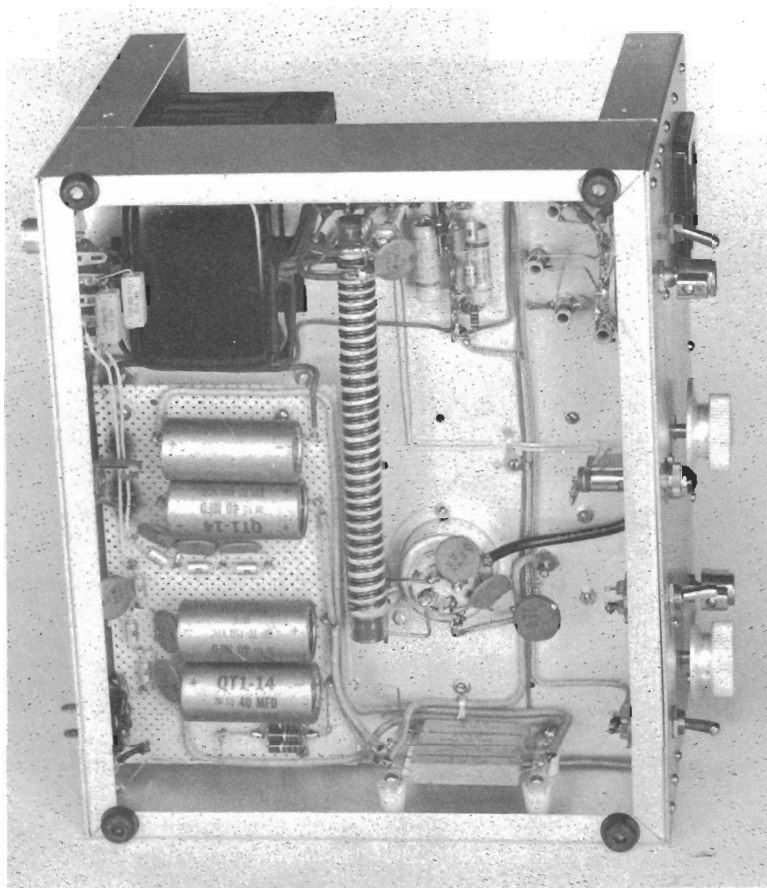


FIG. 4—The 811-A socket is mounted below the chassis in a recessed shell. One end of the homemade filament choke is supported near the socket, and the other end is mounted near the transformer. Four 25,000-ohm bleeder resistors (bottom) and two 50-ohm resistors (upper left) are mounted well away from the plate-supply diodes (left) and bias diodes (top center, to right of filament choke).

A COMPACT 3-500Z AMPLIFIER

The amplifier shown in Figs. 1 through 4 easily handles a kilowatt p.e.p. input at 3000 volts. It has been designed with ease of construction and operation in mind, and to this end as few special parts and machine operations as possible are required. Probably the major operation is adding an arm to the band switch, to ground a plate padding capacitor in the 3.5-Mc. position. This enables a smaller plate tuning capacitor to be used than would be the case if the variable were required to furnish all of the capacitance on this lowest-frequency band.

Referring to the wiring diagram in Fig. 2, the circuit is about as simple as it could be made. No tuned input circuit is used, since it was found that any of the s.s.b. units in the 75- to 100-watts output class could drive it without any trouble. If drive were marginal, as when only 35 watts peak were available, a coupling network might offer a slight advantage. Two r.f. chokes and a 1000-pf. bypass are used in the high-voltage lead because a high-impedance circuit like this is harder to filter than one where the current is higher and the voltage is lower. The plate coil is a standard 500-watt unit that runs cold at a kilowatt c.w. or s.s.b.

The 50,000-ohm resistor in the center tap of the filament transformer biases the tube to cut-off during "stand-by" periods and eliminates the "diode noise" caused by the static plate current. Leads to J_4 and J_5 from the VOX or other control short the resistor during transmit period.

The connections on J_6 are similar to those on the 3-1000Z amplifier shown later in this chapter, with the exception of the lead marked "VM". This variation permits mounting the voltmeter on the transmitter panel instead of in the power supply. The power supply can be the same as that shown for the larger amplifier, with the exception of the power transformer.

A smaller 600-va. unit (BTC-6181) has the same voltage ratings and 60 per cent of the price of the larger transformer; it is heartily recommended unless one plans some day to move up to the 3-1000Z amplifier.

Front and back panels and base plate are all standard unfinished $\frac{1}{8}$ -inch thick aluminum rack panels. They are trimmed to 15 inches. The angle stock holding the pieces together, and furnishing the faces for support of the cover, are $1 \times 1 \times \frac{1}{8}$ -inch Reynolds stock. A short piece is also used for supporting the fan, cut away as shown in Fig. 4.

The tube socket (Johnson 122-275-200) is supported 1 inch above the base on rods threaded into the base. Its three grid pins are bonded together with a $\frac{3}{8}$ -inch wide strap of copper; the strap in turn is bolted to the base.

To conserve space, the filament transformer must be modified so that the leads come out the side. This is done by removing the end bells and drilling a hole in the side through which the leads can be threaded.

To modify switch S_1 , first remove the rear shaft bearing and replace the ceramic insulators with shorter ($\frac{1}{2}$ -inch) ones. Fiber washers or any suitable insulating material can be used here to expose the end of the switch shaft. A brass shaft coupling, cut to a length of $\frac{7}{16}$ inch, is drilled and tapped 6-32 at right angles to the normal set-screw hole. The spring stock ($0.20 \times \frac{3}{8}$ silver solder) is wrapped half around the coupling and fastened at two points with short 6-32 screws through the new hole. The original set screw is left exposed. (Silver solder is available at welding supply houses; the type used here is called "Handy Harmon Easy Flow"). The fixed contact is supported by a ceramic insulator mounted on the base plate.

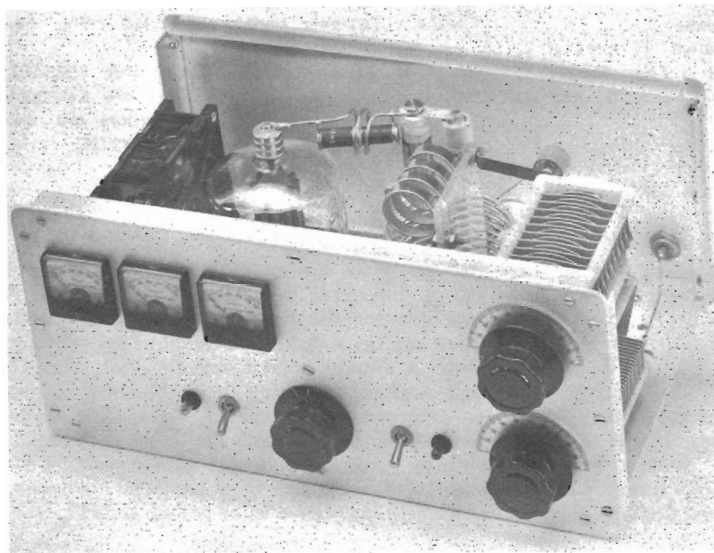


FIG. 1—The compact kilowatt amplifier with its perforated-metal cover removed. Using a 3-500Z in a grounded-grid circuit, it handles a kilowatt p.e.p. input at 3000 volts with ease. The (2-inch) meters monitor plate voltage, grid current and plate current. Panel is 7×15 inches; the bottom plate is $8\frac{3}{4}$ inches wide.

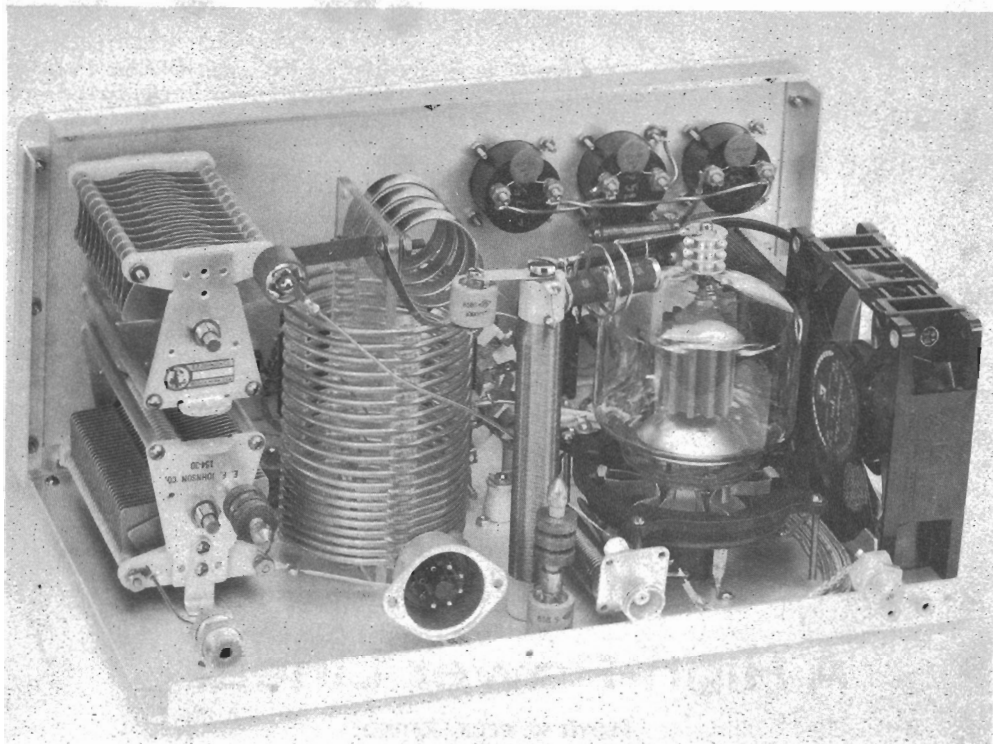


FIG. 3—The rear wall of the compact kilowatt has been removed to reveal the "works." Coaxial receptacles at left are output and input jacks; receptacle at center (near tube) is high-voltage connector. A 50-pf 3.5-Mc. plate loading capacitor can be seen mounted on the plate tuning capacitor (upper left); the 500-pf. 3.5-Mc. output loading capacitor is mounted on the base behind the coil (just visible to right of variable loading capacitor).

Note heavy copper strap from socket to base, grounding grid pins. The fan is mounted on a piece of $\frac{1}{4}$ -inch foam rubber and held in position by two screws through rubber grommets in the vertical plate.

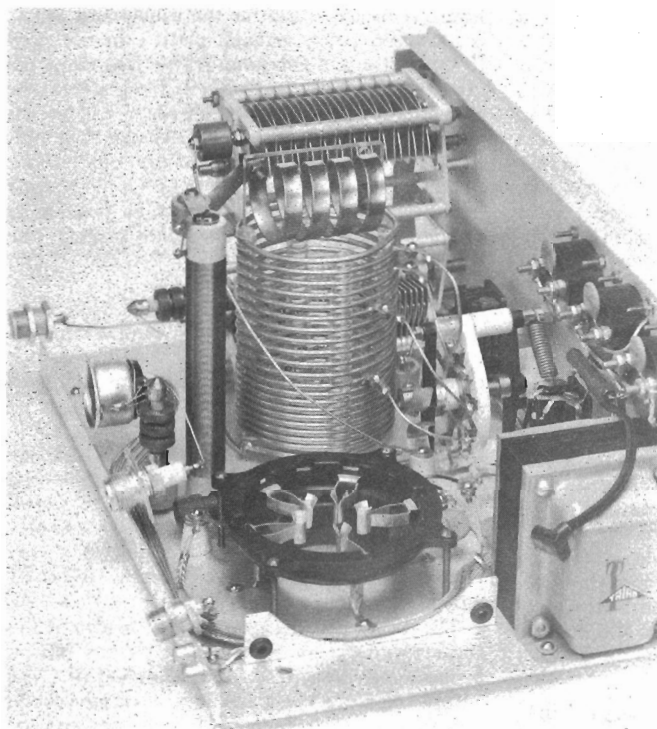


FIG. 4—Close-up view with the tube and fan removed discloses details of switch S_{1B} . It is made from a brass shaft coupling and a length of silver solder; in the 3.5-Mc. position it contacts a fixed arm and grounds the 50-pf. fixed capacitor (upper left).

Mounting plate for fan is trimmed away for maximum ventilation under tube socket.

"A Class-B linear amplifier in the p.e.p. kilowatt category, complete with power supply, in a space barely exceeding 1 cubic foot. The grounded-grid configuration is used with four parallel-connected 811As.

The high-power grounded-grid linear in its homemade cabinet. Controls across the top are for the plate tank capacitor, band switch and loading capacitor. Filament and plate-voltage switches flank the grid and plate milliammeters below. The construction of the cabinet was described in QST as footnoted in the text.



A High-Power Linear

FLOYD K. PECK, K6SNO

Having decided to go all the way with single sideband, the old Class-C amplifier and modulator were sacrificed to the junk box. Then it was decided to see what could be salvaged for a linear amplifier that would give the most output with the available parts. We had a couple of 811As in the old modulator, and a couple of spares, and they were selected for duty as linear amplifiers. Since the exciter was in the 100-watt-output class, it was decided to take maximum advantage of this output and drive the four 811As as grounded-grid amplifiers. The power supply for the old a.m. rig delivered 1250 volts d.c. at 300 m.a., so it fitted our requirements pretty well. The complete circuit of the unit is shown in Fig. 1.

Reducing the Size

As first built, the linear was housed in a cabinet 20 inches wide, 13 inches high and 15 inches deep. It was built on a 17×13×3-inch chassis. In our project to build the compact linear in a cabinet 14 inches wide by 8 inches high and 17 inches deep,¹ the same chassis size was used but the layout was reoriented. The power transformer used is 7 inches high, so it was necessary to submount it since only 5 inches of clearance was available above the chassis. A 5½×6-inch opening corresponding to the base dimension of the transformer was cut in the rear right-hand corner of the chassis, and brackets were made to provide support

2 inches below the chassis. This allows ample clearance for a.c. and high-voltage terminals below chassis.

The 866A rectifier tubes must also be mounted so that their bases are below chassis level. A 5-inch space for the 866As and 811As is provided when ceramic plate caps are used if the bases are submounted so that only the glass portions of the tubes extend above the chassis. The sockets for the four 811As are mounted on a 6×6-inch sheet of ⅜-inch aluminum suspended 1¼ inches below the chassis. Eight ⅜-inch holes were drilled in the chassis in a 2-inch circle around each tube position to provide natural convection for cooling the tubes.

Pi-Network Tank Circuit

A conventional pi-network tank circuit is used, and it was built around the Illumitronic 500-watt coil. The markings on the coil indicate tap points for the band switch, so that no calculations are necessary if a 1250- to 1500-volt power supply is used. About half the turns can be removed from the close-wound end of the coil, which allows it to be physically shortened to mount horizontally within a space of 5 inches. The band switch is a very sturdy one obtained from a surplus BC-375E antenna-tuning unit.

The input tuning capacitor, C₁, is also of the surplus variety, made by Cardwell and having a maximum capacitance of about 500 pf. The output (loading) capacitor is a three-section broadcast-receiver type of 365 pf. per section,

From June, 1961, QST.

¹ See Peck, "Homebrew Custom Designing," QST, April, 1961.

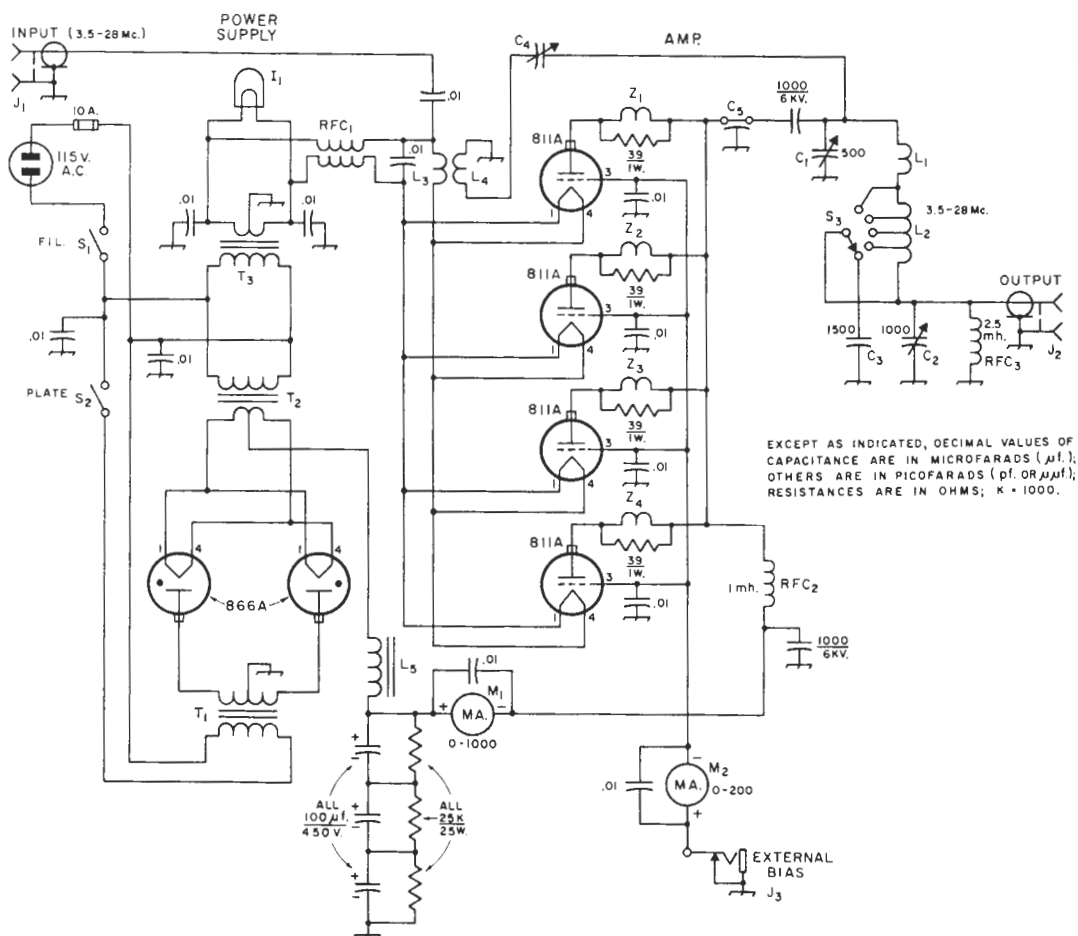


FIG. 1—Circuit of the high-power grounded-grid linear and its built-in power supply. Capacitors not listed below are disk ceramic, except those marked with polarity which are electrolytic. Resistances are in ohms.

C₁—500-pf. 2000-volt variable (Johnson 154-3/500E20 or similar—see text).

C₂—Triple section broadcast replacement variable, 365 pf. per section, sections in parallel.

C₃—2500-volt mica.

C₄—Neutralizing capacitor—approx. 6 pf. 0.06-inch spacing or greater (Bud CE-2028).

C₅—V.h.f. bypass (4-inch length of RG-58/U as connecting lead).

L₁—6.3-volt panel lamp.

J₁, J₂—Coaxial receptacle (SO-239).

J₃—Closed-circuit jack.

L₃, L₂—Pi-network inductor (Illumintron PiDux No. 195-1) approx. inductances in use: 0.4, 0.7, 1, 2.2 and 4.5 μ h., respectively, for 10–80 meter, L₂ wound with No. 8 wire, L₁ wound with $\frac{1}{2}$ -inch copper strap (see text).

L₃—6 turns No. 14, $\frac{1}{2}$ -inch diam., close-wound.

L₄—5 turns insulated hookup wire wound over L₃.

L₅—Filter choke: 5–8 h., 300 ma. (Stancor C-1722 or similar).

M₁—0–1000-ma. d.c. meter.

M₂—0–200-ma. d.c. meter.

RFC₁—Bifilar filament choke (B & W FC-15).

RFC₂—R.f. choke: 1 mh. 600 ma. (National R154-U).

RFC₃—2.5-mh. r.f. choke, 50–100 ma.

S₁, S₂—S.p.s.t. toggle switch.

S₃—Band switch (see text).

T₁—1250-volt (d.c.) 300-ma. plate transformer (Stancor PT-8313 or similar).

T₂—Filament transformer: 2.5 volts, 10 amp. (Stancor P-3024 or similar).

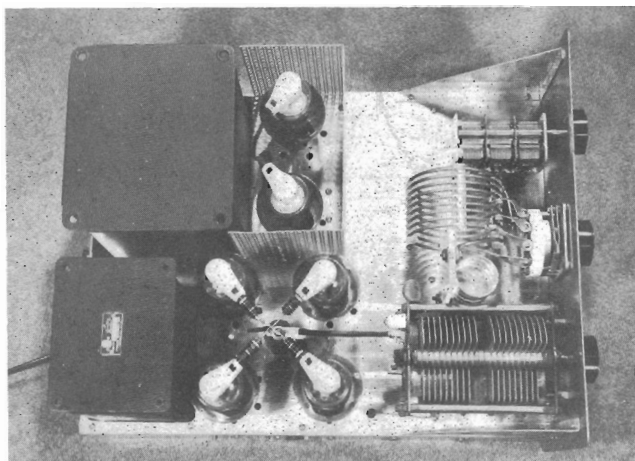
T₃—Filament transformer: 6.3 volts, 16 amp. (Triad F-22A or similar—see text).

Z₁—Z₄ incl.—Parasitic suppressor; 7 turns No. 18 wire, wound on and connected across a 39-ohm 1-watt resistor.

Filament Supply

The filament requirements for the 811As are 6.3 volts at 16 amperes. The old transformer from the modulator, designed to handle a

with the sections connected in parallel. In the 3.5-Mc. position, the band switch connects a 1500-pf. silver-mica fixed capacitor in parallel with the variable loading capacitor.



Components on top of the chassis are easily identified. The power-supply filter choke and submounted high-voltage transformer are at the left-hand end of the chassis. Tubes enclosed in the perforated shield above the four 811As are the 866A rectifiers. To the right are the plate tank capacitor, the pi-network inductor with its switch, and the loading capacitor. The neutralizing connection runs from a stator terminal on the tank capacitor, through a clearance hole in the chassis to the neutralizing capacitor below deck. (Photos by Greg Bethards.)

single pair of 811As, proved incapable of supplying the required voltage through the filament chokes with four tubes in the circuit. The secondary, which turned out to be wound with No. 16 wire, was removed, the turns being carefully counted as they were unwound. A new secondary was wound with No. 14 wire and the number of turns was increased by 10 per cent. The measured voltage at the sockets was then 6.4 volts with a line voltage of 117. There were some qualms about the ability of the primary to hold up under these conditions, but the transformer has operated for over two years with no trouble.

Bias

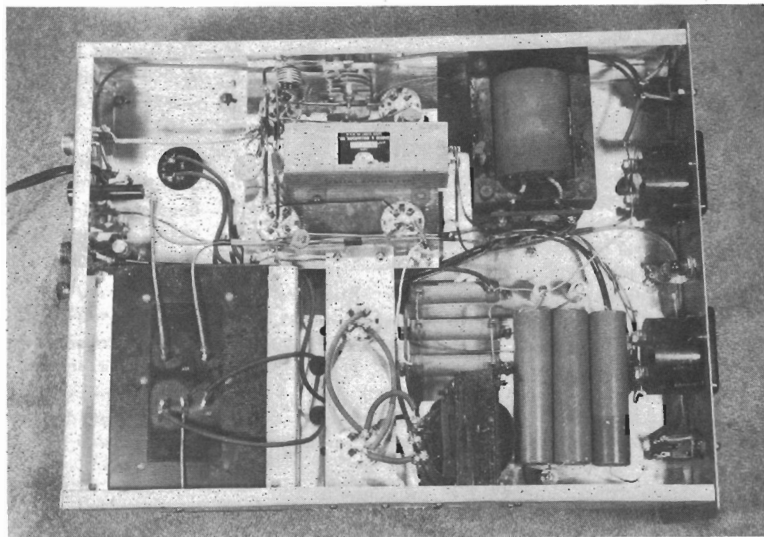
The amplifier operates at zero bias, but the control system is set up so that a relay applies about 100 volts of negative bias from the exciter in the stand-by condition to cut off plate current completely. Without the stand-by bias, the idling current for the four tubes will be

around 110 ma. Complete cutoff on stand-by allows these tubes to operate easily without forced-air cooling and, incidentally, is good insurance against "diode hash" noise while receiving.

Stabilizing

With a parasitic suppressor in the plate lead to each tube, there was no trace of instability in the amplifier, except on 10 meters, without neutralization. To assure yourself that the amplifier is stable, apply the plate voltage without bias, switch from band to band with no load applied and swing the input capacitor through its full travel. There should not be the slightest flicker of either the plate- or grid-current meters with no excitation applied. To correct the instability on 10 meters, a 6-turn coil (L_3), $\frac{1}{2}$ -inch diameter, was inserted in one of the common filament leads from the filament choke. A 5-turn coil (L_4), of hookup wire was wound over this. With a 6-pf. variable neutralizing capacitor (C_4) from L_4

This bottom view shows the submounting of the plate transformer, filter choke, and the rectifier and amplifier tubes. The filament choke (enclosed in a metal box), the neutralizing capacitor and neutralizing coils L_3 and L_4 may be seen in the upper center of the chassis.



to the plate tank circuit, neutralization on 10 meters was easily obtained.

Adjustment

Many articles have been published on the proper loading of linear amplifiers. In nearly all cases, the use of an oscilloscope is recommended. By all means, use a scope for initial tuning if you possibly get your hands on one. Another indispensable piece of equipment is an s.w.r. indicator. In case the scope is not always available, the output indication obtained from the s.w.r. meter can be used to get fairly near to optimum loading.

The following procedure has been checked by a scope to verify the results and was found to be quite satisfactory for this amplifier: Gradually apply carrier from the exciter up to about one half the rated output of the 100-watt-class exciter. Tune the linear amplifier pi-network input and loading capacitors to obtain maximum indication of output with the s.w.r. indicator in forward position. Increase the exciter output on up to full output and again retune the amplifier for maximum indicated output. Many will say that this is the proper loading point for the amplifier, but this has not been found to be true in all cases.

Having proceeded as stated above, reduce the inserted carrier until the plate current drawn by the four 811As is 200 ma. Then,

note the grid-current reading and the ratio of the plate-to-grid current. In this case, with 200 ma. of plate current, the grid current was 40 ma. (a ratio of 5 to 1). Then increase excitation to get 300 ma. a plate current, at which point the grid current should be 60 ma. In the event you reach a point where this ratio changes, further load changes in the amplifier will be required. For example, if the plate current goes to 500 ma. and the grid current required is greater than 100 ma., the amplifier is no longer linear. It has been found that both underloading and overloading will cause this condition. Readjust the output capacitance and reresonate the input capacitor until a linear relationship is attained. Then set the audio gain control so the plate meter of the amplifier never indicates more than 50 per cent of the maximum on voice peaks for single-sideband, suppressed-carrier operation. At this point the signal will be as good as the output of the exciter. No amplifier can improve upon that.

While the power supply for this particular amplifier does not allow it to be driven to a full kilowatt p.e.p., there is room for a 1500-volt (d.c.) transformer that will permit greater output. If a 1500-volt transformer is used, another 100- μ f. 450-volt electrolytic capacitor and 25K 25-watt bleeder resistor should be put in series with the three shown for the 1250-volt supply.

ONE-BAND KILOWATT AMPLIFIERS

Separate kilowatt amplifiers on each of the bands 80 through 10 meters has always been the *ne plus ultra* of transmitter construction. However, space limitations and cost are the two key factors that have prevented many from realizing this goal. The amplifiers to be described are compact and are constructed economically; the builder may wish to construct one amplifier for his favorite band or the group of five for versatile all-band operation. Advantages of the separate-amplifier philosophy include optimum circuit Q for every band, simplified construction and band switching, less chance for tube failure because each amplifier is pretuned, and fast band changing for the contest-minded. The supply voltages remain on all the amplifiers; only the filament and excitation power are switched to the desired final amplifier.

The availability and proven dependability of the 813 make a pair of them the logical choice for the kilowatt amplifier. A shrewd amateur should have no trouble procuring the tubes through surplus channels or by bartering with local hams.

Referring to the circuit diagram, Fig. 2, the

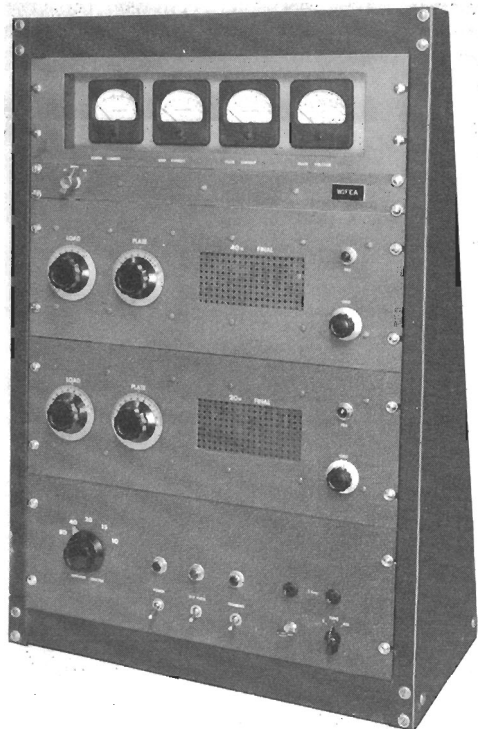


FIG. 1—Individual kilowatt amplifiers for two bands plus complete metering and all control circuits and power supplies (except plate) fit handily into a table rack. Amplifiers for five bands plus the plate supply will mount in floor rack. Band switch at lower left (S_8 in Fig. 2) switches filament supply, excitation and output connections to all amplifiers in use; screen and plate supplies are connected to all amplifiers at all times.

amplifier control unit contains the filament, bias and screen supplies. A 3-position mode switch, S_2 , selects the bias for either Class-AB₁ or -C operation, and in the third position grounds the screen grids, to limit the plate current during initial tuning. Another 3-position switch, S_1 , allows the total or individual screen currents to be read. The latter position is useful in matching tubes. The high-voltage supply should furnish from 1750 to 2250 volts.

Construction

Each amplifier is assembled on a 13×17 -inch aluminum bottom plate. Two $5 \times 13 \times 3$ -inch aluminum chassis are used as the sides of the enclosure. The paint is removed from the back of a 7-inch aluminum rack panel, and a piece of Reynolds cane metal is sandwiched between the panel and the two chassis. A rectangular window in the panel provides additional ventilation and a means for inspecting the color of the tube plates. The top and back of the enclosure are formed from a single piece of cane metal, bent to fit the chassis rear and top. Three lengths of $1 \times 1 \times 1\frac{1}{2}$ -inch aluminum angle stock are used in the corners of the enclosure, as can be seen in Figs. 4 and 5.

The variable tank capacitors, C_4 , are mounted on 1-inch stand-off insulators, to bring the shafts to the proper panel height. In the 10-meter amplifier the capacitor shaft must remain above r.f. ground, and a suitable insulated shaft coupling is used. On the other bands, the rotors of the capacitors are grounded to the chassis through metal straps.

On 20, 15 and 10 meters the tank coils are wound self-supporting of $\frac{1}{4}$ -inch diameter soft-drawn copper tubing, and they are supported by their leads. On 80 and 40 the coils are lengths of Air-Dux stock, and they are supported by small ceramic insulators.

The special plate r.f. chokes, RFC_2 , are constructed by close-winding No. 24 enameled wire on $\frac{1}{4}$ -inch diameter ceramic insulators. Four-inch long insulators (National GS-4) are used on the 80- and 40-meter bands, and 2-inch long insulators (National GS-3) are used on the other bands. In each case the original base of the insulator is removed and the insulator is mounted on a stand-off (Johnson 135-20). The high-voltage lead and the "cold" end of the choke are connected to a soldering lug mounted between the two insulators.

Bridge neutralization is included in the 20-, 15- and 10-meter amplifiers. The neutralizing capacitors are made from two $\frac{1}{2}$ -inch wide aluminum strips 5 inches long. One strip is connected directly to the plate lead at C_3 and the other is supported by a ceramic feed-through insulator that connects to the rotor of C_1 . The amplifiers are neutralized by adjusting the spacing between the aluminum strips.

The metal ring surrounding the base of the

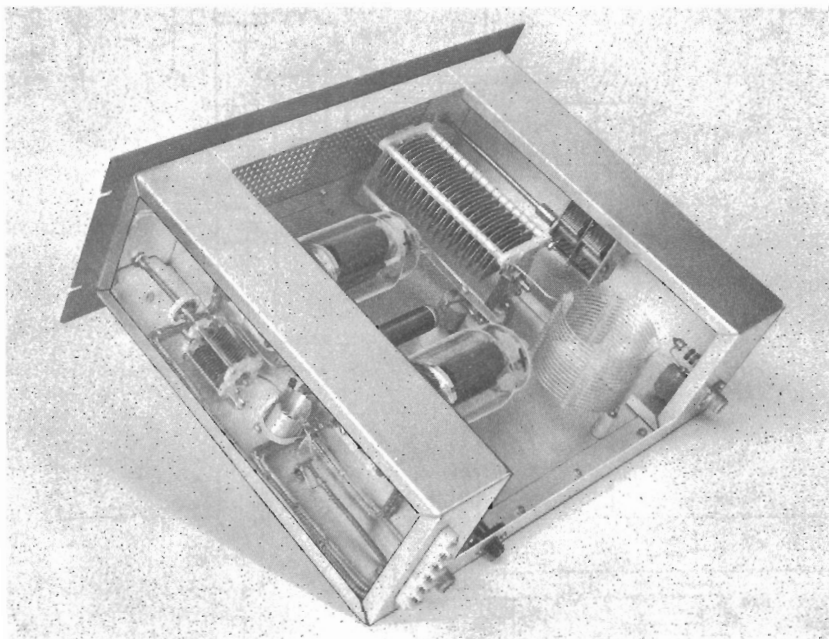


FIG. 3—View of the 80-meter amplifier with its cane-metal covering removed. As in each amplifier, the chassis is made from two $5 \times 13 \times 3$ -inch chassis and a 13×17 -inch base plate. Input and low-voltage leads make up to terminals and jack in center foreground.

813 should be grounded to the chassis. A piece of Eimac Finger Stock or a homemade contact can be used for the purpose.

All power wiring is done with shielded wire and bypassed as described in Chapter Twenty-three. The filament leads should be made from No. 14 (or heavier) shielded wire.

The screen and bias supplies plus station control circuits are built on a rack-mounting chassis (Bud CB-1373) behind a 7-inch panel. In the Class-C position of S_2 , +400 volts is

applied to the screens and -150 is connected to the grids. In the Class-AB₁ position, the screen voltage is increased to 700 and the grid bias is dropped to a value determined by the setting of R_g . This latter setting should be one that gives best linearity without exceeding a no-signal plate input of 150 watts for the two 813s; it depends on the plate voltage available. A heavy bleed on the screen supply helps the regulation.

The unit shown in Fig. 4 uses an Ohmite

COIL AND CAPACITOR TABLE

Band	80	40	20	15	10
C_1	100 μ f. (Johnson 100L15)	100 μ f. (Johnson 100L15)	50 μ f. (Johnson 50L15)	50 μ f. (Johnson 50L15)	50 μ f. (Johnson 50L15)
C_4	150 μ f. (Johnson 150E45)	150 μ f. (Johnson 150E45)	35 μ f. (Johnson 35E45)	35 μ f. (Johnson 35E45)	50 μ f. (Hammarlund MC-50-MX)
C_5	710 μ f. (2-gang 365 μ f.)	325 μ f. (Hammarlund MC-325-M)	325 μ f. (Hammarlund MC-325-M)	325 μ f. (Hammarlund MC-325-M)	325 μ f. (Hammarlund MC-325-M)
C_6	500 μ f. (Centralab TV-207)	100 μ f. (CRL 850S-100N)	—	—	—
L_1	4 t. No. 22*	3 t. No. 22*	2 t. No. 22*	1 t. No. 22*	1 t. No. 22*
L_2	32 t.p.i. No. 24, 1 inch long, 1 inch diam. (B&W 3016)	16 t.p.i. No. 20 1 $\frac{1}{2}$ inch long, 1 inch diam. (B&W 3015)	8 t.p.i. No. 18 1 $\frac{1}{2}$ inch long, 1 inch diam. (B&W 3014)	8 t.p.i. No. 18, $\frac{3}{4}$ inch long, 1 inch diam. (B&W 3014)	8 t.p.i. No. 18, $\frac{1}{2}$ inch long, 1 inch diam. (B&W 3014)
L_3	6 t.p.i. No. 12, 3 inch long, 3 inch diam. (Air Dux 2406)	4 t.p.i. No. 12, 3 $\frac{1}{2}$ inch long, 2 $\frac{1}{2}$ inch diam. (Air Dux 2004)	2 t.p.i. $\frac{1}{2}$ -inch copper tubing, 4 $\frac{1}{2}$ inch long, 2 $\frac{1}{2}$ i.d.	2 t.p.i. $\frac{1}{2}$ -inch copper tubing, 3 inch long, 2 $\frac{1}{2}$ i.d.	2 t.p.i. $\frac{1}{2}$ -inch copper tubing, 2 inch long, 2 $\frac{1}{2}$ i.d. C_1 tap 2 turns.

* Insulated hookup wire, wound over C_7 end of L_1 .

Model 111 switch at S_5 . This is ganged with antenna and excitation switches to permit one-control bandswitching. The relay K_1 is actuated when the plate supply is turned on; when

the relay is open a high bias is applied to the 813s to reduce the plate current to 0 ma. and eliminate receiver noise caused by static plate current.

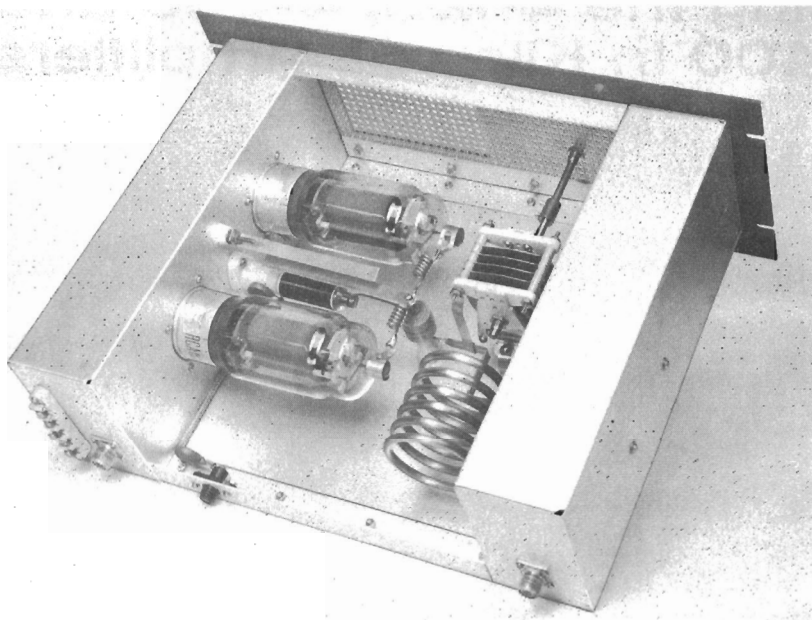


FIG. 4—Top view of the 15-meter amplifier. The neutralizing capacitor consists of two strips of aluminum, supported by the plate-blocking capacitor and a feedthrough insulator. It is mounted over the r.f. choke between the two 813 tubes.

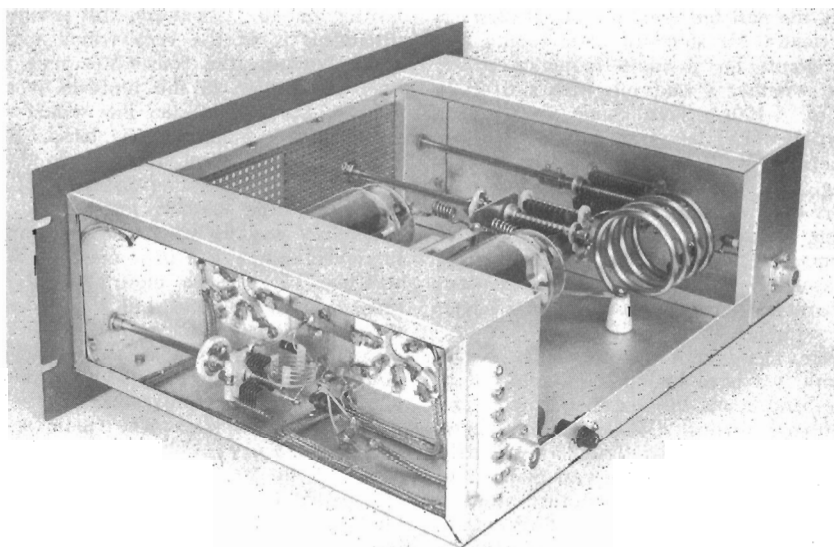
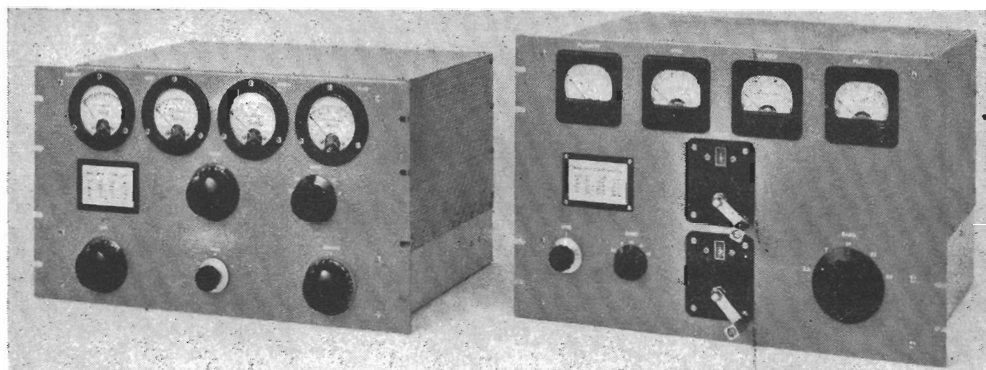


FIG. 5—As in the other amplifiers, the 10-meter final uses shielded wires in the filament, screen, and grid-return circuits. For tuning this amplifier uses a small variable capacitor connected across half of the plate coil, to maintain to favorable L/C ratio.

Using the 4-125, 4-250 and 4-400 in Kilowatt Amplifiers



Two amplifiers built to the same general circuit design and layout. The one on the left uses a pair of 4-125As, that at the right a pair of 4-250As. Both are capable of a kilowatt input on c.w. The 4-250As can handle envelope peaks of 2 kilowatts on sideband.

RAYMOND F. RINAUDO, W6KEV

During the past ten years the pi network has become almost the standard plate tuning and loading circuit for a radio-frequency power amplifier, whether it ends up with a 6146 or a pair of 4-400s. This came about quite naturally when TVI became a problem because the pi network lends itself very nicely to band switching, with tuning and loading done with capacitors—the capacitors, plate switch and coil being located in a comparatively small shield enclosure. All of this was had along with reasonably good harmonic attenuation: second harmonic down 35 to 40 db. and higher harmonics further attenuated.

However, along with the popularity of the pi network, we have had the development of the three-band beam, the multiband dipole and the multiband vertical. While the multiband antenna is a godsend to those with limited acreage, in which category the vast majority of us fall, it serves to bring up another problem because we now find that 35- or 40-db. attenuation of the second harmonic at the amplifier is quite often not enough. The multiband antenna is all too ready to radiate that 20-meter harmonic when the amateur is actually trans-

mitting on 40. Obviously, the antenna under discussion is of the type which requires no tuner between the transmitter and the feed line. A solution to the problem is to put a filter in the transmission line which will pass only the frequencies in one band. But then, when changing to a different band, another filter must be substituted and some of the ease of band change has been lost.

Another way in which the situation can be improved is to use a pi-L network. The pi-L will give 10 to 15 db. more attenuation of the second harmonic than will the pi¹ and even more attenuation of the higher harmonics. This circuit has been used in some commercially-built amateur equipment such as the Collins KWS-1. Further improvement can be had by designing the amplifier plate circuit for a higher loaded *Q*. For example, raising the loaded *Q* from 10 to 20 will increase the harmonic attenuation by 6 db. Unfortunately, one runs into the law of diminishing returns here; the losses in the plate coil begin to be large enough to cause serious heating, and a loaded *Q* of 20 is near the practical upper limit in most cases.

From July, 1962, *QST*.

¹ *Fundamentals of Single Side Band*, Collins Radio Company.

The 4-125A Amplifier

With the harmonic problem in mind, a design was worked out for an amplifier which was to replace the pi-network final then in use. The requirements were as follows:

1) Operation from 3.5 to 28 Mc., band switched.

2) Power input of 1 kw. with 2500 volts on the plate.

3) R.f. power output to feed into a 50-ohm coaxial load.

4) Standard 19-inch rack mounting with a minimum practical panel height.

5) Amplifier enclosed in a shield and incoming power leads bypassed for TVI.

6) Harmonic radiation via the feed line to be minimized.

7) A minimum amount of cash to be involved.

The result of the above design is the 4-125A amplifier shown in the photographs.

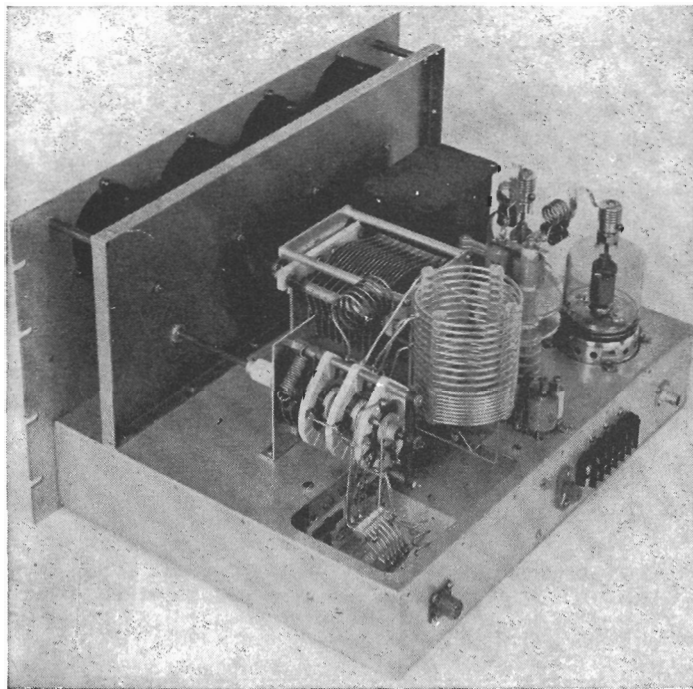
The amplifier uses a pair of neutralized 4-125As in parallel. The grid circuit is tuned, fairly high C , and makes provision for bridge neutralization via bypass condenser C_2 , Fig. 1. The plate circuit is a pi-L network with an operating Q of 15, and plate current is shunt fed. Individual meters are used to measure grid, screen and plate currents and filament voltage.

The amplifier is built on a $13 \times 17 \times 3$ -inch aluminum chassis behind a $10\frac{1}{2} \times 19$ -inch panel. The meters are excluded from the r.f. field by a 7×17 -inch aluminum sub-front panel which is set back two inches from the front panel. The resulting enclosure, which is above the chassis and screened by perforated aluminum, is 11 by 17 by 7 inches. The under-

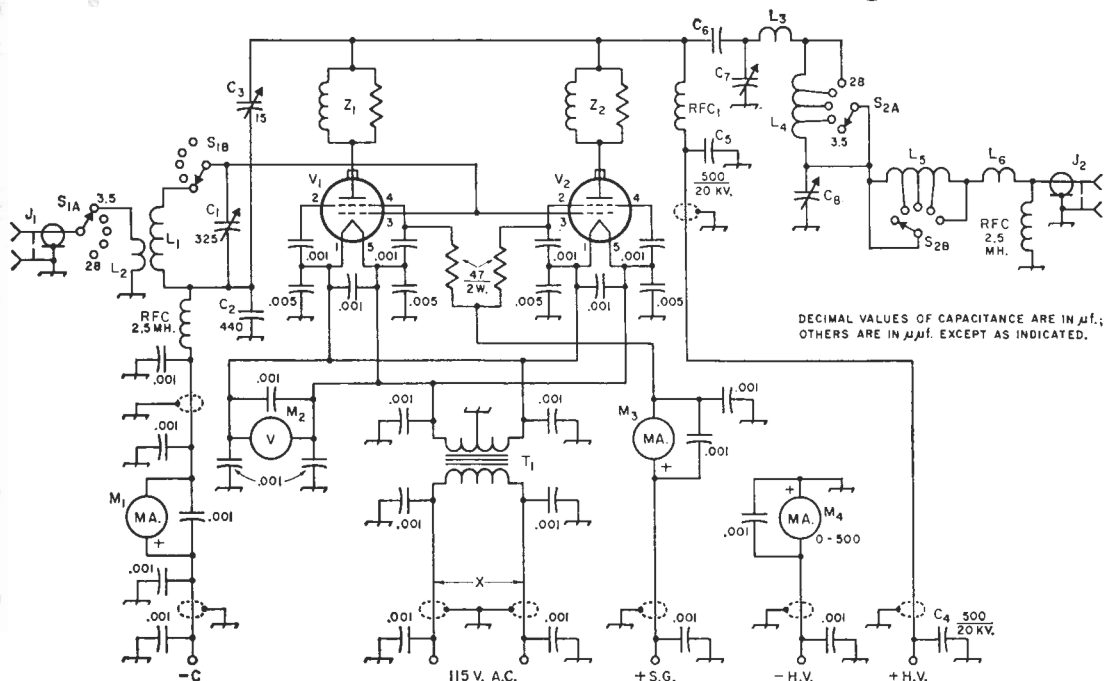
side of the chassis is divided into two units by a shield running from front to back. The grid compartment is 10 by 13 by 3 inches and the output compartment containing the loading capacitors and L net coil is 7 by 13 by 3 inches. The bottom of the chassis is covered by perforated aluminum sheet to allow convection air currents to cool the tubes. No blowers or fans are needed to cool the 4-125As, provided that cooling air is allowed to flow freely past the tubes.

In keeping with requirement (7), maximum use was made of the surplus markets and trades with fellow hams, and the author's own junk boxes were given a thorough going over. No real compromise was made by the use of interior components, but inevitably several of the parts used are either not too commonly-available surplus items or are once-standard parts which are no longer manufactured. But for each of these, a standard commercial part exists which is as good as or better than the one used and will fit into the space available. The standard part is the one given in the parts list. That the use of used and surplus parts paid off is attested to by the fact that the immediate cash outlay was less than \$20! On the other hand, if the reader wants to build the amplifier using all new parts, the cost will be approximately \$235, including tubes.

As mentioned previously, the grid circuit operates with fairly high C . Approximately $300 \mu\text{f.}$ is used on 3.5 Mc., $150 \mu\text{f.}$ on 7 Mc., and proportionally smaller amounts for the higher-frequency bands. A large tuning capacitance is used so that there will be a minimum of clipping of the waveform of the driving signal when the grid is driven positive. A dis-



Chassis view of the 4-125A amplifier. The plate tuning capacitor is at the center. The pi coil for 3.5-21 Mc. is vertical. The 28-Mc. pi coil is mounted horizontally between the band switch and the tank capacitor. The plate r.f. choke and neutralizing capacitor are partially hidden by the plate coil. The filament transformer is at the far end of the chassis between the 4-125As and the sub-front panel.



DECIMAL VALUES OF CAPACITANCE ARE IN μf .; OTHERS ARE IN μf . EXCEPT AS INDICATED.

FIG. 1—The amplifier circuit. Either 4-125As, 4-400As or 4-250As may be used at V_1 and V_2 . See specification below for circuit values that differ with the two types. "X" indicates point where the two cooling-fan motors are connected in the 4-250A amplifier. Shielded wiring in supply leads is continued up to the bypass capacitors nearest the r.f. circuit. All 0.001- and 0.005 μf . capacitors are disk ceramic, 1000-volt rating.

C_1 —320 μf ., 0.0245-inch spacing (Hammarlund MC-325-M).

C_2 —440 μf ., silver mica (two 220- μf . in parallel).

C_3 —Disk neutralizing, 2.2-15 μf . (Millen 15011).

C_4 , C_5 —500- μf ., 20-kv. ceramic (Centralab TV-20).

J_1 , J_2 —Coaxial chassis-mounting connectors.

L_1 —3.5 Mc.: 32 turns no. 20, $\frac{3}{4}$ -inch diam., 16 turns per inch (Air Dux 616T).

7 Mc.: 14 turns No. 20, $\frac{3}{4}$ -inch diam., 16 t.p.i. (Air Dux 616T).

14 Mc.: 11 turns No. 18, $\frac{5}{8}$ -inch diam., 8 t.p.i. (Air Dux 508T).

21 Mc.: 9 turns No. 18, $\frac{1}{2}$ -inch diam., 8 t.p.i. (Air Dux 408T).

28 Mc.: 6 turns No. 18, $\frac{1}{2}$ -inch diam., 8 t.p.i.

(Air Dux 408T).

L_2 —3.5 Mc.: 4 turns insulated hookup wire at cold end of L_1 .

7 Mc.: 3 turns same.

14, 21, and 28 Mc.: 2 turns same.

M_2 —0-8 or 0-10 volts a.c.

M_3 —0-500 ma. d.c.

RFC₁—Transmitting choke (B & W 800, National R-175A, Raypar RL-100).

S_2 —Ceramic, 2 poles, 5 positions (Radio Switch Corp., Marlboro, N.J.); see text.

V_1 , V_2 —4-125A or 4-250A.

Z_1 , Z_2 —4 turns No. 12, $\frac{1}{2}$ -inch diam., $\frac{1}{2}$ inch long, with four 220-ohm, 2-watt composition resistors in parallel.

For 4-125As:

C_6 —0.001- μf ., 20-kv. ceramic (two Centralab TV-20s in parallel).

C_7 —250- μf ., 3000-volt variable (Johnson 154-9).

C_8 —0.001- μf ., 2000-volt variable (two Johnson 154-3 in parallel, ganged).

L_3 —6 turns No. 10, 1-inch diam., $1\frac{1}{2}$ inches long.

L_4 —Vari-pitch Air Dux 2408D4, modified as described in text.

L_5 —Indented Pi Dux 1411A, modified as described in text.

L_6 —4 turns No. 14, $\frac{3}{4}$ -inch diam., $1\frac{1}{4}$ inches long.

M_1 —0-50 ma. d.c.

M_3 —0-100 ma. d.c.

S_1 —1 section, 2 poles, 5 positions (Centralab 2505).

T_1 —5 volts, 13 amp. (Triad F9A or F15U).

For 4-250As or 4-400As

0.002- μf ., 20-kv. ceramic (four Centralab TV-20s in parallel).

300- μf ., 10-kv. variable (Jennings UCS-300).

0.0012- μf ., 3000-volt variable (Jennings UCSL-1200).

See text.

Illumitronic Pi Dux No. 195-2, modified as described in text.

Vari-pitch Air Dux 1608D6, tapped as described in text.

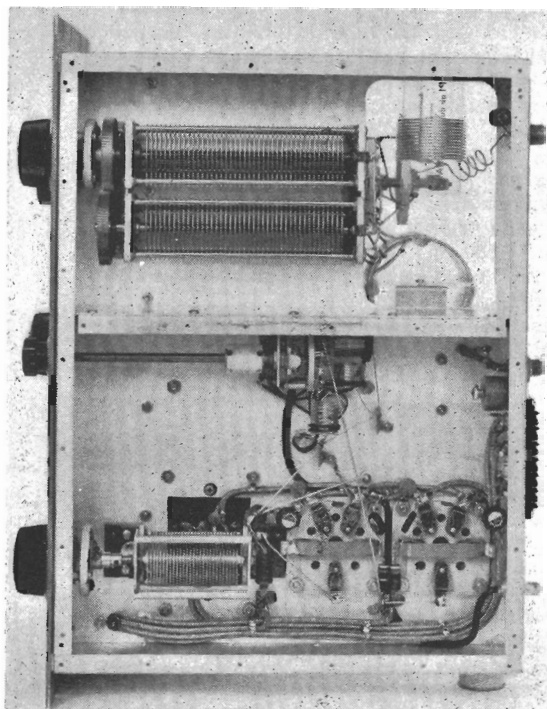
6 turns No. 12, 1-inch diam., 1 inch long.

0-100 ma. d.c.

0-200 ma. d.c.

2 sections, 1 pole per section, 11 positions, 5 positions used (Centralab YD sections with P-270 index assembly); see text.

5 volts, 29 amp. (Stancor P-6492).



Below the 4-125A chassis. The two loading capacitors at the top, ganged together by means of gears, are separated from the grid circuit by an aluminum shield running from the front to the rear of the chassis. The L net coils are directly behind the loading capacitors. The grid band switch and coil are at about the center of the chassis. The grid tuning capacitor is mounted off the chassis by means of bakelite blocks and is directly under the filament transformer.

torted waveform at the grid of an amplifier will mean more harmonic signal in the plate circuit and, hence, a more difficult job to suppress it. For example, in Class C, B or AB₂ operation, during the portion of the cycle that the grids of two 4-125As in parallel are driven positive, the grids look like a resistor of about 1200 ohms to the tuned grid circuit, and that portion of the cycle will be distorted unless precautions are taken to prevent it. Waveform clipping is minimized by using plenty of tuning capacitance in the grid circuit. Of course, for AB₁ operation, the amount of C is not important because the grid is never driven positive and looks like an infinite resistance to the tuned circuit.

The grid tank circuit uses individual coils for each band. A link coil of insulated hook-up wire is wound over the cold end of each coil. The hot ends of the coils and the links are switched by S₁, a 2-section, 5-position switch having one wafer. This switch is mounted on the underside of the chassis by means of an aluminum bracket. The coils are mounted between the appropriate switch ter-

minal and a tie point and are oriented so that there is a minimum of coupling between them.

The tube sockets are mounted on the underside of the chassis, and spring clips on the top of the chassis held by the socket mounting bolts ground the metal tube base shield. Bypassing of the screen and filament terminals is done in the more-or-less standard way. The screen terminals on each socket are connected together by a 1/8-inch wide strip of thin copper. Each screen terminal is then bypassed to the nearest filament terminal with a disk ceramic capacitor. A disk ceramic capacitor is connected between the filament terminals and another is used to bypass one side of the filament to ground. The remaining filament terminal is grounded with a short, heavy lead. Grounding one side of the filament has been found to be helpful in eliminating v.h.f. parasites. Those who expect to use the amplifier for linear service should use bypass capacitors to ground on both filament terminals and ground the filament transformer center tap as shown in Fig. 1. A slightly cleaner signal will result. A 47-ohm, 2-watt carbon resistor is used to feed screen voltage to the screen terminal of each socket and is a parasitic preventive measure.

The pi-L plate tank circuit is made up of individually available coils, capacitors and switch. The switch, which is mounted on the chassis with an aluminum bracket, is made by Radio Switch Corporation of Marlboro, New Jersey. The switch used was bought on the surplus market and has three wafers, each wafer with six contacts. Since the wafers were already there it was decided to make use of them by paralleling the contacts on two wafers and using the parallel combination to switch the coil in the pi portion of the network. The circulating current in the pi coil is about twice as high as that in the L coil. However, the current rating of the switch is 20 amperes, so a single section is all that is really needed to handle the pi coil switching. Also, because six contacts per wafer were available, the sixth contact was used to provide a 3.8-Mc. position; that this is not necessary can be seen by the later description of the 4-250A amplifier. If the builder wishes to retain the 6-position band switch, he should order a Model 86 switch, standard bearing, non-shorting, 30-degree detent, with two Type A wafers. If a 5-position band switch will do, then the builder should order a Model 86 switch, standard bearing, non-shorting, 30-degree detent, with one Type B wafer. The second switch, by virtue of having only one wafer, will cost about three dollars less. The coils used in both the pi and the L are home-brew for 28 Mc. Illumitronic Engineering Pi Dux coils are switched in for the lower-frequency bands. It is of interest to note that as originally built, the pi coil was a Pi Dux 2007A, which is wound with No. 12 wire. After a bit more than two years' use, two of the turns shorted because coil heating had softened the polystyrene insulating supports.

The damaged coil was replaced with a Pi Dux 2408D4, which is made of No. 10 wire. An r.f. choke completes the output circuit to ground for d.c. as a safety precaution.

Drive power is fed into the amplifier through a BNC coaxial receptacle and the output power is taken out by means of a U.H.F. receptacle. Plate voltage is fed in through a Millen 37001 high-voltage terminal.

When the amplifier was completed, it was first tested for parasitics without suppressors of any kind. As is almost always the case with a tetrode or pentode amplifier, it oscillated merrily in the v.h.f. range—at about 150 Mc., as a check with the grid-dip meter showed. The parasitic was killed by the installation of suppressors, Z_1 and Z_2 , in the plate lead to each tube. The test for parasitics is to operate the amplifier with reduced plate and screen voltage and no fixed bias on the control grid, but using a grid leak of about 5000 ohms to ground to develop bias if the amplifier breaks into oscillation. No drive is used and no load is connected to the output. With this amplifier the plate voltage was set at 1000 volts and the screen voltage increased until the plate current was about 200 ma. and the tubes were dissipating about 100 watts each. At this point the screen voltage was between 150 and 200 volts. If an amplifier can be operated in this manner with no current showing on the grid meter, with no change in plate current, and with no detectable r.f. in the amplifier as the grid and plate tuning, loading and band-switch controls are tuned through their full range, then the amplifier can be considered adequately stable. This is a much more severe test than the one often made where full plate

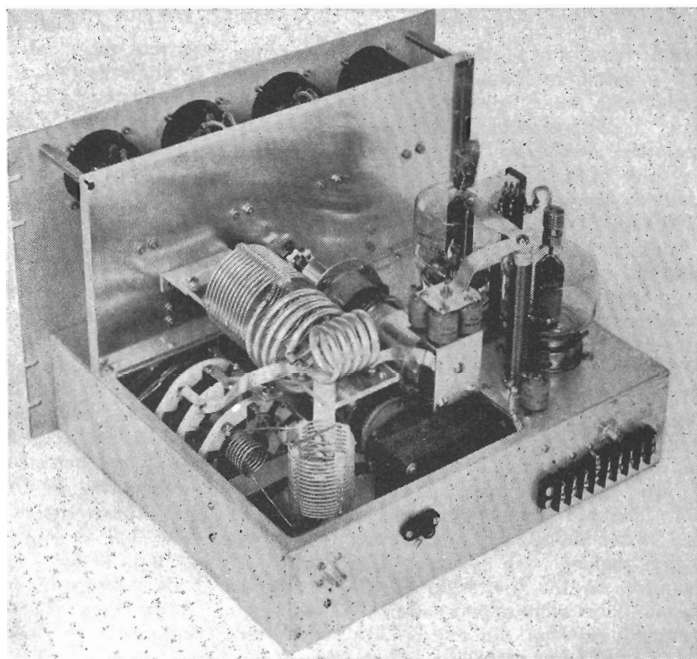
and screen voltages are applied and bias is reduced until the tube or tubes are dissipating full rated power with no excitation.

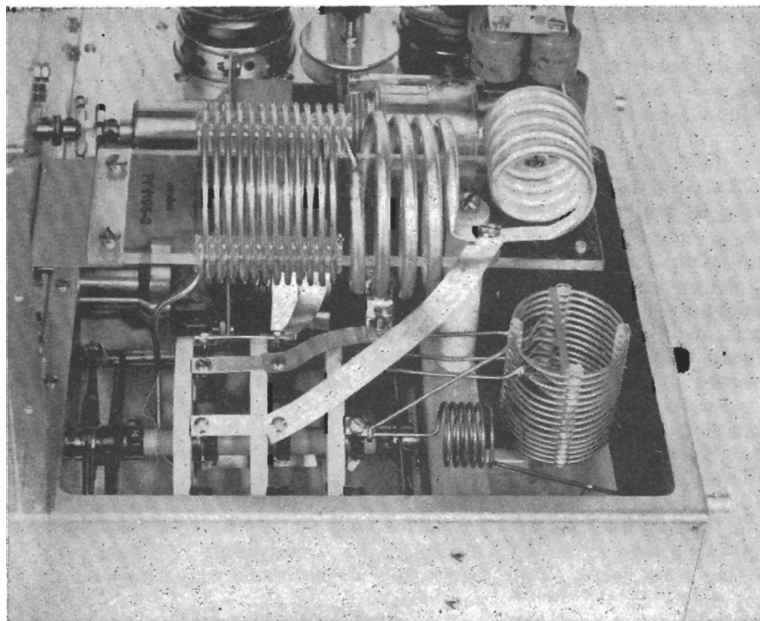
With the components used, the amplifier will operate with up to 3000 plate volts in Class C c.w. or Class AB₁ linear, or up to 1500 volts for Class C a.m. plate-modulated service. Screen voltage for Class C c.w. or plate-modulated a.m. is 350 volts; for AB₁ linear it should be 600 volts. Grid bias should be -100 to -150 volts for Class C c.w., -210 for Class C, plate modulated, and approximately -95 volts for Class AB₁. The exact value of bias for AB₁ should be adjusted for the required idling plate current for the voltage used. Recommended values are as follows, for two tubes: 2000 volts, 85 m.a.; 2500 volts, 70 ma.; 3000 volts, 60 ma.

The screen voltage for Class C operation should come from a separate supply of reasonably good regulation. A series dropping resistor from the plate supply is not recommended. For Class AB₁ service, the screen voltage should be well regulated. While an electronically-regulated supply can be used, the simplest method is to use a string of VR tubes in series with a resistor from the plate supply. The reason that Class AB₁ permits this simple method of getting screen voltage is that the screen current excursions are not very great and are well within the capability of VR tubes.

Both the plate tank coil, L_4 , and the L-network coil, L_3 , are mounted on the chassis by means of aluminum angles bolted to the plastic mounting strip furnished with each coil. L_4 is modified and tapped as follows: Turns are removed from the close-wound end until 18 turns remain. Starting from the end of the coil which

In the 4-250A amplifier the band switch is in a chassis cutout with the pi coil above it. The 28-Mc. L coil and the vertically-mounted 3.5-21 Mc. L coil are between the switch and the rear of the chassis. The plate blocking capacitors are mounted on a bracket held by the vacuum variable plate-tuning capacitor at the center. The plate r.f. choke and its bypass capacitor are beside the 4-250A nearest the rear of the chassis.





Close-up of the 4-250A pi-L plate tank coils and band switch. The horizontally-mounted 6-turn wire coil is L_6 , and the vertically-mounted coil is L_5 .

has the wide-spaced turns, the 21-Mc. tap is at 2 turns, the 14-Mc. tap at the 4th turn, the 7-Mc. tap at the 8th turn, and the 3.8-Mc. tap (if used) at the 16th turn. Since the 16th turn is in the close-wound portion of the coil, it is much easier to make the tap if the turn on each side is pushed in toward the center of the coil.

The L-network coil, L_3 , is modified by removing turns until 13 turns remain. Starting from the L_6 end, taps are placed as follows: 21-Mc. tap at 2 turns, 14-Mc. tap at 5 turns, 7-Mc. tap at 7 turns, 3.8-Mc. tap (if used) at 12 turns.

Non-standard items used are the grid tuning capacitor from the surplus market, the neutralizing capacitor, which National Radio Company no longer makes, the plate tuning capacitor, no longer made by E. F. Johnson, and the loading capacitor, C_8 , which is made up of two capacitors taken from a surplus BC-653 transmitter. The two E. F. Johnson units specified for C_8 will simplify the ganging of the two because they have the shaft out the back as well as the front. The two surplus capacitors did not have this feature and, consequently, gears had to be used for ganging.

The tuning and loading adjustments of the pi-L plate circuit are exactly the same as with a pi network. Plate circuit loading is increased by reducing the capacitance of C_8 . Whenever the loading capacitance is changed, the plate circuit must be retuned to resonance with the plate tuning capacitor, C_7 .

When the amplifier is first tested, it should be neutralized. The neutralizing capacitor, C_3 , is adjusted so that there is about one-half inch spacing between the two plates; then, with

plate and screen voltages off and a load connected, excitation is applied and the grid circuit is tuned to resonance. The excitation level is set so that the grid current is only a few ma. Then plate and screen voltages are applied and the plate circuit is tuned to resonance. Plate-circuit resonance is best indicated by the peaking of the screen-grid current as the plate tuning capacitor is tuned through resonance. The loading control is adjusted so that the screen current is about 60 ma. If the plate input is less than desired, increase the grid drive and plate loading until the correct plate current is flowing with screen current at 60 ma. The plate circuit must be returned to resonance with each change of loading.

The check for neutralization is to tune the plate circuit through resonance, observing both screen and grid currents. When the amplifier is correctly neutralized, the grid-current meter will show a small current peak at the same setting of the plate tuning capacitor that gives a peak in screen current. Neutralization should be done on the 21-Mc. band.

After the amplifier has been neutralized it should be checked for parasitic oscillations, using the procedure given previously. In some cases, parasitics will make it difficult to find the correct neutralization setting. But if construction details are followed, particularly those pertaining to bypassing and the installation of suppressors, parasitic oscillations should not be a problem.

The 4-250A Amplifier

Quite some time after the 4-125A amplifier had been completed and had been operating satisfactorily, a design for a de luxe version

was worked out. In this case, the requirements were the same as before except that the rig had to be capable of 2-kw. p.e.p. input for sideband service, and all the parts used were to be currently-available new items. The result uses 4-250As or 4-400As and is shown in the photographs.

The 4-250A amplifier uses essentially the same circuit as the 4-125A version. However, the plate circuit was designed for an operating Q of 18 instead of 15, in order to take advantage of the heavy-duty plate coil and switch which were to be used. An examination of the photographs shows the similarity of the two rigs in the mechanical layouts and the method of making the shield enclosures. Because all new parts were used, the second amplifier turned out to have a better appearance both inside and out than did the first one.

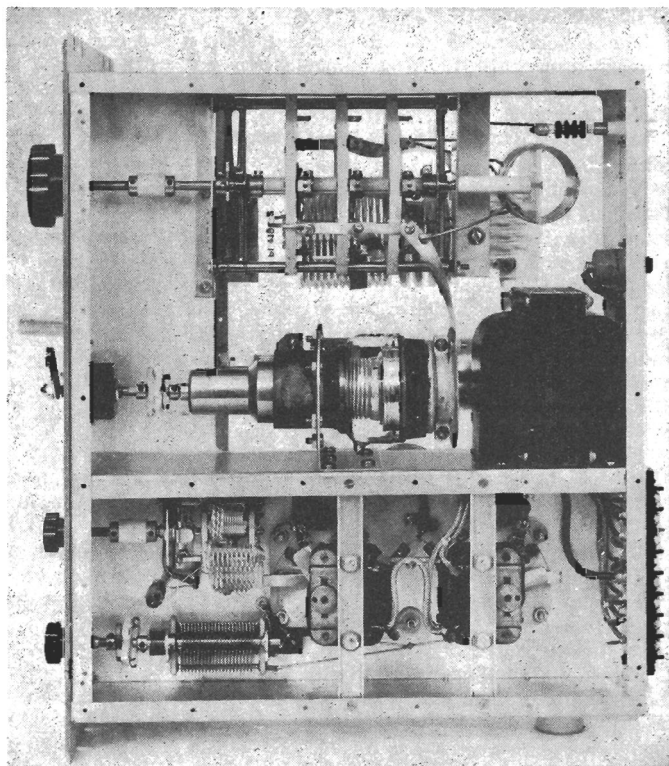
The 4-250A amplifier is built behind a standard 19-inch rack panel 12 $\frac{1}{4}$ inches high. The chassis is 17 by 15 by 4 inches and the shield enclosure above the chassis is 17 by 12 by 7 $\frac{1}{4}$ inches. The vertical sub-panel is set back three inches from the front panel. The grid-circuit compartment is 6 $\frac{1}{4}$ by 15 by 4 inches and is separated from the rest of the under-chassis space by a shield which runs from front to back. The remainder of the underside of the chassis is opened up to the upper compartment by cutting out that portion of the chassis top. This increases the available space for the plate-circuit components and

makes it much easier to connect the various parts together. The filament transformer is mounted in the plate-circuit area on the under-chassis shield which forms the grid compartment. Both the shield enclosure above the chassis and the bottom cover are made of perforated aluminum, which allows convection currents to help keep the tubes and parts cool. The 4-250As require forced-air cooling of the base, and small Barber-Coleman fans are used to blow air directly upward at the tube base pins and through the holes in the tube socket and tube base.

The grid circuit coil turret is made up of commercial coil stock and rotary switches. Two rotary switch wafers are used where only one is really needed to do the necessary switching. The coils are mounted between the wafers on the switch lugs and the axes of all the coils are parallel. The arrangement used with the 4-125As where coils for adjacent bands are at right angles is better, because odd resonances in unused coils are less likely to cause trouble. However, this arrangement has been perfectly satisfactory in performance and is more rugged mechanically.

The cooling-fan motors are mounted on home-brew shock mounts to reduce noise. Rubber grommets with the same spacing as the motor mounting studs are mounted in the support channels which hold the motors, then a sleeve of length equal to the thickness of the grommet is slipped into each grommet. A large

Bottom view of the 4-250A amplifier. The plate band switch, at the top, is mounted on aluminum brackets. The vacuum variable loading capacitor is at the center and the filament transformer is between it and the rear of the chassis. The bracket which supports the loading capacitor also supports the plate tuning capacitor. The grid band-switching turret and tuning capacitor are at the front of the grid compartment. A cooling fan is mounted directly below each tube socket.



washer is placed on each side of the sleeve before the mounting screw is passed through and threaded into the motor mounting stud. Two shock mounts are needed for each motor. The grommet size used is that which fits into a $\frac{3}{8}$ -inch hole. A $\frac{1}{4}$ -inch diameter sleeve $\frac{1}{4}$ inch long is the right size to fit the grommet hole. There is no reason, though, why larger grommets and sleeves cannot be used.

Vacuum variable capacitors are used for plate tuning and loading. These require 24 and 30 turns, respectively, to cover the full capacitance range. Counter dials which read each tenth of a turn are used to drive them. The dials are made by Gates Radio Company, Quincy, Illinois, part No. M3401F. These were chosen because they are r.f. tight and do not require much space behind the panel.

The plate-circuit switch is made by Radio Switch Corporation. It is a Model 88 with 36-degree detent and three Type A wafers. Two of the three wafers are paralleled and switch the pi coil. The remaining one handles the L coil.

The Illumitronic coils used in the plate circuit both require modification. The 28-Mc. pi coil should be removed and replaced with one of slightly greater inductance consisting of 5 turns of $\frac{3}{16}$ -inch copper tubing, 1 $\frac{1}{2}$ inches in diameter and 2 inches long. The remainder, L_4 , of the pi coil should be modified by removing turns from the wire end, leaving 12 $\frac{1}{2}$ turns. Turns are removed from the close-wound end of the L coil, L_5 , until 15 turns remain. The 28-Mc. L coil, L_6 , is home-brew. The taps on the pi coil are placed as follows: 28 Mc.: junction of L_3 and L_4 ; 21 Mc.: 2 $\frac{1}{2}$ turns from the 28-Mc. tap; 14 Mc.: 5 $\frac{1}{2}$ turns from the 28-Mc. tap; 9 Mc.: 9 turns from the 28-Mc. tap. The taps on L_5 are as follows: 28 Mc.: at junction of L_5 and L_6 ; 21 Mc.: 3 turns from the 28-Mc. tap; 14 Mc.: 5 turns from the 28-Mc. tap; 7 Mc.: 9 turns from the 28-Mc. tap. An r.f. choke is used to complete the d.c. circuit to ground at the coax output connector as a safety measure should the plate blocking capacitor, C_6 , break down.

A type BNC receptacle is used to feed drive power into the amplifier and a type C receptacle at the output. The d.c. plate voltage is fed into the amplifier via a Millen high-voltage terminal, type 37001.

Many combinations of plate, screen and bias voltages can be used, as a look at a tube data sheet will show. The following voltages are typical:

	C.W.	A.M.	Phone	AB ₁	Linear	
Plate	2500	2500	3500			volts
Screen	500	400	555			volts
Grid	-150	-200	-105°			volts

* Set to give 45-ma. plate current per tube with no drive power.

The tune-up procedures are the same as for the 4-125A. Also, the amplifier should be

checked for parasitics as described previously. Best linearity is achieved by increasing the loading on the amplifier until the power output just starts to fall off; during this adjustment, the drive power is held constant.

In operation, there is little to choose between the two rigs for the c.w. man. At 1-kw. input on c.w., the amplifiers handle identically; however, the 4-250As are easier to drive. For a 2-kw. p.e.p. input on s.s.b., the 4-250A amplifier stands alone. Which version the builder chooses depends upon his requirements as balanced against the necessary cash outlay. It should be pointed out that a third version combining the better or less-expensive components of the two designs presented could be built around 4-250As and result in an amplifier not costing much more than the strictly economy 4-125As.

Design of the Pi-L Network

The design of the pi-L tank circuits has been covered before in excellent articles presented in *QST*.^{2,3} However, two different approaches are again presented here for those who would like to apply the circuit to transmitters of their own design.

The first method is to use values of components for the pi network with which the builder is already familiar and alter them suitably to make the pi-L work. Figs. 2A and 2B show a pi and a pi-L network, either of which will match a power amplifier tube to a 50-ohm load.

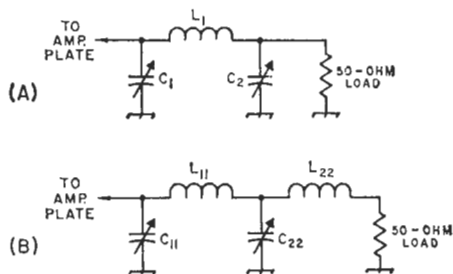


FIG. 2

First, the capacitance and voltage rating of C_1 and C_{11} are exactly the same for both circuits. The capacitance of C_{22} will be about one-half to two-thirds that required for the pi capacitor, C_2 . The voltage rating of C_{22} must be three or four times that required for C_2 . The inductance L_{11} will be greater than L_1 by about 25 per cent. The inductance L_{22} , which has no direct counterpart in the pi, will have an inductance of about one-third to one-half of L_{11} . The circulating currents in L_{11} are the same as in L_1 ; therefore, a coil made of a wire size suitable for a pi net will also be good for a pi-L. The currents flowing in L_{22} are much smaller than those in L_{11} , so it can be made of smaller wire. For example, if L_{11} must be made of No. 10 wire, L_{22} could be made of No. 14 or 16.

This approach will allow the intrepid experimenter to convert his present pi-network output circuit to a pi-L without much pain. But for those who prefer a more formal method, the following is offered:

Just as the pi is designed as two L networks placed back-to-back, the pi-L is designed as three L nets placed

² Miedke, "Pi and Pi-L Design Curves," *QST*, November, 1955.

³ Grammer, "Simplified Design of Impedance-Matching Networks," *QST*, March, April and May, 1957.

back-to-back. In Fig. 3, a pi-L tank circuit is broken down into its three equivalent Ls. The first L matches the desired tube load resistance R_1 to a resistance R_0 and

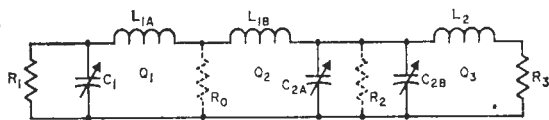


FIG. 3

is composed of C_1 and L_{1A} . The second L matches R_0 to the resistance R_2 and is made up of L_{1B} and C_{2A} . The third L matches R_2 to the load R_3 (the transmission line) and consists of C_{2B} and L_2 . R_1 is determined from the approximate formula:

$$R_1 = \frac{E}{2 \times I}$$

where E = plate voltage applied to the tube
and I = plate current in amperes.

First, the value of Q_1 is selected. Q_1 is the operating Q of the plate circuit and is usually chosen to be between 10 and 20. Knowing R_1 and Q_1 the capacitive reactance X_{C1} of the plate tuning capacitor C_1 is calculated from:

$$X_{C1} = \frac{R_1}{Q_1}$$

Also, calculate R_0 from:

$$R_0 = \frac{R_1}{Q_1^2 + 1}$$

Then, calculate the inductive reactance X_{L1A} of the inductance L_{1A} from:

$$X_{L1A} = R_0 Q_1$$

This completes the calculation of the reactances for the first L network.

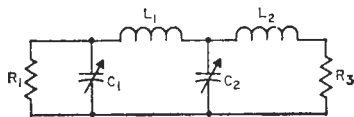


FIG. 4

Before proceeding with the second L network, which consists of L_{1B} and C_{2A} , the value of R_2 should be selected (R_2 must always be greater than R_0 and R_3). Although it is possible to arrive at operating values for L_{1B} and C_{2A} by first selecting Q_2 (the Q of the second L network), it is best, from an equipment designer's viewpoint, to calculate R_2 to match the voltage capability of available tuning capacitors. This is done from:

$$R_2 = \frac{E^2}{P}$$

where E = r.m.s. voltage across R_2
 P = Amplifier power output in watts.

Because the peak voltage must be considered when determining capacitor voltage breakdown (peak voltage equals 1.41 times r.m.s. voltage) and some safety factor is desired, it is best to let E equal one-half the capacitor

breakdown voltage. For a kilowatt transmitter, it is suggested that 1000- to 2000-volt capacitors be considered. Convenient values of power output can be calculated by assuming an efficiency of 75 per cent for a c.w. or plate-modulated amplifier, and 60 per cent for a linear. Don't forget that for an a.m. phone rig, the power output at the crest of a 100-percent-modulated envelope is four times the carrier output.

Having calculated R_2 , proceed with determining Q_2 from:

$$Q_2 = \sqrt{\frac{R_2}{R_0} - 1}$$

Calculate X_{C2A} (capacitive reactance of C_{2A}) from:

$$X_{C2A} = \frac{R_2}{Q_2}$$

Calculate X_{L1B} (inductive reactance of L_{1B}) from:

$$X_{L1B} = R_0 Q_2$$

Next, the capacitive and inductive reactances for the third L network, C_{2B} and L_2 , are calculated. First, calculate Q_3 , the Q of the third L net, from:

$$Q_3 = \sqrt{\frac{R_2}{R_3} - 1}$$

where R_3 is the load that the amplifier will be working into, usually 50 ohms for coax feed lines. It can be almost anything else but must be less than R_2 . Then determine X_{C2B} , the capacitive reactance of C_{2B} , from:

$$X_{C2B} = \frac{R_2}{Q_3}$$

Then calculate X_{L2} , the inductive reactance of L_2 , from:

$$X_{L2} = R_3 Q_3$$

Since the inductances L_{1A} and L_{1B} are in series, these are combined in one coil, L_1 . The inductive reactance is equal to the sum of the separate parts

$$X_{L1} = X_{L1A} + X_{L1B}$$

Similarly, the two capacitors C_{2A} and C_{2B} are in parallel and are combined in one capacitor, C_2 . X_{C2} , the capacitive reactance of C_2 , is obtained by

$$X_{C2} = \frac{X_{C2A} X_{C2B}}{X_{C2A} + X_{C2B}}$$

The actual values for the capacitors and coils can be determined for any frequency from:

$$C = \frac{10^6}{2\pi f X_C}$$

and

$$L = \frac{X_L}{2\pi f}$$

where C = Capacitance in μf .
 L = Inductance in μh .
 f = Frequency in Mc.

The complete pi-L network with the combined inductances and capacitors is shown in Fig. 4.

A HIGH-POWER AMPLIFIER AND POWER SUPPLY

The grounded-grid linear amplifier and power supply shown in Fig. 1 are designed for the amateur power limit in single-sideband operation. The amplifier uses a 3-1000Z triode to handle a p.e.p. input of 2 kw. on peaks. The amplifier and supply each occupy just over 1 cubic foot and are made from readily available components. The 3-1000Z requires a driver capable of supplying at least 65 watts p.e.p.

Referring to the amplifier circuit diagram in Fig. 2, the grid of the triode is grounded for both r.f. and d.c. The cathode is maintained above r.f. ground by feeding it through a home-made filament choke wound on a ferrite rod. Although the input impedance of the grounded-grid 3-1000Z is close to 50 ohms and would provide a good match for a driver with fixed-impedance output, a pi network input circuit, $C_1C_2L_1$ is used to supply some Q to the circuit, for better linearity. The Q is low, however, and once adjusted an input circuit requires no further attention for operation anywhere within its band. For simplicity in the circuit diagram, only one set of capacitors and inductor is shown in Fig. 2.

The plate tank circuit uses a commercial coil assembly (Air Dux 195-2) that has been rearranged to conserve space and fit better into the compact package. The bandswitch is made from the products of two different companies, ganged together to provide an input-circuit 2-pole switch and a plate-circuit single-pole switch.

Antenna changeover relay, K_1 , is included in the amplifier package. During the "receive" condition a 50,000-ohm resistor in the filament-transformer center tap provides self-bias that practically cuts off the plate current; during "transmit" this resistor is short-circuited by one set of contacts on the relay.

Three meters are used in the amplifier. The grid and plate currents are read separately by a 0-500 milliammeter and a 0-1 ammeter. The third meter is a relative-output indicator metering the r.f. voltage at the output. D.c. for the meter is derived from a germanium-diode rectifier connected to a resistive r.f. voltage divider.

Panel switches and associated neon indicators are provided for control of the filament (and blower) and high-voltage power.

Construction

Two identical pieces of $\frac{3}{8}$ -inch thick aluminum, $11\frac{1}{4}$ inches high and $12\frac{1}{4}$ inches wide, are used for the front panel and the rear plate. These may be cut from $12\frac{1}{4}$ -inch rack-panel material (Bud SFA-1837) if no other source is available. The major chassis that supports the tube socket and the filament transformer is a

 L_1 COIL TABLE

Band	C_1, C_2	L_1
80	1600 pf. (Arco VCM-35B162K)	16 t., closewound
40	910 pf. (Arco VCM-20B911K)	8 t., closewound
20	430 pf. (Arco VCM-20B431K)	6 t., closewound
15	300 pf. (Arco VCM-20B301K)	4 t., closewound
10	220 pf. (Arco VCM-20B221K)	4 t., spaced to fill form.

Capacitors are 100-v. silver mica. Inductors wound with No. 16 Formvar or Nyclad on $\frac{1}{2}$ -inch diam. slug-tuned form (National XR-50).

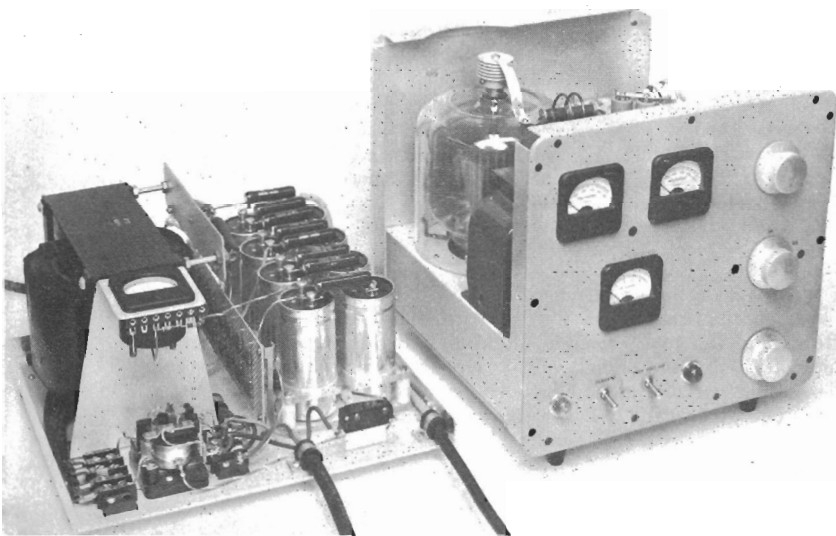
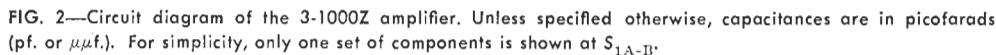


FIG. 1—The 3-1000Z grounded-grid linear is mounted on short legs to allow air to be drawn in at the bottom and blown up past the tube. The meters indicate grid (left) and plate current and relative output (below). Knobs at right (B & W 901) control plate (top), band switching and loading.

Two layers of the heat-radiating plate cap have been removed to allow clearance of the perforated-metal cover.

The companion power supply uses silicon diodes in a voltage-doubling circuit.



L₄—16 turns No. 8, 3¼-inch diam., 4 t.p.i. Tapped
7 turns from L₃ end.

T₁—7½-volt 21-ampere filament transformer (Stancor P-6457). Meters are Simpson Model 127; 1000-pf. 5-kv. capacitors are Centralab 858-S; 1500-pf. feedthrough capacitors are Centralab FT-1500.

Fig. 4). The plate choke, RFC_3 , is mounted on this chassis, with a 1000-pf. 5-kv. ceramic capacitor (Centralab 858-S) near its base; the high-voltage lead is brought from the base of the r.f. choke (and from the capacitor) through the chassis in a ceramic feedthrough insulator. The output-indicator circuitry, consisting of the

5—Amplifier Construction

FIG. 4—A view under the sub chassis of the amplifier. The filament choke can be seen supported off the side wall by ceramic stand-off insulators and plastic cable clamps. A lip on the tube socket (right foreground) has been removed to provide more space and better air flow; the three grid pins of the socket are grounded to the chassis by short straps. Don't try to bend the terminals out of the way before sawing off the lip; remove them entirely.

The resistor mounted on a tie-point strip and visible under the left end of the filament choke is the 50,000-ohm cathode resistor used for stand-by bias; leads from it run in shielded wire over to the chassis wall near the relay. Wires and plug dangling over the side run to the blower (See Fig. 5).

The jack on the rear wall closest to the rear edge of the panel is J_4 , the high-voltage cable jack. A 1000-pf. 5-kv. capacitor is mounted on the chassis just inside this point.

22,000- and 470-ohm resistors, the 1N34A rectifier and the 0.001- μ f. capacitor, is also mounted on the chassis (see Fig. 6). These are mounted on a multiple tie-point strip fastened to the top edge of the chassis near C_4 . The assembly is shielded by a $2\frac{1}{2} \times 2\frac{1}{2} \times 1\frac{1}{2}$ -inch "Minibox" (Bud CU-3000-A).

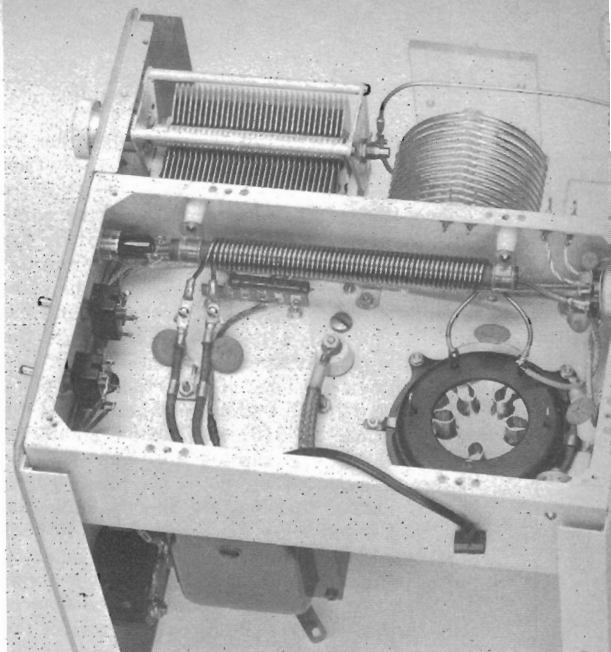
The input circuitry and S_{1A} and S_{1B} are housed in a $4 \times 4 \times 2$ -inch aluminum case (Premier AC-442) held to the main chassis by two $\frac{1}{4}$ -inch panel bearings; the RG-58/U leads to the switches are run through the holes in the bearings. The switch section is mounted on one removable plate of the case; the other plate is not used.

To conserve space and to provide a shaft extension for ganging, switch S_{1C} must be modified slightly. This is done by removing the rear shaft bearing and replacing the two ceramic insulators with shorter ($\frac{1}{2}$ -inch long) ones. If suitable insulators cannot be found in surplus (8-32 tapped holes are required), they can be machined from suitable insulating material. When the rear bearing is replaced, it should first be reversed. The combination of reversing the rear bearing and using a shorter pair of insulators leaves enough shaft extending to take a flexible shaft coupling (Millen 39005). This coupling is connected to a similar coupling on

FIG. 3—Rear view of the 3-1000Z amplifier with the back wall and SK-516 chimney removed. The $4 \times 4 \times 2$ -inch box in the foreground houses the pi-network input circuits and bandswitch sections. S_{1A-B} . The antenna changeover relay is mounted directly below.

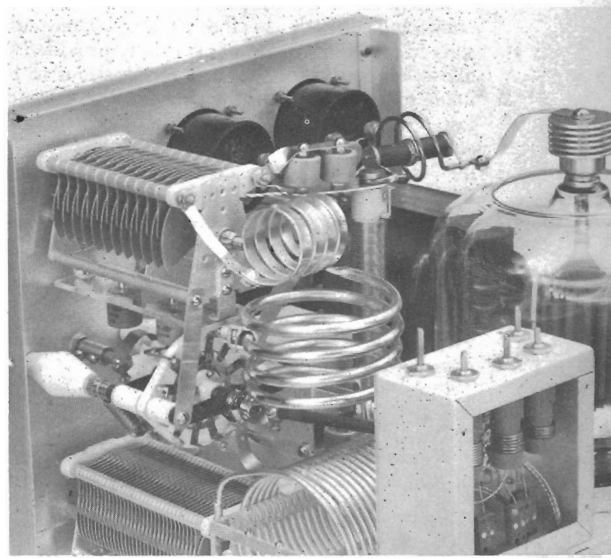
To conserve space and provide a shaft extension for ganging, the plate-circuit switch, S_{1C} , has been modified slightly, as described in the text.

Normally the three meters on the front panel are backed up by a $7 \times 7 \times 2$ -inch aluminum chassis that serves as a shield.



S_{1A-B} through a length of $\frac{1}{4}$ -inch diameter insulating rod. The switch is supported on the panel by four 1-inch high ceramic cone insulators (Johnson 135-501) mounted base-to-base. On the panel, shaft bearings are used for the switch shaft and the two capacitor shafts.

The tank coil assembly is modified by first removing the strap coil and the copper tubing coil from the polystyrene strip that supports them. Then saw a 3-inch long strip from one end and mount it at right angles to the original strip with cement or brass screws (see Fig. 6). Coil L_4 , the wire coil, is supported by the polystyrene strip, which rests on the bottom plate at the outside and on the basic chassis on the inside. The inside end of L_4 is bent up and a loop formed in the end. Coil L_3 is bolted to this loop with a brass 8-32 machine screw, and the tap running to the 20-meter pin on the switch is taken off at the same junction. All coil taps were made of $\frac{1}{8}$ -inch wide strips cut from copper flashing. The coil L_2 is supported



at one end by an end of L_2 and at the other by a copper strap fastened to the stator of C_3 (see Fig. 6).

Two 1000-pf. 5-kv. capacitors and the parasitic suppressor, RFC_3R_1 , are supported by the top of RFC_3 , and a flexible strap runs from the other end of the parasitic suppressor to the plate cap. To avoid contact between the cap and the amplifier cover, two layers must be cut off the top of the plate connector (Eimac HR-8).

The chimney (Eimac SK-516) is held in place around the 3-1000Z by four metal clips, and the socket (Eimac SK-510) is modified slightly as mentioned in the caption for Fig. 4. The blower is mounted on the $12 \times 11\frac{1}{4} \times \frac{1}{8}$ -inch bottom plate so that it is not directly under the tube socket but near the front panel. Rubber (they could be turned wooden) feet attached to the bottom plate support the amplifier above the operating table and allow the free flow of air into the blower. The a.c. connector for the blower (see Figs. 3 and 4) is made from the internal parts of two microphone connectors (Amphenol 80-MC2M and 80-MC2F).

To conserve space, the filament transformer T_1 must be modified so that the leads come out the bottom. This is done by removing the end bells, blocking the original holes with paper and drilling new holes for the leads.

There is a little trick to winding the filament choke, RFC_1 , primarily because the wire is so heavy that it cannot be wound directly on the ferrite rod without springing out. To overcome this, the dual winding of the choke is wound first on a length of $\frac{1}{16}$ -inch wooden dowel. When it is released it will spring out slightly, enough to permit it to be slipped off the dowel and on to the ferrite core. One-half inch nylon cable clamps mounted on 1-inch standoff insulators hold the core in place on the inside wall of the chassis (see Fig. 4). Formvar or Nylclad wire is recommended for the choke because with it there is very little chance that the insulation will be chipped off as the core is inserted in the coils.

It will be noted that the three grid leads are connected directly to the chassis. There are slots in the SK-510 socket especially provided to allow low-inductance ground terminations to be made to each of the grid terminals. The grounding straps are slipped through the slots and soldered to the socket pins.

The $7 \times 7 \times 2$ -inch chassis that shields the three meters is held to the panel by a single screw that threads into a ceramic cone insulator serving as a stud. The stud can be made of metal if suitable taps and drills are available. To clear the shielded wires running up to the meters from the hole in the main chassis, a suitable slot is cut on one side of the shield chassis.

Power Supply

A power supply delivering 2500 to 3000 volts at 400 to 350 ma. will be suitable for use with

the amplifier. The power supply shown in Figs. 1 and 8 is built with sixteen silicon diodes costing 84 cents each (a total of \$13.50), although rectifier units having a p.i.v. rating of 4000 (and a price of \$18 each) are available.

Referring to the circuit diagram of the supply, Fig. 9, a transformer with a dual primary is used, to permit operation from either a 115- or a 230-volt line. The higher voltage is recommended. No fuses are shown; it is expected that the supply will be protected by the fuses (or circuit breakers) in the wall outlet box.

The filter capacitors are called "computer grade" capacitors; the 25K resistors across them

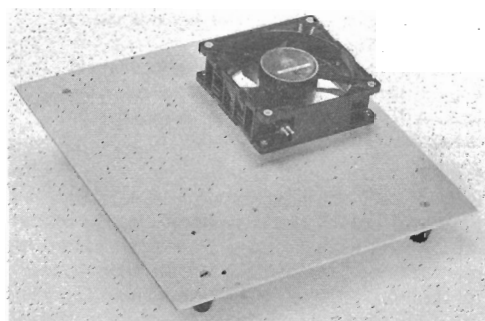


FIG. 5—Blower is mounted on bottom plate of amplifier near the front panel (not directly under the tube). A.c. power connector for blower is made from microphone-connector fittings.

serve both as the bleeder resistor and the equalizing resistors. In operation, the idling current of the amplifier (180 ma.) further bleeds the supply. The 0-5000 voltmeter is included to comply with the FCC regulations. It is a good idea to get into the habit of watching the voltage decay when the power supply is turned off; in this way you are less likely to get mixed up with a residual charge in the capacitors. An interlock switch in series with the relay makes it necessary to replace the cover before turning on the supply.

The 10-ohm resistor between the negative terminal and chassis allows plate-current metering in the negative lead with no difference in potential between power-supply and amplifier chassis.

The power supply construction is not critical, and the main considerations are adequate insulation and safety precautions. The string of silicon diodes and their associated capacitors and resistors are mounted on a $3 \times 9\frac{1}{4}$ -inch strip of prepunched terminal board (Vector 85G24EP), with push-in terminals (Vector T28) serving as tie points. The rectifiers are mounted on one side of the board, the resistors and capacitors on the other. The strip is mounted on the $12 \times 13 \times \frac{1}{8}$ -inch aluminum base plate with a pair of panel brackets (Raytheon MB-128).

The pair of 50-ohm resistors is mounted on a $7\frac{1}{2} \times 1\frac{1}{4}$ -inch strip of pre-punched terminal board, supported by two $\frac{1}{4}$ -20 bolts, 5 inches

long, that replace two of the original transformerbolts. This strip also serves as a stop to prevent the cover and the resistors coming in contact.

The bank of eight 240- μ f. capacitors is insulated from the base plate by a sheet of $\frac{3}{4} \times 9 \times \frac{1}{8}$ -inch clear plastic (Lucite or Plexiglas). A similar sheet with clearance holes is mounted higher and holds the capacitors in place. The 25K bleeder resistors mount on the capacitor terminals.

The high-voltage cable running to the amplifier is a length of RG-8/U terminated in a high-voltage coaxial plug (UG-59B/U). At the power supply end, the braid is peeled back for about a foot on the insulating material, to provide a suitably long leakage path. Disregard of this small point may result in voltage breakdown along the surface of the insulating material. The shield braid is connected to the base plate, which serves as the chassis ground. Wires to the a.c. line should be No. 14 or heavier (a cable marked "14-3 Type SJ 300 V" was used in this unit), and No. 16 wire will suffice for the control wiring.

If desired, a precision resistor can be used for R_p , the voltmeter multiplier. However, selected standard 20-percent resistors will serve as well.

Safety Precautions

A 3000-volt power supply with a 30- μ f. filter capacitor is a lethal device. There is no such thing as a "slight electrical shock" from a power supply like this one. Make absolutely

certain that the voltmeter indication has coasted down to zero before removing the protective cover or touching anything remotely connected to the high-voltage lead. Even then it is a good idea to use a "shorting stick" across the output as a double check.

Adjustment of the Amplifier

An amplifier of this quality and power level deserves the best of treatment, and to that end it is recommended that the operator familiarize himself with its operation by using a dummy load, an oscilloscope and some method of "pulsing" the drive (see Goodman, "Linear Amplifiers and Power Ratings," *QST*, August, 1957). This will enable the operator to work the amplifier at its maximum legal capability with a minimum of spurious radiation.

Lacking the equipment mentioned above, it is possible to approach proper operating conditions by the following rules of thumb. They are intended, however, to serve only as rough guides.

With a sideband exciter set for c.w. operation, feed its output to the amplifier through a length of RG-58/U or RG-8/U. If an s.w.r. indicator (for 50-ohm cable) is available, insert it in the line and switch it to read reflected power. With the filament of the amplifier turned on, adjust L_1 (for the band in use) for a peak in grid current. Do this in turn for each of the bands. The adjustment is a relatively broad one. It may be found that a little reflected power is indicated, but that is not important at this time.

FIG. 6—Another view of the grounded-grid amplifier, showing the output voltmeter (shield cover removed) components mounted on a multiple tie point strip. The cone insulator between the meters receives the screw that holds down the meter shield (chassis).

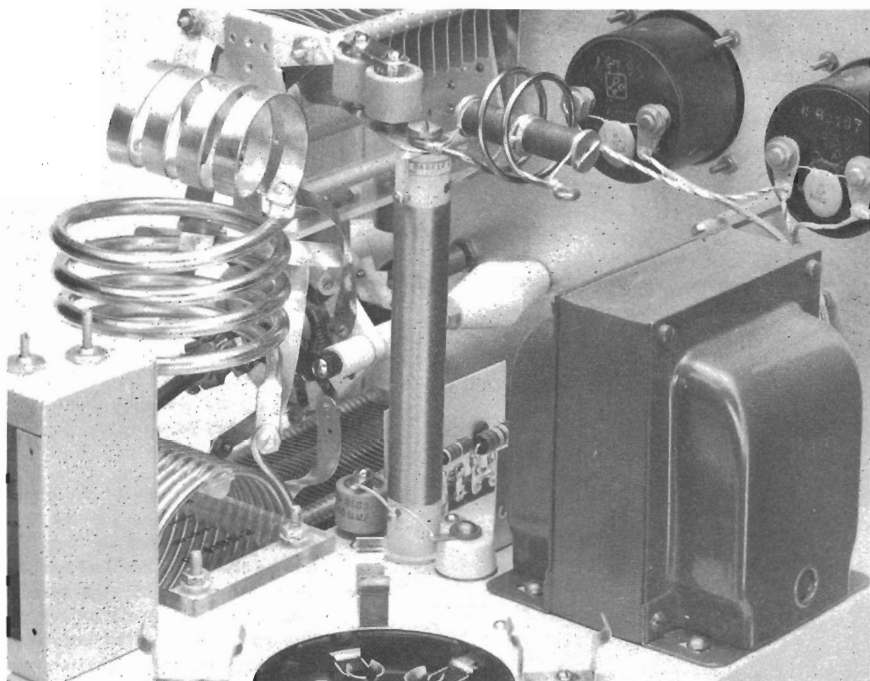


FIG. 7—Design changes for high/low power operation. Unless listed below, component designations and specifications are same as in Fig. 2. For simplification, 1500-pf. ceramic feedthrough capacitors in relay coil and cathode resistor leads are not shown.

(A) Circuit changes when amplifier is used with separate receiver and exciter.

(B) Interconnections when used with separate receiver and exciter.

(C) Circuit changes when amplifier is used with transceiver.

(D) Interconnections when amplifier is used with transceiver.

J_6 —Coaxial receptacle, SO-239.

K_2 —Antenna changeover relay with auxiliary contact, 115-v.a.c. coil, (Advance AH/2C1C/115-VA).

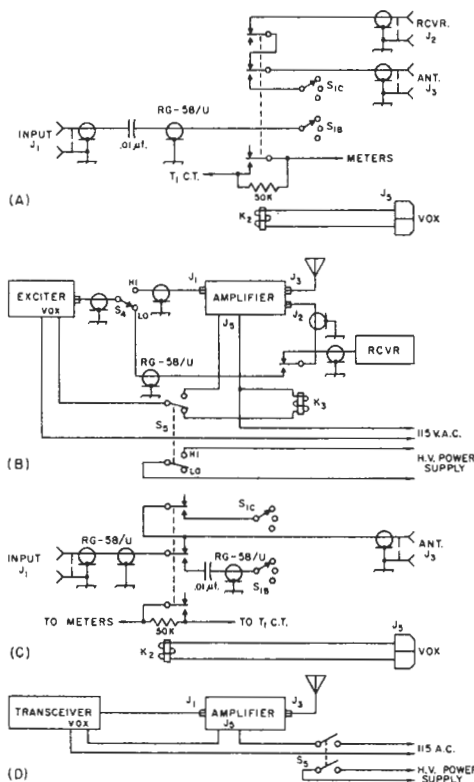
K_3 —Coaxial antenna changeover relay, 115 v.a.c. coil.

S_4 —S.p.d.t. coaxial switch. See text.

S_5 —D.p.s.t. toggle. See text.

Plate voltage can now be applied, but it is recommended that early tests be carried out at half operating voltage, until it has been established that it is possible to tune to the various bands. Never apply plate voltage to the amplifier without a load (dummy or antenna) being connected, because there is danger of burning out CR_1 under these circumstances. Having established that the circuits can be tuned, the amplifier can be tested at full voltage. The loading and excitation (single tone, same as steady carrier or c.w.) should be adjusted to give the readings shown below, with the understanding that these are only general guides and are not strict limits. Notice that these conditions represent tuning to a steady 1 kilowatt input, the only possible legal procedure (without pulsing, which is illegal except into a dummy load). When a set of these conditions has been met, adjust the output of the exciter to drive the amplifier just to an indicated 1 kilowatt plate input on peaks.

Plate Voltage	2500	3000
No-Signal Plate Current	160 ma.	180 ma.
Single-Tone Grid Current	100 ma.	75 ma.



Single-Tone
Plate Current

400 ma. 330 ma.

As a final touch, adjust the input circuits for minimum reflected power.

Although the amplifier should have no v.h.f. parasitic with the suppressor as shown (RFC_2R_1 in Fig. 2), the amplifier should be tested for one. Disconnect the exciter, connect a dummy load to the output, switch to 21 or 28 Mc. and apply filament and then plate power. With one hand on the plate power supply switch, swing the plate capacitor, C_3 , through its range, starting at maximum capacitance. If a parasitic is possible, it will probably show up as C_3 approaches minimum capacitance; it will be indicated by a sudden increase in plate current and the appearance of grid current. If a parasitic does appear, it will be necessary to increase the inductance of RFC_2 (after turning off the plate power!) by pushing the turns together or adding another turn.

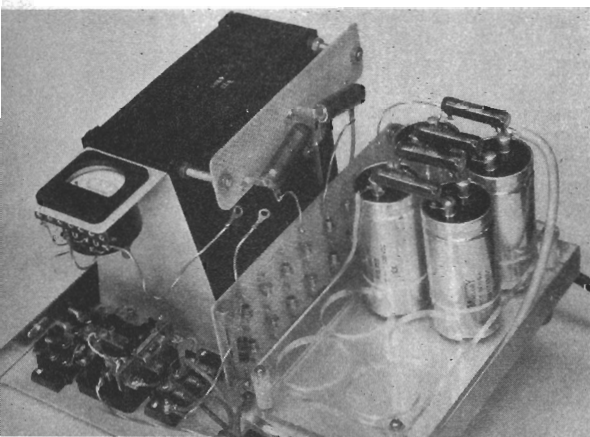


FIG. 8—In this view of the power supply, four filter capacitors have been removed to show how the silicon diodes are mounted on one side of the terminal board; equalizing resistors and capacitors are mounted on the other side. The meter mounting bracket is held to the base plate by two of the bolts that run through the feet of the transformer. Small switch in the foreground is the interlock; control relay is mounted on base plate to left of terminal block.

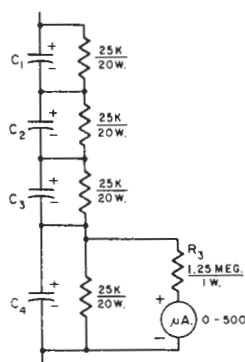
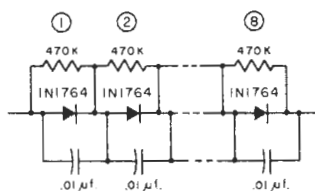
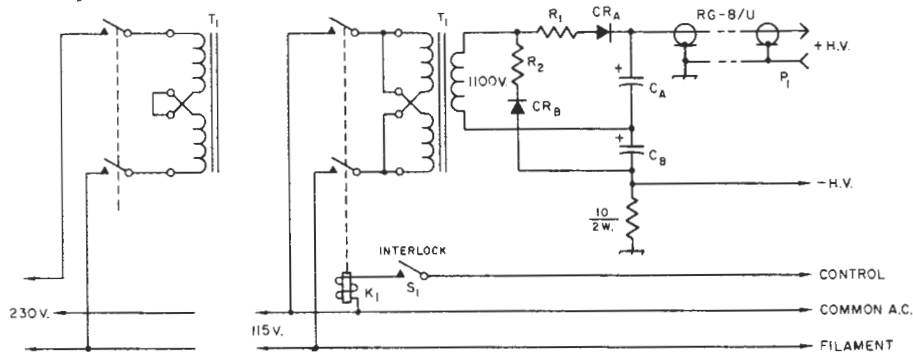


FIG. 9—Schematic diagram of the 3000-volt power supply.

C₁-C₄—240μf. 450-volt electrolytic (Mallory CG241T450D1).

K₁—D.p.s.t. relay, 25-ampere contacts (Potter & Brumfield PR7AY, 115-v.a.c. coil).

P₄—Coaxial plug, UG-59B/U (Amphenol 82-804).

R₁, R₂—50-ohm 25-watt wireground (Ohmite 0200D).

R₃—Selected 0.47- and 0.68-megohm, 1/2 watt, in series.

S₁—S.p.s.t. miniature switch (Acro BRD2-5L).

T₁—1100-v. 1200 v.a. transformer, dual primary (BTC-4905B, Berkshire Transformer Corp., Kent, Conn.)

25K, 20-watt resistors are Ohmite Brown Devil 1845, 470K resistors are 1/2-watt, 0.01- μ f. capacitors are 1000-volt disk ceramic.

Variations in Design

Depending upon the type of operation and the exciter in use, it may be desirable to modify the amplifier design. For example, using an amplifier of this power in an area or on a band where local contacts are often made, it is selfish and uneconomical to run at full power all of the time. By using a different changeover relay than that specified in Fig. 2, it is relatively easy to provide for full-power or exciter-only operation at the flick of a switch.

Referring to Fig. 7, the diagram at (A) shows the revised circuit when a separate exciter and receiver are used. The change involves the addition of a coaxial antenna-changeover relay and, in the amplifier, the use of a relay with one more pole. If this system is used, it should be connected as shown at (B). The coaxial switch, S_4 , need be only a rotary ceramic switch mounted in a small aluminum case

with three SO-239 coaxial receptacles mounted on the side (or sides). The switch S_n can be a d.p.d.t. toggle or, for more convenience, a two-pole rotary switch ganged to S_1 . The two-wire circuit from the VOX (voice-operated) circuit in the exciter should be one that closes on "transmit"

If the exciter is a transceiver, the coaxial changeover relay will be already built-in, and the amplifier modification simplify to those shown in (C), the substitution of a three-pole relay. In use, the connections will be as those shown in (D), and S_5 can become the S_2 (but a d.p.s.t.) of Fig. 2.

In either case, the fact that the input impedance of the amplifier is close to 50 ohms (the required load for the exciter or transceiver and the assumed impedance presented by the antenna feed line) makes for great convenience, since no retuning is required.

"Here is some down-to-earth talk about linear-amplifier meter readings that is "must" reading for all s.s.b. enthusiasts. The case is presented in simple, nontechnical language and with illustrations that clearly demonstrate the basic principles. Although discussed in terms of the plate meter, it also applies to the rectifier-type output meters with which currently-manufactured linear amplifiers are equipped.

Interpreting the Linear-Amplifier Meter Reading

HOWARD F. WRIGHT, JR., W1PNB

In the days of regular a.m. there wasn't much concern about meters. The d.c. plate meter in the final gave most of the answers without complaint. Watch the meter. Tune up the rig. Figure the power input—no strain, no pain! What could be neater? To say that this no longer holds true with a linear amplifier in suppressed-carrier service would be quite an understatement. Strong men have wept bitter tears and spent sleepless nights because of the behavior (or misbehavior) of their linear's plate meter. Why? Simply because most of us seem to find it extremely difficult to modify our nearly complete, all-abiding faith in the value of the plate meter in indicating final amplifier performance.

Let's get down to brass tacks. The attitude of an amateur toward the plate meter of his linear final, under voice conditions, is of great importance. It could, from a broad point of view, mean complete success or partial failure of amateur narrow-band communication techniques.

Why does the d.c. plate milliammeter fall down so badly in indicating the performance of amplifiers in s.s.b. voice service? It's because the meter is no longer able to settle at a steady value as it did in the amplification of unvarying carrier signals. The voice modulation consists of sporadic bursts of energy. They say, "The hand is faster than the eye." If so, the voice is certainly faster than the meter. The meter just doesn't move rapidly enough. It starts to follow the first voice impulse up, but moves so slowly that it meets the signal coming down. Then it tries to follow downward. In this it also fails. If a constant sound is used instead of words, the meter stabilizes at an "average" value. When the signal varies with the syllables of speech, the meter bobbles around. The amount of movement depends upon many factors. Meters can have differing time constants (speeds of response). Different voices

contain varying amounts of "average" power. The amount of swing depends, to some degree, upon the class of amplifier operation: AB₁, AB₂ or B.

Distortion

A "linear" must amplify the signal from its exciter without changing the waveshape of the modulation envelope of the original signal. Any change in this waveshape is distortion. Distortion means that new signals are generated. These new signals result in splatter.

Every linear amplifier has an amplitude point at which it will produce no further undistorted output. Although the driving signal continues to increase, the output no longer increases in exact proportion. While any change of the signal wave-shape at levels other than this maximum value also causes some trouble, it is imperative that the "limiting" ("flattening") point not be exceeded.

The plate meter is basically incapable of indicating the peaks of a voice signal. Any relationship between voice excursions of the plate meter, as it measures d.c. power input, and undistorted unflattened amplifier output is apt to be purely coincidental.

If the meter is such a poor performer, why do we continue to use it? Simply because, when properly interpreted, the meter is still a valuable gadget. It just needs a bit of understanding.

Meter vs. Scope

Articles concerning linear-amplifier adjustment make adequately clear the fact that the oscilloscope is the best tool for indicating performance. Whether the use of this valuable instrument is any more vital to the adjustment of a s.s.b. transmitter than it is to making a conventional a.m. phone station work properly might be a matter for debate. While a gratifying number of amateurs are now using scopes, it would be unrealistic to think that all s.s.b.

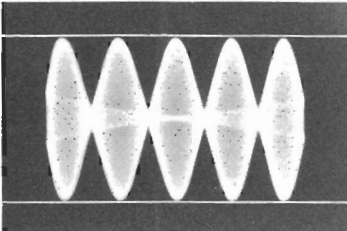
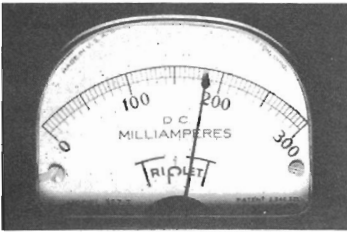


FIG. 1

Each of the accompanying photographs consists of a comparison between the plate milliammeter reading and the output waveshape of a linear amplifier. While each picture shows a different type of signal input, actual amplifier adjustment remains unchanged in all cases. The purpose of the comparison is to demonstrate the action of the average-reading meter as compared to the instantaneous-reading scope while indicating signals of varying waveshapes.

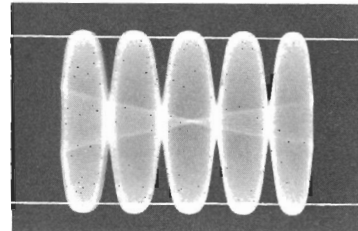
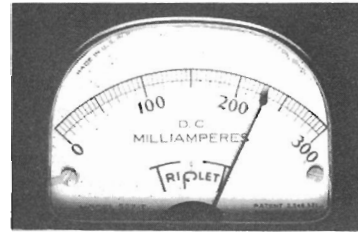


FIG. 2

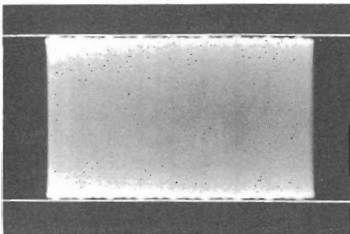
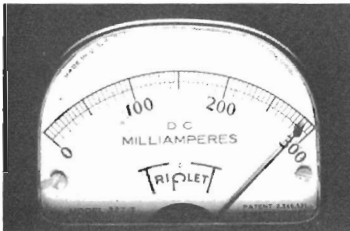


FIG. 3

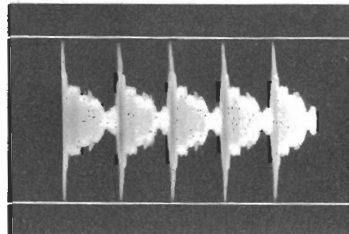
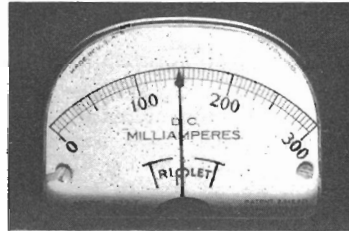


FIG. 4

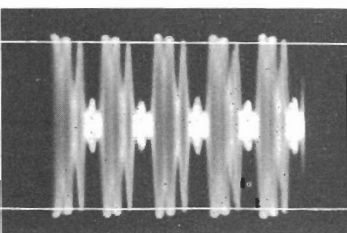
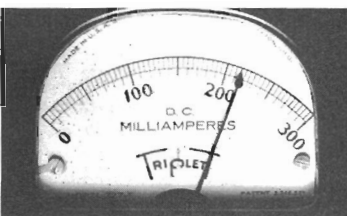


FIG. 5

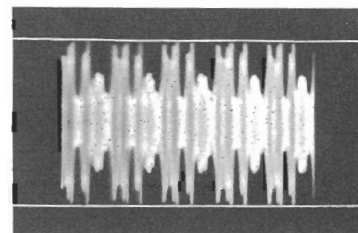
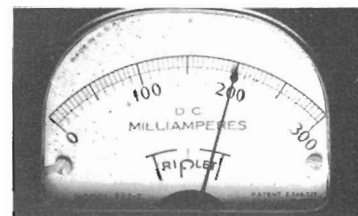


FIG. 6

stations will ever be monitored at all times by operators using such instruments. In fact, an operator who understands what his plate meter means, in conjunction with some form of output indication, can do quite well with no scope.

Each of the accompanying photographs shows a different condition of amplifier operation or type of signal input as seen on both a scope and the d.c. plate milliammeter. The purpose is to demonstrate the action of the average-reading meter as compared to the instantaneous-reading scope. Both continuous signals and voice are used.

Fig. 1 shows a two-tone test signal. This type of signal is used to determine linearity and lack of distortion in all parts of the waveshape. At this time, notice only one thing. The top of the pattern remains unflattened and fills up the space between the limit lines. No splatter caused by distortion of peaks occurs as long as we do not try to exceed the limits while using this set of amplifier conditions. This holds true in all the following pictures regardless of the type of signal input.

Fig. 2 shows the amplifier being driven into distortion on the two-tone signal. The peak linearity capability has been exceeded. Note the flattening of the peaks. Now notice the meter reading as compared to Fig. 1. The meter shows what we all like to see—more power—but the scope indicates that it is distorted power.

Fig. 3 shows the result of introducing a single audio tone into the speech amplifier. It may not look like sine-wave audio, but this is because we no longer have an audio signal. The s.s.b. exciter converted the single tone audio into single frequency r.f.—practically the same as an unmodulated carrier. Look at the plate meter! On steady signal the amplifier has no time to rest. It works regularly—not in spurts as on voice. The meter has a chance to indicate full maximum signal input.

Fig. 4 shows an actual voice waveshape. The sound used was a sustained "O-o-o-h-h-h." Notice that the "peaks" just reach the limit lines. Look at the meter. Oh, how sad! That's the current we must use to calculate power input!

Fig. 5 is the condition where the audio gain is increased to "correct" the low meter reading. Yes, the meter reads more, but take a look at the scope. Those peaks are really flattening. Splatter! Zounds! No escape! A clean signal means a lower meter reading, while greater deflection causes trouble.

Fig. 6 shows the same voice sound, purposely altered by audio compression, to increase the "average-to-peak" ratio. Notice that the meter again shows an increase over the conditions of Fig. 4. This is also due to the flattening of the waveshape. Distortion of the audio is present, but no splatter results if the new frequencies are carefully filtered out early in the exciter. New frequencies resulting from

r.f. linear amplifier flattening cannot be filtered out. They will be transmitted to plague adjoining channels.

Study of the photographs reveals that there is, as previously mentioned, lack of correlation between d.c. meter readings and the type and quality of actual amplifier output.

The situation looks rather gloomy, doesn't it? Is it possible for an average amateur to operate a linear amplifier properly without access to laboratory measuring techniques? Well, the best answer I can give is that hundreds are doing it every day. Perhaps the meters don't give all the necessary indication, but never underestimate the flexibility of an amateur. The trial-and-error system can do wonders.

Splatter

A chain of two or more linear amplifiers, upon construction, is hardly ever able to develop maximum rated output without considerable adjustment. Luckily, s.s.b. transmitting and receiving techniques have the valuable property of making nonlinear amplifier distortion and splatter stick out like a sore thumb. While the same amount of distortion would be partly hidden by the voice sidebands of a double sideband signal and be somewhat obscured by lack of selectivity in an ordinary receiver, such is not the case on s.s.b. There is no such thing as distortion splatter which appears on only one side of the carrier frequency.

A s.s.b. receiver has an opportunity to view, generally unhindered by readable signal from the s.s.b. exciter, the amplitude, nature and frequency spread of nonlinear amplifier splatter appearing on the unwanted sideband. This situation makes possible accurate and worthwhile on the air reports of amplifier performance. In cases of "peak limiting" distortion, one can simply turn down the gain until the person at the receiving end reports a clean "unwanted" sideband. Changes can then be made to try and allow more power without degrading the signal.

Adjustment Without a Scope

This isn't basically an article on linear-amplifier adjustment, but I am going to give an example to demonstrate the proper use of d.c. meters when nothing better is available. The procedure falls into the "cheap and dirty, but rather effective" class.

Although I have mentioned only the "final amplifier," flattening and distortion may also occur in any driver stage. I do not think it necessary, for our purposes, to stress meter readings and waveshapes for other stages. The indications occurring at the output of the final accurately reflect the condition of earlier stages. Of course, in actual practice it is necessary to locate and work on the weak link.

Let's say that I have an amplifier whose specs call for a plate voltage of 1500 and

maximum signal (peak-envelope) current of 300 ma. First, I insert some carrier from the exciter. I tune the grid and plate circuits to resonance as indicated by an output indicator. (Any type of output indicator connected to the feed line will do.) Next, I adjust the coupling of the feed line to the final. The coupling is set for maximum power output at a given plate current at resonance. If the coupling is increased, the d.c. plate current goes up, but the output either remains constant or decreases. This point of maximum output for a given amount of input comes close to the magic point of proper adjustment for all linear amplifiers. The old method of loading by reference to "dip at resonance" is not recommended. Once we have reached the suggested adjustment, we have had it as far as coupling goes. If the d.c. current is less than 300 ma. (for the amplifier under discussion), we simply don't have enough linear drive.

Now, regardless of the power I believe my amplifier should handle, I make a crude check to determine at what point it actually flattens. I vary the amount of carrier insertion, watching mainly the output indication, and note the point at which increasing the carrier no longer results in a rapid increase in output. I now observe the plate meter reading, hoping in this case that it is up to 300 ma. I use this figure to multiply by plate voltage. This is roughly

my maximum-signal linear power available. Suppose my linear drive available limits my actual plate current to 250 ma. My maximum signal power is then only 375 watts instead of the possible 450.

Next, I remove the carrier. I change to voice. This is the point where experience in using a scope counts. I know that my voice, using the average plate meter, will deflect the meter only about half as far, for the same maximum signal power, as the carrier did. Thus, since my steady signal current was 250 ma., I wouldn't expect much over 125 ma. on normal talking. My "meter peak, on voice" or legal power, is only 187 watts ($125 \text{ ma.} \times 1500 \text{ volts}$). My maximum signal or peak envelope power is 375 watts ($250 \text{ ma.} \times 1500 \text{ volts}$).

Actually, your voice might be able to swing the meter somewhat more than mine for the same amount of maximum signal power. If I were using a Class AB amplifier I would expect the meter to swing somewhat higher on voice in relation to its reading on steady signal.

Don't think that I am recommending the above procedure as a replacement for legitimate "two-tone" adjustment with a scope. However, if cautiously used, the procedure is guaranteed to do one thing—to produce a signal vastly more neighborly than one generated in an attempt to wrap the pointer around the pin.

» The first three articles in this chapter deal with a subject that no sideband operator can afford to ignore—testing and adjusting the s.s.b. transmitter for the cleanest possible output.

A Pulsed Two-Tone Test Oscillator

Transistors in a Test Generator for Linear Amplifiers

WALTER LANGE, W1YDS

It has been pointed out that the pulsed two-tone testing of a linear amplifier allows any amplifier to "loaf" along at low average input while being driven to maximum p.e.p. input.¹ The two-tone test pattern is most easily obtained as the double-sideband output from a balanced modulator and a single audio tone. However, nowadays most sideband rigs use a filter to generate the single-sideband signal, and so two similar-amplitude audio tones must be fed to the s.s.b. generator if a two-tone test pattern is to be obtained.

Used with an oscilloscope and a dummy load, the generator described here can be used to make most of the necessary checks on a sideband transmitter or amplifier. With a single audio tone the sideband and carrier suppression can be checked, while the two-tone test, steady or pulsed, gives a visual check on the flat-topping level and the linearity of an amplifier.

To be useful, a two-tone test generator must produce signals of low harmonic content. If it doesn't, the harmonics make a 3-, 4- or 5-tone generator out of it, depending upon the number of harmonics that fall within the pass band of the filter. A number of different transistor-oscillator circuits were tried, but we ended up with Colpitts LC oscillators as giving the best waveforms.

The final circuit is shown in Fig. 1. Each audio oscillator is followed by an emitter follower, to minimize reaction. Control R_1 is included to permit adjusting the 1800-cycle amplitude to match that of the 800-cycle signal. The matched signals are fed to Q_3 , where the setting of an output-level control, R_2 , determines the signal that reaches the output jack or, when pulsing is used, the gate transistor, Q_8 .

An astable (free-running) multivibrator circuit furnishes the gating pulse. The output approximates a square wave, and the repetition rate can be varied between about 50 and 170 p.p.s. The duty cycle or on-off ratio can be controlled by the setting of R_3 ; this also has an effect on the pulse rate, and R_4 is included as a vernier control of the pulse rate and duty cycle, as explained later.

Q_8 is a forward-biased shunt gate. The two-tone signal is developed across R_7 and R_8 in series, and the fraction across R_7 is coupled to the output jacks. However, with no multivibra-

tor pulse arriving at the base of Q_8 , the gate transistor conducts heavily and effectively short-circuits R_7 . A positive pulse from the multivibrator cuts off Q_8 and the two-tone signal appears in the output.

Construction

The two-tone generator is housed in a $5 \times 6 \times 9$ -inch utility cabinet. Battery holders and the two inductors, L_1 and L_2 , are mounted on the base of the cabinet, and the jacks and mode switch S_2 are mounted on the top. The frequency selector switch and the four potentiometers are mounted on the front panel. All of the remaining components are mounted on a $7\frac{1}{2} \times 4\frac{1}{2}$ -inch sheet of prepunched terminal board (Vector 85G24EP) with push-in terminals (Vector T-28). Six 1-inch 6-32 threaded spacers hold the terminal board behind the front panel. The parts arrangement on the board is not critical, but for simplicity in this version the parts layout more or less resembles the circuit diagram.

Testing

When the unit has been constructed and the wiring checked, install the four flashlight batteries in their holders. Connect a length of shielded cable between either J_2 or J_3 and the vertical input terminals of an oscilloscope. With S_2 in the unpulsed position, turn S_1 to the "800" point. Adjust R_2 for maximum output (at least 0.15 volt peak-to-peak). Switch S_1 to "1800" and advance R_1 for maximum gain (about 0.3 volt peak-to-peak). In both cases the scope should show good quality sine waves of negligible distortion.

Adjust the 1800-cycle signal level to the same amplitude as that of the 800-cycle signal. Switching S_1 to "2 tone" should produce a complex waveform of about 0.3 volt peak-to-peak. Turn S_2 to the pulsed position. R_3 and R_4 will vary the "on" time of the pulse from about 40 to 60 per cent.

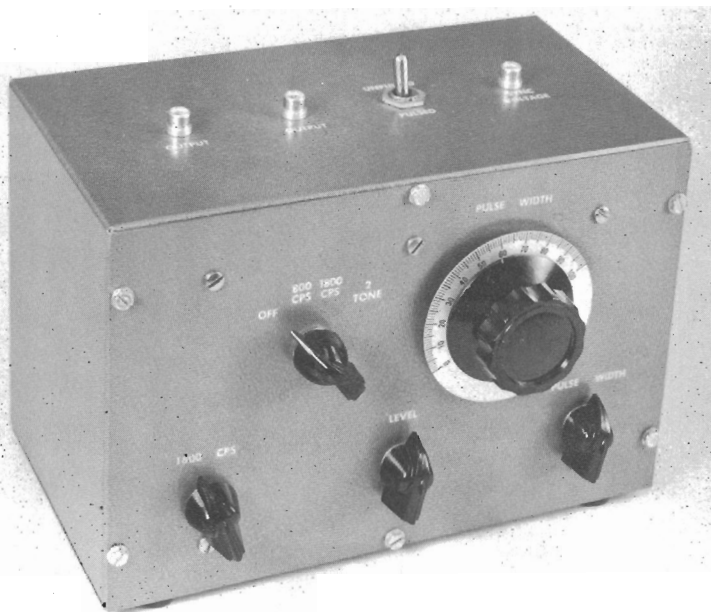
Using the Test Generator

Connect a length of shielded cable between the test oscillator and a sideband exciter. Attach a "T" connector to the transmitter and run a coaxial cable between the dummy load or the Telematch.² Connect a suitable tuned circuit to the vertical deflection plates of the scope, and

¹ From September, 1965, QST.

² Goodman, "Linear Amplifiers and Power Ratings", QST, August, 1957.

² Goodman and Lange, "The Telematch," QST, February, 1965.



Front view of the two-tone test oscillator. The generator is completely self-contained. Battery drain is only 4 ma.

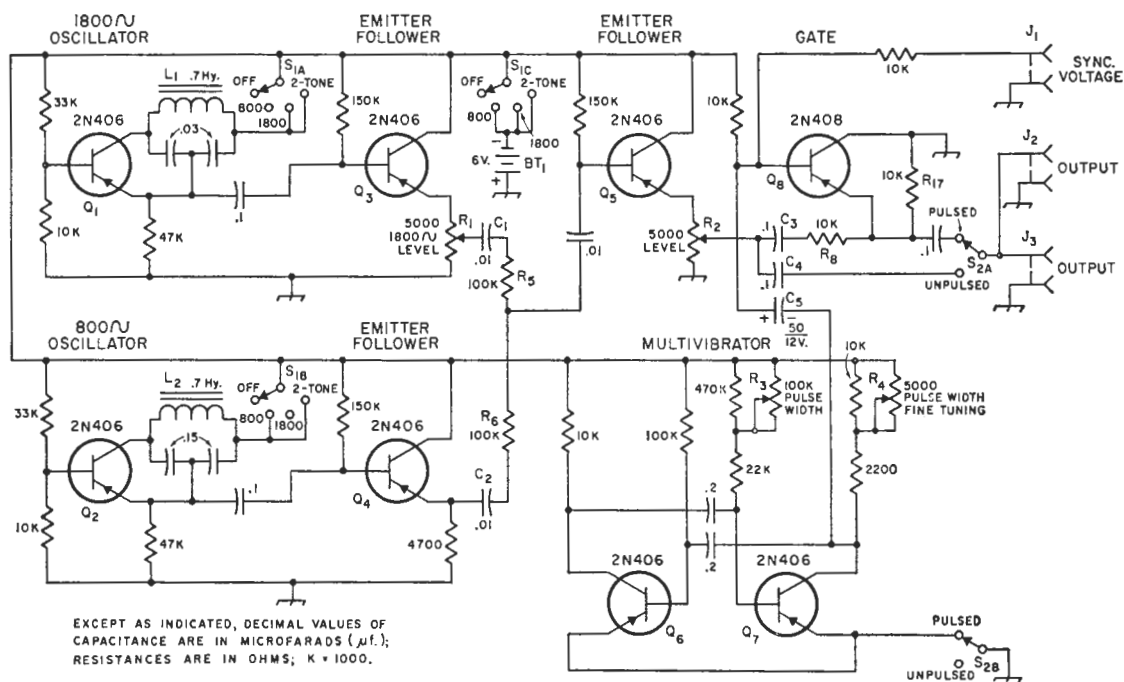


FIG. 1—Circuit diagram of pulsed two-tone test oscillator. All capacitors are tubular paper except C_3 , which is a subminiature electrolytic. Resistors are $\frac{1}{2}$ watt, ± 10 per cent.

BT₁—Four 1.5-volt flashlight batteries (size D) in series.

J₁, J₂, J₃—Phono jacks.

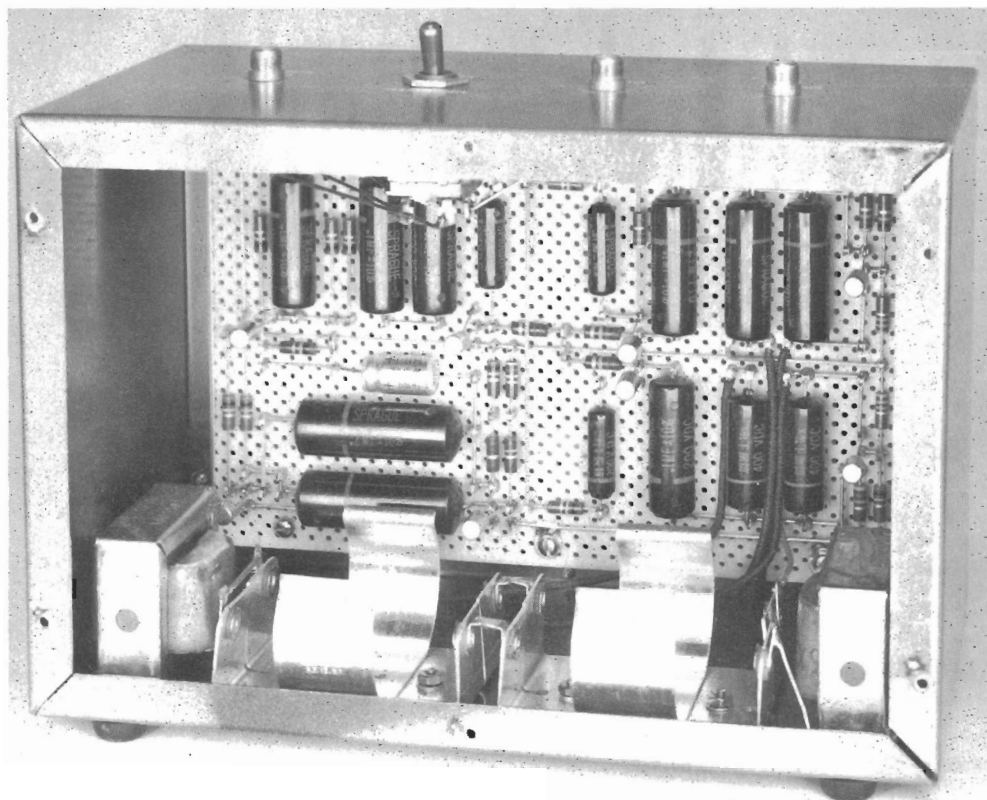
L₁, L₂—0.7-henry 290-ma. filter choke (Knight 64 G 703).

R₁, R₂, R₄—5000-ohm control, linear taper.

R₃—100,000-ohm control, linear taper.

S₁—Rotary switch, 3-pole 4-position, shorting type (Mallory 3134J).

S₂—D.p.d.t. toggle switch.



Interior view of the test oscillator. Parts are arranged on the terminal board in a manner similar to the schematic diagram. The oscillator circuitry is on the right with the three emitter followers towards the center. Multivibrator components are located on the lower left corner of the board with the gate parts just above them. The two double battery holders are Keystone type 176.

link couple this circuit to the open connection on the "T" connector.

Switch the test generator to the 800-cycle unpulsed position. Adjust R_1 and/or the exciter audio gain control for a good size scope pattern, being careful not to overdrive the amplifiers or exceed their ratings. Peak the scope tuned circuit. The desired pattern is a simple rectangle, whose top and bottom edges are as straight as possible (without saturation in an amplifier). Any ripple on the top or bottom indicates insufficient carrier and/or sideband suppression and should be minimized before proceeding. (If the 800-cycle signal has too high second- and third-harmonic components, or if distortion is taking place in the audio section of the sideband exciter, ripple will also appear on the pattern.)

Turn S_1 to "1800" and recheck for ripple. The top and bottom edges of the pattern should be straight lines.

Once the suppression has been checked, a two-tone pattern may be used for checking the linearity of the system. Adjust the 1800-cycle signal level with R_1 to the same amplitude as the 800-cycle tone. If the amplifier is being run

conservatively, it can be operated with the steady two-tone test signal right up to the flat-topping level. As explained in the *Radio Amateur's Handbook*, the things to avoid are flat-topping and non-linear crossovers. Another pattern that is somewhat common but not discussed in the publications mentioned above is something that, at relatively slow sweep speeds, gives the appearance of low-frequency ripple riding across the top and bottom of the two-tone pattern. The several possible causes include 60- or 120-cycle hum modulation, non-linearity in the audio stages, insufficient carrier suppression and insufficient sideband suppression. In other words, the spectrum of the signal includes more than the two frequency components that it should have.

Amplifiers that would be crowded a bit by the steady two-tone test should be checked by the pulse method. Switch the oscilloscope to external sync and connect a length of shielded cable between J_1 and the external sync terminal of the scope. Synchronize the scope sweep to the pulse repetition frequency so that one or two pulses are displayed.

Testing a Sideband Transmitter

DOUGLAS A. BLAKESLEE, W1KLK

Many amateurs are still afraid to "lift the lid" on their sideband transmitters. Granted, a sideband rig is the most complex piece of equipment in most ham shacks, but that is no reason why a hands-off attitude should be so dominant. A large number of amateurs just do not take the time to become familiar with sideband techniques. With a small investment in test equipment and a little practice, any amateur can keep his transmitter in top condition.

Even if your transmitter never blows up, component aging, tube changes and the difference in temperature in your shack between Winter and Summer will affect the performance of balanced modulators and phase-shift networks. These circuits *will* require readjustment from time to time. And, too, even a perfect transmitter can be operated in such a way that it sounds horrible. The damage has been done if you wait until others on the band (or the FCC) inform you that something is wrong with your transmitter.

Test Equipment

To observe the rapidly-changing levels in a sideband transmitter an oscilloscope is absolutely necessary. No meter can keep up with the dynamic variations encountered with the human voice. There are monitor scopes sold that will fill the bill completely, or any shop-type scope which has an internal horizontal sweep generator and external vertical deflection-plate connections may be used with the tuning unit to be described. Several inexpensive scope kits are also available.

An audio generator is the other piece of test equipment required. The standard sort of audio

generator will do; one often can be borrowed from local RTTYers or high-fi buffs, or a simple audio generator may be constructed to give a selection of frequencies.¹ The generator should have good sine-wave output and low distortion. A two-tone generator makes testing even easier. The unit described by W1YDS² is perfect for any of the tests to be described.

For the service-type oscilloscope an r.f. pick-up unit is used to sample the output of the transmitter, and a tuned circuit builds up the r.f. voltage to provide adequate vertical deflection for the scope. See Figs. 1 and 2. The pick up unit is constructed in a $4 \times 2\frac{1}{2} \times 2\frac{1}{2}$ -inch Minibox. The tuning unit has link-coupled input; each link is made by winding two turns of hook up wire around the center of the coil and cementing it down. Solder lugs are used on the ends of the leads from the Miniductor coil and link to facilitate coil changing. The shaft of the variable capacitor must be insulated from ground. In the unit in the photograph, the capacitor is mounted on a $\frac{3}{4}$ -inch stand-off insulator, as is the terminal strip. The chassis of the tuning unit is made from a 4×8 -inch piece of aluminum sheet stock, although a wooden block would do just as well, as you would not need to use the stand off insulators.

Only a small amount of energy is used by the tuning unit, so the pickup unit may be left in the transmitter output line for on-the-air monitoring.

A typical test setup is shown in Fig. 1. All testing should be done with a dummy load, for two-tone tests can make a horrible noise on the

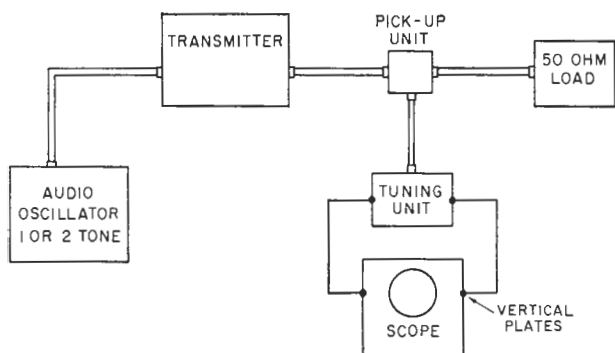


FIG. 1—A typical test setup for a sideband transmitter.

¹ Baxter, "A Transistor Audio Oscillator," *QST*, February, 1965.

² See the article by Walter Lange elsewhere in this chapter.

air. The audio or two-tone generator is connected to the microphone jack of the transmitter, except when a mike is used for speech patterns. The generator should be adjusted so that its output is about at the level of the microphone you normally use. Gain adjustments should be done at the transmitter with the mike gain control. The pick-up unit is inserted between the transmitter and dummy load, and the tuning unit should be placed so short connections can be made to the scope. Don't forget to ground the scope to the tuning unit. A length of RG-58/U or RG-59/U is used to connect the tuning unit to the pick-up unit.

The transmitter to be tested should be tuned up in the c.w. position, or in the sideband position with a single audio tone injected, for normal input. Then adjust the tuning unit to give about half-scale deflection on the scope face, and turn on the horizontal sweep generator in the oscilloscope. Then you are ready to start testing!

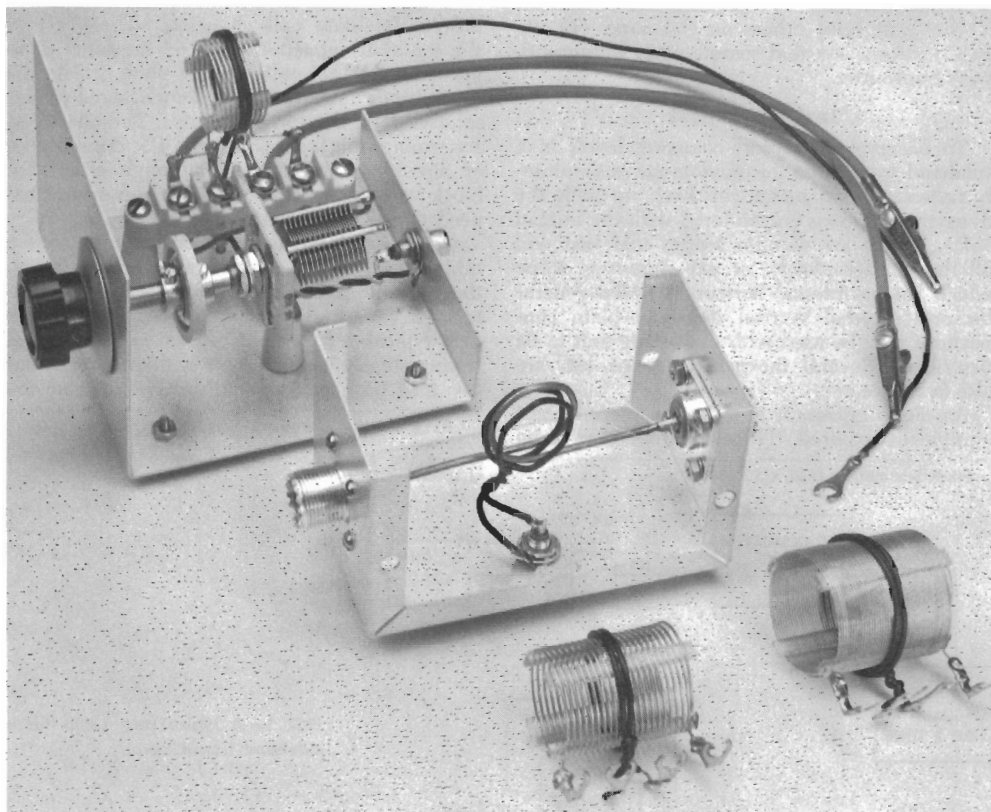
Speech Patterns

Speech patterns offer rather a poor way of telling what is going on in the sideband transmitter because they come and go so fast. Yet

with a little experience one can learn to recognize signs of transmitted carrier and flattening. These are useful later in monitoring on-the-air operation with a scope.

Connect a microphone to the transmitter, set the oscilloscope sweep for about 30 c.p.s. and say a few words. The number "five" will produce a "Christmas tree" pattern similar to Fig. 3A. Each different word will produce a different pattern, which is one of the reasons why speech patterns are so hard to interpret. The important thing here is to observe the peaks to see if they are sharp, as in Fig. 3A. Fig. 3B is the number "five" again but this time the mike gain is set way too high; the final stage is being overdriven resulting in clipping of the voice peaks as the final tube reaches plate-current saturation. Underloading the final stage will produce the same results. Operating a transmitter this way will produce a lot of splatter, making you unpopular with your neighbors on the band. Usually, reducing the gain control a little will remove all signs of flattening. Try different settings of the gain control until you can tell a correct pattern from one showing clipping.

If, when the mike gain is reduced to zero,



The variable capacitor is used to adjust the vertical deflection on the scope. The tuning unit should be mounted near the oscilloscope so short leads to the deflection plates may be used. An extra lead is provided to ground the tuning unit to the scope.

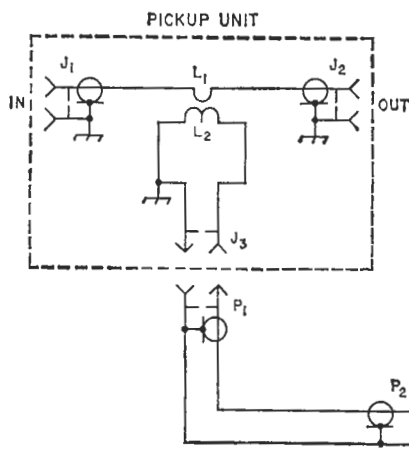
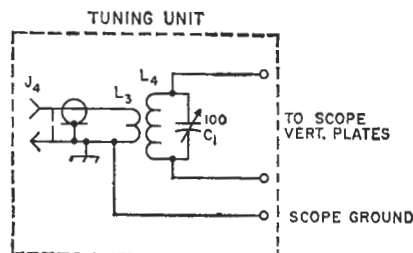


FIG. 2—R.f. sampling and tuning units to provide deflection voltage for the vertical plates of an oscilloscope.



C_1 —100-pf. variable (Hammarlund HF-100).

J_1, J_2 —Coaxial receptacle, chassis-mounting.

J_3, J_4 —Phono connector

L_1 —1 turn No. 14, 1-inch diameter, supported by soldering its leads to the coax receptacles.

L_2 —2 turns insulated hook up wire, placed with one turn either side of L_1 .

L_3 —2 turns of insulated hook up wire wound around

the center of L_4 and ceramic in place.

L_4 —3.5 Mc.: 45 turns No. 24, 1 1/4-inch diam., 32 turns/inch (B&W Miniductor 3020).

7 Mc.: 22 turns No. 20, 1-inch diam., 16 turns/inch (B&W Miniductor 3015).

14-21 Mc.: 9 turns No. 20, 1-inch diam., 16 turns/inch (B&W Miniductor 3015).

P_1, P_2 —Phono plug.

the scope pattern shows you still have some output, you may be transmitting carrier. Adjustment of the balanced modulator, which is covered later, will be necessary.

Two-Tone Tests

A sideband transmitter should be a linear device from mike jack to output connector—for each audio frequency put in you should get out an r.f. frequency, with no distortion of the wave-form. The basis of a two-tone test is that you inject two audio signals, from which you should get out only two r.f. signals. No tube is ever perfectly linear, so some mixing of the two tones will occur, but all of the new signals produced should be so weak in comparison with the main output of the transmitter that you cannot detect their presence in a scope pattern. What you will see is the pattern of two sine-wave signals as they add and subtract, forming peaks and valleys.

A two-tone test's main advantage is that it will produce a stationary pattern that may be examined for defects. It is not easy to tell with your eye exactly what is a pure sine wave on a scope. Complex patterns are even more difficult, so it is a good idea to draw the correct pattern carefully on a piece of tracing paper, which then may be placed over the actual pattern on the scope face for comparison. Remember that this test will show major defects in the transmitter only.

To make the test, apply the output of the

two-tone generator to the mike jack, set the scope sweep for about 200 c.p.s., and check the pattern to see that both tones are of equal level. If they are not equal level, the valleys of the waveform will not meet at a single point on the zero line. Fig. 4A shows the correct pattern; note that the crossover is in the form of an X. Another way to obtain a two-tone test signal is to use a single audio tone and unbalance the carrier to the point where it forms the pattern shown in Fig. 4A.

Examine closely Fig. 4A—this is the correct pattern. Note the clean rounded peaks and straight sides of the envelopes, and again how an X is formed at the crossover. Fig. 2B shows mild flattening of the peaks, and 4C severe flattening. The cause is the same: an amplifier stage being overdriven or underloaded. Cutting the drive level or increasing the loading should result in the Fig. 4A pattern.

Incorrect bias adjustment can also cause a stage to be nonlinear. This defect will show up as rounding of the crossover points as in Fig. 4D. The manufacturer's instruction manual should be consulted for the proper bias value and the location of the bias control. This control should be adjusted for the proper operating bias. Incorrect bias will also show up as high or low values of resting plate current. If a correct resting current and pattern cannot be obtained the tube may be bad and should be replaced.

Fig. 4E indicates what happens when an

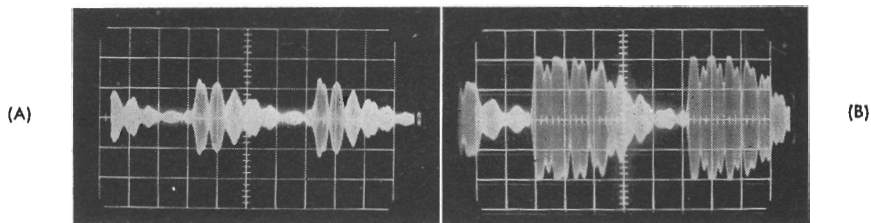


FIG. 3—(A) Speech pattern of a correctly adjusted sideband transmitter. (B) The same transmitter with excessive drive causing peak clipping in the final amplifier.

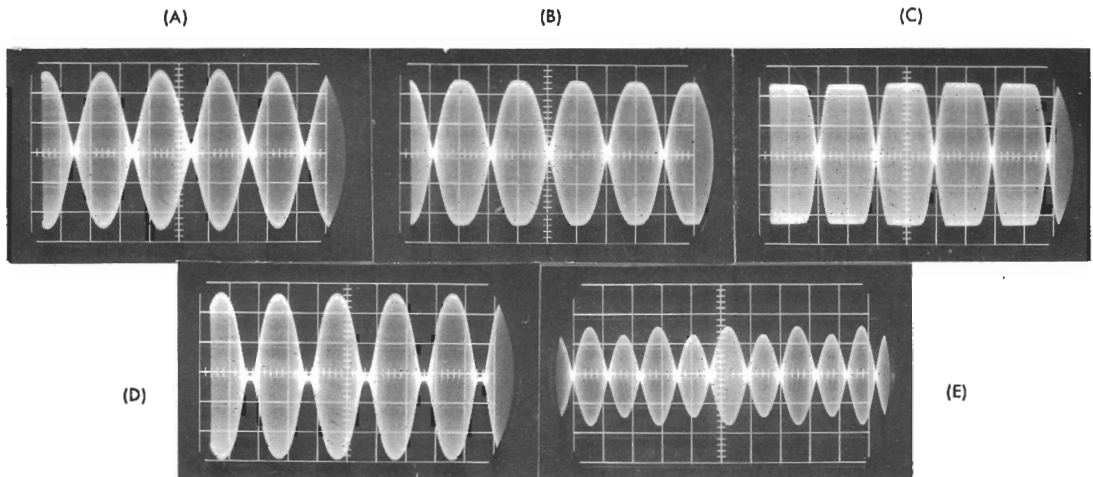


FIG. 4—Sideband two-tone test patterns: (A) a correctly adjusted transmitter, (B) mild peak clipping and (C) severe peak clipping caused by excessive drive or underloading of the amplifier, (D) incorrect amplifier bias causing rounding of the crossover points, (E) pattern with modulation caused by carrier leak-through.

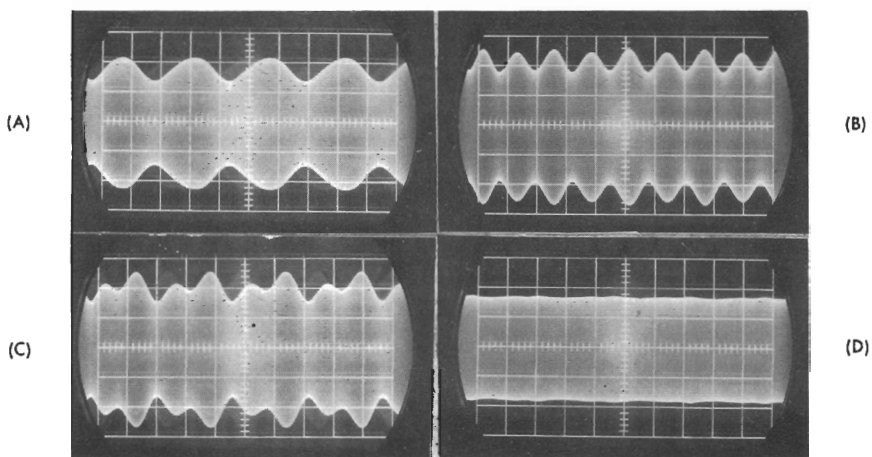


FIG. 5—Phasing-type exciter patterns with single-tone input and constant oscilloscope sweep frequency: (A) carrier leak-through, (B) insufficient unwanted-sideband suppression, (C) both carrier leak-through and unwanted sideband, (D) correct pattern for single-tone input.

external two-tone generator is used and carrier leak-through is also present. The carrier causes the peaks of the two-tone pattern to have different heights. If this happens, you should first null out the carrier, then go back to the two-tone testing.

Carrier Balance

For carrier balance adjustments only one tone is used. The pesky carrier shows up as a sine-wave modulation, similar to what you may have seen in a.m. The carrier-balance control(s) should be adjusted until the sine-wave modulation disappears. Fig. 5A shows the single-tone test with sine-wave modulation caused by a partially suppressed carrier, and Fig. 5D shows the pattern after the carrier has been balanced out.

The location of the carrier-balance controls may be found in the instruction manual if they are not located on the front panel. Phasing rigs usually have two controls, while the filter types have one control and a variable capacitor. In either case the action of these adjustments is somewhat interlocking. The first should be adjusted, then the second, repeating in turn until the carrier is nulled out.

Carrier balance may also be adjusted with the aid of a communications receiver if it has an S meter. The receiver should be coupled to the transmitter so you have a strong, S9 signal. Then adjust the balanced modulator as before for the least amount of indicated signal on the S meter. During this test the mike gain should be reduced to zero, so no modulation appears on the carrier.

The Phasing Exciter

With a single-tone input, the phasing-type exciter owner may find he has a pattern resembling Fig. 5C. This pattern is formed when you have carrier leak-through and also the unwanted sideband is only partially suppressed. The unwanted sideband, like the carrier, shows up as sine-wave modulation, but at twice the frequency. For a scope sweep of about 400

c.p.s. Fig. 5A shows carrier unbalance, Fig. 5B insufficient unwanted-sideband suppression, and Fig. 5C is a combination of the two.

If you need to realign a phasing-type exciter, carefully read the manufacturer's instructions. Different models will require different procedures, so we shall only go over the high points.

Use a single-tone test, and null out the carrier. Carrier balance should be checked again during the alignment of the phasing generator, for it will upset your pattern if any carrier creeps in.

The phasing generator will have audio balance and r.f. phasing adjustments, which may number two, three or four. Injecting a single tone, these controls are manipulated until the unwanted sideband is suppressed, as indicated by the disappearance of ripple from the scope pattern. You should achieve a pattern like Fig. 5D. Then the transmitter should be switched to the other sideband and checked again for suppression of the unwanted. No doubt you will see some sign that in this position you do not have complete suppression of the unwanted sideband. This is probably due to the stray effects in the sideband switching. Thus you must reach a compromise in the setting of the audio balance and r.f. phase that gives good suppression of the unwanted when the transmitter is operated on either sideband. If you have achieved a pattern like Fig. 5D you have carrier and unwanted-sideband suppression of 35 to 40 db. This is about the limit of suppression you can see on an oscilloscope.

If your transmitter has passed all the above tests, you can be sure it is working well. Further tests that will show the small distortion that you cannot see on a scope will require much more advanced techniques, and are beyond the scope of this article.³ In most cases, anything that does not show up in these tests will never be noticed on the air.

³ For more information see *Single Sideband Principles and Circuits*, Pappenfus, Bruene, and Schoenike, McGraw-Hill, Inc., 1964.

PULSED SIGNALS THROUGH SSB TRANSMITTERS

Anyone who builds Walt Lange's pulsed two-tone test generator¹ is going to see some unusual pictures on his scope, and they can stand a little explaining. For example, sending bursts of 800- and 1800-cycle audio through a phasing-type s.s.b. exciter results in patterns as shown in Fig. 1. Here the sweep has been synchronized with the pulse rate, so that the pulses "stand still." It will be found that the two-tone test pattern "walks through" these pulses, since the pulse rate is not usually an exact sub-multiple

of 1000 cycles (the difference between 800 and 1800). The speed of the pattern's walk through the pulse can be controlled by changing the pulse rate and, while the pattern may stand still momentarily, it is not likely to be motionless for more than a few seconds. When testing a linear amplifier, it is the center of the pattern in the burst that is observed, as in the steady two-tone test pattern. However, it is the beginning and ending of the bursts that we found interesting.

Note that the left-hand edge (start of pulse) of Fig. 1 shows a large, sharp spike, of ampli-

¹ Lange, "A Pulsed Two-Tone Test Oscillator," elsewhere in this chapter.

tude almost double that of the normal pattern. As the 1000-cycle pattern walks through the pulse, this spike grows larger and smaller. A similar but smaller spike is generated when the pulse is turned off, and some tailing off or "ringing" can be observed. (If the pulse envelope were a perfect rectangle, the spikes would theoretically be of infinite amplitude. The audio phasing network in the exciter was protected by a low-pass filter that "softened" the pulse.) The spikes are the result of trying to transmit too many different frequency components by s.s.b.; at the instant when all of the components are in phase, the instantaneous amplitude must equal their absolute sum. If such an r.f. signal were to be amplified in a subsequent *linear* amplifier (no clipping allowed), the presence of the spike would place a severe restriction on the amplification of the main body of the signal. There is little energy in the spike, but to clip it, to bring up the level of the main body, would be a distortion of the signal.

Feeding the pulsed two-tone test signal to a filter-type s.s.b. generator gave the envelope of Fig. 2. Here the restricted bandwidth of the filter prevents the spike components from passing through, and as a consequence the start and finish of the pulse are different than they were when the phasing exciter was used.

Readers with long memories may recall that the first article on pulse testing of linear amplifiers² showed sketches of envelope pictures, and the pulsed signals showed no spikes or distortion at the beginning and end of the pulses. There was a very good reason. In the earlier venture the two-tone test signal was obtained directly

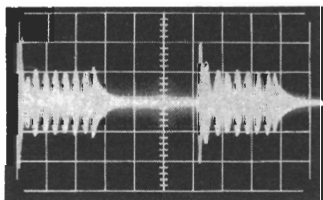


FIG. 1—Pulsed two-tone test pattern obtained with a phasing-type s.s.b. exciter. Note the large spike on the leading edge of the pulse and the lesser one on the trailing edge.

from the output of a balanced modulator with a *single* audio tone applied. The resultant r.f. signal, a pulsed two-tone test signal as far as the amplifier was concerned, was a *double*-sideband signal when it was generated. Double-sideband signals, with or without carrier, can be pulsed in the audio circuits before the modulation process, and there will be no spikes on

the resultant r.f. signal. This is easy to see in the case of an a.m. signal (d.s.b. with carrier), because the r.f. envelope of an a.m. signal is identical to the audio signal used for modulation.

(C.w. or A1 transmission is a double-sideband modulation process. When the r.f. is turned on or off too fast, key circles are produced equally above and below the carrier frequency. An

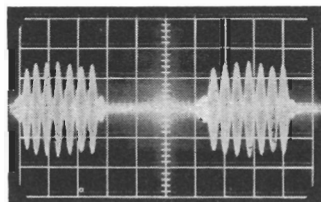


FIG. 2—Pulsed two-tone test pattern obtained with a filter-type s.s.b. exciter.

oscilloscope picture of such a signal never shows an instantaneous value greater than the key-down value, except when the power-supply regulation is poor, and this has nothing to do with the subject under discussion.)

In the case of the d.s.b. suppressed-carrier signal there is no obvious direct relationship between the envelope and the modulating signal, but the amplitude of the r.f. envelope is always proportional to the amplitude of the modulating signal.

An s.s.b. signal insists on being different. And this is why only a small amount of audio clipping can be done in an s.s.b. rig before any advantage is lost. As clipping becomes appreciable and the audio acquires additional higher frequencies, one of two things must happen. In a phasing rig spikes will appear if the additional higher audio frequencies can generate an s.s.b. signal. (The case where they can't is where they are much too high for the audio phase-shift network or where they are completely rejected by a protecting audio filter.) The spikes restrict the power output of a linear amplifier. In a filter rig the same thing happens if the bandwidth is great enough to pass the high-frequency components. If the bandwidth is less (ours was 2.1 kc.) the filter rejects the high-frequency components, the effect of clipping is lost, and one ends up back where he started.

With narrow-band filters, advantageous clipping can be accomplished at r.f. *after* the generation of the s.s.b. signal, but the signal must then be reprocessed by passing it through a second filter.³—WIDX.

² Goodman, "Linear Amplifiers and Power Ratings," *QST*, August, 1957.

³ Squires and Clegg, "Speech Clipping for Single Sideband," *QST*, July, 1964.

» Is that loud signal splattering, distorted, using more spectrum space than it should? Superficially, perhaps yes; but before you jump to the wrong conclusion, learn how to eliminate the misinformation your receiver will hand you if you aren't wary.

Checking Signal Quality With the Receiver

GEORGE GRAMMER, W1DF

Any receiver that will bring in c.w. signals satisfactorily can be used for checking phone signals, either a.m. or s.s.b. The check is purely qualitative, but will go a long way toward the goal of keeping signals clean.

You don't have to know much about your receiver's technical characteristics. It's largely a matter of knowing how to set the controls and knowing what to look for. The "how" is easy; the "what" takes some practice—critical observation and comparison of the various kinds of signals you run across on the air. While there isn't anything complicated about it, the technique differs from that used in ordinary reception.

First, about the receiver's controls. *Turn off the a.g.c.* This is vital. Any variation in receiver gain while you're examining a signal makes it practically impossible to interpret what you hear. Set the audio gain well up and turn the r.f. gain down to the point where the average signal is of moderate strength. Turn on the b.f.o.

Beware of Overloading

Before doing any phone checking you have to find out something about the receiver's ability to handle signals. An easy way is to tune across a c.w. band. When you come to a strong signal, vary the r.f. gain control. If the audio output keeps coming up as you increase the gain, the control is operating in the right region. If the output starts to level off at some point on the gain control, the receiver is beginning to overload. There is a change in the character of the beat note at that point; the tone begins to sound a bit thin or mushy. Also, signals and noise in the background will "bounce" in intensity with the keying of the signal. These effects will readily be recognized after you've heard them a few times. Pick out the strongest signal and set the r.f. gain well below the point where overloading starts. You should still be able to get all the output you need by increasing the audio gain.

Unless the controls are set in this way the

receiver can't handle the stronger incoming signals without overloading. Overloading has to be avoided at all costs if your observations are to be useful.

Adjusting the B.F.O.

Next, set the receiver's selectivity to maximum and turn off the b.f.o.¹ Tune in a c.w. signal by adjusting the tuning control so the response to the background noise is maximum when the sender's key is down. An unmodulated steady carrier can also be used, if such a signal happens to be available.

When the gain controls are adjusted as described, the background noise *increases* when a signal is present. This is opposite to what happens when the a.g.c. is used and the manual r.f. gain is at maximum; in that case the background noise *decreases* when a signal is tuned in.

Finally, turn on the b.f.o. and adjust it to give a beat tone of about 500 cycles on the signal so tuned in. Either side of zero beat can be used.

Checking a Phone Signal

At this point you're ready to take a look at a phone signal. The a.m. broadcast band is a good place to start, if your receiver happens to be one that covers it. Broadcast modulation is likely to be held under proper control, and your object is to find out what the sidebands of a *properly* modulated signal are like.

First, tune in a local b.c. carrier, adjusting the tuning for the selected beat tone. The modulation will sound like a miscellaneous collection of beat tones. Next, tune off about a kilocycle to the side which makes the carrier beat tone rise in frequency. You'll now be well into one of the two sidebands, and if the receiver selectivity is high the carrier beat

¹—It may not always be easy to do this, since the b.f.o. and a.g.c. cannot be controlled independently in some receivers (although it is usually practical to pull out the b.f.o. tube temporarily). Also, receivers with product detectors do not lend themselves to this method of setting the b.f.o. frequency since the detector does not (or should not) function when the b.f.o. is not operating. In such cases the b.f.o. has to be set to give approximately the desired tone on background noise. This is good enough if the selectivity is high.

Based on "Looking at Phone Signals," December, 1962, *QST*, and "Checking Signal Quality With the Receiver," March, 1963, *QST*.

either will be much weaker or will have practically disappeared. Listen carefully to the beat tones that rise and fall with the modulation. They will have a clean, smooth sound—a little hard to describe accurately but easily recognizable after a short listening session. These smooth-sounding beats are “legitimate” sidebands.

If your receiver is an amateur-bands-only affair you can't get this preliminary practice on an a.m. broadcast station, but don't let this discourage you. Any amateur phone signal, a.m. or s.s.b., can be used. Try to find one that isn't fading and is free of interference, so you can be sure you are hearing it alone. Tune through the sideband as described, listening particularly to the lower-level beat tones and ignoring the strong peaks. If there is over-modulation or nonlinearity it is sure to be intermittent and associated only with voice peaks; in between, the lower-amplitude sidebands will be clean.

Bandwidth

If the receiver tuning dial is calibrated closely enough it is possible to get a fairly accurate idea of the transmitted bandwidth by this beat method. Concentrate on those beats which have the same tone for which you set the b.f.o. at the start. Find the frequency setting, farthest from the carrier, at which you get that tone from a sideband component. Then the difference between that dial reading and the dial reading for the carrier is equal to half the signal bandwidth if this was an a.m. signal—half, rather than total, because you've looked at only one of the two sidebands.

Estimating bandwidth by this method requires the ability to concentrate on the right beat tone. Obviously, it is easier to recognize the beats when the receiver has high selectivity, because then the strongest beats will always be around the right tone regardless of the tuning-dial setting.

Analyzing the Process

If you aren't wholly familiar with receiver operation a diagram of this process may help.

FIG. 1—A properly modulated a.m. signal may have, instantaneously, side frequencies distributed something like the pattern in this drawing. The frequency pattern will vary from instant to instant with voice modulation.

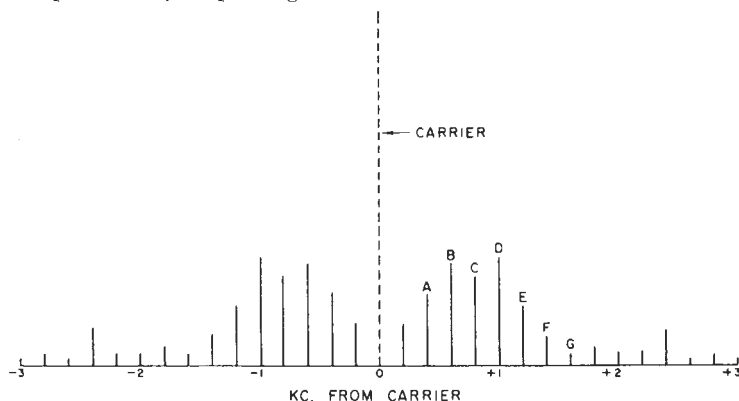


Fig. 1 is typical of the frequency-vs.-amplitude distribution that might exist in a good a.m. phone signal at some instant. Each sideband consists of a series of frequency components associated with a voice sound. These components usually have harmonic relationship, to a close degree, for any given sound; in Fig. 1, all the side frequencies shown are produced by audio tones that are harmonics of 200 cycles. More important, however, is the fact that each sideband consists of a group of *distinct* frequencies. It is not just a continuous mess. Each separate frequency gives a separate, and reasonably stable, beat tone with the receiver's b.f.o.

If the receiver can handle a group of these frequencies without doing injustice to any of them—i.e., without overloading—the individual beat components will stand out just as any one of a similar group of closely spaced c.w. signals will retain its individuality. Sideband components of this sort are generated in a properly modulated transmitter, and sound “clean” with the receiver's b.f.o. on.

By using as much selectivity as the receiver offers, the number of sideband components heard at any one time is narrowed down. In Fig. 2 a curve typical of “500-cycle” selectivity is shown superimposed on the lettered group of sideband components from Fig. 1. The response range shown is 60 db. If the receiver is tuned to the frequency of side component D, the response to that component will be as shown by the vertical line. This response is relative to the carrier-only response; the scale here differs from that of Fig. 1 because the former was plotted to an intensity (voltage or current) scale while Fig. 2 is in decibels. The sideband components labeled B, C, E and F would have the decibel response shown, as a result of the effect of the selectivity on their original amplitudes. Note that A and G are so far down (more than -60 db.) that they do not even show on the graph. This is also true of all components higher in frequency than G and lower in frequency than A, including the carrier.

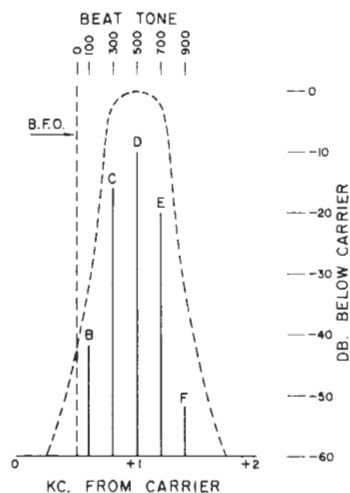


FIG. 2—With high selectivity, only those sideband frequency components to which the receiver is actually tuned will give appreciable response. This drawing shows the relative response a selective receiver would give on the lettered components in Fig. 1. The scale at the top shows the beat tone each component would produce when the b.f.o. is offset 500 cycles from the peak of the selectivity curve. In this case only C, D and E would result in appreciable audio output.

If the receiver's b.f.o. is offset from the selectivity curve by 500 cycles as shown (this was the object of the method of setting the b.f.o. frequency detailed earlier) each sideband component will give a beat tone as shown in the upper scale. The selectivity restricts these tones to a relatively narrow range centering around 500 cycles. This also will be true when the receiver is tuned to other parts of the signal. When this point is appreciated the beat-tone method of checking bandwidth becomes clear.

Practically speaking, a sharply peaked selectivity curve—such as the kind a Q multiplier or the old-type crystal filter give—is best for this type of checking. While your mind can be trained to exclude those tones which differ appreciably from the one for which you originally set the b.f.o., it is easier with a highly peaked selectivity curve because then only a frequency component right on the peak—that is, one that gives the selected beat tone—really stands out.

Splatter

Splatter frequencies arising from overmodulation and nonlinearity tend to have a different character than legitimate sidebands. There is a harshness associated with them that again is hard to describe but not hard to recognize. Listen particularly with the tuning set toward the edge of the band you found to be occupied during normal transmission.

The harshness associated with splatter is the result of a different type of sideband-frequency distribution. The onset of splatter is usually abrupt, giving an effect something like key clicks. Also, the side frequencies it generates are often much more closely spaced than the sideband components of proper voice modulation, so that distinct tones are less easily recognizable.

Checking Incoming Signals

With this background in checking modulation you're in a position to take a look at amateur signals. However, before condemning any signal you hear as not being up to par, ask yourself two questions: First, is there any possibility that the receiver is being overloaded, either by the signal in question or by one that may be far enough removed in frequency so that you aren't aware of its presence? That r.f. gain control setting is important. Second, if there are harsh "burps" indicating splatter from overmodulation or s.s.b. flattening, do they belong to the signal you're blaming? In a crowded band identification of bits and pieces of splatter is sometimes pretty difficult.

In other words, make sure that the signal being checked is the one you're actually hearing, and that no spurious receiver effects are being introduced. An overloaded receiver is worthless as a checking device. Most receivers have so much gain that even a weak signal can be amplified up to overload point unless care is used in holding down the amplification. The lower you can run your r.f.-i.f. gain, the better. And *never* use the a.g.c. when making signal checks.

S.s.b. signals differ from a.m. only in the absence of the carrier and one sideband. Properly generated and amplified, the sideband components will have the same clean sound to them that properly modulated a.m. sidebands do. Overdriving a linear amplifier will result in "burps"—especially noticeable outside the desired-sideband channel and particularly in the undesired-sideband region—just as a.m. overmodulation does.

Since there is supposed to be no carrier with s.s.b., the receiver's b.f.o. must be set up on a c.w. signal or unmodulated carrier as described earlier. This is obviously not the same setting that would be optimum for s.s.b. reception; the b.f.o. frequency is offset by 500 cycles or so from the s.s.b. setting. With this offset, you can easily determine whether any carrier is being transmitted; a continuous carrier will give a steady tone, usually weak compared with the sideband, but nevertheless present. You can also detect a carrier that rises with modulation. It is "keyed" along with the voice, sounding something like slow c.w. with a very soft make and break. This is caused by incomplete carrier balance, which may be a dynamic effect—that is, the carrier may be

quite well balanced out when there is no modulation, but the modulator becomes unbalanced when it is being driven by audio.

With high selectivity it is possible to check the bandwidth of an s.s.b. signal by the beat method, and particularly to see whether there is appreciable output in the undesired sideband region. As shown by Fig. 2, the beat tone that your b.f.o. is adjusted for will predominate only when a sideband component is on the frequency to which the receiver is set. If your mind is trained to exclude any other tones you may hear, you may be sure that you aren't being deceived by instrument errors. The selectivity has to be high enough so that the audio image of the b.f.o. tone is negligible; in other words, you have to have true single-signal c.w. reception.

Transmitter Checking

Of course, all this is only preliminary to the real object—checking your own transmitter. Practice on incoming signals of all types will give you the insight needed for analyzing your own signal. Having found out how to spot defects in others, you're well prepared to find out what, if anything, is wrong with your own.

The requirements for transmitter checking can be stated in simple terms: The transmitter's signal must be reduced in strength to a level well within the receiver's normal signal-handling capability. But transmitter testing has meaning only when the transmitter can deliver its full output, while FCC regulations forbid the extensive one-way transmissions you have to make in finding out what, if anything, is wrong. So testing on the regular antenna is "out." The use of a dummy antenna is mandatory. There are low-cost commercial dummies available, including kits, for practically any legal amateur power level.

Test Setup

The complete test setup is shown in Fig. 3. An essential part of it is the "actuator"—the substitute for *you* in your regular capacity before the microphone. You can't talk and do a good job of listening to your signal at the same time. Neither can you hope to enlist someone else's voice for an extended period. What is needed is an untiring source of audio comparable with what you put into the microphone yourself. Also, if you want to use a speaker instead of headphones in your testing it must be a *silent* source. The ideal actuator is a tape recorder. If you have one, record your own voice and do your testing under conditions as close as possible to actual operation on the air. Recorders usually have preamplifier or external speaker connections, or both, from which audio can be taken, and it requires no circuit diagram to feed one or the other of these outputs into the microphone jack on the transmitter.

If the output voltage level from the recorder is higher than is desirable for going into the microphone preamplifier, cut down the gain in the recorder's amplifier so no stage ahead of the gain control in the transmitter's speech amplifier will be overloaded. If hum becomes bothersome when this is done, it can be overcome by using a simple external attenuator as shown in Fig. 4. R_1 should be about 10 times R_2 , and the sum of the two should equal whatever resistance the preamplifier output of the recorder is intended to work into, if the preamplifier output is used. As this resistance value is fairly high, shielded wire should be used for the connections, in order to avoid stray hum pickup. It may also be necessary to shield the resistors, which can easily be done by wrapping them with aluminum foil over a wrapping of paper for insulation, with the foil connected to the shields on the connecting wires.

If the audio is taken from the speaker output terminals, the total resistance may be of the same order as the voice coil impedance, usually around 8 ohms. The value isn't critical, and as long as a low resistance is used, shielding shouldn't be necessary. Needless to say, the recorder's internal speaker should be shut off if you want to listen with a speaker on your receiver.

If you don't have a recorder there are still other possibilities. A phonograph is one; there are many 100-per-cent voice recordings that are suitable for the purpose. The output of a phono pickup is not generally usable directly, since a crystal or ceramic pickup ordinarily has too much to simulate a microphone and a magnetic has too little. Here again you can take the output from a preamplifier, using an attenuator as in Fig. 3 if necessary. The same type of attenuator can be used directly on a crystal pickup, with resistances totaling something of the order of 1 to 5 megohms. Shielding is a necessity with such high resistances.

Still another source of continuous talk, or very nearly so, is the a.m. broadcast band. Audio can be taken from the speaker voice-coil terminals in a b.c. receiver, but use caution with small power-line radios. Make sure that neither voice-coil terminal is tied to a "hot" a.c.-d.c. chassis before you try this method. The output-voltage problem is the same as with the recorder, and should be handled in the same way. One speaker lead will have to be disconnected from the speaker itself if you want "silent" audio. A transistor set is handy because of its portability and because it will have no hum. Even an old-fashioned crystal receiver can be used.

By one means or another, a suitable actuator can be rigged up at little or no cost. It would be hard to find a household without a radio, and few are without a phonograph. Even the tape recorder is fast becoming a household item.



The Receiver

A normally-shielded transmitter working into a dummy antenna, even if the dummy is not shielded, should not radiate more signal than can be handled by the receiver. No doubt it will be necessary to disconnect the receiving antenna; after all, the "spray" from the transmitter will still be rather strong within a few feet of the set. Here a great deal depends on the over-all shielding, both transmitter and receiver, so it is possible to talk only in general terms. Reread what was said earlier about setting the receiver's controls. You should aim to get the signal pickup down to the point where you can use about the same gain settings on your own signal as you did on distant signals when the receiving antenna was connected. If the receiver, transmitter and dummy antenna are really well shielded, it may be necessary to use a few inches of wire as a receiving antenna in order to get the needed signal strength. If the signal is too strong, try running the antenna trimmer off tune, and if that doesn't do it, try pulling out the r.f. amplifier tube in the receiver—anything that will let you get a moderately-strong signal with the gain settings you found optimum for listening to incoming signals.

One further point needs consideration in using the receiver for monitoring. In s.s.b. testing the load that the transmitter puts on the power line varies with the modulation. This may cause the line voltage to fluctuate, possibly with adverse effects on the receiver's stability. To settle this question, use the receiver normally—i.e., with the antenna connected and an incoming signal tuned in. Pick a frequency sufficiently far from your transmitting test frequency so there is no interference from it.² Let the transmitter operate into the dummy antenna and watch carefully for any shift in naturalness on s.s.b. while your transmitter is being modulated. If the receiver stands this test, you're ready to go. If it doesn't, there is no simple alternative but to try to find an a.c. outlet for the receiver that won't show such large voltage changes. While instability of this sort won't have an apprecia-

² If connecting the antenna to the receiver causes feedback troubles, the transmitter can temporarily be put on a different band, preferably higher in frequency, while the receiver is being checked in this way.

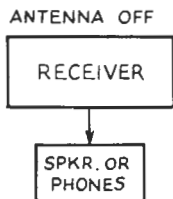


FIG. 3—Setup for using the station receiver for transmitter checking.

ble effect on the bandwidth of the transmitter, as measured by the receiver, it can be misleading when listening to sideband beat tones.

Once you're sure you've eliminated any possibility of receiver overloading and instability, examine your transmitter's signal carefully. Using the highest available selectivity, check the bandwidth as described earlier, and listen particularly for spurious "burps" outside the channel that the signal should occupy legitimately. As you can readily vary the audio gain in the transmitter while listening, it is no problem at all to find the level at which spurious sidebands start to become noticeable. In turn, this level can be observed on the transmitter's meters. Their readings may surprise you in comparison with what you've been seeing in your ordinary operating. But after a test such as this they will take on some real significance, where before you had been working in the dark.

To have the most meaning, the actuating signal should be your own voice, which is why a tape recorder makes such an excellent addition to the test gear. If you have to use other voices, try to avoid those having entirely different pitch and timbre. If a radio is the "actuator," scout around among the disk jockeys and compare the results.

Testing in this way doesn't strain finances, but when done intelligently it will give you all the information you need about your signal. This, and the confidence that your transmissions will stand critical examination, should be more than ample payment for the small trouble and the time off the air.

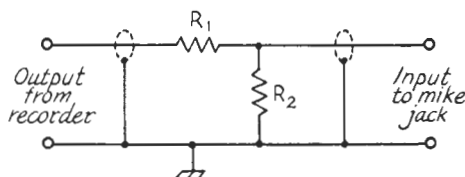


FIG. 4—Simple voltage divider for reducing audio voltage to a manageable level for the transmitter's speech amplifier. Ordinarily R_1 will have about ten times the resistance of R_2 . A variable control having the same over-all resistance can be substituted for the two resistors.

» Several solid-state product detectors are described, and comparisons of their performance are given in practical terms. Any one of the circuits can be used at the end of an i.f. strip for demodulating s.s.b. or c.w. signals. All of the detectors were tried as front ends for simple direct-conversion receivers.

Solid-State Product Detectors

DOUG DEMAW, WICER

The product detector is regarded by some amateurs as a mysterious device which can be used in receivers of inferior performance to cure a host of ills inherent therein. Though a product detector, *per se*, is not a panacea for most receiver ailments, it is an important part of any good s.s.b. receiver. The accepted definition of a product detector is that it is a linear demodulator in which two signals are multiplied together to produce a resultant a.f. signal output. The two input signals are usually the i.f. output and the b.f.o. signals, which are closely related in frequency. (Actually, any b.f.o./diode-detector combination is a product detector, since its output is proportional to the product of the b.f.o. and signal voltages.) A properly-designed product detector should be capable of delivering an a.f. signal that is low in audio distortion and intermodulation (IM) products.

It is believed by some that the measure of a product detector is its inability to deliver an output signal when the b.f.o. is turned off. Although some circuits do exhibit that characteristic, it does not mark them as good product detectors. Conversely, if output still exists from the detector when the b.f.o. is disabled, this does not mean that the circuit is not a product detector while the b.f.o. is operating.

The distortion developed in product detectors is the result of improper b.f.o. injection levels in most cases. Maximum distortion occurs when the b.f.o. and incoming signal levels are the same. The actual value of the b.f.o. injection voltage required depends upon the level of the i.f. signal at the detector input. Ordinarily the i.f. input level is between 0.05 and 0.5 peak volts. The magnitude of the b.f.o. signal should be at least 10 times that value to assure minimum distortion. This means that between 0.5 and 5 volts of b.f.o. injection is required. In some solid-state receivers proportionately lower i.f. and b.f.o. voltage levels are used, but the ratio between them should remain the same.

Depending upon how well balanced a detector circuit is, there may be some b.f.o. energy in its output. All components other than the desired a.f. information should be minimized if good product-detector performance is desired.

The b.f.o. injection level, as mentioned earlier, is set to assure minimum IM distortion. The better the circuit balance, the greater will be the b.f.o. carrier suppression. Ordinary r.f. bypassing will reduce the b.f.o. voltage level at the detector output to an acceptable figure. If too much b.f.o. voltage appears at the detector output, it can reach the i.f. stages of the receiver by being radiated from the circuit wiring. If the unwanted i.f. injection is severe enough, the i.f. stage affected may not have sufficient dynamic range to keep it from overloading. Similarly, excessive b.f.o. voltage may reach the grid, gate, or base of the first audio amplifier stage and cause overloading or desensitization of that part of the receiver. The b.f.o. of any receiver should be well isolated from the rest of the circuit, preferably by means of shielding, to lessen the chance of stray coupling into sections of the receiver other than the detector.

The foregoing attributes, at least, are the criteria for good product-detector performance. They are set forth here to serve as guidelines when building home-made receivers which use product detectors. These rules were observed during the design of the product detectors described in this article.

Some Practical Circuits

Shown in Fig. 1 are four product detector circuits that were built and tested in the ARRL lab. At A, two 1N67A diodes are connected back-to-back in the W1DX adaptation of a vacuum-tube diode circuit designed by Don Norgaard, W6VMH.¹ The main advantage of this circuit is its simplicity and its ability to handle high signal levels without overloading. By using the diodes back to back, the last i.f. transformer can have a high-impedance secondary winding because the high back resistance of CR_2 prevents the detector from loading the secondary of the transformer. If CR_1 were used by itself, it would be necessary to tap the diode down on the secondary of T_1 to minimize loading effects. This circuit has a fairly high noise level, hence requires good i.f. amplification ahead of it to override the noise. In a test circuit a 10- μ v. signal (at the detector input) was

From April, 1969, QST.

¹ QST, July 1948, p. 11.

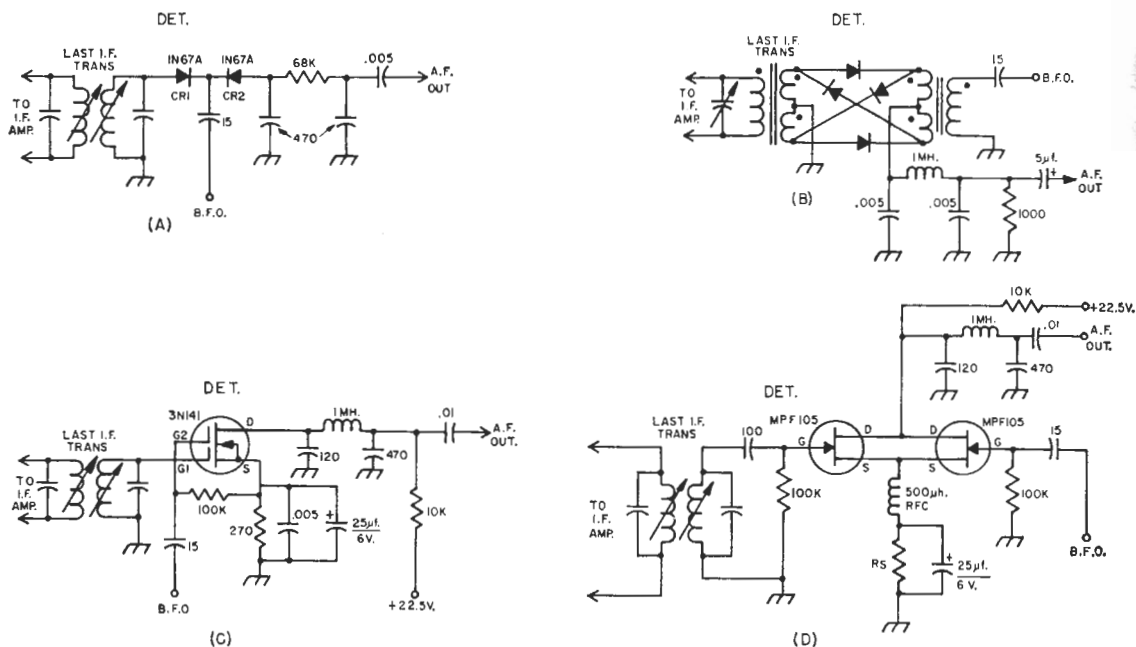


FIG. 1—Examples of four solid-state product detectors. At A, the two 470-pf. capacitors and the 68,000-ohm resistor form a simple r.f. (or i.f.) filter. At B, a 1-mh. r.f. choke and two 0.005- μ f. capacitors comprise the filter. Filtering is provided by the 1-mh. chokes and their associated bypass capacitors in the drain leads of the circuits at C and D. Source-resistor bypassing for audio is used at C and D to boost the a.f. output from the detectors. R_s , at D, should be determined experimentally for best conversion gain and IM characteristics. A value of 1200 ohms worked well in the circuit shown. The circuits at C and D can be operated from a 9- or 12-volt supply by eliminating the 10,000-ohm drain-load resistors and replacing them with small audio chokes whose d.c. resistance does not exceed approximately 1500 ohms.

required to produce an audible c.w. note while feeding the detector output into a typical receiver-type audio channel.

A diode ring demodulator is shown at B, Fig. 1. This circuit is similar to the one described by Hayward and Bingham in November 1968 QST. Trifilar-wound toroidal transformers are used at the input and output sides of the detector to permit proper impedance matching and good circuit balance. B.f.o. carrier suppression is good with this type of detector, and the noise figure is somewhat better than that of the circuit at A. Four 1N67A matched diodes were used in this model, but the hot-carrier diodes used by Hayward might provide superior performance. An RCA CA3019 diode-array IC should be ideal in this type of circuit, offering near-perfect diode matching by virtue of the diodes being on a common silicon chip. With this product detector a 1- μ v. signal produced an audible c.w. note (the same audio amplifier was used with all of the circuits shown in Figs. 1 and 2).

In the circuit of Fig. 1C, a dual-gate MOSFET is shown as a product detector. The 3N141 is an N-channel silicon, depletion-type, dual insulated-gate FET. It has two channels, each with an independent control gate. Designed

specifically for mixer applications, the 3N141 has a wide dynamic range which permits it to handle high signal levels. The signal applied to Gate 2 modulates the input-gate (Gate 1) transfer characteristic. This method is said to be superior to the more conventional square-law mixing method because the latter can be accomplished only in the nonlinear region of the mixer's transfer characteristic. Since mixers and product detectors operate in the same manner, the 3N141 was selected for these tests. Signal isolation between the two gates is good, helping to confine the b.f.o. signal to the desired part of the circuit. With this detector a 0.5- μ v. signal produced a perfectly audible c.w. note at the output of the audio amplifier. The noise figure was extremely low and the conversion gain was good.

Two MPF105 JFETs are used in the product-detector circuit at D, Fig. 1. In this source-follower circuit the gates of the two transistors permit high-impedance input for both the b.f.o. and i.f. signals, while at the same time offering good isolation between the two signals. Connected as shown, this circuit offers some balance through phase cancellation of signal and b.f.o. currents flowing in the common-drain and common-source portions of the circuit. A 1- μ v.

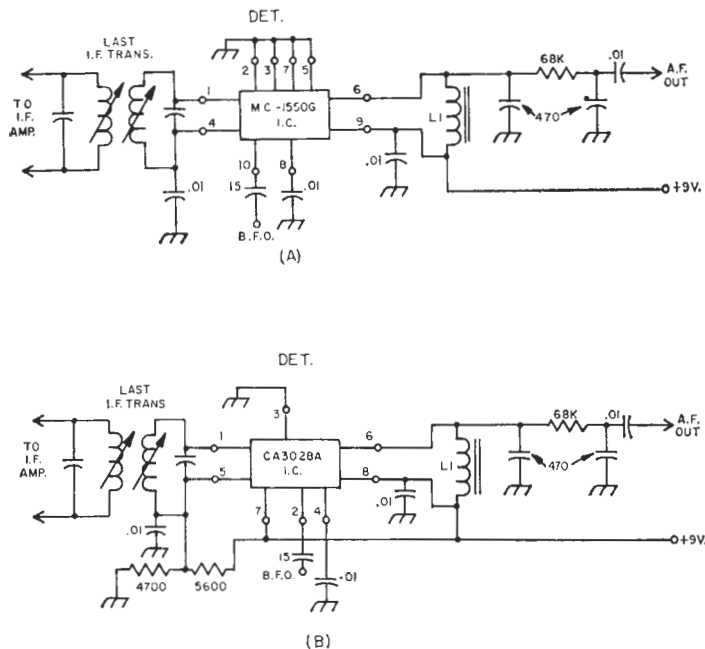


FIG. 2—Circuits for two integrated-circuit product detectors. Though the MC1550G has provision for external biasing it was found that that part of the circuit could be grounded to the chassis with no apparent variation in performance. L_1 is a small a.f. choke. The circuit at B is quite similar to that shown at A, but uses a CA3028A IC. It does use external bias (4700 and 5600 ohm resistors). The bias-divider values were chosen for best overall detector performance and do not follow the ratio recommended by RCA for mixer service. Both of these circuits perform well over a supply-voltage range of 5 to 13 volts with no circuit changes required. The higher supply voltage, however, permits higher signal levels before overloading takes place.

signal was sufficient to produce an audible c.w. note when the output from this detector was passed through the test amplifier. The noise figure was low, but so was the conversion gain.

Integrated-Circuit Detectors

While investigating product-detector performance it seemed like a good idea to try some differential-amplifier ICs in circuits similar to those of Fig. 1. As indicated in Fig. 2, two types were tested. A Motorola MC1550G i.f. amplifier was used in the circuit at A. Its performance proved to be excellent. The outstanding feature was the conversion gain of the detector—far greater than any of the circuits of Fig. 1. This means that the audio amplifier which follows this detector need not have as much gain as that required for the circuits of Fig. 1, or that the i.f. gain ahead of the detector can be somewhat lower than in conventional circuits. This feature should be appealing to those wishing to design simple, low-drain superheterodyne receivers. This detector has good carrier isolation. A $0.1\text{-}\mu\text{v}$. input signal produced an audible c.w. note at the output of the amplifier. The b.f.o. injection level was 1 volt, peak. Very little difference in overall performance was noted while varying the supply voltage from 5 to 12 volts.

An RCA CA3028A differential-amplifier was tried in the circuit of Fig. 2B. This IC is designed for use in i.f. amplifier and mixer circuits. It is quite similar to the MC1550G, but is approximately three dollars lower in cost. Connected as shown, its performance was almost identical to that of the circuit at A. A $0.2\text{-}\mu\text{v}$.

signal provided an audible c.w. note. The conversion gain is quite high with this circuit, offering the same possibilities for use in simple receivers as the circuit at A.

With any of the circuits described in Figs. 1 and 2 the preceding i.f. amplifier chain should have some type of gain control, manual or automatic, to prevent the input signal to the detector from becoming high enough to cause overloading. If these circuits are to be used with existing receivers it may be necessary to lower the gain of the first i.f. stage through modification of the cathode biasing circuit, thus assuring that overloading will not take place at high incoming signal levels.

The b.f.o. used while testing these circuits was similar to the JFET oscillator shown in Fig. 3. Depending upon the size of the injection coupling capacitor, between 0.5 and 5 volts can be obtained for b.f.o. input to the detector. A value of 15 pf. provided the 1-volt level (r.m.s.) used in these tests.

Direct-Conversion Receiver Tests

After going the complete route on product-detector testing, or at least so it seemed, the thought occurred that it would be interesting to try each of the circuits in a direct-conversion receiver lashup similar to that of Hayward and Bingham. Each was tried, and the best overall performance was obtained from the circuit shown in Fig. 3. The 75-meter band was selected for an operating frequency because of its closer proximity to the a.m. broadcast band than any of the other h.f. bands; this permitted the receiver to be tested for cross-modulation from

» A QST article has revived an interest in the direct-conversion technique for reception of radio signals. Simplicity with quite good performance are the advantages. This article describes an experimental receiver using a novel approach to the use of direct conversion for true single-sideband reception.

A Direct-Conversion SSB Receiver

RICHARD S. TAYLOR, WIDAX

An article in November 1968 QST¹ inspired this effort. It is an idea that I rejected some months ago while working with broadband FET mixers because of the high audio gain required. It turns out that the audio gain is not really a problem at all, and that a really effective, yet simple, s.s.b. receiver can be built using direct conversion. Transceive operation seems simple, too.

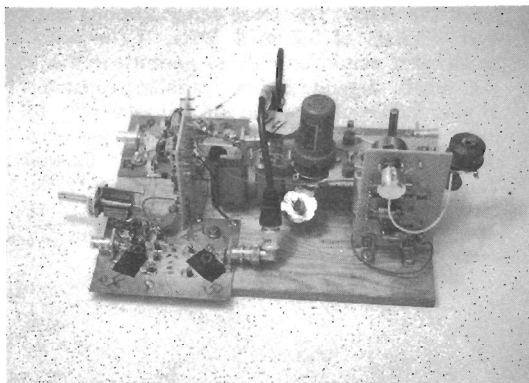
Receiver Operation

Basically, the receiver consists of a pair of balanced mixers operating at the received signal frequency, and a high-gain audio amplifier (Fig. 1). Local-oscillator signals 90 degrees out of phase are applied to the balanced mixers. With the proper choice of audio phase (plus or minus 90 degrees in one leg), one incoming sideband can be rejected and the other enhanced.² A quadrature hybrid network provides the r.f. phase shift, and a Barker and Williamson 2Q4 phase-shift network is used in the audio combiner section.³

Sideband selection is accomplished by reversing the phase of one of the audio channels with respect to the other, as in a phasing transmitter. In fact, the receiver operates as a sort of a phasing transmitter in reverse.

All selectivity is obtained in the audio channel, so a sharp cut-off audio filter is called for.

Assuming the audio response shown in Figure 2A occurs, the r.f. responses of Figs. 2B and 2C result. Since channel balance is not perfect, a spurious response at the unwanted sideband is always obtained, but is far below the desired signal level. A simple analysis based on true



Experimental version of the direct-conversion s.s.b. receiver. Although the article is not intended to present step-by-step construction data, sufficient information is provided so that an experimentally-minded builder can use the principles described by the author.

product detection in the mixers and input signal of the form $\cos(\omega_0 + \Delta\omega)t + \cos(\omega_0 - \Delta\omega)t$ will effectively show this sideband suppression property.⁴

Circuits

A receiver of this type was built for the 14-MHz. amateur band. Fig. 4 shows the block diagram of this receiver. Broadband balanced mixers of standard design were used in the front end. Hot-carrier diodes were used to minimize front-end noise. (See the earlier QST article.¹)

Signals are fed to the two mixers from the input tuner shown in Fig. 3. This tuner performs the dual function of providing front-end selectivity and dividing input power between the I and Q (In-phase and Quadrature) channels.

A Hewlett-Packard Model 606A r.f. signal generator was used for the local oscillator. Any generator capable of delivering about 0.5 volt r.m.s. to 50 ohms and covering 20 meters would do as well.

A lumped-constant 3-db. hybrid network for 14.25 MHz. provides the 90-degree r.f. phase shift (Fig. 5). This network holds a 90-degree phase shift (within the available limits of measurement) from 13.8 to 14.6 MHz. Variation in relative amplitudes between the two outputs over this frequency range is less than 0.8 db.

¹ From September, 1969, QST.

² Hayward and Bingham, "Direct Conversion — a Neglected Technique," QST, November, 1968, p. 15.

³ Norgaard, "Practical Single-Sideband Reception," QST, July 1948, p. 11. Also *Single Sideband for the Radio Amateur*, ARRL, p. 177 in 4th edition.

⁴ The 2Q4 phase shift network is available from Barker and Williamson, Canal St. and Beaver Dam Rd., Bristol, Pa. 19007

⁴ *Single Sideband Principles and Circuits*, Pappenfus, Bruene and Schoenike, McGraw-Hill, pub., 1964, p. 33.

¹ Ibid.

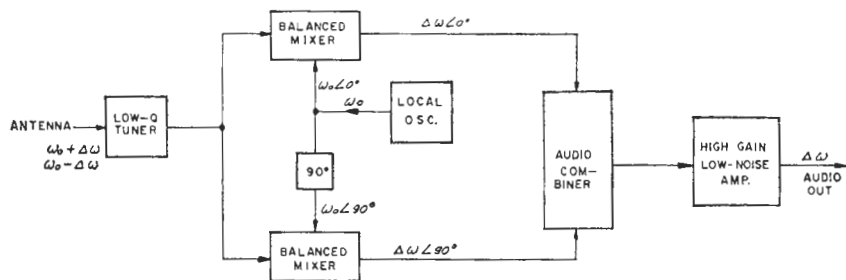


FIG. 1—Block diagram of a basic direct-conversion s.s.b. receiver.

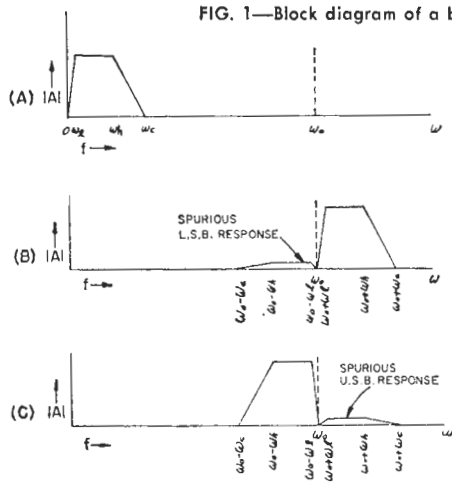


FIG. 2—Audio- to r.f.-bandpass transformation. If the audio response is that shown at A, the resulting responses of the direct-conversion s.s.b. receiver will be those shown at B for upper sideband reception, and at C for lower sideband reception.

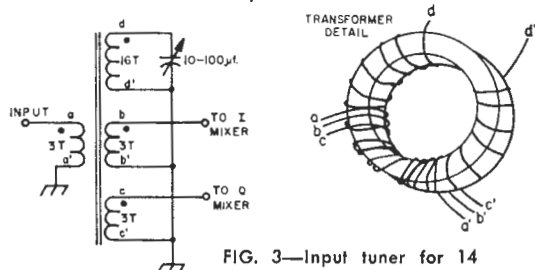


FIG. 3—Input tuner for 14 MHz. The transformer is wound as shown on an Indiana General CF-105 core (Q3 material).

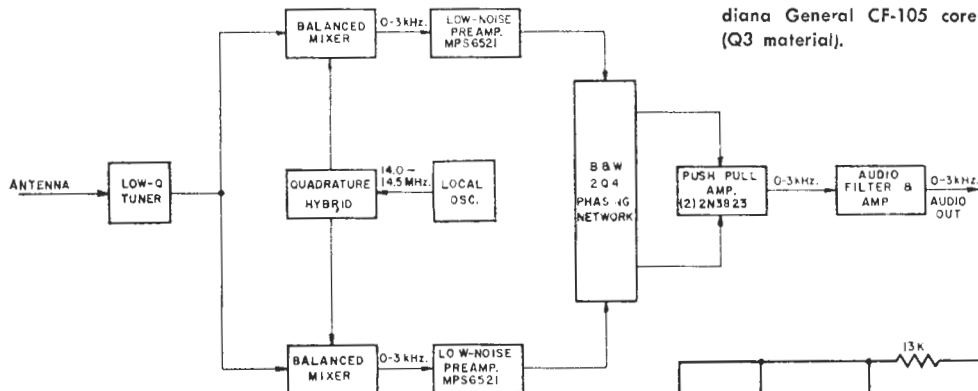


FIG. 4—Block diagram of the 14-MHz. direct-conversion s.s.b. receiver.

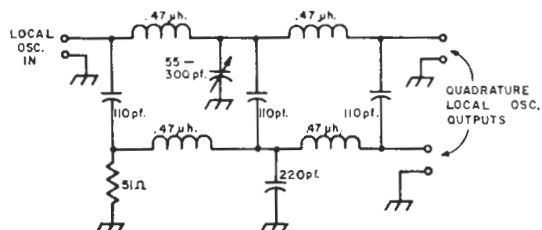


FIG. 5—14.25-MHz quadrature hybrid network. This hybrid holds a 90-degree phase shift from 13.8 to 14.6 MHz.

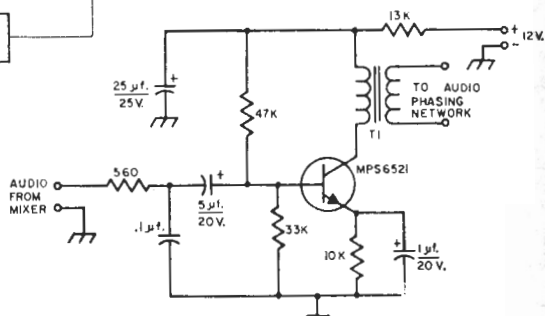


FIG. 6—Low-noise audio preamplifier. The I and Q channels are identical. Resistances are ohms, K = 1000. Capacitors with polarity indicated are electrolytic.

T₁—Audio transformer; primary 22,000 ohms, secondary 600 ohms. See text.

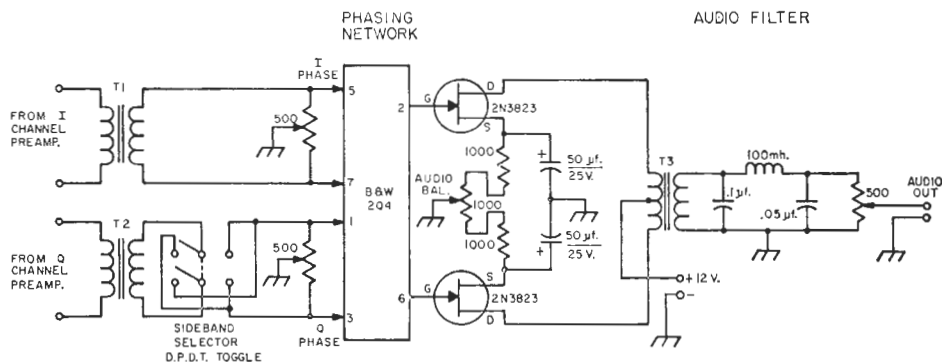


FIG. 7—Audio phasing network. See information referenced in footnote 2 for the setup procedure of a similar network. Resistances are ohms, K = 1000. Capacitors with polarity indicated are electrolytic.

T_1, T_2, T_3 —Audio transformers; primary 22,000 ohms, secondary 600 ohms. See text.

Such performance is adequate for a non-critical application. More stringent requirements would force one to use broadcasting techniques, discussed later.

Mixer outputs are fed to low-noise audio preamplifiers (Fig. 6). The MPS6521 transistor used in the preamplifiers is a plastic low-noise audio device selling for about \$1. With a 600-ohm source, a noise figure of about 2 db. exists at its chosen operating point. Forty db. of gain is obtained from this circuit at one kilohertz. An RC low-pass network provides initial r.f. filtering, and transformer coupling is used to match the audio phase-shift network impedance. All of the audio transformers used in the receiver are of the surplus variety often featured in tube-type sideband rigs. Any audio transformer of the correct turns ratio would do as well.

The preamplifier outputs drive the phasing network in Fig. 7. This network is similar to one described by K2OHF in his discussion of a phasing s.s.b. adapter.⁵ The B&W 2Q4 was designed to drive a grid load, which an FET approximates nicely. The 2N3823 FET is rather expensive but can be replaced by its plastic equivalent, the 2N3819, in this application. Some device shielding may be necessary with the 2N3819, however.

The push-pull amplifier configuration allows I and Q channel summation to be made in the output transformer.

Sufficient gain is provided in the preamplifiers to prevent noise-figure degradation due to loss in the audio phasing network (the network attenuation is considerable—on the order of 30 db.).

A 500-ohm pi-network audio filter follows the mixing amplifier. An upper cut-off frequency of 2.7 kHz. is used and an ultimate roll-off of 16 db. per octave is achieved.

The two-stage audio amplifier following the filter provides more than adequate headphone

volume with high-impedance phones (Fig. 8). The overall midband gain of the audio system is 100 db. The audio response is shown in Fig. 9. This response determines the receiver's selectivity. Low-end roll-off is due primarily to the transformer transfer characteristics.

Setup Adjustments

The receiver setup is similar to that required for a phasing s.s.b. exciter. The audio and r.f. phase shifts must be set to 90 degrees and the audio balance adjusted for minimum unwanted sideband. K2OHF gives a good description of his setup procedure—and much of his technique is applicable here.

After those adjustments are made, listen carefully to a c.w. signal as it is tuned through zero beat. The tone on one side of the carrier will be louder than the other. Retune to the weaker tone and adjust the receiver tuning for a note of about 1 kHz. Adjust the setup controls alternately to null this tone out. Retune to the other side of zero beat to verify enhancement of the other sideband.

As with any phasing rig, some iteration in adjustment must be expected and it is possible to null the wrong sideband and get less than optimum performance. Some patience is called for in making these adjustments. If a good null cannot be obtained, retune to the other sideband and start over again.

Performance

The receiver performs very well considering its simplicity. There is a feeling of "transparency" or "presence," as if the listener were hearing the signals directly rather than through a receiver. This is a result of the excellent linearity and low distortion inherent in the design.

SENSITIVITY

A signal of 0.2 μ V. r.m.s. at 14.2 MHz. is just audible in the headphones at the maximum gain setting. This condition is equivalent to a signal-to-noise ratio of about unity.

⁵ Buhrer, "An S.S.B. Product-Detector Adapter," *QST*, August, 1961, p. 22.

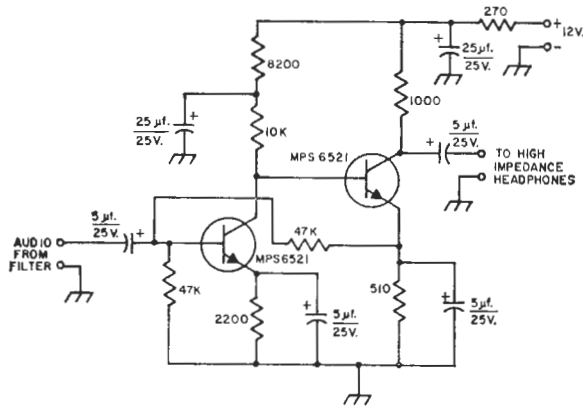


FIG. 8—Output audio amplifier. Resistances are ohms, K = 1000. All capacitors are electrolytic.

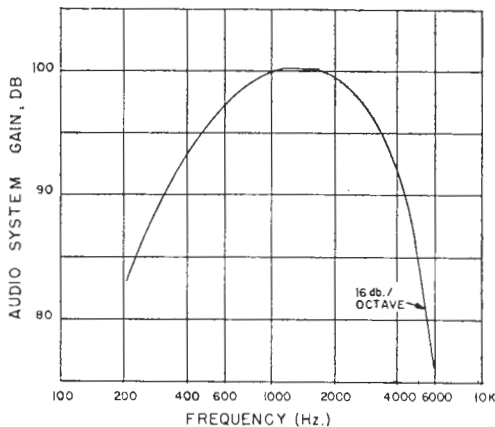


FIG. 9—S.s.b. receiver audio response. The roll-off at the lower frequencies occurs because of the characteristics of the audio transformers used.

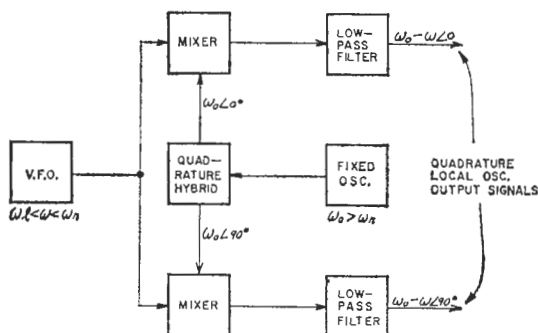


FIG. 10—Broadband quadrature local oscillator.

SELECTIVITY

Selectivity is more than adequate, even if a less-than-optimum audio filter is used. No problems were experienced in reading adjacent weak and strong s.s.b. signals on 20 meters. The nature of the s.s.b. signal helps here, of course, because it is generally limited to a 3-kHz. bandwidth. The adjacent strong signal appears as audio components between 3 and 6 kHz. on which the mind can do some filtering.

On c.w., a narrower or more peaked band-pass would be helpful. Even without it, however, copy was generally good—particularly where the interfering station could be put on the suppressed side of the input spectrum.

SIDEBAND SUPPRESSION

Unwanted sideband suppression is greater than 34 db on the 14.0 to 14.5-MHz. frequency range.

CROSS MODULATION

A 10-μv. signal was placed in the passband. An adjustable amplitude signal was placed 10 kHz. above it (same sideband). No cross modulation was noted until the interfering signal was at 0.2-volt r.m.s. amplitude.

Broadbanding

It is only the quadrature hybrid network that limits the receiver bandwidth or effective frequency range of the receiver. The local oscillator can be broadbanded using the scheme shown in Fig. 10. Two mixers are used in a fashion analogous to the front end of the receiver. A fixed-frequency oscillator drives two mixers. A 90-degree phase shift is provided in the feed to one mixer. The difference products ($\omega_0 - \Delta\omega$) are 90 degrees out of phase at the outputs and are the desired local-oscillator signals. A low-pass filter eliminates the sum component.

By using balanced mixers and broadband transformers, it would seem possible to build a local-oscillator system suitable for the entire high-frequency range (3–30 MHz.) without switching more than the v.f.o.

There are also available now a number of commercial broadband quadrature hybrids covering octave bandwidths. Any of these would broadband the receiver without the need for additional mixing.

Transceive Operation

An s.s.b. output at the signal frequency can be provided by reversing the direction of the audio flow and applying phased audio at the output terminals of the mixers. No r.f. switching is required.

Linear amplification is required after this point, of course, as no more than 10 mw. of r.f. would be available at the mixed output.

Direct conversion is definitely a promising technique. It would be interesting to hear the experiences of others doing similar work.

» As communications receivers go, this one is reasonably simple and straightforward. It combines some of the best features of previous designs, including a high-frequency crystal filter for s.s.b. selectivity, an audio filter for c.w. selectivity, a beam-deflection mixer, dual detectors, audio-derived a.g.c., and a temperature-compensated v.f.o.

The W50MX Communications Receiver

COL. DAVE CURTIS, W50MX



The W50MX receiver. Main tuning dial (J. W. Miller, MD-7) has both 6:1 and 36:1 tuning ratios. A 6:1 planetary drive assembly (Arrow Electronics, Type 4511) is used on the preselector tuning capacitor. The pointer is homemade. (Photos by A/3C Michael M. Steinbacher)

Long-time *QST* readers will recall W1DX's excellent article on receiver design in the January, 1957 issue.¹ At the time it appeared, the article was studied with great interest. Particularly, the point that selectivity belongs as close to the antenna as possible seemed to make a great deal of sense. With the appearance of high-frequency filters at reasonable prices, the author initiated the design of a receiver to utilize this principle. For various reasons, however, this receiver never got beyond the block-diagram stage. A more recent article by W1DX,² which was illustrated with an operational piece of hardware, provided the final push. Serious design and construction followed, and the "W50MX" receiver, described here, is the result. It is a spectacular performer.

Unfortunately, the author's shack is not equipped with test gear adequate to permit performance measurement. Consequently, resort had to be made to subjective comparison, and the opinions of fellow hams. These judgments suggest that the double-conversion re-

ceiver, utilizing a low-frequency second i.f. to obtain selectivity, may be on the way out. The author's second receiver—a 16-tube double-conversion job of sound design—simply cannot compete. In side-by-side tests, using a common antenna, the contrast is remarkable. The new receiver's performance is characterized by a clarity in signal quality, the result of a markedly lower overall noise level. Signals masked to unintelligibility by noise in the older receiver become readable copy. In conditions of reasonably low atmospheric noise, signals appear to pop out of surrounding quiet.

Performance

In more useful specifics, here is how the receiver stacks up:

Sensitivity: Very FB. Digs right down to the noise level on all bands, 80 through 10 meters. The receiver has made possible R5 copy of both ends of a W6/W2 QSO on 40, and of a KL/W4 QSO on 20, using only a finger touching the input connector as an antenna!

Stability: Truly marvelous. From a cold (room-temperature) start, drift is inconsequential after a 15-minute warm-up. Further,

¹ From January, 1968, *QST*.

² Goodman, "What's Wrong with Our Present Receivers?" *QST*, January, 1957.

³ Goodman, "Some Thoughts on Home Receiver Design," *QST*, May, 1965.

the switching arrangement permits leaving the filaments on continuously. When this is done, and heat soaking has occurred, there is no apparent drift after the mode switch is turned to the appropriate "on" position. If there is any drift, it is the other guy!

Selectivity: About right for s.s.b. Gives good single-signal selectivity on c.w.

Mechanical: Can take sharp raps with no noticeable frequency shift.

Birdies: A few. There are one or two of consequence on each band segment, except on 15 meters where there are six (by actual count). These tune sharply, and seldom bother reception. Nevertheless, this is a basic design deficiency which, perhaps, could be overcome by someone who is mathematically inclined and who can select conversion frequencies more intelligently.

A.g.c.: The circuit suggested by W1DX² is the best we have seen. S.s.b. signals ranging from S2 or 3 to 10 over 9 come out of the speaker at quite reasonably similar levels. This is one a.g.c. that will be used most of the time.

Circuit Outline

Interested? Let's have a look at the schematic of Fig. 1. As far as the signal is concerned, this is a single-conversion receiver. The incoming signal is amplified in the single r.f. stage using the pentode section of a 6AZ8. It is then converted to an i.f. of 9 Mc. in a 7360 mixer. A band 2.8 kc. wide is sliced out by a steep-skirted crystal filter, FL_1 . The signal is then amplified through three i.f. stages using 6BA6s, and finally detected by an infinite-impedance detector, V_{3B} , if a.m., or by a 6BY6 product detector, if s.s.b. or c.w. The otherwise conventional audio system includes a selective filter for c.w. work. The a.g.c. system is audio-derived.

The main tuning element is the v.f.o., covering 5 to 5.5 Mc. Bands are changed by altering the frequency of local injection to the signal mixer. This is accomplished by heterodyning signals from the v.f.o. and from the crystal oscillator V_{2A} to produce the required

injection frequency in the output of the heterodyne mixer, V_{2B} . A 3.5-Mc. crystal oscillator, using the triode section of the 6AZ8, provides markers for the low-frequency edges of the bands covered.

The V.F.O.

The v.f.o. is a 6AU6 in a very high-C Colpitts configuration. A differential capacitor, C_8 , in combination with NP0 and N750 fixed capacitors, permits simple and accurate adjustment of temperature compensation. With reasonable attention to mechanical design, and careful adjustment, stability is impressive indeed. This circuit was used in an earlier project,³ and was found to provide stability comparable to that of the BC-221 frequency meter. No small part of the stability is due to the use of the rugged low-torque Miller tuning capacitor.

R.F. Stage and Crystal Calibrator

Air-wound coils are used in the preselector. The gain in this stage appears to be approximately 12 to 15 db. on 80 and 40, dropping off to about 6 to 8 db. on 15 and 10. It does a good job of rejecting i.f. images (none have been found). With some antennas, the gain of this stage may have to be reduced slightly to prevent oscillation on the 80-meter band; on other bands the amplifier is perfectly stable at full gain. Input and output circuits are gang-tuned. Ceramic trimmer C_1 (one for each input coil) is used to adjust tracking.

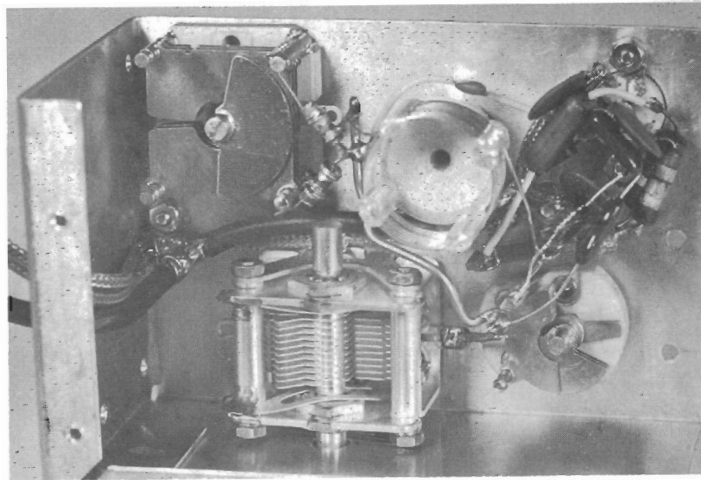
The triode section of the 6AZ8A, V_{1B} , is used in the crystal calibrator. The frequency can be "zeroed in" against a calibrating source by means of C_4 . Notice that the 15-meter band and all ranges of the 10-meter band are covered with a single set of preselector coils.

Signal Mixer

The 7360 performs the mixing function effectively, and contributes inconsequential noise. It does not appear to overload on even the very strongest signals. The mixer gain con-

³ Curtis, "The W4JWV Single-Sideband Exciter," *QST*, January, 1963.

V.f.o. assembly with side-top cover removed. The 6AU6 socket and associated components are at upper right with the band-set capacitor C_7 at lower right. The coil is glued securely to a ceramic standoff insulator. The differential capacitor, C_8 , with temperature-compensating capacitors C_9 and C_{10} attached, is at upper left. Note that all major components and tie points are fastened securely to the same side of the enclosure for maximum mechanical integrity. When mounted on the chassis the right-hand end of the box in this view is at the top, the left-hand end is bolted to the chassis.



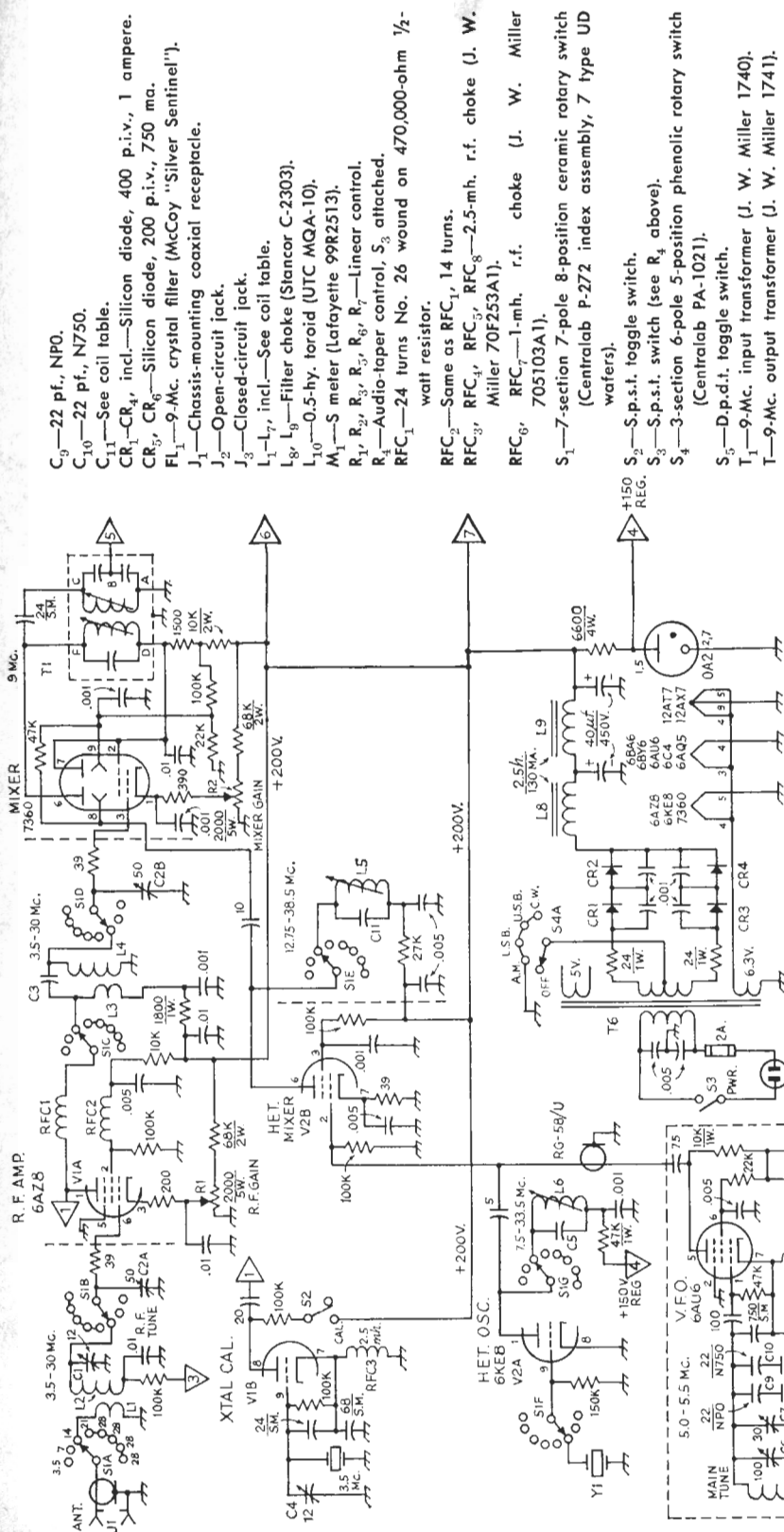
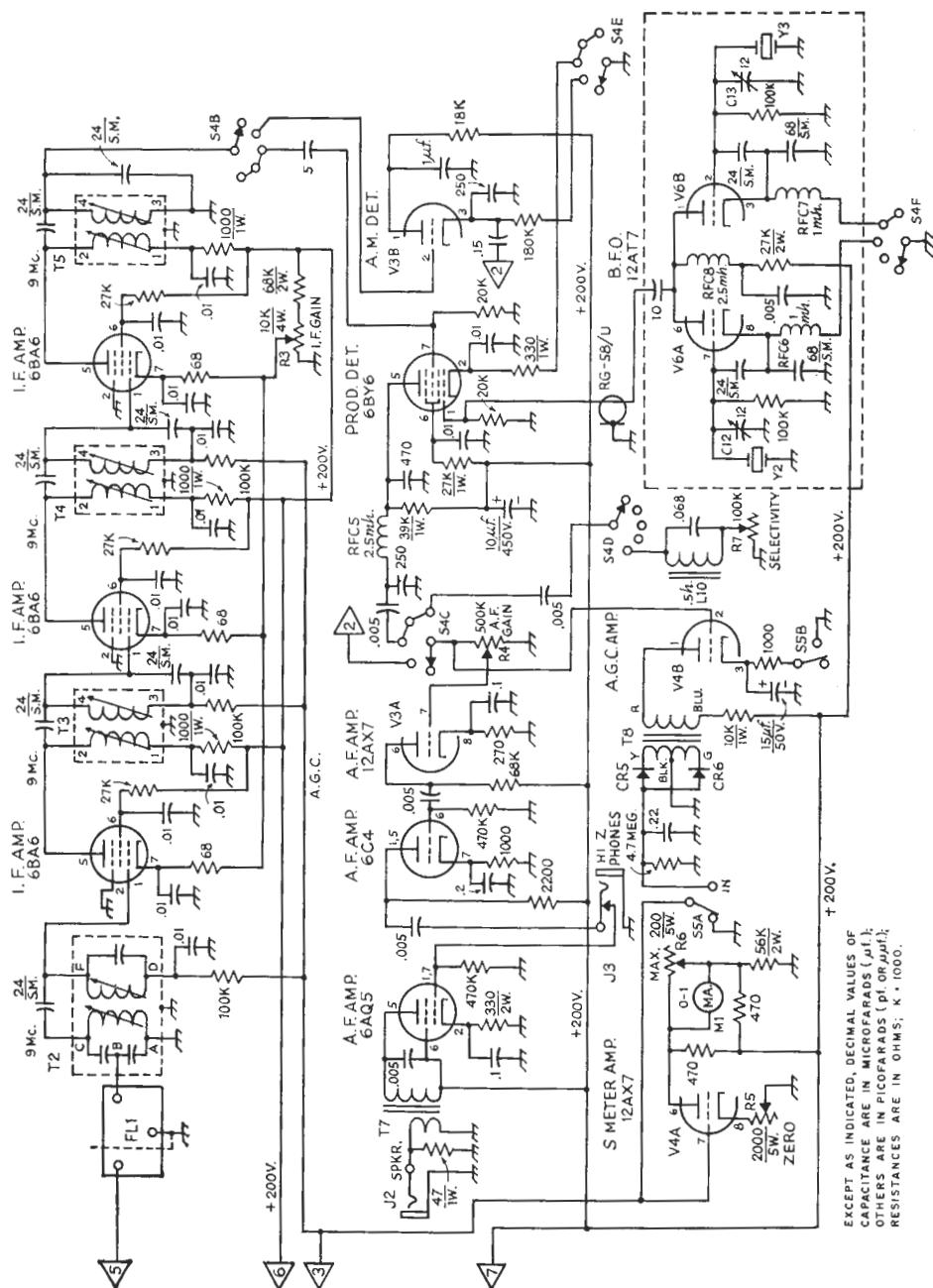


FIG. 1—Circuit of the W5OMX receiver. Unless indicated otherwise, fixed resistors are $\frac{1}{2}$ -watt. Unless identified as silver mica (S.M.) capacitors less than 0.001 μ f. in value are mica or stable ceramic; others are paper, mylar, or disk ceramic, except that polarized capacitors are electrolytic.

C₁—12-pf. ceramic trimmer (one for each L₃ coil).
 C₂—Dual-section air variable, approx. 50 pf. per section (Hammarlund HFD-140 used, with all but 7 rotor plates removed from each section).
 C₃—See coil table.

C₄, C₁₂, C₁₃—Approx. 12-pf. Compression trimmer.
 C₅—See coil table.
 C₆—100-pf. variable (J. W. Miller 2101).
 C₇—30-pf. air trimmer.
 C₈—27-pf. differential capacitor (Johnson 167-32).

C₉—22 pf., NPO.
 C₁₀—22 pf., N750.
 C₁₁—See coil table.
 CR₁—CR₄, incl.—Silicon diode, 400 p.i.v., 1 ampere.
 CR₅, CR₆—Silicon diode, 200 p.i.v., 750 ma.
 FL₁—9-Mc. crystal filter (McCo "Silver Sentinel").
 J₁—Chassis-mounting coaxial receptacle.
 J₂—Open-circuit jack.
 J₃—Closed-circuit jack.
 L₁—L₇, incl.—See coil table.
 L₈, L₉—Filter choke (Stancor C-2303).
 L₁₀—0.5-hy. toroid (UTC MQA-10).
 M₁—S meter (Lafayette 99R2513).
 R₁, R₂, R₃, R₄, R₅, R₆, R₇—Linear control.
 R₈—Audio-taper control, S₃ attached.
 RFC₁—24 turns No. 26 wound on 470,000-ohm $\frac{1}{2}$ -watt resistor.
 RFC₂—Same as RFC₁, 14 turns.
 RFC₃, RFC₄, RFC₅, RFC₆—2.5-mh. r.f. choke (J. W. Miller 70F253A1).
 RFC₇—1-mh. r.f. choke (J. W. Miller 70S103A1).
 S₁—7-section 7-pole 8-position ceramic rotary switch (Centralab P-272 index assembly, 7 type UD wafers).
 S₂—S.p.s.t. toggle switch.
 S₃—S.p.s.t. switch (see R₄ above).
 S₄—3-section 6-pole 5-position phenolic rotary switch (Centralab PA-1021).
 S₅—D.p.d.t. toggle switch.
 T₁—9-Mc. input transformer (J. W. Miller 1740).
 T₂—9-Mc. output transformer (J. W. Miller 1741).
 T₃, T₄, T₅—10.7-Mc. interstage transformer (J. W. Miller 1457). Mount with spade bolts.
 T₆—Power transformer: 550 volts r.m.s., center-tapped, 110 ma.; 6.3 volts 5 amp. (Triad R-112 A, 5-volt winding not used).
 T₇—2-watt audio output transformer (5000 ohms to voice coil) (Burstin-Applebee 178393).
 T₈—Transistor audio input transformer, 5000 ohms to 7500 ohms, center-tapped (Lafayette-Argonne AR-154).
 Y₁—See coil table.
 Y₂—9001.5 kc. (see text).
 Y₃—8998.5 kc. (see text).



trol, R_g , is used to prevent oscillation on 80 meters, and to adjust the overall gain on the other bands. By adjusting the gain at this point, the high-gain i.f. strip may be operated at full amplification at all times for optimum a.g.c. action.

I.F. Amplifier

Since selectivity is provided ahead of the i.f. strip, these stages are designed purely for amplification. The 24-pf. capacitors across the hot ends of the i.f. transformers increase the overall gain spectacularly. A 0.2-volt signal at 9 Mc. injected into this strip ahead of the crystal filter comes out at a whopping 20 to 25 volts. This accounts in a large measure for the rather impressive overall sensitivity of the receiver. The i.f. gain control, R_g , is used only during initial adjustment and testing; therefore it is not mounted on the panel, but on the rear apron of the chassis.

Detectors

The 6BY6 product detector, developed by W6TC for his very efficient HBR receivers,¹ works well at 9 Mc. This circuit has the very desirable feature of accepting a wide range of signal levels with little or no apparent distortion in the audio product. The infinite-impedance detector provides these same advantages in a.m. reception, without overloading the last i.f. transformer as would a diode.

B.F.O.

The b.f.o. uses the two triode sections of a 12AT7 as separate crystal oscillators. The

¹ "Hints & Kinks," QST, June, 1962.

crystals at 9001.5 and 8998.5 kc. (supplied by McCoy with the filter), permit selection of lower and upper sidebands, respectively, by keying the appropriate 12AT7 cathode. These crystals are adjusted to proper frequency by trimmers C_{12} and C_{13} .

Audio Section

Three stages of audio provide generous output to high-impedance phones or a speaker. You can hear signals on this receiver over the QRN of all but the noisiest "harmonics"! In the c.w. mode, a high-Q audio filter, composed of toroid L_{10} and its related capacitor, permits peaking the beat note at approximately 1000 cycles. Substitution of a different value of capacitance will move the resonant frequency to your choice of pitch. Selectivity may be varied by adjustment of R_7 .

A.G.C.

The a.g.c. circuit amplifies and full-wave rectifies audio from either detector, and controls the r.f. amplifier and all three i.f. stages. It is remarkably effective, and makes the multi-party s.s.b. ragchew a real pleasure. (Those who enjoy fiddling with knobs probably won't like it!) The fast-attack/slow-decay characteristics which result from the component values suggested by W1DX have proven to be very close to the ideal.

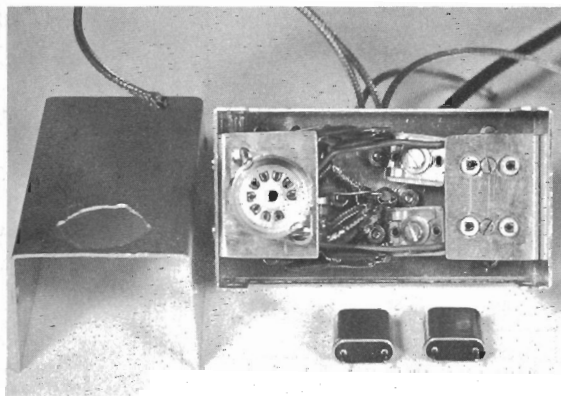
The S-meter and power-supply circuits should be familiar to most readers. S-meter adjustments are made at the rear of the chassis. Silicon rectifiers are used in the power supply, and a voltage-regulated tap supplies the v.f.o. and heterodyne oscillator.

Muting

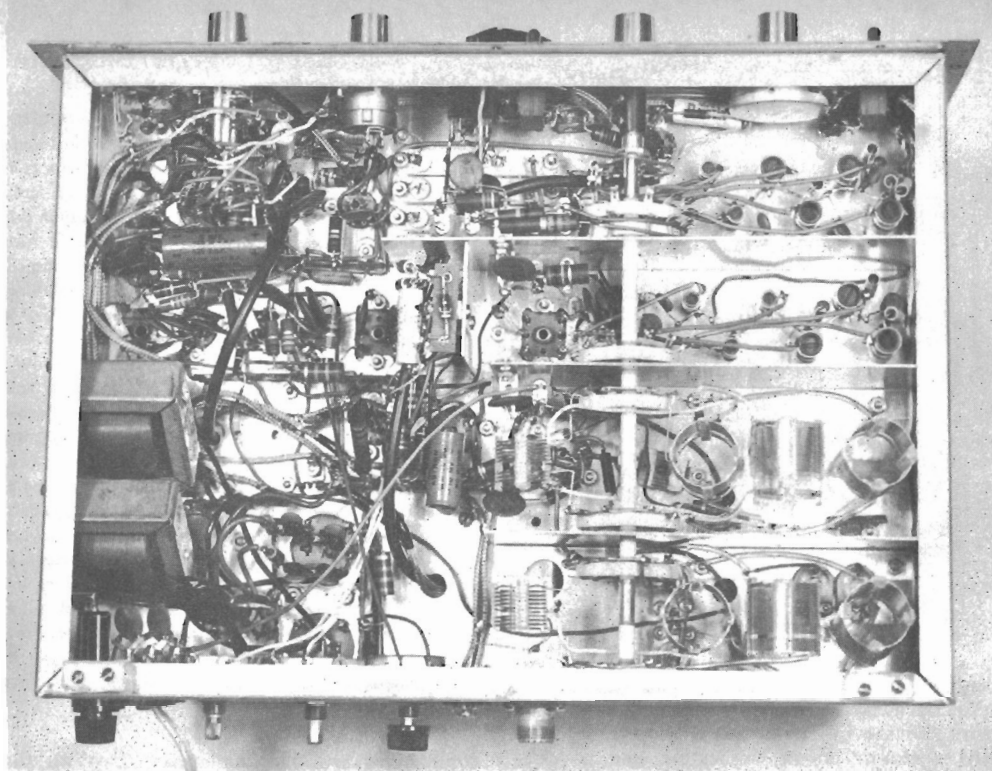
You will note that no provision for muting is indicated in the circuit schematic. Three possible arrangements are suggested. Your choice should be based upon how you intend to connect the receiver into the overall station setup. If you intend to monitor your transmissions on the receiver, and use an antenna relay that grounds the receiver input on transmit, break the plus B or cathode connection of V_{1A} , and insert the muting switch and remote connections at this point. If you have side-tone monitoring, you can cut off the receiver entirely by breaking the plus B or cathode connection of the 7360 mixer, and inserting the muting connections at that point. Finally, and perhaps the best of all, although additional components will be required, use the muting arrangement suggested by W1DX.²

Construction

Viewed in its entirety, the construction of a receiver of this complexity may appear to be an overwhelming task. Certainly, it would be a very ambitious first project. However, for anyone with sufficient experience and skill to do the minor fabrication and locate sensibly



B.f.o. assembly with tube, cover and crystals removed. The sockets for the crystals and the 12AT7 are mounted on small aluminum brackets, the small components underneath being wired prior to final assembly. The crystals are plugged in internally and require no clearance holes in the cover. Crystals trimmers C_{12} and C_{13} are fastened to the bottom of the Minibox enclosure, at the center. The shielded leads and output coax cable leave the enclosure through tightly-fitting holes to minimize r.f. leakage.



Bottom view showing band switch and coil compartments. The pair of close-spaced wafers at the top switch the heterodyne-oscillator coils and crystals. The single wafer below switches the heterodyne-mixer coils. The signal-mixer coils are in the next compartment, switched by the pair of widely-spaced wafers. R.f.-stage coils are in the bottom compartment; one of the two switching wafers is hidden by the lip of the chassis. The 40- and 80-meter air-wound coils are cemented to platforms made of polystyrene sheet. The higher-frequency coils are supported on switch terminals. The mode switch is in the upper left-hand corner, filter chokes in the lower left-hand corner. A.g.c. components are mounted on tie points on the short vertical shield near the center.

the many small components, it should be a feasible undertaking. The primary ingredients of successful homebrew construction seem to be patience, a willingness to take one step at a time, and the interest to keep going. If you have these talents, you can probably build a receiver of the same superlative performance as the one described. And it should be better looking; this one is the final result of many, many component substitutions in the search for optimum performance.⁵

The following paragraphs contain construction and alignment suggestions, roughly in the order followed by the author. Additional information may be obtained by a careful study of the several illustrations and accompanying explanatory captions.

The receiver is built on a 10 × 14 × 3-inch aluminum chassis which fits into the 11 × 15 × 9-inch cabinet (Wyco CR-7725). An additional 10 × 17 × 3-inch chassis (the smallest size obtainable made from 16-gauge stock) was purchased as a source of material for the v.f.o. enclosure and shielding partitions.

⁵ To assist those who wish to duplicate this project, the author will provide full-size templates for chassis and front panel, an enlarged schematic, complete parts list, and 8 × 10-inch enlargements of the four primary illustrations, at a cost of \$4.50, postpaid.

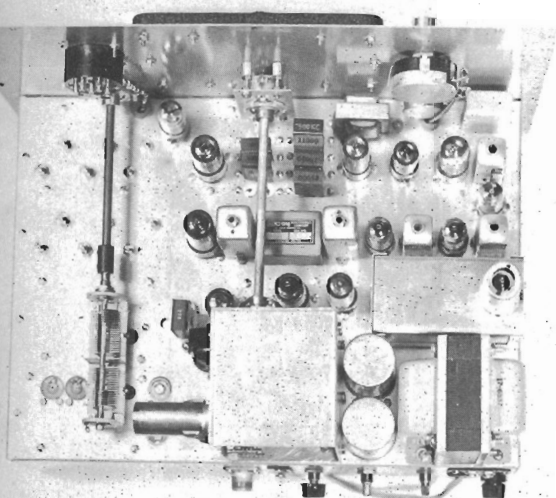
Assembling the V.F.O. and B.F.O.

Make the v.f.o. first. The main part of the enclosure was made from a corner of the spare chassis. Its dimensions are 4½ by 3¾ by 3 inches. The side/top cover was cut from adjacent spare chassis material. (The rear of the box is left open.) Mount the major components all on one side of the box, as shown in the detail photo, to minimize frequency changes with mechanical stress. Care should be used in locating the tuning capacitor so that its extended shaft will be high enough above the chassis to clear the McCoy filter in the i.f. strip (see top-chassis photo), and yet not be so high that it will crowd the dial too close to the upper edge of the panel.

The b.f.o. components are assembled in a 1½ × 2½ × 4-inch Minibox. Construction is detailed in one of the photographs.

Band Switch

Before starting to lay out the component pattern on the chassis, the under-chassis shields should be cut, using material from the aprons of the spare chassis. The longer shield has a length of 8¼ inches; the other two are 7 inches long. Then they should be placed temporarily in the chassis while their positions



are adjusted. Space them apart suitably to provide adequate room for the coils, and measure the spacing accurately. Make a mark on the rearmost shield, indicating the distance that the switch shaft will be placed from the end of the chassis. Mark and drill the switch-shaft and mounting holes in the three partitions, using extreme care to see that they are as identically located as possible. Make the holes reasonably oversized. Then assemble the switch and shields as a unit, using spacers on the switch assembly rods to obtain the partition spacings measured earlier. Do not tighten the assembly nuts more than finger tight. Place the assembly in the chassis, and press down firmly on the shields while the assembly nuts are tightened. Spot the shield mounting holes, remove the assembly, and drill the holes. Avoid any mounting holes in the area that will be occupied by the v.f.o. box, since this box must rest flat on the chassis. (The b.f.o. assembly can be raised on spacers to clear any mounting screws in its area.) Additional holes that should be drilled in the shields are one in each of the shields, below and to the left (in the bottom view) of the switch wafers (for wires), one in the upper left-hand corner of the second shield, and another in the same relative position in the first shield (for tie-point strips). A $\frac{3}{8}$ -inch hole should be drilled in the first shield, to the left of the short vertical shield. This will be used to pass the coax feed line from the v.f.o. to the heterodyne mixer, and some of the power leads. The corners of the partitions that rest in the fold of the chassis should be cut off to allow passage of wiring between the panel and the rear of the chassis.

Chassis Layout

Once the shield locations have been determined, the positions of the two main rows of

Top chassis view of the W5OMX receiver. Mounted in two groups in the upper left-hand corner of the chassis are the slug-tuned coils L_6 (top) and L_5 (below). In the lower left-hand corner are the preselector tuning capacitor and the C_1 tracking trimmers. The i.f. strip runs across the center with components in logical order, starting with the 7360 mixer, and turns vertically at T_4 , ending at T_5 . Proceeding to the left from T_5 are the two detector tubes, the 6AQ5 audio output tube, the heterodyne crystals, the 6KE8 and the 6C4 (above). The audio output transformer and c.w. filter toroid are to either side of the mixer-gain (top) and c.w.-selectivity controls. Occupying the lower right-hand corner of the chassis are the v.f.o. and b.f.o. units, and power-supply components. Immediately above the v.f.o. compartment are the calibrator crystal (with hole for access to trimmer C_4 just to the left), the 6AZ8, the 12AX7 a.g.c. tube (V_4), and the VR tube. Along the rear apron are the antenna connector, speaker terminals, i.f. gain control, S-meter controls, muting terminals, and fuse holder.

components will become apparent. With the v.f.o. subassembly placed with its rear edge flush with the rear edge of the chassis, and the shaft of the tuning capacitor central on the chassis, the location of surrounding components can be spotted. In locating the preselector tuning capacitor, place it far enough toward the edge of the chassis to assure space for its dial on the panel.

After all hole centers have been marked and hit with a center punch, the various holes may be drilled or cut. The author used a nibbler to cut the i.f.-transformer holes to approximate size, and finished up with a file.

Before mounting any components on the chassis, fasten the panel temporarily in place, and place the shafts of the v.f.o. and preselector tuning capacitors against the back of the panel while you mark the shaft heights.

Wiring

Tie-point strips should be located liberally on the underside of the chassis, convenient to tube sockets and related components. It is advisable also to place grounding solder lugs on most of the mounting screws. You may not use all of them, but it is much more convenient to install them as you mount the components, than later on when space becomes scarce as the wiring progresses.

Power-supply and filament connections should be made first. Thereafter, the wiring procedure is not particularly critical. Installation of the preselector coils can be left as a last operation, after the v.f.o. and b.f.o. circuits have been adjusted. To make sure that no connection is overlooked, it is a good idea to mark the schematic with a colored pencil as each connection is completed. The author

wired the front circuits first, working toward the rear of the chassis. Following standard practice, long leads, particularly those connecting front-panel controls and switches to components at the rear of the chassis, may be made with shielded wire. This practice permits fastening the leads solidly in place by soldering the shield to conveniently-located soldering lugs along the way. Shielded wire should also be used for all a.f. grid leads to avoid unpleasant feedback problems. R.f. bypass-capacitor leads should be as short as possible, using the center post of the related tube socket as a common grounding point.

Testing the V.F.O. and B.F.O.

The v.f.o. tuning range should be checked first with all tubes except the v.f.o. voltage-regulator tube out of their sockets. After power has been turned on and the v.f.o. allowed to warm up, a v.t.v.m. with an r.f. probe should show about 2 volts at the output coupling capacitor. The v.f.o. frequency can be checked by comparing it with the signal from a calibrated source, such as a BC-221 frequency meter, or a general-coverage receiver. Set C_5 at about midpoint. Set the tuning capacitor C_6 at about 3 degrees from maximum capacitance. Then adjust C_7 to bring the frequency to 5.0 Mc. Turn C_6 to about 3 degrees from minimum capacitance, and check the frequency again. If the frequency is higher than 5.5 Mc., spread the end turns of the coil apart, and repeat the process. If the frequency is too low, squeeze a few of the turns slightly closer together, and repeat the process. It should be possible to arrive at an adjustment where the 5-to-5.5 Mc. band occupies about 95 per cent of the dial, with the band central on the dial.

Plug in the b.f.o. tube and check the r.f. output voltage. It should be about the same as from the v.f.o., i.e., 2 volts.

Checking the Audio Section

Plug in the audio tubes. With speaker or headphones connected, and the a.f. gain control near maximum, a sharp click, when the top end of the gain control is touched with the lead of a pencil, will tell you that the audio stages are working.

I.F. Alignment

Plug in the 7360 mixer and i.f. tubes. Connect the r.f. probe at the arm of S_{1B} . Introduce a 9-Mc. signal at the input to the last i.f. stage. The author used the crystal calibrator as the source, with a 9-Mc. crystal, borrowed from his s.s.b. exciter, plugged into the calibrator. The 20-pf. calibrator coupling capacitor was temporarily disconnected from Pin 1 of the 6AZ8, and connected by means of an extension lead to Pin 1 of the last 6BA6 i.f. tube. (A reasonably-accurately-calibrated r.f. signal generator may be used, if available.)

Tune T_5 for maximum output. Move the signal source to Pin 1 of the second i.f. tube, and adjust T_4 . Do the same with the first i.f. tube and T_3 . You will probably have to reduce the i.f. gain as you move down the i.f. strip to avoid burning out the diode in the probe. Introduce the signal at the output connection of the crystal filter, and adjust T_2 . Finally, inject the signal at Pin 3 of the 7360 mixer, and adjust T_1 . (If you are using an r.f. signal generator, you may have to jockey the frequency slightly to hit the center of the crystal-filter passband.) Reconnect the calibrator coupling capacitor to the plate of the 6AZ8.

S-Meter Adjustment

The next step is to adjust the S-meter circuit, since it will be used in adjusting the preselector. With V_4 out of its socket, adjust R_6 for full-scale S-meter reading. Plug in V_4 . Allow the tube to warm up and, with the a.g.c. switch off, adjust R_5 for a zero reading.

Heterodyne Tuning

Now plug in the 6KE8, and adjust each slug-tuned coil (L_6) for approximately 3 to 4 volts as measured with the r.f. probe at the "hot" end of the coil. The lower-frequency crystals are capable of producing much more than 4 volts; the higher-frequency crystals may not provide quite 4 volts. Tune for all you can get up to a maximum of 4 volts.

Using a grid-dip oscillator, tune the heterodyne-mixer coils (L_5) to the frequencies listed in the coil table. Be sure that the band switch is set to the band corresponding to the coil you are checking, because the stray capacitance may vary with the switch position.

Preselector Alignment

Alignment of the preselector coils can now be undertaken. The author built the preselector coils for 80 meters first, and aligned the front end on this band before proceeding to the higher-frequency bands, in order. However, it need not be done this way. The alignment procedure is the same for all bands. The important consideration in making the coils is to keep L_2 and L_3 as nearly identical as possible, including lead length and proximity to chassis and shields.

With a set of coils in place, introduce a signal near band center at the antenna connector. Set the v.f.o. to mid scale, and the mode switch to one of the sideband positions. Adjust C_2 , and the slug of L_3 for maximum S-meter reading. Then tune the preselector slowly across the signal. If the signal peaks at two dial settings, it means that the circuits are not tracking. By cautious adjustment of C_1 , and the turn spacing of either L_2 or L_4 , a condition should be found where only a single S-meter peak occurs as C_2 is tuned

COIL TABLE

<i>L</i> ₂ , <i>L</i> ₄									
Band	Turns	Wire Size	Diam. In.	T.P.I.	<i>L</i> ₁ Turns	<i>L</i> ₁ / <i>L</i> ₂ Space	<i>L</i> ₃ Turns	<i>L</i> ₃ / <i>L</i> ₄ Space	<i>C</i> ₃ pf.
80	50	24	1	32	6	2 t.	6	10 t.	None
40	22	24	1	32	6	2 t.	3	10 t.	None
20	12	20	1	16	4½	1 t.	3	6 t.	5
10-15	6	20	¾	16	3	1 t.	3	2 t.	5

<i>L</i> ₅					<i>L</i> ₆				
Band	Freq. Mc.	<i>L</i> μh. (Nom.)	Type	<i>Y</i> ₁ Mc.	Freq. Mc.	<i>L</i> μh. (Nom.)	Type	<i>C</i> ₅ pf.	<i>C</i> ₁₁ pf.
80	12.75	3.3	21A336	7.5	7.5	6.8	21A686	45	25
40	16.25	2.2	21A226	11.0	11.0	3.3	21A336	30	20
20	23.25	1.5	21A156	18.0	18.0	2.2	21A226	20	10
15	30.25	1	20A106	25.0	25.0	1.5	21A156	None	None
	37.5	0.82	20A827	32.00 33.50	32.25	1	20A106	None	None
10	38.5	0.82	20A827	33.0 33.5	33.25	1	20A106	None	None

*L*₇—8 turns No. 20, 1-inch diam., 16 t.p.i.

*L*₁/*L*₂, and *L*₃/*L*₄ (as well as *L*₇) are of Miniductor, Air Dux, or Polycoid stock, with the indicated number of turns removed to provide spacing between the main coils and the coupling links.

*L*₅ and *L*₆ are iron-slug coils (phenolic). Type numbers are J. W. Miller (suffix RBI). Those with prefix 20 are ¼-inch diam.; prefix 21 indicates ⅜-inch diam.

across the signal. (The paragraphs on r.f. alignment in the "Receiving Systems" chapter of the ARRL *Handbook* explain how this is done.)

Temperature Compensation

To adjust the v.f.o. temperature compensation, the most stable frequency source you can get is required. The crystal calibrator will do nicely. Allow the receiver to warm up thoroughly; leave it on for at least an hour or two. Tune the receiver to zero beat with the calibrator. Then, as drift occurs, adjust *C*₈ slightly, and bring the receiver back to zero beat with *C*₇. Continue to do this until no drift is apparent.

B.F.O. Adjustment

Remove the cover of the b.f.o. enclosure, and adjust trimmers *C*₁₂ and *C*₁₃ for optimum s.s.b. reception. Most 80- and 40-meter stations use l.s.b., while those operating in the higher bands use u.s.b. (Most c.w. operators prefer the u.s.b. position.) The b.f.o. frequency is adjusted so that it falls only high enough on the filter slope to assure adequate low-frequency response. With this adjustment,

the "other side" of a c.w. signal simply is not there.

V.F.O. Calibration

After checking to make sure that the 5-to-5.5-Mc. band is still centered on the dial, the dial may be calibrated (0 to 500, and 500 to 0) against a standard, such as a BC-221 frequency meter. The tuning should be found to be close to linear. A single calibration for all bands requires the exact crystal frequencies listed in the Table. Crystals not too far off on the high side can be "rubbered in" with a small compression trimmer in parallel with the crystal. Crystals on the low side must be ground or etched in. (The 3.5-Mc. band-edge marker will provide a reference.) Otherwise, *C*₇ in the v.f.o. will have to be retrimmed each time bands are changed, zeroing the v.f.o. against the calibrator with the v.f.o. dial set at the previously-calibrated zero mark.

Before placing the receiver in the cabinet, punch 4 or 5 holes through the bottom, and along the top back of the cabinet for air circulation.

You should now be able to make R5 copy of signals that your ham friend down the block may not be able to hear. Congratulations!

» Described here is a totally solid-state instrument designed to have wide dynamic range, freedom from spurious responses, and good sensitivity and selectivity. While such features are certainly not uncommon in modern-day receivers, the author's method of received-frequency conversion is a novel one for the home constructor! The high-frequency local oscillator, using a frequency-synthesis technique, is phase locked to a linear master oscillator operating at a much lower frequency. This approach affords, with single conversion, the stability and band-spread advantages of a dual- or triple-conversion receiver having a crystal-controlled first oscillator and tunable first i-f, while the possibility of "birdies" from mixing products is greatly reduced.

An Engineer's Ham-Band Receiver

RUDOLF FISCHER, DL6WD



Front view of the receiver. At the upper left from the tuning-dial assembly is the receiver incremental-tuning control and the S meter, and at the upper right is the afc loop sync. push button. Within the dial assembly appear the dial counter in the window at the left, and the phase-lock meter indicator in the window at the right (an edge-wise meter is used). The home made dial was inspired by the design used in the HRO series of receivers.

This article describes a solid-state receiver which utilizes a somewhat unconventional conversion principle for optimum overall performance. Today's modern electronic components were incorporated for realization of this project, which offers the following highlights:

- 1) Single-conversion principle with very selective 9-MHz crystal filters.
- 2) MOSFET front end for good cross-modulation performance.
- 3) Double-balanced diode mixer with hot carrier diodes.
- 4) Electronically tuned front end using varicap diodes.
- 5) Phase-locked precision-tuned hf oscillator.
- 6) Complete coverage of all ham bands 160 through 10 meters.

7) Transceive output signal 9 MHz above the received frequency plus receiver incremental tuning.

8) Extensive use of linear integrated circuits and toroidal inductors.

Block Diagram

Fig. 1 shows the main layout of the receiver. The upper portion of the diagram is the basic receiving system and the lower portion is used for generation of the main conversion frequency.

The rf signal from the antenna is amplified in a cascode amplifier followed by a source follower which provides impedance matching to the balanced diode mixer. The intermediate frequency of 9 MHz is filtered by one of three crystal filters selectable for a-m, ssb, or cw reception and then passes to a 3-stage i-f amplifier. Separate demodulators for a-m and ssb/cw are provided. For ssb/cw reception a

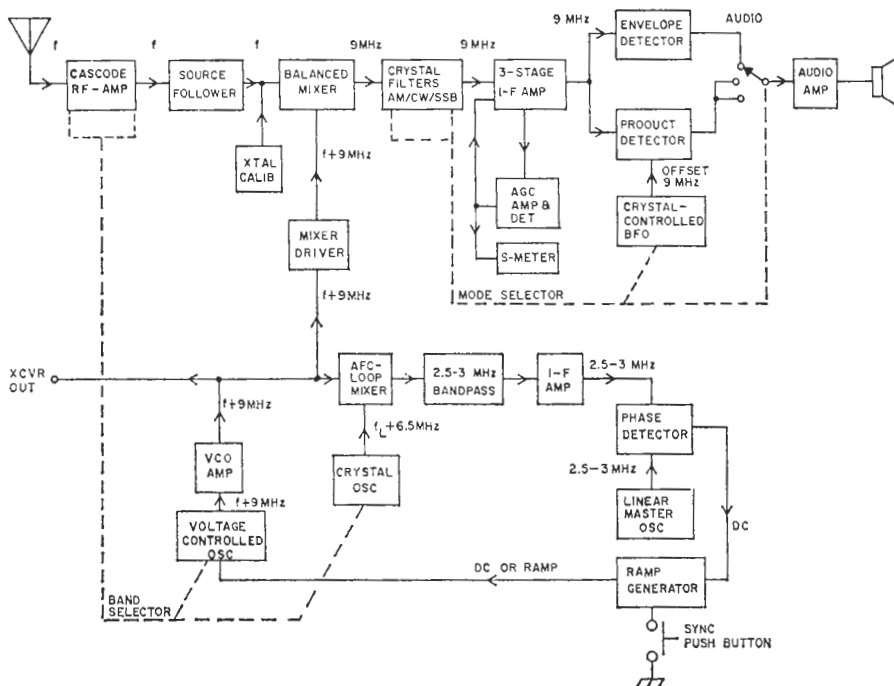


FIG. 1—Block diagram of the receiver. Dashed lines show band-switched and mode-switched stages. The frequencies of receiver operation are shown between appropriate blocks.

f —received frequency, MHz.

f_L —lower frequency limit of amateur band segment in use (1.8, 3.5 . . . 28.5, 29.0, or 29.5 MHz).

crystal-controlled BFO feeds the injected carrier to the product detector. A part of the i-f signal is amplified separately for agc generation and for the S-meter circuit. The detected output passes to a 2-watt audio amplifier which feeds a low-impedance speaker or dynamic headsets. A 1.75-MHz calibration oscillator can be switched on for checking the low-frequency band edges.

The receiver's main conversion oscillator is a so-called VCO, a voltage-controlled oscillator, which is phase locked to a lower-frequency linear master oscillator. This VCO is simply a free-running oscillator, the frequency of which depends on the reverse dc voltage applied to a varicap, which is an integral part of the oscillator tank circuit. The VCO frequency can be voltage-tuned through the range of frequencies 9.0 to 9.5 MHz above the lower frequency limit of the received ham band segment. Phase locking is obtained in the following manner: The VCO frequency is mixed with a crystal frequency 6.5 MHz higher than the low-frequency end of the received band. This results in an i-f of 2.5 to 3 MHz (with 500-kHz band coverage available from this layout). This i-f signal passes through a band-pass filter, is then amplified in a broadband amplifier, and fed to a phase detector. This phase detector gives a dc error output voltage when the phases (and thus frequencies!) of

the i-f and LMO signals are deviating from each other. In the phase-locked condition, any frequency drift of the VCO causes an error voltage at the phase-detector output, correcting the VCO frequency within microseconds. The VCO frequency therefore follows exactly that of the LMO, and perfect stability is achieved although the VCO operates at a much higher frequency. This is called an automatic frequency control or afc loop.

By pressing the sync push button, the VCO frequency is swept through its spectrum by a ramp or low-frequency sawtooth waveform voltage. Locking to the LMO frequency is obtained by the phase detector error signal at the moment of phase coincidence.

The main advantage of this manner of frequency generation is that the VCO can be designed for high spectral purity (or low spurious content), so problems associated with unwanted mixing products are reduced considerably. Broad frequency coverage can be obtained by combining a simple VCO and a crystal oscillator to give the proper afc loop intermediate frequency. The disadvantage of course is the more complex circuitry and higher cost. Commercial frequency synthesizers are using this principle to generate stable frequencies up into the GHz range.

Figs. 2 through 5 show the schematic diagrams of the various parts of the receiver.

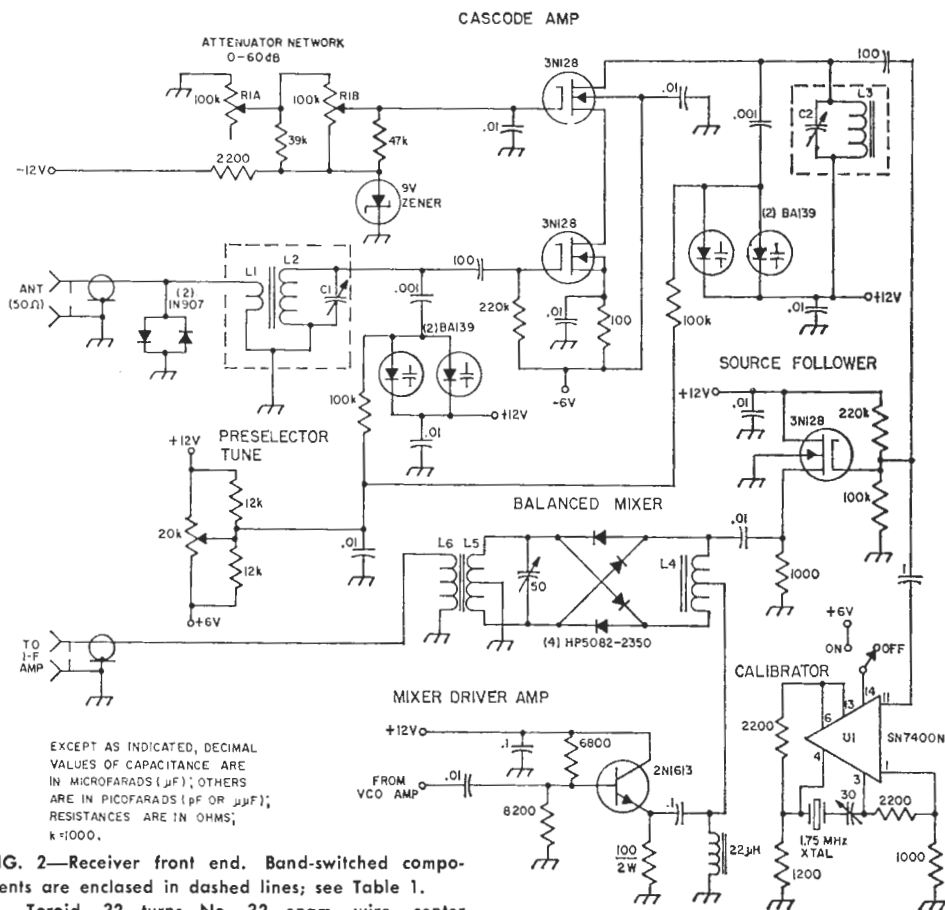


FIG. 2—Receiver front end. Band-switched components are enclosed in dashed lines; see Table 1.

L_4 —Toroid, 32 turns No. 32 enam. wire, center tapped, wound on Indiana General Ferramic Q2 core, form CF102.

L_5 —Toroid, 16 turns No. 32 enam. wire, center tapped, wound on Indiana General Ferramic Q3 core, form CF102.

L_6 —Toroid, wound over L_5 ; 20 turns No. 32 enam. wire.

R_1 —Dual control, each section 100,000 ohms linear taper.

Table I (page 200) gives the data for constructing or selecting the band-switched components. The following paragraphs provide operational information about each section of the receiver.

Receiver Front End

The rf signal coming from the low-impedance antenna is clamped by a pair of fast-acting silicon diodes, which protect the receiver against excessive signals from the station transmitter. If the broad-band signal voltages from the antenna are in excess of 200 mV, which never happened with the author's various antennas, protection should be made in another manner, such as an rf attenuator, to prevent cross-modulation effects.

Inductive coupling is used to the input circuit. Winding ratios were carefully adjusted with a noise generator. Input and drain circuits are wound on separate ferrite toroids for

each band and are peaked by small concentric trimmer capacitors. Band tuning is done electronically by two paralleled varicaps, completely eliminating the mechanical problems associated with a ganged tuning capacitor. Reverse voltage of 0 to 6 volts is applied to the varicaps from the panel-mounted PRESELECTOR TUNE potentiometer. When using varicap diodes in a tuned circuit, care should be taken that the reverse voltage applied to the diodes is an order of magnitude larger than the peak rf voltage across them. Otherwise, curious effects will be observed when the rf voltage modulates the varicaps or drives them into conductivity. The author used German BA139 varicaps from Siemens Company for his receiver. These diodes offer a capacitance variation of 3 to 20 pF for a 20- to 1-volt reverse voltage, and a Q of about 400 at 30 MHz. An equivalent varicap should be the VC615 Evicap from Easton Corp., 25 Locust

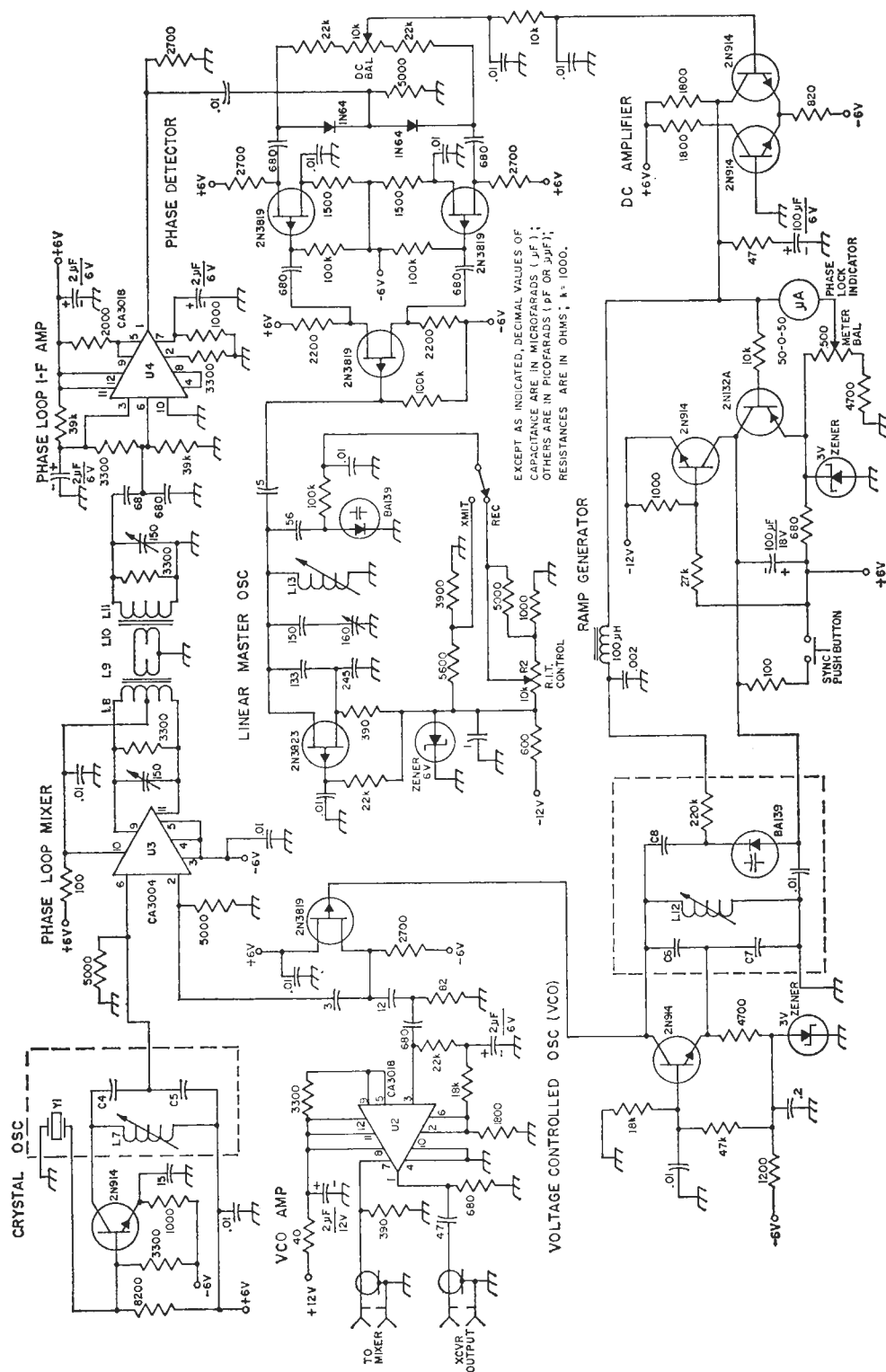


FIG. 3—Phase-locked main-conversion-oscillator system. Capacitors with polarity indicated are electrolytic. Band-switched components are enclosed in dashed lines; see Table 1.

L_8 , L_{11} —Toroid, 40 turns No. 32 enam. wire, center tapped, wound on Indiana General Ferramic Q2 core, form CF102.

L_9 , L_{10} —Toroid, wound over L_8 or L_{11} ; 4 turns No. 32 enam.

L_{13} —20 μH; 42 turns No. 30 enam. wire closewound on slug-tuned ceramic form ½-inch dia. 1-inch long.

R_2 —Receiver incremental tuning; 10,000-ohm 10-turn control (Spectrol model 162 or equiv.).

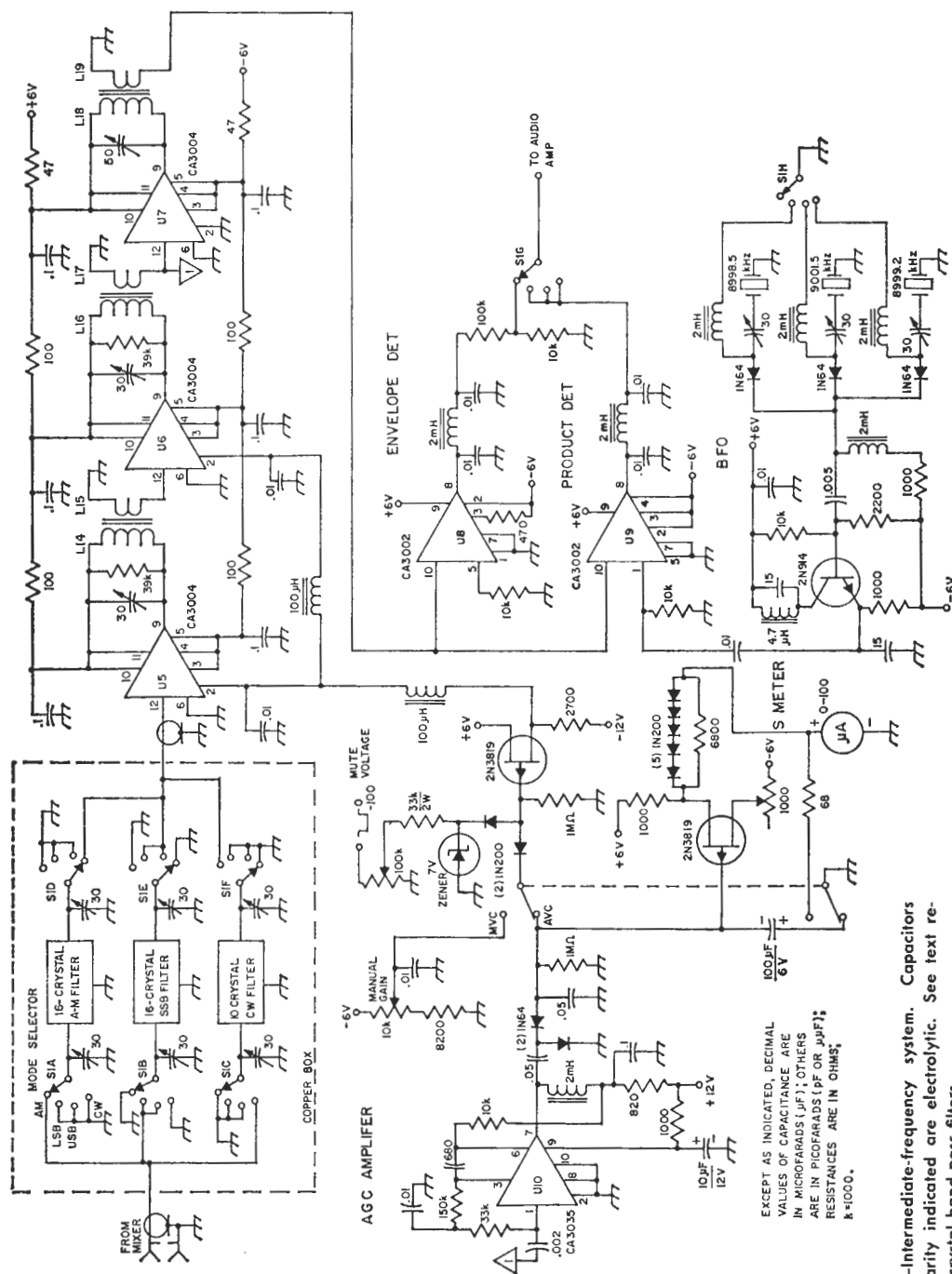
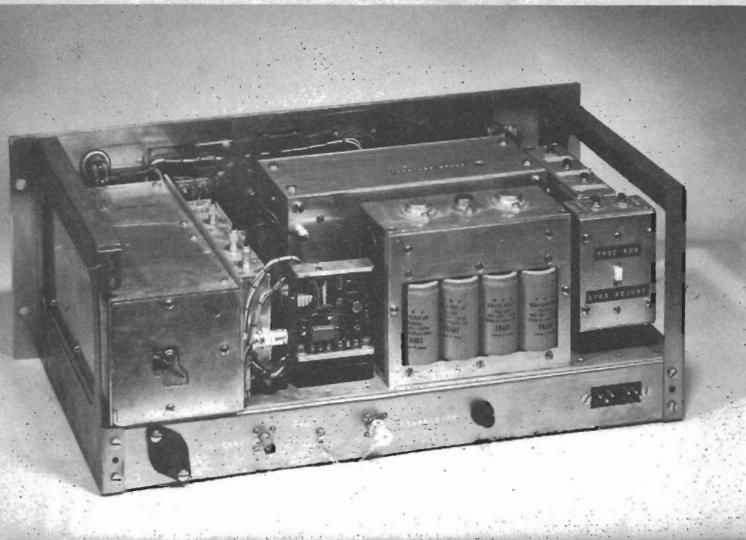


FIG. 4—Intermediate-frequency system. Capacitors with polarity indicated are electrolytic. See text regarding crystal band-pass filters.

L_{13} , L_{16} , L_{18} —Toroid, 28 turns No. 32 enam. wire, wound on Indiana General Ferramic Q3 core, form CF102.

L_{15} , L_{17} , L_{19} —Toroid, wound over L_{14} , L_{16} , and L_{18} respectively; 9 turns No. 32 enam. wire.

S_1 —Miniature ceramic rotary, 3 sections, 9 poles, 5 positions, non-shorting; 8 poles and 4 positions used (Centralab PA2029 or equiv.).



Rear view of the receiver, showing block construction and heavy shielding. The large compartment at the left of the chassis is the rf assembly. To the right of center may be seen the four electrolytic capacitors in the power supply. The assembly forward of the supply is the LMO.

St., Haverhill, Mass. 01830.

Having carefully checked preamplifier performance of transistor front ends, a cascode configuration with two 3N128 MOSFETs was chosen. Its gain of 35 db at 30 MHz is by far too high to be usable, but when it is reduced to 10 to 20 db by applying reverse bias to the gate of the upper transistor, this results in a cross-modulation characteristic comparable to vacuum tubes. Measurements were performed with two modulated signal generators 50 kHz apart; the desired signal was 10 μ V and within the receiver's bandwidth. The interfering signal could be as high as 100 mV before noticeable cross modulation occurred. The gain-reduction voltage is derived from a resistor network around a double-ganged potentiometer to get a nearly linear calibration of 0 to 60 dB on this "attenuator." This eliminates an attenuation network at the receiver's antenna input, as it is possible to reduce preamplifier gain below unity.

An FET source follower provides matching to the low-impedance balanced diode mixer, which is a ring modulator. The mixer is driven by a toroidal broad-band balancing transformer, L_1 , which covers 1.5 to 30 MHz. A quartet of Hewlett-Packard hot-carrier diodes 5082-2350 was used for the ring modulator, but the less expensive type 5082-2800 was tested and performs as well at these frequencies. These are ultrafast switching diodes operating on majority carrier conduction and are perfectly suitable for wide-dynamic-range mixer applications. It should be noted that diode mixers have gain below unity. In this layout, conversion loss was measured to be around 8 dB, and proper impedance matching in the mixer is of utmost importance so as not to degrade the overall noise figure. The

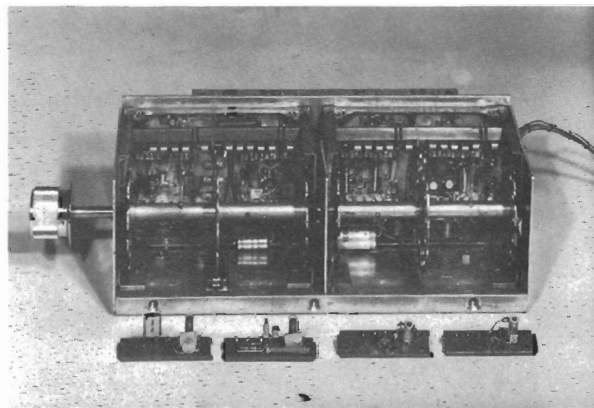
mixer output is coupled through a tuned and balanced transformer, which steps up the mixer impedance of 300 ohms to 500 ohms to match the succeeding crystal filters.

Mixer performance largely depends on the drive level applied to the diodes. A 4-mW signal is delivered from the 2N1613 power transistor to the input transformer's center tap. Signal levels up to 800 mV at the mixer's input can be processed without noticeable cross-modulation. In this front end, any cross modulation which occurred originated from the preamplifier, and not from the mixer. (I am still looking for someone having experience with negative feedback in a low-gain MOSFET cascode amplifier.) Dynamic range of the receiver was measured to be around 80 dB and sensitivity on all bands was better than 0.1 μ V for a 10-dB signal-plus-noise to noise ratio for a 30%-modulated a-m signal; on two bands it was as low as 40 nanovolts.

The calibration oscillator utilizes an SN7400 quadruple-NAND digital IC, having a square-wave output of 5 volts peak-to-peak. Because of its good long-term stability, a 1.75-MHz HC-27/U crystal is used to check the low band edges. The trimmer in series with the crystal is for adjusting it to the proper frequency.

To obtain the shortest leads to all band-switched components, the front end and the phase-locked oscillator (described later) were

The rf assembly built into two ganged TV turret tuners. From left to right are the crystal oscillator, voltage-controlled oscillator, balanced diode mixer, and MOSFET cascode amplifier. A set of bandswitched components is shown in front of the assembly.



built into two ganged TV turret tuners. (See the preceding photograph.)

Phase-Locked Main Conversion Oscillator

The functional principle of the main conversion oscillator was already discussed. More details will be given here for the circuit shown in Fig. 3. This system is completely self-working, providing two output signals for feeding the mixer and for transceiver purposes.

The crystal oscillator has a tuned collector circuit with the crystal connected between the transistor's base and ground. Series-resonant crystals, regardless of whether fundamental or overtone ones, can be used in this oscillator.

The voltage-controlled oscillator is a simple Colpitts, with the varicap in series with C_s forming a part of the tank capacitance. As C_s determines the VCO's tuning range, it is band-switched together with the inductor and the varicap. Using a separate varicap for each band was a precaution with regard to VCO tuning capability, but proved not to be necessary. (The tuning range of the VCO is 8.5 to 10 MHz above the received ham-band segment, thus leaving a margin of 500 kHz on both sides.)

The VCO output passes to an isolation amplifier and is then fed to the integrated-circuit VCO amplifier. An amplitude-frequency correction network at the output of the 2N3819 holds the output level constant regardless of the band being used. The transceiver output is capable of delivering 250 mV rms into a 50-ohm load at 2-dB regulation.

Frequencies of the crystal oscillator and the VCO are mixed in the CA3004 phase-loop mixer, U_3 . The bandpass filter following the mixer consists of two inductively coupled toroidal circuits damped by parallel resistors to obtain a flat passband from 2.5 to 3.0 MHz. U_4 , an IC broad-band amplifier having 35 dB gain, brings the level up to 1.5 volts rms to feed the phase detector.

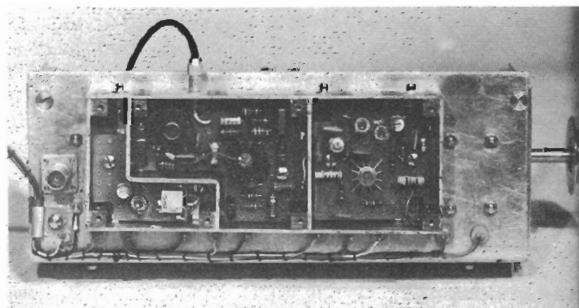
The linear master oscillator is an FET Colpitts tuned by an old BC-221 variable capacitor. Only 80% of the capacitor's tuning range is used to cover the 2.5- to 3.0-MHz range, and maximum frequency error could be held below ± 500 Hz with respect to linear dial markings. Final adjustment for linearity was made by trimming the outer rotor plates. All capacitors in the tuning circuit are 750-ppm/degree negative temperature coefficient, resulting in an overall frequency drift below 5 ppm/degree for the complete assembly at room temperature. The linear master oscillator would be difficult to duplicate without a frequency counter being at hand; there are excellent LMOs from Collins on the market.

Receiver incremental tuning is done during receiving periods by a varicap fed from a 10-turn Helipot, giving 2 kHz per revolution at midband.

The phase detector must be a broad-band

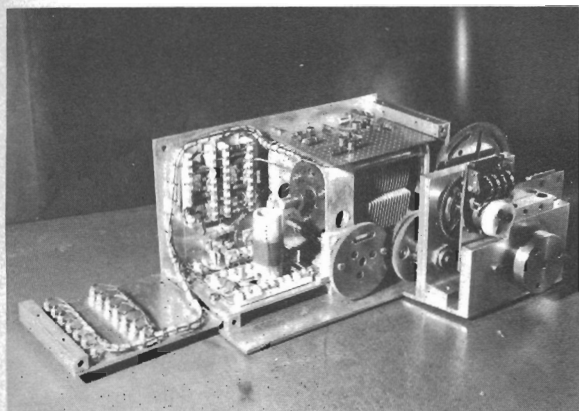
version to cover the 2.5- to 3.0-MHz range. Three inexpensive FETs are used, the first to obtain two LMO signal voltages 180 degrees out of phase. The other two FETs amplify the frequency to be detected up to 2 volts rms, and the amplified signal is fed to the phase-sensing diodes. Within the so-called capture range, the phase-detector output voltage is positive for frequencies lower and negative for frequencies higher than that of the LMO. Graphically plotted this is an S-shaped curve, with the peaks only fractions of a hertz on either side of the LMO frequency.

Only an i-f falling inside these close peaks will generate an error output signal of the correct polarity to lock the afc loop. As this cannot happen by itself, tuning into the capture range must be done by another device, the ramp generator. The phase-detector output passes through a simple inverting dc amplifier, the output of which feeds the ramp generator, and simultaneously feeds the cathode of the varicap in the VCO. The ramp-generator out-



Rear view of rf assembly showing from right to left the afc loop mixer, VCO amplifier, and mixer driver amplifier. Heat sinks with radial fins have been placed over the CA3018 IC and the 2N1613 transistor.

put feeds the anode of the same varicap. Pressing the sync push button shorts the 100- μ F electrolytic capacitor in the ramp generator with a 100-ohm resistor. Releasing the button starts the 2N914 transistor to act as a constant-current source which slowly charges this capacitor from +6 towards -12 volts, thus forming a linear ramp voltage. This ramp voltage tunes the frequency of the VCO through its spectrum at a rate of approximately 500 kHz per second, starting from the low-frequency end. At the moment the afc-loop i-f, which is also being swept through its spectrum by this procedure, tends to cross the capture range of the phase detector, a negative output voltage from the dc amplifier starts conduction of the 2N132A transistor. Before, this transistor was held below cutoff by the Zener diode in its emitter. Its conduction now stops the ramp voltage from rising any more, by clamping it to a fixed level. This is the moment the VCO locks in.



Linear master oscillator assembly, with the BC-221 tuning capacitors, reduction gear train and counter dial. The phase detector circuit is constructed on the back panel, and the afc loop i-f amplifier, on perforated board, is mounted on top of the assembly.

Once locking has been achieved after each band change, the frequency of the VCO stays like the "Rock of Gibraltar," even at 38 and 39 MHz. A small center-reading meter signals the locked condition. Note the series combination of a 47-ohm resistor and a 100- μ F electrolytic capacitor at the dc amplifier's output. This is a very important network to give the proper time constant within the afc loop for low phase jitter and good locking capability.

Of course the SYNC push button might have been replaced by a band-change activated contact, but it was provided to demonstrate the small difference in comparison with normal receivers, and so it stayed there.

The I-F System

The incoming i-f signals are filtered by high-frequency crystal filters having a center fre-

quency of 9 MHz. One of three high-performance filters for a-m, ssb and cw reception is selected by the mode switch. Trimmer capacitors at the input and output terminal of each filter are provided for adjusting passband ripple to minimum. The mode selector switch with the crystal filters positioned around it was built into a copper box, leaving only coaxial connectors emerging for the cables to the mixer and i-f amplifier. It is very important that, within the box, two separate contact sections be used for input and output terminals with careful shielding between them, and that all unused filter terminals be grounded. Otherwise, stopband attenuation would be determined primarily by stray coupling around the filters, and not by the filters themselves.

The crystal filters used by the author are computer-designed laboratory models from a German crystal manufacturer having bandwidths of 5.0, 2.4 and 0.4 kHz for a-m, ssb, and cw reception, respectively. A total of 42 crystals are incorporated, giving 140-dB stopband attenuation for the a-m and ssb filters,

TABLE I
Band-switched components

Component	Band					
	160	80	40	20	15	10 (A through D)
L ₁	6 turns } Ferramic Q2—	5 turns } Ferramic Q3—	3 turns } Ferramic Q3—	2½ turns } Ferramic Q3—	2 turns } Ferramic Q3—	1½ turns } Ferramic QR—
L ₂	57 turns toroid	53 turns toroid	30 turns toroid	15 turns toroid	10 turns toroid	8 turns toroid
C ₁	10—60-pF trimmer	10—60-pF trimmer	3—30-pF trimmer	3—30-pF trimmer	2—20-pF trimmer	2—20-pF trimmer
L ₃	57 turns, Q2—toroid	53 turns, Q3—toroid	30 turns, Q3—toroid	15 turns, Q3—toroid	10 turns, Q3—toroid	8 turns, Q3—toroid
C ₂	10—60-pF trimmer	10—60-pF trimmer	3—30-pF trimmer	3—30-pF trimmer	2—20-pF trimmer	2—20-pF trimmer
L ₂ *	28 turns	23 turns	40 turns	24 turns	18 turns	14 turns
C ₃	47 pF	47 pF	30 pF	10 pF	10 pF	10 pF
C ₄	470 pF	470 pF	390 pF	120 pF	120 pF	120 pF
L ₁₂ *	26 turns	22 turns	19 turns	14 turns	12 turns	10 turns
C ₅	33 pF	26 pF	13 pF	10 pF	8 pF	6 pF
C ₇	27 pF	27 pF	27 pF	27 pF	27 pF	27 pF
C ₈	56 pF	39 pF	27 pF	15 pF	10 pF	6 pF
Y ₁	8.3 MHz	10.0 MHz	13.5 MHz	20.5 MHz	27.5 MHz	(A) 34.5 MHz (B) 35.0 MHz (C) 35.5 MHz (D) 36.0 MHz

* L₂ and L₁₂ are wound on slug-tuned phenolic forms ½-in. dia., ½-in. long. Tuning is done on 160- and 80-meter bands by high- μ slug cores, on 40- through 10-meter bands by low- μ slugs. All inductors are wound with enameled No. 32 wire. Toroids are wound on Indiana General Ferramic Q2 or Q3 cores, form CF102.

and 90 dB for the cw filter.¹ Any 8-crystal filter designed for low insertion loss and having a good shape factor will do well in such a receiver.² A very important aspect: stop-band attenuation of the filters should be larger than the receiver's dynamic range!

Thanks to RCA's application literature, the i-f amplifier could be built easily with linear integrated circuits. A 3-stage amplifier using CA3004s gives a 65-dB voltage gain, where the first two amplifier stages are controlled by either manual or automatic gain control. Tuned toroidal transformers are used for inter-stage coupling and impedance matching.

The last i-f amplifier feeds both the envelope detector for a-m signals and the product detector for ssb/cw reception. The crystal-controlled BFO was built around a 2N914 transistor. A 1.7-volt rms signal is delivered to the product detector for optimum performance. A trick used to select the appropriate BFO crystal for lsb, usb or cw operations is electronic remote switching, by gating one of the three diodes in series with the crystals. Series trimmers are provided for adjusting the crystal frequencies to the desired point on the filter skirt.

A part of the i-f voltage is picked up after the 2nd i-f amplifier, amplified 40 dB in the agc amplifier CA3035, U_{10} , and rectified by a voltage doubler. This agc voltage is fed via a gate diode to the 2N3819 source follower, which provides gain control of the 1st and 2nd i-f amplifier stages. The gate of this transistor can be switched from the agc signal to the manual gain potentiometer. The available receiver muting voltage of -100 volts is clamped by a Zener diode and fed to a second gate diode to shut down the receiver during transmitting periods. The 100-k Ω potentiometer is set for a convenient level during transmitting periods.

The agc voltage is fed simultaneously to another 2N3819 FET which is a dc amplifier with a bridged S meter. The resistor-diode network in series with the meter provides linearity. A 50- μ V signal at the antenna input was fixed to be S9, and a fairly linear calibration of 5 dB per S unit could be obtained.

When the receiver is operated with agc, a 100- μ F capacitor is switched in, giving a fast-attack, slow-decay characteristic, with a decay time constant of 5 seconds. When the receiver is switched to manual gain control, a scale with a calibration of 0 to 20 dB is used. The receiver is then run in the linear amplification mode, and the dynamic range is slightly higher than 20 dB. That is to say, for the linear mode (mgc) and the logarithmic mode (agc) two different calibrations on the meter are used.

¹ Careful compartment shielding and double-shielded interconnecting coaxial cables are required in order to realize the benefits of stopband attenuations of the magnitude described by the author. In practice, even with careful physical layouts, shielding to provide greater than 90- or 100-dB isolation between input and output circuits is difficult to obtain.—Editor.

² Suitable 9-MHz crystal filters, manufactured by KVG in West Germany, are available in the U.S. from Mr. Henry Ingwersen, Spectrum International, Topsfield, Mass. 01983.—Editor

Audio System and Power Supply

Details on the power supply and audio amplifier will be omitted here for simplification. A transformer with two 18-volt secondary windings feeds two bridge rectifiers and two 10,000- μ F filtering capacitors to obtain positive and negative supply voltages. Four Zener-stabilized series regulators deliver outputs of -12, -6, +6, and +12 volts. Attention should be paid to residual hum, which should be kept below 2 or 3 mV. Total power drain of the receiver is less than 20 watts.

Mechanical Layout

When constructing a receiver of this complexity, one never knows beforehand whether a specific new circuit will work satisfactorily or not (and several did not!). Separate blocks for the power supply, audio system, LMO, i-f system, and front end were constructed. Each block was tested outside the arrangement and inserted for final alignment. A disgusting but nevertheless very important problem to avoid is unwanted coupling. Supply voltages are therefore applied with feedthrough capacitors to each block and in some cases to each amplifier stage. Rf interconnections were made with small-diameter coaxial cable and appropriate connectors.

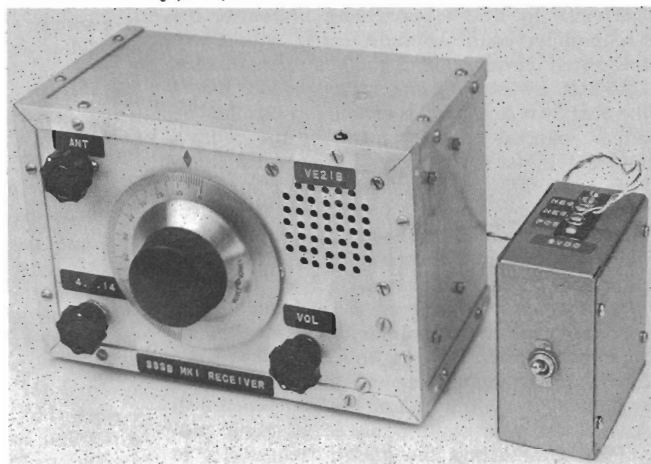
The main tuning gear was built around an old BC-221 tuning capacitor and reduction gear. The counter dial and tuning knob are the result of four weeks of labor. The counter dial reads tens of KHz, where the main tuning knob has a calibration of 200 Hz per division, from an HRO inspiration. The front panel was covered with a special 3-layer plastic which allowed direct engraving of the function titles.

The metal cabinet and chassis frame is a commercial one. By the way, this transistor receiver has a weight of 52 lb; excessive shielding pays in electrical performance, but not in weight!

References

1. "The Hot Carrier Diode, Theory, Design and Application," Hewlett-Packard Associates, Application Note 907.
2. Carlson, F. M., "Application Considerations for the RCA 3N128 VHF MOS Field Effect Transistor," RCA Application Note AN-3193.
3. "RCA Linear Integrated Circuits Application Notes," Technical Series Documentation IC-41.
4. Hacks, J., "Treffericherheit bei KW-Empfängern," Rohde & Schwarz Publication 10, 1958.

» The complexity and cost of a receiver can be kept at a minimum by designing the set for only one or two bands. VE2IB takes this approach in his simple SSB receiver, the SSSB Mark I. In addition, by having most of the receiver's gain in the audio channel, he has developed an easy-to-align set that is relatively free from overload.



Front view of the two-band receiver and 9-volt power supply. The receiver cabinet is homemade from angle and sheet aluminum. A commercial steel utility box is used to house the power supply.

A Simple Transistor Receiver for 20- and 75-Meter Sideband

C. A. LAMONTAGNE, VE2IB

The unit shown in the photographs and diagrams is the outcome of my attempt to build a transistor 20-meter s.s.b. receiver that would meet the following requirements:

- 1) Low cost.
- 2) Simple to construct.
- 3) Easy to adjust without elaborate test equipment.
- 4) Materials locally available and easy to get.

All of these objectives have been achieved, and 75-meter coverage has been included as a bonus. The circuit, a simple single-conversion superheterodyne, consists of an emitter-follower mixer, a half-lattice crystal filter, a single i.f. stage at 9 Mc., a product detector, a low-level audio stage, a commercial audio-amplifier module and a loudspeaker. Also included are a two-transistor tunable oscillator and a crystal-controlled b.f.o.

Because the receiver doesn't have an r.f. stage, switching complications have been avoided, and cross modulation is no problem. No r.f. shielding is necessary in the receiver, since toroids, which have practically no external fields, are used for all the r.f. coils.

The receiver can be powered from either flashlight cells or a separate 9-volt supply that operates off the a.c. line.

Circuit Details

Fig. 1 is a schematic diagram of the receiver. Single-tuned antenna circuits are used because they are adequate at my location. A 9-Mc. i.f. trap is included in the antenna lead, but it may not be sufficient to eliminate 9-Mc. feedthrough in some locations. If such is the case, double-tuned antenna circuits should be employed. Separate padding capacitors, C_1 and C_2 , are used to avoid mistuning the antenna trimmer, C_3 , to 9 Mc; however, the padders 140-pf. are not strictly necessary, and a single variable should work well at C_3 if ultimate simplicity is required.

Q_1 , the emitter-follower mixer, developed from my attempt to use an emitter follower to isolate the antenna from a diode mixer. The diode was inadvertently shorted out, and mixing continued to take place. Investigation resulted in the circuit shown. Q_1 has plenty of gain, is quiet and not prone to cross modulation. Note, however, that this stage may oscillate if too low a value of emitter resistor is used or if Q_1 's base lead is not tapped down far enough on the input coil in use.

The v.f.o. is a two-transistor affair that operates from 5.0 to 5.4 Mc. An emitter follower, Q_2 feeds a grounded-base amplifier, Q_3 , through a shared emitter resistor. Because the input

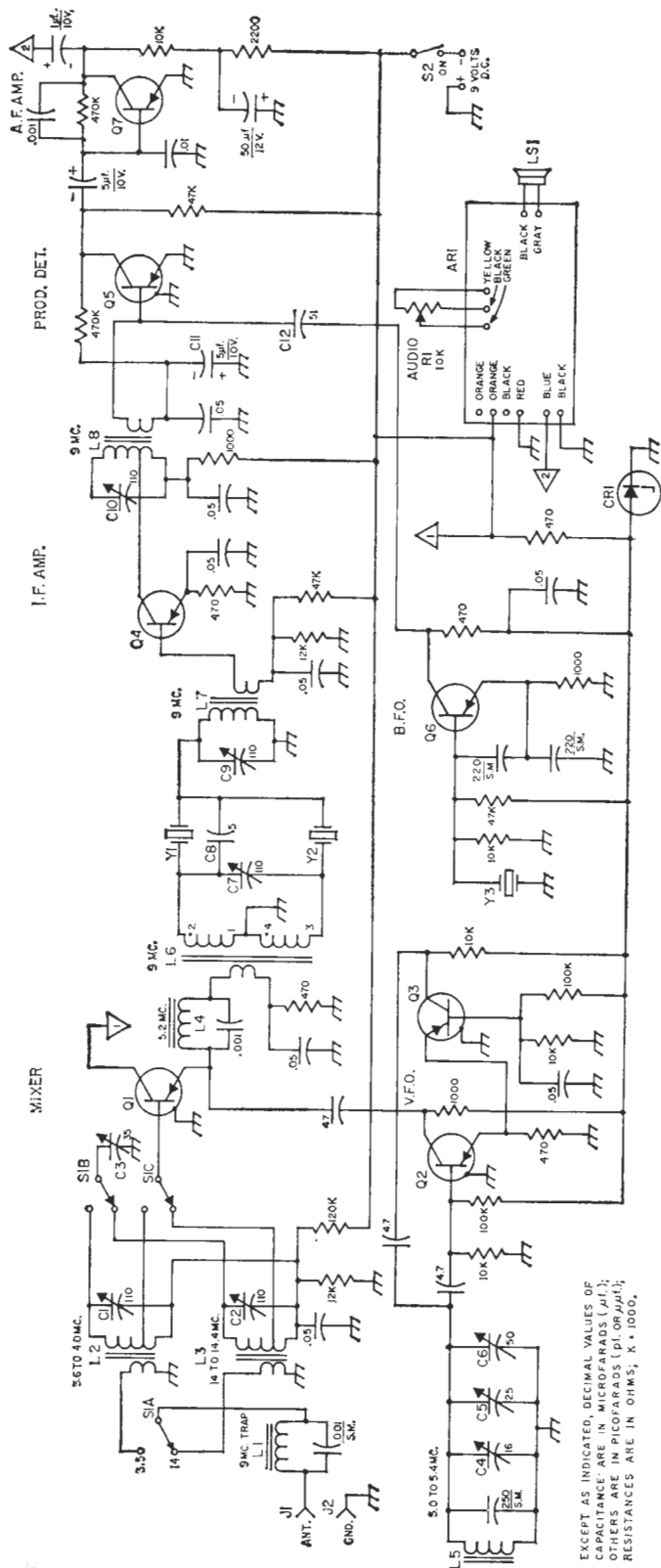


FIG. 1—Circuit of the two-band s.s.b. receiver. Fixed capacitors marked SM are silver mica; those marked with polarity are electrolytic; 0.05- and 0.01- μ F units are Mylar; all others are ceramic. Coils are wound with No. 28 enameled wire on toroid forms made from powdered iron slugs removed from National XR-50 coil forms; details are given in the text. Fixed resistors are $\frac{1}{2}$ -watt composition.

AR₁—Audio amplifier module (Lafayette Radio 99 HP037).

C₁, C₂, C₇, C₉, C₁₀—11–110-pf. ceramic trimmer (Erie 503–000 P3P0 62R or equivalent).

C₃—35-pf. variable (Hammarlund MAPC-35-B).

C₄—N300 ceramic trimmer, 3–16 pf. (Erie 503–000 B2P0 19R or equivalent).

C₅—NPO ceramic trimmer, 4.5–25 pf. (Erie 503–000 COP0 25R or equivalent).

C₆—50-pf. variable (Hammarlund HF-50).

C₈—5-pf. disk ceramic.

C₁₁—5- μ F, 10-volt electrolytic.

C₁₂—See text.

CR₁—7-volt Zener, $\frac{1}{4}$ watt or greater (Sarkes Tarzian VR7A or equivalent).

J₁, J₂—Alligator clip.

L₁—6 turns.

L₂—Secondary, 50 turns, tapped 10 turns from cold end; primary, 4 turns around cold end of secondary.

L₃—Secondary, 13 turns, tapped 2 turns above cold end; primary, 2 turns around cold end of secondary.

L₄—12 turns.

L₅—30 turns.

L₆—Secondary, two 15-turn coils, bifilar wound, end 2 parallel with end 4, and end 1 parallel with end 3; primary, 2 turns midway between beginning and end of bifilar winding (at the center tap).

L₇—Primary, 30 turns; secondary, 2 turns around cold end of primary.

L₈—Primary, 30 turns, tapped 10 turns from cold end; secondary, 3 turns around cold end of primary.

LS₁—10 ohms, 0.5 watt, $\frac{1}{2}$ 1/2-inch cone.

Q₁, Q₄, Q₆—2N1225 used, RCA SK3006 suitable.

Q₂, Q₃, Q₅—2N1396 used, RCA SK3006 suitable.

Q₇—2N2613 used, RCA SK3004 suitable.

R₁—10,000-ohm volume control, audio taper, with s.p.s.t. switch.

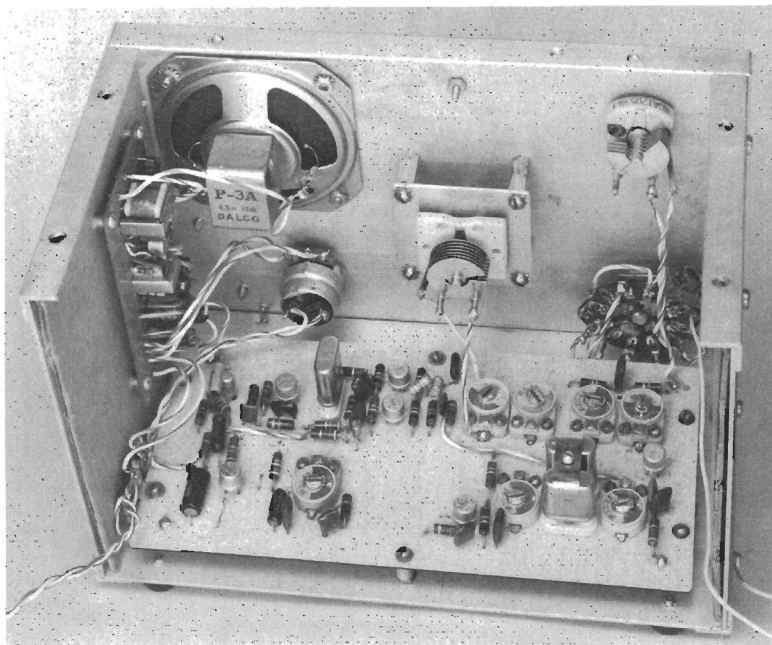
S₁—Phenolic rotary, 1 section, 4 poles, 2 positions.

S₂—Part of R₁.

Y₁, Y₃—27.005-Mc. third-overtone crystal (CB Channel 4).

Y₂—27.015-Mc. third-overtone crystal (CB Channel 5).

Top view of the SSSB Mark I. The audio module is at the left; to its right on the circuit board, from left to right, are the audio amplifier, b.f.o., v.f.o. and mixer. In the foreground, from left to right, are the product detector and the i.f. amplifier.



impedance of the emitter follower and the output impedance of the grounded-base amplifier are both moderately high, no capacitive divider or coil taps are necessary for impedance matching. Since each stage has its input in phase with its output, feedback to sustain oscillation is easy to obtain. One 4.7-pf. ceramic capacitor couples the input of Q_2 to the 5.0- to 5.4-Mc. tuned circuit, and another 4.7-pf. capacitor couples the output of Q_3 to the same circuit.

The oscillator signal is fed to the mixer by capacitively coupling the collector of Q_2 to the emitter of Q_1 . Normally Q_2 's collector would be grounded for r.f.; however, by leaving the collector hot and using it as a coupling point, you have a receiver in which antenna trimmer adjustments do not pull the oscillator frequency.

A low-impedance link in the emitter lead of Q_1 is used to couple the 9-Mc. output of the mixer to the half-lattice crystal filter. The filter is conventional with one exception; it uses two 27-Mc. third-overtone CB crystals, Y_1 and Y_2 , on their fundamental frequencies near 9 Mc. Y_1 was for CB Channel 4 (27.005 Mc.) and Y_2 was for CB Channel 5 (27.015 Mc.)

The 9-Mc. i.f. stage, Q_4 , is conventional in every way. If the input link is kept small, neutralization is not necessary.

Because there is little amplification ahead of it, the product detector, Q_5 , is a critical stage in this receiver. The circuit is essentially that of a mixer which has had particular attention paid to audio bypassing. Without C_{11} , a 5- μ f. electrolytic at the cold end of the base circuit, Q_5 is insensitive and produces a great deal of audio hiss.

Q_6 , the crystal-controlled b.f.o., is based on a

circuit described by Stoner.¹ Y_3 , a 27.005-Mc. third-overtone CB crystal is used on its 9-Mc. fundamental in a series-resonant mode. If a few CB crystals can be tried, the one that results in the best sounding audio should be retained. Because no provision has been made for adjusting the crystal frequency or switching crystals, only u.s.b. signals can be received on 20 meters and only l.s.b. signals on 75. If you want to vary the crystal frequency, a trimmer can be inserted in series with Y_3 .

The b.f.o. signal can be coupled to the product detector in one of two ways: either a gimmick capacitor can be used between the collector of Q_6 and the hot end of L_8 's primary, or a 51-pf. capacitor, C_{12} , can be connected between J_6 's collector and the base of Q_5 . If the first method is used, two insulated hookup wires that are twisted together for about $\frac{1}{4}$ inch will make a suitable capacitor.

A 7-volt Zener CR_1 , is used to regulate the collector supply voltage of both the b.f.o. and the tunable oscillator.

To make up for the gain usually found in an r.f. stage and additional i.f. stages, an audio amplifier, Q_7 , is used between the product detector and the audio-amplifier module, AR_1 . RC coupling and filtering are used for simplicity and to reduce the chances of hum pickup.

Power Supply Details

The power supply circuit is shown in Fig. 2. It consists of a voltage doubler followed by a transistor series regulator. A Zener diode, CR_1 ,

¹ Stoner, *Transistor Transmitters for the Amateur*, 1st Edition, page 47.

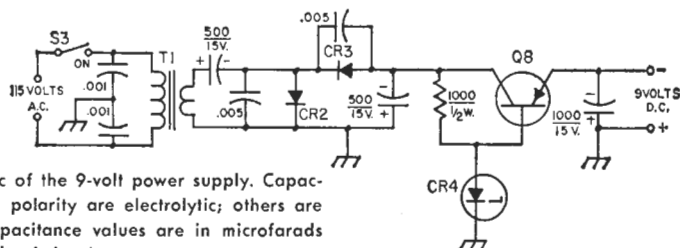


FIG. 2.—Schematic of the 9-volt power supply. Capacitors marked with polarity are electrolytic; others are disk ceramic. Capacitance values are in microfarads (μf); resistance value is in ohms.

CR₂, CR₃—400-p.i.v., 1000-ma. silicon diode (International) Rectifier 5A4 or equivalent).

CR₄—9.1-volt, 1-watt Zener (General Electric Z4XL9.1B or equivalent).

Q_g—2N1038

S₂—S.p.s.t. toggle.

T₁—6.3 v. at 0.6 ampere.

provides a more-or-less constant voltage reference for the regulator transistor, Q_s .

Since so much of the receiver's gain is at audio frequencies where hum pickup can be a problem, the power supply is housed in a separate $4 \times 4 \times 2$ -inch steel utility cabinet. It seemed easier to shield the a.c. power supply by putting it in a separate steel box than to shield the receiver. Although aluminum provides no a.c. shielding, it is used in the receiver itself because it is easier to work with.

For portable operation, six 1.5-volt flashlight cells in series can be used to power the set. The usual 9-volt transistor battery isn't recommended, as it doesn't have enough power to supply the rather large current demands of the audio module's Class B stage at high signal levels.

Construction

The SSSB Mark I required no fancy construction work of any kind. A 4 × 8-inch piece of kitchen-counter laminate, obtained from the local hardware store, is used as a circuit board for Q_1 through Q_7 . Except for the toroids, which are glued with Glyptal cement to the underside of the laminate, components are mounted on top of the board, and their leads are fed through small holes drilled at convenient places. The leads are clipped underneath the board to a length of about $\frac{3}{4}$ inch. Connections are made by running hookup wire to the component leads and looping the wire around them. With this type of construction, it's easy to make changes and substitutions. Crossovers are kept at a minimum with a little forethought.

Except for the i.f. input and output coils, which should be at least two inches apart, the placement of the components is noncritical. For the most part, the various stages are arranged in sequence, input to output, around the outside of the board. It was found convenient to locate most of the toroids on the underside of the board in the open spaces below the ceramic trimmers.

No chassis is used in the SSSB Mark I. Instead, the circuit board is supported above the cabinet bottom plate by six $\frac{1}{2}$ -in. spacers. The $5\frac{1}{2} \times 6 \times 9$ -inch cabinet is made from pieces of sheet and angle aluminum that are bolted together with self-tapping screws; however, a

commercial utility cabinet will probably work just as well. C_6 , the main tuning capacitor, is mounted on a 2-inch square aluminum plate that is spaced from the front panel with four $\frac{3}{4}$ -inch long tapped metal pillars. The pillars are secured by the same screws that bolt a National type N Velvet-Vernier dial mechanism to the front of the cabinet.

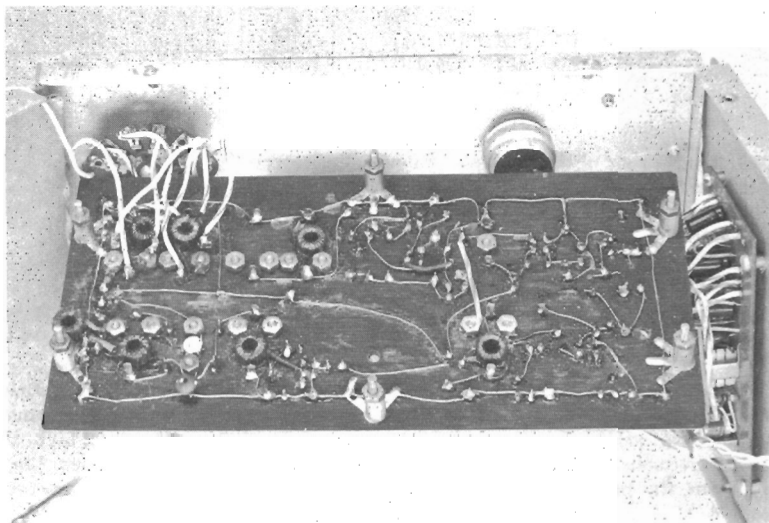
All the toroids in the receiver are made from ½-inch diameter powdered iron slugs removed from old National XR-50 coil forms. With a hacksaw, the slugs were scored around the circumference about ¼ inch from the end that has a slot for a screwdriver blade. A wafer was split off each slug by rolling the slug under a knife blade held in the hacksaw cut. Fortunately, most wafers break off evenly. Next a small piece of softwood was drilled with a bit whose diameter was the same as the slug's, and a wafer was inserted in the hole with the slotted end of the wafer facing up. A small vise was used to compress the wood and hold the wafer tight. With a ⅝-inch diameter high speed bit, a hand-held electric drill was used to cut a hole through the center of the wafer. The hole was enlarged to about ⅝ inch either with sandpaper rolled around a drill bit or with a very small grindstone of the kind mounted on a shank for use with an electric drill or a polishing machine. Side pressure must be kept to a minimum or the toroid will crack. Sandpaper was used to smooth the edges and to reduce the form to the desired ⅝-inch width. Unfortunately, the hacksaw blades, drill bits and round files, used to form homemade toroids are soon blunted.

Other powdered-iron slugs may be used if they are designed for high frequency operation. A rough indication of a slug's high frequency performance as a core material can be had by inserting the slug into a high-frequency grid-dip-meter coil. If the meter reading drops, the material is lossy at the indicated frequency and should not be used.²

If one doesn't wish to operate on coil slugs, commercial toroid forms may be used. Indiana General CF-101 cores of Q2 material or CF-102 cores of Q3 material can be substituted for

² Actually the g.d.o. will be tuned to a frequency that is lower and possibly a great deal lower than that indicated by the instrument dial.

Bottom view of the circuit board. Clockwise around the board from the lower right corner of the photograph are L_8 , L_7 , L_6 , L_4 , L_3 , L_2 and L_5 . L_1 , the only toroid not shown in this view, is mounted on the band switch.



homemade forms.³ With these types, the number of turns per coil will be approximately the same as given in the parts list.

Alignment

Three things are required to align the receiver: a grid-dip meter, a general-coverage receiver and a lot of patience. All toroids should be grid dipped after they are wired in place, but prior to being glued to the circuit board. Turns should be removed or added until a dip at the required frequency is obtained. L_1 and L_4 should be checked with their respective fixed capacitors in place; L_2 , L_3 , L_6 , L_7 and L_8 should be dipped with their respective trimmers set at half capacitance or greater; and L_5 should be adjusted as described below. Link coupling is used between the g.d.o. and the toroid being checked. Sufficient coupling can be provided by connecting a two-turn link around the toroid to a one-turn link around the g.d.o. coil.

The v.f.o. can be aligned by tuning the general-coverage receiver to 5.4 Mc., setting C_4 and C_6 at minimum capacitance, and tuning C_5 until resonance is achieved. Then C_6 can be varied and the frequency range of the v.f.o. checked. If the range is too small, C_5 should be decreased in capacitance and more turns added to L_5 to bring the circuit back to resonance. If the tuning range is too large, increase C_5 and remove turns from L_5 . Temperature compensation is a matter of increasing the value of the N300 ceramic trimmer, C_4 , while decreasing the value of C_5 to retain calibration until drift is minimized.⁴ I didn't bother much with

this, as the receiver drift is very slight without compensation.

For alignment purposes, the g.d.o. is used as a signal source first at 9 Mc. and later at 14 and 4 Mc. Provided the b.f.o. is working, it should be possible to hear a strong note at all these frequencies. Except for the filter and b.f.o. alignment, it's not important that the crystal frequencies be exact. In regard to choosing the crystals, it is best to obtain a few 27.005- and 27.015-Mc. units from a CB enthusiast or a friendly service shop. A number of 27.005-Mc. crystals should be tried and interchanged between the filter and the b.f.o. until good-sounding ungarbled sideband signals can be tuned in on the 20-meter band. Several crystals should be tried in the filter, and various values of capacitance (1 to 15 pf.) should be used at C_8 until a good passband for sideband signals is achieved. I know that these cut-and-try methods will make most engineers shudder, but these ways seem to work for me.

Before putting the receiver into regular station use, some means should be found for protecting the mixer transistor from strong r.f. fields. Some possible methods were discussed in *QST* recently.⁵

Performance

Considering its simplicity and low cost, the SSSB Mark I does a good job of receiving DX here in Quebec. Its sensitivity and image rejection aren't quite as good as a set that has an r.f. stage, but the receiver pulls in all but the weakest s.s.b. signals.⁶ What more could be asked for the time and money involved?

³ Indiana General Corp., Electronics Div., Keasbey, N.J. 08832. Address all correspondence to Termag Corp., 88-06 Van Wyck Expressway, Jamaica, N.Y. 11418, authorized distributors for IGC. Ask for price bulletin No. 101, bulletin No. 101A, and data sheets for Q1, Q2 and Q3 Ferramic materials.

⁴ For C_1 to be effective, the other capacitors in the circuit should be low-temperature-coefficient types.

⁵ Emerson, "Save Those Transistors!" *QST*, October, 1967.

⁶ One simple way to improve the image rejection of the receiver is to use higher C input circuits at L_2 and L_3 .

» This transmitting converter provides up to 120 watts p.e.p. input for operation in the 6-meter band. It can take its driving signal from any low-power 28-MHz. c.w. or s.s.b. exciter, and requires less than 5 watts of excitation to develop full input power. Output from the transmitting converter is fed directly into the antenna system, or it can be used to drive a high-power linear amplifier.

A Transmitting Converter for 50 MHz

H. GORDON DOUGLAS, W8PMK

More and more v.h.f. enthusiasts are turning to s.s.b. as a means for extending their communications range over that achieved with a.m. Many operators already have an h.f. band s.s.b. transmitter in the station lineup. A practical solution to getting on v.h.f. s.s.b. is to build a transmitting converter and use the h.f. s.s.b. transmitter as the heart of the system. Because a Heath HX-10 s.s.b. transmitter was in use at W8PMK, this approach was taken to get on 6- and 2-meter s.s.b. The equipment described has seen daily use for more than a year. It has provided reliable service, with a minimum amount of normal maintenance. Similar units are shown in the first photograph: one for 50 and one for 144 MHz.

The Circuit

Referring to Fig. 1, the triode section of V_{1A} is used as an overtone oscillator on 22 MHz. It operates at low power and has good frequency stability. Output from the oscillator is amplified by the remaining half of the 6EA8, which operates at a low power level by virtue of rather high values of series resistance in the plate and screen leads. The values were chosen experimentally to provide only enough 22-MHz. drive to excite the mixer, V_2 . Too much drive results in objectionable intermodulation distortion in the mixer, a common cause of poor signal quality in s.s.b. transmitters. Bandpass coupling is used between the mixer and the driver, V_3 , to help reduce spurious responses from the 12BY7A. S.s.b. or c.w. energy is supplied to the mixer cathode across a 100-ohm resistor which provides cathode bias for V_2 while serving as a load for the s.s.b. exciter, connected to J_1 .

The driver stage, V_3 , is operated Class AB and gets its bias from a 120-ohm cathode resistor. It is neutralized by means of C_1 , a 1½-inch length of No. 16 wire. This is soldered to the low side of L_4 , and its free end is brought through a hole in the chassis and placed adjacent to the tube envelope, parallel to its plate. Neutralization is adjusted by moving the wire to or from the tube, as needed, until stability is



Front view of the transmitting converter. The equipment described in this article is seen at the left.

achieved. A metal shield divides the plate side of the driver from its input tuned circuit, and aids in stabilizing the stage.

Link coupling, L_6 and L_7 , is used between the driver and the grid circuit of the final amplifier. Neutralization is again used (C_4 and C_5) to assure good stability. C_4 and C_5 are 3-inch lengths of stiff wire, cross-connected as indicated in Fig. 1. The stubs are placed adjacent to the 6146B tube envelopes, using the technique described for the driver. Parasitic chokes Z_1 and Z_2 are installed at the plate terminals of V_4 and V_5 as a further measure of assurance against spurious output. All power leads going into the driver-plate/p.a.-grid compartment are decoupled by means of feedthrough capacitors.

Construction Hints

An 8 × 12 × 2½-inch steel chassis is used as a foundation for this equipment. An aluminum chassis could have been used, and it is quite likely that a much smaller chassis would have served, had it been available at the time the project was started.

A homemade panel, measuring 8 by 10 inches, was cut from ⅛-inch thick aluminum stock and mounted on the front end of the chassis. Another plate of the same dimensions was attached to the rear apron of the chassis for use as a back. Perforated aluminum was then formed as a cover for the assembly. It is attached to the sides of the chassis by means of sheet-metal screws. An aluminum bottom plate is used on

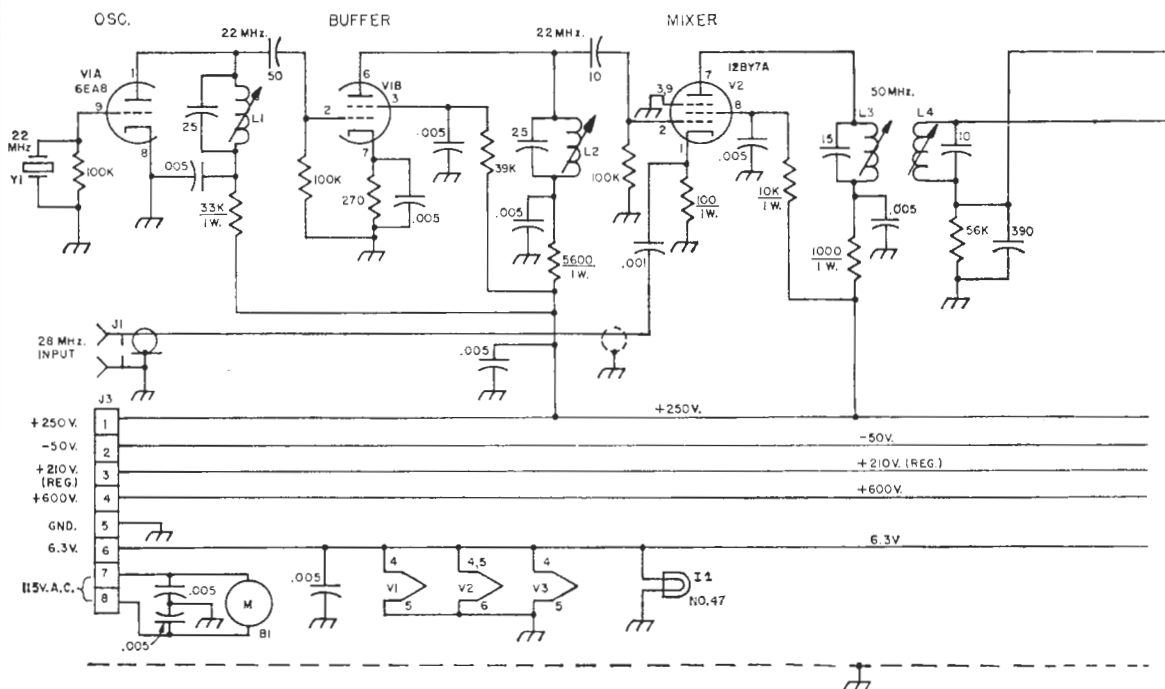


FIG. 1—Schematic of the 6-meter transmitting converter. Fixed-value capacitors are disk ceramic. FT = Feed-through type. Fixed-value resistors are 1/2-watt composition unless noted differently.

B₁—Medium- or high-speed cooling fan.

C₁, C₄, C₅—Neutralization capacitors (see text).

C₂—20-pf. miniature variable (E. F. Johnson 160-110 suitable).

C₃, C₆—25-pf-per-section butterfly (E. F. Johnson 167-22 suitable).

C₇—50-pf. miniature variable (E. F. Johnson 148-4 suitable).

I₁—Panel indicator, No. 47 lamp.

J₁—Phono jack.

J₂—BNC-style chassis connector.

J₃—8-circuit male chassis-mount connector.

the chassis to help confine the r.f. and to prevent accidental contact with the operating voltages of the unit.

V₁, V₂, and the grid circuit of V₃ are on one side of the bottom of the chassis, and are isolated from the plate circuit of V₃, and the grid circuit of the p.a. stage, by means of a flashing-copper divider. The feedthrough capacitors for power-lead decoupling are mounted on the copper partition.

All resistors and capacitor leads are kept as short as practical, and ground connections are made directly to the chassis. Keep all ground leads short. Dress all power leads along the chassis and away from the r.f. leads of the equipment. The signal-carrying leads should be as short and direct as the builder can make them, keeping the input and output circuits of each stage as well isolated from one another as practical.

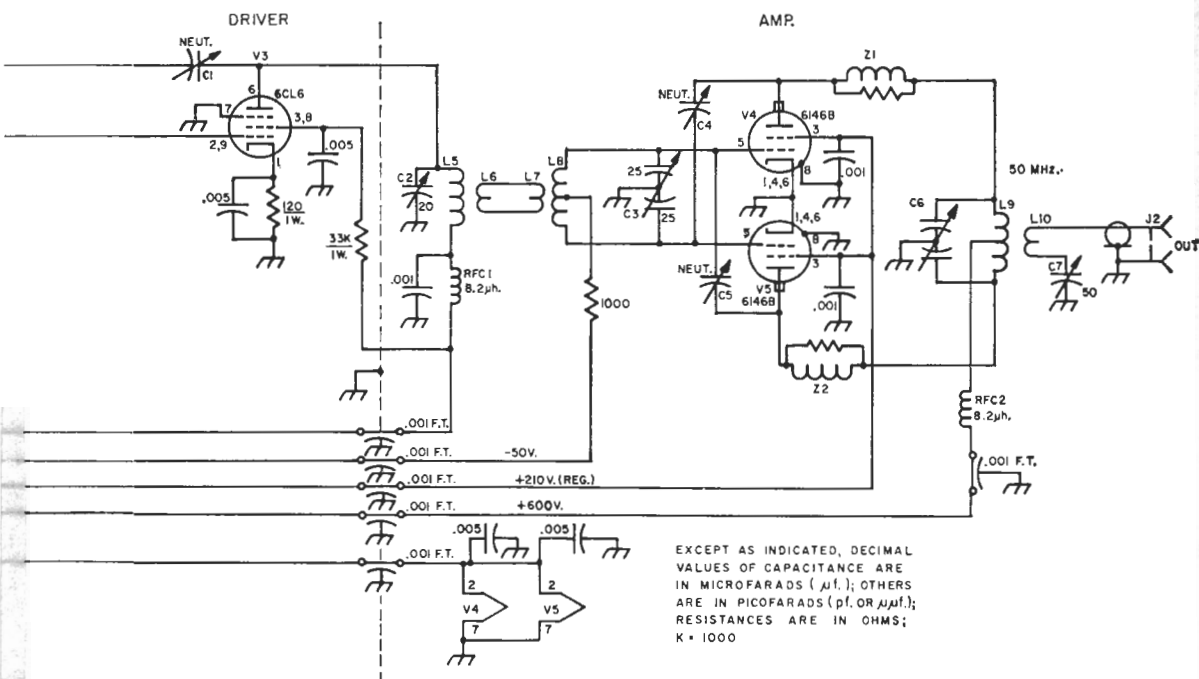
The bandpass-coupling circuit in the mixer plate uses two 3/8-inch diameter slug-tuned coil forms, mounted with their centers 1 inch apart. L₅ and L₈ are wound on short pieces of 1/2-inch plastic rod. Air-wound coils can be substituted if the builder wishes. The coils used by the author are doped with TV high-voltage cement

so that the turns will stay in place. A suitable length of Miniductor coil stock could also be used at L₅ and L₈. L₈ is supported from the grid pins of the 6146B sockets by its leads.

The plate circuit of the output stage is above the chassis. L₉ is supported by the stator posts of C₆, which is mounted on the front panel. Its rotor terminal is grounded to the chassis by means of 3/8-inch wide copper strap. L₁₀ is supported by a 1-inch high ceramic cone insulator, and by the stator post of C₇. The loading capacitor, C₇, is mounted at right angles to C₆ on a small aluminum L bracket. It is adjusted from outside the case, through one of the holes in the cover. RFC₂ is mounted between the center tap on L₉ and a 0.001-μf. feedthrough capacitor terminal. The latter is used to bring the plate voltage through the chassis. A high-speed cooling fan blows an air stream across the 6146Bs to help assure longer tube life.

Tune-Up and Operation

The operating voltages for this unit can be taken from the h.f. s.s.b. rig if desired, or a separate power supply can be built for this purpose. In the writer's installation, the voltages are taken from the HX-10. The HX-10's meter-



L_1, L_2 —1.7- to 2.7- μ h. adjustable inductor (J. W. Miller 4503).

L_3, L_4 —0.99- to 1.5- μ h. adjustable inductor (J. W. Miller 4403).

L_5 —8 turns No. 16, $\frac{1}{2}$ -inch diam., 2 inches long, center tapped (see text).

L_6, L_7 —2 turns insulated hookup wire; L_6 over B-plus end of L_5 , L_7 over center of L_5 .

L_8 —10 turns No. 18 enam., center-tapped, $\frac{1}{2}$ -inch diam., $\frac{3}{4}$ -inch long, on ceramic form.

L_9 —8 turns No. 12, $\frac{3}{4}$ -inch diam., 3 inches long, center-tapped.

L_{10} —2 turns No. 16 enam., $\frac{1}{2}$ -inch diam. Insert in center of L_9 (see photo).

RFC_1, RFC_2 —8.2- μ h. r.f. choke (Millen 34300-8.2 or equivalent).

Y_1 —22-MHz. overtone crystal (International Crystal Co.).

Z_1, Z_2 —Parasitic suppressor. 5 turns No. 18 wire over body of 56-ohm, 1-watt carbon resistor.

ing system is used also, permitting monitoring of the plate current for V_4 and V_5 , and for reading relative r.f. output from the transmitting converter. Details are not given here since most builders will no doubt prefer to install a 300-ma. d.c. meter in the plate line to the 6146s, making it a permanent part of the assembly. Relative r.f. output can be observed on the station s.w.r. bridge.

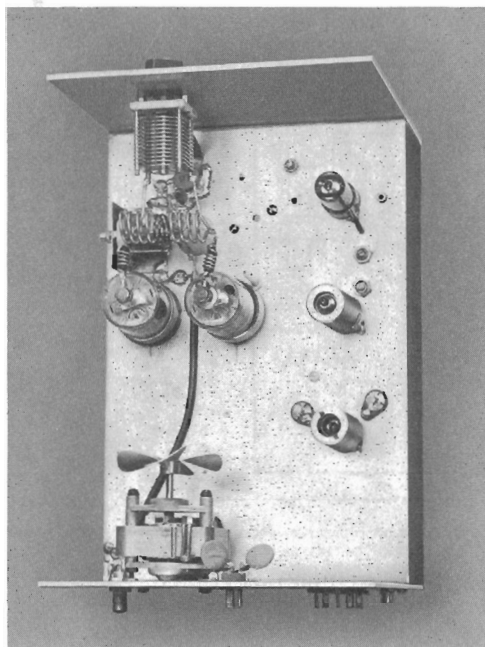
Initial adjustments to the equipment should be made with the operating voltages removed from V_3 , V_4 , and V_5 . This will prevent damage to those tubes should instability be present. Start by coupling a wavemeter to L_1 and tuning its slug until the crystal oscillates, as indicated by the wavemeter. Turn the slug of L_1 two or three additional revolutions outward (slug moving out of coil). This will help the oscillator to start readily each time. Next, adjust L_2 for peak response, still using the wavemeter.

It will be necessary to reduce the output from the 28-MHz. s.s.b. exciter before conducting the remainder of the tests. This can best be done by adding a toggle switch to the rear

apron of the exciter and wiring it in series with the screen supply to the amplifier stage of the exciter. Plenty of drive will be available from most exciters, with screen voltage removed in this way. For h.f. operation, merely restore the screen voltage. Alternatively, the output stage of the exciter can be disabled and link coupling used between the exciter's driver stage and J_1 of the transmitting converter. No more than 5 watts of c.w. or s.s.b. drive will be needed, and quite likely the actual driving power will be on the order of 2 to 3 watts.

The next step is to apply drive at J_1 and adjust L_3 and L_4 for peak output at 50 MHz., still using a wavemeter as an indicator. After this has been done, apply operating voltage to V_3 and tune its output tank for peak indication on the wavemeter, also at 50 MHz. If output on frequencies other than 50 MHz. is evident when using the wavemeter (check from 25 to at least 150 MHz.), check for instability in the driver stage, adjusting C_1 closer to or farther away from the tube envelope for optimum neutralization.

Drive and operating voltages can next be



Looking at the top side of the chassis, the amplifier section is at the upper left, with the cooling fan below. V_1 , V_2 , and V_3 are in a row along the right side of the chassis.

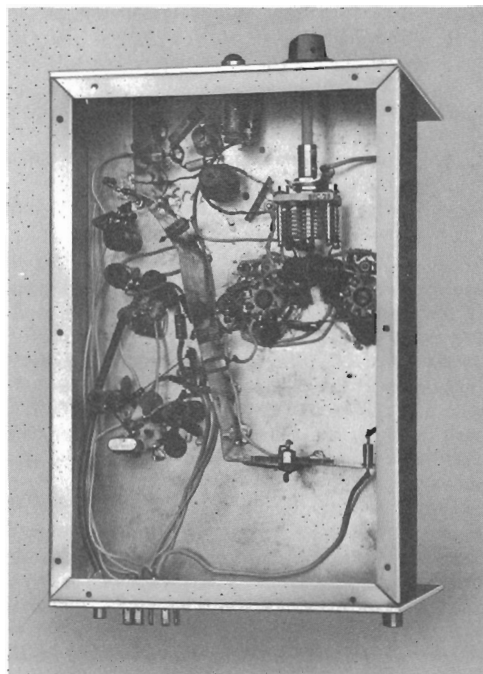
applied to the 6146Bs. A 50-ohm dummy load should be connected to J_2 during this test. A power-output indicator can be installed between J_2 and the dummy to facilitate tuning for maximum r.f. output. Assuming that a plate-current meter is in the 600-volt line to the 6146B plates, apply sufficient carrier from the 28-MHz. exciter to cause a rise in plate current from the resting value—approximately 50 ma.—to approximately 75 ma. Tune C_3 for maximum indicated plate current, then tune C_6 for a dip in plate current, and for maximum output. Next, adjust L_2 , L_3 , L_4 and C_2 for maximum drive to the 6146Bs. As C_6 is tuned through its range there should be no sudden changes in p.a. plate current; just the normal dip, if the amplifier is stable. If instability is noted, adjust C_4 and C_5 (a little at a time) until the amplifier is neutralized. It may be necessary to repeak C_3 after the foregoing adjustments.

C_7 is the loading capacitor. It should be adjusted so that the plate-current dip of the amplifier is rather broad, indicating tight coupling to the load. This condition is mandatory if a high-quality s.s.b. signal is to be had. Once the initial tune-up is completed, the 28-MHz. drive can be increased until the plate current (at dip) is 200 ma., the correct value for 120 watts c.w. input. During s.s.b. operation the plate current should not be allowed to exceed this value. Best linearity will be obtained only if

the s.s.b. drive is held below the point where the 6146Bs begin to draw grid current. For lower-power operation, merely reduce the level of drive from the 28-MHz. exciter. A.m. operation is possible with this equipment, if the 28-MHz. exciter has a.m. capability. During a.m. operation do not let the plate current of the 6146Bs exceed approximately 75 to 100 ma. if the maximum AB₁ ratings are to be observed. Too much plate current will damage the tubes. Provisions should be made externally to cut off the screen voltages from V_3 , P_4 and V_5 during receive. A relay can be used for this purpose.

Results

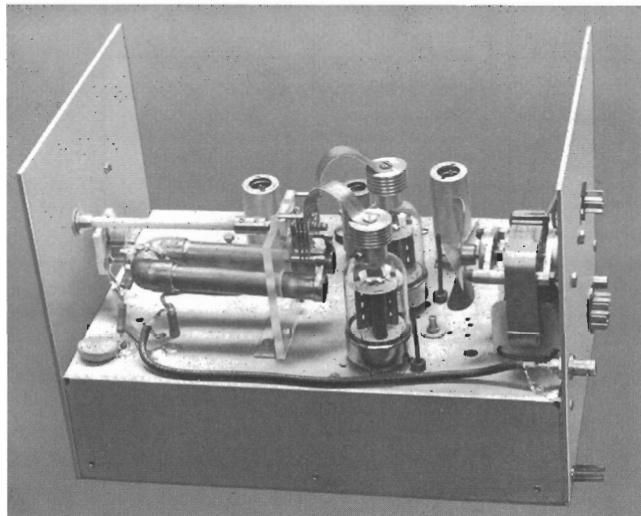
This equipment has been very effective in working DX on s.s.b. and c.w. For local operation each evening, often in the a.m. mode, the power level is kept down to less than 10 watts input—an aid to TVI reduction which is quite worth while. No TVI other than fundamental overloading has been experienced. This type of TVI can be encountered on the lower TV channels with *any* 6-meter rig. A 6-meter Transmatch is used between the transmitter and the feed line to help reduce harmonics and other spurious radiation from the antenna. Acknowledgement is given here to WA8COT for his help in designing and testing this equipment.



Looking into the bottom of the chassis, V_1 , V_2 , V_3 , and their associated circuits are on the left side of the copper shield which divides the chassis. The plate circuit of the driver, V_3 , upper center, and the grid circuit of the amplifier, right, are on the same side of the partition.

» Designed to be driven by a low-power 28-MHz. s.s.b./c.w. exciter and requires less than 5 watts 28-MHz. energy to develop its rated input power, approximately 100 watts. If conventional v.h.f. construction techniques are followed, the builder should have no difficulty in putting this equipment into s.s.b. or c.w. service on 144 MHz.

A Medium-Power Transmitting Converter for 144 MHz



Top-chassis view of the 2-meter transmitting converter. The amplifier plate lines are in the left foreground. The oscillator section, mixer and driver are in a line along the far side of the chassis.

H. GORDON DOUGLAS, W8PMK

Considerable similarity can be seen between this transmitting converter and the 6-meter model described on pg. 207. The tube lineup is the same with the exception of the mixer stage, which in this instance is a 6GM6. Use of this tube instead of a 12BY7A was based on its lower output capacitance—a factor when designing tuned circuits for 144 MHz. The basic circuit is similar to one used by Hall, K1PRB,¹ in the transmitting section of his transverter. The main difference is in the amplifier section. Hall used an 829B. The writer uses push-pull 6146Bs, duplicating an amplifier described in QST.² Operating voltages can be taken from the h.f. s.s.b. exciter, or a separate power supply can be used. A Heath HX-10 supplies the driving signal and operating voltages for the converter at this station.

Circuit Information

A 58-MHz. overtone crystal is used in the oscillator circuit at V_{1A} . The high crystal frequency minimizes the chance for spurious responses, which often result when higher orders

of multiplication are used in the oscillator chain. V_{1B} operates as a double and supplies 116-MHz. injection to the mixer, V_2 . The 28-MHz. driving signal is injected into the mixer cathode by means of tuned circuit L_4 .

The driver stage is neutralized by means of C_3 , a 1-inch length of hookup wire. This is soldered to Pin 2 of the tube socket, and the free end of the wire is placed near the coil lead connected to Pin 6 of V_3 . The spacing is varied until the stage is stabilized. This may not be needed and should be used only if instability is observed in the driver. The driver operates Class A and gets its bias from a 120-ohm cathode resistor.

Bandpass coupling is used between the driver plate and the input of the amplifier. L_7 and L_8 are mounted parallel to one another, $\frac{1}{8}$ inch apart. The screen voltage is reduced from that used in the 6-meter model. This was done to make it easier to derate the tubes at 144 MHz., as recommended by the manufacturer. Input should be reduced about 28 percent to assure their normal life span at this frequency.

Construction

The same chassis, panel, and cover dimensions as in the 6-meter version are used here. A copper partition separates the driver plate and grid

From March, 1969, QST

¹ Hall, "A 100-Watt 2-Meter Transmit-Receive Converter," QST, January 1966.

² DeMaw, "An All-Mode Amplifier for 2 Meters," QST, September 1966.

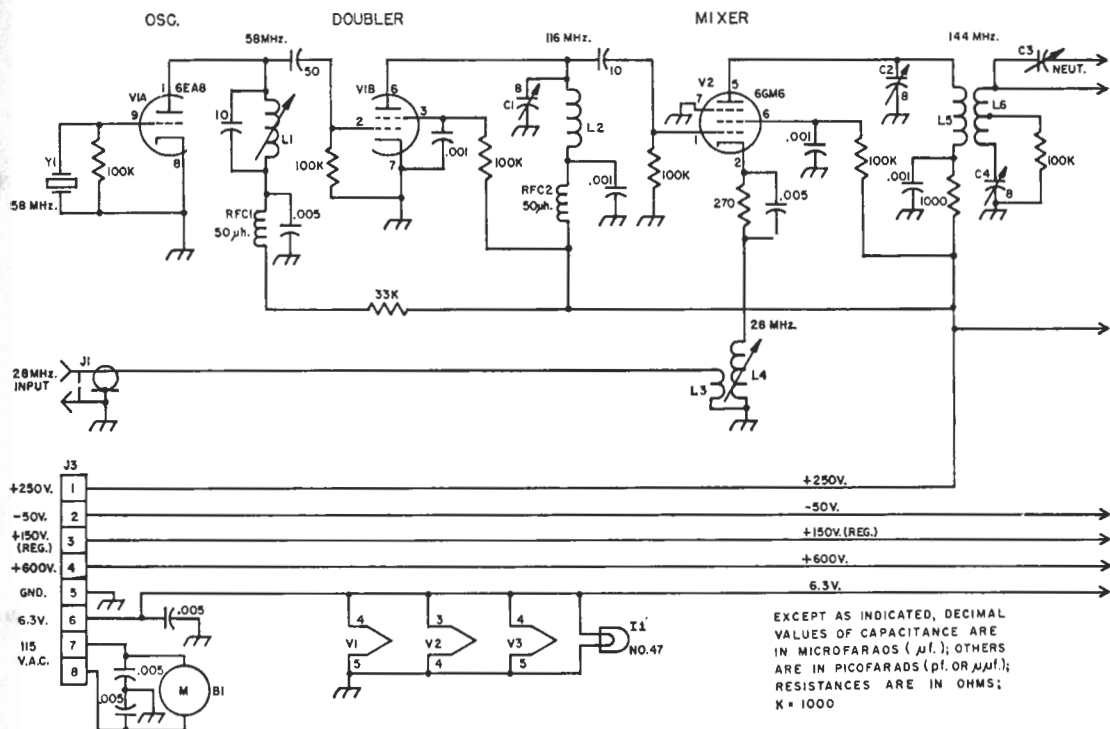


FIG. 1—Schematic of the W8PMK transmitting converter. Fixed-value capacitors are disk ceramic. FT = Feed-through. BM = Button Mica. Resistors are $\frac{1}{2}$ -watt carbon unless otherwise noted.

B₁—Medium- or high-speed tube-cooling fan.
 C₁, C₂, C₄—8-pf. plastic tubular variable.
 C₃, C₆, C₇—Neutralizing capacitor (see text).
 C₅—10-pf. miniature variable (Johnson 160-104 suitable).
 C₈—Butterfly-type variable (Johnson 167-22 with two rotor and three stator plates removed from each section).

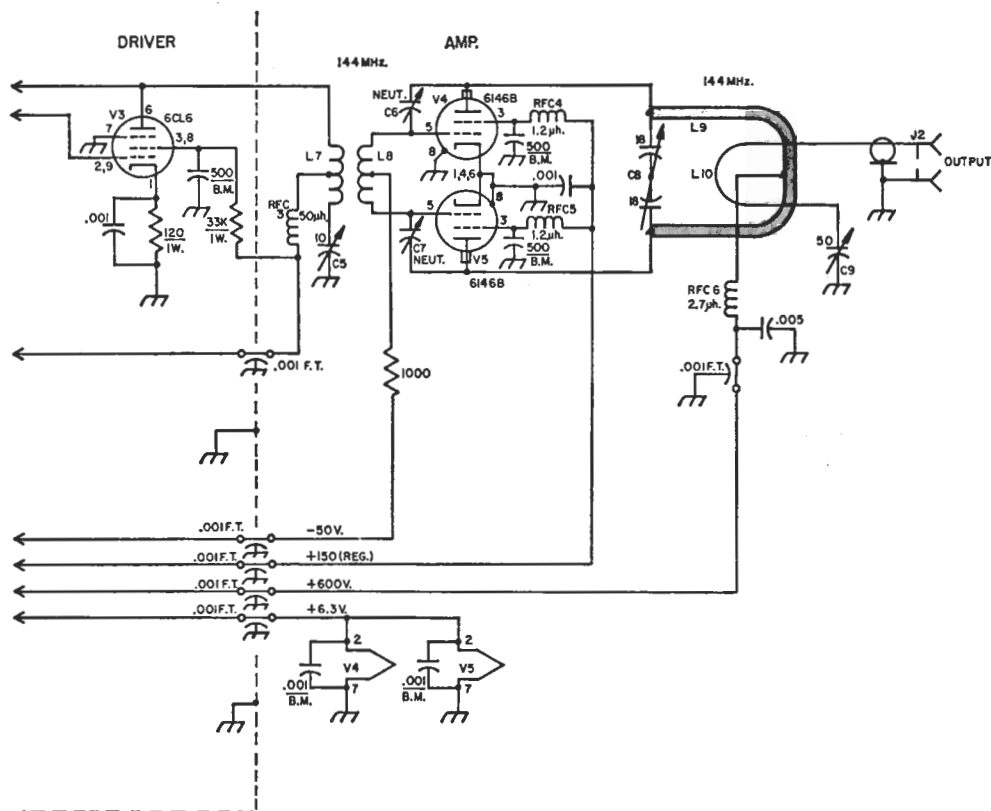
C₉—50-pf. miniature variable (Johnson 167-3 suitable).
 I₁—Panel indicator, No. 47 lamp.
 J₁—Phono jack.
 J₂—BNC-type chassis-mount connector.
 L₁—8 turns No. 26 enam., close-wound on $\frac{1}{4}$ -inch slug-tuned form (Miller 4500-4 form suitable).
 L₂—4 turns No. 20, $\frac{3}{8}$ -inch diam., $\frac{3}{4}$ inch long.



Bottom view—The oscillator, doubler, mixer and driver stages are below the copper partition which divides the chassis. The driver plate and amplifier grid coils are parallel to each other, just to the right of the amplifier tube sockets.

circuits to aid stability of the driver and amplifier. Operating voltages for the driver plate circuit, and for the entire amplifier section, are decoupled by means of 0.001- μ f. feedthrough capacitors. The capacitors are mounted on the copper divider (see photo).

Unlike the 6-meter version, this unit uses copper lines for the amplifier plate circuit. The large-diameter tubing provides a higher tank-circuit Q, and efficiency. The rotor of C₈ is not returned to ground. This practice aids amplifier balance and stability at 144 MHz. C₈ is mounted on a Plexiglas plate. A piece of $\frac{1}{4}$ -inch diameter insulating rod is used as a tuning shaft, to isolate the rotor from ground. Forced-air cooling helps prolong the life span of the tubes. A small fan is mounted near the tubes and can be seen in the photo. Heat-dissipating plate connectors offer additional help in this regard.



L_3 —2 turns insulated hookup wire over ground end of L_4 .

L_4 —1.6-3.1- μ h. variable inductor; 13 turns No. 26 enam. on $\frac{3}{8}$ -inch iron-slug form.

L_5 —5 turns No. 20, $\frac{3}{8}$ -inch diam., $\frac{3}{4}$ inch long.

L_6 —6 turns No. 20, insulated, $\frac{1}{2}$ -inch diam., $\frac{3}{4}$ inch long, center-tapped. Mount with C_4 end against B-plus end of L_5 .

L_7 —5 turns No. 14 enam., $\frac{3}{4}$ -inch diam., $1\frac{1}{4}$ inches long, center-tapped.

L_8 —5 turns No. 14 insulated wire (or $\frac{1}{8}$ -inch wide copper strip), $\frac{3}{8}$ -inch diam., 2 inches long,

center-tapped. Adjust turn spacing for maximum grid drive.

L_9 —See footnote 2.

L_{10} —See footnote 2.

RFC_1 — RFC_3 , inc.—50 μ h. r.f. choke (Millen 34300-50 suitable).

RFC_4 , RFC_5 —1.2- μ h. r.f. choke (Millen 34300-1.2 suitable).

RFC_6 —2.7- μ h. r.f. choke (Millen 34300-2.7 suitable).

Y_1 —58-MHz. overtone crystal (International Crystal Co.).

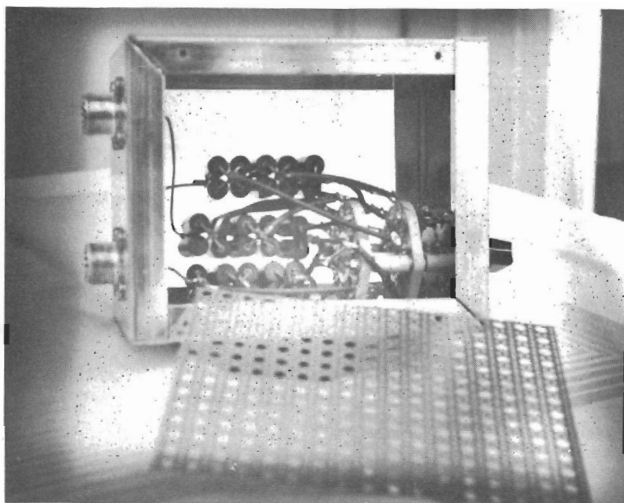
Testing and Use

Rough alignment of the tuned circuits can be done with the help of a grid-dip oscillator. Final touching up of the various stages can be carried out by observing the amplifier plate current and adjusting all of the tuned circuits (except the amplifier tank) for maximum input. This should be done with reduced drive from the 28-MHz. exciter so that plate current will not be excessive. Adjustment of the amplifier neutralization wires is the same as in the 6-meter model, though it should be noted that the wires are not cross-connected for conventional neutralization. The technique used here may be necessary when 6146-type tubes are used at frequencies above approximately 70 MHz., their normal self-neutralizing frequency. The same is true of the driver stage, V_3 .

Because of the derating factor mentioned earlier, the plate current of this amplifier should be limited to 150 ma. or less at dip. Tight coupling to the load is necessary during s.s.b. operation to assure a quality signal. The dip in plate current should be quite shallow, 10 ma. or so, at resonance. The loading capacitor, C_9 , should be adjusted to provide the foregoing condition. No more than two or three watts of 28-MHz. drive should be needed to develop the specified input power to the amplifier.

Provisions should be made for cutting off V_3 , V_4 , and V_5 during receive. This can be done by opening their screen-supply leads with a set of relay contacts. Other methods, such as fixed bias cutoff, can also be used. We'll leave that part of the project up to the constructor.

» The attenuator described in this article provides an answer to a common problem—what to do with the excess driver power when a 50-to-100-watt exciter is used for exciting a grid-driven tetrode linear amplifier.



The step attenuator is assembled in a standard 3×4×5-inch aluminum box fitted with perforated aluminum covers.

(Photos by K9BJA)

A Step-Type R.F. Attenuator

EUGENE A. HUBBELL, W9ERU

It came as a distinct shock to the author to find that there are times when it is necessary to throw away some of that precious r.f. energy from a transmitter. This realization came about with the advent of s.s.b. in the shack.

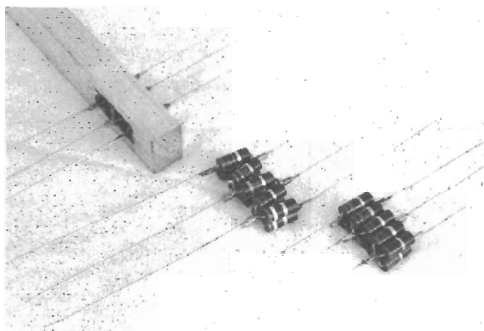
The Pacemaker sideband exciter has a normal output of 60 to 70 watts, as indicated by an M. C. Jones Micro-Match s.w.r. bridge. The Johnson Kilowatt, operating in Class AB₁, requires only 2 to 3 watts of driving power. It is not recommended that the Pacemaker be operated at lower than full input, because the signal-to-noise and signal-to-carrier ratios will suffer. The problem, therefore, was how to drop the full output of the Pacemaker to the necessary drive level for the Kilowatt.

A 50-watt noninductive dummy load resistor was used across the grid-input line when driving the amplifier on a.m. and c.w. With this dummy load in the circuit, the Pacemaker drive was still far in excess of what was needed. Apparently, some sort of variable attenuator would be nice, but nothing practical which would present a constant load to the Pacemaker turned up in any of the various handbooks. So a step-type attenuator seemed the practical solution.

A suitable formula for H, T, and L pads was found in Nilson & Hornug's *Practical Radio Communications*. The T pad was chosen, mostly because the resulting unit would be usable in either direction without regard to how it was hooked up. The first design was for one with five steps of 0, 3, 6, 9 and 12-db. attenuation. It was built and found to operate satisfactorily, except that it did not provide quite enough attenuation on 75-meter phone. So a redesigned unit was built, adding another step which brought the maximum attenuation to 15 db. The description to follow is of this unit.

Circuit

The basic T-pad configuration is shown in the inset of Fig. 1, while the main diagram shows the practical circuit used in the construction of the unit. A resistance of 50 ohms was chosen for input and output impedances because results would be easy to check with the 50-ohm Micro-Match, the Pacemaker would operate into this load, and the grid-input circuit could be swamped down to this value with the dummy load mentioned above. With the dummy load next to the Pacemaker and the step-type attenuator on the grid input, the dummy load current does not pass through the attenuator. Values of resistance for the three



Wooden jig and "bundles" of resistors.

that they touch, but do not overlap. Remove the wood left between the holes, and saw a slot in each end of the hole pattern, between the two holes at the end. Insert the ten resistors making up the group designated R_1 in Fig. 1. Clamp the wooden jig on the resistors with a C clamp, so they will not slip. Cut off the long leads to about $\frac{1}{2}$ inch, and bend into bundles of four leads, except for the end pairs, where the leads will be of two each. Into each bundle of four insert the end of a 6-inch length of No. 16 tinned wire and solder securely. Also solder 6-inch length of this wire to the paired leads at the end. Note the finished assemblies shown in the photograph, and make up similar bundles for groups R_2 and R_3 .

Switch Assembly

The resistor bundles may now be soldered to the switch, and this can be done outside the box, if a little care is used to see that the result will not interfere with the box sides. The two bundles forming R_1 and R_2 will extend directly back of the switch, and be soldered to the two-pole, five- (or six-) position section. (The confusion as to identification of this section is because we are making use of an "off" position not considered as usable by the manufacturer.) The bundle of resistors making up R_3 is placed just above the switch, and soldered to the appropriate terminals of the single-pole, six-position section. A jumper connects the terminals of the R_1 and R_2 groups together, and the latter to the two rotor connections on the two-pole section and also to the rotor on the single-pole section. Then the whole assembly is inserted in the box and the switch secured with its nut. The two remaining terminals of R_1 and R_2 are soldered to the two coax jacks, and the remaining terminal of R_3 is grounded to a lug on one of the bolts securing one of these coax jacks. The circuitry calls for leads between the

hot terminals of the two coax jacks and switch points, and these are added last.

Testing

The wiring may be checked by an ohmmeter. Put a 50-ohm resistor across one coax jack, and the ohmmeter across the other. Rotating the switch should show very little change in the ohmmeter reading at any switch position, and the reading should be just about 50 ohms. Checking with a Micro-Match and a good dummy load will show a barely perceptible increase in s.w.r. when the attenuator is added to the circuit. Placing the Micro-Match between the dummy load and the attenuator, the following output readings were obtained:

Step	10-Watt Input	100-Watt Input
0 db.	10 watts	100 watts
3 db.	5.5 watts	46 watts
6 db.	2.9 watts	21 watts
9 db.	1.5 watts	10 watts
12 db.	0.85 watt	5 watts
15 db.	0.5 watt	3 watts

The strict accuracy of these readings is somewhat doubtful because of the difficulty in reading the Micro-Match accurately. The scales vary considerably between the 100-watt and 10-watt levels, and the power level may vary considerably from the nominal value. The results obtained agree very well with calculated values, considering the fact that the resistors are not exactly what is needed, and are of ten-per-cent tolerance.

Using the Attenuator

In use, the attenuator does all that is required. The Pacemaker is tuned up with the attenuator in the zero db. position, but with the 50-ohm dummy load on the output, and the grid circuit of the Kilowatt detuned. When the Pacemaker is properly loaded, the grid circuit is tuned to resonance and enough attenuation introduced to prevent overloading in the Class C mode of operation (20-ma. grid current). The plate circuit is tuned to resonance and properly loaded for Class C operation. Then the mode switch is turned to s.s.b., which places the amplifier in Class AB₁, and the attenuator is set to the position where s.s.b. modulation peaks show only a slight indication of grid current on the final-amplifier grids. It is even handy for a.m. and c.w. work when the drive runs too high at a given setting of the 32V-1 output circuits. No trouble has been experienced with heating of the resistors, as long as the drive is not left on continuously over a minute or so. Try one; it works!

» Such commonly used parameters as peak envelope power (p.e.p.) and instantaneous peak power are defined and described in this text. A description of a speech model (peak-to-average level) in relation to the s.s.b. voice envelope is also offered. Because of the vast amount of published material available on this subject, only a cursory compilation of facts and data can be presented.

Ordinary and Processed Speech in SSB Application

HAROLD G. COLLINS, W6JES

Because of the high-amplitude peaks in the voice waveform, the natural, unprocessed voice does not provide maximum utilization of an s.s.b. suppressed-carrier communications system. It is quite apparent that many off-the-shelf processing units are becoming available for use with existing transmitters. These units, in one way or another, claim to increase talk power of the s.s.b. transmitter by increasing the average-to-peak levels in the modulation signal and hence in the r.f. envelope. With the apparent increase of interest in the subject, some of the more-simple techniques that are presently employed or are readily applicable are summarized here.

The Fundamental S.S.B. R.F. Envelope

The power developing capability of a given transmitter in single-sideband service depends upon the nature of the signal waveform being transmitted. The most fundamental s.s.b. waveform is generated from a single-frequency audio sine-wave tone. This single tone is translated in the s.s.b. generator to produce a single radio-frequency output.

The term that describes power in an s.s.b. r.f. envelope is *peak envelope power* or simply p.e.p. It is defined as "the average power of the highest amplitude signal measured over one r.f. cycle," or "the mean power developed at the crest of the modulation envelope," in other words, the keydown power developed within allowable distortion limits.

As an example, for a given transmitter with four 811As, the maximum single-tone operating level occurs when plate current reaches 700–800 ma. With this value of plate current and a plate voltage of 1600 volts, the r.f. power output is nearly 800 watts with a plate power input of 1250 watts. The r.f. output power is the mean power value (p.e.p.) previously stated. With 800 watts into a 50-ohm load, an r.f. current I_{rf} of 4 amp. r.m.s. will flow. An r.f. voltage E_{rf} of 200 volts r.m.s. is developed across the load.

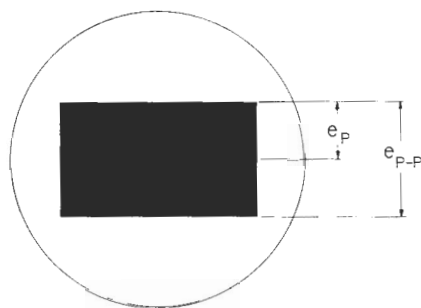


FIG. 1—S.s.b. envelope using single-tone modulation, as displayed on an oscilloscope. Power in the envelope (assuming a resistive load) is a time average value, over one r.f. cycle, and is equal to $e_p^2/2R_{load}$. P.e.p.-to-average ratio is 0 db.

If one monitors the above r.f. envelope on an oscilloscope connected across the 50-ohm load, the instantaneous r.f. peak-to-peak voltage waveform of Fig. 1 is observed. One half the envelope height is the instantaneous peak envelope voltage; that is, $\sqrt{2} E_{rf}$. In the 811A amplifier above, if $E_{rf} = 200$ volts when obviously the peak voltage is $\sqrt{2} \times 200$. A value of *instantaneous r.f. peak power* can be calculated which is

$$\frac{(\sqrt{2} \times 200)^2}{50} = \frac{2 \times 40000}{50} = 1600 \text{ watts}$$

The instantaneous r.f. peak power is 2 times p.e.p. or 3 db. greater in the s.s.b. envelope.

One quantity of interest remains, the transmitter p.e.p. input power during single-tone modulation. It is simply the d.c. plate voltage \times the d.c. plate current, or in the above example 1600 volts \times 0.780 amp. or 1250 watts. The basic reference values are thus established.

Speech and the S.S.B. Voice Envelope

The ratio of instantaneous peak-to-average power in speech is somewhat dependent upon the percent of the time that the peak is allowed to be exceeded. For instance, if we can specify

that the real peaks of speech can exceed some lower level that we choose to call the "peak" for 1% of the time, then the ratio of instantaneous peak-to-average speech power is about 14.5 db. This figure is for a continuous discourse. It is the figure used as the statistical peak-to-average power ratio, and it is the value used also in the following discussion as the peak-to-average power ratio for *unprocessed speech*.

Now let us apply audio to an s.s.b. transmitter (with a voice waveform such as you would see on a scope placed at the a.f. input) and modulate the transmitter to its p.e.p. capability, as seen in Fig. 2. We find the *peak-to-average power ratio* measured in the audio signal is essentially the same as the *p.e.p.-to-average power ratio* in the s.s.b. voice envelope. Hence, translating the voice signal to s.s.b. increases the measured voice peak-to-average power ratio by an amount equal to the instantaneous peak-to-average power ratio of the r.f. sine-wave envelope which is 3 db. (see previous section on fundamental r.f. envelope). What this is saying is that a peak-to-average power audio voice waveform of 14.5 db translated to an s.s.b. voice waveform will result in an *instantaneous r.f. peak-to-average power* of 17.5 db. The p.e.p.-to-average power in the r.f. wave remains 14.5 db., a ratio of 28/1.

Employing our above 811A amplifier, again the p.e.p. output is 800 watts. Hence, if we just talk our amplifier to the p.e.p. limit, as would be observed on the r.f. monitor scope by observing instantaneous r.f. peaks, the average r.f. power output P_{out} is 800/28 or nearly 30 watts. This can best be observed by monitoring the r.f. current in a 50-ohm load with a thermocouple r.f. ammeter. It has been assumed in the above discussion that the amplifier supply voltage shows negligible change from minimum-signal to maximum-signal plate current.

The thermocouple r.f. ammeter indicates the r.m.s. value of current for any signal-envelope waveshape. The r.f. current in its varying state with modulation applied should average between $\frac{1}{2}$ to 1 ampere in the above example. Unfortunately, cranking up the mike gain causes linear-amplifier limiting or flat-topping to occur. This does, however, increase the average power output and hence the loudness at the receiving station. Along with this increase in average power comes a broadening of the transmitted bandwidth in the form of distortion products caused by nonlinear operation. In an unprocessed-modulation s.s.b. transmitter, generation of the odd-order products by opening up the mike gain becomes annoying to intolerable in adjacent channels rather rapidly.

Speech Processing

If one could increase the average power in the modulation envelope while effectively limiting the modulation envelope at the p.e.p. capa-

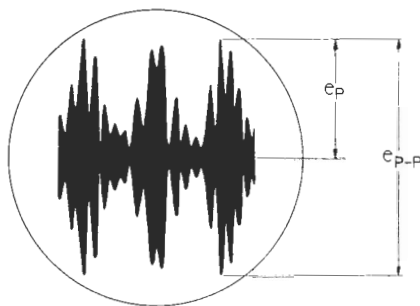


FIG. 2—S.s.b. r.f. envelope using unprocessed voice modulation. P.e.p. is the mean power developed at the crest of the modulation envelope, i.e. $e_p^2/2R_{load}$.

bility, and minimize adjacent channel interference by reducing the number of generated distortion products, then an increase in the transmitter's effectiveness would be obtained. To say it another way, talk power would be increased if the average-to-peak power ratio of the inband modulation envelope was greater.

In order to increase the average-to-peak power ratio, it is necessary to modify the signal waveform. This inevitably leads to distortion of the voice signal, and reduces intelligibility. One measure of intelligibility is the effective received signal-to-noise ratio. If distortion is treated as if it affects intelligibility the same as noise, then the effective received signal-to-noise ratio is

$$\frac{S}{N_{eff}} = \frac{S_{av}}{(N + D)}$$

where S_{av} = The average sideband power at the receiver.

D = Effective noise power produced by the distortion products of the processor.

N = The average noise power from all other sources.

Since S_{av} and D are both functions of the processor limiting level, intelligibility can be maximized by choosing the proper level of limiting for a given level of noise power N . From this we conclude that the best method of speech processing is the one that gives the greatest increase in talk power with the least distortion.

The most useful way of evaluating the benefits of a processor is to observe the improvement in the speech intelligibility threshold in the presence of noise with the same peak power as the unprocessed signal. Intelligibility threshold is defined as the condition where a connected discourse is just barely understandable in the presence of white noise which is limited to the signal bandwidth. Thus intelligibility threshold improvement, or ITI, may be defined as the effective signal-to-noise ratio with processing, to the signal-to-noise ratio without processing. From the above, it then follows that the

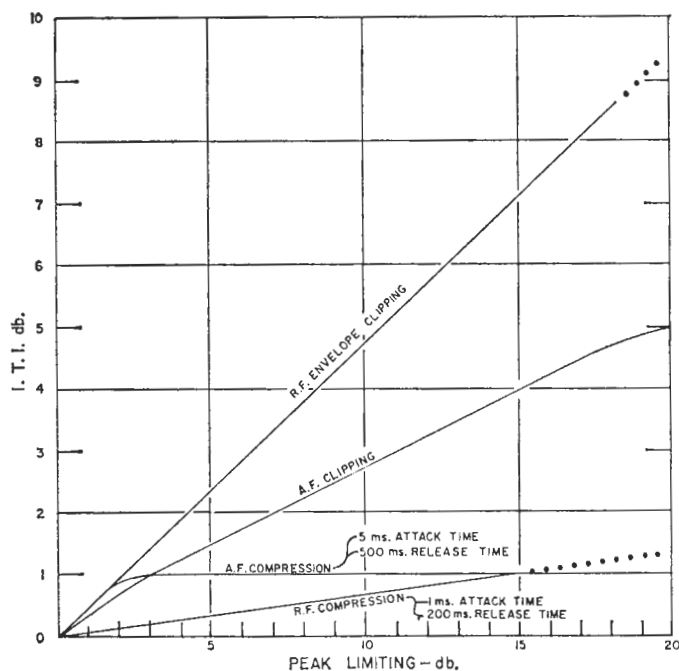


FIG. 3—Comparison of effectiveness of simple speech processing techniques applied to s.s.b. transmission. Typical compression and release times are indicated.

intelligibility threshold improvement, ITI, may be expressed by

$$ITI \text{ (in db.)} = 10 \log_{10} \frac{S_{av}}{S_{oav}} - 10 \log_{10} \frac{(N + D)}{N}$$

where S_{oav} = average sideband power at the receiver without processing. Since S_{oav} and N are considered as constant and equal bases for all systems, we conclude that the best method of speech processing (i.e. the method yielding the greatest ITI) is the one that maximizes the ratio of S_{av}/D .

Note that the above threshold improvement *cannot* be directly measured by the receiver S-meter. An increased S-meter (i.e. average) reading is in response to greater average sideband power at the receiver which also includes quantity D produced by processor limiting. A simple means of measuring ITI is by the use of a calibrated attenuator at the station receiver antenna point. The received test signal, with the processor switched out, is reduced to the threshold level where it becomes barely understandable. The processor is then switched in and the test signal is now reduced further by the attenuator to where it again becomes barely understandable. The increase in attenuation is thus a reasonable value for ITI.

Fig. 3 shows a comparison of four methods of speech processing used in s.s.b. work. This figure illustrates the intelligibility threshold improvement or db. improvement in signal-to-noise ratio at the receiver vs. db. of peak limiting in single-sideband speech. The r.f. compressor has a 1-ms. attack time and 200-ms. release time, which is typical of a.l.c. circuits.

As these time constants are shortened, the a.l.c. circuit ultimately approaches that of r.f. clipping. The audio compressor's attack time is of the order of 5 ms., and it has a release time of 500 ms. As these time constants are shortened, the performance approaches that of an audio clipper.

Speech Clipping

Investigations have shown that speech remains intelligible under conditions of heavy peak clipping or what is termed infinite clipping. This process retains only the zero cross-overs and all envelope amplitude information is therefore destroyed. The fact that infinitely clipped speech is recognizable at all may be surprising until it is remembered that speech sounds consist of the fundamental frequency plus harmonics. Clipping will produce more harmonics which may upset the original harmonic structure of the sound, however not enough to destroy recognition of the original harmonic relationship of the sound.

It may be recalled that s.s.b. has been criticized in the past for its theoretical inability to transmit an audio square wave. Such would require an infinite amplitude capability, infinite bandwidth, and preservation of amplitude and phase relationships. In any practical modulation system, bandwidth is limited, and this reduces the required s.s.b. p.e.p.-to-average power ratio. This is particularly true when filtering is incorporated after clipping to restore original signal bandwidth. Table I illustrates the important relationships.

Clipping may also be accomplished at s.s.b. r.f. or i.f. frequencies. This has the advantage

that fewer in-band distortion products are created for a given amount of clipping, as harmonic distortion would then appear at multiples of the r.f. frequency. An r.f. filter following the clipper is necessary to maintain original bandwidth.

A.F. Clipping

Consider a speech wave that is infinitely clipped at audio frequencies, band-limited from 300-2700 Hz. and is used to modulate an s.s.b. transmitter. With an r.f. passband of $f_c + 300$ to $f_c + 2700$ Hz., the highest harmonic retained is the ninth; $2700/300 = 9$. From Table I, the p.e.p.-to-average power ratio of the s.s.b. envelope will not exceed 2.7/1 or 4.3 db. This would result in an increased average power output of 14.5 db. - 4.3 db. = 10.2 db. Again in the example of the 811A amplifier, the measured output with speech clipping, using the r.f. thermo-couple ammeter and load resistance technique would be about 300 watts (2.4 + amp.).

Observing the above transmitted signal from the receiver side, it was shown experimentally that 15 db. of clipping (with a relatively constant level speech) improves the speech intelligibility threshold by nearly 4 db. Increasing the clipping level to 25 db. gives an additional 1.5 db. improvement. Fig. 3 shows about 5 db. improvement with 20 db. of clipping. The intelligibility threshold improvement is less than the 10 db. gain, and it is evident that the distortion produced degrades the intelligibility about the same as noise.

R.F. Clipping

Infinite clipping of the s.s.b. signal followed by only enough filtering to remove the r.f. harmonics would give a p.e.p.-to-average power ratio of nearly 0 db. The result would approximate a constant-amplitude frequency-modulated r.f. sine wave. Increased filtering to remove out-of-band intermodulation distortion products will raise the p.e.p.-to-average power ratio toward 3 db. Like a.f. clipping the intelligibility threshold improvement is less than the theoretical gain in average-to-p.e.p. ratio of the transmitter. The expected improvement in intelligibility threshold if no distortion masking is

present is: 14.5 db. - 3 db. = 11.5 db. In the example of the 811A amplifier the measured power output with r.f. speech clipping using the thermocouple ammeter technique is in the region of 400 watts (3 db. less than p.e.p. capability). The r.f. ammeter would indicate a level of 2.8 amp. under these conditions.

Evaluation of s.s.b. clipping from the receive side with constant-level speech, and filtering to restore the original bandwidth, results in an improved intelligibility threshold of 4.5 db. with the clipping threshold set 10 db. below the signal peak. Increasing the clipping level to 18 db. gives an additional 4 db. improvement, or 8.5 db. total increase. The difference between expected and measured intelligibility threshold value is less with s.s.b. clipping than with the earlier discussed audio clipping. This verifies that distortion produced by r.f. clipping is less than that caused by a.f. clipping.

Speech Compression

Because of the syllabic character of speech, it would seem that an optimum compression system could operate to maintain a flat envelope. Peak volume compression acts much like delayed a.g.c. and is capable of producing a flat output once the compression threshold is exceeded. Ideally, the flat part of the compression characteristic should have a range of 35 to 40 db. Practically however, background noise very often reduces the range of the lower limit of the input. Compression much beyond 20 db. with high background noise will offer little improvement.

Compression may be accomplished at audio frequencies or at the s.s.b. transmitter's r.f. or i.f. amplifiers. In practice, fast attack a.g.c. is desired to prevent overshoot on steep voice wavefront signals (1 ms. or less). Too slow a release means the a.g.c. cannot follow rapidly decaying syllables. The d.c. control voltage does not change appreciably from one cycle to the next. Weak syllables following a strong one are then also compressed.

A.F. Compression

A common a.f. compression attack time is in the region of 10 ms. and release time in the order of 300 ms. In order to follow the envelope of speech, considerably shorter attack and release times are necessary for an effectively improved intelligibility threshold. The ideal attack time for a syllabic compressor's a.g.c. loop should be 1 ms. or less. The release time should be faster than 10 ms. Difficulties arise with such a fast release time constant, as feedback by the a.g.c. line can be regenerative or degenerative depending on the phase shift in the control loop.

Slow time constant compression is valuable for maintaining relatively constant average volume speech, but such a compressor adds little to the intelligibility threshold at the receiver, only about 1-2 db.

The faster time constant syllabic compressor,

TABLE I

Relationship of Power in a Band-limited Square Wave and its S.S.B. Envelope

A.F. Square Wave		S.S.B. Envelope	
Maximum Harmonic	Pk./Av. Power	P.E.P./Av. Pwr. Ratio	Db.
Fifth	1.5	2.0	3.0
Seventh	1.47	2.4	3.8
Ninth	1.46	2.7	4.3
Fifteenth	1.4	3.4	5.3

on the other hand, can give up to 5-5% db. improvement in the intelligibility threshold. This level requires a rather high degree of peak compression, between 24-40 db. As is shown in Fig. 3 the intelligibility threshold improvement of audio compression ultimately will approach that of audio clipping. Then the speech compressor behaves just like a clipper with the same threshold because the output waveshape is identical. The amount of distortion produced is the same as that produced by conventional clipping with the same limiting threshold.

R.F. Compression

S.s.b. transmitters generally utilize converters to translate the audio frequencies up to an r.f. band. After translation, the r.f. s.s.b. signal is linearly amplified by class A and class B stages to the required power level. Examination of the envelope of the r.f. signal for s.s.b. speech shows a variation in amplitude at the syllabic rate not too unlike that of the envelope of the speech.

The most common type of speech processing presently employed in today's s.s.b. transmitters is a.l.c. (automatic load or level control). This form of processor has been employed over the last 10-12 years. It has found wide acceptance in maintaining the peak r.f. output of the transmitter at a relatively constant level when the modulation varies over a considerable range. Present a.l.c. systems by nature of their design time constants offer a limited increase in transmitted average-to-p.e.p. power ratio. A value in the region of 2-5% db. is typical.

In the example of the 811A amplifier, the measured power output, using the thermocouple-ammeter technique, is increased to the region of 45-100 watts. The r.f. ammeter would indicate an average level between 1-1½ amp. r.m.s. under these conditions. Examining the above transmitted signal from the receiver side, 15 db. of peak a.l.c. improves the intelligibility threshold by 1 db. to as much as 3 db. The intelligibility threshold improvement is again less than the transmitted increase, as was noted in the preceding processing techniques.

An r.f. syllabic compressor can be made to function in the i.f. stages of the transmitter. Sufficient a.g.c. is employed to flatten out the amplifier output characteristic above the threshold. This requires not only fast attack to follow the envelope rise but also fast decay to follow the fall of the envelope. The frequency ratio between the r.f. s.s.b. voice signal and the voice syllabic rate is of course much greater than the ratio between the voice signal (audio) and the syllabic rate. Gain control time constants can be made much faster without the hazard of regenerative signal feedback and resultant severe distortion and even oscillation.

An r.f. syllabic compressor will naturally initiate distortion of the s.s.b. envelope in relation to the voice signal. Harmonic distortion generated in the r.f. envelope will of course be in multiples of the r.f. frequency and therefore lie

outside the r.f. passband. No even-order distortion products will appear in the output. This has the advantage that fewer in-band distortion products are created for a given level of compression. An r.f. filter in the output removes IM-distortion products falling outside the original passband.

With such circuit application and operating in the region of 40 db. peak compression, the intelligibility threshold for constant level speech can be made to improve nearly 6 db. This compares with the 8.5 db. improvement obtained by nearly 20 db. of r.f. clipping followed by filtering.

Since clipping produces as good a result as compression, why is r.f. compression considered in view of its greater complexity? For a given improvement in marginal intelligibility the r.f. compressed wave has about 6 db. less third-order IM-distortion and about 12 db. less fifth-order IM-distortion than the r.f. clipped wave. Thus, under nonmarginal conditions the r.f. compressed signal should have the better quality.

Conclusion

From the previous discussion it is seen that speech processing increases transmitted average-

TABLE II
Parameters of Described Speech
Processing Methods

Processing Technique	Processor Peak Limiting Db.	Trans- mitter Gain in Av.-To- P.E.P. Output Db.	Intelli- gibility Threshold Improve- ment ITI, Db.
A.F. Clipping (300-2700 Hz. bandpass)	25 15	10.2 7.5*	5½ 4
Common A.F. Compression (Slow time constant)	20	—	1
Syllabic A.F. Comp. (300-2700 Hz. bandpass)	25-40	8-10*	5-5½
A.L.C. (Fast attack time, slow release time)	15	2-5½	1-3
R.F. Clipping (300-2700 Hz. bandpass)	18 10	11.5 6*	8.5 4.5
R.F. Syllabic Compression (Fast attack and release time)	40	—	6

NOTE: unprocessed speech is given in text as having a peak-to-average power ratio of 14.5 db. Values marked by asterisk* are approximate from author's measurements and observations.

to-p.e.p. power, thereby improving the intelligibility threshold. This has been determined experimentally in actual communications set-ups by careful examination of what is presented as intelligibility in the receiver output.

Audio clippers (and fast syllabic a.f. compressors) add harmonic distortion as well as intermodulation distortion in the signal passband; r.f. clippers (and fast syllabic r.f. compressors) generate harmonic distortion at frequencies that are multiples of the r.f. filter frequency and are therefore rejected. Only odd-order products falling within the filter passband will appear in the output. Hence for these reasons, r.f. processors "sound better" than do a.f. processors.

In the absence of noise, speech processing (r.f. or a.f.) changes the character of the sound. Hard limiting produces noticeable distortion. Low-level sounds on most consonants have an increased prominence. Breath sounds and background noise also assume greater significance. Use of large amounts of clipping or syllabic compression makes a noise cancelling microphone a very desirable item to suppress ambient background noise.

Listed in Table II is a summary of parameters for the processing techniques previously described. A conclusion readily drawn from observing Table II is that the greatest intelligibility threshold improvement is obtained from an r.f. clipper employing about a 20 db. clipping level. Intelligibility threshold values are justified on empirical grounds. Caution should be exercised in extrapolating results; e.g., if a.f. and r.f. compression are employed together, only the larger improvement factor should be considered as they are not directly additive.

Today most of the speech processing is taking

place outboard from the transmitter or by adapters incorporated into existing equipment. Solid-state devices lend themselves ideally in the design of effective compact units. From this it seems only logical to presume as well that the next generation of s.s.b. equipment will have speech processing (other than just a.l.c.) included in the design concept.

The author wishes to extend appreciation and thanks to Mr. Stanley Brown and to Mr. Clarence Munsey, K6IV, for their helpful suggestions in reviewing this writing. Certainly the greatest credit must be given to those authors listed in the bibliography, for without their work this article would probably not have evolved.

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» A controlled amount of clipping added to compression gives a better overall result, in speech processing, than either alone.

A Solid-State Speech Processor

Compression and Clipping for Increased Voice Power

JOSEPH J. SPADARO, WB2EY2

The object of this speech processor is to raise the average-to-peak power level in the speech spectrum with negligible change in audio quality. This will result in a corresponding increase in the average r.f. power, when used with a transmitter. It creates greater output from an s.s.b. transmitter, which normally is limited by the p.e.p. rating of the output tubes, and it permits much greater modulation power for a.m. rigs.

The proper use of this speech processor can double the d.c. power into the final of a modern s.s.b. transmitter without degradation to the audio quality.

The January 1969 issue of *QST* carried a splendid article explaining the virtues of speech processing.¹ The article treated rather thoroughly the relative merits of r.f. clipping, audio clipping, and speech amplitude compression. It showed that theoretically the average-to-peak value of a transmitted signal can be enhanced considerably by the addition of r.f. or audio clipping of high-amplitude pulses. It also showed that r.f. and audio clipping, when properly applied, result in much greater effective power increase, or punch, than can be obtained from compression alone.

This article prompted the author to develop the audio speech processor described here. Previously, he had successfully built a speech compressor,² but experience showed that the compressor when used properly with a modern s.s.b. transmitter having effective a.l.c. circuits, offered only a small improvement in transmitter punch. The enhancement was primarily in maintaining a constant microphone level and was most advantageous only when the overall speech-input level was not kept reasonably constant.

The present speech processor employs two fundamental principles: level speech output through the use of automatic gain control (a.g.c.), and clipping of audio pulses at a predetermined amplitude to raise the average-to-peak level.³

These functions are best related when the a.g.c. (or compression) precedes the amplitude clipper. Under these conditions more consistent clipping can be achieved since the audio level fed into the clipper is comparatively constant. To translate this concept into hardware, it was decided to add a clipper circuit to the output of the speech compressor already in use.

Adding a Clipper

The typical clipping voltage for germanium diodes is about 0.2 volt. Therefore, adding diodes directly at the output of this particular compressor circuit would prove ineffective since the compressor output is about this same level. It is necessary to add another stage to the circuit to act as a voltage amplifier and raise the audio amplitude high enough for clipping by the diodes.

After careful scrutiny of the compressor circuit it was realized that the transistor stage following the output transistor Q_2 in the original compressor is, in fact, a voltage amplifier. This transistor is in the compressor feedback loop and serves as a convenient signal takeoff point for subsequent clipping.

Compressor Circuit

Several minor and one major improvement were made to the compressor circuit. Fig. 1 is a schematic of the speech processor, a portion of which includes the compressor circuit. All components used are essentially equivalent to those originally specified in August 1965 *QST*. The 1N270 diodes have very low forward resistance and are better than the 1N34s. The 20- and 25- μ f. capacitors were replaced by 22 μ f. capacitors were replaced by 22 μ f., 10-volt types for uniformity. The single-pole single-throw switch was replaced with a triple-pole double-throw slide switch. This enables the operator to bypass the processor without the need for disconnecting cables. Also, R_2 is now a fixed 10,000-ohm carbon resistor.

One major improvement was the addition of a 0.002- μ f. ceramic capacitor directly from the base of transistor Q_1 to ground. Although great care was taken in providing a good ground for the unit, the author was plagued by r.f. feedback, detected as random clicking pulses during transmission. The capacitor completely stopped the feedback.

¹ From November, 1969, *QST*.

² Collins, "Ordinary and Processed Speech in S.S.B. Application," *QST*, January, 1969.

³ Moate, "An Audio Peak Limiter for Voice Transmission," *QST*, August, 1965.

⁴ The only previous design using this combination of which the author is aware was described by J. L. Tonne, W5SUC, in September 1956 *QST*. Appropriately, the article was titled "Compression and Clipping."

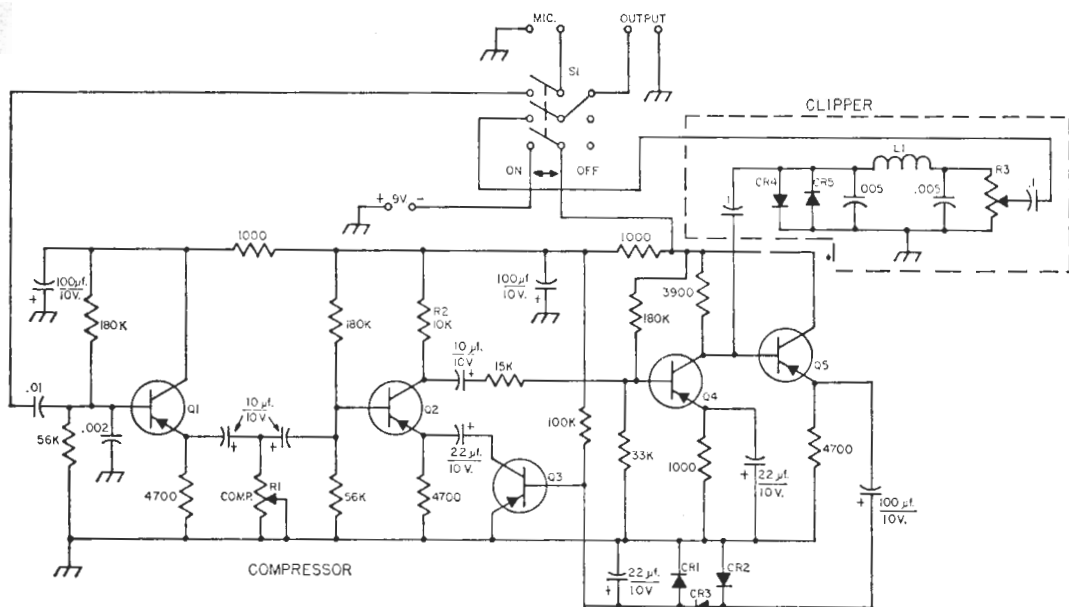


FIG. 1—Circuit diagram of the speech compressor. Capacitances are in μf .; capacitors with polarity marked are electrolytic, others are ceramic. Resistances are in ohms ($K = 1000$); fixed resistors are $\frac{1}{2}$ -watt composition.

CR_1 — CR_5 incl.—1N270 or equivalent.

L_1 —3-3.5 henrys, miniature type desirable.

Q_1 — Q_5 , incl.—2N1375 or equivalent.

R_1 —10,000-ohm control, audio taper.

R_2 —For text reference.

R_3 —50,000-ohm control, linear taper.

S_1 —3-pole double-throw slide switch.

A single 9-volt transistor radio battery serves as the power supply, and will last a reasonably long time even for the most ambitious operator.

Clipper Circuit

The clipper, which is coupled to the collector of Q_4 through a 0.1- μf . capacitor, consists of two diodes back-to-back followed by a constant- k filter which cuts off all frequencies above 2500 to 3000 Hz. A 50,000-ohm shunt resistor is added, the object of which is to provide a relatively constant high-impedance termination for the filter under varying load conditions. The use of a 50,000-ohm linear potentiometer for the termination permits adjusting the output level to the transmitter. The output of the processor can then be matched to the microphone output so that switching from one to the other will not require readjustment of the transmitter microphone gain control. The output range of the processor is greater than that of the original compressor.

The filter includes a UTC DOT-8 choke, which has proved very effective. However, any choke having an inductance of about 3 henrys should perform satisfactorily.

The compression and clipping levels could be independently adjusted by using a 50,000-ohm potentiometer after Q_4 and the clipper. However, the additional control is not necessary nor is it recommended, as the processor works extremely well without it.

Observations of the audio output on an oscil-

loscope verified that the clipping is "soft." That is, when the amplitude is clipped it does not result in a sharply-squared-off wave-form, but rather a compressed or shortened version of the original pulse. This feature is considered to be the result of the combination of a.g.c. (compression) and clipping, together with the low-pass filter, and the final result is an audio output essentially free of harmonics above the filter cutoff frequency.

Construction

As in the construction of most audio devices, layout of components is not critical. For example, the circuit components can be laid out physically just as they are depicted in the schematic, with satisfactory results. Layout details are left to the reader. Other than good construction practices there are no restrictions about component placement.

In the original version the author, having had previously constructed the compressor merely added the clipper circuit on a separate 4-prong tie-point standoff. The processor was combined with a phone patch, being housed in a 3 \times 4 \times 5-inch Minibox. (Incidentally, phone-patch quality is considerably enhanced when channeled through the processor.) A later version, built on prepunched board using flea clips as tie-points, is shown in Fig. 2.

The most important point to remember during construction is a good circuit ground solidly connected to the chassis ground.

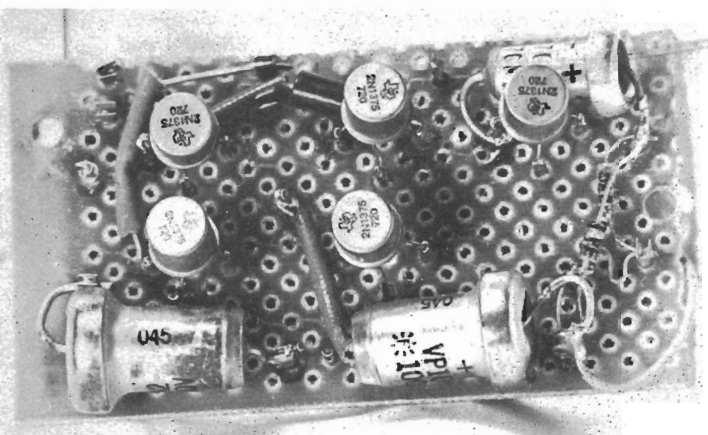


FIG. 2—One side of a punched-board assembly of the processor. The three 100- μ f. electrolytic capacitors and the five transistors are visible. Other components are mounted on the reverse side.

Adjustment

The limitations of the processor must be realized. The best that it can do is double your signal strength, when using good-quality equipment that already includes such features as a.l.c. Beyond this point there will be noticeable deterioration of the voice quality and an objectionable increase in background noise. The optimum setting is just below the level where these two conditions occur.

Only one adjustment actually exists, since R_3 merely serves to set the output initially. The author's experience is that the optimum adjustment for R_1 is about $\frac{1}{4}$ to $\frac{1}{2}$ turn (100 ohms on an audio-taper potentiometer). Once set, this adjustment should satisfy all operating conditions from armchair ragchewing to pulling out the rare ones. A condition that can justify a higher setting of R_1 is when an operator is distant from the microphone, provided that background noise is nonexistent. Tests performed on the air verified that although more d.c. plate current could be drawn, the punch power no longer increased when the limiting control R_1 reached the point of noticeable distortion.

Adjustment of the processor can be facilitated if you are equipped with a hi-fi system. By connecting the processor to the auxiliary input of the amplifier and attaching headphones to a suitable output, one can actually hear the optimum adjustment while speaking.

Performance

Due to the complexity of voice waveshapes a quantitative evaluation of the processor is beyond the scope of this article. However, certain significant observations have been made which, though qualitative, verify the claim made at the outset—that the processor can increase the punch of your high-quality rig up to two times.

FIG. 3—Oscillograms of the word "eye" with (above) no processing and (below) processing with the adjustments made as described in the text. The more nearly constant amplitude of the peaks in the lower trace is an indication of the increase in average power.

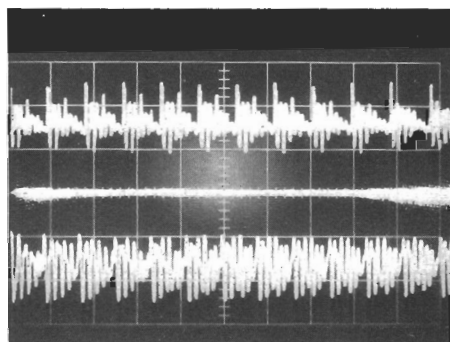
The performance of this unit has been verified by (1) significant additional heat dissipation in the power-output tubes,⁴ (2) doubling of the estimated average d.c. plate current, (3) glowing reports from distant receivers, and (4) actual visual verification on an oscilloscope.

Fig. 3 shows a picture of audio signals made on a Tectronix 549 storage oscilloscope. Sweep rate was set at 10 milliseconds/cm and the amplitude was set at 0.2 v./cm. The scope could independently record and store a trace on either the upper or lower half of the screen. The scope was triggered with the voice sound "eye" which was simultaneously recorded on the screen. The upper trace was made using a straight microphone and the lower trace using the processor. The increase in average-to-peak ratio is easily recognizable in the lower trace. These traces are also representative of patterns observed on the scope during conversational speech. The amplitude variation was always much more uniform when using the processor and the observed average-to-peak ratio was always higher. All scope patterns were observed with the processor adjusted just below the point where distortion or background noise are objectionable, as discussed earlier.

Acknowledgments

Thanks to Mr. Alfred Jorysz, electrical engineer, for assistance in developing the circuit, to "Pappy" (H. Bohning), W2MZV, for his confirmatory tests and to B. Walsh, WB2QLR, for his assistance.

⁴ This is an extremely important consideration in the use of any speech processor which increases the average-to-peak power ratio. Many transmitters simply are not built to stand the effects of increased average input, either in the final-amplifier tube or tubes or in the power supply. If heating in the final tube is the limiting factor, adding a cooling fan may be a satisfactory answer. The author added one in his own equipment. —Editor.



» Increased talk power on s.s.b. at the same peak-envelope power, with a reduction in out-of-channel splatter. R.f. clipping and filtering is the answer. And it doesn't take complicated circuits.

R. F. Clippers for SSB

WILLIAM SABIN, WØ1YH

It has been verified, through extensive tests, that the ordinary voice contains high-amplitude peaks which are about 14 db. greater than the average level. It is also recognized that voice-communication systems are often peak-power limited. For example, a certain single-sideband linear amplifier will handle 2000 watts peak power before it flat-tops, even though it can handle 1000 watts average. From these facts, one infers that the natural voice may not provide the maximum utilization of the equipment. Let's say it another way: the *talk power*¹ is not as great as it could be if the peak-to-average ratio were not so high. If we can elevate the average level without overloading on peaks, the weaker components of the voice become more prominent. The result is that in a noisy, or cluttered, channel a higher level of *articulation*, or understandability, is achieved.

In order to reduce the peak-to-average ratio it is necessary to modify the signal waveform. This inevitably leads to distortion of the voice signal. This distortion reduces the articulation. From all of this we conclude that the best method of speech *processing* is the one that gives the greatest increase of talk power and the least distortion.

A figure of merit for speech processors is the ratio of peak signal to r.m.s. noise which is required for a given degree of articulation. For example, one system may require a ratio of 10 db. for an articulation index of 0.3. If an audio compressor is used, perhaps only 7 db. will be required.

Fig. 1 is a comparison of three different methods of speech processing used in s.s.b. work. The audio compressor had an attack time of 0.005 seconds and a release time of 0.5 seconds. As these time constants are shortened, the performance approaches that of the audio clipper. The r.f. compressor had 0.001 seconds attack time and 0.2 seconds release time (typical of a.l.c. circuits). As these time constants are shortened, the performance approaches that of the r.f. clipper.

Not all the possible schemes are shown in Fig. 1. This article is concerned only with "simple" techniques. Moreover, the type of

audio limiting described by Schleicher², which might be called "simple," has not been considered here.

Fig. 1 shows an advantage of several db. for r.f. clipping, for 20 db. of processing, over its nearest competitor. This article is concerned with this type of processing. At this point, previous material should be reviewed.^{3,4,5}

Sideband Clipper Performance

Fig. 2 is a block diagram of the setup that was used to study the r.f. clipper. The measurement equipment was all of very high quality. The transmitter and receiver were standard amateur gear with the clipper unit onboarded.

Fig. 4 shows a two-tone envelope, measured right at the point of clipping, after various amounts of clipping. Note that the peak-envelope signal remains nearly equal to the single-tone envelope of Fig. 5. The clipped envelope is quite flat. At large amounts of clipping, the peak-envelope signal is nearly the same as the average. That is, the peak-to-average ratio of the envelope approaches zero db.

Fig. 6 shows the output of the second filter for the single tone and for a two-tone signal

² Schleicher, "A Passive Limiter," *QST*, December, 1966.

³ Craiglow, et al. *IRE Trans. Audio*, Nov. 1961.

⁴ Squires and Clegg, "Speech Clipping for Single Sideband," *QST*, July 1964.

⁵ Pappenfus, et al. *Single Sideband Principles and Practice*, Chapter 20. (McGraw-Hill).

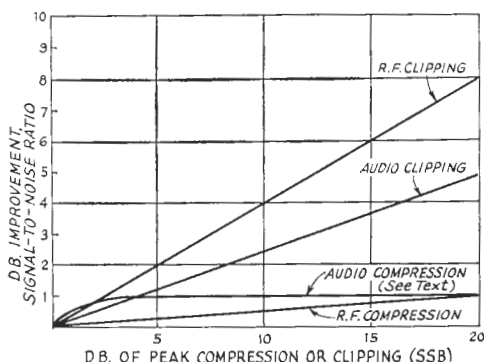
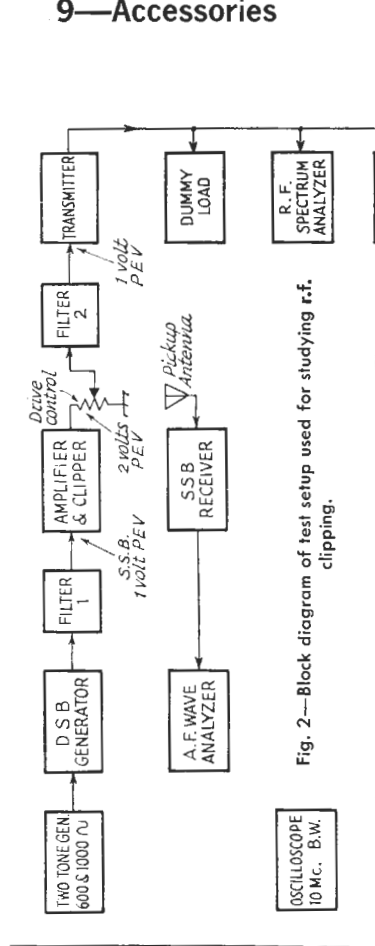


FIG. 1—Comparison of simple single-sideband speech processing methods.

From July, 1967, *QST*.

¹ *QST* January, 1954, page 19.



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μ f.); OTHERS ARE IN PICOFARADS (p.f. OR $\mu\mu$ f.); RESISTANCES ARE IN OHMS; K = 1000.

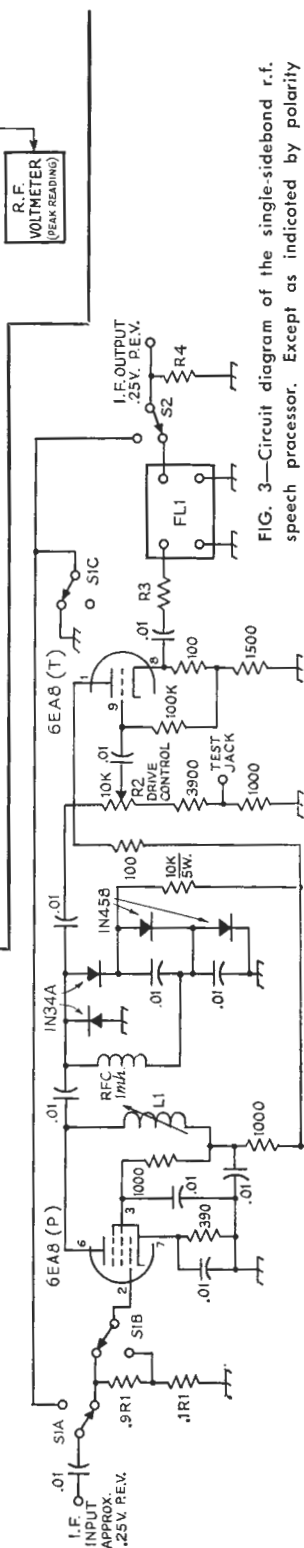


FIG. 3—Circuit diagram of the single-sideband r.f. speech processor. Except as indicated by polarity (electrolytic), capacitors are disk ceramic; fixed resistors are $\frac{1}{2}$ -watt composition.

C_1, C_2 —A.C. type disk ceramic.

FL₁—Sideband filter; should be same type as in exciter.

I_1 —Slug tuned, to resonate at exciter's intermediate frequency. For high-frequency i.f.'s (e.g., 9 Mc.) I_1 can resonate with the stray circuit capacitance, which is of the order of 25 pf. For 455-kc. i.f., additional capacitance such as 240 pf. can be used in parallel with I_1 . Approximate inductance values are 15 μ h. for 9 Mc. and 500 μ h. for 455 kc.

R_1 —Load resistance for filter (in exciter) as recommended by filter manufacturer. Sections

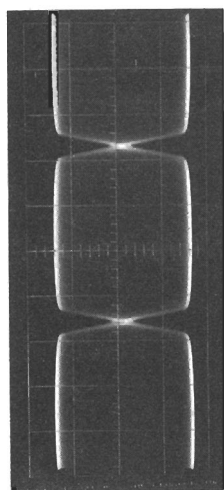
R_2 —10,000-ohm composition control, linear taper.
 R_3 —Filter manufacturer's recommendation for source impedance, less 500 ohms (output impedance of cathode follower).

R_f —Filter load resistance recommended by manufacturer for FL_1 . (0.1 megohm for Collins mechanical filter).

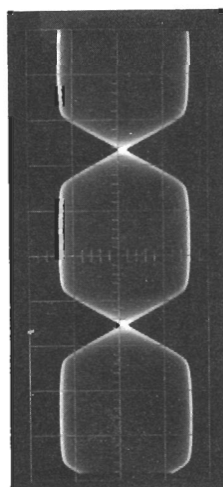
S_1 —3-pole, 2-position ceramic switch.

S_1 —1-pole, 2-position ceramic switch.

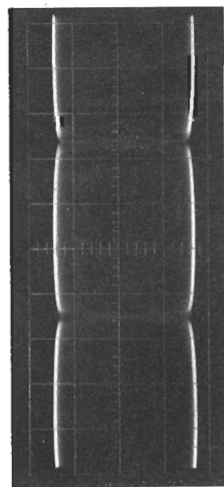
T₁—250 volts c.t., 25 ma. (Knight 54 A 2008).



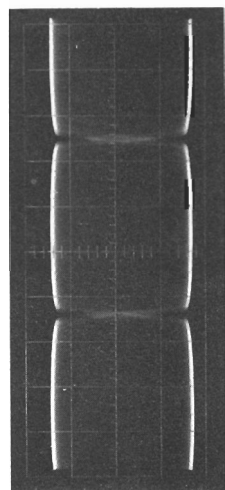
(A)



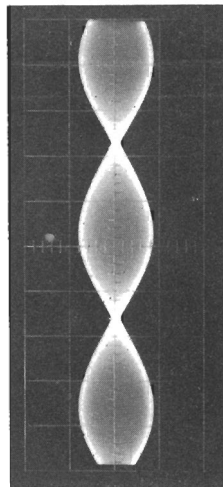
(B)



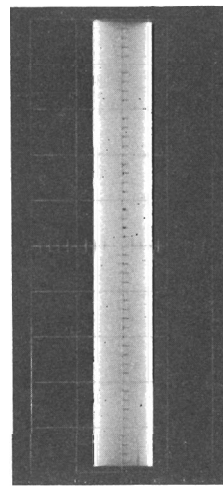
(C)



(D)



(E)



(A)

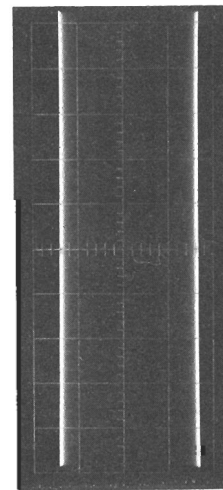


FIG. 5—Single-tone envelope measured at test jack.

FIG. 4—Two-tone envelope patterns with various degrees of clipping. All patterns taken at the test jack at the clipper output (Fig. 3), using tones of 600 and 1000 c.p.s. A—at clipping threshold; B—5 db. of clipping; C—10 db. of clipping; D—15 db. of clipping; E—20 db. of clipping.

FIG. 6—Envelope patterns at filter output: A—single tone; B—two-tone envelope, same frequencies as Fig. 4, with 20 db. of clipping.

TABLE I

Wave Analyzer Data Receiver Output

	Two-Tone Source	S.S.B. Receiver	0-db. R.F. Clipping	10-db. R.F. Clipping	20-db. R.F. Clipping
f_A 600 c.p.s.	0 db.*	0 db.*	0 db.*	0 db.*	0 db.*
f_B 1000 c.p.s.	0 db.*	0 db.*	0 db.*	0 db.*	0 db.*
$2f_A$ 1200 c.p.s. (2nd harm.)	-70 db.	-50 db.	-48 db.	-46 db.	-42 db.
$2f_B$ 2000 c.p.s. (2nd harm.)	-80 db.	-50 db.	-48 db.	-46 db.	-41 db.
$2f_B - f_A$ 1400 c.p.s. (3rd order)	-75 db.	-35 db.	-26 db.	-11 db.	-10 db.
$3f_B - 2f_A$ 1800 c.p.s. (5th order)	-80 db.	-45 db.	-40 db.	-20 db.	-17 db.

* Normalized

with 20 db. of clipping. Several interesting things are illustrated by these photos. For one thing, the heavily-clipped two-tone signal is restored to almost the original form, except for distortion products which lie in the passband of the filter. Secondly, the peak-to-average ratio of the restored waveform is almost equal to its original 3 db. These things happen because the filter rejects all of the out-of-band distortion products.

Now, from the photographs, we make the following measurements. At the input to the filter, we see that the two-tone peak (20 db. of clipping) is 2 percent higher than the single tone. At the output of the filter, the two-tone is 31 percent higher than the single tone. The net increase through the filter is 29 percent. Theory predicts 27 percent. Voice waveforms exhibit approximately the same effect, and this fact must be considered when adjusting the transmitter. If we tune up and adjust clipping on a single tone, we will flat-top like mad on voice. This situation is aggravated by the higher d.c. plate and screen voltages which are available on speech as compared with single tone.

We now look at the output of the receiver with a wave analyzer. Thus we are duplicating an actual communications setup and examining what comes out of the loudspeaker. We are looking at *in-band* distortion products. The results are listed in Table I. The quality of the two-tone audio source and the receiver are included, for reference. Looking first at the column for zero db. of clipping, we see that the harmonic distortion is very good, and the third and fifth order intermodulation products are pretty good. Proceeding to the 10-db. column, we see that the harmonic distortion is hardly affected, but the intermodulation products get larger. We have now put the finger on why r.f. clippers sound better than audio clippers. Audio clippers add a lot of harmonic distortion as well as "intermod." In the r.f. clipper, the harmonic distortion appears at multiples of the filter frequency, and is therefore rejected.

Proceeding to the 20-db. column, we see that the intermod products increase only slightly. This in itself is very interesting and is substan-

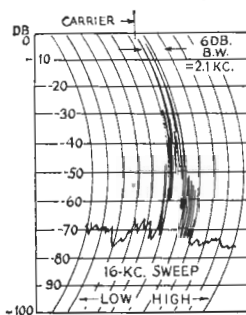


FIG. 7—Spectrum distribution of two-tone signal, exciter only, no clipping. Filter 6-db. bandwidth 2.1 kc. Sweep width 16 kc.

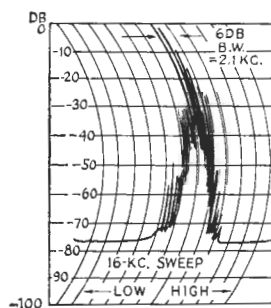


FIG. 8—Two-tone spectrum, 20-db. clipping with circuit shown in Fig. 3. Filter 6-db. bandwidth 2.1 kc. Sweep width 16 kc.

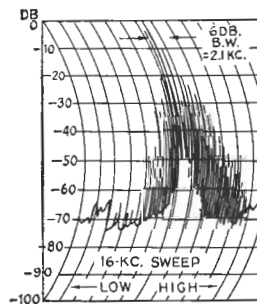
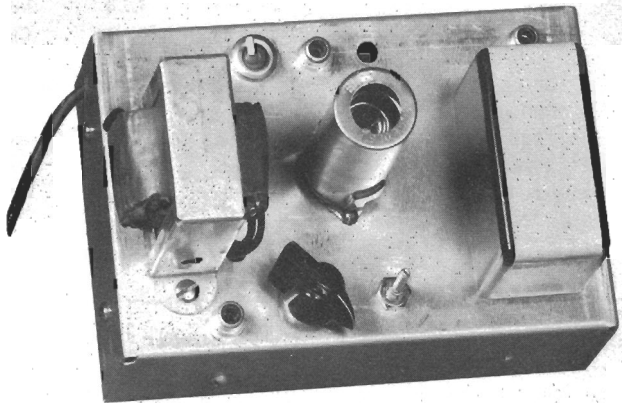


FIG. 9—Two-tone spectrum, 20-db. clipping, no output filter (measured at test jack in Fig. 3). Filter 6-db. bandwidth 2.1 kc. Sweep width 16 kc.

The outboard s.s.b. clipper built by the author is on a $6 \times 4 \times 2$ -inch chassis. The filter is at the right; the test jack, drive control, and slug-tuned coil are along the edge at the bottom. The chassis wall at the top in this photo is bolted to the rear of the exciter cabinet in normal use. This model does not have the bypass switch S_1 in Fig. 3.



tiated fairly well by listening tests. The harmonic distortion is degraded a little, because in this particular test we deliberately drove the audio circuitry in the exciter harder. The data shows that this is all right, as long as we don't overdrive the balanced modulator or run into hum problems.

Figs. 7, 8 and 9 are spectrum sweeps (16 kc. wide) measured at the output of the transmitter. The filter response is indicated. We see that with 20 db. of clipping, Fig. 8, the out-of-band performance is acceptable, in terms of present-day amateur practice. The particular exciter used for these tests was a fairly inexpensive unit, and does not represent the best that can be had. Aside from all of this, it is a fact that the complete absence of flat-topping, despite the high average level, is a great big plus mark for the r.f. clipper.

It is interesting to see what the "garbage" looks like if we leave out the second filter. See Fig. 9. Surprisingly enough, the intermod. products are down at least 65 db. at each end of the 16-kc. sweep. If higher pitched tones had been used, the results would have been worse looking. Nevertheless, the higher-order products do drop off rather quickly. Still, the second filter should be of good quality, and should have low leak-through, outside the pass band.

Transmitter Adjustment

When first starting out with a clipper it is easy to get confused. Because of the irregular behavior of meters and scope patterns with speech signals, it takes some headscratching and experience to know what is going on.

The best place to begin is to tune up the rig on a two-tone audio signal with the clipper in the "tune" position. The level of the two-tone signal should be set just below the point at which clipping begins (see Fig. 4A). This measurement is made with a scope at the clipper test jack. The drive control and all succeeding adjustments may then be set for best output

and linearity, in the usual way. If a.l.c. is used the peak envelope signal should be set, using the drive control, right at the point where a.l.c. begins. If the peak-envelope voltmeter to be described later is used, it may now be adjusted for ten percent deflection.

Now switch over to microphone input. Turn up the mike gain until the same peak output is indicated on the output monitor scope. The a.l.c. meter will flicker occasionally and the peak-envelope voltmeter will show occasional activity. Now turn the switch on the clipper to the "clip" position. You now have about 20 db. of peak clipping. While talking steadily into the microphone, adjust the drive control so that the proper peak level is maintained. The peak-envelope meter will kick up on scale very frequently and the final plate meter will kick up to a level which is just about equal to that obtained in the two-tone test. That is to say, on voice peaks the peak-to-average ratio will approach 3 db. for short intervals.⁶ The ratio over a longer period will be about 9 db.

The mike gain control can be used to make minor adjustment of clipping. A little experience will enable the operator to set the gain for good quality and lively meter action. Avoid the temptation to run the gain way up. The rig won't flat-top, of course, but distortion and room noise can become excessive and communications effectiveness is hardly improved at all.

Any change of gain after the clipper due to mistuning or line-voltage changes will cause the peak level to become too high or too low. The peak voltmeter makes a good monitor for this condition. If a.l.c. is applied to the exciter after the clipper, it can be used to adjust the drive level. One or two db. of fast-attack, slow-release type a.l.c. should be used. If a.l.c. is applied ahead of the clipper, it will be worthless.

The Clipper Circuit

The signal at the input of the clipper unit

⁶ See page 160, *Single Sideband for the Radio Amateur*, 3d ed., 1962, "Power Ratings," by L. Norton.

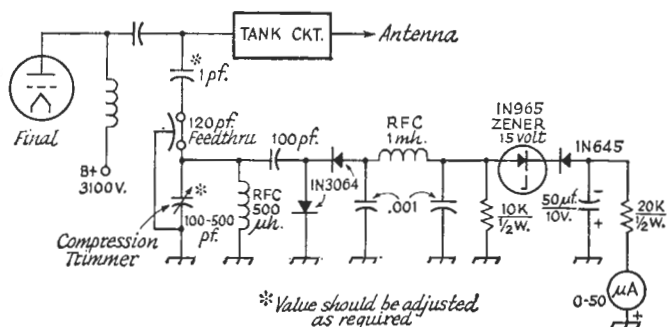


FIG. 10—Peak-envelope voltmeter circuit. Resistances are in ohms; K = 1000. Capacitor with polarity marked is electrolytic; other fixed capacitors are ceramic or mica.

should be about 0.25 volt peak envelope. The signal at the clipper diodes should be enough to make up the loss in the drive control, cathode follower and filter, and to drive the right to full output. In the exciter used by the author, one volt peak envelope was enough. The diode clipper circuit shown gives flat clipping, good symmetry and freedom from rectification effects which can spoil the clipping symmetry during the transient conditions encountered in speech wave-forms. An important feature is that the clipping level is independent of line-voltage fluctuations. Also, hard, flat clipping prevents overloading the transmitter on the strongest peaks.

From the above it can be seen that for 20 db. of clipping, a total voltage increase of forty (32 db.) is required. The amplifier should be biased so that the control grid does not go positive on peaks. The inductor L_1 in the plate circuit resonates with stray capacitance, about 25 pf., at the i.f. frequency. This adjustment may be made with a small one-tone input. Before clipping, the plate load is essentially 15,000 ohms. During clipping the plate load drops to less than 100 ohms. After all adjustments are complete, recheck the resonance of the tuner circuit.

Finally, on c. w. it is best to switch the clipper out of action altogether.

Construction

The clipper unit is built on a small chassis which is bolted to the exciter cabinet for good grounding. The coax leads must be kept very short to prevent distortion of the filter response. Leakage paths around the filter must be eliminated by careful shielding and lead dress. The filter should be kept away from both a.c. and d.c. magnetic fields. A soft-iron cover for the filter may be needed. Stray r.f. from high-power amplifiers should not be allowed to sneak in and gum up the works.

Peak-Envelope Voltmeter⁷

Fig. 10 shows that the peak-envelope voltmeter consists of an adjustable capacitive divider from the plate of the amplifier to ground, a rectifier, an r.f. filter, a zener diode and isolation diode, a 50- μ f. storage capacitor, and a

0-1 voltmeter. The principle of operation is that if the peak-envelope signal exceeds the zener voltage, the capacitor will charge up quickly, causing the meter to "hang up." After several syllabic peaks the meter will give a pretty good indication of peak-envelope voltage. The adjustable capacitor should be set for about ten percent deflection of the meter. The quantity being measured is the r.f. voltage supplied to the plate load. After the initial adjustment, previously described, it makes a good tuning and loading indicator. A glance at its activity also tells you if the drive and clipping are adequate (or too much).

Results

When you build a gadget like this, you start out in a wave of optimism. You are going to knock 'em dead on the DX phone bands. However, after battling it out with the gang, you become a realist again. (*Will I ever beat those stacked Yagis in Saddle Brook??*)

The following statements tell the story for a good 20-db. r.f. clipper:

- 1) If you start out with good, clean audio and a high-quality microphone, the signal has a crisp, clean communications quality with a very high degree of articulation. You get a lot of "solid copy" and "perfectly readable" reports.
- 2) Clipping is most effective under very weak-signal conditions. It is also helpful (but apparently no panacea) in the big pileups. For ordinary state-wide ragchews it provides effortless readability. The clipper has definitely improved the author's competitive capability in DX chasing.
- 3) The clipper provides a very excellent way of keeping amplifiers linear and clean, even in the "heat of battle."
- 4) In the author's opinion, and according to the best laboratory data available, the r.f. clipper is significantly better than any simple audio compressor.
- 5) It also requires that you dig into the exciter and do a little work. You don't just "plug it in."
- 6) An additional sideband filter is required. They are somewhat expensive.

Acknowledgments

Many thanks to Mr. R. Craiglow for reviewing this material and to Mr. Larry Wilson, W0KVL, who helped out with the measurements.

⁷ Bruene, "Directional Wattmeters", QST, April 1959.

» Automatic transmit-receive switches are the answer to break-in operation with one antenna. This electronic switch uses a twin-triode, with an "all-band tank" for tuning. As a consequence, the switch also provides some gain.

An Electronic T-R Antenna Switch

EDWARD ARVONIO, W3LYP

After 4½ years on s.s.b. I felt the need for a t.r. switch that would replace the old coaxial relay and give worthwhile gain on all bands with low noise and no TVI. The t.r. switch described here will give a gain of better than 20 db. on all bands, and its operation as a t.r. switch leaves little to be desired. The tuning control has to be set only once whenever you change bands. With proper shielding and filtering, it was possible to eliminate TVI completely. At present I operate mainly on 21-Mc. s.s.b., and no TVI is caused by the switch.

Referring to the circuit in Fig. 1, one section of a 6BZ7 is used as a grounded-grid amplifier. Its plate circuit is tuned by an "all-band tank" that requires no switching. The output is coupled to the receiver through the second section of the 6BZ7, operated as a cathode follower. Operating bias for the input section is obtained by the d.c. drop across the 2.5-mh. r.f. choke; when the transmitter is on, a high

bias is developed across the 470K grid return.

The choice of tube for the switch came out of many experiments. I chose a tube that would have a low noise figure and would stand up under 1-kw. s.s.b. conditions without burning out. Let me not mislead you at this point; it is possible to blow the tube under certain operating conditions. It is possible to blow the tube if you operate the switch without an antenna load or operate with an s.w.r. of more than 3 at a kw. input. After checks of several makes of tubes, it was found that RCA 6BZ7s were the only ones that would stand up with a kilowatt transmitter. If lower power is contemplated, any brand will probably do.¹

¹ Manufacturers do not rate their tubes for r.f. voltages between heater and cathode, and the 200-volt d.c. rating for the 6BZ7 is not applicable. W3LYP's findings are reported here because the t.r. switch is a useful device, but using it at power levels above several hundred watts can only be considered as a calculated risk. At higher power levels two tubes should be used, with the heaters fed from separate windings, so that the heaters can be tied to the cathodes.—Ed.

From QST, October, 1957.

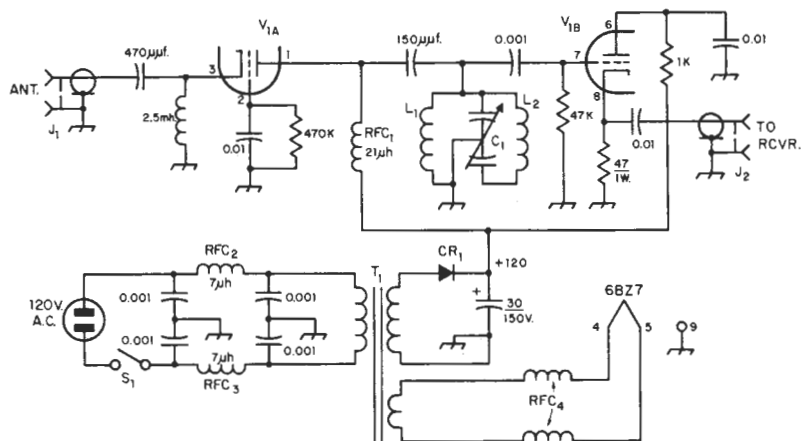


FIG. 1—Schematic diagram of the electronic t.r. switch. Capacitances are in μ f. unless otherwise noted. Resistors are $\frac{1}{2}$ watt unless otherwise specified.

C1—450-pf. per-section, broadcast-receiver type.

CR1—130-volt 65-ma. selenium rectifier (Federal 1002A or equiv.).

J1, J2—Cable connectors, SO-239.

L1—19 turns, 1-inch diam., 32 t.p.i. (B & W 3016).

L2—23 turns, $\frac{1}{2}$ -inch diam., 16 t.p.i. (B & W 3003).

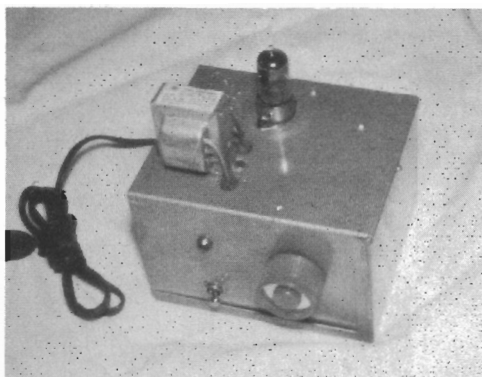
RFC1—Ohmite Z-28 or equiv.

RFC2, RFC3—Ohmite Z-50 or equiv.

RFC4—Bifilar winding. See text.

V1—6BZ7. See text.

T1—115-v. secondary at 15 ma., 6.3 volts at 0.6 amp. (Triad R-54X or equiv.).



This electronic transmit-receive switch works on all amateur bands down to 10 meters. No switching is required to change bands.

A bifilar winding is used in the heater circuit of the 6BZ7 to reduce the heater-cathode capacitance at V_{1A} . Shown as RFC_4 in Fig. 1, it was made by putting two parallel windings of No. 26 enameled on a $\frac{1}{2}$ -inch diameter form $1\frac{1}{2}$ inches long. The form can be a piece of hard wood or fiber rod or tubing, with the

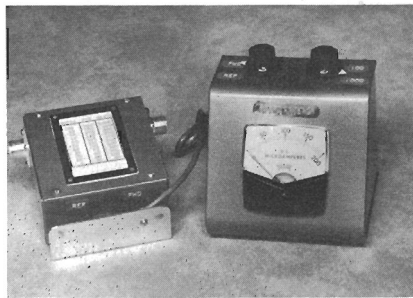
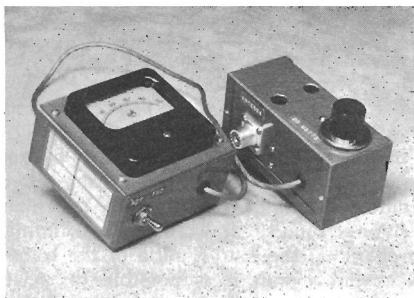
wires anchored through small holes at the ends of the form.

The switch was built in a $4 \times 5 \times 3$ -inch utility box, with the transformer and tube on top of the chassis and the remaining components inside. The tube socket was mounted close to the input connector J_1 . A little trouble with oscillation of the grounded-grid section was encountered when the unit was first tried, but this was curbed by connecting a lead from the rotor of C_1 to a common ground point instead of relying upon the chassis for a ground return.

It has been found that when the switch was installed in some ham stations a loss of gain occurred when the transmitter was connected to the switch, but at no time did the gain go below unity. This loss of gain only occurs when the "suck-out point" of the transmitter output circuit occurs at the frequency to which the receiver is tuned.² It has been my finding that by changing the L -to- C ratio of the transmitter's output circuit it is possible to move the suck-out point sufficiently to overcome this difficulty. It takes only a small change to correct the situation.

² See Campbell, "Some Variations in T.R. Switch Performance," *QST*, May, 1956.

» Practically all ham radio stations use some type of rf power meter or SWR indicator for tuneup and transmission line matching. This article points out some of the problems which are frequently encountered in designing and building reflected-power meters and SWR bridges. Examples of practical in-line rf watt-meters are given here, along with complete details for building a unit that will provide two power ranges, forward and reflected, for use from 3.5 to 30 MHz.



In-Line RF Power Metering

DOUG DEMAU, WICER

It is neither costly or difficult to build an rf wattmeter. And, if the instrument is equipped with a few additional components it can be switched to read reflected power as well as forward power. With the foregoing feature the instrument can be used as an SWR meter for antenna matching and transmatch adjustments.

Perhaps the most difficult task faced by the constructor is that of calibrating the power meter for whatever wattage range he desires to have. The least difficult method is to use a commercial wattmeter as a standard. If one is not available, the power output of the test transmitter can be computed by means of an rf ammeter in series with a 50-ohm dummy load, using the standard formula, $P = I^2 R$. Or, if one is not interested in obtaining power readings the bridge can be used solely as an SWR indicator, as is done with the Monimatch-style SWR bridge.¹

The advantage of the circuits shown here over those of Monimatch bridges is that these instruments are not frequency-sensitive. Monimatch indicators become more sensitive as the operating frequency is increased, thus making it impractical to calibrate them in watts for more than one band, or for more than one portion of a given band. The units described here are more sensitive than Monimatches are. This makes it possible to calibrate them for power levels as low as 1 watt, full scale, in any part of the hf spectrum.

All of the circuits shown in this article are similar to the basic one which was described in an earlier issue of *QST*.² Some of the circuits are those of commercial power meters, and are used to illustrate variations in the basic Bruene design. The reader may wish to experiment with some of these circuits.

Design Philosophy

Referring to the circuit of Fig. 1B, the circuit used by Collins Radio Company, the transmission-line center conductor passes through the center of a toroid core and becomes the primary of T_1 . The multi-turn winding on the core functions as the transformer secondary. Current flowing through the line-wire primary induces a voltage in the secondary which causes a current to flow through resistors R_5 and R_6 . The voltage drops across these resistors are equal in amplitude, but 180 degrees out of phase with respect to common or ground. They are thus, for practical purposes, respectively in and out of phase with the line current. Capacitive voltage dividers, C_3C_7 and C_4C_8 , are connected across the line to obtain equal-amplitude voltages in phase with the line voltage, the division ratio being adjusted so that these voltages match the voltage drops across R_5 and R_6 in amplitude. (As the current/voltage ratio in the line depends on the load, this can be done only for a particular value of load impedance. Load values chosen for this standardiza-

From December, 1969, *QST*.

¹ "Monimatch Mark II," *QST*, Feb. 1957.

² Bruene, "An Inside Picture of Directional Wattmeters," *QST*, April 1959.

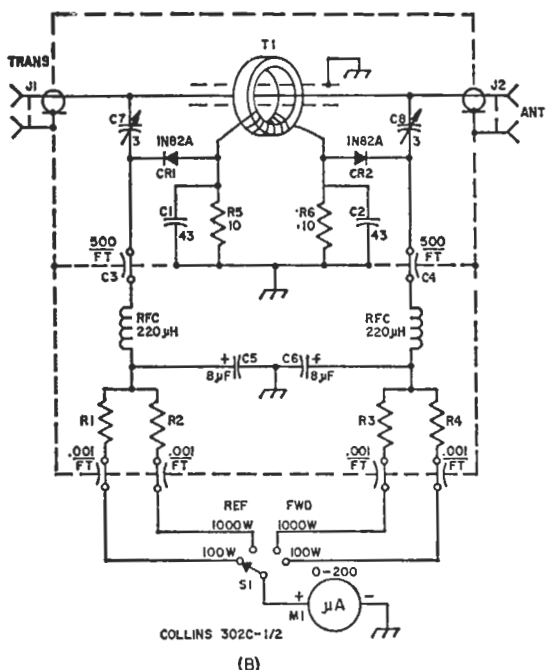
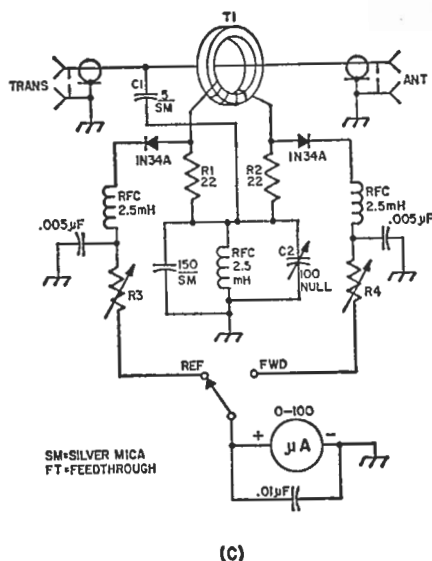
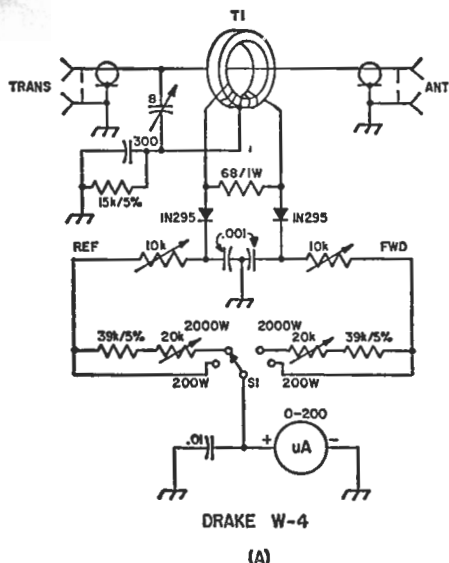


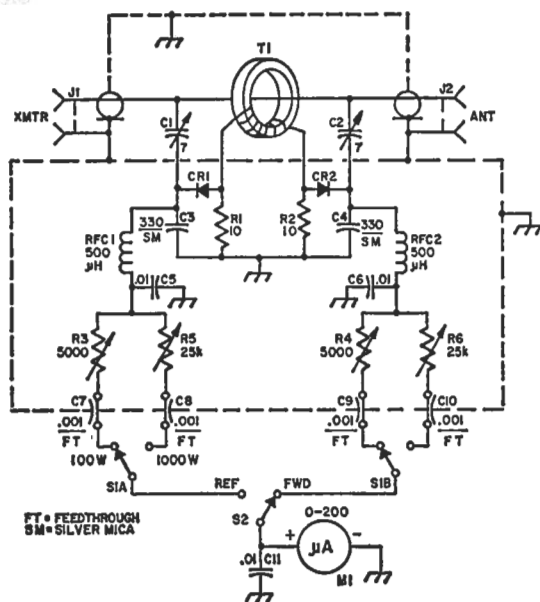
FIG. 1—Schematic diagrams of typical in-line power meters. At A, the R. L. Drake W-4 instrument. It uses a center-tapped transformer at T_1 , and has but one capacitive voltage divider in the sensing circuit. The circuit at B is discussed in the text, and is used by Collins Radio Company. The capacitive voltage dividers in this circuit use two 500-pF feedthrough capacitors in place of the silver-mica capacitors specified in Fig. 2. Capacitors C_5 and C_6 permit a charge time that enables the meter to read near-peak power on ssb. Calibrating resistances R_1 through R_4 are factory selected. The circuit at C is similar to one used by Comdel in their power meter. In this circuit C_1 is a fixed-value (small) capacitor, and the bridge is nulled by the larger capacitor in the divider, C_2 .

tion are pure resistances that match the characteristic impedance of the transmission line with which the bridge is to be used, 50 or 75 ohms usually.) Under these conditions, the voltages rectified by CR_1 and CR_2 represent, in the one case, the vector *sum* of the voltages caused by the line current and voltage, and in the other, the vector *difference*. With respect to the resistance for which the circuit has been set up, the sum is proportional to the forward component of a traveling wave such as occurs on a transmission line, and the difference is proportional to the reflected component.

The Collins circuit uses two 8- μ F capacitors, C_5 and C_6 , to permit the meter to approach the PEP level during ssb operation. The dc voltages in the forward and reflected lines charge the capacitors to permit a near-peak reading. The discharge rate is set by the series calibrating resistors, R_1 through R_4 , and is dependent upon which of them is switched into the metering line at a given time. The circuit of Fig. 1B uses two 43-pF capacitors, C_1 and C_2 , to cancel the inductive reactances of R_5 and R_6 . Such reactance may become manifest at the high end of the range for which the instrument is built. If reactance is present in that part of the circuit the meter readings may not be accurate, especially at 10 and 15 meters. The capacitors were not needed in the circuit of Fig. 2, perhaps because the resistor leads were very short when they were mounted on the etched-circuit board.

Some Design Hints

It is important that the layout of any rf bridge be as symmetrical as possible if good balance is to be had. The circuit-board layout for the instrument of Fig. 2 meets this requirement. Also, the input and output ports of the equipment should be isolated from the remain-



Top view of the rf head for the circuit of Fig. 2. A flashing-copper shield isolates the through-line and T_1 from the rest of the circuit. The second shield (thicker) is not required and can be eliminated from the circuit. If a 2000-watt scale is desired, fixed-value resistors of approximately 22,000 ohms can be connected in series with high-range printed-circuit controls. Or, the 25,000-ohm controls shown here can be replaced by 50,000-ohm units.

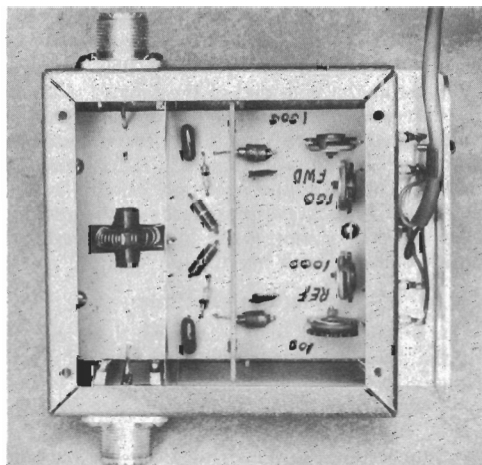


FIG. 2—Schematic diagram of a practical power watt-meter. A calibration scale for M_1 is shown also. Fixed-value resistors are $\frac{1}{2}$ -watt composition. Fixed-value capacitors are disk ceramic unless otherwise noted. Decimal-value capacitances are in μ F. Others are pF. Resistance is in ohms; $k=1000$.

C_1, C_2 —1.3 to 6.7-pF. miniature (E. F. Johnson 189-502-4. Available from Newark Electronics, Chicago, Ill.).

C_3 — C_{11} , inc.—Numbered for circuit-board identification.

CR_1, CR_2 —Small-signal germanium diode. IN34A, etc. (see text).

J_1, J_2 —Chassis-mount coax connector of builder's choice. Type SO-239 used here.

M_1 —0 to 200- μ A meter (Triplett type 330-M used here).

R_1, R_2 —Matched 10-ohm resistors (see text).

R_3, R_4 —5000-ohm printed-circuit carbon control (IRC R502-B).

R_5, R_6 —25,000-ohm printed-circuit carbon control (IRC R252-B).

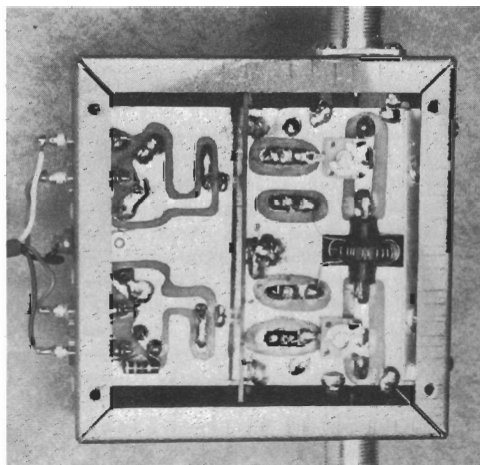
RFC_1, RFC_2 —500- μ H rf choke (Millen 34300-500 or similar).

S_1 —Dpdt single-section phenolic wafer switch (Mallory 3222J).

S_2 —Spdt phenolic wafer switch (Centralab 1460).

T_1 —Toroidal transformer; 35 turns of No. 26 enam. wire to cover entire core of Amidon T-68-2 toroid (Amidon Assoc., 12033 Otsego St., N. Hollywood, Ca. 91607).

Bottom view of the rf head for the circuit of Fig. 2. The fixed-value resistor at the lower left does not belong in the circuit, but was added as a shunt for one of the calibrating controls which was too high in value—a 50,000-ohm unit that was on hand. The shield partition shown here proved unnecessary and can be eliminated.



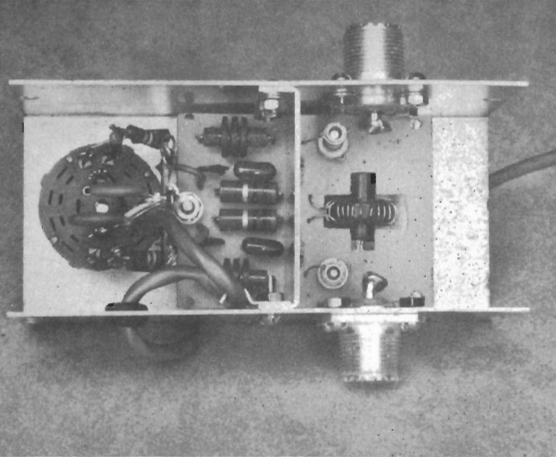


FIG. 3—Inside view of the 5- and 50-watt power meter rf head. Component values are the same as in the circuit of Fig. 2, except for the calibrating resistances (see text). An aluminum shield isolates the through-line and toroid from the remainder of the circuit. A 50- μ A meter is used in this model.

der of the circuit so that only the sampling circuits feed voltage to the bridge. A shield across the end of the box which contains the input and output jacks, and the interconnecting line between them, is necessary. If stray rf gets into the bridge circuit it will be impossible to obtain a complete zero reflected-power reading on M_1 even though a 1:1 SWR exists.

Referring again to Fig. 2, resistors R_1 and R_2 should be selected for the best null reading when adjusting the bridge into a resistive 50- or 75-ohm load. Normally, the value will be somewhere between 10 and 47 ohms. The 10-ohm value worked well with the home-made instruments shown here. It was found that half-watt resistors exhibited somewhat less inductive reactance at 30 MHz than did some one-watt units tried. R_1 and R_2 should be as closely matched in resistance as possible. They need not be exactly 10 ohms, so a vtvm can be used to match them. The resistors used for the circuit of Fig. 2 were actually 10.5 ohms each, and were chosen from an assortment of "10-percenters" on hand.

Silver-mica capacitors C_2 and C_4 were close enough in value so that special selection was not required. There should be enough leeway in the ranges of C_1 and C_2 to compensate for any difference in the values of the 330-pF capacitors. Ideally, however, C_3 and C_4 should be matched in value.

Diodes CR_1 and CR_2 should also be matched

for best results. An ohmmeter can be used to select a pair of diodes whose forward dc resistances are within a couple of ohms of being the same. Similarly, the back resistances of the diodes can be matched. The matched diodes will help to assure equal meter readings when the bridge is reversed. (The bridge should be perfectly bilateral in its performance characteristics.) Germanium diodes are used in the bridges described here, but silicon diodes can also be used. Silicon diodes conduct at a higher voltage than germanium diodes do—approximately 0.7 volt—and will not work too well in low-power wattmeters. Some silicon diodes were tried, but ceased to conduct at approximately 8 watts in the circuit of Fig. 2. This effect can cause misleading results when low values of reflected power are present during antenna adjustments. The SWR can appear to be zero when actually it isn't. The germanium diodes conduct at approximately 0.3 volt, making them more suitable for low-power readings.

Any meter whose full-scale reading is between 50 microamperes and 1 milliampere can be used at M_1 . The more sensitive the meter, the more difficult it will be to get an absolute reflected-power reading of zero. Some residual current will flow in the bridge circuit no matter how carefully the circuit is balanced, and a sensitive instrument will detect this current flow. Also, the more sensitive the meter, the larger will have to be the calibrating resistances, R_3 through R_6 , to provide high-power readings. A 0 to 200-microampere meter represents a good compromise for power ranges between 100 and 2000 watts.

Construction

The power meter of Fig. 2 is built in two sections. The rf circuit and the calibrating resistors are housed in a 4 \times 4 \times 2-inch aluminum utility box. All components other than J_1 , J_2 , and the feedthrough capacitors, are assembled on the etched-circuit board.³ Switches S_1 and S_2 , and the meter, M_1 , are installed in a sloping-panel utility box which measures 5 \times 4 inches. Four-conductor shielded cable—the shield serving as the common lead—is used to join the two pieces. There is no reason why the entire instrument cannot be housed in one container, but it is sometimes awkward to have coaxial cables attach to a unit that occupies a prominent place in the operating position. Built as shown, the two-piece instrument permits the rf pickup head to be concealed behind the

³ The etched-circuit board pattern and parts layout sheet for this power meter are available from ARRL Hq. Send 25 cents and a SASE. Ready-made circuit boards can be purchased from Stafford Electronics, 427 S. Benbow Rd., Greensboro, N.C. 27401.

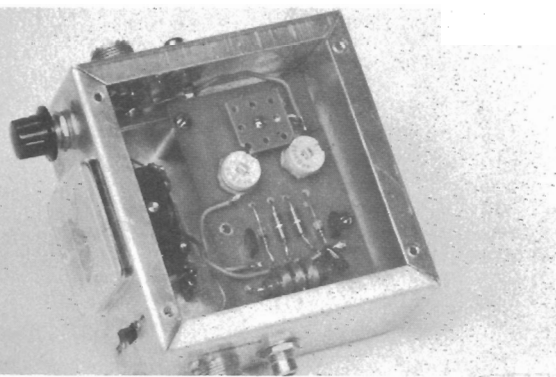


FIG. 4—Inside view of a 3-watt power meter for QRP rigs. Its circuit is given on page 16 of June 1969 QST. Ceramic trimmers are used for nulling the bridge. Type SO-239 connectors are paralleled with phono jacks to add versatility. A 4 \times 4 \times 2-inch utility box houses the entire unit.

transmitter, while the control head can be mounted where it is accessible to the operator.

Toroidal transformer T_1 fits into a cutout area on the circuit board. A 1-inch long section of RG-8/U cable—vinyl jacket and shield braid removed—provides a snug fit in the center hole of the toroid, and is used to complete the line between J_1 and J_2 . The inner coordinator of the RG-8/U section solders to the circuit board, thus holding T_1 in place.

A flashing-copper shield divides T_1 and its center-conductor line from the remainder of the circuit. This partition is shown in dotted lines in Fig. 2. It is mounted on the non-foil side of the circuit board and is secured at each end to solder lugs which are mounted under the retaining screws for J_1 and J_2 .

The circuit board is held in place, at the end near T_1 , by means of an aluminum L bracket. The circuit-board end nearest the feedthrough capacitors is held in place by a No. 6 spade bolt. A solder lug is mounted under the No. 6 nut (outside the case) which secures the spade bolt. The lug serves as a connection point for the common lead between the rf head and the control box. Two solder lugs are mounted under the bottom two retaining screws of each coax connector. The free ends of the lugs are soldered to the copper foil of the circuit board.

A partition is visible in the foil-side view of the rf head. It can be eliminated if desired, since it did not prove necessary when the unit was tested. Similarly, an extra shield partition is shown on the top side of the board. It too can be eliminated, for it turned out to be unnecessary. The flashing-copper shield discussed earlier is the only one required for the circuit of Fig. 2.

Check-out and Tuneup

Once the instrument is wired and ready to test it should be inspected for unwanted solder bridges between the circuit-board foils. It is usually a good idea to scrape out the rosin buildup between the foils, and this can be done with the blade of a small screwdriver. A continuity check for "opens" and "shorts" should also be made before power is applied to the unit. Make certain that the diodes are installed for the correct polarity—the banded ends (cathodes) toward C_1 and C_2 .

Connect a noninductive 50-ohm dummy load to J_2 . A Heath Antenna or similar load will serve nicely for adjustment purposes. Place S_2

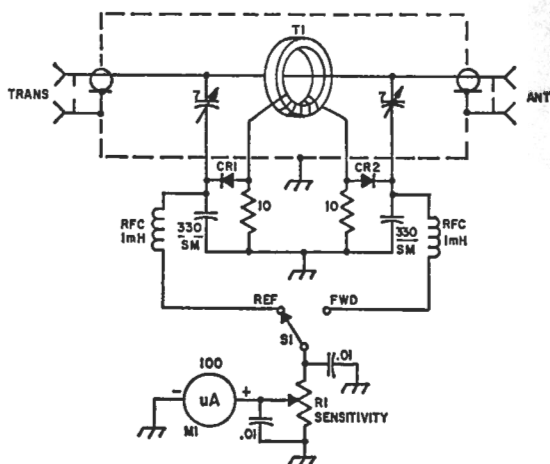
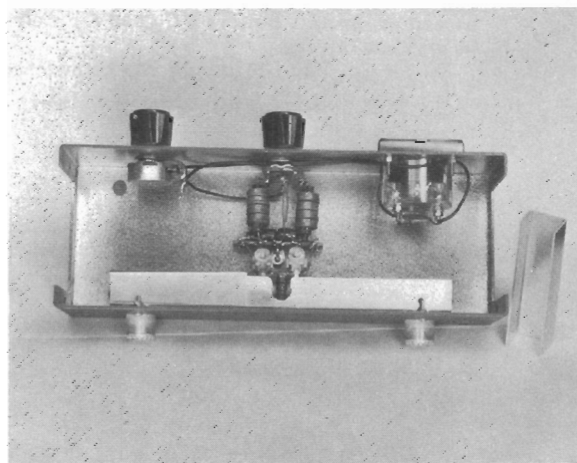


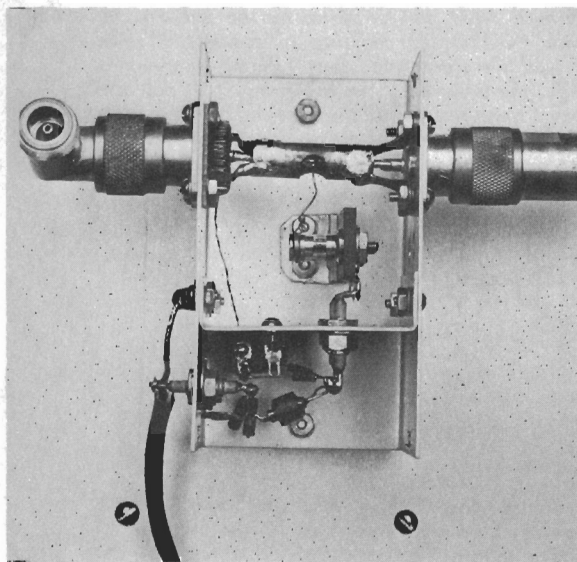
FIG. 5—Schematic diagram of the modified Heath HM-15 SWR meter. CR_1 , CR_2 , M_1 , R_1 and S_1 are original components from the HM-15. T_1 is the same as in Fig. 2. See text for additional details.

in the FORWARD position, and set S_1 for the 100-watt range. An rf ammeter or calibrated power meter should be connected between J_2 and the dummy load during the tests, providing power calibration points against which to plot the scale of M_1 . Apply transmitter output power to J_1 , gradually, until M_1 begins to deflect upward. Increase transmitter power and adjust R_1 so that a full-scale meter reading occurs when 100 watts is indicated on the rf ammeter or other standard in use. Next, switch S_2 to REFLECTED and turn the transmitter off. Temporarily short across R_3 , turn the transmitter on, and gradually increase power until a meter reading is noted. With an insulated screwdriver adjust C_2 for a null in the meter reading.

The next step is to reverse the coax connections to J_1 and J_2 . Place S_2 in the REFLECTED position and apply transmitter power until the meter reads full scale at 100 watts output. In this mode the REFLECTED position actually reads forward power because the bridge is reversed. Calibrating resistance R_3 is set to obtain 100

Inside view of the modified Heath HM-15 bridge. The new components are grouped at the center of the chassis on a 5-lug terminal strip. The nulling capacitors are connected between the inner line and the terminal strip. Press-fit aluminum shield covers are slipped over the trough line to aid in rf isolation. One cover is in place; the other is at the right of the photo. Improved shielding might be effected by installing an aluminum plate between the terminal strip and the two nulling capacitors.





Inside view of a 2000-watt power meter built by W1KLK. This bridge is patterned after the circuit of Fig. 1B. Point-to-point wiring is used throughout, thus avoiding the need for a circuit board. Two piston trimmers are used for the nulling capacitors and are mounted one above the other on a phenolic block. The two 500-pF feedthrough capacitors are part of the capacitive voltage dividers.

watts full scale during this adjustment. Now, switch S_2 to FORWARD and temporarily place a short across R_4 . Adjust C_1 for a null reading on M_1 . Repeat the foregoing steps until no further improvement can be obtained. It will not be necessary to repeat the nulling adjustments on the 1000-watt range, but R_5 and R_6 will have to be adjusted to provide a full-scale meter reading at 1000 watts. If insufficient meter deflection is available for nulling adjustments on the 100-watt range, it may be necessary to adjust C_1 and C_2 at some power level higher than 100 watts. If the capacitors tune through a null, but the meter will not drop all the way to zero, chances are that some rf is leaking into the bridge circuit through stray coupling. If so, it may be necessary to experiment with the shielding of the through-line section of the rf head. If only a small residual reading is noted it will be of minor importance and can be ignored. In the circuit of Fig. 2 there remained approximately one half a meter division when the null was reached, and this occurred only on the 1000-watt range. Since this was representative of less than 2 watts of power it was deemed inconsequential.

With the component values given in Fig. 2 the meter readings track for both power ranges. That is, the 10-watt level on the 100-watt range, and the 100-watt point on the 1000-watt range fall at the same place on the meter scale, and so on. This no doubt results from the fact that the diodes are conducting in the most linear portion of their curve. Ordinarily, this

desirable condition does not exist, making it necessary to plot separate scales for the different power ranges.

Tests indicate that the SWR caused by insertion of the power meter in the transmission line is negligible. It was checked at 28 MHz and no reflected-power could be noted on a Bird wattmeter. Similarly, the insertion loss was so low that it could not be measured with ordinary instruments.

Other Circuits

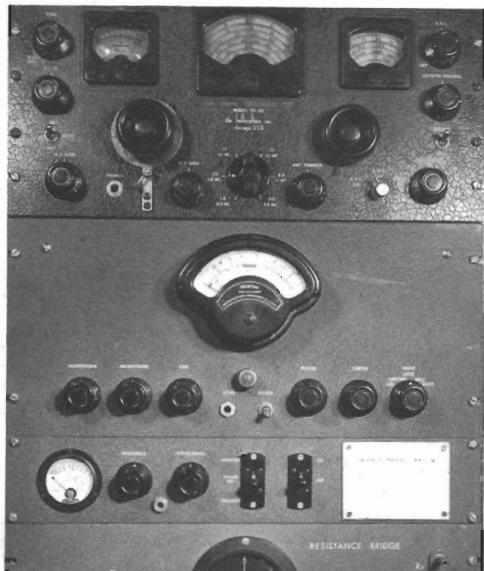
Additional circuits and photos are shown for variations in the basic design used at Fig. 2. A low-power model, having scales for 5 and 50 watts, is shown in Fig. 3. It uses fixed-value resistors for meter calibration. The required values of resistance were first determined by temporarily inserting a potentiometer in the meter line, obtaining the required full-scale reading, then substituting fixed-value resistances of the proper ohmage. The meter readings for the two power ranges do not track in this model.

A low-power meter was designed for use with the QRP transmitter described in June 1969 QST. It is shown in Fig. 4, and has a full-scale calibration of 3 watts. To obtain additional sensitivity, the primary of the toroidal transformer consists of a one-turn link instead of the single wire that would normally pass through the hole in the toroid core.

Some experiments were conducted to see if a Heath HM-15 SWR bridge (a Monimatch type) could be modified to work in a Bruene circuit. The results were satisfactory, and the circuit is given in Fig. 5. No attempt was made to obtain a calibration scale for the meter. The unit is being used as a simple SWR indicator, but now has better sensitivity in the lower part of the hf spectrum—7 watts, full scale, from 3.5 to 30 MHz. Also, the instrument is no longer "frequency-conscious" as was the case before modification. The original pick-up lines were discarded, the FWD-REV panel switch was rotated 180 degrees so that the labels were correct for the new circuit, and press-fit shield covers were installed on the trough line as shown in the photo. A power scale could be plotted by setting the sensitivity control in a fixed position—possibly replacing the existing control with a screwdriver-adjust type. A new 100- μ A meter could be installed to provide a better scale for calibration in watts.

It was necessary to dismantle the trough line so that the toroidal transformer could be slipped over the inner line. A few wraps of mylar tape were wound over the center of the inner line to insulate the toroid winding from the line, and to provide a snug fit to keep the toroid in place. The trough was notched out with a nibbling tool to allow clearance for the toroidal transformer. Additional shielding can be added between the line and the rest of the circuit to further assure a zero meter reading in the reflected position.

» Effective January 1, 1969, tariffs set up by the Bell System in response to an FCC Order covering interconnection with privately-owned facilities went into effect. Phone patching is no longer an under-the-table business, but bringing it out into the open also brings some responsibilities that the amateur must shoulder. Here's the dope, including the description of a phone-patch circuit that meets the technical requirements.



The telephone companies associated with the A.T.&T. Co. (The "Bell System") recently filed tariff changes with their regulatory bodies that are of importance because they permit interconnection on a completely legitimate basis—on the condition of course, that all applicable tariff provisions are complied with. Independent telephone companies (those not associated with the Bell System) will determine their own policies in this matter but they will probably choose to make similar offerings. While the new tariff provisions offer several interconnection arrangements for both voice and data equipment, only the voice equipment interconnection will be discussed here.

Of interest to the amateur operator is the offering whereby the telephone company will provide a voice connecting arrangement to be associated with a telephone set. The company will provide a device known as a "Voice Coupler" and a key or switch to connect and disconnect the coupler. In Illinois, the rate established for the connecting arrangement includes an initial charge of \$5.00 and a monthly charge of \$0.50. The rate and service became effective on January 1, 1969. The availability and rates in other states are subject to the approval of each state's regulatory body. As a matter of practical operating convenience, an amateur operator would want the telephone and voice

Phone Patching

GEORGE P. SCHLEICHER, W9NLT

coupler close to the operating position. For most of us that would mean that we would have to order an extension telephone (at the regular tariff rates) in addition to the connecting arrangement.

Operation

In operation the telephone set would be used to originate an outgoing or answer an incoming telephone call. The switch would then be operated to complete the circuit to the voice coupler. Options may be available that will permit the handset transmitter and/or receiver to be disabled when the coupler is in use. (Disabling the handset transmitter might be desirable so as to avoid the pickup of room noise; some operators might prefer to leave the handset receiver in operation as it would be convenient for monitoring the conversation in progress through the radio facility). Connection to the voice coupler is simple. It is equipped with a jack that will accept a standard ¼-inch phone plug such as a PL-55 or equivalent. See Fig. 1. Coupler input may be balanced or have one side grounded.

The voice coupler will remain telephone company property and its circuitry is subject to change from time to time. It can be expected to contain a number of electrical components that will include a transformer for isolation purpose, diodes for limiting electrical levels, and some capacitors. The options mentioned above may not be available in every case and they may change over a period of time.

User's Responsibility

An amateur using the new interconnection privileges will be responsible for operating in accordance with the FCC rules, for conforming to all provisions of the applicable tariffs of the telephone company serving him, and for meeting certain technical criteria for the protection of the telephone network. We should all have a copy of the FCC rules applying to Amateur Service somewhere handy for reference. Telephone company tariffs are available at most of



The voice coupler, to the left of the telephone, is supplied by the phone company, ready for the phone patch to be plugged in. The coupler is normally fixed to a wall or desk and serves also as a connecting block. Lifting the left hookswitch plunger connects the coupler and disables the handset.

their business offices as well as people who will explain any provisions that might be difficult to understand. The technical criteria have been filed with the FCC in the A.T.&T. Company's tariff No. FCC 263. It may also appear in local tariffs.

Briefly, the electrical limitations at the point of interconnection as defined in the tariff are as follows:

- The average voice level¹ to be applied at the point of interconnection will be specified by the telephone company so that the voice signal power will be received at the serving central office at a level that is not stronger than -12 dbm.; the power level at the point of interconnection will not be in excess of one milliwatt (0 dbm.) in any case
- Power in the band from 3.995 to 4.005 kilohertz shall be at least 18 db. below the average level as specified above.
- Power in the band from 4.0 to 10.0 kilohertz shall not exceed -16 dbm.
- Power in the band from 10.0 to 25.0 kilohertz shall not exceed -24 dbm.
- Power in the band from 25.0 to 40.0 kilohertz shall not exceed -36 dbm.
- Power in the band above 40.0 kilohertz shall not exceed -50 dbm.
- At no time should signal power be present *only* in the band from 2.450 to 2.750 kilohertz; when present, it must not exceed the power that is present simultaneously in the 0.8 to 2.450 kilohertz band.

The amateur's equipment should be designed

so as to have an internal output impedance of 900 ohms at the point of connection with the voice coupler. Its maximum output power should be no more than zero dbm. (0.95 volt, r.m.s.) and be adjustable down to -12 dbm. (0.238 volt, r.m.s.) at the voice-coupler terminals. Operating levels should be carefully observed, as excessive levels will cause distortion that results from clipping in the voice coupler. The amateur's equipment must not apply any d.c. potential to the coupler.

The Problem

Amateurs are familiar with radio systems which inherently provide separate paths for the two directions of transmission. The wire line equivalent of this arrangement involves the use of 4 wires; one 2-wire circuit for sending and another 2-wire circuit for receiving. Normal residence telephone service (in the U.S.) uses the same two conductors for both transmitting and receiving. To interconnect radio sending and receiving equipment with such a line requires that we achieve three objectives. First, we must comply with all of the technical limitations in the tariff. Second, our equipment must be able to recognize which direction of transmission is intended (the "land" party sending or receiving) and condition itself accordingly. Third, there will be a significant difference in the level of the signal that is received over the telephone line from one call to another as compared with the level of the signal sent over it from the amateur's equipment.

PROVIDED BY AMATEUR

PROVIDED BY TELEPHONE CO.

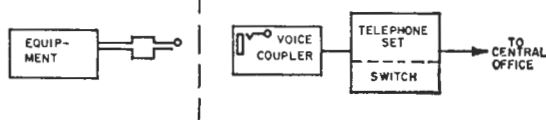


FIG. 1—Interconnection arrangement.

Taking the level problem first, we know that our sending level will be somewhere between zero and -12 dbm; it is to be specified by the telephone company. The received level can be expected to arrive over a much wider range. If the telephone communication is between two locations that are served by the same telephone central office and they are close to it, the received level might be in the order of -10 dbm. depending on the type of instruments, central office equipment, cables, etc. On the other hand, the received level on a long distance call might occasionally drop to the -30 s or even the -40 s.

The other serious problem is that of recognizing which direction of transmission should be open. Of course, we can always put a switch in the voice path and operate it manually, perhaps in connection with our carrier control. Such a method is cumbersome, however, and would preclude the rapid back and forth exchange of comment that amateurs are likely to be transmitting.

¹ The term "voice level" as used here means the peak level averaged over a 3-second interval as read with a meter having standardized speed of response, such as a "v.u." meter or a fast-acting rectifier-type meter.

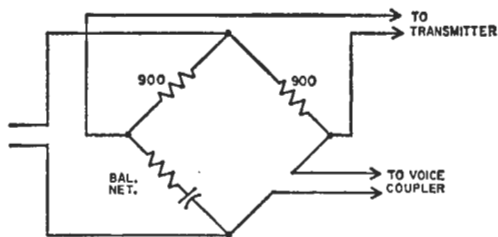


FIG. 2—Wheatstone Bridge hybrid.

An automatic device that will put the transmitter in an operative condition is called a "VOX" circuit; the letters refer to "voice operated transmission." Whether the voice signals come from a microphone or a telephone line, there is a need to inhibit the VOX circuit so that it will not respond to signals coming from the station receiver. Such added circuitry is often referred to as an "anti-VOX" feature. It now appears that we need an electrical circuit that will interconnect our receiver output and our transmitter input with the telephone voice coupler in a way that will greatly reduce the amount of audio energy that goes directly from the receiver into the transmitter. By doing this we will be able to use simpler VOX and anti-VOX circuits and their adjustment will be far less critical. A number of adequate VOX and anti-VOX circuits have already been described in amateur radio publications. Information on combining circuits, usually called "Hybrids," is pretty scarce, however, and so several kinds are described below.

While VOX equipment is not the subject of this article it is so closely related to the success or failure of interconnection that a few suggestions are in order. The VOX and anti-VOX circuits should give control of the transmission direction to the "land-line" party. A

voice signal from the telephone line should disable the receiving path and enable the transmitting path, in that sequence. When the land party stops talking the transmitting path should be disabled and the receiving path enabled, in that order. Switching of the transmission direction should take place in about 100 milliseconds. Good station design will permit a two-way conversation to take place over an a.m. or s.s.b. radio system: even though transmitting and receiving are done on the same frequency.

Hybrid Circuits

A number of different hybrid circuits have been developed. All of them rely on dividing voice-energy currents that come from the receiver between the wire line and an identical dummy load or "balancing network." The transmitter is connected to the two current paths in such a way that these currents and/or voltages are zero at the transmitter connection. Energy from the wire line will be divided among the various circuit elements, one of which will be the transmitter audio path. Hybrid balance requires that the network simulate the electrical impedance at the voice coupler terminals as closely as possible. A simple R-C series circuit will usually be adequate, but unusual conditions might require the addition of an inductive element.

Perfect balance at all frequencies will never be achieved, of course, due to the complexity of the impedance of the telephone set and line as seen at the terminals of the voice coupler. (Energy reflected from impedance irregularities in the line can also be troublesome). A well-designed hybrid should be capable of reducing the level of any one frequency from the receiver by over 50 db. as measured at the transmitter connection and it should be capable of reducing the level of all frequencies in the voice band by more than 36 db. By achieving this degree of reduction we cause the unwanted signal from the receiver to arrive at the transmitter at a level that is no higher—and hopefully well below—the level of the signal from the telephone line. *This effect is what really simplifies the design of our VOX and anti-VOX circuits.*

The simplest hybrid circuit is an adaptation of the Wheatstone bridge. It is easily constructed with few elements. The basic design is shown in Fig. 2. Good results can be obtained with this circuit but its use generally requires that the transmitter and receiver audio circuits be well balanced to ground or isolated by well-balanced transformers.

Another form of the Wheatstone bridge circuit can be made with a transformer that has three windings, two of which are center-tapped. The transformer must be connected as shown in Fig. 3. The taps on the windings must be accurate, however, and each of the windings must be well balanced to ground and to the other windings. Proper polarity of the two tapped windings is required, otherwise the output voltage will be zero across all three branches!

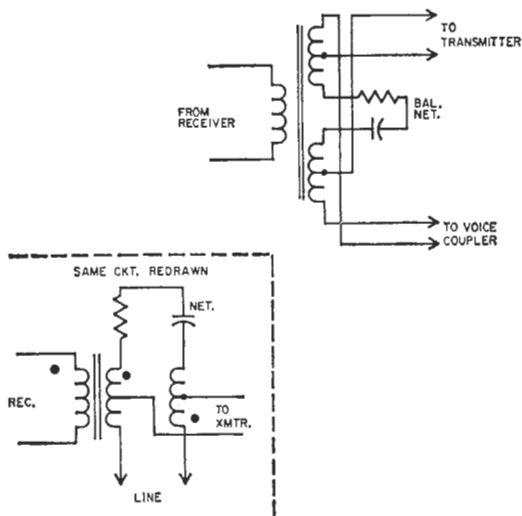
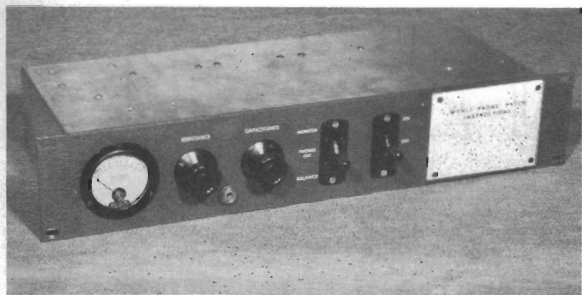


FIG. 3—Transformer hybrid.



W9NLT's hybrid includes a surplus db. meter for level indication. The relay-rack size panel provides ample room for the resistance and capacitance controls in the balancing network, as well as the on-off and monitoring-control switches. Since only voice frequencies are present, the physical layout can be practically any that suits the constructor.

A third hybrid configuration involves two transformers, each having three windings. While it is best if the transformers are well balanced and identical, the hybrid is not critical in this respect as long as the balancing network can be adjusted to compensate for the differences. The transformers would be wired as shown in Fig. 4. Energy from the receiver is coupled through the first transformer to the line and the balancing network. Currents in these two paths are made equal in magnitude and phase by adjustment of the network. The transformer windings are poled so that the magnetizing forces in the core of the second transformer cancel, resulting in no output toward the transmitter. Voice energy from the telephone line is divided between the transmitting and receiving branches of the circuit. The advantages of this circuit are that minor variations in the transformers are easily compensated for by network adjustment and that the transformers provide isolation between all of the branches of the hybrid.

Building A Hybrid

It is not difficult to build a hybrid that will be useful in making legitimate phone patches and while doing so, comply with all of the technical requirements that are now listed in the tariffs. My hybrid uses the two-transformer ar-

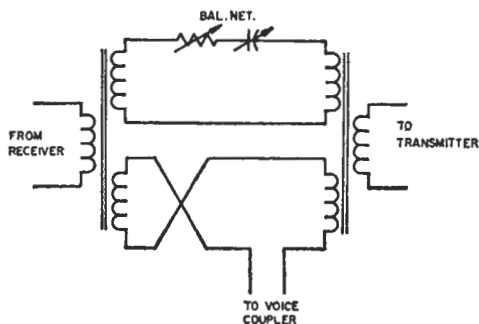


FIG. 4—Two-transformer hybrid.

range and includes such refinements as a level-indicating meter, provision for a fixed attenuator for meeting telephone company level requirements, earphone connections for monitoring and balancing, and finally a low-pass filter. The circuit is shown in Fig. 5. The reasons for some of my choices may be of interest. I found that the best quality audio transformers that were available at low cost were the "telephone repeating coils" offered by several surplus stores who are QST advertisers. Most of these transformers have split primary and secondary windings consisting of either four or six individual windings; turns ratios are commonly 1:1, 1:1½ or 1:2. Almost any pair of transformers may be used if they are of the same type. Fig. 6 shows one of the more common terminal numbering plans. The 9-10 and the 11-12 windings are seldom used alone; they usually are connected in series with the 3-4 and 7-8 windings if they are used at all.

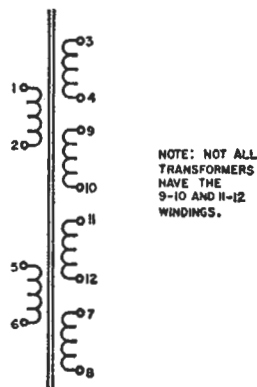


FIG. 6—Repeating-coil terminal numbering system.

If you are not sure of the quality of your transformers, a simple test can be made to determine which transformer and which of its windings will give the best balance. Assume that you have a 4-winding transformer of the type shown in Fig. 6. Its primary terminals are numbered 1-2 and 5-6; the secondary terminals are 3-4 and 7-8. Connect 2 to 5 and 4 to 7; also connect 1 to 6, shorting the primary. Supply a 1-kHz. tone at about 1-milliwatt level to the primary, connecting the source to terminals 1-6 and 2-5. If the transformer primary is well balanced, the level measured at terminals 3 and 8 will be -70 dbm. or lower. Reverse the transformer, feeding power to the now-short-circuited secondary and measuring the power in the primary at terminals 1 and 6. This arrangement determines the degree to which the secondary winding is balanced. Whether testing one transformer or a half-dozen, you will find one primary or secondary winding that is better balanced than all of the others. Use that transformer to supply energy to the transmitter branch of the hybrid (T_2 in Fig. 5). The winding that showed the best

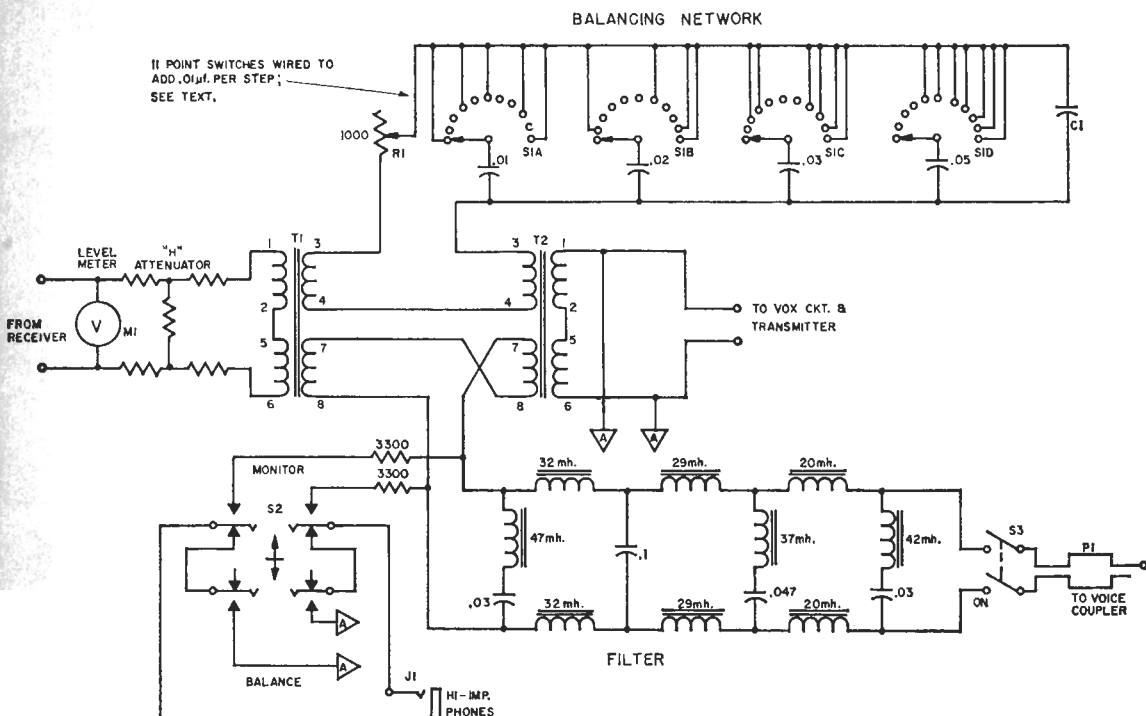


FIG. 5—Circuit diagram of complete hybrid. Capacitances are in $\mu\text{f.}$; capacitors are paper or mylar, 50 volts. Resistances are in ohms; resistors are $\frac{1}{2}$ watt. See text, Tables I and II, and Appendix for discussion of component types and values.

C_1 —App. 2 $\mu\text{f.}$; see text.

J_1 —Insulated phone jack.

M_1 —Volume-level indicator meter.

P_1 —Stondord phone plug.

R_1 —Linear control.

S_1 —Rotary, 4 sections, 1 pole per section, 11 contacts per section.

S_2 —4-pole double-throw key switch.

S_3 —2-pole single-throw toggle or key switch.

T_1, T_2 —Audio, 1:1, split windings.

balance should be split for connection to the line and balance-network. A transformer that shows less than about 55 db. loss when tested this way should not be used on the transmitter side of the hybrid; it might be OK on the other (T_1) side, however.

Filters

A filter is recommended at the point of connection with a voice coupler. Properly designed and built, a filter offers several advantages to the amateur. It will assist greatly in meeting the technical limitations for which the amateur is responsible and it can help to achieve good balance in the hybrid. That is because a properly designed filter will present a more constant impedance to the hybrid than will the voice coupler. It will not completely isolate the hybrid from impedance changes beyond the coupler, however. Some hams might prefer to connect the filter between the receiver and the hybrid, in which position it will meet all technical requirements as well as limiting the range of frequencies over which the hybrid must be balanced.

Inductors for filters are of odd values and can be costly if purchased new. I recommend ac-

quiring a dozen or so of the 44- and 88-millihenry "loading coils" that are showing up in surplus. My design for a filter uses inductance values that can be obtained by modifying these coils.² The toroids selected can be the smallest size available (if you have a choice) for either value of inductance since the voice power is low and there will be no d.c. present that might tend to cause saturation. If you use this kind of inductor, remember that the two windings must be connected in series aiding in order to obtain the rated inductance.

My trip to the surplus store also netted a couple of lever-type switches and a fast acting "db." meter designed to be bridged across a 600-ohm circuit. All of the other parts—panel, chassis, terminal boards, etc.—came from normal sources, some by way of "Ye Olde Junke Boxx". In fact, a large instruction card on the front panel of my unit is really there to hide some unused holes. It ought to be emphasized that no part of the hybrid is so special that other parts could not be substituted if good judgment is used in their selection. Component tolerances of $\pm 20\%$ are acceptable except in the

² Wetherhold, "Inductance And Q Of Modified Surplus Toroidal Inductors", *QST* Sept., 1968.

filter, where components should be within 5% of the values that are specified.

Levels

The loss of this kind of hybrid will be found to be close to 4.5 db., plus the loss of the filter. Convenient levels for metering will probably be in the order of zero to +3 dbm., because of the scales on available meters. An attenuator may be required between the metering point and the voice coupler to reduce the voice level to that specified by the telephone company. If it is placed between the metering point and the hybrid it will minimize the energy level that the hybrid must balance for the benefit of the transmitter branch. It could also be located between the hybrid and the filter, in which case it would simplify the work of balancing the hybrid by isolating much of the reactance in the load. This arrangement is desirable, however, only if the improvement in balance that can be obtained (by placing the attenuator here) exceeds *twice* the loss of the attenuator when both are expressed in db. Resistor values for various attenuator losses are given in Table I below. The values are computed ones; the nearest commercial values may be used in construction.

Practical Details

Two switches are provided on the finished unit; one disconnects it from the voice coupler and one is used to switch an earphone jack from a monitoring position across the line to a position across the transmitter audio connection. This permits the phones to be used during the course of a patched communication both for monitoring and to determine that the network is adjusted for optimum during the progress of the call.

The balance network is arranged with adjustable resistance and capacitance values. A 1000-ohm wirewound pot is used as the resis-

TABLE I

Resistance values for 600-ohm "H" attenuators; multiply all values by 1.5 for 900-ohm attenuators.

Loss	Series Arms	Shunt Arms
1 db.	17 ohms	5200 ohms
2	34	2580
3	51	1700
4	67	1260
5	84	986
6	100	800
7	115	670
8	129	568
9	145	488
10	156	422

Note: Four resistors of the value shown for the "series" arms are required.

TABLE II

Step	Total Capacity	Capacitor Unit			
		0.01	0.02	0.03	0.05
1	0.01	×			
2	0.02		×		
3	0.03			×	
4	0.04	×		×	
5	0.05				×
6	0.06	×			×
7	0.07		×		×
8	0.08			×	×
9	0.09	×		×	×
10	0.10		×	×	×
11	0.11	×	×	×	×

Note: An "X" indicates the connection of a capacitor unit on the step shown.

tance element. Four capacitors and a four-deck, eleven point switch are arranged to give eleven values of capacitance ranging from 0.01 to 0.11 μ f. in increments of 0.01 μ f. That is not normally enough to balance the hybrid so additional capacitance is placed in parallel with that in the switch when the hybrid is initially balanced. The small increments are about right for precise adjustment of the balance net. The way the switch is connected is indicated in Table II and Fig. 5.

Balancing The Hybrid

After constructing the phone patch it will be necessary to balance the hybrid. This means that the balancing network is adjusted to match the impedance that is connected to the "line" side of the hybrid. If the hybrid is completely unbalanced, the coupling from the receiver branch to the transmitter branch will be maximum. As the hybrid is brought into balance the energy that reaches the transmitter branch from the receiver is reduced. The amount of reduction can be expressed in decibels and this figure used as a measure of the hybrid's performance. With high-quality transformers and single-frequency testing, hybrid balance in the order of 70 db. can be obtained.

It is a good idea to test the hybrid before it is connected to the line. One way to make such a test is as follows: Connect the receiver branch to an oscillator or the station receiver. Feed a tone of 1 to 2 kilohertz into the hybrid at a level of one milliwatt. Disconnect the filter, leaving the line side of the hybrid open; short circuit the balance network and read the level of the signal at the transmitter leg. With this arrangement the transmitter branch is connected to the receiver branch through the two transformers. Good transformers of 1:1 ratio will have a loss of 0.5 to 1.5 db. each, so the level reading should be between -1 and -3

dbm. Now connect a resistance of about 900 ohms (give or take 100) to the line side of the hybrid and another of similar value to the network side. The resistive element in the balance network can be used, as one of the resistances should be adjustable to compensate for small differences in the transformers. Measure the signal level in the transmitter branch, adjusting the variable resistor for minimum level. The hybrid should now be in balance and the level should have dropped to -50 dbm, or lower. A hybrid that fails this test should be checked for incorrect wiring or a defective coil; a good hybrid should meet the above test at any voice-band frequency. After the hybrid has been tested the wiring should be restored to normal.

The next step is to balance the hybrid with the filter and coupler connected. If you have an impedance bridge you can save time by measuring the impedance that the filter and coupler present to the hybrid. Without the bridge some old fashioned cut-and-try will have to be used. The impedance that will have to be balanced cannot be estimated because it will depend on many variables. The filter, coupler and telephone set all contribute to the total impedance and yet each of them is largely transparent, in the sense that its input impedance is dependent on the load impedance that is connected to its output. Most of the elements mentioned usually have an impedance that is capacitive. If you choose to leave the receiver in the telephone handset operative it could shift the total impedance so that it would appear to be inductive. No two situations will be alike, but here are some suggestions: As a starting

point, use 900 ohms in series with 2 μ f. in the balancing network. A capacitor decade will help to speed the work along. Remember that a condition of balance must be achieved with a telephone connection established, so you may want to make some advance arrangements with a patient and understanding next door neighbor or a cross-town ham that you can talk to on the air while testing. If you find that the balance seems to be only fair and that the capacitance required in the network exceeds 3 μ f., it may be an indication that inductance will have to be added to the network. After good balance has been obtained at some frequency in the middle of the voice band, be sure to check the low and high ends. As a rule, a simple series circuit will be adequate in the balance network. Unusual circumstances, such as a change from positive to negative reactance with frequency, may require a more complicated network. Be careful to observe all technical limitations when testing.

If you are located close to the telephone central office you may find that hybrid balance changes somewhat from call to call. Or you may discover that there is one best setting for local calls and another for calls into other exchange areas. If you live some distance away from the office you may be able to use one setting of the balancing network for all calls. If this is the case you will want to consider leaving the switches and variable controls out of the assembly. With no need to adjust controls, you might be able to locate the hybrid more advantageously.

A Phone Patch

DOUGLAS A. BLAKESLEE, W1K1K

Although the phone patch shown in the photographs was intended to be used with the popular Collins S/Line equipment directly, the same circuitry will work well with all current amateur equipment. Two such units were constructed, one being used with the author's 75S1/32S1, and the other by WA1HJZ with a B & W 6100 transmitter and Hallicrafters SX101 receiver. Excellent results have been obtained in both installations.

Design

The phone patch is based on a simple resistance bridge, which has been used by Collins in their commercial designs and in some hardware for the space program. This bridge circuit provides isolation between the transmitter and receiver, while connecting both to the telephone lines. An attenuator is provided between the bridge circuit and the telephone line to reduce the output level of the patch and to provide a

fairly constant impedance for the bridge to "look into." Rf filtering is used to prevent trouble with pickup on the telephone line, a problem that shows up all too often in stations running the legal limit. An attenuator is also used on the receiver input so that the receiver gain does not have to be reset when switching from speaker to patch operation.

A heterodyne at about 2600 Hz can produce accidental disconnect on long-distance calls, as the telephone company uses in-band signalling on this frequency for control purposes. The chances are perhaps slight in normal ssb operation that you will have a heterodyne on just the right frequency, but it is possible. The Collins receivers will provide some attenuation of 2600 Hz, as this is down the slope of the receiver's mechanical filter, but the exact amount of attenuation will depend on the slope of the particular filter used and the BFO crystal frequency.

As both of these factors can change with aging, a trap was added to the patch to insure compliance with telephone company tariff requirements. A surplus 88-mH toroid coil is used

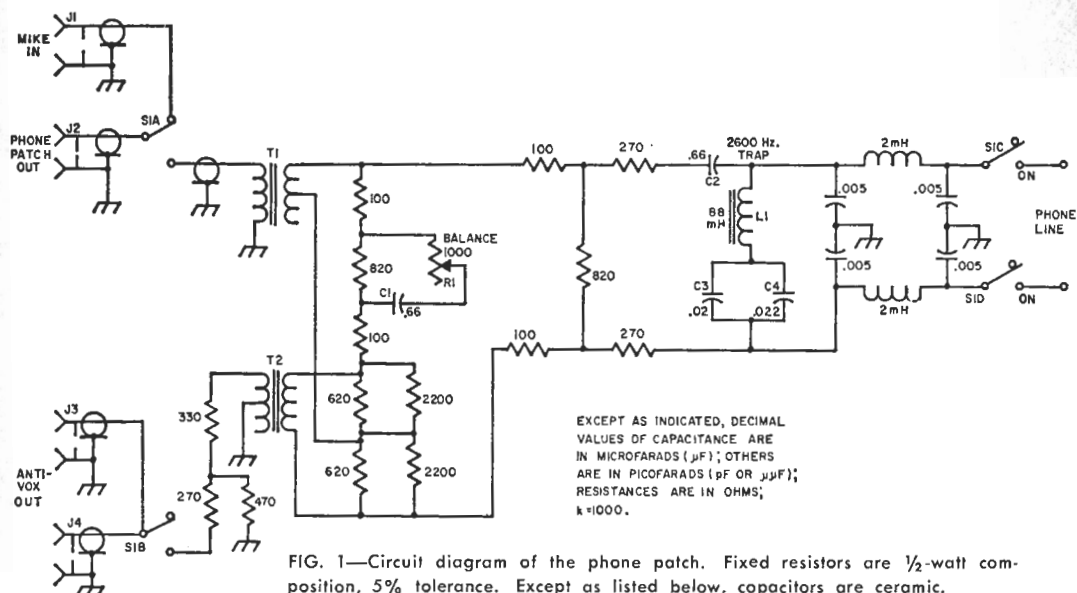


FIG. 1—Circuit diagram of the phone patch. Fixed resistors are 1/2-watt composition, 5% tolerance. Except as listed below, capacitors are ceramic.

C_1, C_2 —200-volt paper, each formed by connecting 0.22 μF units in parallel.

C₂, C₁—Paper; see text.

J₁—J₄ inc.—Phono jacks.

L₁—88 mH toroid, telep

1-800-368-6868, telephone service (see main text).

R₁—Linear control.

S₁—Rotary, 4 poles, 2 positions.

T₁, T₂—Audio, 1500-ohm primary, 500-ohm c.t. secondary (Stancor TA-28).

with two paper capacitors to resonate at about 2600 Hz. If an audio generator and handful of capacitors are available, C_3 and C_4 can be selected to give the best notch at the desired frequency. Normally, the attenuation of the receiver's filter plus the trap is more than necessary, so the adjustment of the trap need not be critical.

Construction

The patch is constructed in an LMB Minibox (W-2C). All small components are mounted on a piece of Vectorbord. A little thought should

be used in the layout so that the circuit "flows" from input to output. A miniature version was tried, but it did not exhibit the excellent isolation of the larger unit, probably because of coupling between the closely packed components. The transformers are mounted with their cores at an angle—when mounted side by side there was some coupling between them. Shielded transformers might be a good idea, but their prices are staggering. The jacks and plugs used can be any of the popular types—use whatever you have as a shack standard.

Adjustment is also simple. Plug the patch

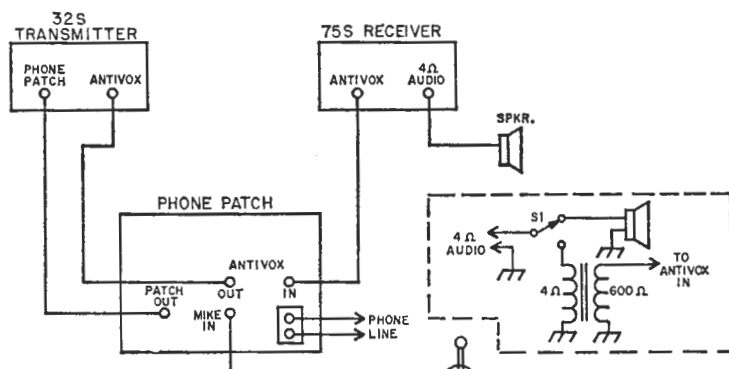
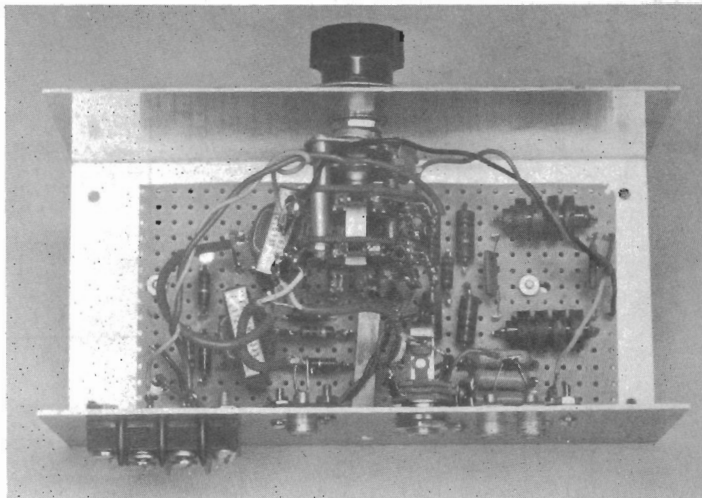


FIG. 2.—Phone patch-S/Line interconnections. With this equipment the load on the receiver's speaker output makes it unnecessary to disconnect the speaker when the patch is switched in. With other receivers it may be desirable to include an extra switch section on S_1 to provide 600-ohm audio (through a transformer) from the receiver while disconnecting the speaker. A suitable circuit is shown in the dotted enclosure; T_1 is a 4-ohm primary, 600-ohm secondary audio transformer.

The patch is assembled on perforated board inside a $3\frac{3}{8} \times 7 \times 3\frac{1}{2}$ -inch box. Jacks and connector block are on the rear wall (bottom in this view). The 2600-Hz trap shown in Fig. 1 was installed after this photograph was taken.



into the telephone company's coupler. Call a friend and ask him to leave his phone off the hook while you make a few adjustments. Connect an audio oscillator to the receiver input, J_3 , and an oscilloscope to the transmitter output jack, J_2 . Turn the patch on and adjust the balance control, R_1 , for the minimum pattern height on the oscilloscope. If the pattern goes to zero, increase the audio oscillator gain and adjust R_1 until you are sure you have the best null. If no test equipment is available, you can make a fair null adjustment by connecting your shack receiver to the receiver input jack, tuning in a loud, steady carrier (or the receiver's calibrator) and connecting a pair of high impedance headphones to the transmitter output jack. Listen on the headphones and adjust control R_1 for minimum signal.

The easiest way to monitor patches is with your own telephone, so leave your phone off the hook while making the null adjustment. Slight improvements may be possible in the null by adjusting C_1 to balance with your particular phone line. This patch has been tried in several locations, but little operational improvement was actually obtained by playing with C_1 .

The actual isolation obtained between transmitter and receiver is about 50 dB, which is about all you can use, as the signal to noise ratio is not much better than this on the average telephone call.

Operation

The operation of a phone patch is affected by many nontechnical factors, some of which are learned the hard way in practical use.

Normally, when requested to make a patch, the operator will place a long-distance station-to-station (or person-to-person) collect call. Unless you live in a downtown metropolitan area, you seldom make local phone patches. The person answering the phone should be briefed on what is happening. The conversation can take place only one way at a time, which is a difficult concept for many people to grasp, so

take it slow. Also, get the person's name for your log—it's your transmitter they are going to use, and FCC regulations require you to log the name of anyone permitted to speak over your microphone.

Flip the patch on, and you're off and running. Monitor the conversations closely—again, it's your transmitter and license. People say the strangest things!

VOX or PTT switching can be another problem area. The unit has sufficient isolation to work well with VOX circuits. However, telephone calls, especially long distance calls, arrive at widely different levels, and almost all phone patches are taking place over considerable distances, on channels that are seldom free of QRM for long. These factors add up to push-to-talk operation by the phone patcher. VOX is tricky at best (and often unusable) on a channel with some QRM, connected into a telephone circuit that often has occasional strange noises. For proper control and best patch operation, we recommend PTT operation.

One final point to remember is that a phone patch is a textbook example of third-party traffic, and thus can only be run to countries allowing such traffic. Keep a list of such countries handy, as requests will occasionally be received from stations in countries with which third-party traffic is prohibited.

Phone patching can be a useful public service and means a lot to the persons involved, as received cards and letters will indicate. The phone company benefits, too, as was shown on a patch from another sea-bound sailor to his wife. The wife answered by saying that she was glad to hear from him again, but unless he had got a raise to stop calling—it was running some \$20 per call on her phone bill!

But don't overdo it. A phone patch from Miami to New York on Sunday has no place on the air, not when you can call direct for a dollar or less. A patch under these circumstances is not a public service; rather, it's a public nuisance.

Appendix

POWER RATINGS OF S.S.B. TRANSMITTERS

Fig. 1 is more or less typical of a few voice-frequency cycles of the modulation envelope of a single-sideband signal. Two amplitude values associated with it are of particular interest. One is the *maximum peak amplitude*, the greatest amplitude reached by the envelope at any time. The other is the *average amplitude*, which is the average of all the amplitude values contained in the envelope over some significant period of time, such as the time of one syllable of speech.

The power contained in the signal at the *maximum peak amplitude* is the basic transmitter rating. It is called the *peak-envelope power*, abbreviated p.e.p. The peak-envelope power of a given transmitter is intimately related to the distortion considered tolerable (see Bruene, "Distortion in Single-Sideband Linear Amplifiers", page 116). The lower the signal-to-distortion ratio the lower the attainable peak-envelope power, as a general rule. For splatter reduction, an S/D ratio of 25 db. is considered a border-line minimum, and higher figures are desirable.

The signal power, S , in the standard definition of S/D ratio is the power in *one* tone of a two-tone test signal. This is 3 db. below the peak-envelope power in the same signal. Manufacturers of amateur s.s.b. equipment usually base their published S/D ratios on p.e.p., thereby getting an S/D ratio that looks 3 db. better than one based on the standard definition. In comparing distortion-product ratings

of different transmitters or amplifiers, first make sure that the ratios have the same base.

Peak vs. Average Power

Envelope peaks occur only sporadically during voice transmission, and have no direct relationship with meter readings. The meters respond to the amplitude (current or voltage) of the signal averaged over several cycles of the modulation envelope. (This is true in practically all cases, even though the transmitter's r.f. output meter may be *calibrated* in watts. Unfortunately, such a calibration means little in voice transmission since the meter can be calibrated in watts only by using a sine-wave signal—which a voice-modulated signal definitely is not.)

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

D. C. Input

FCC regulations require that the transmitter power be rated in terms of the d.c. input to the final stage. Most s.s.b. final amplifiers are operated Class AB₁ or AB₂, so that the plate

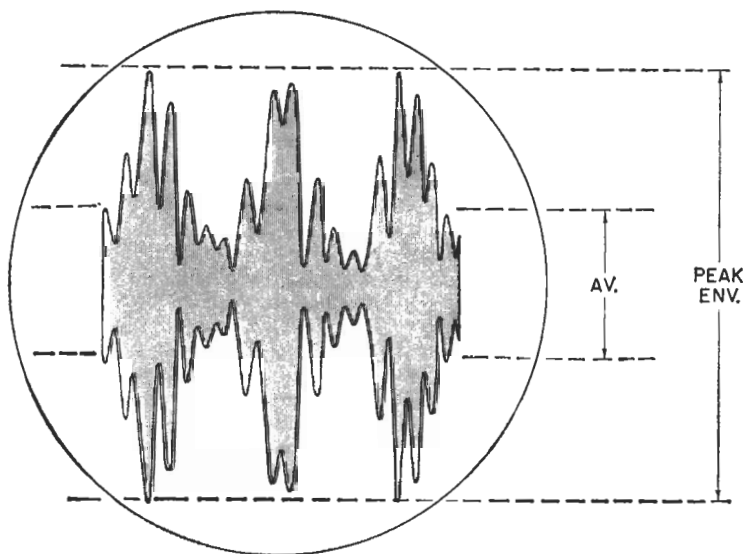


FIG. 1—A typical s.s.b. voice-modulated signal might have an envelope of the general nature shown, where the r.f. amplitude (current or voltage) is plotted as a function of time, which increases to the right horizontally.

current during modulation varies upward from a "resting" or no-signal value that is generally chosen to minimize distortion. There will be a peak-envelope value of plate current that, when multiplied by the d.c. plate voltage, represents the instantaneous tube power input required to produce the peak-envelope output. This is the "peak-envelope d.c. input" or "p.e.p. input". It does not register on any meter in the transmitter. Meters cannot move fast enough to show it—and even if they did, the eye couldn't follow. What the plate meter *does* read is the plate current averaged over several modulation-envelope cycles. This multiplied by the d.c. plate voltage is the number of watts input required to produce the *average* power output described earlier.

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

The 2-kilowatt p.e.p. input rating can be interpreted in this way: The amplifier can

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

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

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

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

FCC REGULATIONS

In the FCC regulations, single sideband is considered the same as A3 (amplitude modulation) and any frequencies allocated for A3 emission may be used for s.s.b. All requirements for station operation and equipment, as outlined in the amateur regulations, must be met. The full text of the current regulations is contained in the latest edition of *The Radio Amateur's License Manual*, published by A.R.R.L.

Amplification of a few points on which questions arise occasionally may be helpful. Numbers in parentheses refer to the applicable section of the FCC Regulations.

Power Input (97.67)—The maximum authorized power is one kilowatt d.c. input, which is interpreted to be the product of the d.c. plate voltage by the d.c. plate current during the largest voice peaks registered on the plate-current meter. The meter must be one having a time constant of not over 0.25 second (practically all panel-type milliammeters and ammeters used in amateur equipment meet this requirement). The one-

kilowatt power limitation is an inclusive one applying to *all* stages supplying power to the antenna; e.g., if the final amplifier is a grounded-grid stage, the sum of the inputs to the driver stage and final stage must be used in determining the power input, since the driver stage contributes some r.f. output power to the antenna.

Station Identification (97.87)—The rules are clear on this, but sometimes are overlooked. In a round table, it is not necessary to transmit the call of every station in the group when signing your own call. In such case it is permissible to say "W9XYZ and the group, this is W9ZYX"; or, if the group is a net having a distinctive name, "W9XYZ and the Butterfly Net, this is W9ZYX." Neither is it necessary to call and sign every time your transmitter goes on, *provided* no transmission is more than three minutes long. It is necessary to transmit the call of the station with whom you are communicating (or the name of the net) followed by your own call (a) if you make a trans-

mission more than ten minutes in length, in which case you must sign at each ten-minute interval, and (b) every ten minutes when communication is by a series of short exchanges. If the station with whom you are communicating, or another station in a group contact, happens to be transmitting when the ten-minute period is up, you must identify yourself at the first opportunity.

DEFINITIONS OF COMMONLY-USED TERMS

Automatic Level Control (a.l.c.)—A system for automatically reducing the gain of a low-level exciter stage when the final amplifier drive approaches the peak-envelope point, thus preventing overdriving and resultant splatter.

Automatic Load Control—Same as automatic level control.

Anti-Trip—A method for preventing sound from the loudspeaker from actuating the VOX system.

Diode Hash—Noise generated in the plate circuit of a tube when its plate is taking current but there is no driving signal. Used particularly with reference to noise generated in a final stage idling during receiving periods.

Envelope Detector—A detector whose output is the envelope of a modulated signal applied to its input circuit. The conventional diode detector is an example.

Modulation Envelope—Outline formed by joining the tips of r.f. current or voltage trace, during modulation, when the r.f. amplitude is plotted graphically against time. It shows the low-frequency (audio frequency, in the case of voice modulation) variations in the amplitude of the r.f. signal.

Peak-Envelope Power—Instantaneous power at the peak of the modulation cycle. See Page 249 for detailed discussion.

Product Detector—A detector whose output is proportional to the product of the amplitudes of two signals applied simultaneously. In the usual case, the amplitude and frequency of one signal are fixed; this signal is generated by the beat-frequency oscillator (b.f.o.) in a superheterodyne receiver. The second signal is the one being received; its amplitude may vary over a wide range but its frequency, in s.s.b. reception, must have a definite relationship to the b.f.o. frequency for proper demodulation.

Resting Current—The plate current of an amplifier in the absence of a driving signal. Also called idling current.

Slicer—A device, usually in the form of an adapter for a receiver, which permits the receiver to respond to one sideband while suppressing the other. As the name is popularly applied, it refers to a device using the

Log Keeping (97.103)—Logs must be kept as described in detail in the regulations. However, in round-table or net communication it is necessary to enter only the calls of stations actually worked, together with the time at which each is first contacted. The time at which you sign out of the round table or net is sufficient for the termination time of all contacts.

phasing system for selectable sideband reception.

Splatter—Frequency components generated by the transmitter which fall outside the *necessary* channel, as a result of distortion in amplifier or frequency-converter stages.

Spurious—Same as splatter.

Two-Tone Test—A method of testing an s.s.b. transmitter for linearity and power output, using two frequencies spaced apart by some moderately-low audio frequency and having the same amplitude.

Unwanted (sideband)—The sideband that is suppressed in a single-sideband transmitter. If the suppression is not complete, there will be output from the transmitter in the unwanted-sideband channel. This is not generally classed as splatter, since it arises from a different cause, but is nevertheless equally undesirable.

VOX—Voice-controlled send-receive switching, generally by means of an electronically-controlled relay, the system being actuated by the presence or absence of sound output from the microphone.

Demodulation—Detection; the process of recovering the original modulating signal from the incoming r.f. signal.

Double Sideband—Commonly used to describe a signal having the two sidebands associated with amplitude modulation, but the carrier suppressed; the output of a balanced modulator.

Linearity—Relative ability of a device to produce, in its output circuit, a true replica of the signal applied to its input circuit. As applied to amplifiers and frequency converters driven by modulated signals, the term linearity means the ability of the device to reproduce the modulation envelope (not necessarily the r.f. waveform) without distortion. In a modulator, linearity means the ability to generate a modulation envelope that reproduces the modulating signal without distortion.

MOX—Manually-operated send-receive switch.

PTT—Manually-operated send-receive switching with push-button control. The push-button usually is part of the microphone assembly.

LINEAR-AMPLIFIER TUBE-OPERATION DATA FOR SINGLE SIDEBAND—GROUNDED-CATHODE CIRCUIT

Tube	Class	Plate Voltage	Screen Voltage	D.C. Grid Voltage ¹	Zero-Sig. D.C. Plate Current	Max. Sig. D.C. Plate Current	Zero-Sig. D.C. Screen Current	Max. Sig. D.C. Screen Current	Peak R.F. Grid Voltage	Max. Sig. D.C. Current ²	Max. Sig. Driving Power ³	Max. Screen Dissipation	Max. Rated Grid Dissipation	Avg. Plate Dissipation	Max. Sig. P.E.P. Output
2E26	AB ₁	500	200	— 25	9	45	—	10	25	0	0	2.5	—	—	15
6146	AB ₁	600	200	— 50	14	115	.5	14	50	0	0	3	—	25	47
6146B	AB ₁	750	195	— 50	12	110	.5	13	50	0	0	3	—	25	60
807	AB ₁	600	300	— 34	18	70	.3	8	34	—	0	3.5	—	25	28
1625	AB ₁	750	300	— 35	15	70	.3	8	35	—	0	3.5	—	30	35
8298	AB ₁	750	200	— 21	20	100	—	20	42	0	0	7	—	40	55
8072	AB ₁	700	250	— 20	100	205	—	16	20	0	0	8	—	—	80
1500	500	— 90	30	83	—	—	—	5	70	—	—	10	—	—	60
2000	500	— 105	20	75	—	—	—	3	80	—	—	10	—	—	85
2500	400	— 85	15	66	—	—	—	3	77	—	—	10	—	—	100
3000	400	— 90	15	60	—	—	—	3	77	—	—	10	—	—	120
811-A	B	1000	—	—	22	175	—	—	93	—	3.8	—	—	65	124
1250	—	—	0	27	175	—	—	—	88	13	3.0	—	—	65	155
1500	—	—	4.5	16	157	—	—	—	85	—	2.2	—	—	65	170
8121	AB ₁	1500	250	— 20	100	210	—	10	20	0	0	8	—	—	170
PL-177A	AB ₁	1500	600	— 110	30	175	0	8	108	0	0	10	—	110	140
PL-177WA	AB ₁	2000	600	— 115	25	175	0	7	112	0	0	10	—	125	210
7094	AB ₁	1500	400	— 65	30	200	—	35	65	0	—	20	—	—	185
2000	400	— 65	30	200	—	—	—	35	60	0	4	20	—	—	250
2500	750 ⁵	— 95	25	145	—	—	—	27	90	0	0	—	—	—	245
813	AB ₂	2250	750 ⁵	— 90	23	158	.8	29	115	—	.1	22	—	100	258
2500	750 ⁵	— 95	18	180	.6	—	—	28	118	—	.2	22	—	125	325
4-125A	AB ₁	2000	615	— 105	40	135 (100) ⁴	—	14 (4.0) ⁴	105	0	0	20	—	—	150
2500	555	— 100	35	120 (85) ⁴	—	—	—	10 (3.0) ⁴	100	0	0	20	—	—	180
3000	510	— 95	30	105 (75) ⁴	—	—	—	8.0 (1.5) ⁴	95	0	0	20	—	—	200
1500	350	— 41	44	200	0	17	0	17	141	9	1.25	20	5	125	175
2000	350	— 45	36	150	0	130	0	3	105	7	.7	20	5	175	175
2500	350	— 43	47	130	0	130	0	3	89	6	.5	20	5	125	200
7034/4X150A	AB ₁	1000	300	— 50	50	225	0	11	50	0	0	12	—	—	115
1800	300	— 50	50	225	0	225	0	11	50	0	0	12	—	—	200
4X250B	AB ₁	1500	350	— 55	100	250	—	8	50	0	0	12	—	—	215
2000	350	— 55	100	250	—	—	—	5	50	0	0	12	—	—	300
7703/4CX250B	AB ₁	1500	350	— 55	83	250	—	30	55	0	0	12	—	—	200
2000	350	— 55	83	250	—	—	—	30	55	0	0	12	—	—	295
8122	AB ₁	2000	400	— 35	100	335	—	10	35	0	0	8	—	—	380
2500	600	— 115	65	230 (170) ⁴	—	—	—	15 (3.5) ⁴	115	0	0	35	—	—	335
3000	600	— 110	55	210 (150) ⁴	—	—	—	12 (2.5) ⁴	110	0	0	35	—	—	400
3500	555	— 105	45	185 (130) ⁴	—	—	—	9.5 (2.0) ⁴	105	0	0	35	—	—	425
4000	510	— 100	40	165 (115) ⁴	—	—	—	7.5 (1.5) ⁴	100	0	0	35	—	—	450
4-250A/5D22	AB ₂	1500	300	— 48	50	243	0	17	96	11	1.1	35	10	150	214
2000	300	— 48	60	255	0	13	—	13	99	12	1.2	35	10	185	325
2500	300	— 51	60	250	0	12	—	12	100	11	1.1	35	10	205	420
3000	300	— 53	63	237	0	17	—	17	99	10	1	35	10	190	520

7580	AB ₁	2000	400	—	77	—	35	77	0	0	12	—	400
7580W/	AB ₁	1500	350	—	62	—	—	62	0	0	12	—	262
4CX250R	AB ₁	2000	400	—	80	—	—	80	0	0	12	—	470
PL-175A	AB ₁	2500	750	—	143	—	35	143	0	0	25	—	265
	AB ₁	3000	750	—	150	—	29	150	0	0	25	—	305
	AB ₁	3500	750	—	160	—	24	160	0	0	25	—	680
PL-8295/172	AB ₁	2000	500 ⁶	—	110	—	43	110	0	0	30	—	1040
PL-8432	AB ₁	2500	500 ⁶	—	115	—	40	115	0	0	30	—	1260
	AB ₁	3000	500 ⁶	—	115	—	39	115	0	0	30	—	1390
4CX1000A	AB ₁	2000	325	—	60	—	35	60	—	0	12	0	1020
	AB ₁	3000	325	—	60	—	35	60	—	0	12	0	1680

¹ Approximate; adjust to give stated zero-signal plate current.
² In Class AB₁ operation there may be a small amount of grid current (1 ma. or less) when the grid is driven to zero voltage. This is caused by contact potential and initial velocity of electrons emitted by the cathode.
³ Driving-power figures do not include circuit losses which must be supplied by the driver.
⁴ Values in parentheses are with two-tone test signal.
⁵ 0 v. suppressor grid.
⁶ +35 v. suppressor grid.

LINEAR-AMPLIFIER TUBE-OPERATION DATA FOR SINGLE SIDEBAND—GROUND-GRID CIRCUIT

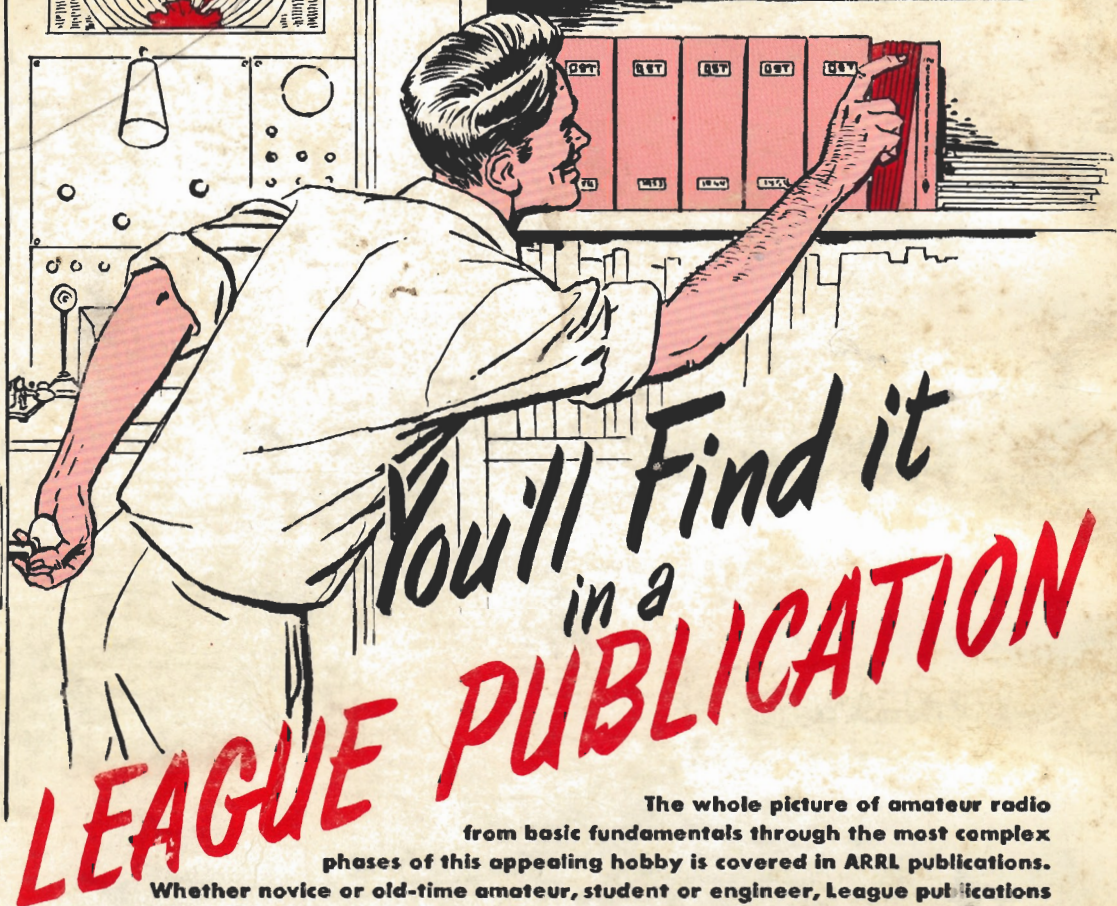
Tube	Plate Voltage	D.C. Grid Voltage	Zero-Sig. D.C. Plate Current	Max.-Sig. D.C. Plate Current	Peak R.F. Grid Voltage	Max.-Sig. D.C. Grid Current	Max.-Sig. D.C. Screen Current	Approx. Input Impedance	Max. Sig. Driving Power	Max.-Sig. Useful Power Output
4-125A ¹	2000	0	10	105	—	55	30	340	16	145
	2500	0	15	110	—	55	30	340	16	190
	3000	0	20	115	—	55	30	340	16	240
4-400A ¹	2000	0	70	265	—	100	55	160	38	325
	2500	0	80	270	—	100	55	150	39	435
	3000	0	90	280	—	100	55	140	40	555
3-500Z	2000	0	95	400(270) ³	—	130(80) ³	—	—	—	500 ¹
	2500	0	130	400(280) ³	—	120(70) ³	—	—	—	600 ²
	3000	0	160	370	—	115	—	—	—	750
PL-6569	2500	—60 ⁶	40	300	180	80	—	300	70 ²	550
	3500	—90 ⁶	30	270	220	68	—	300	75 ²	760
	4000	—105 ⁶	24	250	205	42	—	300	60 ²	800
PL-6580	2500	—50	60	350	195	95	—	300	75 ²	610
	3500	—85	45	300	210	65	—	300	68 ²	765
	4000	—100	40	300	230	65	—	300	72 ²	910
3-1000Z	2500	0	162	800(550) ³	—	254(147) ³	—	—	—	1050 ⁴
	3000	0	240	670	—	300	—	—	65	1360

¹ Grid and screen connected together.
² Includes bias loss, grid dissipation, and feed-through power.
³ Two-tone signal.
⁴ Minimum distortion products.
⁵ Minimum distortion products of 1 k.w. p.e.p. input.
⁶ Approximate; adjust to give stated zero-signal plate current.

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