SINGLE SIDEBAND

for the Radio Amateur

$2.50

A DIGEST OF AUTHORITATIVE ARTICLES ON AMATEUR RADIO SINGLE SIDEBAND

ARRL

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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 h.f. communication services—which up to that time had made almost no use of the system for general multifrequency communication. Today, "s.s.b." has almost supplanted the long-standard amplitude-modulation method in the lower-frequency amateur bands, and is making rapid inroads into v.h.f. communication.

The pages of *QST* have carried the complete story of this transition from a.m. to s.s.b., covering the whole field of amateur s.s.b. 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 s.s.b. 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 s.s.b. 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.

John Huntoon
General Manager, A.R.R.L.

Newington, Conn.

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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?

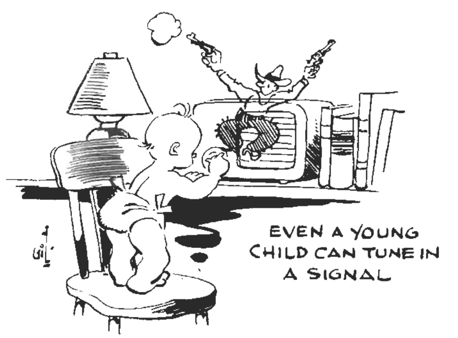
There are several contributing factors. Back in the early 30s there was not as much interest in phone as there is now. The usual receiver was a regenerative one, with or without r.f. stage, and superheterodynes were rather rare. The commercial point-to-point sideband stations used crystal-controlled 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. All radio is complex these days, and we are conditioned to expect it.

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

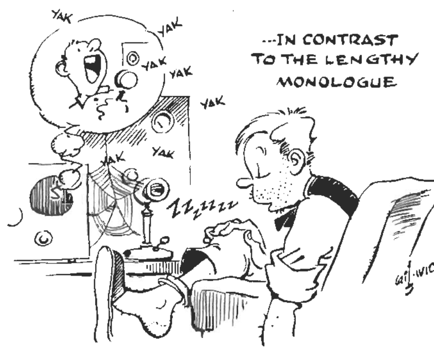


requires a certain amount of skill and understanding to tune in a single-sideband transmission, because a slight error in tuning will turn the signal into "monkey chatter." Also, if the signal is allowed to overload the receiver, it will become distorted and impossible to tune in with any setting of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

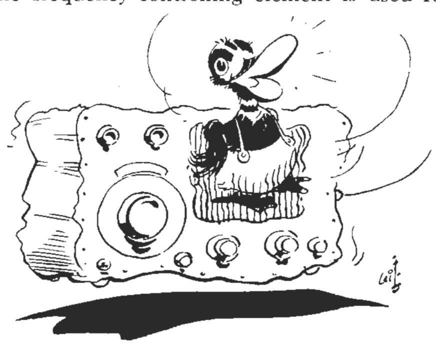
Receivers

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 roundtable. 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.



TO AVOID DISTORTION IN THE RECEIVER, LINEARITY MUST BE MAINTAINED

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 suballocations 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.

» Having a clear picture of a phone signal is the first requirement for understanding what single sideband is all about. If you're a raw newcomer to s.s.b., start here, then skip to "How To Tune in a Single-Sideband Signal" and try it on your own receiver.

How To Visualize a Phone Signal

BYRON GOODMAN, WIDX

The usual description of amplitude-modulated telephony, with its "modulation envelopes" and "percentage of modulation," doesn't prepare you for further understanding. With a background of classical a.m. theory, it becomes practically impossible to form a mental picture of "suppressed carrier," "single sideband," and even plain c.w. In this article we hope to present a picture that will make it easy for you to understand "sideband" techniques.

We will start with the initial statement that to understand phone you must first understand c.w. Practically everyone knows that an unmodulated carrier and a c.w. signal with the key held down are the same thing. Any way you tune them in on a receiver they act the same. On a panoramic receiver they look the same. *Any* test you can make of them at the receiving location will give the same result. *They are the same.* Furthermore, if they are stable they take up *no* room in the spectrum! Oh, sure, you tune in one or the other on your receiver, with the b.f.o. on, and you can hear it over several dial divisions. Turn your b.f.o. off and the S-meter on, and the signal gives a reading over a range of several kilocycles. But neither of these effects proves that the signal is broad—it only indicates that your receiver doesn't have infinite selectivity. By definition, 14,200.000 and 14,200.010 kc. aren't the same frequency, so they must be different. Actually, they differ by 10 cycles, and a receiver or other device that could separate signals 10 cycles apart could separate these two.

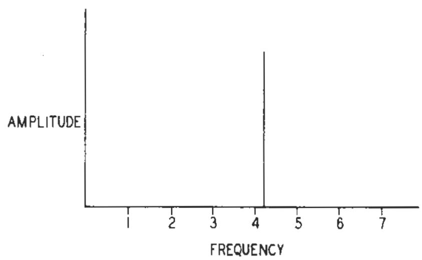


FIG. 1—A representation of a single radio frequency, shown by plotting amplitude against frequency. A steady signal by itself takes up no room in the spectrum.

From *QST*, July, 1950.

All this leads us to the first step in visualizing signals. Any single r.f. signal can be represented by an infinitely-thin vertical line on a plot of amplitude *vs.* frequency. Fig. 1 is such a representation, except that the draftsman couldn't draw an infinitely-thin line that would show on the paper, and we had to settle for a finite-thickness line. The frequency can be read from the "Frequency" scale, and the amplitude from the "Amplitude" scale. The taller the line, the greater the amplitude. Don't worry about the units—the frequency scale could be megacycles, or even cycles at some part of the spectrum. Your panoramic receiver would show such a picture if it had infinite selectivity. If your receiver had infinite selectivity, the S-meter would indicate the amplitude at *one* setting of the tuning knob as you tuned across the frequency range shown, and *nothing* at any other setting.

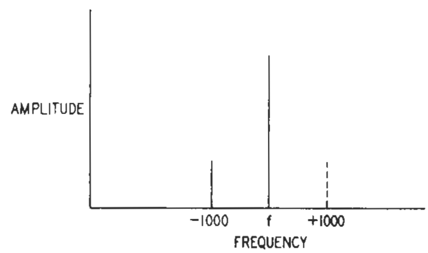


FIG. 2—Two radio signals, separated by 1000 cycles, will give a 1000-cycle audio signal when they are mixed in a detector or other nonlinear circuit.

Two Signals

Suppose now that we wish to transmit some intelligence, and let's say that the intelligence is a simple 1000-cycle tone. One way to do it would be to set up another transmitter on a frequency exactly 1000 cycles removed from the first frequency. It could be higher or lower—it wouldn't matter so long as the separation was exactly 1000 cycles. A practical receiver (one that doesn't have infinite selectivity) would receive both signals simultaneously when tuned to or near the correct frequency, and the audio output of the receiver would be the 1000-cycle beat between the two signals. This is hardly a difficult thing to understand—you don't have to operate long in a phone band before you meet up with "heterodyne QRM," which is exactly the same thing. Such a signal can be represented as in the drawing of Fig. 2.

In Fig. 2, the alternative signal that would also give a 1000-cycle beat is shown as a dotted line. However, we would still be transmitting our 1000-cycle intelligence if we used three transmitters separated as shown in Fig. 3. The signals removed 1000 cycles from the center frequency give 1000-cycle beats in the receiver, and the audio output from the receiver is 1000 cycles, the intelligence we are transmitting.

"Ah, yes," you say. "But what about the 2000-cycle beat between the two outside frequencies? They're separated by 2000 cycles, and you will get a beat between them."

Right you are. Except for one special case, where the proper phase relations exist, this 2000-cycle beat would show up. But the spurious effect is somewhat reduced when the center signal is made large in proportion to the other signals. Thus if we didn't wish to introduce

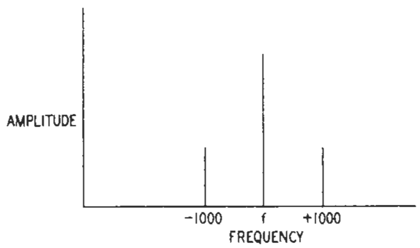


FIG. 3—A representation of two weak radio signals and a stronger signal between them. This also represents the output of a radio transmitter modulated by a 1000-cycle tone.

some extraneous or false intelligence at the receiver, we would have to hold the phase relations exactly right or keep the amplitudes of the outer signals far below the amplitude of the center signal.

Obviously, using three transmitters to transmit this 1000-cycle intelligence is doing things the hard way, and fortunately it isn't necessary. All we have to do at the transmitter, which we will assume is generating a single signal as in Fig. 1, is to beat (or "mix" or "modulate"—they're all the same) this signal with a 1000-cycle signal. As in any beating or mixing or modulating or heterodyning process, the output consists of the original two signals, and two new ones, the sum and difference frequencies. The 1000-cycle audio signal isn't radiated but the others are. The resultant signal is exactly the same as the one we got in Fig. 3 using three separate transmitters! Being the same signal, it gives the same result in a receiver. And, fortunately for the art, the phase relations are right to eliminate the spurious 2000-cycle beat mentioned earlier. Sure, you know that when you mix signals like this you get such a signal—that's what your phone rig does—but you call it "modulation." But when you do the same thing in a receiver, you call it "heterodyning" or "mixing" or "beating." Silly, isn't it? Let's use the word "modulate" from now on, remembering that *the smaller-amplitude signal modulates the larger one,* and that we may run into new products if the signal we are modulating isn't large compared to the modulating signal.

At the start we said you had to understand c.w. to follow this discussion. Let's see why that is so. Suppose, for some strange reasons, that the sole purpose of radio communication was to transmit a 1000-cycle tone. Obviously we could do it in the manners just described, either by setting up three transmitters properly phased or by modulating the output from a single transmitter with 1000 cycles. Sooner or later some bright gentleman would come up with the idea that it isn't necessary to transmit the three signals of Fig. 3. Instead, you could transmit a single signal as in Fig. 1 and incorporate a to-be-modulated signal in the receiver. Duty-bound to receive only 1000-cycle intelligence, we could set up this to-be-modulated signal 1000 cycles higher or lower than the transmitted signal. Every time the transmitter was turned on, we would get the 1000-cycle tone, and in every respect we would have the same communicating ability that we had when the signal of Fig. 3 was working into a receiver where there was no to-be-modulated signal. That is exactly what we do in c.w. communications circuits, except that the receiving operator selects the tone and we complicate the matter by superimposing further intelligence in the form of a code made up of short and long signals and spaces.

The greater the amplitude of the incoming signal the more it modulates the local signal (beat oscillator) and the louder the audio output becomes. If we are to avoid beats between two or more *different* signals present in the receiver passband, the local signal (beat oscillator) must have a much greater amplitude than the incoming signals, just as in the 3-signal case described earlier.

Carriers and Sidebands

Now let's tie in these concepts to the sideband bugaboo. The big husky signal that all the other signals modulate has been—and still is—called the "carrier." As you can see now, it isn't a carrier at all, because it doesn't carry anything. In a c.w. receiver you call it the "beat oscillator," even though it does exactly the same thing as a transmitted carrier and might well be called a "local carrier."

The carrier by itself conveys no intelligence. The intelligence is contained in the smaller signals and is obtained from their modulating action on the carrier. These smaller signals are called "side frequencies," and a band of them is called a "sideband," reasonable names that have no confusing aliases or synonyms.

In a communications system based on the modulation of a large signal by a smaller one (a.m. or c.w.), the amplitude of the audio output from the receiver is proportional to the amplitude of the side frequencies. The frequency of the output is determined by the frequency difference between the carrier and the side frequencies. The carrier conveys no intelligence, so it doesn't have to be transmitted and might very well be supplied at the receiver. What could be simpler?

Complex Modulation

It should be obvious that we don't have to confine ourselves to 1000-cycle tones. The modulating signal might well be a complex signal, made up of different frequencies, without modifying the basic concept one iota. For example, if our purpose were to transmit simultaneously a 2500-cycle tone and a 1000-cycle tone of greater amplitude, we could set up five transmitters as shown in Fig. 4, with careful control of the relative phases so as not to have some 1500-, 2000-, 3500- and 5000-cycle signals in the receiver output. Or we could modulate the carrier with the 1000- and 2500-cycle signals and get exactly the same thing. The

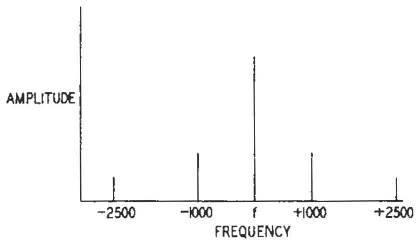


FIG. 4—A representation of five radio signals or a single transmitter modulated by 1000- and 2500-cycle tones. The 1000-cycle tone has almost twice the amplitude of the 2500-cycle tone.

effect at the receiver would be the same. Speech and music are more complex than just two tones, but the principle is identical. The complete a.m. signal consists of the steady carrier and the two sidebands. The individual side frequencies in the sidebands are determined by the individual components that exist in the audio modulating signal at the instant under consideration. In an a.m. transmitter the audio frequencies modulate (you could say "are beat against" or "are mixed with") the carrier and generate corresponding side frequencies through what you call "modulators" and "modulated amplifiers." You would be just as correct, if not more so, to call your modulator a "power amplifier" and your modulated amplifier a "mixer" (as you would in a receiver).

Tricks with Sidebands

Since the carrier conveys no intelligence, it should be possible to dispense with it and introduce it at the receiver. This will save transmitter power and reduce heterodyne QRM. Unfortunately, if both sidebands are received at the detector where the carrier is introduced, the carrier has to have exactly the correct phase relationship with the sidebands if distortion is to be avoided. Since exact phase relationship precludes even the slightest frequency error, such a system is workable only with very complicated receiving techniques. However, if only one sideband is present at the detector, there is no need for an exact phase relationship and there can be some frequency error without destroying intelligibility. The extra sideband can be removed either at the transmitter or at the receiver—one is *single-sideband transmission* and the other is *single-sideband reception*. Thus we could get rid of heterodyne QRM in our bands if no one transmitted the carrier but only one or two sidebands, but the double-sideband signals would require single-sideband receivers at the receiving end.

When the carrier is eliminated at the transmitter and reinserted at the receiver, its frequency must be set rather carefully. For example, if it is set 100 cycles off, there will be an error of 100 cycles in all of the received audio signals. This is of no importance in radiotelegraphy, but in voice work manual receiver tuning for a single-sideband suppressed-carrier signal is somewhat critical. (There are electronic means for simplifying this tuning, provided a weak carrier is transmitted to give a clue to the exact setting of the carrier, but they are not used in amateur communication).

The minimum possible bandwidth of a modulated signal is the bandwidth of one sideband. Ordinary a.m. signals use at least twice this bandwidth because both sidebands are transmitted. Claims that some methods of amplitude modulation result in narrower signals than others are obviously ridiculous—any normal system resulting in double sidebands will give the bandwidth of any other, provided, of course, that the same modulating signals are used and that both systems are in proper adjustment. Out of adjustment, they can only result in still greater bandwidths.

That's about the whole sugar-coated story. Think of modulation, beats, heterodyning and mixing as exactly the same thing, and forget about carriers transporting audio and all of the other misconceptions, and you will be able to understand any new techniques thrown at you. Visualize the audio signal modulating the carrier to generate sidebands, and (at the receiver) the sidebands modulating the carrier to produce the audio signal, and it should all begin to make sense.

» 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, ex-W2KUJ

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

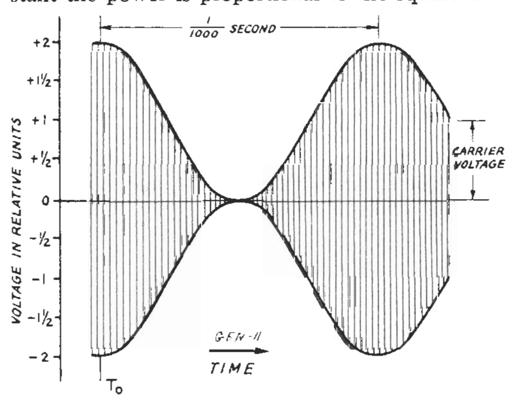


FIG. 2—Envelope of carrier 100% modulated by a 1000-cycle sine wave.

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.

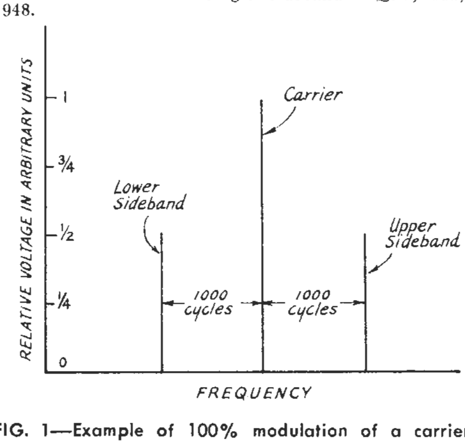


FIG. 1—Example of 100% modulation of a carrier by a single tone of 1000 cycles per second.

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!

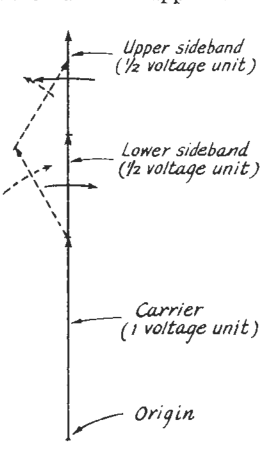


FIG. 3—Vector diagram of 100% modulation of an a.m. carrier at the instant (corresponding to T_0 in Fig. 2) when peak conditions exist. The broken vectors show the relationships at an instant when the modulating signal is somewhat below its peak.

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 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 tht 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

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

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 modulation 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 ½ 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.

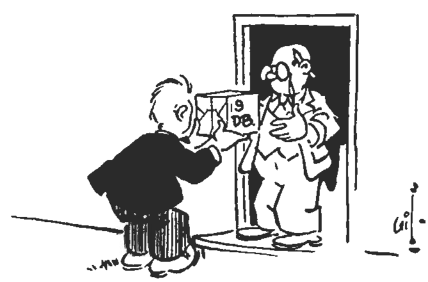
It doesn't take much figuring to see that if it requires twice as much single-sideband power as it does double-sideband power, to get the same signal output power from a receiver with the noise power output half as much for single-sideband operation as for double-sideband operation, nothing has been 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 eleminate 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 transmission. 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.

» The elements of single-sideband reception are simple, but often confusing to those whose only previous experience has been with the "BCL" type of reception used for a.m. transmissions. Read "How To Visualize a Phone Signal" in conjunction with this article.

How To Tune In a Single-Sideband Signal

BYRON GOODMAN, WIDX

Receiving an s.s.b. signal properly is a lot easier to do if you have a mental picture of what's going on. Let's assume that an instantaneous picture of a 25-kc. section of the 75-meter subband looks like Fig. 1. Your receiver can be considered a sort of "peephole" that you slide back and forth across the band. If you were going to build a working model of this receiver-operation picture, you would cut out a long strip of cardboard, as shown in Fig. 2A, and notch it as shown. The width and shape of the notch varies somewhat with the type of receiver—the more selectivity you have, the narrower this notch would be. Your working model would consist of this cardboard strip laid on Fig. 1. Turning the tuning knob of the receiver corresponds to sliding this strip back and forth across the band. When the tuning scale on your receiver indicates "3903," it corresponds to the notch being centered on 3903 kc. on Fig. 1, and if the notch is, say, 8 kc. wide all you could see (and hear) would be "Signal A" and a bit of "Signal B" that might show through. With the notch centered on 3911 kc. you would see (and hear) only "Signal B," and with the receiver (cardboard scale) centered anywhere from 3918 to 3921 kc. you could see all of Signal C.

After you have moved the notched cardboard mentally across the band a few times, you're ready for the next step. Forgetting the band for a minute, visualize the notched cardboard with a small piece of celluloid mounted on it. This celluloid has a single vertical line scribed on it, representing the beat-oscillator frequency. A working model would look like Fig. 2B. Assembled on your receiver model, it would look like Fig. 2C. Your b.f.o. adjustment on your

From "Tuning and Checking S.S.B. Signals," *QST* October, 1950.

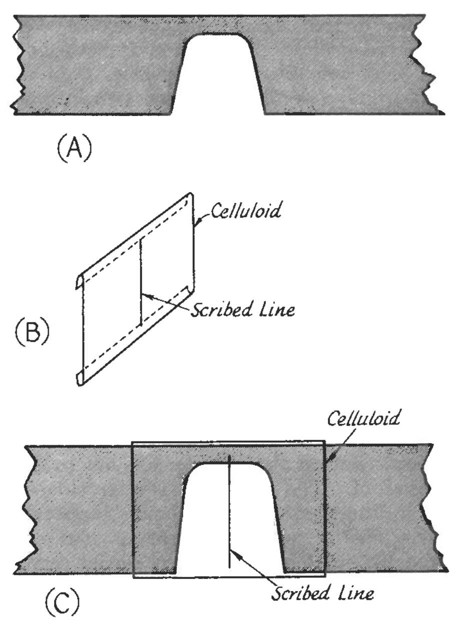


FIG. 2—Parts required for model receiver to be used with Fig. 1. The strip at A could be a piece of cardboard, notched as shown to represent the "passband." A celluloid slider with a line scribed on it, to represent the b.f.o. frequency, would represent the b.f.o., as shown at B. With b.f.o. on, the celluloid slider would be clipped on the cardboard strip, as shown in C. Thus A represents a receiver with b.f.o. off, and C with b.f.o. on.

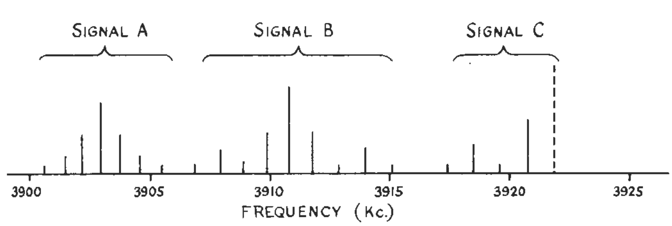


FIG. 1—An instantaneous frequency-vs.-amplitude representation of a portion of an amateur band. Signals A and B are a.m. signals, and C is a single-sideband suppressed-carrier signal. Signal C is using the lower sideband, and the suppressed carrier is represented by the dashed line.

receiver is the same as moving this celluloid clip with respect to the notch on the cardboard, but your tuning knob moves the cardboard strip *and* the celludoid together. This simply means that the relationship between b.f.o frequency (line on celluloid) and the receiver passband (notch in cardboard) is constant with receiver tuning.

Your receiver with b.f.o. off looks like Fig. 2A. It looks like Fig. 2C when the b.f.o. is on.

Now you're ready to tune in that s.s.b. station, represented by Signal C in Fig. 1. With b.f.o. off, tune your receiver until Signal C is centered in the passband. As mentioned before, any setting between 3918 and 3921 would allow him to come through, and he would be centered at 3919.5 kc. You can do this with the a.g.c. on, telling when you have him centered by the point where he kicks the S-meter the highest, or you can do it with the a.g.c. off and with the r.f. gain backed down, in which case you tune aurally for maximum sound on peaks. In any event, center him and then turn the r.f. gain down, a.g.c. off, the audio gain up, and then turn on the b.f.o. Vary the b.f.o. frequency slowly back and forth until the speech becomes recognizable and you can copy the voice. This corresponds to sliding the celluloid scale back and forth until you have the scribed line exactly or very close to superimposed on the dashed line, as shown in Fig. 3 (The dashed line represents the suppressed carrier.)[1]

It should now be obvious that if the b.f.o. were originally set on the proper side of the passband, you could have done the tuning with the main dial alone, and this is generally a little easier to do, particularly with receivers with slow tuning rates. On some communication receivers, however, the b.f.o. tuning rate is slower than the main-dial rate, and that is why we described it this way. Since some s.s.b. stations use the upper sideband and some the lower, it is also apparent that setting the b.f.o. with respect to the passband for one s.s.b. signal is not necessarily correct for another, but it will be right for all s.s.b. signals using that same (upper or lower) sideband.

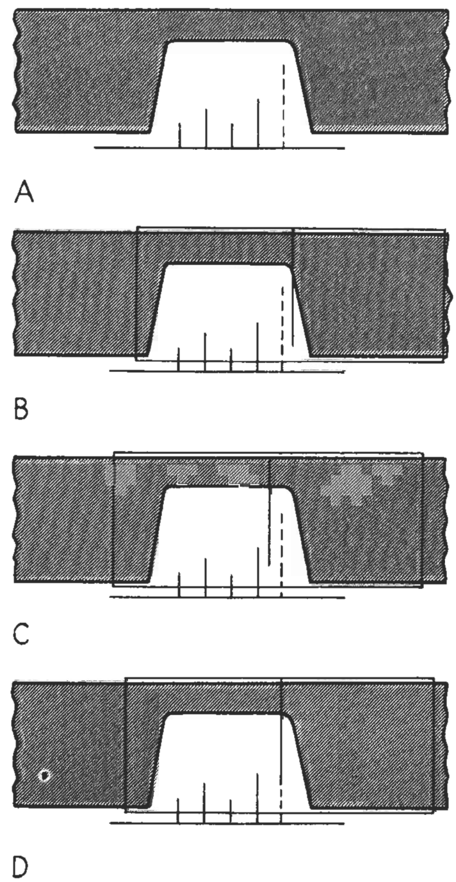


FIG. 3—(A) First step is to center the s.s.b. signal in the pass band. The b.f.o. is off.

(B) The b.f.o. is turned on but the frequency is wrong. The s.s.b. signal will sound garbled, high-pitched, feminine.

(C) The b.f.o. is wrong in the other direction. Signal will sound garbled, low-pitched, inverted.

(D) The b.f.o. coincides with suppressed-carrier frequency. Signal will be clean if receiver isn't overloading (and if the signal was good to start with).

[1] The tuning instructions given here apply to "conventional" receivers not especially equipped for s.s.b. reception. In these, the a.g.c. system is generally not usable with the b.f.o. turned on, since the b.f.o. voltage actuates the a.g.c. and reduces the receiver gain to an undesirably low value. Modern receivers using product detectors usually have a.g.c. systems that can be used with the b.f.o. on; often, the b.f.o. frequency is not adjustable but is switch-controlled for upper- and lower-sideband reception. In these, too, it may not be possible to turn off the a.g.c. from the front panel. Nevertheless, the principles of tuning are the same as described.

» 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, W2KUJ

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 "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.

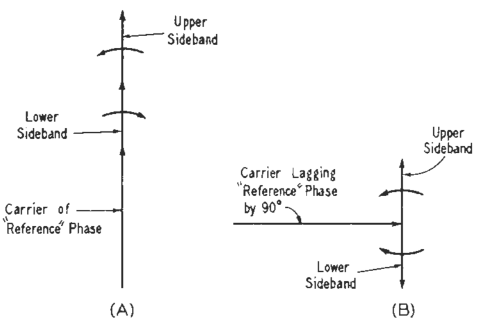


FIG. 1—The carrier and sideband relationship required to generate a single-sideband signal by the "phasing" or "balancing" method. The modulating signal in B leads the modulating signal in A by 90°. When the two signals represented by A and B are combined, the upper sidebands add and the lower side bands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

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 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.

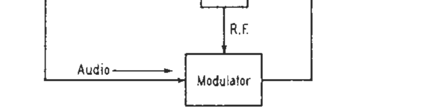


FIG. 2—A block diagram showing the circuits required to generate a single sideband by the method of Fig. 1. This is an impractical method because there is no known means for obtaining the 90° audio shift over a wide range of frequencies.

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

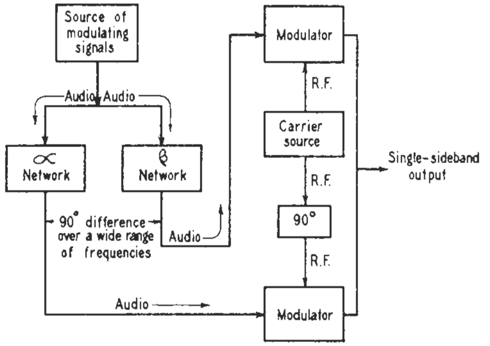


FIG. 3—The system outlined in Fig. 2 becomes practical by using two audio channels (α and β networks) with a constant phase *difference* of 90°.

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 ± 2° 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

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

and for $\delta = 2°$,

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

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 a 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,

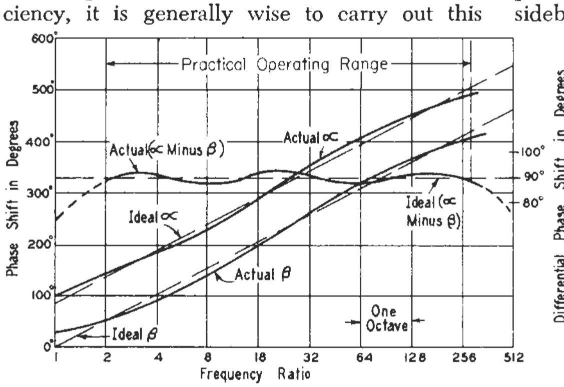


FIG. 4—This plot shows how the 90° difference between the α and β networks is maintained over a wide frequency range. The scale for the phase difference is given on the right-hand side of the graph.

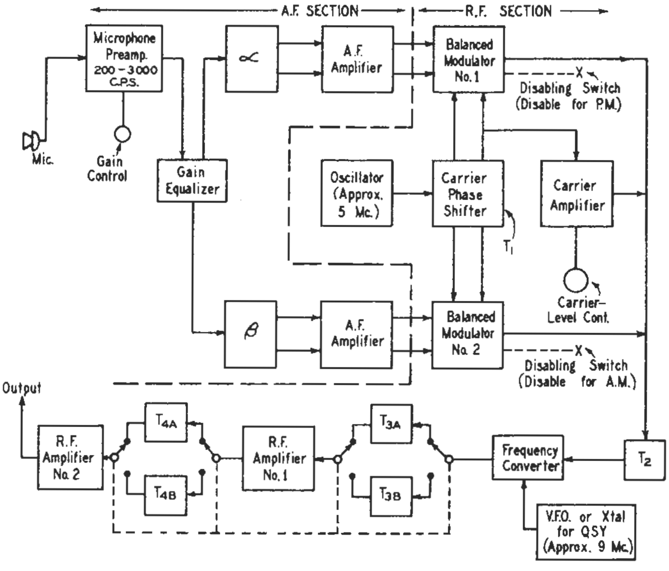


FIG. 5—Block diagram of an exciter capable of generating s.s.b., a.m. and p.m. signals on either the 75- or 20-meter bands. Table I gives a description of the various components.

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 AB_1). 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-sizable 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-frequency 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 intelligibilty. 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

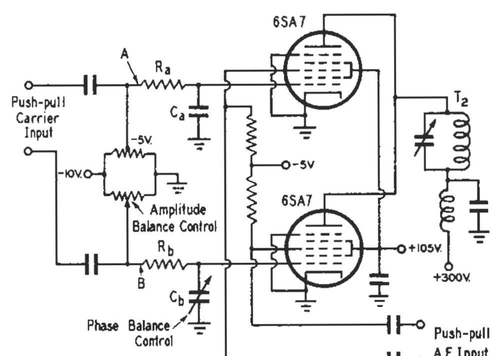


FIG. 6—A typical balanced modulator, using 6SA7 tubes. Provision is included for obtaining amplitude and phase balance of the r.f. (carrier) input.

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 signal, 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 $\pm 10°$ —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	6SJ7 tube.
T_2	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	6SA7 converter tube.
R.F. amplifier No. 1	6AK6 beam tube. Operates as Class A amplifier.
R. F. amplifier No. 2	807 beom-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.

» Outside their application in balanced modulators, diode modulators are not usually encountered in amateur work. Here is a simple explanation of how the diode modulator works, combined with an explanation of some basic electrical principles.

Diode Modulators

BYRON GOODMAN, W1DX

Before single sideband, amateurs had little or no contact with diodes used as modulators. While they had been used for years as *de*-modulators—"detectors" is the common word—there was never any reason to consider their use in the allied function of modulator. Their use as modulators is old hat to the commercials, however, particularly in the field of carrier telephony.

But before a discussion of diodes, let's review some of our basic concepts and terminology, because it will help us to understand a few things later on. You are familiar with the plot of an alternating current or voltage with respect to time. This is shown in Fig. 1A, where the time is represented along the horizontal axis and the amplitude is shown on the vertical. An alternating current or voltage of a single frequency is called a "sine" (or "cosine") wave, from the trigonometric function that defines the instantaneous values. It is symmetrical about the zero-amplitude axis, the positive peaks extending as far above as the negative peaks do below. Along the time axis, the distance between similar parts of the wave is a time equal to $1/f$, where f is the frequency. If the wave in Fig. 1A is to represent a 1000-cycle wave, $1/f$ is 0.001 second, but if it were a 100-kc. wave, $1/f$ is 0.00001 second. Drawn to the same scale, the 1000-cycle and 100 kc. waves might look as in Fig. 1B. But remember that the *shape* is always the same, and that only the scale changes. It's something like those trick mirrors in a penny arcade—they change the scale in one or the other dimension.

One very important thing to remember from the preceding paragraph is that a single-frequency a.c. wave is always symmetrical about the zero axis. If it isn't symmetrical, it isn't a single-frequency affair. Take, for example, the waveform shown in Fig. 1C. At first glance it looks exactly the same as that in Fig. 1A, with the zero-amplitude axis displaced. (That's just what it is.) But it no longer represents a pure a.c. wave, because it doesn't satisfy our definition of being symmetrical about the zero-amplitude axis. Instead, it is now a representation of the a.c. wave of Fig. 1A plus a d.c. (zero-frequency) component. It is obtained by adding the a.c. wave to a steady d.c. value, as shown. The polarity never goes negative, in contrast to the pure a.c. wave where the polarity is negative half the time. (Of course, the d.c. component could be negative, in which case the polarity would never go positive; or the d.c. component could be less than the peak value of the a.c., in which case the wave would fall on both sides of the zero-amplitude axis, but not symmetrically.)

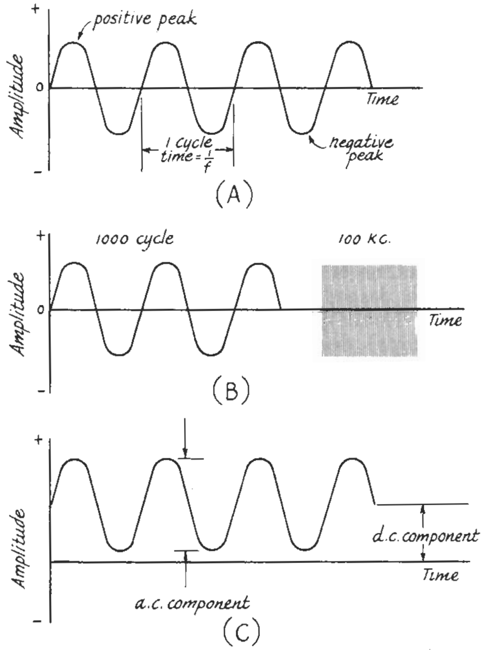


FIG. 1—The old sine wave, familiar to one and all, is shown at (A). It is a plot of amplitude *vs.* time of a single-frequency a.c. wave.

Two different frequencies drawn to the same time-base scale look entirely different, because the higher-frequency cycles are necessarily crowded (B). The shape is the same, however—only the scale is different.

A pure single-frequency a.c. wave must swing equally above and below the axis—if it doesn't, it has a "d.c. component" (C).

This a.c. wave with a d.c. component is easy to come by, and exists in many places throughout radio equipment. The current in an audio amplifier is of this type, where the d.c. component is the steady value of plate current and the a.c. component is the audio signal. But there is one more thing we should know—and remember—about it. If the d.c.-

From *QST,* April, 1953.

plus-a.c. signal is coupled to anything, like a load or another stage, through a capacitor or a transformer, only the a.c. component appears at the load. This should be obvious, of course—the capacitor or transformer cannot pass the d.c., and anything passing through the capacitor or transformer must swing equally about the zero-amplitude axis. Thus the signal of Fig. 1C passing through a capacitor or transformer—or "a.c. coupler"—will appear as Fig. 1A.

Envelopes

Before we settle down to the main business at hand, there is one more aspect of a.c. that we should review. The signals in Fig. 1 were drawn for only a few cycles, for convenience and ease of studying, but we should worry a little about how they start and stop. Suppose we examine a 100-kc. signal that builds up slowly (instead of instantaneously as in Fig. 1B) and then decays slowly. It might look as in Fig. 2A. The first few (and the last few) cycles do not have the same peak-to-peak amplitude that the main bulk of the cycles do. The outline of the 100-kc. wave is represented by the dashed line and is called the "envelope." Notice particularly that this dashed line (envelope) does not represent the instantaneous value of the wave, but only the limits of its peak-to-peak excursions. It is, however, symmetrical about the axis, and must always be so if no d.c. component is present.

Fig. 2B should be a familiar picture. It represents the 100-kc. signal we have been using "modulated" by our 1000-cycle signal. Actually, the only a.c. signal drawn here is the 100-kc. "carrier," although we immediately recognize that the envelope has the form of our 1000-cycle signal. The amplitudes of the 100-kc. cycles are changing from time to time. Notice also that, looking at the *half* r.f. cycles above the zero-amplitude axis, the outline bears a strong resemblance to Fig. 1C, except that in Fig. 2B the envelope replaces the signal, and the (half) carrier amplitude replaces the d.c. component. The same picture, flopped over, appears below the zero-amplitude axis, and the envelope is symmetrical about this axis, as it was in Fig. 2A. Remember that the only a.c. existing here has a frequency of 100 kc. (and some 99- and 101-kc. side frequencies that we won't discuss right now), and that there is no 1000-cycle component that we could find with a wave analyzer.

But consider the signal in Fig. 2C. Here a 1000-cycle signal and a 100-kc. signal exist in the same circuit. It is no longer symmetrical about the zero-amplitude axis. Instead, one signal is "super-imposed" on the other, and a wave analyzer or tuned circuit could select one or the other quite easily. This is the basic difference between this "superimposed" wave and the "modulated" wave of Fig. 2B. In the superimposed waves, the peak-to-peak amplitude of each 100-kc. cycle is the same as that of the previous cycle, even though the excursion above and below the zero-amplitude axis is not always the same. And the envelope is not symmetrical about the zero-amplitude axis—it is as though the 1000-cycle signal had become the axis (dashed line).

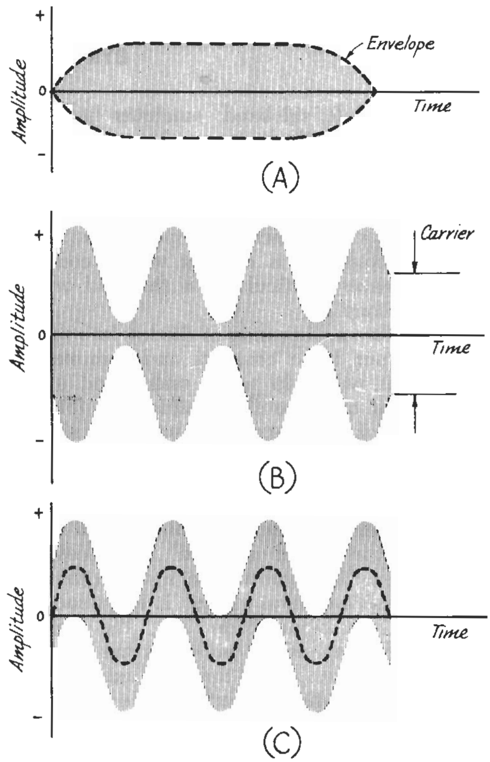


FIG. 2—High-frequency waves don't start and stop instantaneously, and the outline of their rise and fall is called the "envelope" (A). Each cycle swings equally above and below the axis, however.

The familiar envelope of a "modulated" wave is shown at (B), with the less-familiar pattern of "superimposed" waves at (C).

Now that you can recognize the difference between superimposed signals and modulated signals, and know the effects of a.c. couplings, we are ready to talk about the mechanics of modulation in a diode.

Modulation

If we feed the superimposed signals of Fig. 2C into a resistor (or into a good Class A or Class B amplifier of such bandwidth as to pass 1000 cycles and 100 kc.), they will come out looking exactly the same as they did at the input. But suppose we use the circuit of Fig. 3A, and feed them into a diode? The action can be analyzed by plotting the effect in the diode, as in Fig. 3B. Whenever the 100-kc. applied voltage swings to the right (is positive), the diode conducts and a half cycle of r.f. passes through R_1. Plotted against time,

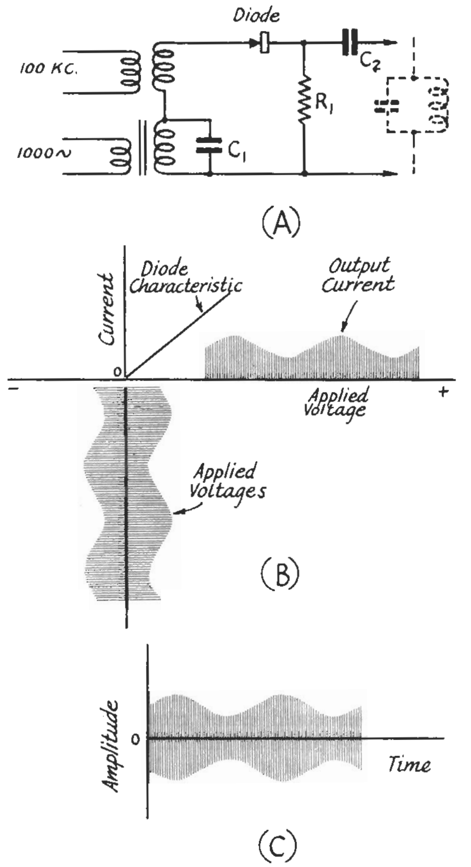


FIG. 3—A basic diode-modulator circuit is shown in (A). C_1 and C_2 are bypasses for the 100-kc. signal. The modulator action is shown at (B), where the envelope of the superimposed signals becomes a modulated envelope in the output. The a.c. coupling in the output of the modulator, and the tuned circuit, convert the "output current" envelope of (B) to the modulated-wave envelope of (C).

they would appear as the "output current" shown to the right of the diode characteristic. When the voltage swings negative, the diode will not conduct and no output current appears.

So far we have only half cycles of 100-kc. r.f., all swinging up from zero to an amplitude determined by the 1000-cycle signal that was superimposed on the original signal. You know that half cycles of any frequency contain *harmonics* of that frequency, so we can expect that the current through R_1 is made up of a 1000-cycle component, a 100-kc. component, and some harmonics of 100 kc. (There are also those side frequencies we mentioned earlier, but they are close to 100 kc. and its harmonics, and we will again ignore them in this discussion.) If now we connect a parallel circuit tuned to 100 kc. on the other side of C_2 (as shown by the dotted lines), only the 100-kc. energy will appear across it, the other components being rejected by the selectivity of the circuit. The voltage across this tuned circuit will appear as in Fig. 3C, since the a.c. coupling (through C_2) has made it necessary that each 100-kc. cycle swing as much below the axis as above. This figure we recognize as a modulated wave.

The diode characteristic shown in Fig. 3B is much too good to be true, and in practice it isn't a straight line from zero on up. A practical characteristic has some curvature, and so the usual practice in diode modulators is to use a large r.f. signal and a small audio signal. This has the effect of doing the actual work of modulating on a small relatively-straight portion of the diode characteristic, and means that you can't use a high percentage of modulation without runnning into distortion of the envelope. In the applications where diode modulators are used, we try to hold the distortion down as low as possible.

Balanced Modulators

A balanced modulator is a device for obtaining the side-frequency components of modulation without passing the carrier. In single-sideband transmitters this is done prior to removing one of the sidebands with highly-selective circuits. While balanced modulators may take several different forms, they all serve the same basic purpose, and the various circuits involving diodes differ only in the frequency components (harmonics) that appear in the output.

Two of the common circuits are those shown in Fig. 4. It is apparent in both that the carrier frequency cannot appear in the output

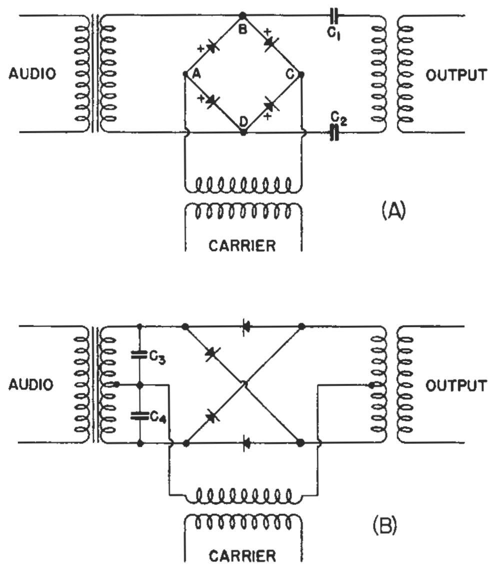


FIG. 4—Two common diode balanced-modulator circuits are (A) the bridge and (B) the ring. C_1, C_2, C_3 and C_4 are r.f. by-pass capacitors, used to complete r.f. paths without short-circuiting the audio.

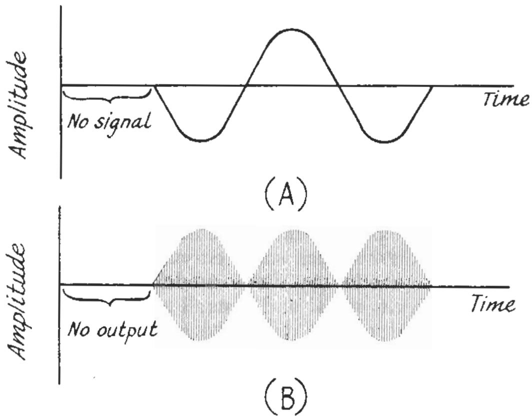


FIG. 5—A modulating signal as in (A) gives an r.f. output from a balanced modulator as in (B).

because the net effect of the carrier across the output is zero, when there is no audio signal.

Now suppose that we disconnect the audio transformer and connect a small battery across points *B* and *D* in Fig. 4A, the positive terminal to *B*. Diodes *AB* and *CD* will be "biased back" by the amount of the battery voltage, and they will not conduct r.f. (of the proper polarity) until the r.f. voltage exceeds this bias value. The other two diodes, *BC* and *AD*, will conduct readily, however, and over more than half the r.f. cycle, because they are biased "forward." Since the one set of diodes is conducting better than the other, the circuit is no longer balanced, and r.f. will appear across the output. The fact that these are approximately half cycles of r.f. flowing through the diodes shouldn't bother you—remember that this is an a.c.-coupled affair and the r.f. will be normal full cycles in the output. The more voltage applied, the more the unbalance, and the more r.f. there is in the output. When the polarity of the bias is reversed, the diodes *BC* and *AD* will be biased "back," and diodes *AB* and *CD* will be the easier paths.

Since the output depends upon the voltage across points *B* and *D*, if we reconnect our audio transformer and apply a single audio frequency, the r.f. output will appear in proportion to the audio voltage and regardless of its instantaneous polarity. Thus we will obtain an output like that of Fig. 5B when an audio voltage like that of Fig. 5A is applied. This pattern is that of the "two-tone" test signal, but it should be apparent that it is the envelope pattern of a balanced modulator when a single modulating frequency is used. It will also occur to the reader that the balanced-modulator action could have been described simply on the basis of a balanced bridge being upset by the action of the audio, without any introduction explaining something about normal modulators and a.c.

Except that this isn't the complete story. One thing these envelope patterns can't show is the resultant frequency "spectrum" of the modulated wave. For example, the frequency spectrum of the envelope shown in Fig. 5B, when generated in a balanced modulator, consists of two side frequencies, separated from the (eliminated) carrier by the modulation frequency. In the case we have been speaking about, the spectrum of this signal would show two side frequencies, 99 and 101 kc., with no energy at the (eliminated) carrier frequency of 100 kc. Such an *envelope* pattern can be generated in a normal modulator, by modulating with a complex wave that could be obtained from a full-wave rectifier and adjusting the modulation percentage to exactly 100. In this case, however, the spectrum would consist of the carrier at 100 kc. and side-frequency components spaced at 1000-cycle intervals out to 10 or 15 kc. either side of 100 kc. Hence, although the envelopes could look the same, the spectrums could differ greatly—the difference is in the phase of the r.f. cycles and the lack or presence of a carrier. In the balanced modulator, the phase of the r.f. in the output is reversed as the modulating signal passes through zero, because one pair of diodes takes over the job from the other and routes the r.f. differently from its source to the output.

Practical Considerations

It has already been mentioned that the ratio of modulating voltage to carrier voltage should be low in a diode modulator if the distortion products are to be held to a low value, and this is equally true in the balanced-modulator application. Normal practice is to make the carrier voltage at least 10 to 20 times the peak modulating voltage. For germanium crystals the r.f. voltage is usually on the order of 2 to 6 volts. The inherent carrier balance will sometimes run as high as 30 db. without any balancing adjustments, and with balancing (through circuits shown in any practical description) it will run to 60 or 70 db. Sideband energy is equal to the modulator power delivered, minus the resistance losses in the diodes, and these losses will run from 2 to 10 db. depending upon the carrier frequency.

» *An improved type of balanced modulator was the design objective of the beam deflection tube described here. This article tells how the 7360 can be put to work as balanced modulator or balanced mixer.*

S.S.B. Circuits Using the 7360 Beam-Deflection Tube

H. C. VANCE, K2FF

The 7360 was originally developed to provide a high degree of stable carrier suppression when used as a balanced modulator in single-sideband service. More than 60 db. of carrier suppression has been obtained with it as a balanced modulator in s.s.b. exciters of both the filter and phasing types. It is of course equally valuable in double-sideband suppressed carrier service.

In this article circuits will be described which make use of the 7360 as a balanced modulator and a frequency mixer. Balanced modulator circuits will be shown for both the filter and phasing methods of s.s.b. generation.

Fig. 1 is a cross-section sketch of the main elements of the tube. The single flat cathode, control grid and screen grid form an electron gun which generates, controls and accelerates a ribbon or sheet beam of electrons. The screen grid and the two deflecting electrodes act as a converging electron lens to focus this beam.

Varying the bias or signal voltage on the control grid varies the plate current as in a conventional tube. The *total* plate current to the two plates, at a given plate voltage, is determined by the voltages applied to the control grid and the screen grid. The *division* of the total plate current between the two plates is determined by the differences in voltage between the two deflecting electrodes.

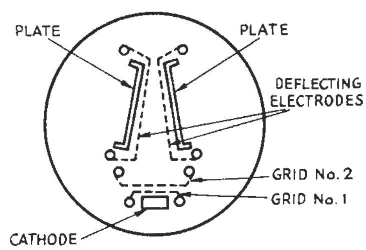


FIG. 1—Arrangement of electrodes in the beam-deflection modulator tube, type 7360.

Mechanical-Filter Type S.S.B. Generator

Now, bearing these brief fundamentals in mind, let's see how they can be applied in a balanced modulator, using a mechanical filter to obtain a single-sideband signal. A 455-kc. circuit for this purpose is shown in Fig. 2.

The 7360 beam-deflection tube as used in this circuit combines two basic functions—it generates its own 456.85-kc. carrier as a crystal-controlled oscillator, and it also functions as a balanced modulator which delivers both sidebands without the carrier to the mechanical filter. The filter suppresses one sideband and delivers the other to its output circuit.

The control and screen grids of the 7360, together with its cathode, are used in a self-oscillating circuit. It is also entirely practical, of course, to supply the carrier to the control grid from a separate oscillator, if desired, as will be described later.

In the self-oscillatory circuit shown in Fig. 2 the 1N34A diode connected between the control grid and the ground side of the cathode-bias resistor acts as a clamp to prevent the voltage on the control grid from going positive or even to zero volts. As operated here the most positive excursion of the control grid is to −2.5 volts. This prevents excessive values of screen current from flowing and gives better modulation linearity and carrier frequency stability.

We now have a single-ended carrier input to our push-pull plate output. In order to suppress this carrier in the push-pull output circuit both ends of the output circuit must go equally positive and then equally negative at exactly the same times. That is, the amplitudes of the two voltages, one from each plate, must be exactly equal and the two voltages must be in exactly the same phase in order to balance out or cancel the carrier completely. Amplitude balance is obtained very simply by varying the d.c. voltage difference between the two deflecting electrodes by means of the amplitude-balance potentiometer, R_1.

Since a phase unbalance of only 1 degree makes it impossible to obtain more than about 40 db. of carrier suppression, special pains were taken to provide good methods of obtaining stable phase balance. In the method used in Fig. 2 the phase angle of the load circuits is controlled by varying the resistance to ground

From March, 1960, *QST*.

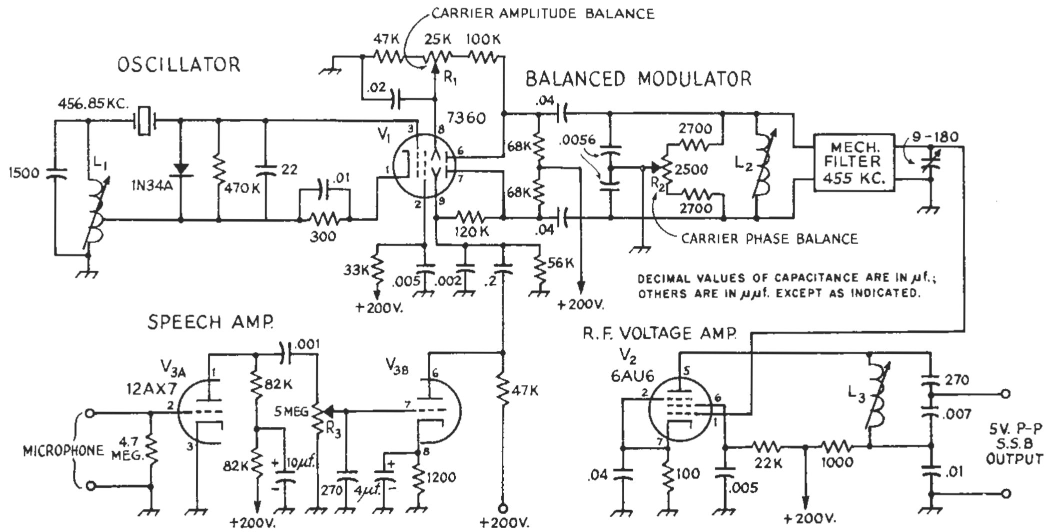


FIG. 2—Filter-type s.s.b. generator using the 7360 as a combined self-excited oscillator and balanced modulator. Fixed resistors are ½ watt except as indicated.

L_1—App. 88 µh., adjustable (see text).
L_2—App. 50 µh., adjustable (see text).
L_3—App. 450 µh., adjustable.
R_1, R_2—Composition control, linear taper.
R_2—Composition control, audio taper.

from the two ends of the capacitance-center-tapped plate tank circuit by means of the phase-balance potentiometer, R_2. This method is best suited for relatively low-frequency operation. A small differential capacitor across the plate tank circuit is best suited for phase balancing in higher-frequency circuits.

These two balance controls allowed better than 60 db. of carrier suppression to be obtained from the balanced modulator. An additional 20 db. of suppression was obtained in this filter circuit because the carrier was located at a point 20 db. down on the filter frequency-response curve. This gave a total of approximately 80 db. suppression!

Negative Feedback

You will notice that each deflecting electrode receives its d.c. voltage (approximately 25 volts) from a tap on a resistance voltage divider from each plate to ground. A desirable by-product of this resistance coupling from plate to deflecting electrode is negative feedback. Additional negative feedback, which tends to correct any unbalance between the two plate currents, is obtained from the two 68,000-ohm plate resistors. If desired, still more negative feedback can be obtained by reducing the capacitance of the two 0.04-µf. plate blocking capacitors so that they offer a relatively high reactance to audio frequencies but relatively low reactance to the carrier frequency being used. This negative feedback improves modulation linearity and reduces hum and microphonics originating in the balanced modulator.

The audio signal is fed to this balanced modulator at a relatively high impedance—special transformers to obtain a low-impedance driving source are not required. Furthermore, a push-pull audio source is not required—a single-ended feed is used. With 10 volts r.f. on the control grid, 2.8 volts a.f. deflecting voltage is required. Both values are peak-to-peak.

This arrangement is possible because of the fundamental characteristic of the tube mentioned initially—the voltage difference between the two deflecting electrodes controls the division of the total plate current between the two plates. This gives an intrinsically balanced push-pull output from a single-ended input. Also, a 180-degree phase reversal can be obtained by simply switching the audio input from one deflecting electrode to the other.

In order not to overload the mechanical filter the gain of this circuit was reduced by shunting the mechanical filter input winding with a separate slug-tuned inductance, L_2, as shown in Fig. 2. This allowed about 1.5 volts input to the filter. Approximately 0.5 volt peak-to-peak output was obtained from the filter.

This shunting inductance, L_2, consisted of 63 turns of No. 36 enameled wire, close-wound in a single layer on a tube 9/32 inch in outside diameter. A ¼-inch-diameter iron slug in the tube allowed the plate tank circuit to be resonated at the crystal frequency.

The tank coil, L_1, for the 456.85-kc. crystal oscillator must be of the high-Q ferrite-core variety. One of the kind normally used as an oscillator coil with the 6BE6 converter tube for the standard a.m. broadcast band was used in our tests. The cathode excitation can be obtained by a tap on the coil. The cathode tap point should be located above ground 13 to 15 percent of the total number of turns in the coil as a starting point. If possible, the r.f. voltage between grid and cathode should be measured with a high input-impedance v.t.v.m. equipped with an r.f. probe, and the tap point varied so as to obtain 10 volts r.f. peak-to-peak (3.5 volts r.m.s.) between the grid and cathode.

When using 1500 pf. total oscillator tank capacitance, provision should be made for varying the coil inductance above and below about 88.5 μh. by a percentage a little larger than the capacitance tolerance percentage of the tank capacitor used, in order to resonate the coil-capacitor combination at the crystal frequency. Varying the tuning of the tank circuit around the resonance point will vary the oscillator frequency slightly.

Voltage Amplifier

A voltage amplifier suitable for raising the 0.5-volt output from the mechanical filter to a more usable level consists of a 6AU6 stage, Fig. 2, that has a capacitance voltage divider as a part of its plate tank circuit. With the constants shown, it can provide an s.s.b. peak-to-peak output of about 5 volts. However, any output up to about 150 volts can be obtained by changing the capacitance ratio of the two voltage-divider capacitors that are connected in series across the 6AU6 plate tank circuit. When this ratio is changed the resultant capacitance of the two capacitors in series must remain constant so the L/C ratio of the tank circuit is not changed too much.

V.F.O.-Mixer Circuit

Fig. 3 shows a schematic of a v.f.o.-mixer unit. Its resemblance to the balanced modulator circuit is quite evident. Here the front end again functions as an oscillator, except that it is of the variable-frequency type.

The modulating signal is the s.s.b. output from the 6AU6 stage described above. Again the modulation is applied to only one deflecting electrode, the other being at r.f. ground due to the 0.005-μf, bypass capacitor.

The mixer tank circuit employs a center-tapped, bifilar-wound inductor with the 68,000-ohm feedback resistors in its center-tapped connections to the dc. plate voltage supply. This mixer output transformer, T_1 in Fig. 3, was constructed as follows for our tests on 3.9 Mc.:

Primary—Bifilar-wound on ½-inch diameter tube, tuned with a ¼-inch slug; winding length, ⅜ inch. Two wires wound parallel to each other on tube, 23½ turns of each wire (47 total). No. 34 wire, single Teflon insulation if possible (silk insulation can be used if necessary). The dielectric properties of the insulation on the wire are important because in a bifilar winding the distributed capacitance is relatively high and is a part of the tank capacitance. This accounts for the relatively low value of 22 pf. shown for the tank capacitor in this circuit.

Secondary—Twenty-six turns of No. 32 wire with Formex insulation, close-wound in a single layer. The spacing of this winding from one end of the primary winding should be adjusted so as to obtain satisfactory bandpass between 3.8 and 4.0 Mc. Approximately 190 pf. was required to resonate the secondary to 3.9 Mc.

Without the carrier amplitude-balance control, R_4, shown in Fig. 3, the balanced load circuit provides 20 to 25 db. suppression of the v.f.o. carrier. Including the carrier amplitude-balance control allows about 40 db. total v.f.o. carrier suppression, thus simplifying the selectivity requirements of the output circuit.

The grid and cathode connections of the v.f.o. are tapped down on the inductor so as to reduce the coupling between the tube and the tank circuit, and thus improve stability and obtain the correct r.f. voltages on the tube elements.

The v.f.o. tank coil, L_4, in Fig. 3, consists of 15 turns of No. 22 enameled wire spaced uniformly in a winding 0.6 inch long on a 1-inch diameter coil form. No slug was used. The grid tap should be located 7½ turns above the ground end of the coil. The cathode tap is 1⁹⁄₁₀ turns above the ground end of the coil. Actually, in building L_4, it was first determined that the coil specified above required a total length of wire measuring 121 cm. The taps were then soldered to this length of wire before winding it on the coil form, in order to avoid melting the polystyrene form with the hot soldering iron. The cathode tap was located 15.1 cm

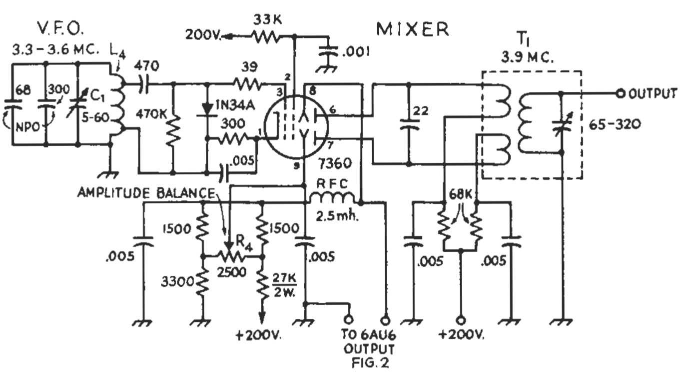


FIG. 3—Combined v.f.o. mixer for frequency conversion from 455 kc. to 3.8-4.0 Mc. Fixed resistors are ½ watt except as indicated. Decimal values of capacitance are in μf., others are in pf.
C_1—Variable, 5–60 pf.
L_1—See text.
R_4—Composition control, linear taper.
T_1—See text.

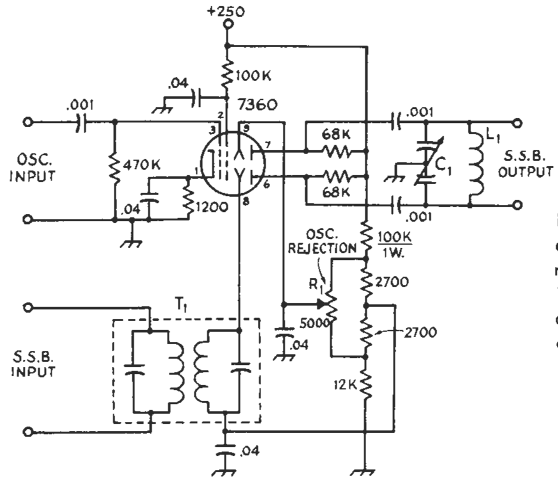


FIG. 4—Balanced mixer circuit with separate excitation. Capacitances are in μf.; fixed resistors are ½ watt except as indicated. Tuned-circuit constants depend on frequency; ordinary L/C ratios may be used for L_1C_1 and in the s.s.b. input transformer, T_1. R_1 is a linear-taper composition potentiometer.

(12½ per cent of the total wire length) from the ground end of the wire and the grid tap was located 60.5 cm (50 per cent of the total wire length) from the ground end. At 3.5 Mc. this coil had a Q of 150 and required 400 pf. to resonate it. The Q was measured with the coil shield in place.

7360 Mixer With Separate Oscillator

A generalized mixer circuit for use with an external r.f. oscillator is shown in Fig. 4. Here the s.s.b. input is shown fed from a two-winding transformer instead of from a capacitive tap on the preceding tank circuit, as was used in Fig. 2.

Since the 7360 is not used as a self-oscillator the 1N34A diode clamp is not used and the cathode bias resistor is changed from 300 ohms to 1200 ohms. The r.f. oscillator input to the control grid must be adjusted to be between 5 and 10 volts peak-to-peak, measured between control grid and cathode, for best results. A 0.04-μf. r.f. bypass capacitor effectively grounds one deflecting electrode (pin 9) so the s.s.b. input is single-ended between the other deflecting electrode and ground.

In this mixer circuit the 68,000-ohm d.c. feedback or plate-current equalizing resistors are connected as shunt feed resistors to the two plates, the same as was shown in the balanced modulator circuit in Fig. 2, instead of being in series with the center-tapped connections to the d.c. plate voltage supply as was shown in the mixer circuit of Fig. 3. This difference allows L_1, the mixer plate tank coil, to be a simple untapped coil instead of requiring a bifilar-wound coil as is the case when the d.c. feedback resistors are connected in series with the split, center-tapped coil connections to the plate voltage supply. The capacitance and inductance values of T_1 and C_1 and L_1 in Fig. 4 will depend upon the input and output frequencies involved.

The combination of the two circuits shown in Figs. 2 and 3 will provide single-sideband output in the range between 3.8 and 4 Mc. Since these circuits were for the purpose of obtaining characteristics and specifications, as was previously stated, they do not include all of the facilities that might be required for actual amateur operation on the air, particularly as regards switching between upper and lower sideband.

Sideband switching can be obtained by any of the normal methods. For example, a simple method would be to employ two crystals in the carrier oscillator circuit of the balanced modulator, one for upper sideband and the other for lower sideband, with a switch for instant choice. Band switching would require an additional mixer stage to heterodyne the v.f.o.-mixer output to the various other bands.

R.F. Phasing-Type S.S.B. Generator

Fig. 5 shows the schematic of an r.f. phasing exciter circuit for 455 kc. which gives a peak-to-peak output of about 4 volts single sideband. Here you see the usual r.f. and a.f. 90-degree phase-shift networks, an audio amplifier, and two 7360 tubes as balanced modulators.

The audio circuits include two sideband-balance potentiometers—one for adjusting the input voltage ratio to the audio phase shift network, R_3, and one for audio amplifier balance, R_4. Each balanced modulator has its own carrier amplitude-balance potentiometer which controls the d.c. bias voltage on one of its deflecting electrodes, as was done in the filter rig.

The outputs of the two balanced modulators are combined in a common push-pull tank circuit. Over-all r.f. phase balance is obtained in this tank circuit by the use of a differential capacitor connected across the tank circuit. This type of phase-balancing circuit is better suited to the higher carrier frequencies generally used in phasing-type exciters.

The r.f. phase-shift network used here is of the simple R-C bridge variety. Any of the

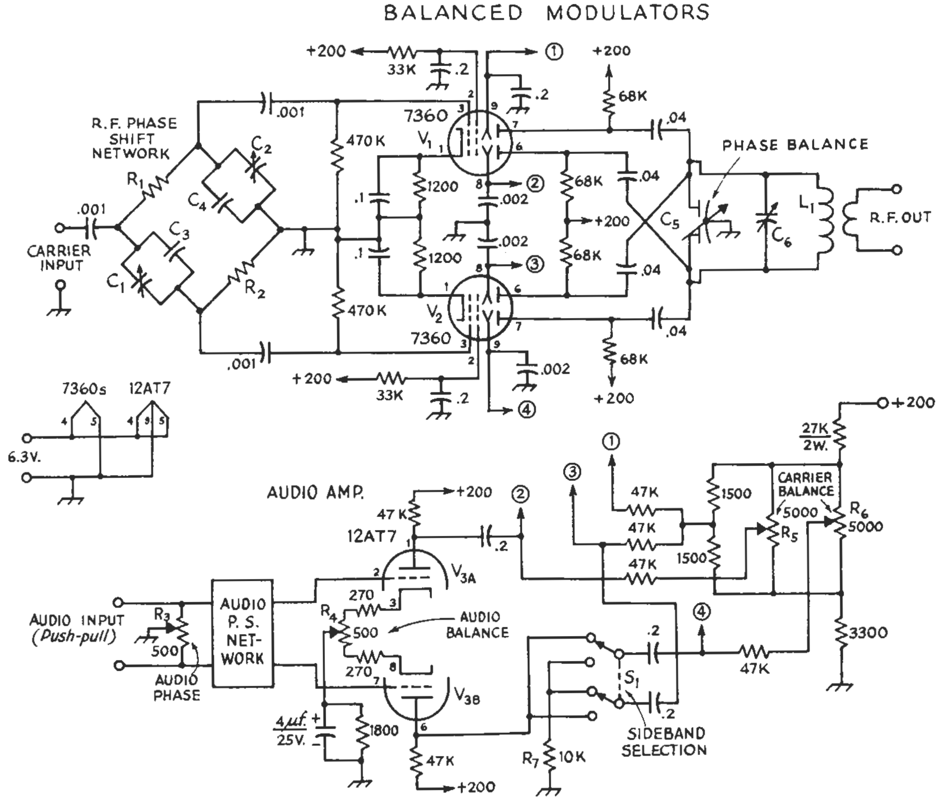


FIG. 5—Phasing-type s.s.b. generator using two 7360s as balanced modulators. Capacitances are in pf.; fixed resistors are ½ watt except as indicated.

C_1, C_2—Trimmers, ceramic or air, approximately 25 pf.

C_3, C_4—Value dependent on frequency and R_1R_2. Reactance at operating frequency should be approximately equal to the resistance of R_1 and R_2.

C_5—Differential capacitor, approximately 25 pf. per section.

C_6—Variable, to resonate with L_1 at output frequency.

L_1—To resonate with C_6 at output frequency. Conventional L/C ratio may be used.

S_1—D.p.d.t. toggle or rotary.

R_1, R_2—Non-inductive, ½ or 1 watt, values to be equal within close tolerances. Actual resistance not critical, but should be low to minimize effect of stray capacitances. Resistances of the order of 100 ohms are satisfactory.

R_3, R_4—500-ohm composition control, linear taper.

R_5, R_6—5000-ohm composition control, linear taper.

R_7—App. 10,000 ohms, ½ watt (see text).

other usual types of networks could be employed instead if desired.

As in the filter exciter, better than 60 db. suppression of the carrier was obtained by proper adjustment of the r.f. phase and amplitude balancing controls.

Suppression of the unwanted sideband is limited by the degree of accuracy with which the audio phase-shift network maintains an exact difference of 90 degrees in phase between the two branches of the audio system, over the entire range of audio frequencies fed to the audio phase-shift network. With the audio phase-shift network used in our experimental setup carrier plus unwanted sideband measured about 40 db. below the wanted sideband.

Switching the single-ended audio input to one of the balanced modulators from one deflecting electrode to the other allows a ready choice of upper or lower sideband outputs. This is done by switching the audio input to one balanced modulator, V_2, from one deflecting electrode to the other, through one arm of S_1. The other arm of S_1 connects a 10,000-ohm resistor, R_7, between the other deflecting electrode and ground through a d.c. blocking capacitor, in order to preserve better balance. The exact value of R_7 should be adjusted for best balance stability since various wiring layouts, and particularly various audio tube types, will require different values.

As in the case of the filter circuit, additional stages and functions would be required for a complete s.s.b. exciter, including a v.f.o.-mixer stage and a crystal oscillator mixer stage to heterodyne the signal to various bands.

» A concise presentation of the principles of product detection —which, like linear amplification, is an operating state rather than a circuit as such.

Product Detectors

MURRAY G. CROSBY, W2CSY

The detector in many communications receivers is of the diode type, its design requirements being those of amplitude modulation. The beat frequency oscillator has been somewhat of an afterthought so that c.w. telegraph reception could be included in the features of the receiver. The usual design with respect to the beat frequency oscillator was to couple a small amount of oscillator voltage into the i.f. circuits so that the oscillator and incoming signal together formed a beat note in the detector. In many cases the amount of beat frequency oscillator voltage injected into the detector was very low. Such an arrangement functioned fairly satisfactorily for c.w. telegraph reception since that type of reception may be accomplished when the signal at the detector is stronger than the local beat frequency oscillator. This type of oscillator injection produces a certain amount of limiting, which does not distort c.w. telegraph signals and many tend to smooth out fading.

When suppressed-carrier single-sideband signals are being received, the b.f.o. must be used to supply the carrier. However, when this is done under detection circumstances which were set up to receive c.w. telegraph, distortion may result. When such a diode detector is used for single-sideband reception the only possible adjustment is to turn the audio gain control as high as possible and turn the i.f. or r.f. gain control as low as possible. Under these circumstances the level of the signal entering the detector may be equal to or less than that of the b.f.o. and proper reception may be obtained. Sometimes this adjustment is impossible, however, because the audio gain must be turned up so high that hum is excessive. As a consequence the normal condition of a large number of ordinary communications receivers is that they inherently distort in single-sideband reception.

Increasing the b.f.o. injection is a considerable help and may make single-sideband reception possible in many cases. However, it may introduce new problems such as pulling of the local oscillator in reception of c.w. signals or hum and microphonics.

The ideal answer to the situation is a good "product" detector, designed for proper reception of single-sideband transmission. In addition, a sideband filter is very desirable, since by eliminating one sideband a large amount of interference may be rejected.

The Product Detector

Fig. 1 shows the basic arrangement of a product detector. Its general nature is that it has two separate inputs. One of these inputs is used for the sidebands and the other for the carrier oscillator. It is called a "product" detector because the audio output is a mathematical product of the two separate inputs which are fed to the tube grids in the detector circuit. The only audio output is that which results from coaction of the local carrier oscillator and the incoming sidebands. That is, the audio output is comprised of beat notes or heterodynes between the carrier oscillator and individual sidebands of the incoming signal, and there is no detection of the signals applied to the signal grid when the carrier oscillator is switched off.

A good test of a product detector is to switch off the carrier oscillator and listen to determine if there is any detection of a modulated signal applied to the signal grid. The best product detectors reject detection, when the carrier oscillator is switched off, to the extent of 40 or 50 db. For instance, assume a double-sideband a.m. signal is coming in on the sideband input. If the carrier oscillator is switched off there will be no audio output, or at least it will be negligibly low. When the carrier oscillator is turned on there will be reception of the double-sideband signal if the local carrier oscillator is synchronized with the

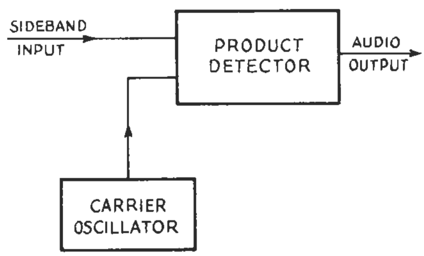


FIG. 1—Basic arrangement of the product detector. Two inputs, sideband and local (carrier) oscillator, combine in the detector to produce an audio output proportional to the product of the applied voltages.

From "Reception With Product Detectors," *QST*, May, 1956.

incoming carrier. If the local carrier oscillator and incoming carrier oscillator are not synchronized, the detection is unintelligible and all that comes out of the detector is a beat note between the carrier and local oscillator, together with distortion from the sidebands.

S.S.B., A.M., P.M. and C.W. Detection

The product detector not only is the ideal detector for single-sideband signals but is also ideal for the reception of double-sideband amplitude-modulation signals, double-sideband phase-modulation signals, and c.w. signals. With double-sideband reception it offers the advantages of exalted-carrier detection with its reduced fading distortion.[1] The same advantages are present for the reception of p.m. of the type using a peak phase deviation of one radian or 57 degrees. For the reception of p.m., it is a better detector than an f.m. discriminator since the product detector is less susceptible to selective fading.

For c.w. reception the product detector has the advantage of producing a clean, undistorted beat note. It is more linear than the ordinary diode detector operating with a low value of b.f.o. injection. This gives the maximum signal-to-noise ratio at all times. Also, when used in conjunction with a sideband filter in the i.f. circuit, "single-signal" reception is obtained so that the audio image is rejected. The product detector is therefore the ideal c.w. detector.

Dual-Grid Product Detectors

Fig. 2 shows the pentagrid converter type of product detector. In this type the oscillator and signal grids provide the two inputs. Electrode voltages, signal level, and carrier level are adjusted so there is a minimum of detection when the carrier oscillator is switched off and a modulated signal is fed to the sideband grid. The measurement that might be made would be to turn the oscillator on, feed an unmodulated carrier to the sideband grid and measure the audio output. Then switch the oscillator off and measure the level of audio output when a 50 per cent amplitude modulated signal is applied to the sideband input. The difference between the two levels should be at least 25 db. to produce good exalted-carrier product detection.

A disadvantage of the pentagrid converter type of product detector is the variation between tubes. Replacement of the tube may require readjustment of element voltages. However, if signal and oscillator levels are kept low this effect can be minimized.

The 6BN6 f.m. detector operated in the non-limiting condition may be used as a product detector in the same manner as the ordinary pentagrid converter tubes.

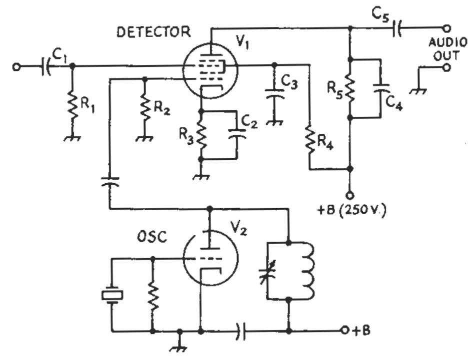


FIG. 2—Product detector using a pentagrid converter tube. Circuit values are conventional for the frequency, and electrode voltages are normal for the particular types of tubes used. V_1 may be a 6BA7, 6BE6, 6SA7, etc. V_2 may be a 6C4, one section of a 12AU7, 6SN7GT, etc. Any convenient oscillator circuit may be employed. For 455 kc. and a 6BE6, suggested values are:

C_1—100 pf. to 001 μf.
C_2—0.01 to 0.1 μf.
C_3—0.1 μf.
C_4—100 to 500 pf.
C_5—Depending on audio load resistance (0.01 satisfactory for 0.5- to 1-megohm load).
R_1—0.5 to 1 megohm.
R_2—20,000 ohms.
R_3—150 to 300 ohms.
R_4—22,000 ohms.
R_5—50,000 ohms.

Oscillator amplitude should be adjusted so that not more than 10 volts r.m.s. is applied to the No. 1 grid of the converter.

Triple-Triode Product Detector

Fig. 3 shows the triple-triode type of product detector.[2] It uses three triodes such as those in the 12AU7 twin-triode tubes. A particularly convenient arrangement is to use two 12AU7s with three of the triodes in the detector and the fourth either used for the carrier oscillator or as an audio amplifier. Various arrangements of resistances and biases may be worked out. The one shown uses a common cathode resistor of 1000 ohms and a bypassed cathode resistor connecting the output triode with the two cathode-follower input triodes. In effect, the arrangement is two cathode followers which receive the sideband input and the carrier oscillator input, respectively, and a cathode-driven output tube which gives the detected audio output. With this arrangement a sideband input of 0.25 to 0.5 volt r.m.s. and a carrier oscillator input of about 3.5 volts will produce proper operation. This gives about one-half volt audio output.

If desired, the 2700-ohm self-biasing resistor

[1] In exalted-carrier reception the carrier is amplified considerably more than the sidebands, before detection. This reduces the distortion that results from selective fading when the carrier amplitude fades below the sideband amplitude.—*Ed.*

[2] M. G. Crosby, U.S. Patent No. 2,470,240, May 17, 1949.

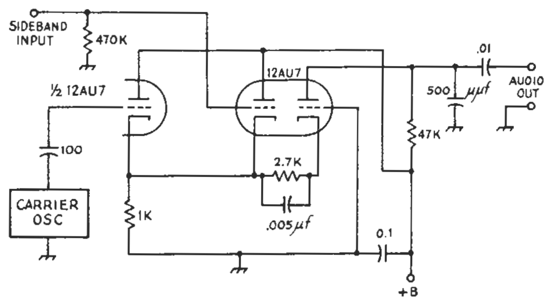


FIG. 3—Triple-triode product detector. Fixed bias may be substituted for the cathode bias (2700-ohm) resistor on the output triode for fine adjustment of the detector operating point, as explained in the text. In the practical circuit, a d.c. grid return should be provided for the triode section connected to the carrier oscillator. With capacitive coupling as shown, a resistor of 0.1 megohm or more may be used.

may be removed and all cathodes tied together, in which case fixed bias must be applied to the output triode grid. This bias runs in the order of a few volts and is usually negative. However, it is best to adjust it both negative and positive to minimize the output when a modulated signal is applied to the sideband input with the carrier oscillator off. In other words, make the same adjustment of switching off the carrier oscillator and adjusting the bias so that detection is minimum from the sideband-input grid.

» Using a 7360 beam deflection tube results in a much improved product detector for sideband reception. The tube provides greater audio output voltage, lower intermodulation distortion, and the feature of impulse noise limiting.

The 7360 As A Product Detector

JOHN M. FILIPCZAK, K2BTM

Although pentagrid converters are basically product-detection devices, they have some inherent limitations. Characteristics of the pentagrid-coverter tube are such that small changes in element voltages can shift tube operation out of the "center of the linear range" under large-signal conditions. The pentagrid product detector, for example, has the carrier-insertion signal applied to grid No. 1 and the modulated sideband signal to grid No. 3. Because of the electronic interaction existing between grid No. 1 and grid No. 3, pentagrid converter tubes are seldom used to generate their own beat-frequency-oscillator signals in product-detector circuits.

A second method of product detection: i.e., the popular Crosby system,[1] uses two dual-triode units which require additional socket space and components. The limitations of both systems can be circumvented by the use of the 7360.

The 7360 is a grid-controlled beam-deflection tube having a cathode, control grid, screen grid, two deflecting electrodes, and two plates in a nine-pin miniature envelope. The tube was specifically designed for application in such sideband circuits as balanced modulators, balanced mixers, product detectors, and frequency converters.[2]

The tube structure is such that the total beam current is determined by the voltage applied to grid No. 1 and grid No. 2. The difference in voltage between the deflecting electrodes determines the amount of beam current collected by each plate. In balanced operation, the beam current is divided equally between the two plates. When signals are applied to grid No. 1 and one of the deflecting electrodes, the resultant output contains signal components produced by the product of the input signals. Therefore, if the modulated signal is applied to one of the deflecting electrodes and the carrier insertion is applied to grid No. 1, the resultant output contains the desired audio component.

A typical circuit for use with a separate beat-frequency oscillator is shown in Fig. 1.

Condensed from December, 1960, *QST*.

[1] Crosby, "Product Detectors," page 34.

[2] Vance, "S.S.B. Circuits Using the 7360," page 29.

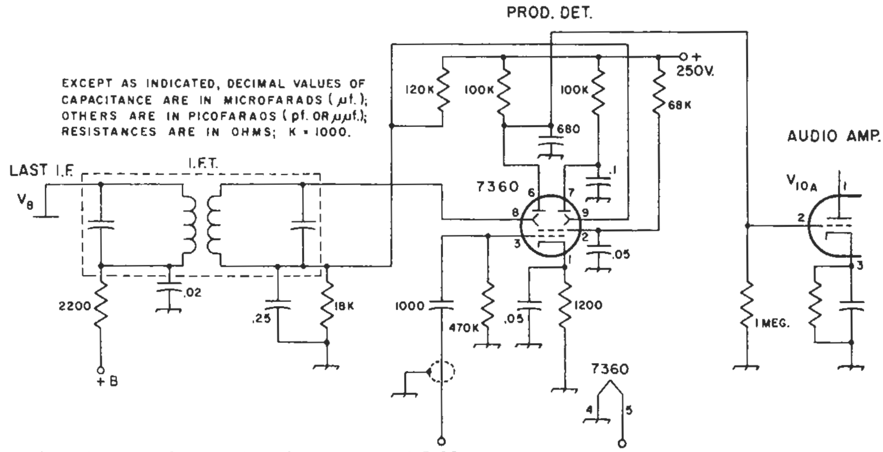


FIG. 1—Product detector circuit using the 7360. Resistors are ½-watt composition.

The signal to be detected is taken from the secondary of the last i.f. transformer; the i.f. amplifier can be one operating at any of the commonly-used intermediate frequencies. Audio output is taken from one of the plates and applied to the grid of an audio voltage amplifier.

The b.f.o. output should be adjusted to bring the voltage to 10 volts peak-to-peak on grid No. 1 of the 7360. This adjustment is preferably made with the aid of an oscilloscope. If a scope is not available, the b.f.o. voltage may be adjusted for maximum undistorted audio-output signal.

A quick operating check of the completed circuit is simple to perform. Turn on the receiver, tune in an s.s.b. signal and adjust the b.f.o. for clear reception. Switching off the b.f.o. at this point should result in negligible audio output.

Self Excitation

If desired, the 7360 may be used to provide its own b.f.o. excitation. The circuit is shown in Fig. 2. C_1 is used to adjust the signal level on grid No. 1 to a value of from 5 to 10 volts peak-to-peak with respect to the cathode. This adjustment shifts the b.f.o. frequency somewhat, but the shift can be compensated for by adjustment of the capacitor in the b.f.o. transformer. The pitch control is the front-panel control and is used to zero beat the incoming signal. The grid-clamping diode CR_1 prevents the No. 1 grid from approaching too closely to zero voltage, at which point distortion would result. Because the detector conversion gain of the 7360 in the product-detector circuit is about 6, the audio-output stage can be driven directly in most cases.

Noise Limiting

Another feature of the 7360 is its excellent noise-limiting capabilities. The normal signal voltage appearing at the deflecting electrodes should be limited to a maximum value of 8 volts peak-to-peak. If this voltage becomes larger, the audio signal becomes slightly clipped. Noise pulses ten times greater than the s.s.b. signal were only twice the peak audio signal after detection. It is recommended that the deflecting-electrode signal be kept near the maximum of 8 volts peak-to-peak to take advantage of this signal-limiting feature.

Precautions

The 7360, like other types of beam-deflection tubes, is affected by stray magnetic fields. Variations in magnetic fields cause corresponding variations in plate currents, and upset the tube's exceptionally good balance. Therefore, a tube shield (iron or steel) is recommended for most applications.

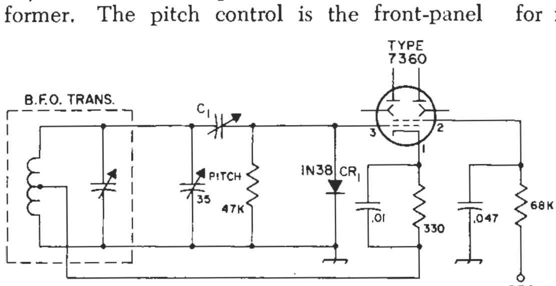


FIG. 2—B.f.o. circuit for self excitation.
C_1—30-pf. ceramic or mica trimmer.
CR_1—1N38 germanium diode.

» 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, ex-W4OLL

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 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.

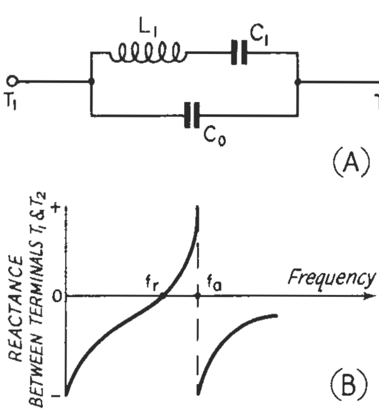


FIG. 1—The equivalent electrical circuit of a piezoelectric crystal (A). The reactance varies with frequency as in (B).

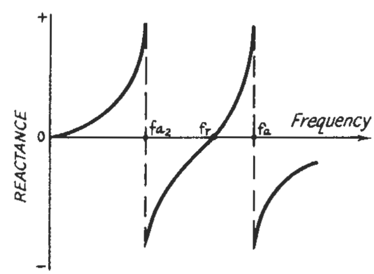


FIG. 2—Reactance plotted against frequency for crystal shunted by an inductance.

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

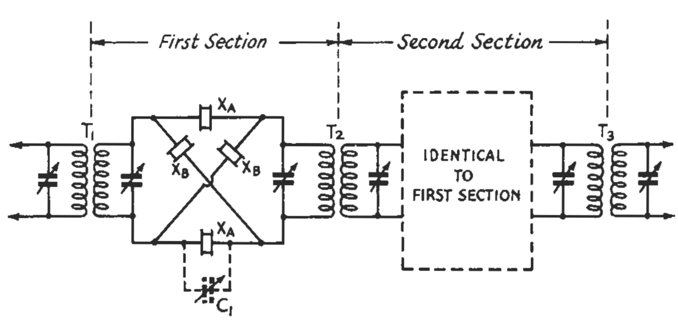


FIG. 3—Schematic diagram of a two-section crystal lattice filter.
T_1, T_2, T_3—Replacement-type i.f. transformers.
X_A, X_B—Matched pairs of FT-241 crystals (see text).
C_1—1- to 5-pf ceramic capacitor (see text).

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

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 (× 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 overspreading 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 attenuation ($f_{\infty}1$, $f_{\infty}2$, 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.

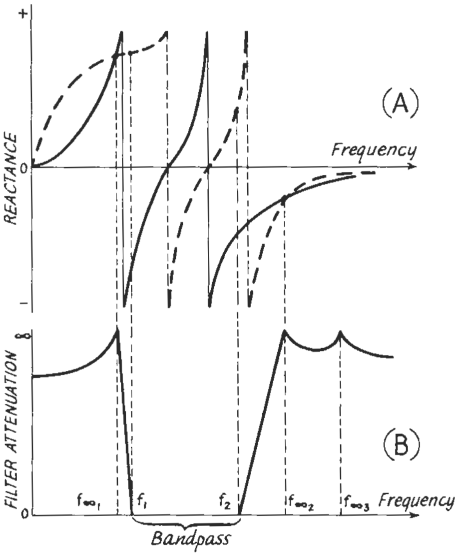


FIG. 4—The reactance-vs.-frequency characteristic for the two pairs of crystals in a lattice filter section is shown at (A). The resultant attenuation characteristic is shown in (B).

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

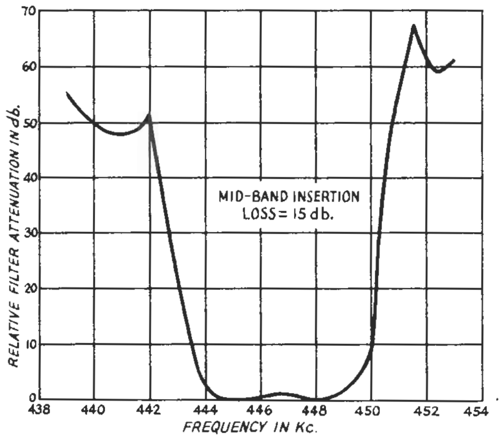


FIG. 5—Attenuation characteristic of an experimental crystal lattice filter (two sections) suitable for receiver use. The crystals were Channels 40 and 42 of the FT-241 series.

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 question 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.

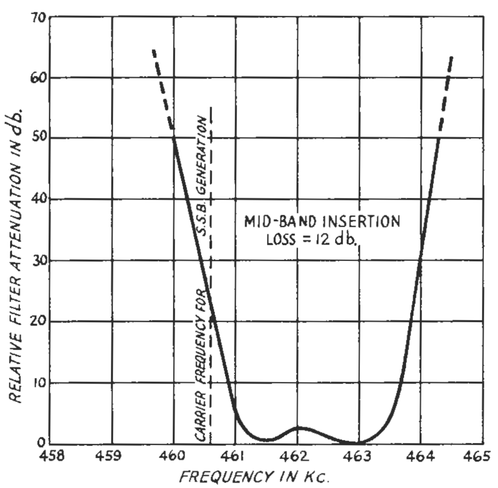
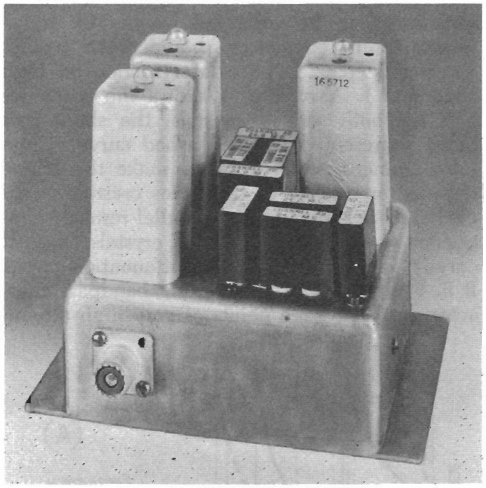


FIG. 6—Attenuation characteristic of an experimental crystal filter (two sections) suitable for a s.s.b. receiver or transmitter. The crystals were Channels 49 and 50 of the FT-241 series.

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 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" capacitor [1] 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.



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

[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 W3TLN [6] 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_o 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-

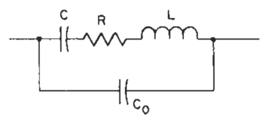


FIG. 1—Equivalent circuit of a crystal. C and L are the motional capacitance and inductance of the crystal, and R represents the frictional loss. C_0 is the electrode and holder capacitance shunting the crystal.

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.

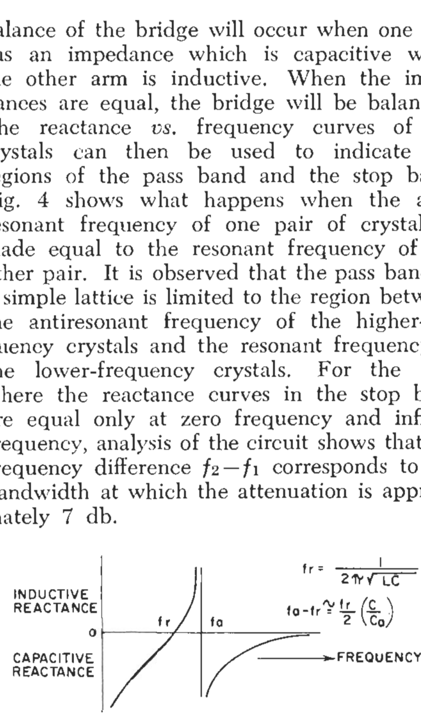


FIG. 2—Reactance vs. frequency characteristics of a crystal. The series-resonant frequency, f_r, is that of C and L. The anti-(parallel) resonant frequency, f_a, is that of the circuit formed by C and L in one branch and C_0 in the other.

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.

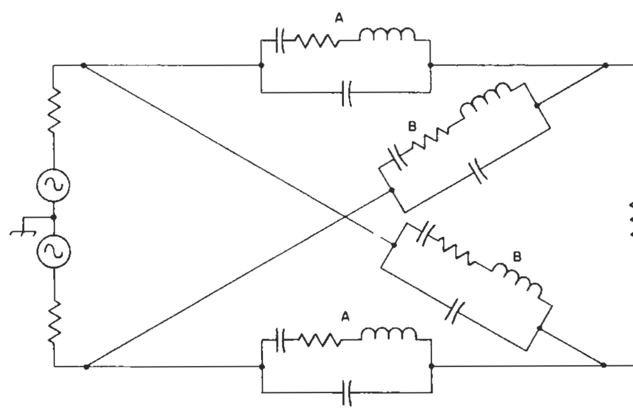


FIG. 3—Equivalent circuit of a full lattice crystal filter. The series crystals, *A*, are the same frequency, as are the shunt crystals, *B*. Note that the lattice could be redrawn as a bridge circuit.

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 tap[7] 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

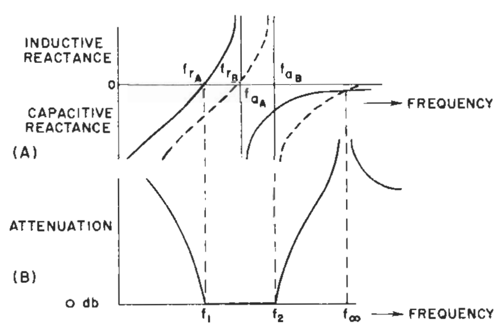


FIG. 4—(A) Reactance characteristics of crystals *A* and *B* in the lattice of Fig. 3. In the type of filter described in this article, the series-resonant frequency of the higher-frequency crystal, f_{rB}, is made equal to the antiresonant frequency of the lower-frequency crystal, f_{aA}. (B) Attenuation 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.

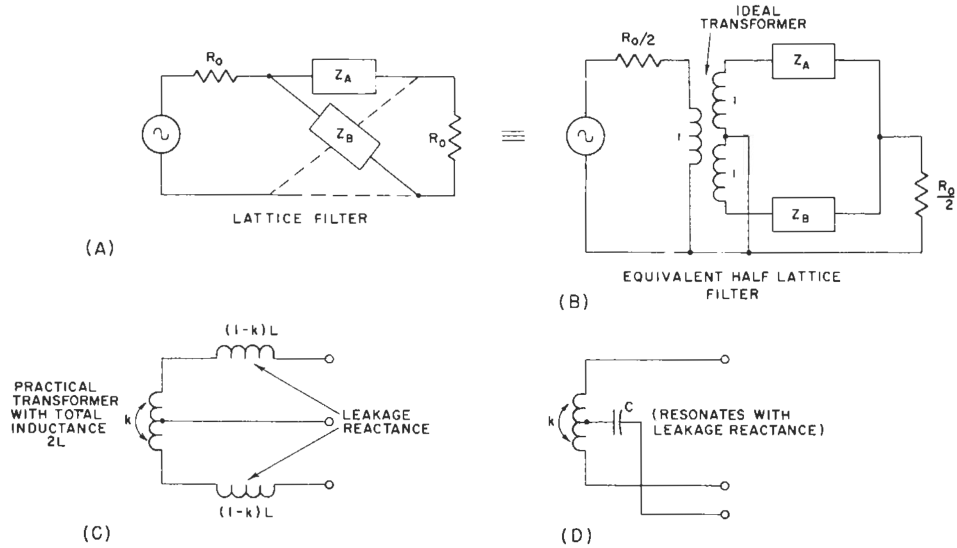


FIG. 5—(A) Basic circuit of a lattice filter and (B) an equivalent half-lattice filter. The transformer is an ideal one with unity coupling between windings. (C) Diagram of a practical transformer showing the leakage inductance which results from a coupling coefficient, k, less than unity. (D) How the 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 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 pastband characteristics may be distorted.

[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.*

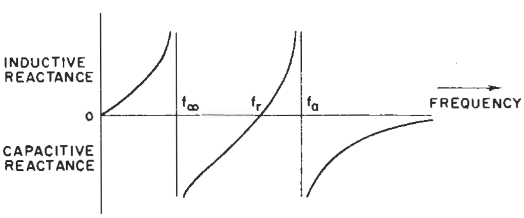


FIG. 6—Reactance vs. frequency characteristics of a crystal and inductance, L_o, in parallel. f_r, the resonant frequency of C and L, the motional capacitance and inductance of the crystal, is the same as in Fig. 2. f_a, however, is higher than the antiresonant frequency of the crystal alone, since L_o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

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_o, of 14.7 pf. The spacing between the resonant and antiresonant frequencies 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_o. This would require the addition of a *negative* capacitance of 14.7/2=7.35 pf. If an inductance, L_o, in parallel with capacitance C_o is considered the susceptance of the parallel combination is given by

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

where f_∞ is the antiresonant frequency of L_o and C_o. (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_o alone, $2\pi f C_o$, by a factor of one half. This means that $1 - \frac{f_\infty^2}{f^2}$ must equal ½, so L_o and C_o must be antiresonant at 0.707 times the filter frequency, or

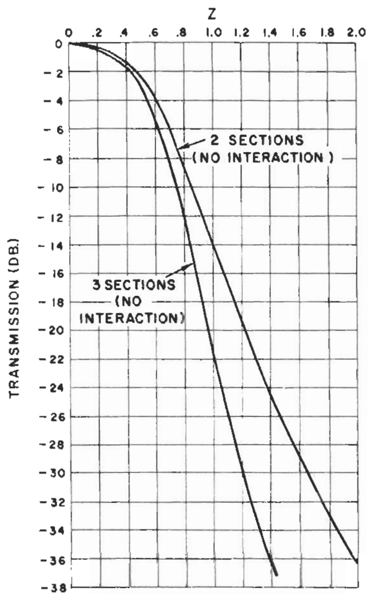


FIG. 7—Theoretical selectivity characteristics of two and three cascaded identical filter sections when interaction due to mismatch between the sections is ignored. Z is the normalized frequency variable defined in the text. It is proportional to the frequency difference from the center of the pass band.

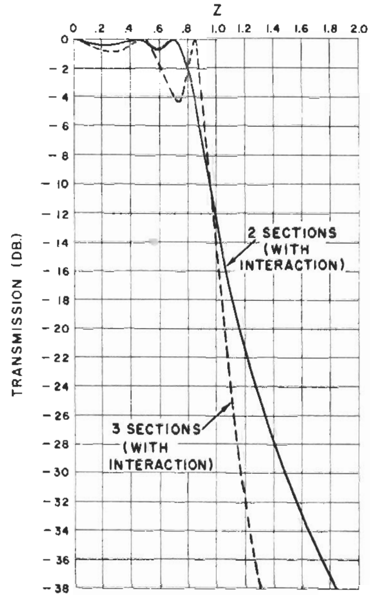


FIG. 8—Theoretical selectivity characteristics of two and three cascaded identical filter sections when interaction is taken into account. These curves are steeper than those of Fig. 7, but there is considerable ripple in the pass band.

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 pf., $1 - \dfrac{f_\infty^2}{f^2} = 7.35/47.7 = 0.154$, and f_∞ 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 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 sections [9] 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.

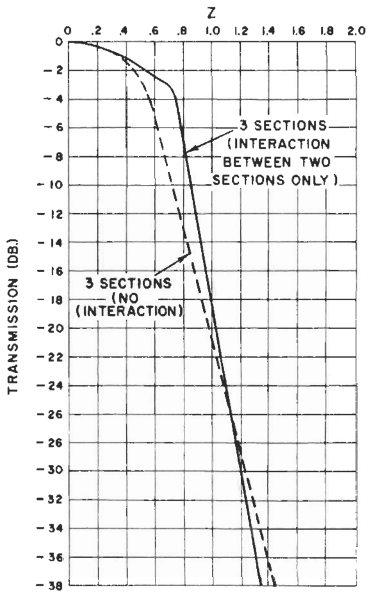


FIG. 9—The solid curve shows the selectivity of a three-section filter with interoction between two sections only. This design is an excellent compromise between those of Figs. 7 and 8. The dashed curve drawn for comparison is for a three-section filter without interaction.

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 $Z = 0.91 \times \frac{1500}{1200} = 1.138$. At this value of Z, the rejection is 26.5 db. At the low audio end of the undesir^d 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 characteristics[10] 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_o 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

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

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

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

or

$$C = 2\frac{\Delta f}{f_r}\left(C_1 + C_o\right)$$

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

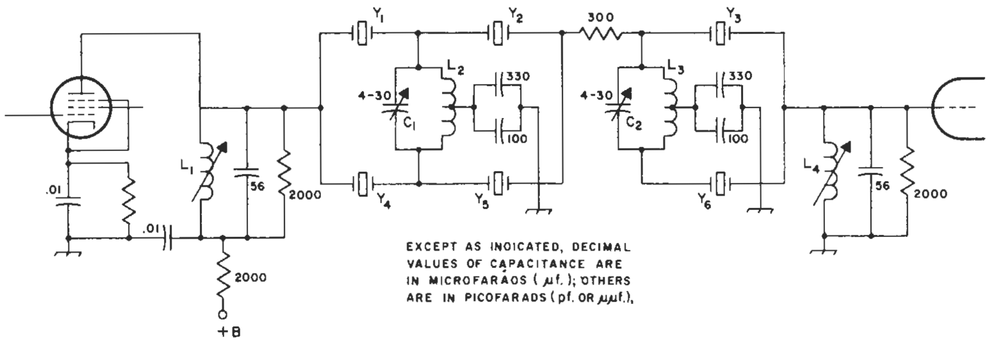


FIG. 10—Circuit diagram of a filter designed according to the methods of this article. Resistances are in ohms, and resistors are ½-watt composition; capacitors are disk ceramic except at noted.

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]

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

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_∞ 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 $Q/2f_oC$) 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 circuit [13] 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.

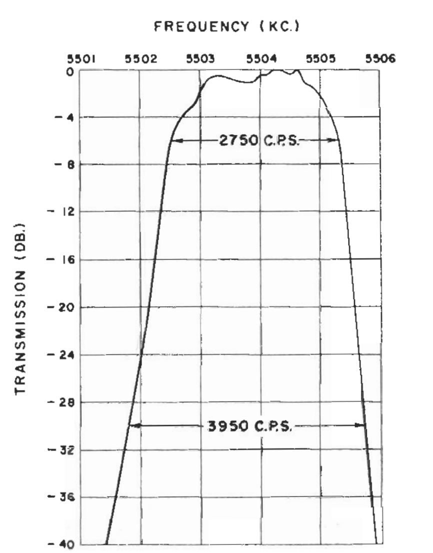


FIG. 11—Measured selectivity characteristic of the filter described in this article when aligned at 5505.2 kc. The 6-db. bandwidth is 2750 c.p.s., and the 30 db./6 db. shape factor is 1.44.

[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 Kosowsky [1] boils a lot of "long-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.

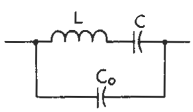


FIG. 1—The equivalent circuit of a crystal. *L* and *C* are the electrical equivalents of mechanical constants of the crystal, while C_0 is the shunting capacitonce of the electrodes and holder.

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 resonances 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-

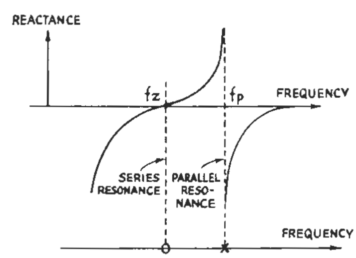


FIG. 2—Reactance characteristics of a crystal. The series-resonant frequency, f_z, is that of *L* and *C* (Fig. 1) in series; the parallel-resonant frequency, f_p, is the resonant frequency of the parallel circuit formed by *L* and *C* in one branch and C_0 in the other.

From January, 1959, *QST*.

[1] 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*.

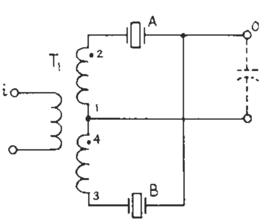


FIG. 3—The half-lattice crystal filter. Crystals A and B should be chosen so that the parallel-resonant frequency of one is the same as the series-resonant frequency of the other. Very tight coupling between the two halves of the secondary of T_1, is required for optimum results.

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-

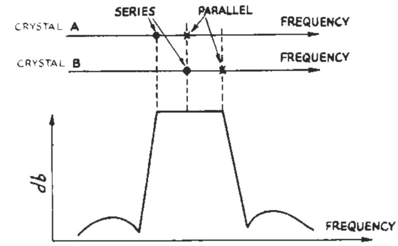


FIG. 4—The theoretical attenuation-vs.-frequency curve of a half-lattice filter shows a flat pass band between the lower series-resonant frequency and higher parallel-resonant frequency of the pair of crystals.

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,

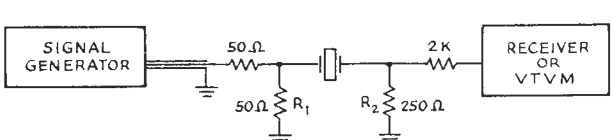


FIG. 5—Setup for measuring the series- and parallel-resonant frequencies (or pole-zero spacing) of a crystal.

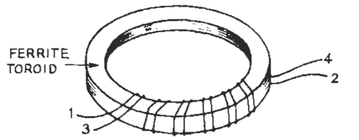


FIG. 6—Bifilar winding on a toroidal core.

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. 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 ¾-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 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

[2] *I.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.*

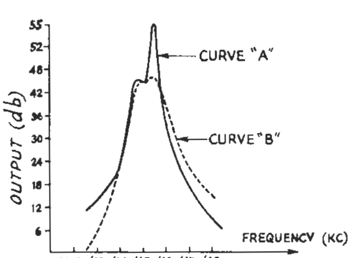


FIG. 7—Measured attenuation curves of a half-lattice filter using two nominal 5645-kc. crystals having series-resonant frequencies separated by 560 cycles. A—without resistance termination; B—with 10,000-ohm terminating resistor. In taking the data for these curves and those shown in Figs. 8, 9 and 11,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

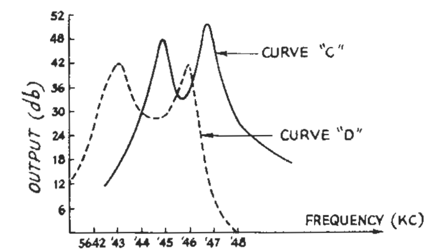


FIG. 8—Attenuation curves of half-lattice filter with crystals of the same nominal frequency as in Fig. 7, but with 1.5-kc. separation. C—with 0.5-megohm terminating resistor; D—shunt coil added across the output to resonate with capacitance present at that point.

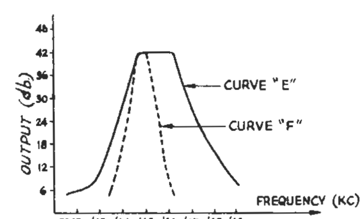


FIG. 9—E—half-lattice filter using same crystals as in Fig. 8, with 1500-ohm terminating resistor. F—using two nominal 5645-kc. crystals separated 300 cycles, with 3900-ohm terminating resistor.

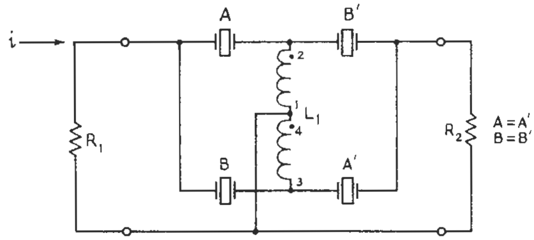


FIG. 10—Half-lattice filters cascaded in a back-to-back arrangement. The theoretical curve of such a filter has increased skirt selectivity and fewer spurious responses, as compared with a simple half lattice, but the same pass band as the simple circuit.

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

This was sufficiently encouraging, so I dug out the ammonium bifluoride[3] 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).

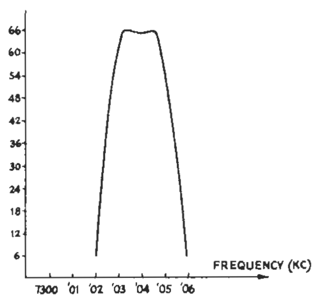


FIG. 11—Attenuation curve of filter using four nominal 7300-kc. crystals, pairs separated 1.5 kc., in the circuit of Fig. 10.

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 bandwith here is only half that obtained with the half-lattice. It should work nicely with the plated crystals, however.

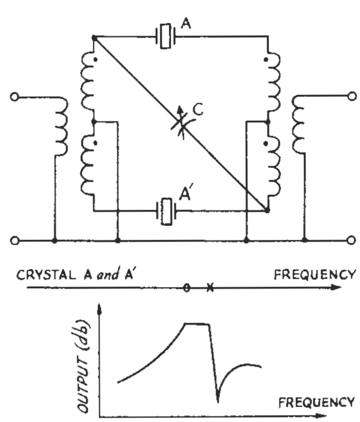


FIG. 12—An asymmetrical filter and theoretical attenuation curve.

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

» The crystal-etching technique described in this article has been used successfully by many experimenters, particularly those needing closely-matched pairs or sets for use in high-frequency filters. The etching solution is dangerous, but will cause no trouble if the instructions are followed carefully.

A Safe Method for Etching Crystals

ALBERT J. NEWLAND, W2IHW

Even after more than ten years, the surplus market still has an abundant supply of crystals at very low prices. Many of them will not multiply into any amateur band, unfortunately, and those that do are so widely used that surplus-frequency QRM is a common complaint on the v.h.f. bands. This article describes a safe and almost effortless way of raising the frequency of the pressure-mounted type of crystal, such as those found in the popular FT-243 holders.

Crystals may be raised in frequency by grinding with a fine abrasive. Valve-grinding compound, jewelers' rouge and various scouring powders may be used, provided that the work is done on a completely flat surface. A piece of plate glass is often used. This method has some disadvantages, however, one of them being occasional loss of crystal activity. Acid etching is a more satisfactory method, but has been little used by amateurs because of the dangers inherent in the handling of the solutions needed. The method of handling the ammonium bifluoride described here is not only safe, but it is an accurate and easy way of putting your surplus crystals on frequencies you won't have to share with hundreds of others when the v.h.f. bands are open for long-distance work.

From January, 1958, *QST*.

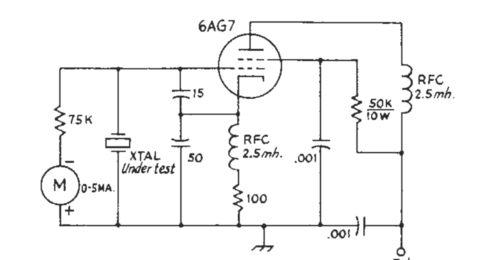


FIG. 1—Crystal activity may be checked by connecting the crystal in a circuit that has no tunable elements and measuring the grid current developed. The oscillation frequency in this circuit may be somewhat different from that when the station transmitter is used, so the activity-checking frequency should not be relied on when band edges are involved.

Equipment and Materials Needed

An oscillator for checking the activity of the crystals is very desirable, though it is by no means a necessity. The oscillator in the station transmitter will be usable for a frequency check, but if it has tunable circuits the setting of these will affect the grid current reading; consequently, it provides no accurate check on crystal activity. If the oscillator of Fig. 1 is used before etching is done it will serve as a reference for judging future activity of the crystal. The frequency of oscillation is not necessarily the same as it will be when the crystal is used in the station transmitter, however, so the activity-checking oscillator should not be relied on completely if you are etching for a precise frequency.

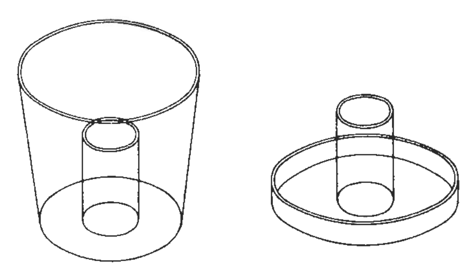


FIG. 2—The two-container method for handling the etching solution. The small plastic container is cemented inside the larger, or to its cover, to prevent spilling.

Now we need a teaspoon of ammonium bifluoride salt and some special containers and tools for handling it and the crystals. When the solution is made it becomes dangerous to handle and requires special precautions. It attacks glass, as well as crystal quartz, so it must be used and stored in plastic containers, and handled with plastic tools. The ammonium bifluoride may be obtained from a chemical supply house, or in some cases from drugstores.

Safe methods for handling the solution and crystals are shown in Fig. 2. A small plastic container of the type used for pill dispensing in drug-stores holds the solution. This container is cemented inside a larger one, which may be the sort often used for packaging various kinds of food. Use plastic cement or acetone. As an alternate method, the small container may be cemented inside the cover of the larger one, and the main part of the large container used as a cover. It can then also serve as a water dish, to be used in the washing process. Either way we have a method that involves the use of only a very small amount of the solution, and provision for storing it in a way that will make spilling unlikely.[1]

About one teaspoon of the ammonium bifluoride salt should be placed in the small container, which is then filled about three-quarters full with water. Let the solution stand, or stir it occasionally with a plastic rod, until the crystals are completely dissolved.

The crystals should be handled with plastic tweezers, which are probably available commercially, but we made our own as shown in Fig. 3. A plastic rod slit down the center will

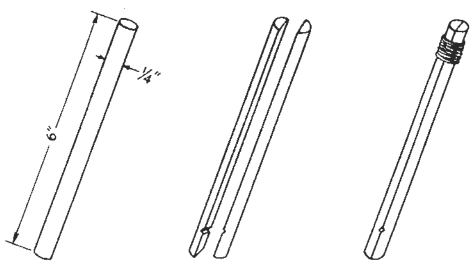


FIG. 3—Plastic tweezers of the normally-closed type for handling the crystals may be made readily from plastic rod.

do very nicely. The tweezers will be the normally-closed type, tension being maintained by wrapping the top ends together with a rubber band. The inner surfaces should be filed to form a notch, the and the crystal is held by the edges, inserted in this notch. Immersion of the crystal in the solution in the tweezers allows the solution to reach both surfaces equally and eliminates probing for the crystal in the solution when you want to remove it.

A container of water for washing and a clean cloth for drying are also needed. Paper tissues may be used, provided that they are the plain untreated type. The final tools required are your receiver and some method of checking frequency accurately. Now let's go to work.

Etching and Checking

Select a crystal that will require a large change in frequency for the initial try. First check its activity and frequency of oscillation. Now remove the crystal from the holder and place it in the tweezer slots. Immerse the crystal (in tweezers) in the solution for exactly one minute. Remove it from the solution and place it in the water container at once. Agitate for fifteen seconds or so. Now, slide the crystal onto a clean cloth and wipe it dry, being careful not to touch the crystal with the fingers. When the crystal is completely dry, slide it back into the holder and recheck the frequency and activity. Plot a graph of frequency and etching time, as shown in Fig. 4, increasing the time in the solution until a satisfactory curve has been developed.

The solution strength and the surface finish of the crystal will affect the etching rate. If the rate is too fast or slow, add more water or salt, as required. After three years of use, the author's original solution has a nonlinear etching rate of approximately 100 kc. in 10 hours. In the event that a decrease in crystal activity is observed, try cleaning the pressure plates and holder.

Remember always that you are handling a dangerous solution. Arrange the working area so that there is no possibility of spilling. Should the solution come in contact with the skin, wash it off at once with plenty of water. The small amount of solution required with this method makes safe disposal easy. If you want to discard the solution (even though it apparently can be kept and used indefinitely) pour it down the drain and flush with plenty of water at once.

[1] An alternative method is suggested by WØZJY. He uses a container of such size that the crystals will not lie flat in it. He drills holes in the bottom of this container and uses it as a basket to immerse the crystals in the solution. The container and crystals are removed and inserted in water for washing. More solution is needed in this way, but some workers may prefer it to the tweezer method.—*Ed.*

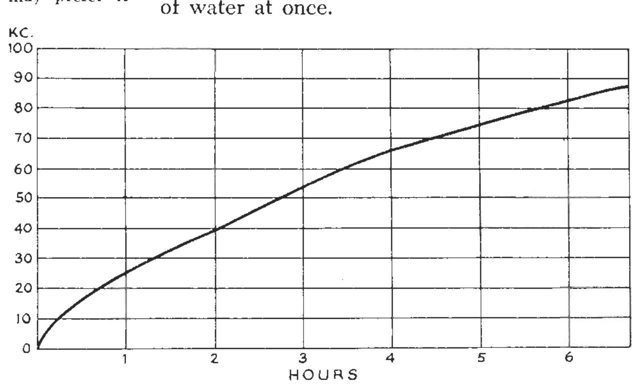
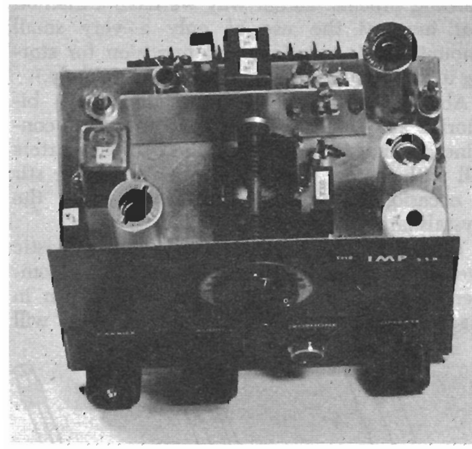


FIG. 4—Typical curve for etching solution. Frequency change is plotted against time, as a guide for future work.

» When a single-sideband generator is stripped to essentials, there isn't much to it; the complications pile on when assorted accessory equipment is added. Here's a basic unit that will get you off to a good start on s.s.b. Built mostly from odds and ends of surplus, including the crystals, it doesn't leave much room for argument on the question of economy.



The "Imp" uses a simple crystal filter and VXO frequency control to put a single-sideband signal on the 14-Mc. band. The 5×7-inch chassis shown in this photograph contains the entire r.f. and audio circuits of the exciter. Output from the 6CL6 amplifier is about 1 watt.

The "Imp"—a 3-Tube Filter Rig

JOSEPH S. GALESKI, JR., W4IMP

The desire to do a little experimenting with high-frequency crystal filters and VXOs, prompted the development of the "Imp"; I needed an exciter with a minimum number of tubes to use as a laboratory for my experimentation.

The results have been most encouraging. The three tubes and filter generate a very acceptable s.s.b. signal, with variable frequency and a watt or so of output to drive a linear amplifier.

For purposes of simplification this exciter is designed to operate only on 20 meters. However, by the proper choice of filter frequency, VXO crystal, and suitable modification of the three coils it can be made for any band. Components are readily obtainable on the surplus market and substitutions are quite in order where necessary. I was able to purchase crystals for less than twenty-five cents each. The modulation transformer can be any small plate-to-line unit with a turns ratio of about six or eight to one, such as the output transformer from an ARC-5 receiver.

Every effort has been made to keep circuits simple and with as few parts as possible. These circuits are not original with me and complete descriptions can be found in the handbooks.

Circuit and Construction

The triode section of V_1, Fig. 1, is used as an untuned crystal oscillator to feed carrier to the diode balanced modulator. The pentode section of this same tube will deliver enough audio from a crystal microphone to upset the modulator balance and furnish a double-sideband signal to the filter, which passes only the upper sideband to the triode mixer, V_{2A}. The pentode section, V_{2B}, is a variable-frequency crystal oscillator which supplies the mixing signal to the grid of V_{2A}. About 10- to 12-kc. shift can be expected from an 8-Mc. crystal. The 6CL6 amplifier, V_3, uses tuned tanks in both the grid and plate circuits to provide adequate selectivity.

Construction is straightforward. A 5×7-inch chassis was used, with the filter mounted on top. A shield separates it from the VXO tuning capacitor. A reasonable effort should be made to keep the circuits separated. If the unit is not to be put in a metal box, I would suggest putting a shield can over the carrier crystal and over the filter, because hand capacitance tends to throw the carrier balance out of kilter.

The selection of crystals for the filter permits a wide latitude of frequencies. However, the harmonics of the filter frequency and of the mixing frequency should be well removed from the desired 20-meter output.

Selecting Crystals

On the surplus market are several groups of 5- to 9-Mc. crystals that have a frequency difference of 1.7 kc. I obtained about ten at

From May, 1960, *QST*.

5773.3 and ten more at 5775 for experimenting, but I now feel that for a similar project seven at 5773.3 and three at 5775 would be enough. While the crystals are marked as having these frequencies few of them are "on the nose," and you will find that they will differ from one another by as much as a kilocycle.

Mark each of the 5773.3 crystals with an identifying letter and determine the *relative* frequency of each by inserting them one at a time in the crystal socket of V_1 and tuning them in on your receiver. If your receiver covers only the ham bands, use a second crystal at approximately 8500 kc. in the VXO to bring the sum frequency to the 20-meter band. A difference in audio tone against the receiver b.f.o. will permit you to get the crystals in order of frequency from highest to lowest. Record this order by the letters previously marked on them.

Select two of the lower-frequency crystals of the 5773.3-kc. group having a separation of a couple of hundred cycles or so and call the lower one Y_4 and the higher Y_2. You will later use one of the remaining crystals of this group for Y_1. Use one 5775-kc. crystal for Y_3. Peak T_1 and the trimmers on T_2 with a 5775-kc. crystal at Y_1.

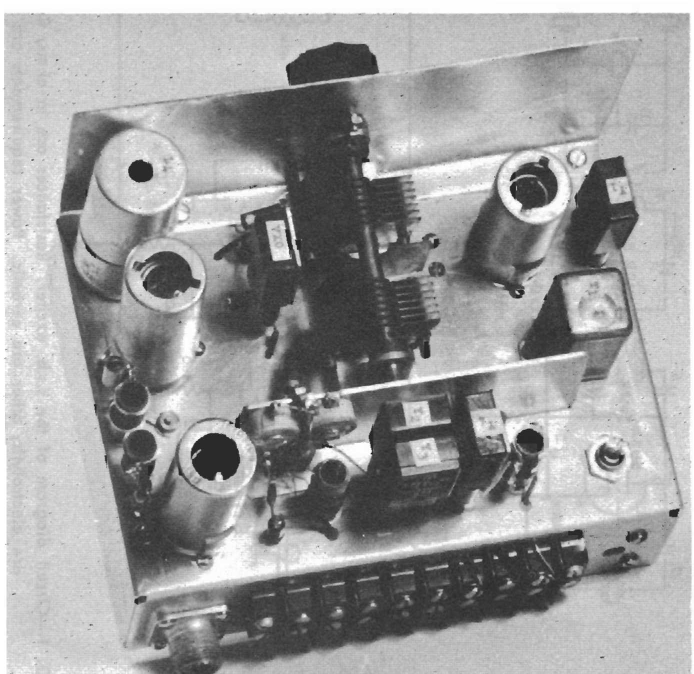
Circuit and Filter Alignment

The three tuned circuits, L_2, L_3, and L_4, can best be aligned by first removing both Y_1 and the VXO crystal and then, with a signal generator set at 14,300 kc. connected to the grid of V_{2A}, peaking the coils. An alternate method would be to use a 7150-kc. crystal in the VXO and peak the coils on its second harmonic. This procedure should be followed to avoid the possibility of alignment of the coils on a harmonic of the VXO or a harmonic of Y_1.

Alignment of the filter is the next step, and a BC-221 frequency meter or other slow-tuning oscillator is necessary. I used a 221 on its low range, which gives approximately 30 dial divisions per kilocycle. Insert a crystal about 150 to 225 kc. lower than the passband frequency at Y_1; this would be in the 5550- to 5625-kc. range. Exact frequency matters little as long as the 221 output and the temporary Y_1 add to tune across the filter passband. A difference frequency may also be used if you remember that in such case *increasing* the 221 frequency *decreases* the resultant frequency.

Connect a capacitance of a few pf. between the output terminal of the 221 and a shielded lead running to the arm of the carrier-balance potentiometer, R_2, which should be turned to one end of its rotation. Remove the 6CL6 from its socket and connect a lead from the ungrounded end of L_3 to your receiver antenna terminal. You should be able to get an S-meter reading on the 20-meter band. If the meter goes off scale, loosen the coupling between the Imp and the receiver until a mid-range reading is obtained.

Tune the 221 so that the output frequency of the diode balanced modulator, which is now acting as a diode mixer, sweeps across the filter passband. Keep the receiver in tune with the signal and observe the action of the S meter. It takes a little practice, but after a



Behind the panel. Most of the parts are from surplus. L_4 is in the can (from a roll of film) at the upper left. Following down along the left edge of the chassis are the output tube, V_3, the mixer-amplifier coupling coils, L_2L_3, and the mixer-VXO tube, V_2. The VXO crystal is alongside the tuning capacitor, which is 100 pf. per section with 100 pf. fixed in series with each section to give the 50 pf. specified in Fig. 1. T_2 is on the coil form at the left near the rear edge of the chassis; its associated trimmers, C_2 and C_3, are mounted on the shield alongside. The filter crystals and T_1 are also near the rear edge of the chassis. The carrier crystal is at the right in the far corner; V_1 is alongside, followed by the audio transformer, T_1, and, in the lower right-hand corner, the carrier balance control, R_2.

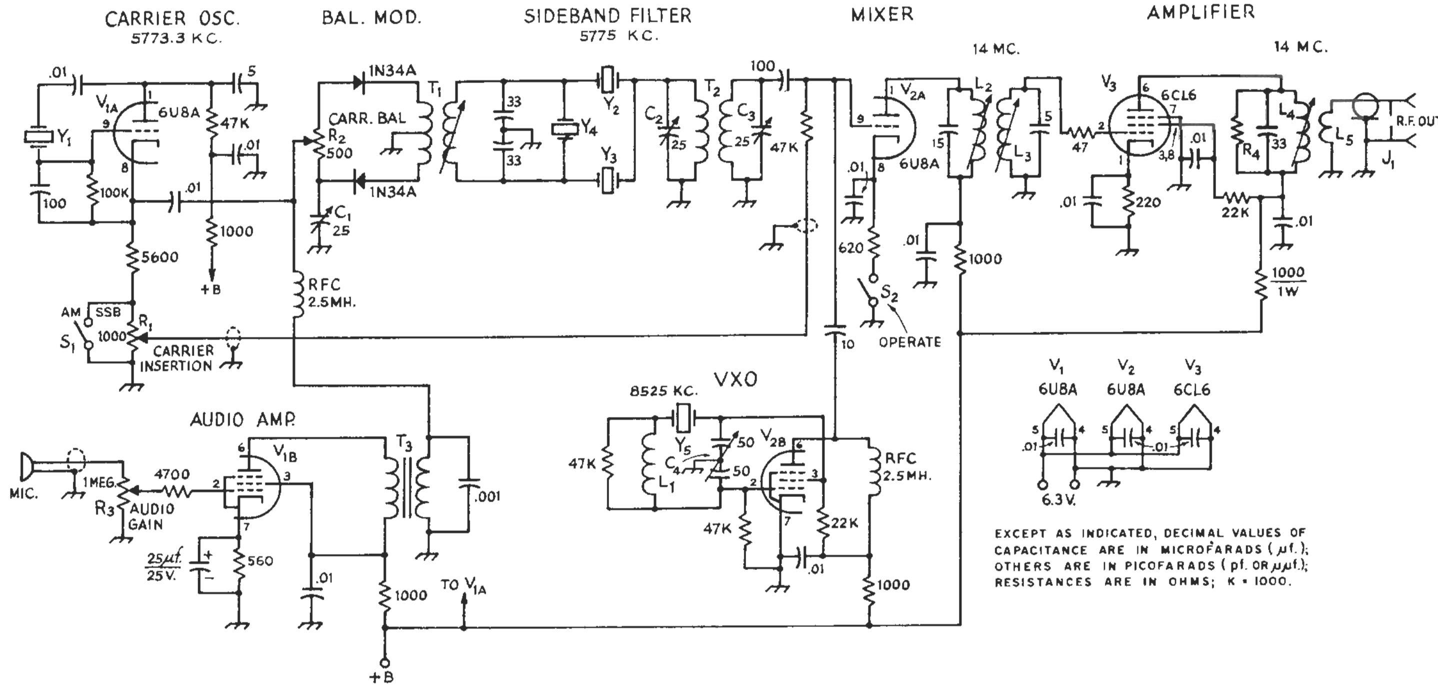


FIG. 1—Circuit diagram of the s.s.b. exciter. Resistances are in ohms; fixed composition resistors are ½ watt except as indicated. Fixed capacitors with polarities marked are electrolytic; other are ceramic. Power requirements are 6.3 volts at 1.6 amp. for tube heaters and 250 to 300 volts at 50 ma. for plates.

C_1, C_2, C_3—4.5-25 pf ceramic trimmer (Centralab 822-AZ).

C_4—50 pf. per section (Hammarlund MCD-50-M).

J_1—Coax connector, chassis mounting.

L_1—22 turns No. 22 enam. close-wound on 1½-inch diam. form. Modify as necessary to give desired VXO frequency shift.

L_2, L_3—22 turns No. 22 enam. close-wound on ⅜-inch diam. slug-tuned form. L_2 and L_3 mounted side by side with ¾-inch spacing, center to center.

L_4—20 turns No. 22 enam. close-wound on ½-inch diam. tuned form.

L_5—5 turns same as L_4, wound at cold end of L_4.

R_1—1000-ohm potentiometer, linear taper.

R_2—500-ohm potentiometer, linear taper.

R_3—1-megohm control, audio taper.

R_4—25,000 to 50,000 ohms, 2 watts, as needed for swamping and for stabilizing the 6CL6 amplifier.

S_1—S.p.s.t. mounted on R_1.

S_2—Rotary, single-throw, with additional poles as needed for controlling external circuits.

T_1—Tuned winding: 60 turns No. 28 enam. scramble-wound to length of 7/16 inch on ⅜-inch diam. slug-tuned form. Primary winding: 8 bifilar turns on same form close to tuned winding.

T_2—Each winding 50 turns No. 28 enam. scramble-wound to length of ⅜ inch on ⅜-inch form (no slug); windings spaced 3/16 inch between adjacent ends.

T_3—Plate-to-line audio transformer, approx. 20,000 ohms to 500–600 ohms (Stancor A-3250, ARC-5 receiver output, or similar).

Y_1, Y_2, Y_4—5773.3 kc., surplus FT-243 type (see text).

Y_3—5775 kc., surplus FT-243 type (see text).

Y_5—8525 kc., surplus FT-243 type.

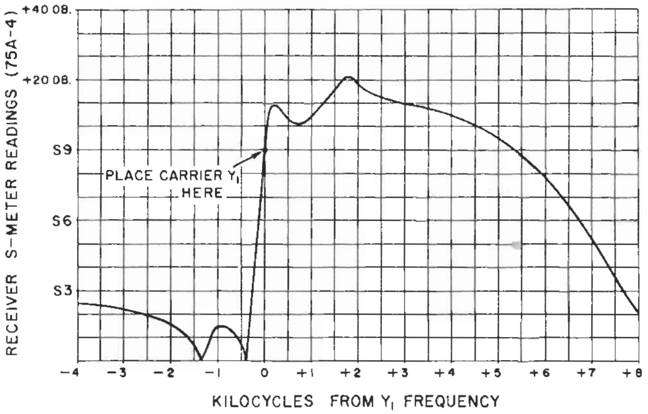


FIG. 2—Pass-band of crystal filter use in the Imp, in terms of S-meter readings on the 75A-4 receiver used by W4IMP. The frequency measurements were made by using a BC-221 frequency meter as a signal source.

few moments of using one hand on the receiver and one hand on the frequency meter this process becomes quite easy. You should be able to observe a definite increase in S-meter readings within the passband and a decreased reading outside of the passband.

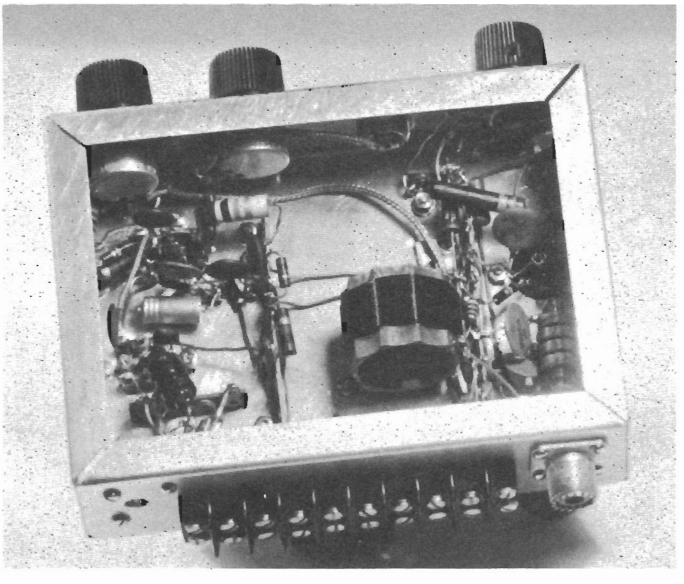
Using a sheet of graph paper, plot the S-meter readings on the vertical scale against 500-cycle dial settings from the 221 calibration book on the horizontal scale. Run a series of points and sketch in the curve. After you have plotted one or two of these curves you will be able to visualize what happens to the passband by watching the S-meter action after each adjustment of the filter trimmers. It will only be necessary to plot the final curve for your records.

The filter passband of the Imp is shown in Fig. 2. It has a very sharp cutoff on the low-frequency side and is suitable as a filter for the upper sideband for transmission, but is too wide for receiving purposes. The curve has a dip and a bump or so, but they do not seem to affect the speech quality too adversely. Final filter adjustment will be a compromise between flatness of passband and maximum suppression of the unwanted sideband.

Carrier Balance

There should be little trouble with the carrier balance. If the trimmer, C_1, does not add to the carrier suppression that can be obtained by adjusting R_2, connect it at the other diode. This is a matter of cut and try. You will find that different crystals at Y_1 require different settings of R_2 and C_1. Any r.f. indicator, such as an r.f. probe and v.t.v.m. or a receiver S meter, can be used for setting the balance. Be sure S_1 is closed.

Selecting Y_1 is also a bit of cut and try. If



The large coil is L_1, in the VXO circuit. Knob-adjusted controls are, left to right, carrier insertion, audio gain, and operate switch. The microphone jack is between the latter two. The extra contacts of the operate switch, S_2, are brought out to the terminal strip on the rear edge of the chassis. These can be tied in with a linear amplifier and other accessory equipment as the operator may desire.

its frequency is too low you will find that the sideband suppression is excellent, but the signal is difficult to copy because the low voice frequencies are cut off by the filter. If it is too high, the signal will sound fine, but you've lost suppression of the unwanted sideband. Don't be afraid to move the frequency around a bit by loading the crystal with a pencil mark. The final frequency of Y_1 should be as low as possible consistent with good voice quality.

Other Bands

Operation on other bands may be accomplished by using this same filter. For example, *lower*-sideband output at the high-frequency end of the 75-meter band can be realized by (1) replacing the VXO r.f. plate choke with a parallel-tuned circuit at 9760, (2) using a 4880-kc. VXO crystal, and (3) changing L_2, L_3, and L_4 to tune to 3980 kc. You could leave the plate choke alone and obtain a fundamental crystal at about 9760 kc.

I made an attempt at 15 meters using a 7825-kc. crystal, doubling in the VXO tank to 15,650 to give exciter output at about 21,423. It worked fine except that L_2, L_3, and L_4 did not give sufficient selectivity for adequate attenuation of the third harmonic of 7825 kc. Construction of a filter at about 4125 will permit using an 8650 crystal for better rejection of harmonics in the tuned circuits.

Afterthoughts

I would like to say here for the benefit of those without access to a BC-221 that they should not lose heart. Any existing v.f.o. can be used if it is given additional bandspread with a trimmer so that a 180-degree turn of the dial will cover about 10 kc. It doesn't even have to tune the filter frequency. Use the heterodyne principle as described above with the BC-221. After all, in this case we want to know only that the passband has the desired shape. A VXO on a separate chassis could also be used.

Since only one crystal, Y_3, is needed for the higher channel, all filter crystals may be purchased for the same frequency and a couple etched or ground up 1½ to 2 kc. This job is easier to do than one can imagine.

» Here is a complete multiband transmitter, power supply, VOX and all. Countless amateurs have built it since its presentation in QST some years ago, and various of its circuit ideas have been picked up by others for incorporation in their own designs.

A Sideband Package

GEORGE K. BIGLER, W6TEU

When construction of this exciter was started, it was hoped it would have a few features that previously described units lacked. Some of the features which appeared desirable, after building several smaller rigs, were:

1. Bandswitching, all-band c.w.-a.m.-s.s.b. operation
2. Sideband selection without carrier shift
3. Voice control with a loudspeaker
4. Peak limiter
5. Ample driving power for a kilowatt final
6. Good frequency calibration
7. Complete self-contained unit.

All of these features are included in the exciter to be described.

Construction and alignment should not be very difficult or time-consuming for a ham with a reasonable amount of construction experience.

Circuit Description

Referring to the block diagram in Fig. 1, a crystal oscillator, V_{1A}, is used to feed a cathode follower, V_{1B}, which drives the diode balanced modulator for carrier suppression. The double-sideband signal is fed through the two-section crystal filter, where the lower sideband is rejected. The remaining upper sideband is amplified by V_2 and fed to the mixer, V_3. The unmodulated 450-kc. carrier signal from the cathode follower is also fed to the grid of the frequency doubler, V_{4A}, doubled to 900 kc. and fed to V_{4B}. V_{4B} is used as either a doubler to 1.8 Mc. or as a tripler to 2.7 Mc., depending on the position of the sideband selector switch. Thus the output of V_{4B} is either four or six times the crystal oscillator frequency. When the fourth harmonic from V_{4B} is mixed with the sideband signal in V_3, the frequency sum can be taken at the output of the mixer to give an upper sideband signal at five times the crystal oscillator frequency. When the sixth harmonic is mixed with the sideband signal, the frequency difference can be taken to give the same resultant output frequency with the opposite sideband.

The output of the sideband generator chassis is therefore at a fixed frequency of five times the original frequency, but with sideband selection available. This same principle can be applied to a fixed oscillator, such as the b.f.o. in a receiver, for a selectable-sideband system.

The sideband signal at approximately 2.25 Mc. is fed to V_{101}, where it is mixed with the v.f.o. running from 5.25 to 6.25 Mc., to give a frequency difference output from 3.0 to 4.0 Mc.

To deliver a clean signal to the final mixer, a double tuned circuit is used, which is gang tuned with the v.f.o. The final mixer is used to convert the 3.0- to 4.0-Mc. signal to the desired band. By choosing the crystals for final conversion so that 3.0 Mc. goes to the low even megacycle end of each band, a single dial calibration from 0 to 1.0 Mc. can be used on all bands. It is only necessary to mentally add the megacycles of the band in use to the dial reading to get the frequency of operation. By putting the conversion oscillator on the low side in each case, sideband reversal is eliminated. Harmonics of the 3.0- to 4.0-Mc. signal in the higher bands will fall outside the pass band where they are easily suppressed in the tuned circuits. On 10 meters, the band is divided into two sections, 28.0 to 29.0 and 29.0 to 30.0 Mc.

After final conversion the signal is amplified by V_{105} and V_{106} to about 30 watts peak output. Some of the r.f. output from the final stage is fed to the peak limiter V_{201}. The d.c. voltage developed when a preset peak level is exceeded is fed back as a control voltage to V_2, where the remote cut-off characteristic of the tube allows the gain of the stage to be modified.

V_{301} is used as a two-stage speech amplifier to feed the balanced modulator. V_{302} and V_{303}, in conjunction with a crystal diode, furnish voice control with a loudspeaker.

Construction and Alignment of Sideband Generator

The sideband generator is constructed on a 5 × 9 × 1½-inch chassis. Before construction is started the i.f. transformers can be modified as mentioned in Fig. 2. Transformer T_1 for the balanced modulator is altered as follows: Remove all the wire from the winding on the free end of the dowel. Double enough of this wire on itself to make a 25-turn jumble-wound bifilar winding ¼ inch from the remaining winding. Join the finish of one 25-turn winding to the start of the other and ground this junction to the bolt at the top of the transformer which will pass through the hole in the can. The remaining ends are soldered to the unused trimmer terminals for tie-points. Apply coil dope and reassemble.

After construction of the unit is completed,

From *QST*, June, 1958.

the multipliers should be aligned first. This can best be done with an r.f. probe on a v.t.v.m.

Tune T_6 to the second harmonic of the l.f. oscillator, about 900 kc., and then tune T_7 to 2.7 Mc. with S_1 open. Close S_1 and tune the two 10-100 pf. trimmers (mounted on the chassis) to resonate T_7 to 1.8 Mc. The output delivered to the mixer should measure about 12 to 15 volts in both positions of S_1. Next, insert carrier by advancing the 5000-ohm control and then peak the grid winding of T_4, indicating resonance by measuring the r.f. at a grid of V_3. With the r.f. probe on J_1, tune T_5 to 2.25 Mc. Several peaks will be noted, but the correct one will be obtained when switching S_1 results in little output change and turning down the carrier control reduces the signal to nearly zero. The output should run about 1½ to 2 volts with full carrier insertion, measured across L_{101} (Fig. 4).

Alignment of the crystal filter is next. Referring to Fig. 3, it is seen that the filter has a very sharp cut-off near the carrier frequency due to the shunt crystals Y_3 and Y_6. Since the carrier is always on the same side of the filter, the characteristic can be shaped for better attenuation near the carrier frequency.

In selecting crystals for the filter, an adequate supply should be obtained. At a dime apiece this shouldn't break the bank and will save time in alignment. The exact channels named in Fig. 2 are not required, but they should be adjacent and the lower frequency should not be below 440 kc. This will assure adequate 2nd-harmonic rejection of the selectable sideband signal output in the 3- to 4-Mc. channel when the dial is near the 4-Mc. end.

A preliminary check on the crystals can be made by trying them for oscillation in the oscillator. Those that oscillate can be further checked for frequency of the peaks by connecting them between the antenna post of a BC-221 frequency meter and an r.f. probe of a v.t.v.m. Pair the crystals according to the series peaks. Only two upper-channel crystals are required, and by choosing the highest-frequency pair the pass band will be broadest. The lower-channel crystal that oscillates at the lowest frequency will probably be best for the oscillator, the next higher pair for the shunt crystals Y_3 and Y_6, and the next higher pair for the series crystals Y_2 and Y_5.

The filter is aligned as follows: With the power off the unit, connect the BC-221 output terminal to the output winding of T_1 through a 5-pf. capacitor. With the BC-221 set in the center of the pass band (455 kc.) peak the circuit with the r.f. probe on the transformer terminal. Move the leads and repeat for the windings of T_2, T_3 and the plate winding of T_4. Remove the crystal from the oscillator and connect the BC-221 output to the crystal socket.

Remove the 6X8 multiplier, V_4, and temporarily ground the amplifier grid resistor at (E) and the audio feed to the diode modulator at (A). With the balanced modulator unbalanced by turning the arm of the 500-ohm

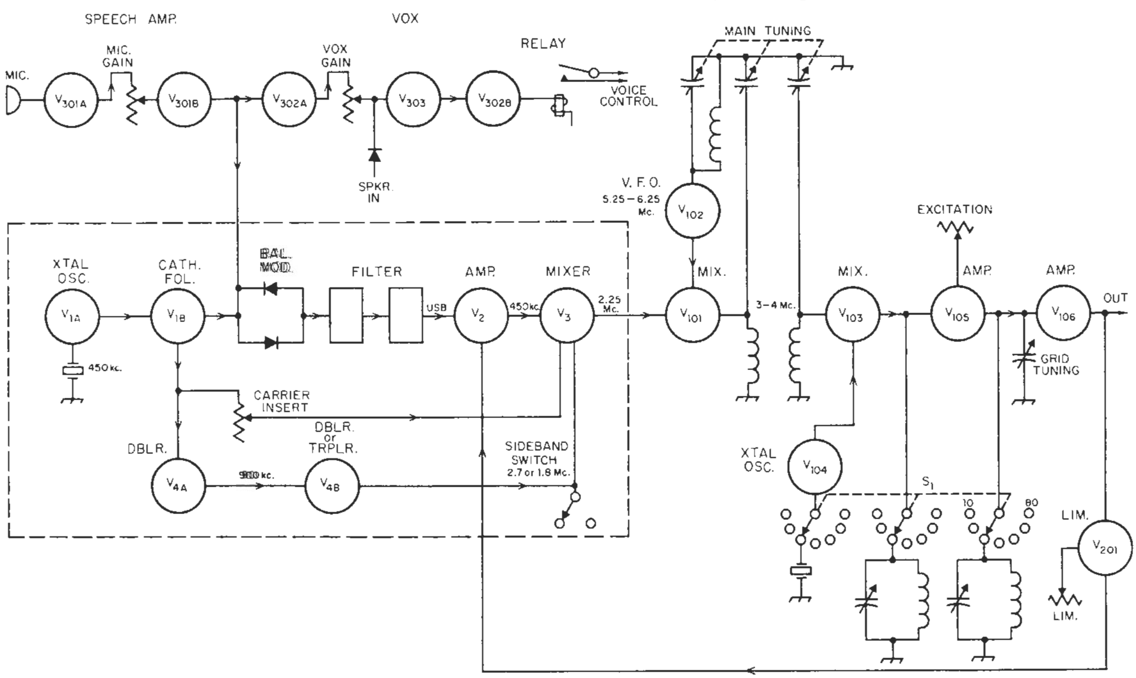
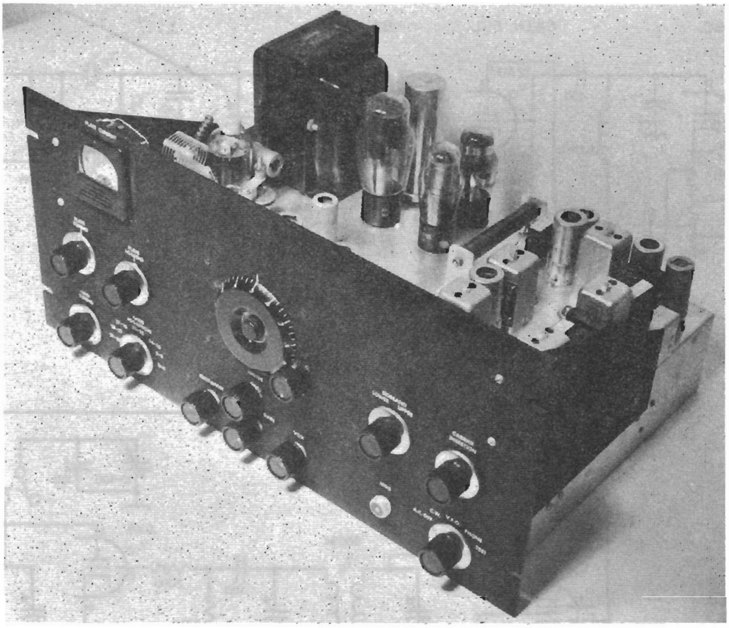


FIG. 1—Block diagram of the sideband exciter. The sideband generator proper is enclosed by dashed lines; this section of the exciter is built on a separate chassis. Sideband selection is obtained by using either the 4th or 6th harmonic of the low-frequency oscillator to convert the signal to 2.25 Mc.



This complete sideband transmitter uses a filter made from inexpensive low-frequency crystals. The frequency control utilizes the tuning gang from an ARC-5 receiver. The dial has been recalibrated to read 0 to 1.0 Mc.

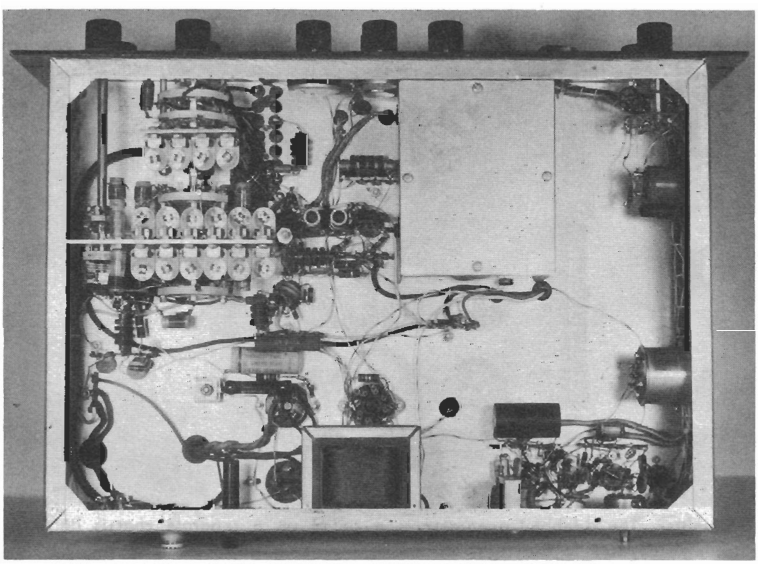
potentiometer to one end, apply power and connect the r.f. probe to the plate of V_2. Peak all trimmers in the filter.

The passband and sideband rejection of the filter can be checked by moving the BC-221 across the filter frequencies. By small trimmer adjustments and rechecking, a suitable curve should be obtained. The curve of this particular unit is shown in Fig. 3. In constructing five such filters, similar curves have been obtained in each case without resorting to crystal grinding or overly-tedious selection.

After filter alignment is complete, replace the 6X8, the oscillator crystal, and unground point (A). With an audio oscillator fed to the balanced modulator through an ARC-5 receiver output transformer, which will be mounted later on the main chassis, retune the T_4 windings with the r.f. probe connected to the output winding of T_5 at J_1.

The pass band can be checked, which will give an opportunity to judge the relative position of the crystal oscillator frequency. Modification of the 10-pf. capacitor at the plate of V_{1A} may be necessary to move the oscillator to the right frequency.

With the audio input at zero, balance the 500-ohm potentiometer for minimum carrier. The 50-300 pf. mica trimmer should aid the balance. If not, connect it to the other side of



The v.f.o. coil L_{102}, is housed in its own shield section (upper right of center) and the two coils L_{103} and L_{104} are mounted nearby at the left. Output jack, fuse, J_{401}, J_{402} and the bias potentiometer are mounted on rear apron of chassis.

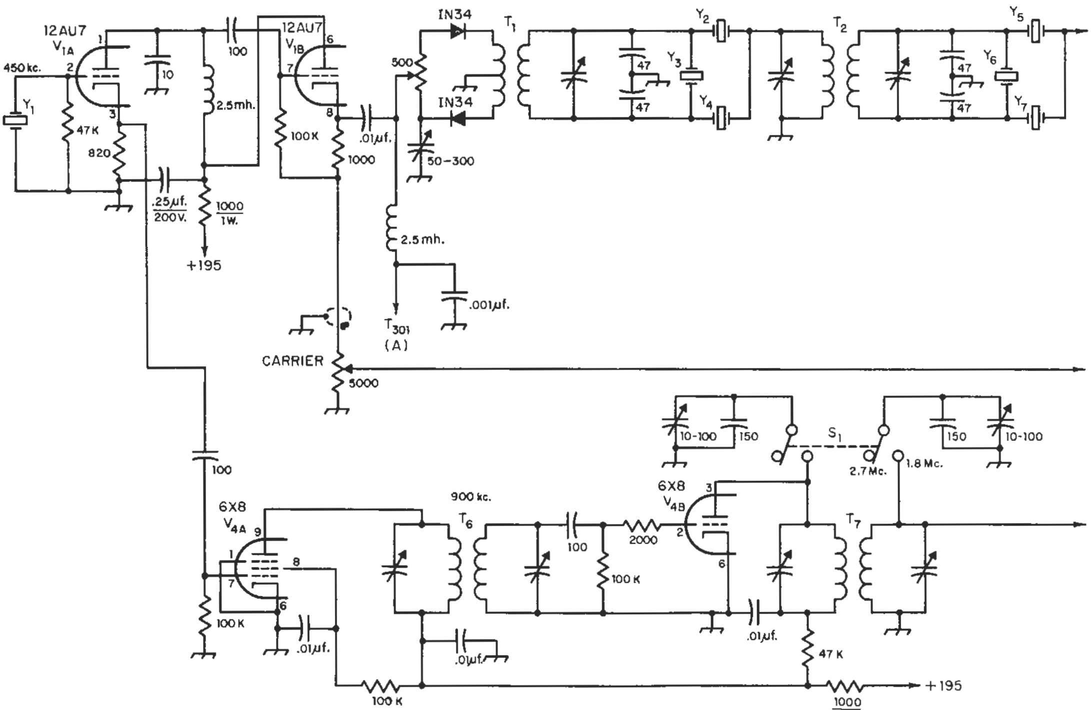
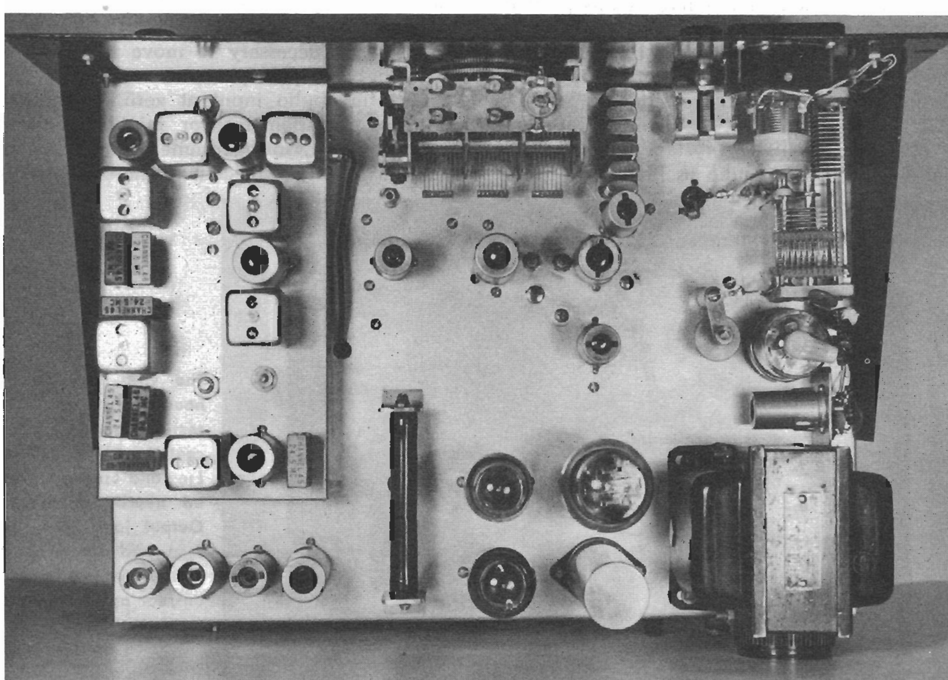


FIG. 2—(See facing page.)



Top view of the sideband transmitter. The sideband generator is mounted on a separate chassis (left) for better shielding and easy testing. Crystals to the right of the tuning gang (top center) heterodyne the signal to the output frequency. Speech and VOX at lower left.

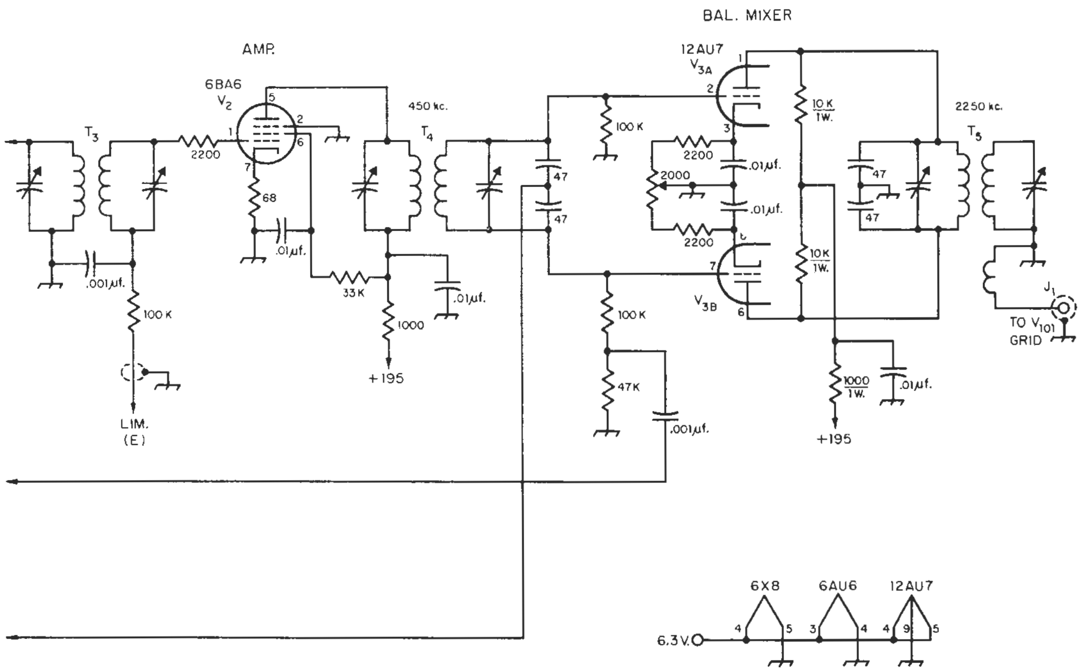


FIG. 2—Circuit diagram of the sideband generator section. Unless otherwise indicated, capacitances are in pf. resistances are ohms, resistors are ½ watt.

S_1—D.p.d.t. rotary switch.

T_1—Modified 455-kc. i.f. transformer. See text. (Miller 112C1).

T_2, T_3, T_4—455-kc. i.f. transformer (Miller 112C1).

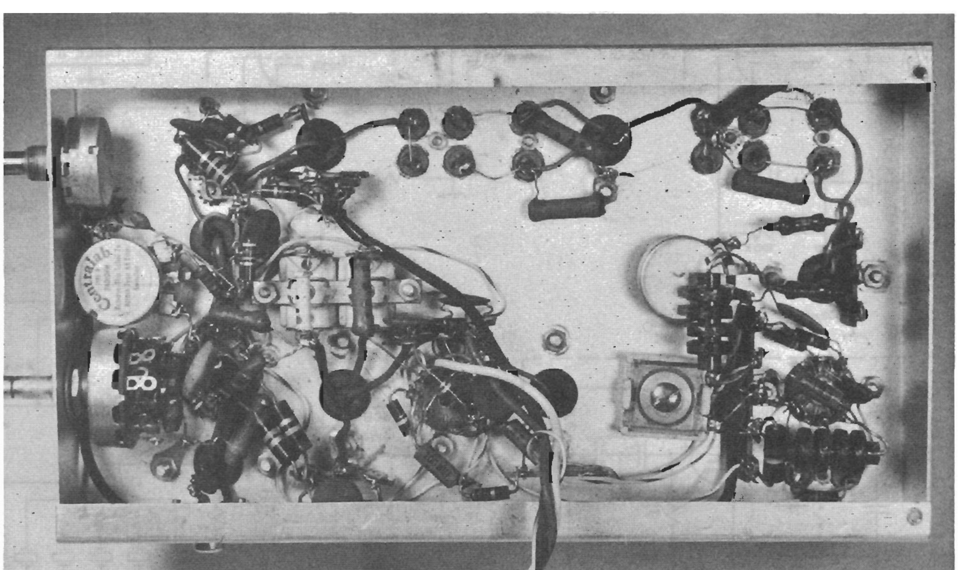
T_5—2.25-Mc. transformer, made by removing 8½ feet of wire from each winding of 1500 kc. transformer (Miller 112W1) Coupling coil is 4 turns over rim of secondary coil.

T_6—900-kc. i.f. transformer, made by removing 28 feet of wire from each winding of 455-kc. transformer (Miller 112C1).

T_7—2.7-Mc. transformer, made by removing 9 feet of wire from each coil of 1500-kc. transformer (Miller 112W1).

Y_1, Y_2, Y_3, Y_5, Y_6—453.7-kc. crystal. Surplus, marked "Channel 45, 24.5 Mc."

Y_4, Y_7—455.6-kc. crystal. Surplus, marked "Channel 46, 24.6 Mc."



Bottom view of the sideband generator section. The panel controls are sideband selection switch and corrier insertion potentiometer.

the potentiometer. Mixer balance is obtained with the 2000-ohm potentiometer at V_3, but this is not a critical adjustment.

Main Chassis Construction

Before laying out the main chassis it will be easier to build the bandswitch assembly. This is built in two parts: first the conversion oscillator and second the mixer/amplifier.

The oscillator section is built using three Centralab GGD and two Centralab YD switch sections assembled on a P-272 index assembly. A 1¾ by 2¾-inch aluminum plate is mounted on the rear using the switch assembly bolts. This plate holds the four trimmers. The switch shaft is cut off behind the assembly so that a metal ¼-inch shaft coupling can be attached for coupling to the mixer/amplifier section. This facilitates removal of the mixer/amplifier switch assembly.

Referring to Fig. 4, S_{1B}, S_{1C} and S_{1D} are the oscillator switching sections. Since the crystal oscillator is a screen feedback type, when no output circuit is connected, fundamental output will result. This is used for conversion to 40 and 20 meters with crystal frequencies of 4.0 and 11.0 Mc. respectively. Converting to the higher bands requires an output circuit tuned to the second harmonic of the crystal.

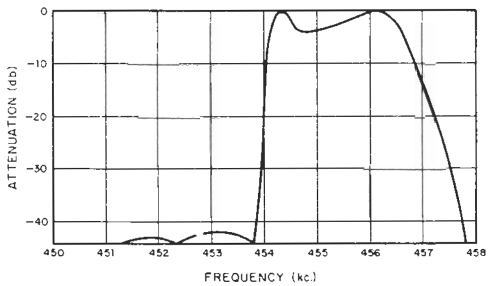
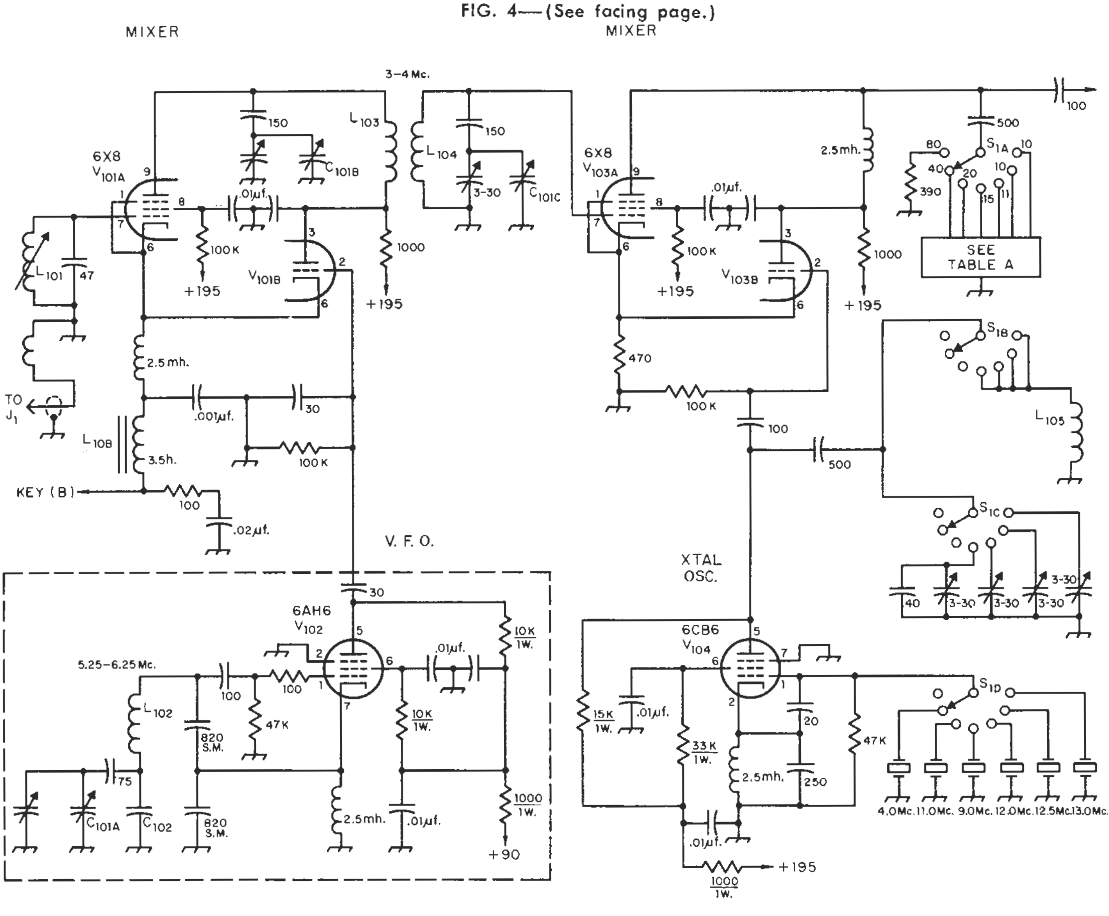


FIG. 3—Pass band of crystal filter after correct adjustment.

Therefore, the coil L_{105} is switched in by S_{1B} while S_{1C} connects a trimmer on each of the upper bands for resonating L_{105} to 18.0, 24.0, 25.0, and 26.0 Mc. These convert the 3.0-Mc. sideband signal to 21.0, 27.0, 28.0, and 29.0 Mc. (The 24-Mc. crystal used for the former 11-meter band, may be omitted.—*Ed.*)

The mixer/amplifier switch assembly is built on a 3 × 6 inch L-shaped bracket with a ½-inch mounting foot, as shown in one of the photographs. One GGD switch section is mounted on each side of the plate. The amplifier, or rear side is shown. The coils are mounted by threading them on a 1-inch brass

FIG. 4—(See facing page.)



bolt with head removed, which passes through the plate. Before threading the coils on, a nut is placed on each side of the plate, with a soldering lug under the mixer side. The poly forms can be softened enough for easy threading by filling the hole in them with coil dope. For mechanical convenience, a 100-pf. capacitor is connected directly from the mixer plate to the amplifier grid. Then only one lead is required from the common switch terminal to the mixer plate, making easier installation possible. The same arrangement is used between the 6AK6 and the 6146.

All tuned circuits on the amplifier side are insulated from ground and bypassed by a 300-pf. capacitor, across which the neutralizing voltage for the 6AK6 and the 6146 is developed.

Since mixing in V_{103A} is not required on 80 meters, a resistor is used for the output load on this band. On 40 meters a series trap (see Table A) is used to shunt out the second harmonic of the 4.0-Mc. oscillator. To short out all unused coils, one of the spare switch terminals is wired across to the common side of the tuned circuits.

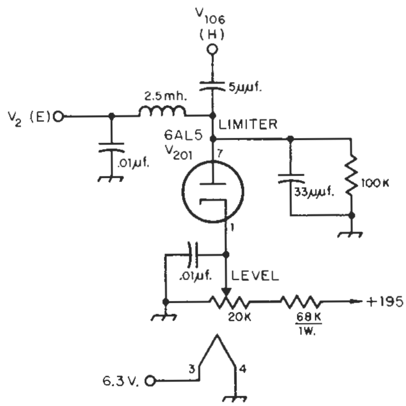


FIG. 5—The limiter circuits samples the r.f. output and reduces the r.f. gain if the signal exceeds the bias level.

After completion of the bandswitching units, they can be mounted on the main chassis. The piece of shaft cut from the oscillator section is slid through the mixer-amplifier section and

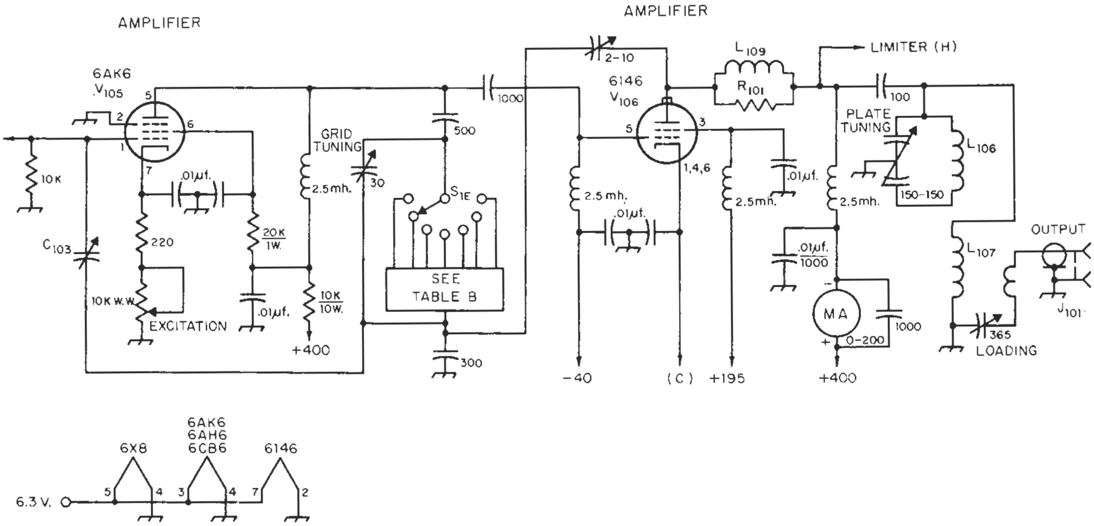


FIG. 4—Diagram of the r.f. circuit following the sideband generator of Fig. 2. Unless otherwise indicated, capacitances are in pf., resistances are in ohms, resistors are ½ watt.

C_{101}—Triple 150 pf. variable (from 3- to 6-Mc. ARC-5 receiver).

C_{102}—4.7 pf. N330 and 30 pf. NPO in parallel.

C_{103}—Small capacitor made by winding several turns of insulated wire around lead to pin of V_{105}. Adjust by changing number of turns.

L_{101}—80 turns No. 36 enam. on ⅜-inch diam. slug-tuned form (Miller 4400). Link is 3 turns wound over bottom end.

L_{102}—24 turns No. 26 enam. on 1-inch diam. threaded ceramic form, with half-turn loop for adjustment. See text. (National XR-60 with slug removed.)

L_{103}, L_{104}—40 turns No. 34 on ½-inch diam. slug-tuned form. (National XR-50). Spaced ¾-inch on centers.

L_{105}—17 turns No. 18 enam. on ¼-inch diam. form.

L_{106}—9 turns No. 18 wound 8 t.p.i. 1-inch diam. (B&W 3014).

L_{107}—21 turns No. 20 wound 16 t.p.i., 1-inch diam. B&W 3015). Link is 8 turns No. 18 enam. over cold end.

L_{108}—3.5-h. choke (No. 5634 from ARC-5 receiver.)

L_{109}—4 turns No. 18 wound on R_{101}.

R_{101}—100 ohms, 2 watts.

S_1—See text. S_{1A} and S_{1E} wired to short unused mixer and amplifier coils.

attached to the shaft coupling at the rear of the oscillator section.

The underside of the v.f.o. section is enclosed in a 3 × 4 × 5-inch box. The main tuning capacitor and diode modulation transformer were taken from a 3- to 6-Mc. ARC-5 receiver. V.f.o. coil L_{102} is made with an extra half-turn loop through the inside of the form, for fine adjustment of the inductance. The 2500-ohm dropping resistor in the power supply (Fig. 7) is mounted on top of the chassis, since it dissipates about 25 watts. New mounting brackets will protect the terminals from accidental contact. The limiter, V_{201}, is mounted on an L-shaped bracket between the 6146 and the power transformer, to prevent coupling the output circuit under the chassis.

Main Chassis Alignment

Attach the sideband generator chassis to the main chassis using sheet metal screws at the rear and the panel bushings at the front. With the band switch on 80 meters, carrier control full on, set the operation switch S_2 to v.f.o. The output winding of T_5 should be repeaked after connecting it to V_{101A} through the shielded lead. The output of the v.f.o. is reduced through a capacitance divider to about 1.4-2.0 volts, to prevent mixer overloading.

With the main tuning capacitor C_{101} set at a half turn from minimum, adjust the trimmer on C_{101A} until the mixer output is heard on a receiver at 4.0 Mc. Peak the other trimmers on C_{101} by connecting an r.f. probe to the cathode of V_{103}. It may be necessary to listen first in a receiver for the peak until enough signal is getting through the double-tuned circuit to indicate on the meter. Close capacitor C_{101} to three turns from maximum and tune the slugs of L_{103} and L_{104} for maximum output. Repeat until the circuits track with the v.f.o., as indicated by the r.f. probe on the cathode of V_{103}.

The v.f.o. tuned circuit values have been chosen to give a reasonably linear dial, with 4.0 Mc. falling ½ turn from minimum and 3.0 Mc. occurring about three turns from maximum. If a general-coverage receiver with a calibrator is available, the 3.0-Mc. end can be checked at this point; otherwise it can be checked indirectly later.

Table A—Mixer Coils

Each coil wound on 1½-inch long ⅜-inch diam. polystyrene rod.
Each coil shunted by 3–30-pf. trimmer.

Band	Coil
40 m.	33 turns No. 30 enam., shunted by trap of 80 turns No. 36 enam. on ¼-inch diam. slug-tuned form (Miller 4500) in series with 10 pf.
20 m.	24 turns No. 26 enam.
15 m.	12 turns No. 22 enam.
*11 m.	10 turns No. 18 enam.
10 m.	9 turns No. 18 enam.
10 m.	9 turns No. 18 enam.

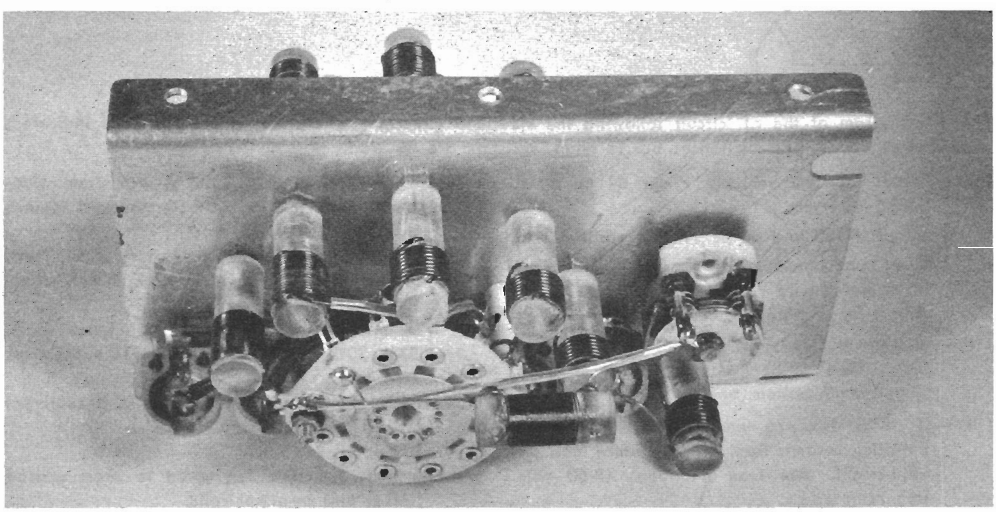
Table B—6AK6 Coils

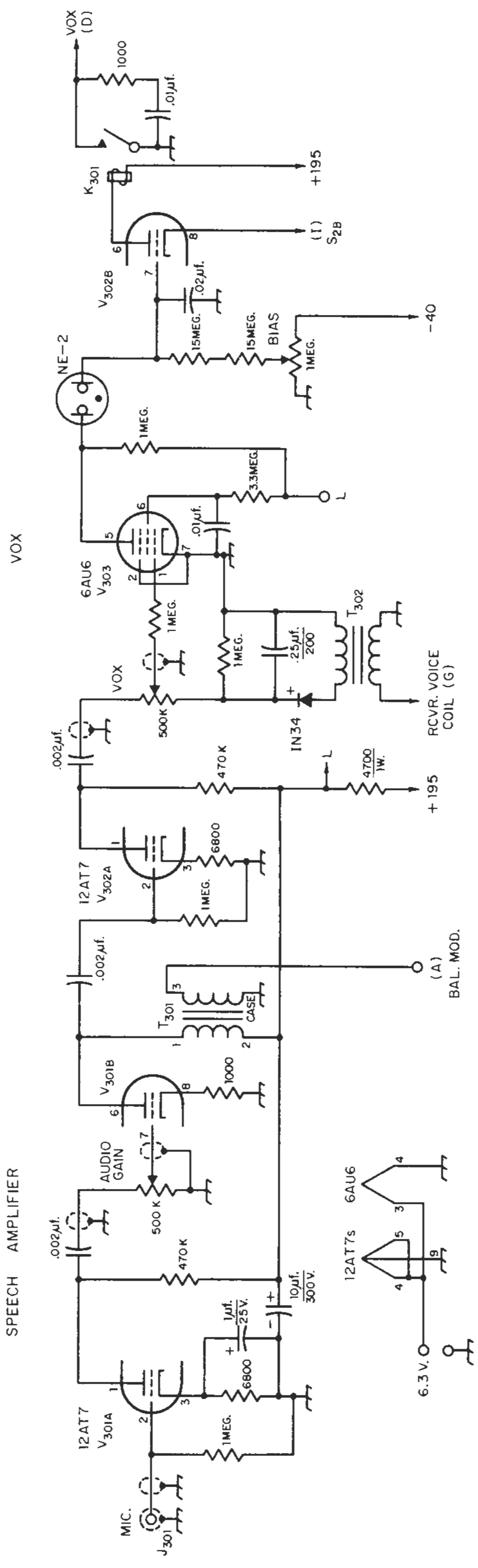
Each coil wound on 1-inch long ⅜-inch diam. polystyrene rod. Each coil shunted by 3–30-pf. trimmer unless otherwise noted.

Band	Coil
80 m.	70 turns No. 36 enam., shunted by fixed 30 pf.
40 m.	33 turns No. 30 enam.
20 m.	17 turns No. 22 enam.
15 m.	9 turns No. 18 enam.
*11 m.	8 turns No. 18 enam.
10 m.	7 turns No. 18 enam.
10 m.	7 turns No. 18 enam.

*This range may be omitted.

Assembly of the coils described in Table B.





The 6AK6 output tuning can next be checked over the 80-meter band to make sure that the 30-pf. panel control will tune the range. Next check the neutralization of the 6146 by connecting the r.f. probe in the antenna output connector. (The cathode of the 6146 is open in this position of S_2.) With grid and plate tuning capacitors peaked for maximum output, set the 2-10 pf. neutralizing capacitor for minimum output. If this does not fall within the range of the neutralizing capacitor, change the value of the 300 pf. capacitor.

To neutralize the 6AK6, disconnect the B+ at the 10K 10-watt resistor and, with the r.f. probe on the plate, adjust C_{103} for minimum feed-through. This completes 80-meter alignment and a rough neutralization which should be checked later on a higher band. Output can be checked by putting the operation switch to "test" with carrier inserted.

With the band switch on 40 meters, check the output of the crystal oscillator with the r.f. probe. With the receiver dial at 7.0 Mc., tune the v.f.o. dial until the signal from the mixer is heard about three turns from maximum capacity. (The 80-meter second harmonic will be heard with the capacitor about half open.)

If this calibration does not fall fairly close, with the bottom cover on, move the half-turn loop in L_{102} until it does. Set the main dial to about the center of the band as indicated by the receiver, with the 30-pf. grid-tuning capacitor in mid-range, and peak the active trimmer capacitors at S_{1A} and S_{1E}. Check the plate tuning of the 6146 at the same time. Repeat in the center of each band with the trimmers for the band, first peaking the oscillator trimmer on the four upper bands. Uniform output across each band should be obtained by retuning the grid and plate of the 6146 with the panel controls.

To tune the 8.0-Mc. trap used on 40 meters, set the band switch on 40 meters, S_2 to Test, and remove the 12AU7 balanced mixer, V_3. With the dual 150-pf. plate tuning capacitor near minimum capacity, output should be obtained as indicated by the r.f. probe. Adjust the slug in the series-trap inductor until the output goes through a minimum, which should be nearly zero. Recheck the tuning of the 3-30-pf. trimmer at 7150 kc. with the 12AU7 back in its socket and again recheck the trap if it is necessary to move the trimmer.

FIG. 6—Speech amplifier and VOX circuits. Unless otherwise indicated, resistances are in ohms, resistors are ½ watt.

K_{301}—5000-ohm s.p.d.t. relay (Potter-Brumfield SM5LS).

T_{301}—8-to-1 turns ratio output transformer (ARC-5 receiver output, No. 5631).

T_{302}—7K-to-50-ohm plate-to-line transformer, reversed (Triad A-51X).

J_{301}—Microphone jack (Amphenol 75-PC1M).

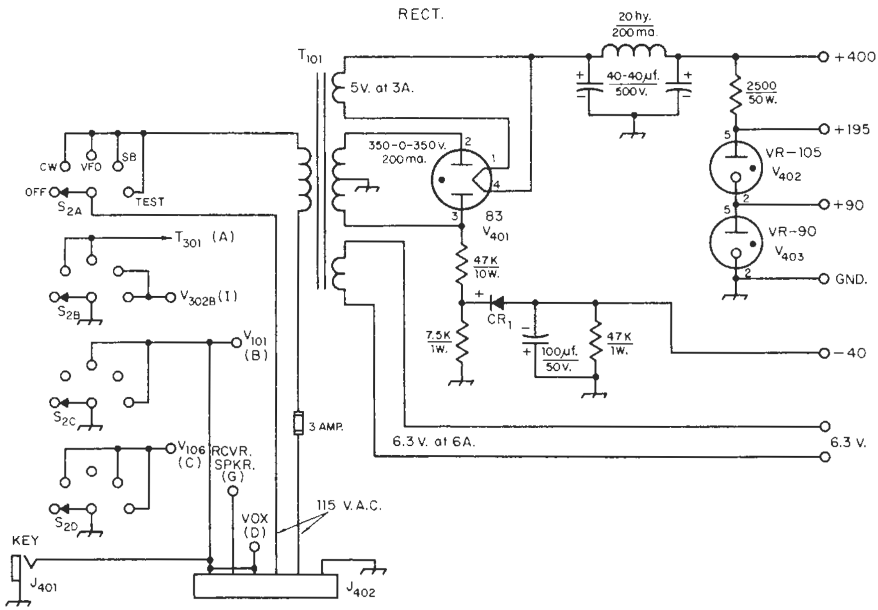


FIG. 7—The power supply section.

J401—Open-circuit phone jack.
J402—6-prong tube socket for external connections.
S2—Made of one 2-pole 5-position (4 used) shorting phenolic switch section (Centralab C) and one 2-pole 5-position (4 used) non-shorting phenolic switch section (Centralab K) on one indexing assembly (Centralab P-121). Shorting section is S2A-B.
CR1—130-volt 65-ma. selenium.
T101—Replacement power transformer (Thordarson 22R07).

Recheck neutralization on one of the high-frequency bands and alignment is complete.

A temporary dial calibration can be placed on tape run around the edge of the main dial. Make the .5-1.0 portion at 80 meters and the 0-.5 on 15 meters if a general coverage receiver is not available.

Audio and Voice Control

The audio and voice control circuits are mounted behind the sideband generator chassis on the main chassis. A small plug-in relay is used for silent control of whatever circuits are desired. The audio gain and voice control gain (VOX) potentiometer leads are brought to the front panel through two-wire shielded cables. A positive voltage is developed across the 0.25-µf. capacitor from rectified receiver audio, which prevents speaker operation of the voice relay. The shielded input lead is grounded at both V_{301A} and the mike input jack. After the audio wiring is complete, with the VOX pot set the bias to zero, by turning the arm to the ground end. The relay should close. Back the bias pot off from the ground end until the relay opens.

With the mike gain and VOX gain pots turned to suitable values, the relay should close instantly with speech but hold over by an amount determined by the *RC* constant in the grid of V_{302B}.

Main Tuning Dial

After the unit is complete, a permanent tuning dial can be made. The one shown is from an ARC-5 transmitter and gives a larger scale than a receiver dial. The old calibration was removed by slipping the dial over the shaft of a grinder and tightening a nut. The grinder was turned on and sandpaper held against the dial until the old calibration was gone. Four coats of flat black spray paint were applied, with fine sanding between coats. The plain dial was attached to the capacitor shaft so a small scratch could be placed every 25 kc. The short line decals are easily lined up by cutting them about ½ inch long and turning the excess over the back of the dial. After leaving the dial overnight, lacquer thinner was applied sparingly to remove the binder. This leaves shiny spots on the dial which are eliminated by spraying a coat of clear plastic over the whole dial. The result is a professional-looking dial with very little expense.

Some Notes on the "Sideband Package"

BERNARD WHITE, W3CVS

Since the construction and alignment of the Sideband Package are not as simple as in the straightforward transmitters most of us have been used to building in the past, some of the experiences here at W3CVS may be helpful to those who are planning to build it or may be experiencing troubles with it.

Alignment

One of the first problems we ran into after completing the filter section was insufficient output as measured across L_{101}. The author calls for 1½ to 2 volts, but we were only able to measure a little more than 1 volt. By placing the tip of a soldering gun into the hollow cores of the dowels in T_1 and T_3, it was possible to melt the wax enough to move the windings closer together by about ¼ inch. This was enough to raise the output voltage to the required figure.

In attempting to align the various transformers and coils throughout the exciter, it was found to be far simpler and more certain to set up these adjustments by using a grid-dip meter. This method avoids the possibility of aligning on a harmonic, which is very easy to do. In the writer's exciter it was not necessary to use some of the 3-30-pf. trimmers across the 6AK6 coils because the distributed capacitance, along with the grid tuning capacitor, was sufficient to tune the coils to the proper frequency with the grid capacitor just about in the middle of its range.

Stabilizing

After completing the exciter and giving it its first tryout, it was found that the 6AK6 and 6146 stages were very unstable because of feedback. The 6AK6 stage was cleaned up somewhat by shortening and carefully dressing the leads to the socket. This stage oscillates very easily, and particular attention should be paid in wiring the socket to keep the grid and plate connections as far apart as possible. By mounting a metal plate across the socket between the grid and plate prongs and grounding it, this stage was made completely stable on all frequencies. Finally, with careful neutralizing, there was no indication of voltage output on a v.t.v.m. with its r.f. probe touched on the plate connection of the 6AK6 tube, at any frequency.

The limiter d.c. leads were also shielded. The lead from the plate of the limiter tube should go directly to the plate of the 6146, of course, and should not run under the chassis.

Coil L_{105} was found to be contributing some feedback. This was corrected by placing the coil in an aluminum box.

Control Circuit

In the final operation of the exciter, tube noise from the 6146 could be heard in the receiver during standby, since this stage was not completely cut off as originally wired. To cure this the control circuit was rewired as shown in Fig. 1, so the cathode of the 6146 would be disconnected from ground during standby.

The 6146 stage was next tackled, and it was quickly found that the coax running from the link output coil to the J_{101} connector was contributing a considerable amount of feedback through being terminated under the chassis by means of a feed-through terminal. This terminal was removed and the coax brought directly to the top of the chassis, where it was terminated by a stand-off insulator to make connection to the link.

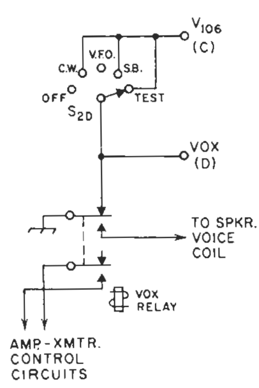


FIG. 1—Modified control circuit using a d.p.d.t. relay in the VOX circuit. This eliminates noise from the final amplifier during receiving, short-circuits the receiver speaker voice coil during transmitting, and provides a control circuit for a linear amplifier, antenna relay and other accessories.

From February, 1960, *QST*.

Those who wish to control the receiver and transmitter by the VOX relay alone may be interested to know that the Allied Control Company [1] manufactures a 5000-ohm plug-in relay of the same type and size as the P-B type 5MSLS, except that it is double-pole double-throw. It is known as the type RSHX-51 and comes with a matching socket. It is also available as a wired-in relay. The contacts are rated at 2 amperes. Using this relay it is possible to take advantage of the extra contact to disable the voice coil circuit of the speaker during transmitting.

[1] Allied Control Company, Inc., 2 East End Ave., New York, N. Y.

General

More than enough gain is available, and it is very easy to overdrive the amplifier if the gain control is advanced too far. It is suggested that a scope be used in conjunction with a 1000-cycle oscillator to conduct a two-tone test for determining the correct settings for the controls on all bands.

On-the-air comments have been most gratifying. Reports indicate that the pass-band characteristics are excellent, and the voice quality is more like broadcast quality than the usual s.s.b. audio sound. The v.f.o. has also been commented on as being very stable with practically no drift.

MORE "SIDEBAND PACKAGE" NOTES

Following three years of active s.s.b. operation on 75, 20, 15 and 10 meters, and an extensive overhaul of my "S.S.B. Package," built in 1958 from W6TEU's original article in June, 1958, *QST*, I have several suggestions which should be of help to anyone constructing this excellent exciter.

1) The use of high-quality potentiometers such as Ohmite CA will prevent failure and allow better operation. I had to replace the carrier-insertion control (a replacement-type radio and TV control) because of failure, and changing to an Ohmite CA also cured an annoying reduction in carrier on a.m. and c.w. as the exciter warmed up. Apparently the carrier-insertion control increased in resistance due to heating caused by the cathode current of V_{1B}.

2) The 8th harmonic of the carrier oscillator at 3640 kc., unless attenuated, will leak out of the sideband generator, be amplified in the 3.5-Mc. part of the exciter and will appear as an extra signal on each band. This extra frequency will also beat with the desired frequency if the two are close in frequency, resulting in an unacceptable signal. After much experimentation I have discovered a simple and very effective way of attenuating this unwanted harmonic. A parallel trap, using a low-inductance slug-tuned coil in parallel with a large silver-mica capacitor (I used 470 pf.) tuned to 3640 kc., should be connected in the plate lead of the 6X8 frequency multiplier, V_{4B}. Using a small inductance and relatively large capacitor seems to cause less detuning and attenuation of the desired multiplier frequencies. This treatment reduces the harmonic many db., but does not eliminate it. Complete elimination may be obtained by placing a series trap, consisting of a 4-30-pf. ceramic trimmer in series with a large inductance (TV peaking coil), between the grid of V_{101A} and ground, across coil L_{101}. This trap detunes L_{101}, so the capacitance used should be small. By working back and forth between L_{101} and the series trap capacitor, the harmonic may be completely eliminated while the 2250-kc. s.s.b. signal is not attenuated. Use a receiver tuned to 3640 kc. with a test probe connected to the antenna input for an indicator.

3) Very careful shielding and placement of components in the final mixer and amplifier stages is necessary to make the exciter completely stable. I would recommend that these stages, as well as the band-switching coils, be enclosed in shield compartments. Carefully located and dressed leads are a must for stable operation. Shielding coil L_{101} resulted in a very definite improvement in operation.

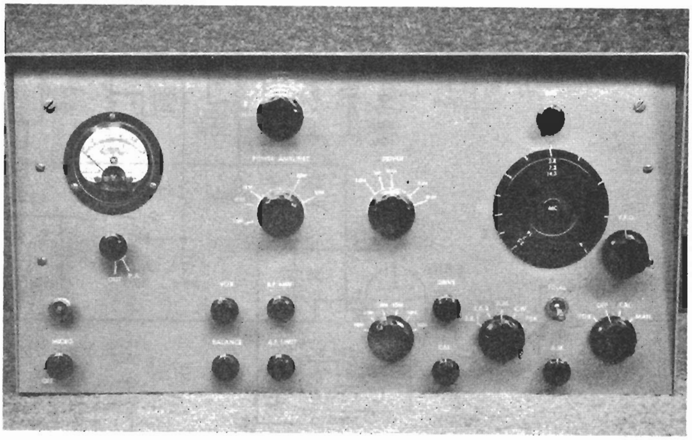
4) In my exciter, heat has been a problem and a blower is required to cool the unit—which, incidentally, is mounted in a well-ventilated hinged-lid cabinet. I plan to replace the 83 rectifier by silicon rectifiers. This should result in much less heat being generated as well as better efficiency and a reduction in receiver noise.

The "Sideband Package" is used at VE3BHQ to drive a pair of 811s in grounded grid (*QST*, January, 1960). The output of the exciter is exactly right for this linear. Reports on all bands verify the excellent sideband suppression and almost complete carrier suppression claimed for this exciter. Audio quality reports are excellent and compare favorably with the most expensive commercial amateur transmitters. Frequency stability is remarkable, with practically no drift after the first several minutes of operation. I would not recommend this exciter as a project for a beginner; however, for a person with some construction experience it is well worth while.

—Farncomo Le Gresley, VE3BHQ

From "Technical Correspondence", *QST*, May, 1962.

» 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 (S_1). On the upper portion of the panel, near center, are the loading control (C_{25}) 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 difficut to prove that the purchase of commercial gear will "pay off" in the end. The advocators 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 W6TEU[1] and an adaptation by K4EEU,[2] looked especially interesting. Although the basic signal-generating

From November, 1962, *QST*.

[1] Bigler, "A Sideband Package," page 59.
[2] Kelley, "A Phasing-Type Sidebander," page 80.

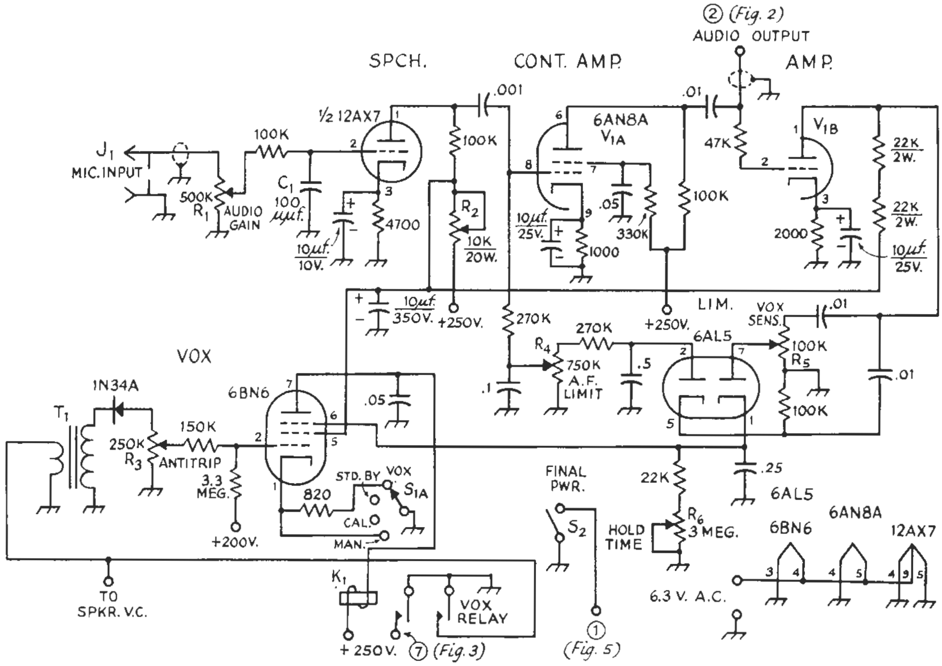


FIG. 1—Audio and VOX-control circuits. Resistances are in ohms and fixed resistors used by the author are rated (conservatively in most cases) at 1 watt, unless indicated otherwise. Except for C_1, capacitances are in μf. Capacitors are 600-volt paper except for those marked with polarity, which are electrolytic.

C_1—Mica.

J_1—Microphone connector.

K_1—D.p.s.t. relay, 5000-ohm coil (Potter & Brumfield LM11). Use series resistance, if necessary, to limit coil current to rated 6.3 ma.

R_1—Audio-taper control.

R_2—Slider adjustable. Set for 105 volts.

R_3, R_4, R_5, R_6—Linear-taper control.

S_1—3-pole 4-position rotary switch (CRL 1415, one pole not used). See Fig. 3 for other sections.

S_2—S.p.s.t. toggle switch.

T_1—Universal output transformer, 10,000 ohms to voice coil.

circuits are somewhat different, you will notice a strong resemblance between the author's exciter and the two just mentioned. The problem of what to do in the audio section was solved by the "Omnivox," which was designed by W4PFQ.[3] His circuit was used almost intact. It includes a.f. limiting in addition to standard features such as VOX and antitrip.

Low-Level Sections

The circuit of the audio section is shown in Fig. 1. The microphone output is amplified by one section of a 12AX7 (second triode is not used) and the pentode section of a 6AN8. The output goes to the 7360 balanced modulator and the 6BE6 a.m. modulator (Fig. 2). The output is further amplified by the triode section of the 6AN8. The output of this section is then rectified by the two sections of a 6AL5. The negative d.c. output of one rectifier is fed to the control grid of the pentode section of the 6AN8. The amount is adjustable by a potentiometer, R_4, which is the a.f. limit control. The amount of this limiting is adjustable over a wide range. The second rectifier section produces a positive d.c. output which is applied to one of the grids of the 6BN6 relay tube. When this voltage is sufficiently large, the tube normally conducts and the VOX relay becomes energized. Voltage from the receiver speaker circuit is stepped up in T_1, rectified, and the

[3] Hoase, "The Omnivox," *G.E. Ham News,* Jan.-Feb., 1961.

TABLE I

Bandswitching Coils

Band	L_8	L_9 (Number of Turns)	L_{10}
80	23	60*	60*
40	21	33**	33**
20	15	21	21
15	10	21	21
10	6	9	8
10	Same as above		

All coils are close-wound on ⅜-inch polystyrene rod with No. 22 enameled wire, except * wound with No. 30, and ** wound with No. 26.

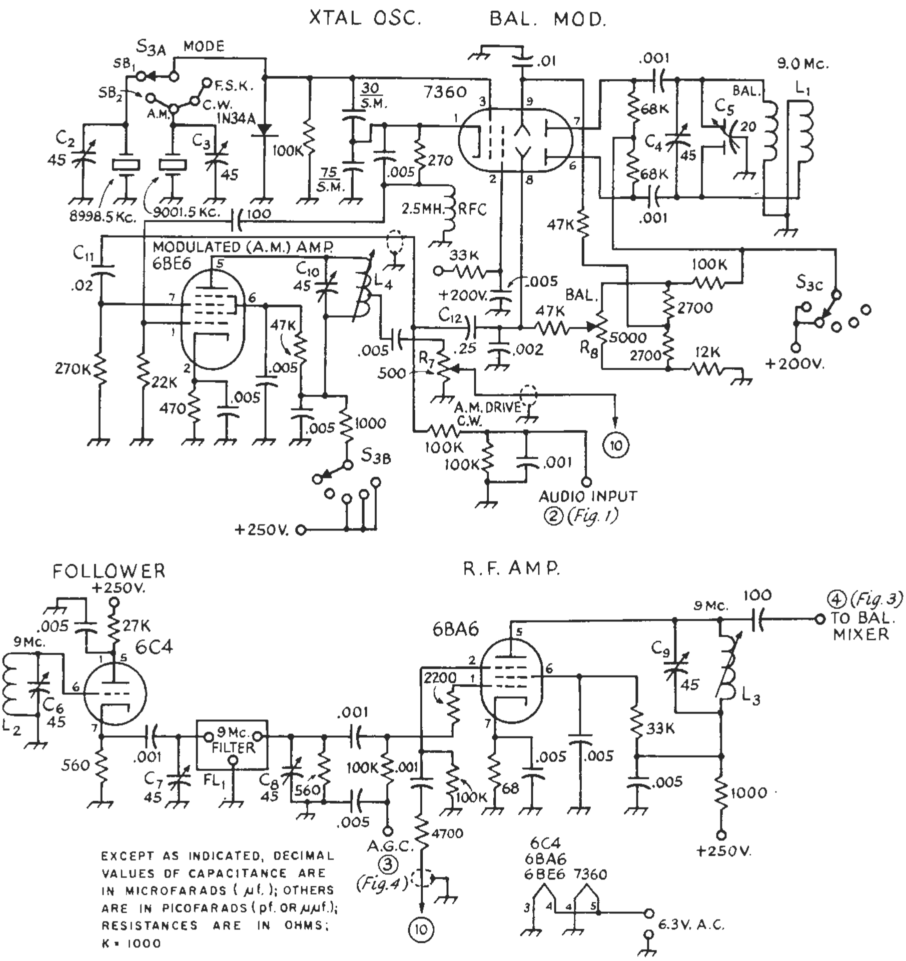


FIG. 2—Sideband and a.m. generator circuits. Resistances are in ohms and fixed resistors used by the author are rated (conservatively in most cases) at 1 watt unless indicated otherwise. Fixed capacitors of less than 0.001 μf. are mica or silver mica (SM); others are disk ceramic, except as listed below.

C_2, C_3, C_4, C_6, C_7, C_8, C_9, C_{10}—7—45-pf. ceramic trimmer (Centralab 822-BN or equivalent).

C_5—Differential capacitor (Johnson 19MA11/160-311).

C_{11}, C_{12}—Paper.

FL_1—Sideband filter (McCoy Electronics 32 B1).

L_1—32 turns No. 26 enam., bifilar-wound on 3/8-inch polystyrene rod.

L_2—38 turns No. 26 enam., close-wound on 3/8-inch polystyrene rod. Form is placed parallel to form of L_1, forms spaced 3/4 inch center to center.

L_3—30 turns No. 26 enam., close-wound on 3/8-inch ceramic iron-slug form (Miller 4400 form).

L_4—Same as L_3, tap at 7 turns from ground end.

R_7—Carbon control, linear taper.

R_8—Linear-taper control.

S_3—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,

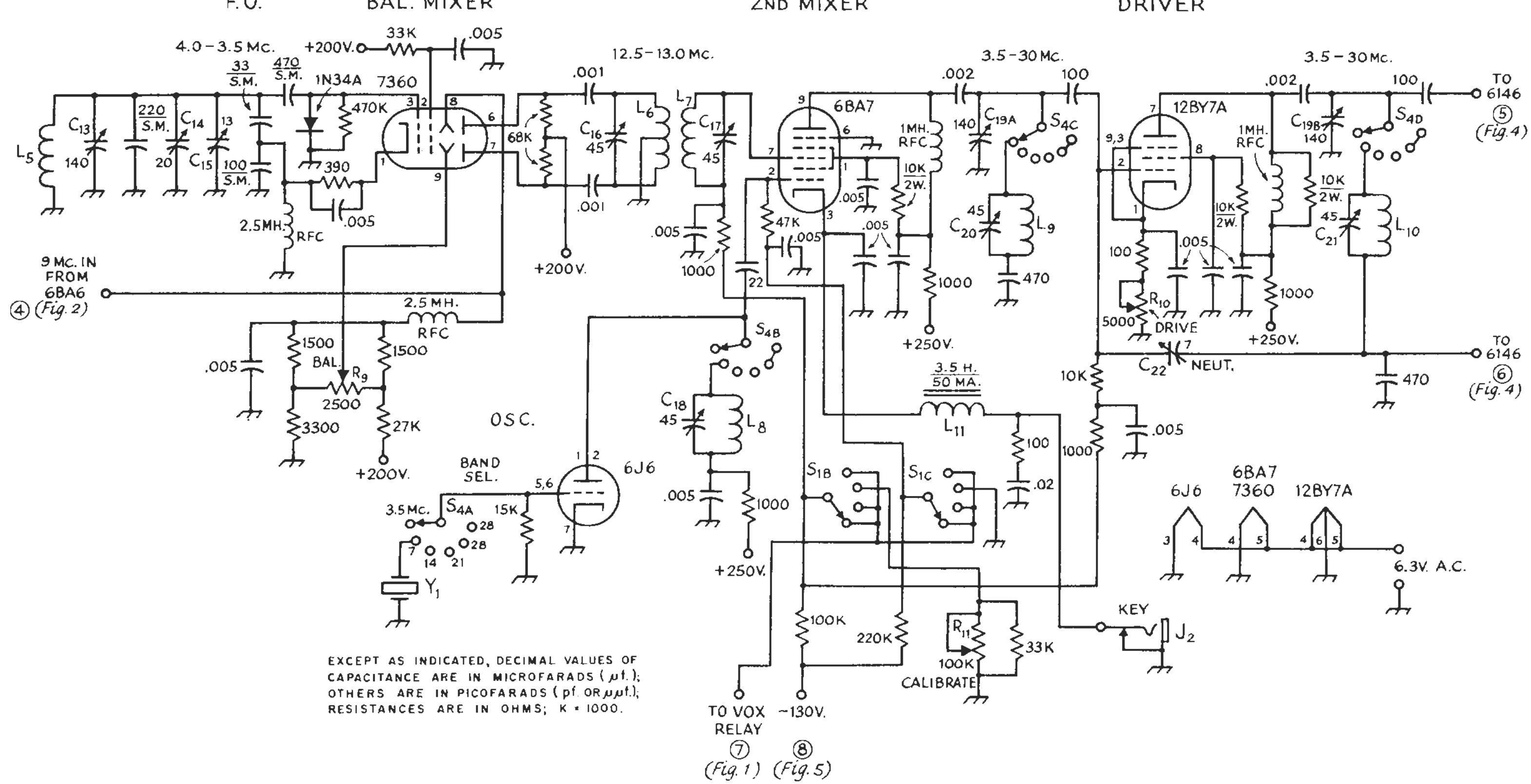


FIG. 3—V.f.o., band-heterodyning and driver circuits. Resistances are in ohms, and fixed resistors used by the author are rated (conservatively in most cases) at 1 watt. Fixed capacitors of less than 0.001 μf. are mica or silver mica (SM); others are disk ceramic. Triode sections of 6J6 are connected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

C_{13}—Midget two-bearing 140-pf. air variable.
C_{14}—N300 5–20-pf. ceramic trimmer (Erie TS-D).
C_{15}—NPO 2.5–13-pf. ceramic trimmer (CRL 822-BZ).
C_{16}, C_{17}, C_{18}, C_{20}, C_{21}—7–45-pf. ceramic trimmer (CRL 822-BN). In bandswitching circuits, a similar capacitor is connected across each coil.
C_{19}—Dual 140-pf. air variable.
C_{22}—NPO 1.5–7-pf. ceramic trimmer (CRL 822-EZ).
J_2—Closed-circuit jack.
L_5—26 turns No. 22 enam., close-wound on ½-inch low-loss bakelite form (National XR-50 form with slug removed).
L_6, L_7—26 turns No. 26 enam., close-wound on ⅜-inch polystyrene rod, forms mounted as described for L_1L_2.
L_8, L_9, L_{10}—See coil table.
L_{11}—Filter choke (Stancor C-1080).
R_9, R_{10}, R_{11}—Linear control.
S_1—See Fig. 1.
S_4—4-pole 6-position ceramic rotary switch (CRL P-272 index head, 4 type XD wafers).
Y_1—16.5 Mc. for 2.5-Mc. output.
20.0 Mc. for 7-Mc. output
27.0 Mc. for 14-Mc. output.
34.0 Mc. for 21-Mc. output.
41.0 Mc. for 28–28.5-Mc. output.
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.

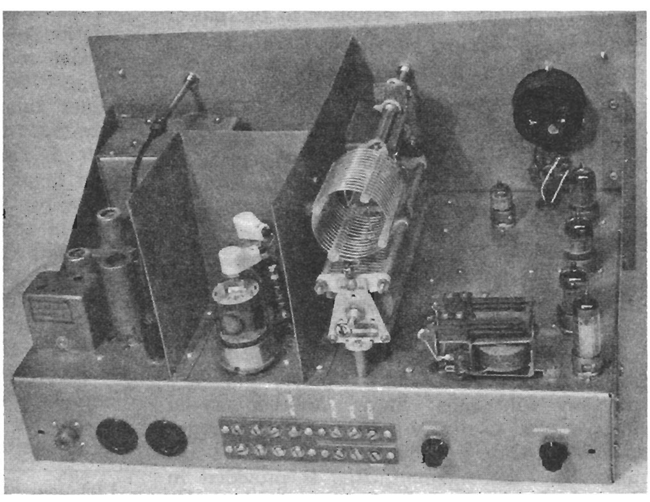
In this exciter, some of the output of the 9-Mc. crystal oscillator is fed to grid No. 1 of a 6BE6 r.f. amplifier. Audio is fed to grid No. 3. The plate is tuned to 9 Mc. and the output is a standard a.m. signal. Proper adjustment of the cathode resistor and the audio input is necessary to obtain the 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 products 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.

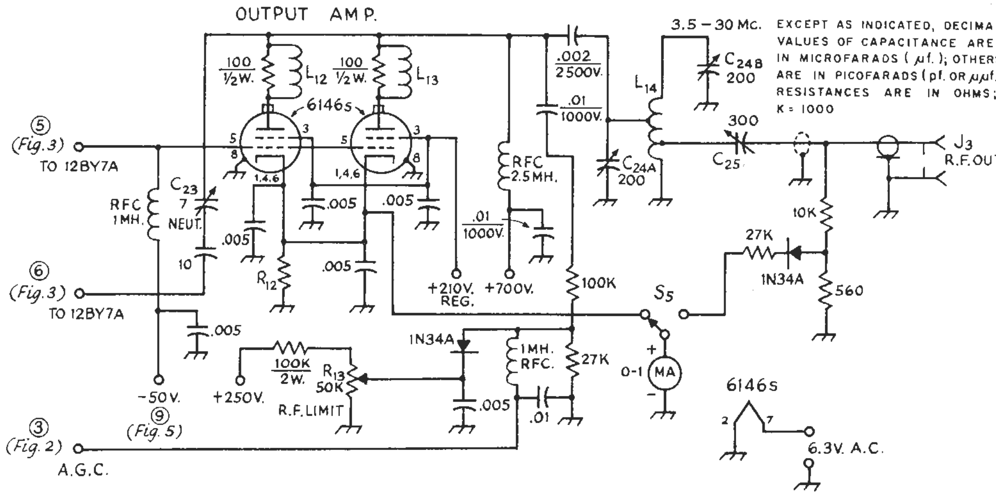


FIG. 4—Circuit of the output amplifiers. Fixed capacitors of less than 0.005 μf. are mica; others are disk ceramic. Resistances are in ohms and resistors are 1 watt unless indicated otherwise.

C_{23}—Same as C_{22}.
C_{24}—Dual 200-pf. 2000-volt variable (Johnson 200FD20/155–505).
C_{25}—Midget air variable (Hammarlund MC-325-M).
J_3—Chassis-mounting coax receptacle.
L_{12}, L_{13}—5 turns No. 20 close-wound on associated resistor.
L_{14}—15 turns No. 14, 2-inch diam., 6 turns per inch, tapped at 3 and 8 turns from ground and (Air Dux 1606T stock).
R_{12}—300-times meter shunt.
R_{13}—Linear control.
S_5—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, C_{19A}, adjustable from the front panel.

Following the 6BA7 is a 12BY7 which is tuned by a second section (C_{19B}) of the same variable capacitor used for the plate of the 6BA7. (The 470-pf. capacitor from L_9 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, R_{13}, 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, R_{12}, 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.

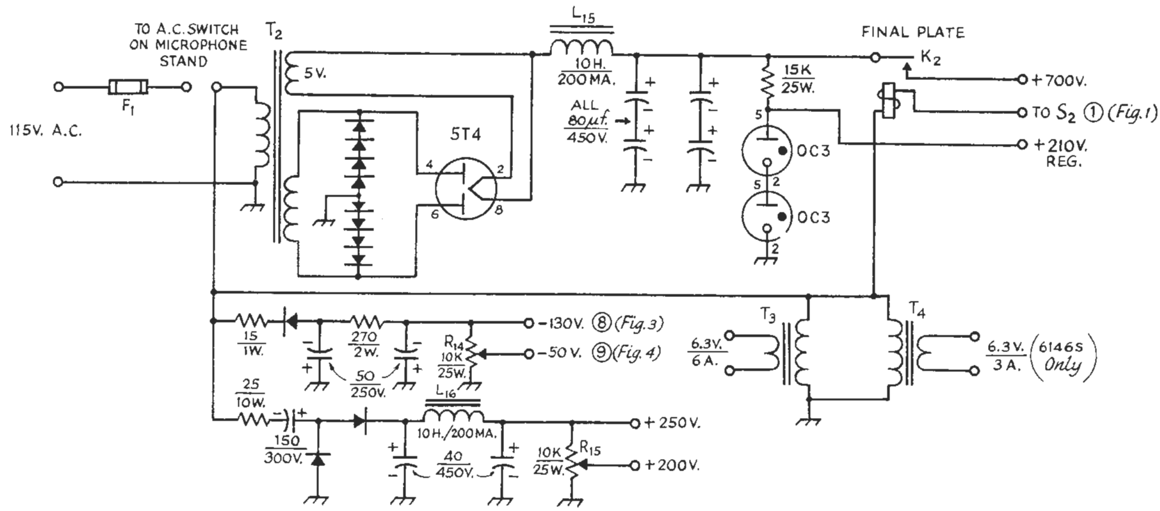


FIG. 5—Power-supply circuits. Capacitances are in μf. and capacitors are electrolytic. Resistances are in ohms. All solid-state rectifiers are 130-volt a.c., 500-ma. d.c. silicon units (Sarkes-Tarzian M-500). See text regarding use of polarized a.c. line plug.

F_1—Fuse, 5 amp.
K_2—S.p.s.t. 115-v. a.c. relay (Potter & Brumfield KT11A or similar).
L_{15}, L_{16}—Filter choke (Stancor C-2705 or similar).
R_{14}, R_{15}—With adjustable slider.
T_2—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).
T_3—6.3-volt 6-amp. filament transformer.
T_4—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.

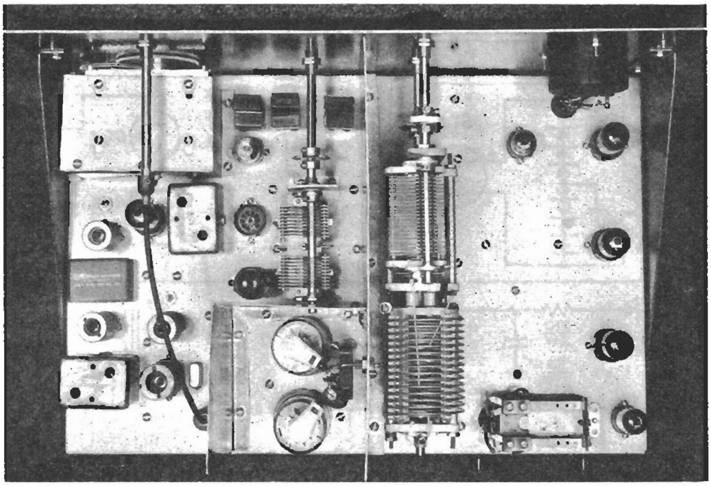
For transmitting, the -130-volt bias must be removed, and the receiver speaker shorted. This function is performed by the VOX relay. Talking into the microphone will cause the relay to close. The relay will also be energized if the function switch is placed in the manual position. One set of contacts 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 ⅜-inch polystyrene rod as are the bandswitching 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 ⅛-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 ¾ 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.

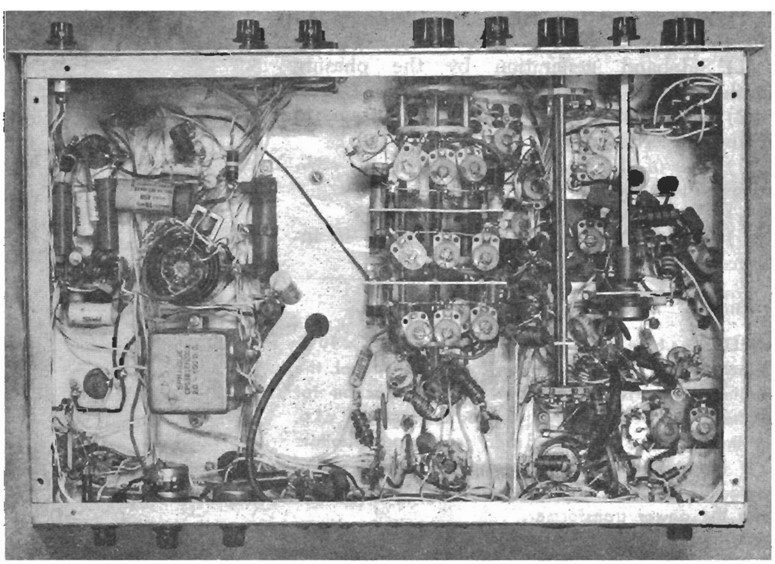
The band-switching circuits of the 12BY7 are adjusted by shifting the probe to the grids of the 6146s, setting C_{19} to the recorded points in succession, and adjusting the C_{21} trimmers for maximum readings.

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.

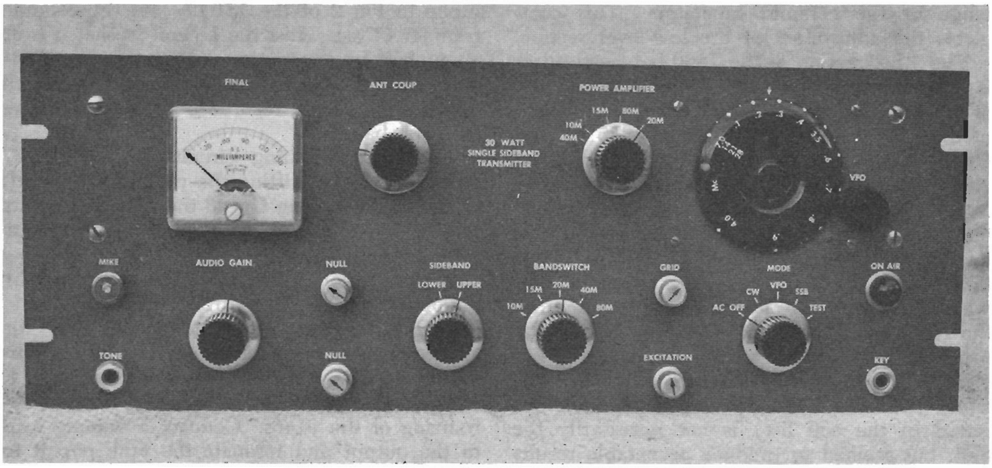
The preceding alignment information is necessarily brief and hits only the high spots. The previously-cited articles give additional information which should be useful. It is assumed that anyone with enough experience to build such an exciter would have no difficulty with its alignment.

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.

» 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 W6TEU[1] 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.

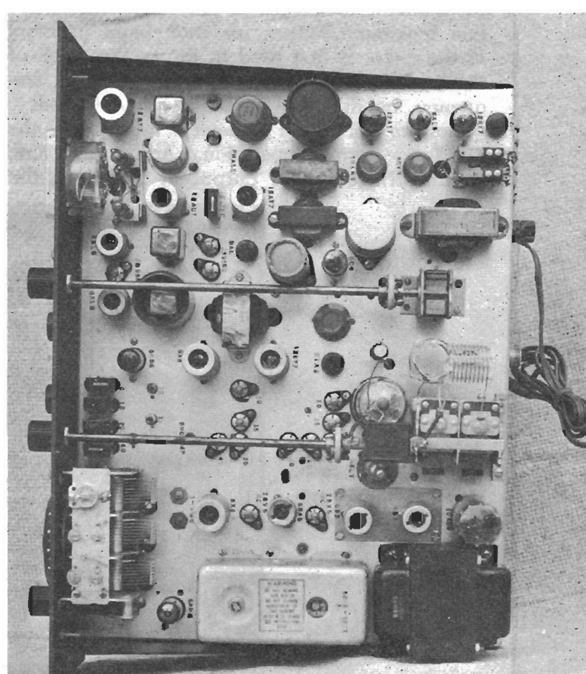
4) Provision for side tone on c.w.

5) Modification of the power supply to increase the power output.

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

From November, 1959, *QST*.

[1] Bigler, "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.

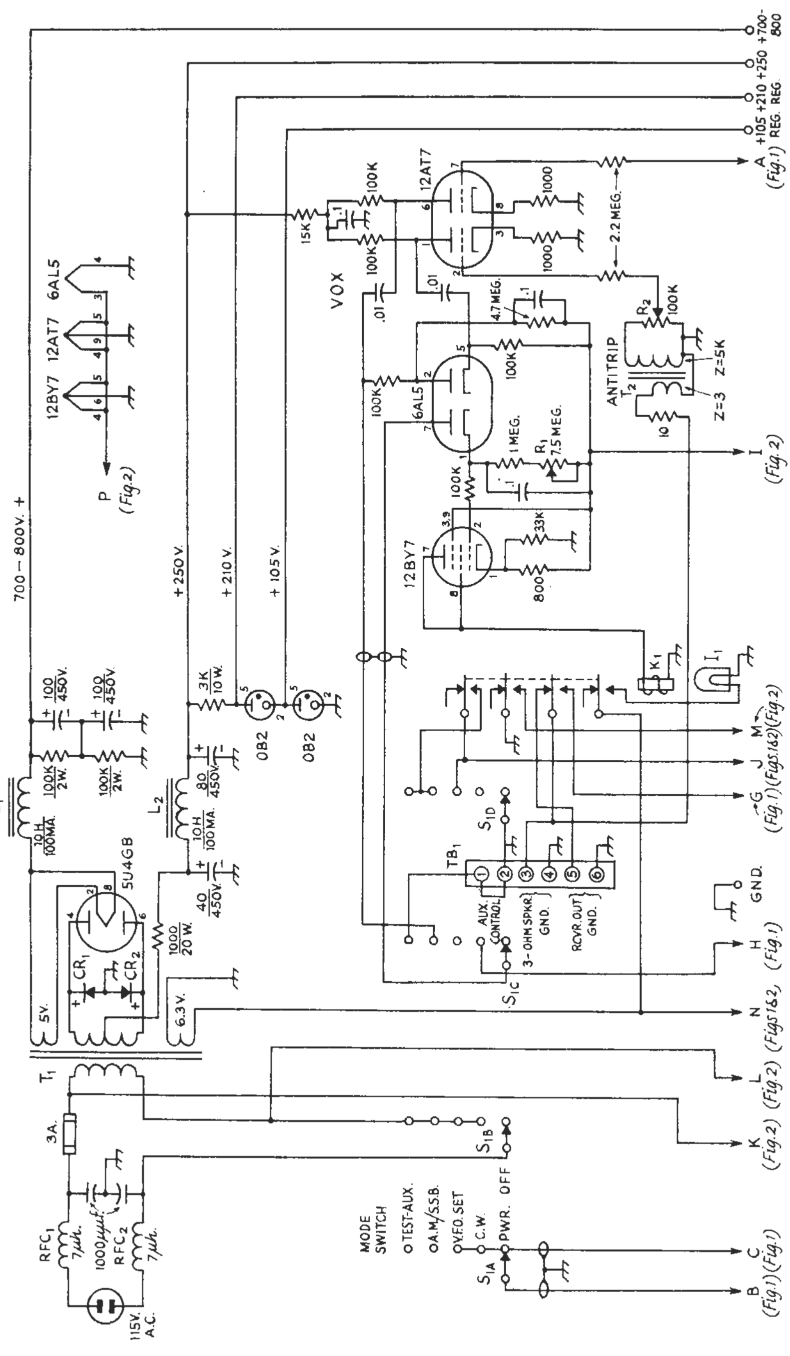


FIG. 3—VOX, power-supply and control circuits of K4EEU's s.s.b. exciter.

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 1/2 watt unless indicated otherwise.

CR_1, CR_2—Silicon rectifier, 840 volts r.m.s., 425 ma. (Sarkes Tarzian 120SM).

I_1—6.3-volt dial lamp.

K_1—4 p.d.t. 5000-ohm relay (Potter & Brumfield GPD coil and GP17 contact assembly).

L_1, L_2—10.5-hy. 110-ma. filter choke (Stancor-C1001).

R_1, R_2—Potentiometer, carbon.

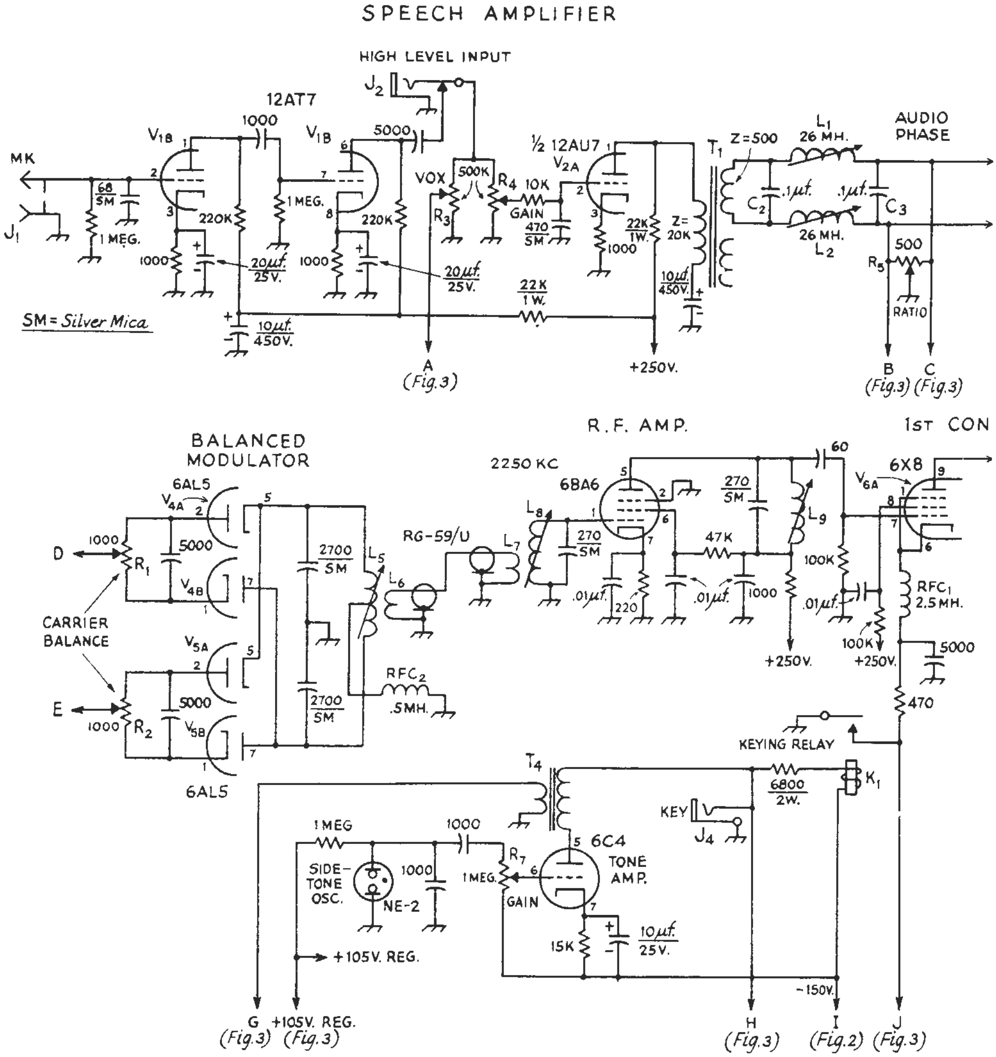
S_1—4-pole 5-position rotary switch (Centralab PA-1013 or PA-2011).

T_1—Power transformer: 700 v.c.t., 160 ma.; 5 volts, 3 amp.; 6.3 volts, 5 amp. (Triad R-16A*).

T_2—5-watt output transformer, 5000 ohms to 3.2-ohm voice coil.

RFC_1, RFC_2—7-μh. 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.



If all parts and surplus items are purchased as needed especially for this transmitter, the cost will be less than $150. A moderately well-stocked spare parts supply will reduce this figure considerably and, if care is taken in construction, the builder will be rewarded with a transmitter equal in most respects to commercial exciters costing several times as much.

A project such as this is not for the Novice, but the ham with experience in building transmitters and superhet receivers should be well qualified.

Circuit Discussion

The sideband signal is generated at 2250 kc. with the circuit used in "S.S.B. Jr."[2] and amplified to a level of about 2 volts by a 6BA6. It is then heterodyned with a v.f.o. signal operating from 5250 to 6250 kc. The resulting 3- to 4-Mc. heterodyne is selected by tuned band-pass coils and further amplified by linear amplifiers for straight-through operation on 80 meters. For other bands the 80-meter signal is again converted by crystal oscillators to the correct frequencies. Output frequencies of the crystal-controlled oscillator in the second converter are chosen so that the low-frequency end of the first-converter range is used for the higher-frequency bands. As examples, the 4-Mc. output from the 6CB6 in the second converter beats with the 3-to-3.3-Mc. output-of the first converter to cover the 7-to-7.3-Mc. range; the second harmonic of the 9-Mc. crystal (18 Mc.) beats with the 3-to-3.45-Mc. portion of the first converter range to produce output from 21 to 21.45 Mc., etc. The high-frequency end of the first-converter range is used only for direct 80-meter output and for the high-frequency end of the 10-meter band.

The V.F.O.

If you have a BC-458 chassis you may start by removing all the audio and r.f. circuitry up to the 12A6 mixer, and using these parts in

[2] "S.S.B. Jr.," *G.E. Ham News, Nov.-Dec.*, 1950 (Vol. 5, No. 6).

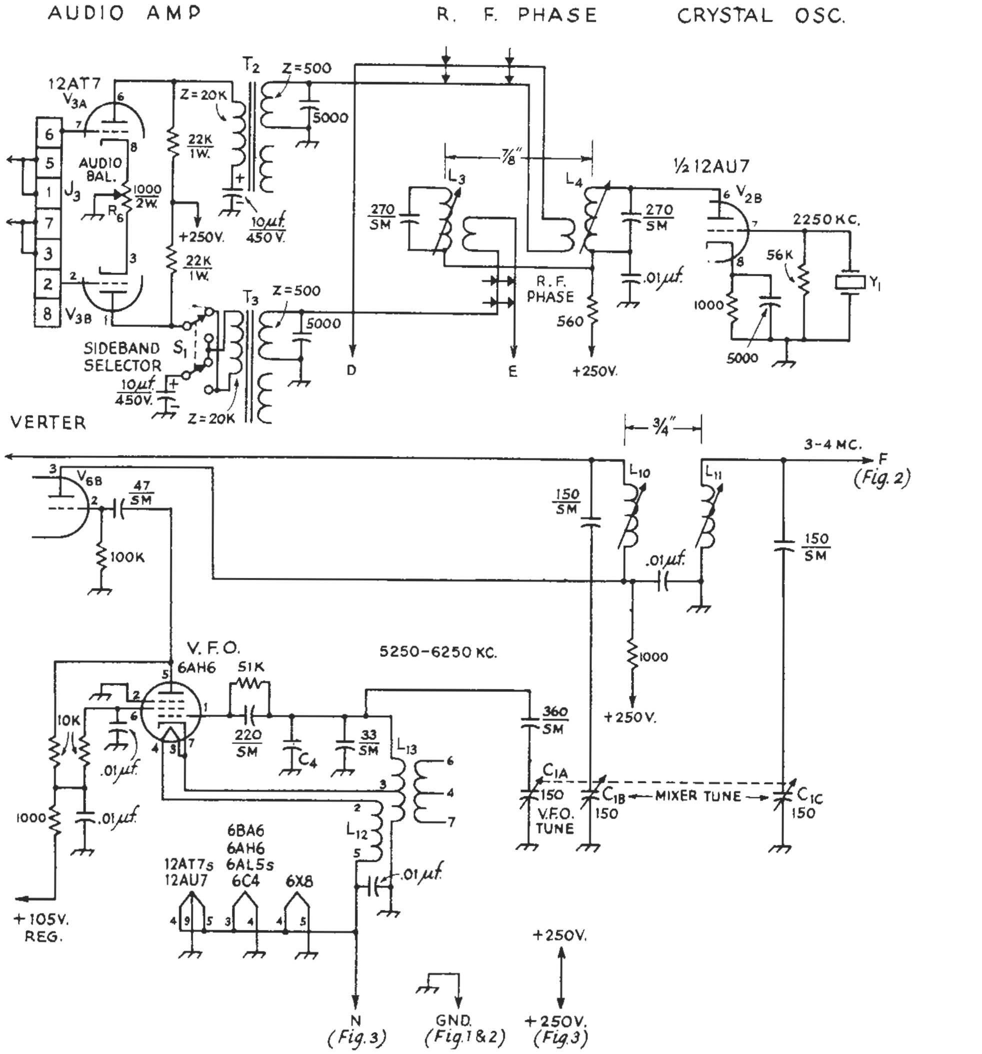


FIG. 1—Sideband-generator, 4-Mc.-output, and c.w.-keying sections of the 30-watt s.s.b. transmitter. Unless otherwise marked, capacitances are in pf. Capacitors with polarity markings are electrolytic. Other fixed capacitors not marked SM may be mica or ceramic. Unless indicated otherwise, resistors are ½ watt.

C_1—Three-gang capacitor, see text.
C_2, C_3—Disk ceramic (Centralab DDA104).
C_4—Fixed air padder in BC-458.
J_1—Microphone connector.
J_2—Closed-circuit jack.
J_3—Octal tube socket (for B & W 350 phaseshift net.)
J_4—Open-circuit jack.
K_1—2500-ohm relay, s.p.s.t. contacts (Potter & Brumfield LB-5).
L_1, L_2—26 mh. TV-width coil, (Thordarson WC-19).
L_3,* L_4,* L_8,* L_9*—45 turns No. 28 enam., adjusted to resonate at 2250 kc. (2-turn links on L_3, L_4. Arrows indicate twisted pair).
L_5*—16 turns No. 22 enam. double-spaced, c. t., adjusted to resonate at 2250 kc.
L_6—2 turns over center of L_5.
L_7—6 turns at ground end of L_8.
L_{10}, L_{11}—40 turns No. 34 enam., close-wound on ½-inch diam. iron-slug form (National XR-50).
L_{12}, L_{13}—Oscillator coil unit from BC-458.
R_1, R_2, R_5, R_6—Linear-taper potentiometer.
R_3, R_4, R_7—Audio-taper potentiometer.
RFC_1—2.5-mh. r.f. choke (National R-100S or equiv.)
RFC_2—0.5-mh. r.f. choke (National R-50).
S_1—D.p.d.t. rotary switch (Centralab 1462).
T_1, T_2, T_3—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.
T_4—5-watt tube-to-voice coil (3.2 ohms) output transformer (Thordarson 24S50 or equivalent).
Y_1—2250-kc. crystal.
* Wound on 5/16-inch diam. iron-slug form from surplus unit. If 3/8-inch forms are used, turns should be decreased by about 25 per cent. If ¼-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.

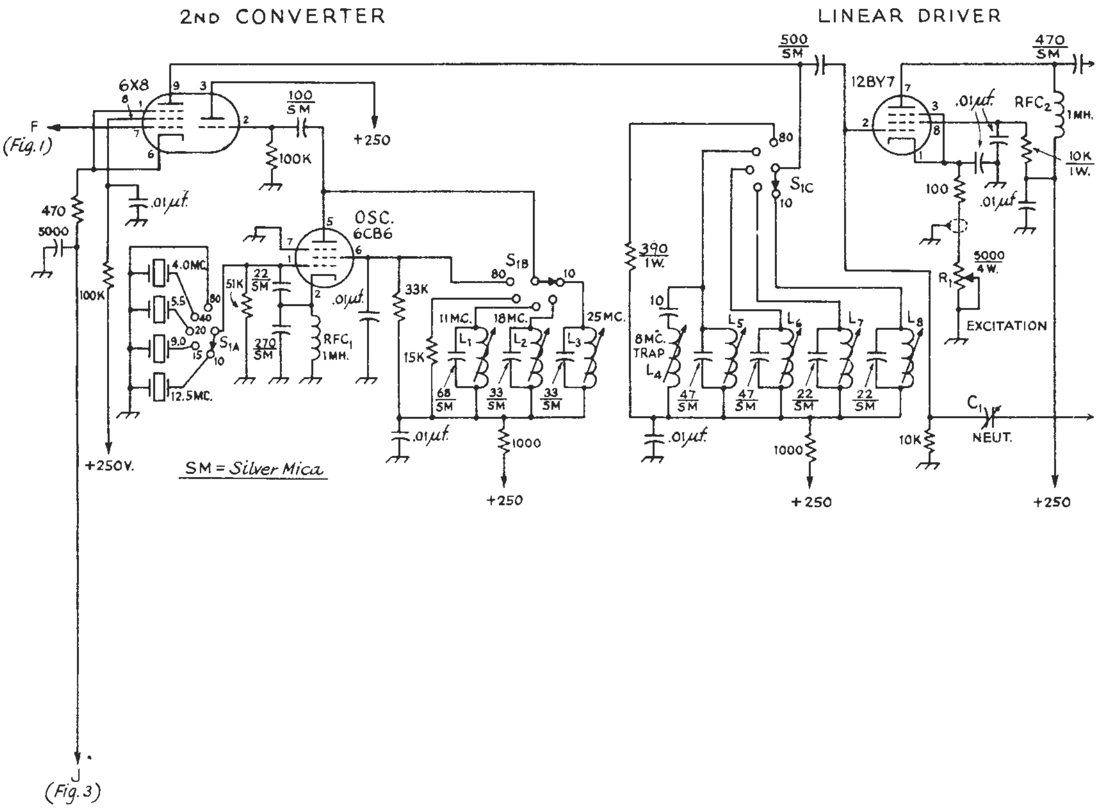


FIG. 2—Second converter, driver and final-amplifier circuits of K4EEU's s.s.b. exciter. Unless otherwise indicated, capacitances are in pf. Fixed capacitors marked SM should be silver mica or NPO ceramic; those with polarity markings are electrolytic. Other fixed capacitors are disk ceramic. Resistors are ½ watt unless indicated otherwise, and resistances are in ohms.

C_1—Neutralizing capacitor—insulated wires twisted together to form capacitor.

C_2—30-pf. variable, midget receiving type.

C_3—Neutralizing capacitor, 0.5 to 5 pf. (Millen 15001, Bud NC-1929 or similar).

C_4—Dual 150-pf. variable capacitor (Bud CE-2046 or surplus).

C_5—365-pf. variable, broadcast-replacement type.

CR_1—50-ma. selenium rectifier.

J_1—Chassis-mounting coax receptacle (SO-239).

L_1—15 turns No. 22.

L_2—13 turns No. 22.

L_3—10 turns No. 22.

L_4—100 turns No. 34.

L_5—30 turns No. 26.

L_6—20 turns No. 22.

L_7—10 turns No. 22.

L_8—7 turns No. 22.

L_9—90 turns No. 34.

L_{10}—20 turns No. 26.

L_{11}—20 turns No. 22.

L_{12}—9 turns No. 22.

L_{13}—4 turns No. 22.

Note: In reference to the above coils, slug is adjusted in each case to resonance at the circuit frequency indicated in the diagram. See * in caption Fig. 1.

L_{14}—9 turns No. 18, 1-inch diam., 1⅛ inches long (Air Dux 808T).

L_{15}—21 turns No. 20, 1-inch diam., 1¼ inches long (Air Dux 816T).

L_{16}—8 turns No. 18, 1¼-inch diam., ¾ inch long (Air Dux 1010T).

Note: C_4, L_{14} and L_{15} form a multiband tuner.

M_1—3-inch-square milliammeter.

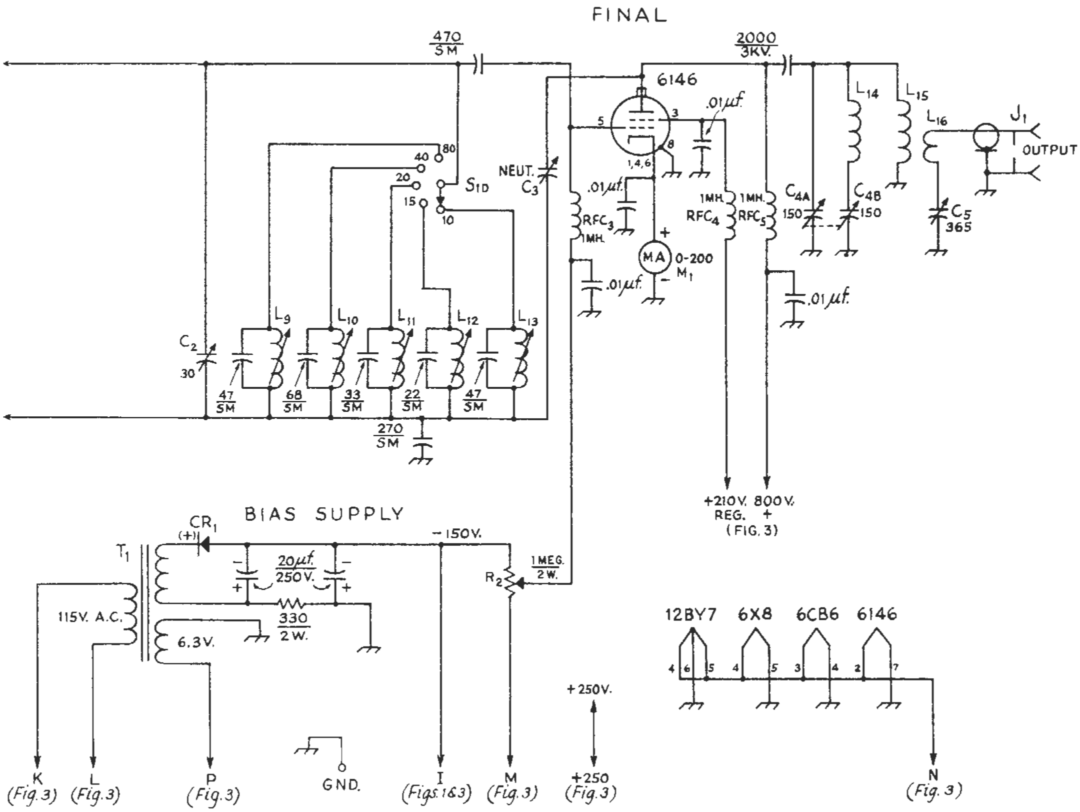
RFC_1, RFC_2, RFC_3, RFC_4—1-mh. 100-ma. r.f. choke.

RFC_5—1-mh. 300-ma. r.f. choke.

S_1—Band switch made from Centralab Switchkit components as follows: Index assembly—PA-302; S_{1A} and S_{1B} combined on single PA-5 switch section (two of three circuits used); S_{1C}, S_{1D}—each one PA-41 or PA-18 section (all unused contacts connected and shorted out, 5 of 11 positions used).

T_1—Power transformer: 125 volts, 50 ma.; 6.3 volts, 2 amp. (Stancor PA-8421 or similar).

the new transmitter. The BC-458 has an excellent v.f.o. and this, too, is used. The ceramic coil, fixed air padding capacitor, and the shield are re-used in the new v.f.o. The small green fixed capacitor on the air padder is removed. The old tuning variable is not used, and the triple section tuning capacitor from an R-26/ARC-5 (3-6 Mc.) receiver is substituted.



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 grommeted 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

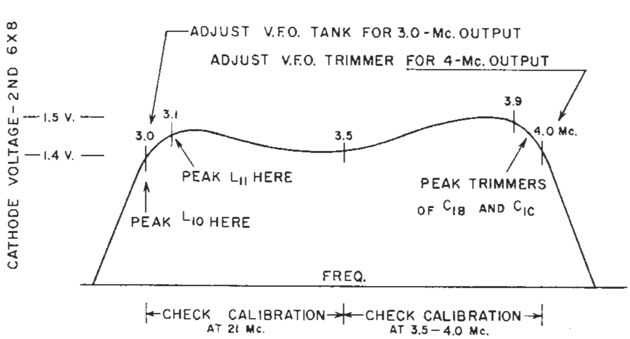
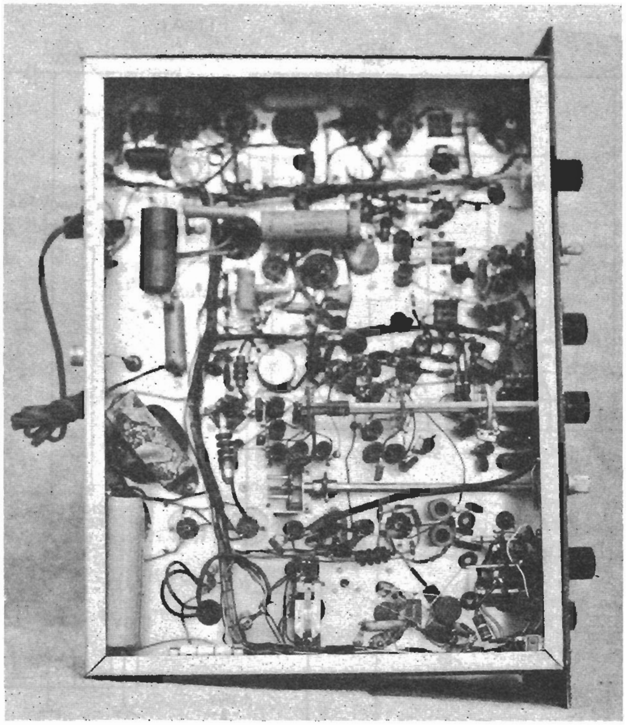


FIG. 4—Response curve of the band-pass coupler L_{10} L_{11} (Fig. 1) when correctly adjusted. Adjustment is discussed in the text.



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.

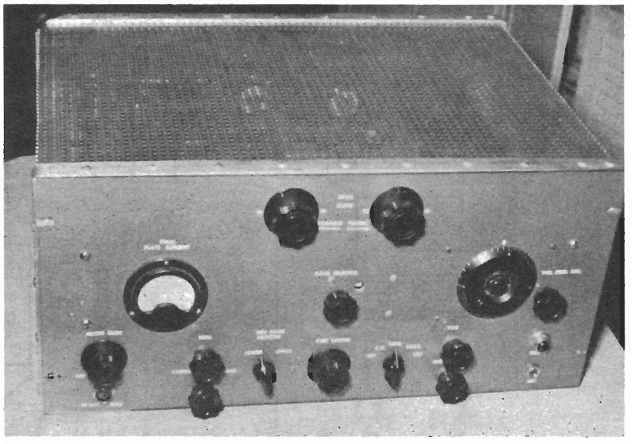
Exciter Coils

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.

» *A good example of the selective approach to building your own. Picking up one idea here, another there, the author comes out with a design that meets his own special requirements. You can do likewise.*



The panel is of standard rack width and 8¾ inches high. From left to right along the bottom are the microphone connector and audio gain control, the two carrier-balance controls, the sideband-selector switch, driver tuning control, mode switch, VOX and excitation controls, pilot lamp and key jack. The band switch is at the center of the panel, and the final-amplifier tuning and loading controls are at the top. The ARC-5 dial at the right controls the v.f.o.

Another Phasing-Type S.S.B. Exciter

RICHARD L. EVANS, K9YHT

After deciding to get my feet wet in s.s.b., previous designs were reviewed to determine what type of exciter might appeal to me most. The accompanying diagrams and photos indicate the result. The unit combines features of three previously described designs as well as a few variations of my own.

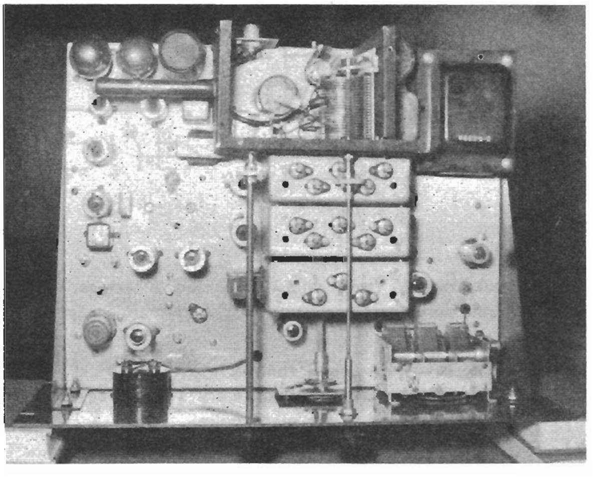
The circuit is basically that of the exciter described by K4EEU.[1] However, the beam-deflection-tube balanced modulator described by K2FF,[2] and the v.f.o. and VOX circuits of W6TEU's Sideband Package[3] have been inserted in place of the corresponding sections used by K4EEU. In addition, pi-network output has been substituted for the multiband tank arrangement used by both W6TEU and K4EEU. A 6CL6 drives the 6146 final. Sufficient output is obtained from the balanced modulator to drive the driver stage directly without the need for an intermediate stage. Reference should be made to the three articles mentioned for any details that are not made clear here.

From September, 1962, *QST*.

[1] Kelley, "A Phasing-Type Sidebander," page 80.

[2] Vance, "S.S.B. Circuits Using the 7360 Beam-Deflection Tube," page 29.

[3] Bigler, "A Sideband Package," page 59.



Top view of K9YHT's s.s.b. exciter. The sideband generator occupies the left-hand side of the chassis, with the remainder devoted to the v.f.o., band-heterodyning circuits, final amplifier and power supply.

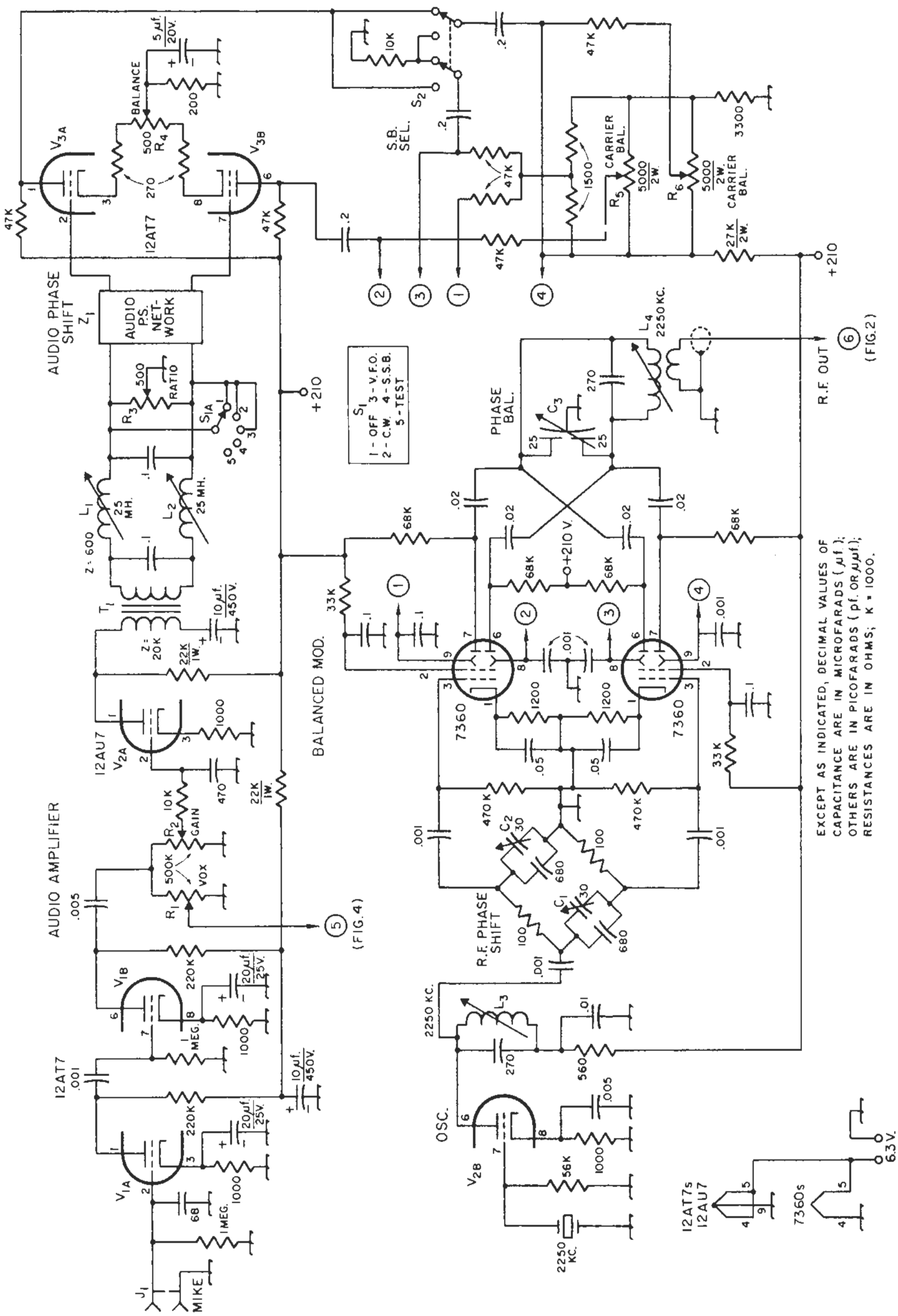
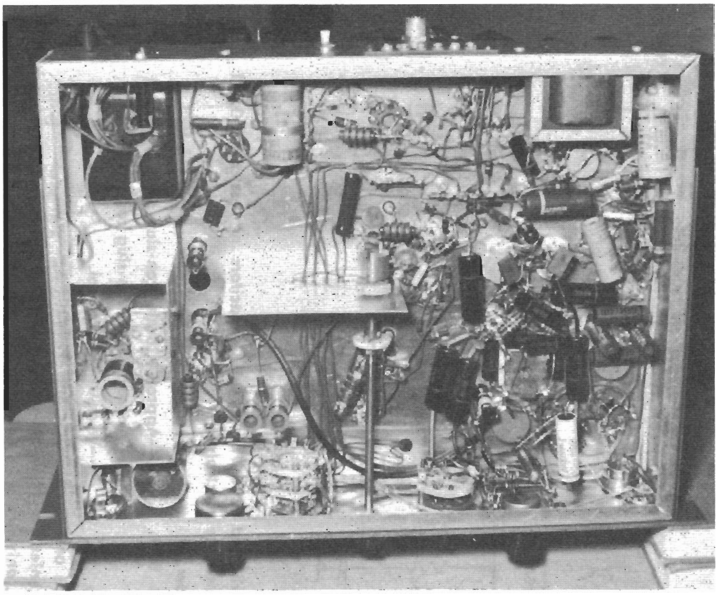


FIG. 1—Circuit of K9YHT's sideband generator. Resistors are ½ watt unless indicated otherwise. Fixed capacitors of less than 0.001 μf. should be mica or stable ceramic; others may be disk ceramic except those marked with polarity which are electrolytic.

C_1, C_2—Miniature var. (Johnson 30M8/160–130).
C_3—Differential capacitor (Johnson 25LA15/167–32).
J_1—Microphone connector.
L_1, L_2—4–30-mh. iron-slug coil (adjust to 25 mh.).
L_3, L_4—18μh. iron-slug coil; L_4 has 3-turn link.
R_1, R_5, R_6—Linear-taper control, R_2—Audio-taper.
R_3, R_4—Wire-wound control.
S_1—Four-pole 5-position rotary switch (see Figs. 2, 3 and 4 for other poles).
S_2—D.p.d.t. rotary switch.
T_1—Tube-to-line transformer (Triad A-53X).
Z_1—Audio phase-shift network (B & W type 350).



Bottom view showing the v.f.o. coil and trimmers in the shielding compartment in the lower left-hand corner. The loosely coupled coils between the two converters are just above the mode switch. The driver tuning capacitor is mounted on the bracket at the center of the chassis.

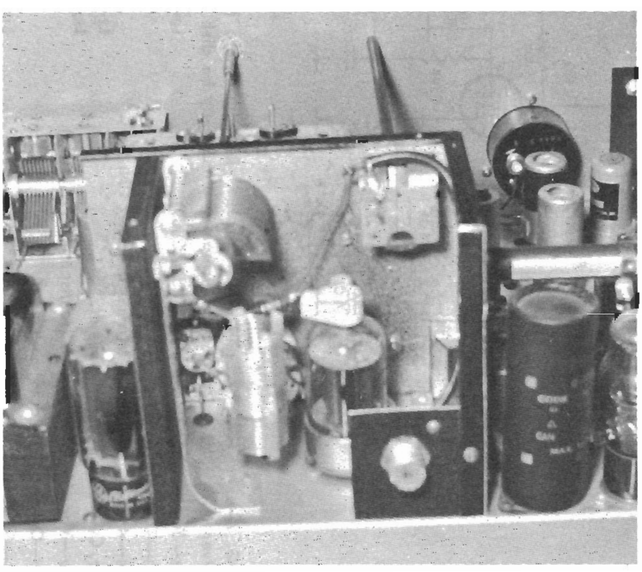
Construction

The unit is assembled on a $17 \times 13 \times 3$-inch chassis. Many of the components used were from surplus or old TV receivers as suggested in the reference articles. Suitable standard-component substitutes are listed here along with the diagrams. The low-level r.f. coils were wound as described in the previous articles. Approximate inductance values are given here so that those who prefer not to wind their own can make a selection from available ready wound coils in the instances in which such coils make acceptable substitutes. Power-supply components were salvaged from an old TV set.

Referring to the photographs, the left-hand side of the chassis is devoted to the sideband generator. The two VR tubes, a filter capacitor and the voltage-divider resistor R_9 (Fig. 3) are mounted at the rear. Immediately in front of the resistor are the 12AT7, 6AU6 and relay of the VOX system. (The transformer in front of the relay is the audio output transformer, T_4.)

In the audio section, strung out along the left-hand edge of the chassis are the 12AT7, 12AU7 (crystal alongside), T_1, and the plug-in audio phase-shift network with the 12AT7 audio output tube to the right.

The r.f. section occupies the right-hand half of the chassis. The three-section v.f.o./first-converter tuning capacitor, mounted against the



Close-up shot of the 6146 final amplifier. The pi-network coil is mounted vertically alongside the tube. Tuning and loading capacitors are above on the front wall of the enclosure. The tube to the left is the power rectifier. VR tubes are to the right.

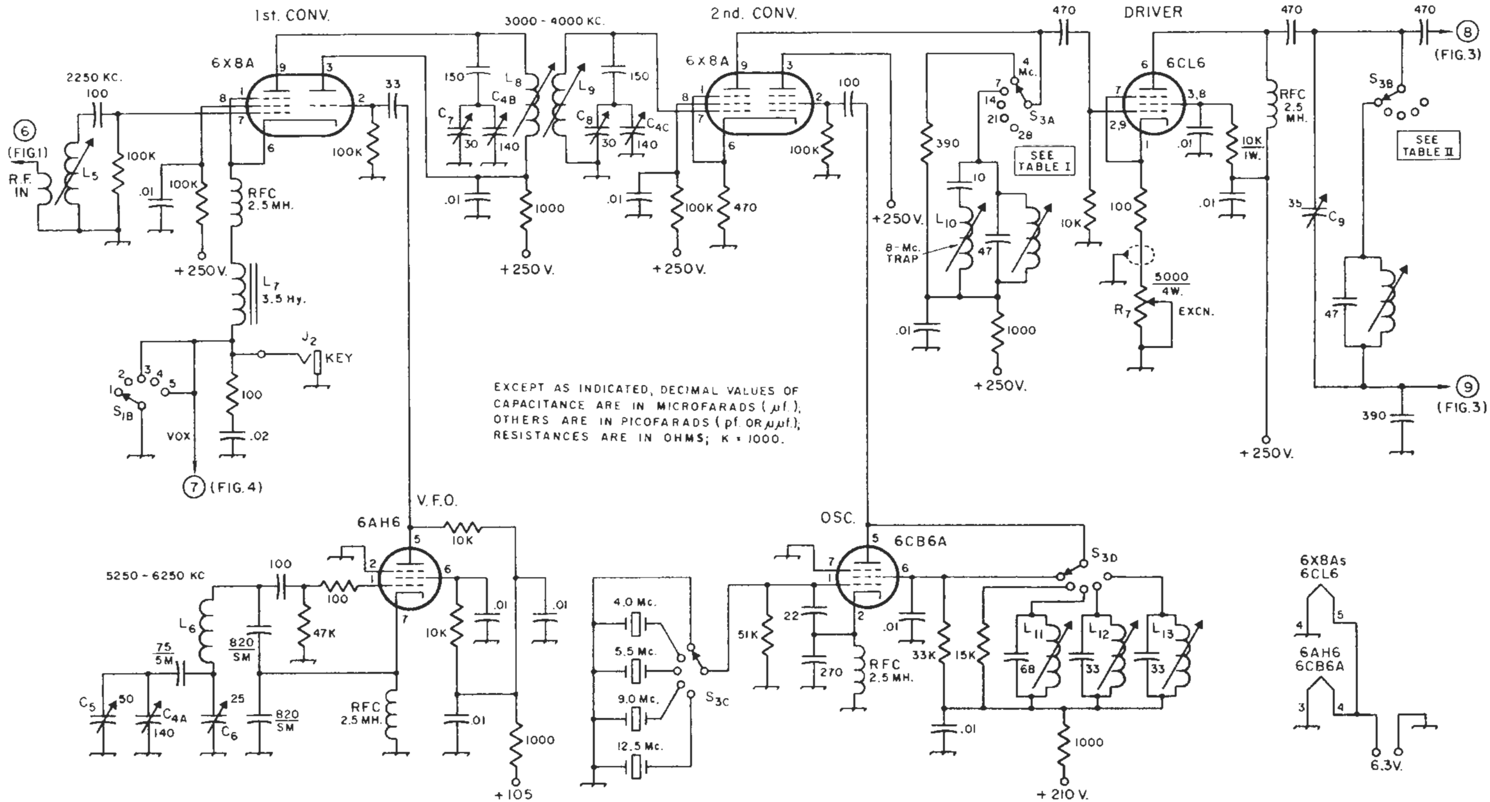


FIG. 2—Band-converter and driver circuits. Resistances are in ohms and resistors are ½ watt unless indicated otherwise. Fixed capacitors of less than 0.001 μf. are silver mica (SM), mica or stable ceramic; others may be disk ceramic except those marked with polarity which are electrolytic.

C_4—Triple-gang air variable (Bud MC889).

C_5—10–50-pf. zero-temp. trimmer (CRL 823-BZ).

C_6—4–25-pf. neg.-temp. trimmer (CRL 822-CN).

C_7, C_8—3–30-pf. mica trimmer.

C_9—Variable air capacitor (Johnson 35R12/149–2).

J_2—Open-circuit jack.

L_5—Same as L_4.

L_6—Approx. 20 μh.: 30 turns No. 26, 1-inch diam., 1 inch long (National XR-60 ceramic form with slug removed).

L_7—3.5-hy. filter choke (Stancor C-1080).

L_8, L_9—40 turns No. 34 enam., close-wound. on ½-inch iron-slug form (National XR-50 form). Forms are mounted parallel, spaced ¾ inch center to center.

L_{10}—Approx. 40-μh. iron-slug coil; used on 7 Mc. only.

L_{11}—Approx. 2.5-μh. iron-slug coil (11 Mc.).

L_{12}—Approx. 1.8-μh. iron-slug coil (18 Mc.).

L_{13}—Approx. 1-μh. iron-slug coil (25 Mc.).

R_7—Wire-wound control.

S_1—See Fig. 1.

S_3—Six-pole 5-position ceramic rotary switch (see Fig. 3 for other sections).

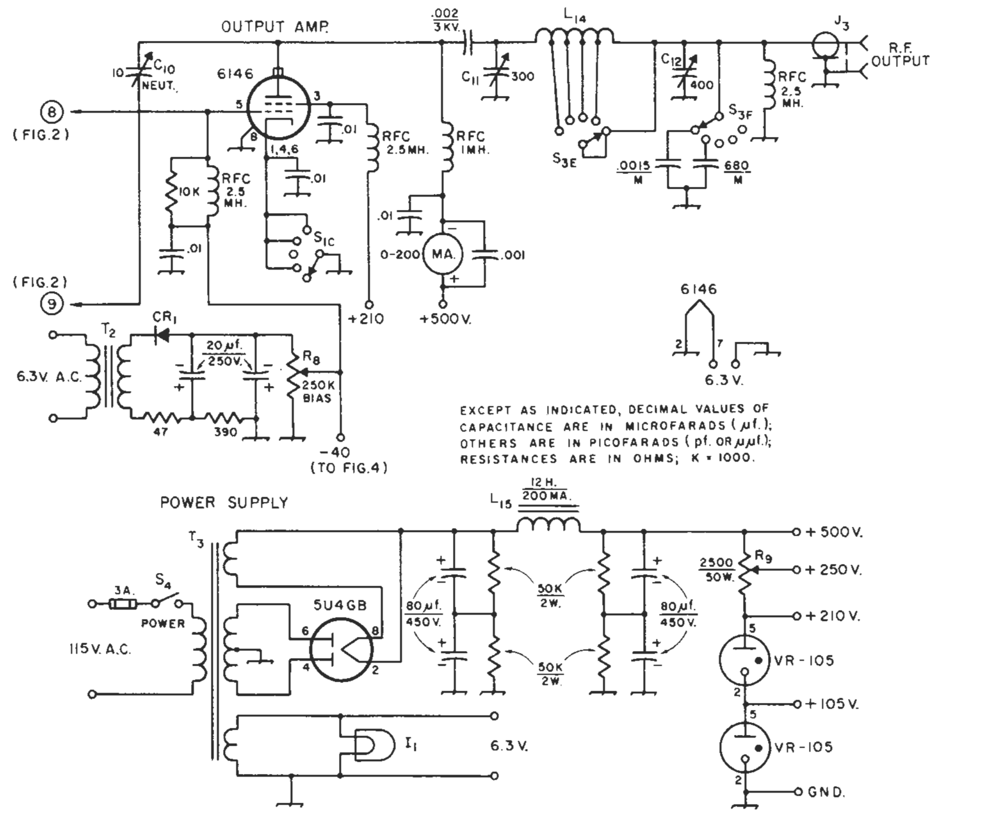


FIG. 3—Circuits of the output amplifier and power supply. Resistances are in ohms, and resistors are ½ watt unless indicated otherwise. Fixed capacitors marked wtih polarity are electrolytic; others are disk ceramic except those marked M which are mica.

C_{10}—Neutralizing capacitor (Bud NC-853, Hammarlund NZ-10).
C_{11}—Air variable, 1000-volt rating (National TMS-300).
C_{12}—Air variable, broadcast-replacement type.
CR_1—130-volt 20-ma. selenium rectifier.
I_1—6.3-volt lamp.
J_3—Chassis-mounting coaxial connector.
L_{14}—34 turns No. 18, 1-inch diam., 14 turns at 10 t.p.i. (C_{11} end), plus 20 turns at 20 t.p.i. (Illumitronics Pi-Dux 820-D-10) tapped at 7, 11, 14, and 21 turns from C_{11} end.
L_{15}—Filter choke (Chicago RC-12200).
R_8—Linear-taper control.
R_9—Slider adjustable.
S_1—See Fig. 1.
S_3—See Fig. 2.
S_4—S.p.s.t. toggle switch.
T_2—6.3-volt 1-amp. filament transformer.
T_3—Power transformer: 800 volts c.t., 200 ma.; 5 volts, 3 amp.; 6.3 volts, 5 amp. (Stancor PC8412).

right-hand end of the panel, is one found in the 3-6-Mc. ARC-5 Command receiver. The ceramic-form v.f.o. coil and trimmer capacitors are below deck in a shielding box. The 6AH6 is to the rear of the tuning capacitor, near the right-hand edge of the chassis.

TABLE I
Second-Converter Coils

Band (Mc.)	*C (pf.)*	*L (μh.)*
7	47	8
14	47	2
21	22	1.5
28	22	0.8

The 6CB6 crystal-oscillator, second-converter, and driver circuits are in individual shielding boxes near the center of the chassis. The boxes were salvaged from a surplus unit. They have feedthrough insulators in the bottoms which extend below the chassis, making connections convenient. The slug-tuned coils associated with each stage are mounted in the top of the box in each case, where they are readily accessible for adjustment.

The crystal-oscillator box is the one nearest the panel, with the 6CB6 mounted in front.

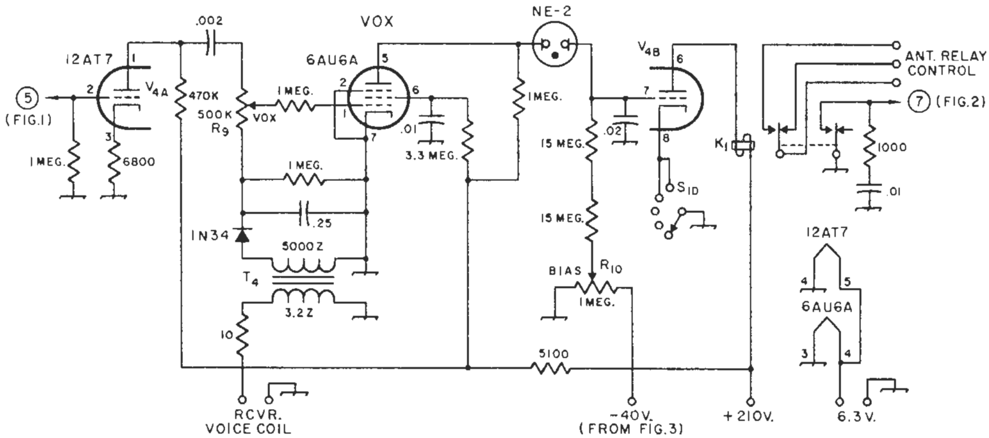


FIG. 4—VOX control circuit. Capacitances are in μf. Resistances are in ohms and resistors are ½ watt unless indicated otherwise. Capacitors are paper or ceramic.

K_1—5000-ohm d.p.d.t. relay.
R_9, R_{10}—Linear control.
S_1—See Fig. 1.
T_4—Output transformers: 5000 ohms to voice coil.

Two octal tube sockets set in the left-hand end of the compartment serve as receptacles for the four crystals. The 6X8 first-converter tube is mounted near the right-hand end of this box.

The central box contains the second-converter output circuits, with the second 6X8 mounted near the left-hand end. The output circuits of the driver stage are in the third box, and the 6CL6 is mounted close to the left-hand end.

The final-amplifier components are mounted in a larger enclosure at the rear of the chassis. The pi-network coil is mounted with its axis parallel to the 6146. The tuning and loading capacitors are mounted above on the front wall of the compartment. Long shafts extend to the controls on the panel.

The individual ceramic wafers of the band switch are mounted within the compartment of the circuit with which each is associated. There are two wafers in the 6CB6 oscillator box, and two in the final-amplifier compartment. The two remaining enclosures have one wafer each. The wafers are mounted so that the shaft holes will line up accurately when the compartments are mounted on the chassis. Instead of the usual pointer for indicating the switch positions, the switch shaft is fitted with a dial whose band markings appear through a small hole drilled in the panel.

The cabinet is homemade of aluminum angle and perforated sheets fastened together with rivets.

Adjustment

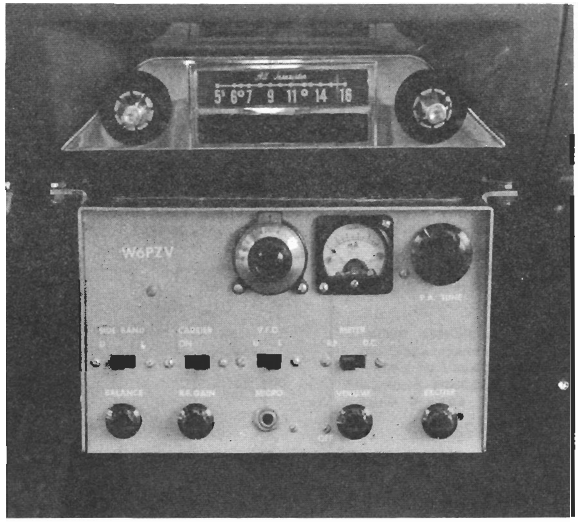
Alignment and adjustment procedures are covered in the articles previously mentioned. Carrier and unwanted-sideband suppression were easily obtained and appear to be entirely adequate, although no actual measurements have been made. Initial on-the-air results have been most gratifying and signal reports have been universally good. My thanks to the authors of the reference texts for their ideas and assistance.

TABLE II

Driver Coils

Band (Mc.)	*C (pf.)*	*L (µh.)*
3.5	47	19
7	68	4
14	33	1.5
21	22	0.7
28	47	0.3

» *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-Mc. Mobile S.S.B. 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 exciter [1] 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.

From August, 1963, *QST*.

[1] Isaacs, "Filter-Type Sidebander," page 71.

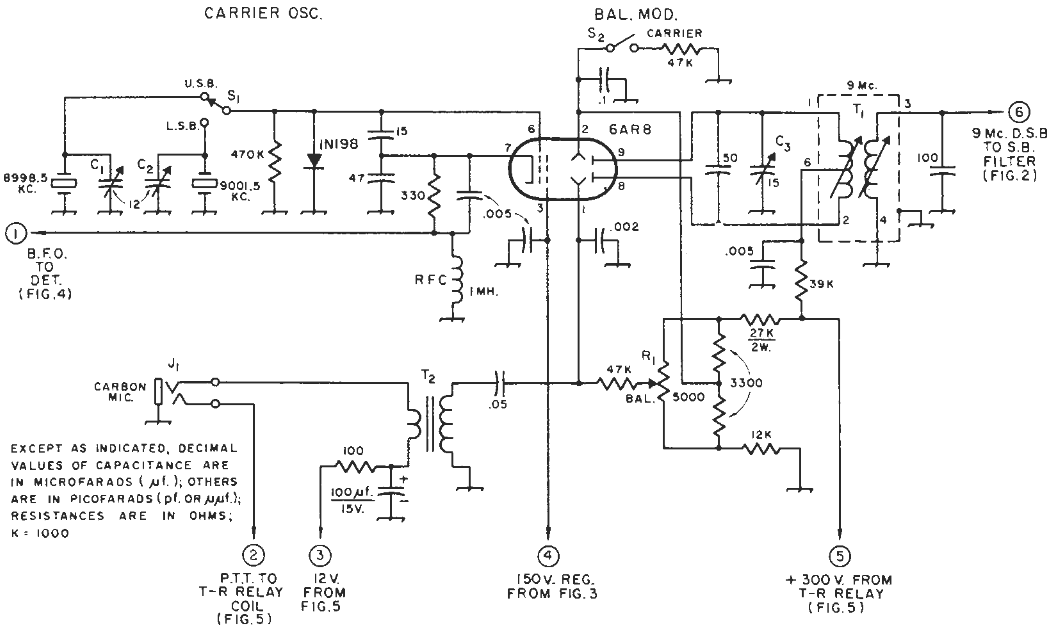


FIG. 1—Carrier-oscillator and balanced-modulator circuits. Fixed capacitors of decimal value are ceramic; others are silver mica, except polarity marking indicates electrolytic. Unless otherwise specified, resistors are ½-watt composition. See Fig. 5 for heater connections.

C_1, C_2—3–12-pf. ceramic, zero temp. coefficient (Centralab 822-FZ).

C_3—Air trimmer (Johnson 160–107, or similar).

J_1—Three-circuit jack.

R_1—Linear control.

S_1—S.p.d.t. slide switch (Carling S60B or equivalent).

S_2—S.p.s.t., or same as S_1.

T_1—10.7-Mc. f.m. discriminator transformer; see text (Meissner 17–3494).

T_2—Microphone transformer, 200 to 500K (UTC "Ouncer" O–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 (C_1 and C_2) 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 T_1 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 T_1 through 0.001-μf. 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

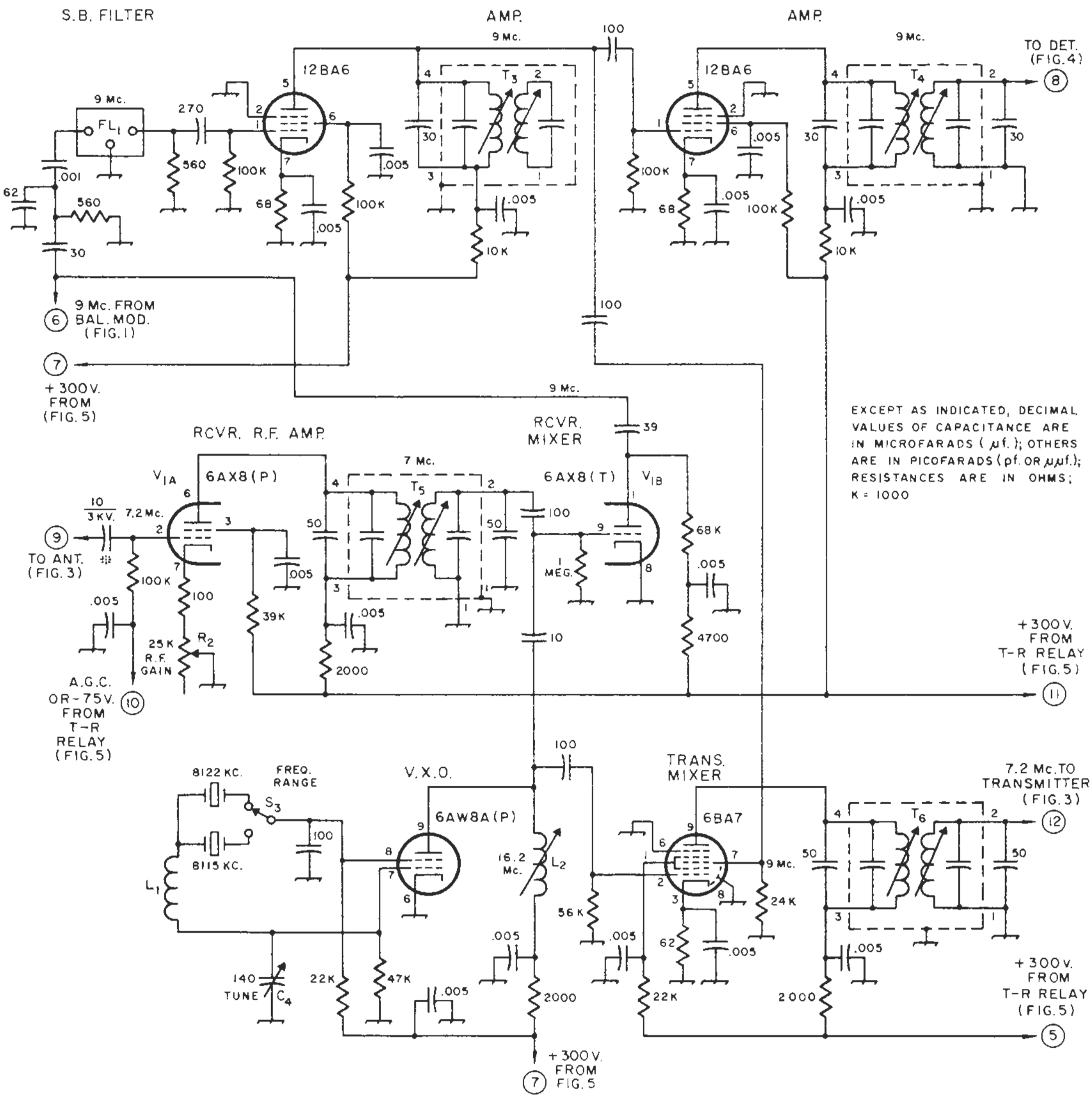


FIG. 2—I.f. and frequency-conversion circuits. Fixed capacitors of decimal value are ceramic; others are silver mica, except * indicates ceramic. Resistors are ½-watt composition. See Fig. 5 for heater connections.

C_4—Air-trimmer variable (Hammarlund APC-140-B).

FL_1—9 Mc. crystal filter (McCoy 32B1).

L_1—20 μh.—Approx. 45 turns No. 30 enam., close-wound on form taken from National R-33 100 μh. r.f. choke. See text.

L_2—18 turns No. 22 enam., close-wound on ⅜-inch iron-slug form.

R_2—Linear control.

S_3—Same as S_1.

T_3, T_4, T_5, T_6—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 volage for the microphone is taken from the 12-volt circuit through an *RC* filter. I have

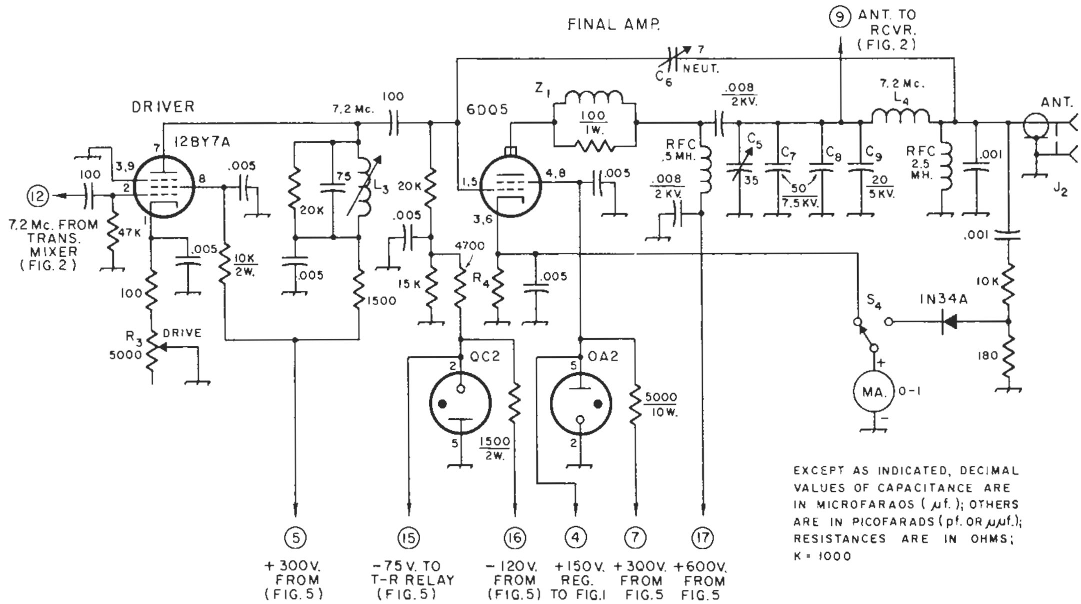


FIG. 3—Transmitting driver and final-amplifier circuits. Antenna connection to receiving r.f. amplifier is taken from the input side of the transmitting pi network. On transmit, the t.r. relay (K_1 Fig. 5) applies —75 volts bias to the receiving r.f. amplifier to protect it from the transmitter signal. Capacitors of decimal value are ceramic; others are silver mica.

C_5—Double-spaced midget variable (Hammarlund MC-35-SX or similar).

C_6—NPO ceramic trimmer, 1.5–7 pf. (Centralab 822-EZ).

C_7, C_8—7500-volt ceramic (Centralab 850S-50Z).

C_9—5000-volt ceramic (Centralab 853-20Z).

J_2—Coaxial plug-in connector (UG1051/U).

L_3—24 turns No. 26 enam. on 3/8-inch iron-slug form.

L_4—18 turns No. 20, 1 1/2 inches long on ceramic form 1 1/8 inches in diam.

R_3—Wire-wound control.

R_4—200 times meter shunt.

S_4—Same as S_1.

Z_1—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-kc. 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

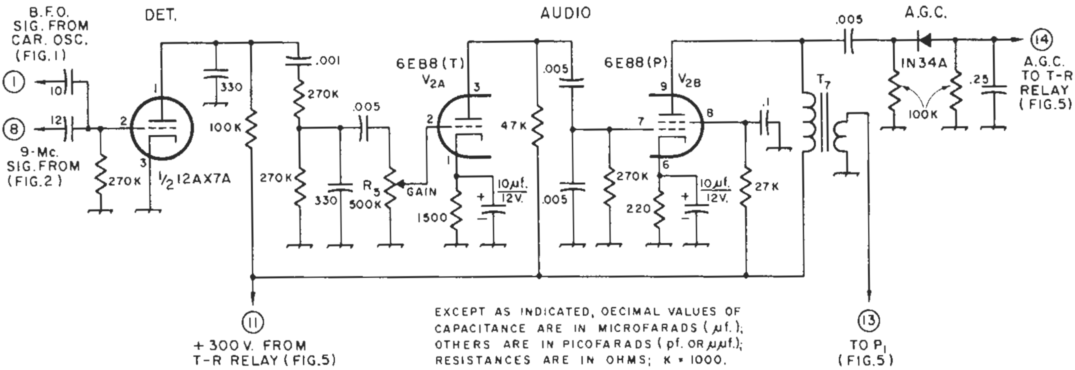


FIG. 4—Detector and receiving audio circuits. The b.f.o. signal is obtained from the carrier oscillator. Rectified audio supplies a.g.c. for the receiver r.f. stage. Capacitors of decimal value are ceramic or paper; others are mica, except polarity markings indicate electrolytic. Fixed resistors are ½-watt composition. See Fig. 5 for heater connections.

R_5—Audio taper control.

T_7—Output transformer, 10,000 ohms to 7 ohms (Thordarson 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.

L_1 is made by taking a 100-μh. 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 μh.

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 C_5, C_7, C_8 and C_9 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 R_3 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 (R_4) 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

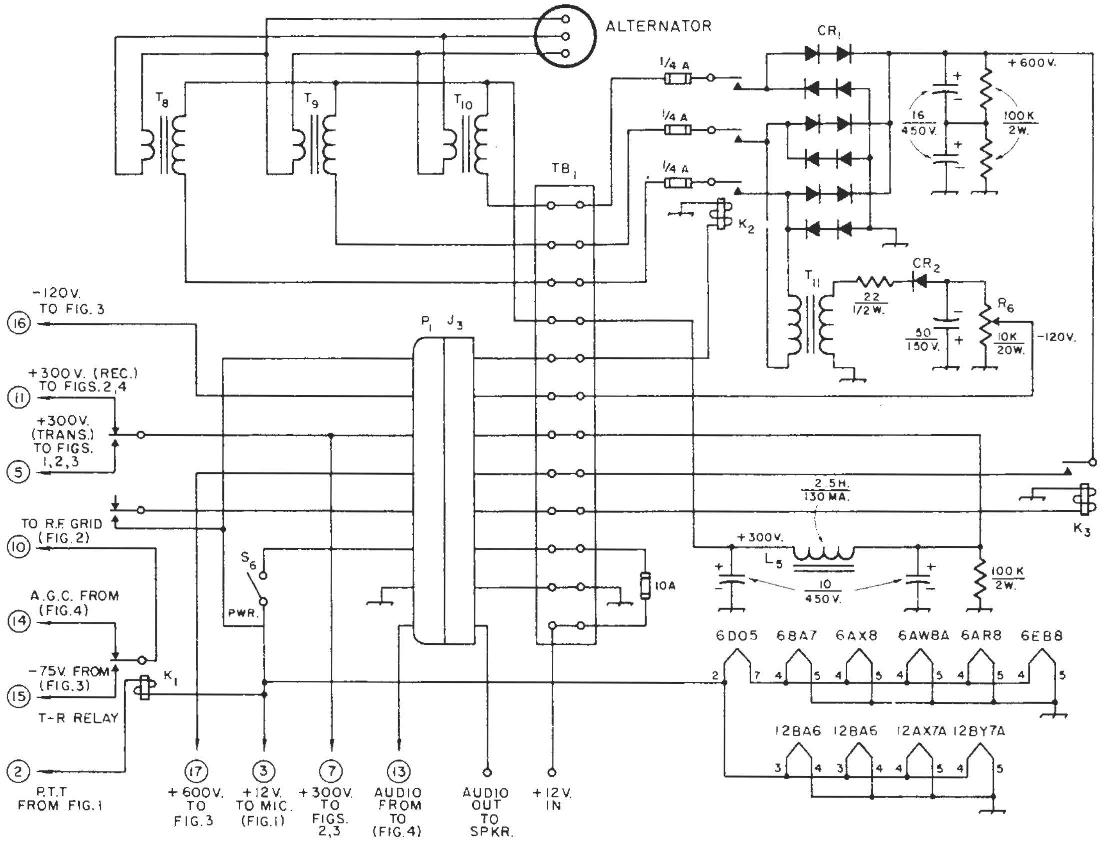


FIG. 5—Power and control circuits. Capacitances are in μf., and resistances in ohms. Capacitors are electrolytic.

CR_1—12 silicon units, 400 p.i.v., 750 ma. each unit.
CR_2—Silicon rectifier, 400 p.i.v., 750 ma.
J_3—Octal socket.
K_1—Subminiature 12-volt 3-pole double-throw d.c. relay (Potter & Brumfield MG17D, one pole not used).
K_2—12-volt 3-pole double-throw d.c. relay (Potter & Brumfield KA14DY).
K_3—12-volt s.p.d.t. d.c. relay (Potter & Brumfield KA5DY).
L_5—Filter choke (Stancor C2303 or equivalent).
P_1—Male octal plug.
R_6—Slider adjustable.
S_6—S.p.s.t. attached to R_5, Fig. 4.
T_8, T_9, T_{10}—See text.
T_{11}—230-volt 40-ma. power transformer, 115-volt primary used as secondary, center tap on 230-volt winding not used (Triad R-29A or similar).
TB_1—Terminal board.

the same as the frequency used on transmit, the sideband relationship is the same for both modes. That is, when S_1 is set for l.s.b., the lower sideband will be transmitted and received. On receive, the transceiver is tuned until the received voice signal sounds natural. The transmitting frequency will then be very close to that of the received signal.

Output from the audio amplifier which follows the detector is adequate for mobile operation. I have a switch mounted under the dash which transfers the speaker from the broadcast receiver to the transceiver.

A.g.c. voltage is obtained by rectifying the audio drop across a resistor shunting the output-amplifier load. The rectified voltage appears across a 100K resistor shunted by a 0.25-μf. capacitor. A larger capacitor at this point would provide smoother operation, but I did not have space for it. A.g.c. is applied to the r.f. stage (Fig. 2) only, but it is effective and eliminates the necessity for reaching for the manual gain control when a strong signal hits the front end.

Power Supply

Fig. 5 shows the circuit of the power supply and control system. The power-supply is a little unusual, and should be of interest to anyone contemplating a mobile sideband transceiver. I use the alternator which is standard equipment in my car. It is rated at

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

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.

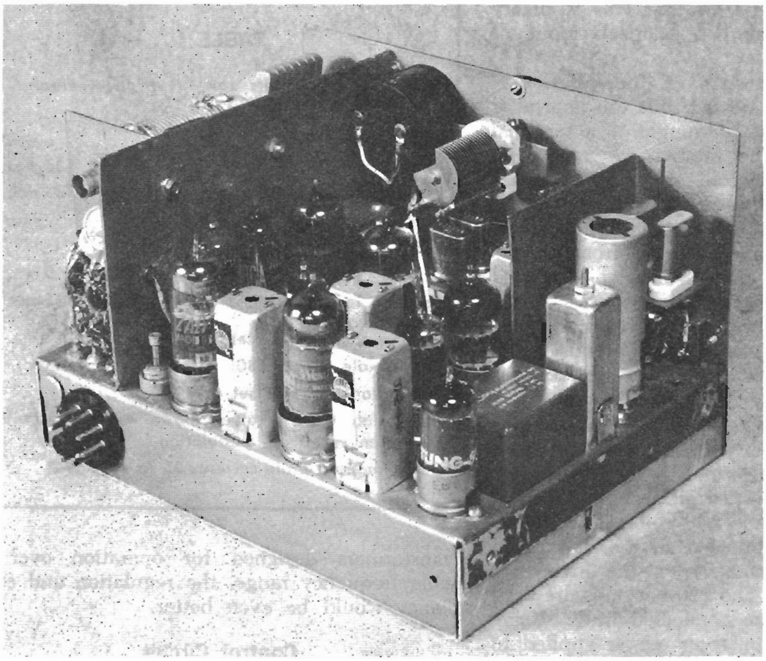
Control Circuit

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 6⅛ 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.

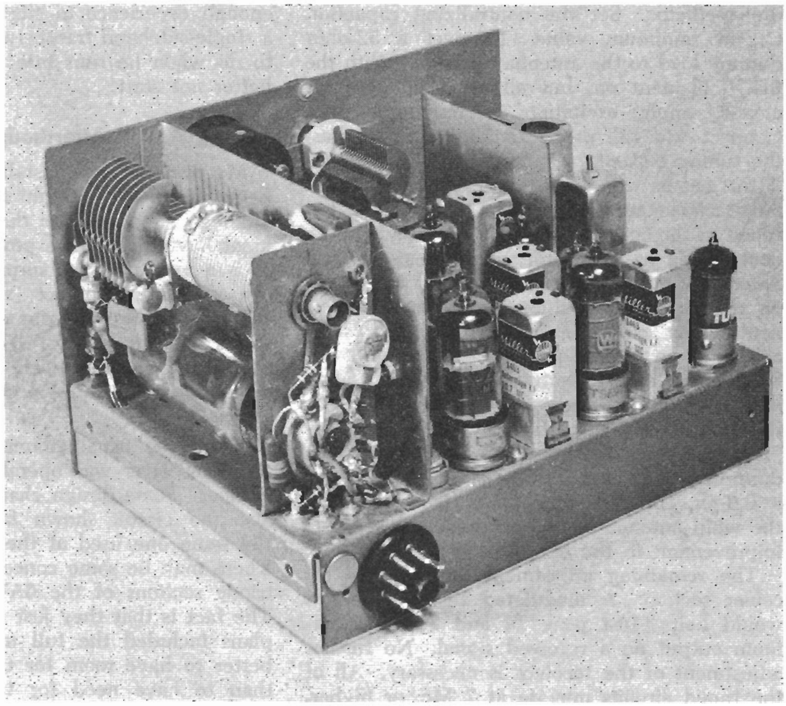
The cabinet, which included the chassis, was purchased in the unpainted condition. All of the holes were cut and the ventilating grilles riveted on, and then all of the metal pieces were painted with hammer-tone paint from an aerosol can. It is not difficult to get a good-looking paint job this way, 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 repeak 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 neutralizating 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-μh. 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.

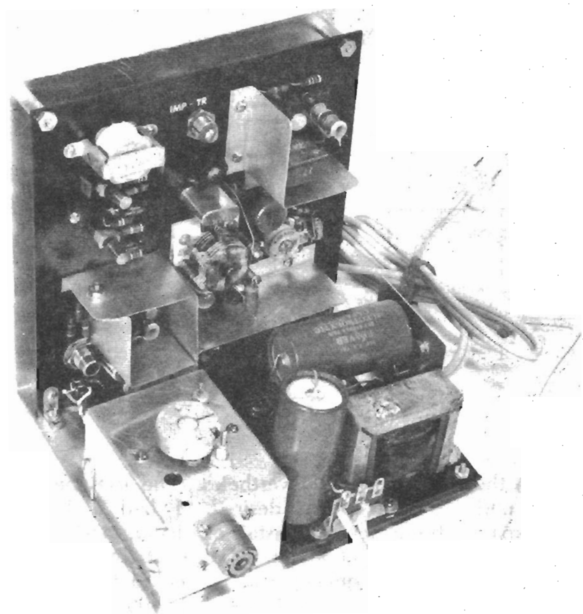
Afterthoughts

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.

» *For the experimenter who likes to go solid-state—a transistor version of the popular little "Imp" exciter.*



The Imp-TR, built in experimental fashion, is assembled on an L-shaped piece of aluminum. The power supply and the 21-Mc. linear amplifier are separate units, mounted side by side on the horizontal part of the L in this view.

The "Imp-TR"

JOSEPH S. GALESKI, JR., W4IMP

Transistors, as well as other solid-state devices, have been greatly improved in recent years Units are available that are capable of handling high frequencies at a reasonable power level. While some are still quite expensive, costs are dropping rapidly.

Before building an all-transistor multiband s.s.b. exciter, I felt that it would be necessary to gain experience. Twenty-one megacycles was chosen to be the output frequency of a simple filter-type exciter. If I could make it perform well on this frequency, then 10 through 80 meters in a bandswitching v.f.o. rig should not be too great an undertaking at a later date.

The Circuit

The general plan of the "Imp" exciter [1] with an added stage of audio was followed by replacing each of the tube sections with a transistor. Interpretation of the circuit diagram is quite easy if one regards the collector as analogous to the plate of a vacuum-tube triode, the base as the grid, and the emitter as the cathode.

In Fig. 1, crystal oscillator Q_1 feeds a carrier to the balanced modulator through the link of L_1, which serves as a convenient means of coupling to the diodes. R_1 and C_1 are the carrier balance controls. If C_1 does not add to the carrier suppression obtainable from R_1 alone, it should be moved to the other side of the potentiometer. The 4700-ohm resistor from the L_2 link to chassis provides the necessary return path for the diodes.

Two stages of audio supply ample voltage to the balanced modulator. A transformer was originally used as a matching device for a high-impedance crystal microphone, but was later abandoned in favor of the circuit shown. It worked, but generated objectionable noise. The photographs were taken before this change was made. A step-down output transformer is not used because of the relatively low collector impedance of Q_3.

A few surplus type FT-243 crystals marked

From December, 1961, *QST*.

[1] Galeski, "The 'Imp'—a 3-Tube Filter Rig," page 54.

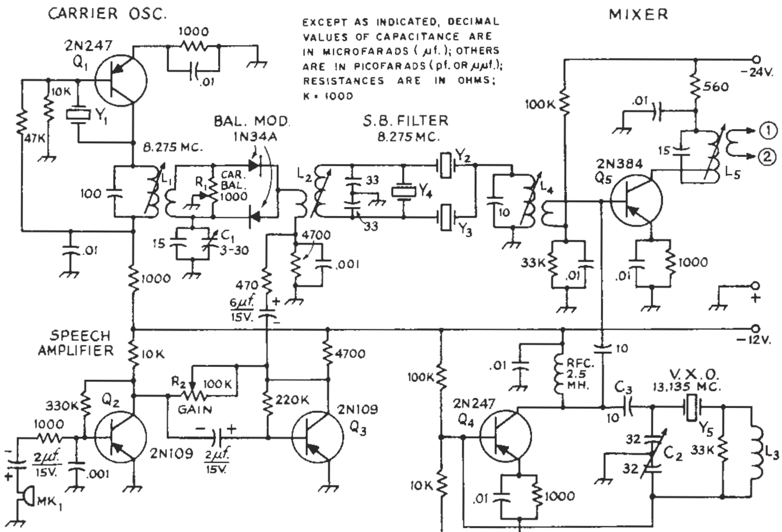
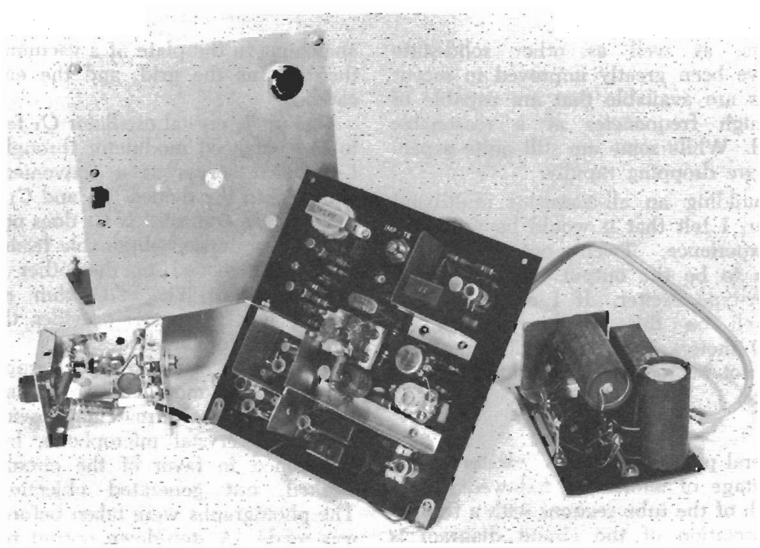


FIG. 1—(Continued on facing page.)

8275 kc. were obtained for the filter. As in the original "Imp," two were selected for Y_2 and Y_4 that were a couple of hundred cycles apart in frequency. The lower was used at Y_4 as the shunt crystal. A third was ground up about 2 kc. for Y_3 and a fourth was later brought down a bit with a pencil mark for Y_1. The alignment procedure and the result to be expected were fully discussed in another article,[1] and need not be repeated here. My methods and the resulting passband were essentially the same.

A VXO with transistors works well; however, output falls off as the crystal is pulled lower in frequency. Nevertheless, the variable feature is still to be desired. I had a 13.135-Mc. fundamental crystal on hand, and with this circuit get about 20-kc. shift. Notice that the filter frequency was chosen so that the output sum frequency is about 21,400 kc. The 10-pf. capacitor, C_3, serves to prevent loading of the crystal by the transistor. Without it, the available frequency shift is much smaller. Of several high-frequency transistors, the 2N247 gave the greatest frequency coverage. W6BAF suggested that a toroid be used at L_3 for the sake of compactness, so a quarter-inch slice of a ½-inch-diameter powdered-iron tuning slug was drilled



Sections comprising the complete transmitter have been separated in this photograph to show the individual assemblies. The "panel" is the part of the aluminum L containing the slide switch and microphone connector.

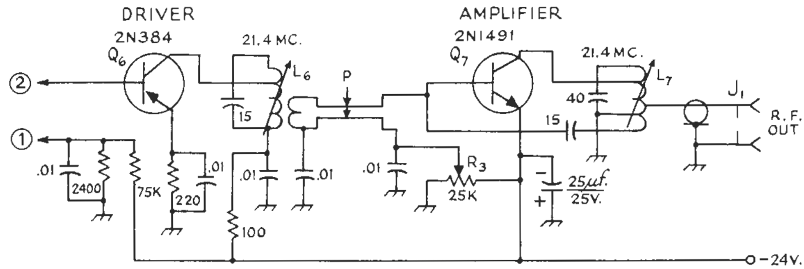


FIG. 1 (Continued)—Circuit of the Imp-TR 21-Mc. transistor s.s.b. exciter. Resistances are in ohms; fixed resistors are ½ watt. Capacitors with polarities marked are electrolytic; other fixed capacitors are ceramic. All coils except L_3 are wound on slug-tuned ceramic forms, ¼-inch diameter (CTC type PLS6/E or similar), with windings doped or lacquered. The lead from L_6 link to Q_7 is a twisted pair made from hookup wire.

C_1—3–30-pf. ceramic trimmer shunted with additional capacitance, if necessary, to enable balancing out carrier.
C_2—Split-stator, 32 pf. per section, surplus. Any small unit with 25 pf. to 50 pf. per section may be substituted.
C_3—10 pf.; see text.
J_1—Coaxial connector, chassis mounting.
L_1—25 turns No. 28 enam., close-wound; link 7 turns. (All links wound over cold end).
L_2—40 turns No. 34 enam., close-wound; link 15 turns.
L_3—33 turns No. 28 on homemade toroid core. See text.
L_4—40 turns No. 34 enam., close-wound; link 6 turns.
L_5—24 turns No. 28 enam., close-wound; tapped 5 turns from hot end; link 4 turns.
L_6—Same as L_5 with 3-turn link.
L_7—16 turns No. 28 enam., close-wound; tapped at 4, 8, and 12 turns from hot end.
MK_1—Dynamic microphone, low impedance.
R_1—1000 ohms, linear taper, miniature (Lafayette VC-32 or similar).
R_2—0.1 megohm, audio taper, miniature.
R_3—25,000 ohms, linear taper, miniature.
Y_1-Y_4 inc.—8275-kc. crystals (surplus FT1243).
Y_5—13.135-Mc. fundamental.

with a ¼-inch hole to make the necessary doughnut-shaped form.

Problems that were anticipated with the simple mixer fortunately did not develop. The base of Q_5 is fed from a link on the filter coil L_4 and from the VXO through a small capacitor. I realize that the match here is not optimum, but results seem adequate. Among other possibilities, we might include replacing the r.f. choke in the VXO by a tuned circuit and feeding either the base or the emitter of the mixer from a link. The collector of Q_5 is tapped down on the tank coil, L_5, for a better impedance match.

The first r.f. amplifier, Q_6, is similar to the mixer in circuitry except that it is driven from a single r.f. source. It is interesting to note here that the r.f. voltage developed across L_6 is about 6 to 8 volts r.m.s. as measured on a v.t.v.m. This is enough to drive some of the smaller vacuum tubes.

In my opinion, the key to a successful transistorized exciter is enough power output to drive a high-power amplifier without having to resort to an intervening heat-producing vacuum-tube stage. The 2N1491 has a typical alpha-cutoff frequency of 250 Mc. and a free-air heat dissipation of one-half watt. This can be increased to three watts with a heat sink. Maximum collector-to-emitter voltage is 30 volts.

Several circuits were tried for the 2N1491 power amplifier. My greatest problem was parasitic oscillation. (This will sound familiar to those who have experience in building vacuum-tube s.s.b. exciters!) As usual, the simplest circuit seemed to perform the best. An output link on L_7 was abandoned in favor of a tap on the coil because stability was better.

Power Supply

To take full advantage of the 2N1491, a d.c. supply of 24 to 30 volts for the collector was necessary.[2] Twelve volts, well filtered, was needed for the low-level stages. The available literature was searched and it was decided to

[2] This is pushing the transistor rating a bit, since the 30-volt collector-emitter maximum applies to *peak*, not d.c. supply, voltage. The peak voltage is the sum of the d.c. and r.f. voltages, and may approach twice the d.c. voltage. It would be safer to use no more than 15 to 20 volts d.c. This of course reduces the available output.—*Editor.*

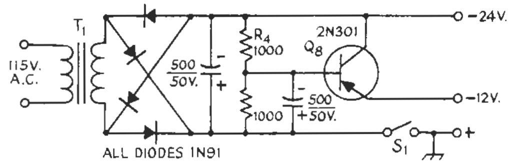


FIG. 2—Circuit of the power supply. Capacitors are electrolytic; capacitances are in µf. Resistors are ½ watt.

R_4—See text.
S_1—S.p.s.t. slide switch.
T_1—Small filament transformer with secondary removed and rewound with No. 28 enam. to give 18 to 20 volts a.c.; see text.

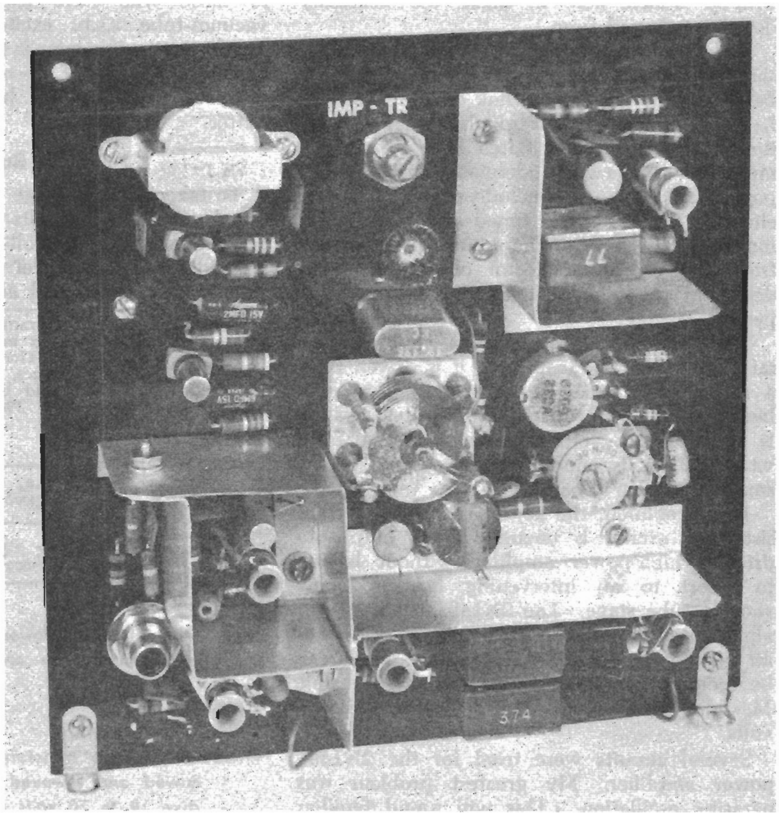
use a full-wave bridge rectifier with an electronic filter.[3] A transformer with about a 20-volt secondary was needed, so a new secondary was wound on a small filament transformer. The full voltage output of the first filter capacitor in the circuit shown in Fig. 2 is used for the mixer and amplifiers. The voltage divider supplying the base of the 2N301 determines the value of the low voltage. R_4 was adjusted to give 12 volts with the entire exciter and the amplifiers connected. The effective capacitance across the 12-volt output is equal to the current gain of the transistor multiplied by the base capacitor, and this is in the order of 25,000 microfarads.[3] No hum was evident when the supply was connected to my transistorized receiver. The regulation was satisfactory and the use of zener diodes was considered to be unnecessary. Although the output voltages are not completely cut off with S_1 open, the system works satisfactorily for send-receive, preventing the transmitter from "hanging on" while the filter capacitors discharge. A double-pole switch breaking the −12 and −24-volt leads could be substituted.

Construction

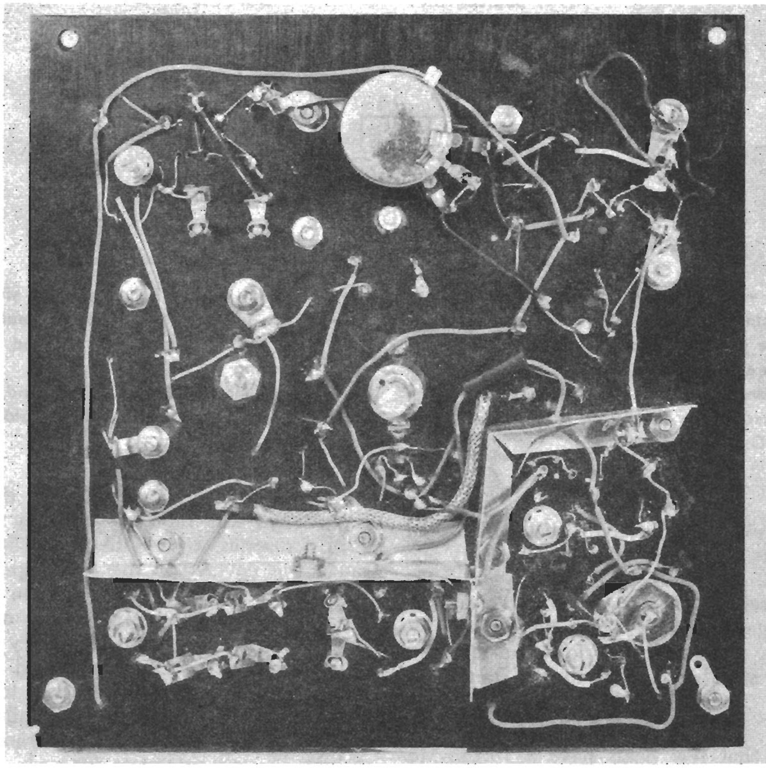
Transistors seem to be associated with printed circuits, but printed or etched circuits are not too practical when only one of a kind is to be constructed. I liked the form and appearance of etched-circuit construction, so for no other reason I used an "unprinted" printed circuit. A 6 × 6-inch square of 1/16-inch bakelite board was obtained and the general layout determined. A mounting border was marked off all the way around, and the remainder of the board divided into nine equal squares. The original thought was to use one square for the carrier oscillator, one for the balanced modulator, two for the filter, and so on. It didn't quite work out that way, but the intention is evident from the photographs. All parts, including transistor sockets, were mounted by drilling small holes for the leads and inserting them from the top of the board. Wiring was done on the underside. As each stage progressed, thin aluminum partitions were located where it was thought they might be helpful in isolating the various stages. All partitions were connected to a common ground. Each stage was tested as soon as it had been constructed.

The power amplifier was built into a small Minibox and is not unusual in construction. The extra holes evident in the photographs were the result of the fact that many circuits have been tried. A heat sink was made for the 2N1491 by drilling out a piece of aluminum to fit the case and tightened by using the slot

[3] *Motorola Power Transistor Handbook,* first edition, p. 154.



Close-up of the r.f. and audio assembly. The audio section is at the upper left. (As explained in the text, the audio transformer in the upper-left corner is no longer used.) The mixer and driver are surrounded by the shielding at the lower left. The crystal-controlled carrier oscillator is in the upper right corner with the balanced modulator immediately below it. The sideband filter is along the lower edge. Tuning capacitor and crystal in the center are part of the VXO circuit.



The "wrong" side of the mounting board doesn't show after the transmitter is completely assembled, being held just behind the panel by short mounting pillars. Wiring is direct, with no attempt at making it pretty.

and screw arrangement which can be seen in the illustrations.

A piece of aluminum sheet 6 by 12 inches was bent into an "L" shape. The exciter board was mounted vertically with the shaft of C_2 brought through the front panel. The carrier-balance control was cut off flush with the panel and slotted. The amplifier was mounted on the rear section along with the power supply that was constructed to fit the remaining space.

The power supply was mounted on a small piece of bakelite, which provided a convenient method of isolating an aluminum plate for the power transistor heat sink so that no sockets or insulating washer had to be used. Small clips were used for power-supply connections to facilitate disassembly.

Adjustment

A v.t.v.m. with an r.f. probe was used to check that the oscillators were working and that the output was of satisfactory magnitude. Don't fear that these tiny transistors will have so little output that you can't find it. Actually, 4 volts r.m.s. was measured at the collectors of Q_1 and Q_4.

The bias resistors for Q_1, Q_2, Q_3 and Q_4 are chosen to give about 1-ma. resting current. In the case of Q_1 and Q_4, a voltmeter across the emitter resistors provides an easy way to monitor the emitter current; one volt across the 1000-ohm resistor indicates 1 milliampere of current. The resting currents of the mixer, Q_5, and amplifier, Q_6, should be 1½ to 2 ma. Final-amplifier resting current is adjusted with the potentiometer, R_3, to 12 ma. Caution should be used to insure that the maximum emitter-to-base voltage rating (1 volt) is not exceeded when adjusting R_3. Be sure the control arm is at the emitter end before applying voltage to the amplifier. I strongly advise metering the 2N1491 at all times until you are sure that you have the circuit behaving itself.

Tune-up was done on the board first. The signal was located in the receiver and the carrier nulled. An audio tone of about 1500 cycles was fed to the audio input and the receiver's S meter watched for peak output while L_5 and L_6 were tuned. The amplifier was then connected to a dummy load and a similar procedure followed for it. The current kicks up a little on voice peaks, to about one-watt input, and the output is approximately ½ watt. This is not enough to "drive" my Thunderbolt, but it does make the plate meter very nervous, indicating several hundred watts input to the 4-400's.

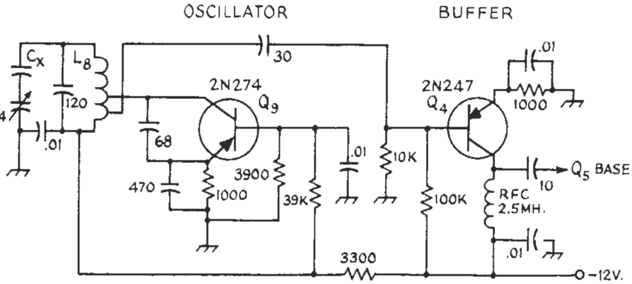


FIG. 3—The v.f.o. circuit. This substitutes for the variable crystal oscillator circuit shown in Fig. 1. Decimal value of capacitance in μf.; others are in pf. Resistances are in ohms, resistors are ½ watt. By-pass capacitors (0.01 μf.) are ceramic; others are silver mica except for the 10-pf. coupling capacitor (ceramic) to Q_5.

C_X—(See text) 5-pf. NPO ceramic or silver mica.

C_4—32 pf. (single section of original capacitor). Any small variable can be used by properly choosing C_X.

L_8—12 turns No. 28 enam., close-wound, slug-tuned; taps at 3 and 6 turns from cold end. Form same as in Fig. 1.

On the Air

Quite a few contacts have been made on 21 Mc. Reports have been good as to quality and suppression. HC1AGI was able to read the exciter "barefoot." Extra contacts on my antenna change-over relay were used to activate the exciter. Since there is no warm-up time, all voltages may be removed during standby.

A V.F.O.

I realize that few hams have a 13.135-Mc. crystal in the junk box, and that a suitable v.f.o. circuit would be needed for the all-band project. Having seen an interesting article on a stable v.f.o.,[4] I decided to adapt it to the Imp-TR.

The crystal Y_5, VXO coil L_3 with its swamping resistor, and capacitor C_3 were removed from the board. The rest of the circuit was left intact except to remove the lead between the base of Q_4 and C_2.

Coil L_8 of Fig. 3 was mounted in the location formerly held by L_3, and the transistor socket for Q_9 in place of the crystal Y_5. The remaining parts were placed as space permitted. Only a single-section variable is needed, so one half of the split-stator capacitor was used. Since 32 pf. was too much capacitance for the frequency change needed, the effective capacitance was brought down by the use of 5 pf. in series, C_X. With it, a 75-kc. tuning range was covered in the 180 degrees of shaft rotation. Altering C_X will change the band-spread. Unlike the VXO, the output was constant over the desired frequency range.

In operation the v.f.o. performed beautifully. Stations being worked were asked to watch for drift, but they detected none. While no actual measurements were made, I can say that stability is very adequate for s.s.b. There is a slight drift from junction heating when voltage is first applied to the oscillator. This seems to be small, occurs quickly, and no station being worked could detect it. Even this might be eliminated by permitting the v.f.o. to run continuously.

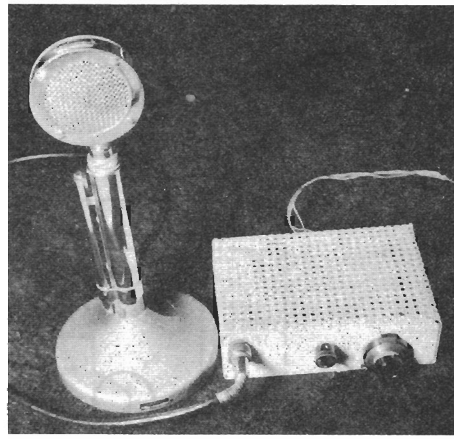
Acknowledgments and Remarks

My thanks to ZS5DN for portions of the circuitry and assistance in getting started. Mr. K. M. Aitken of RCA provided information and other help which was greatly appreciated. Experimentation is still being continued, but the all-band job is getting closer.

In general, the remarks regarding "other bands" for the tube version hold for the Imp-TR. Care should be used to prevent the carrier and mixer frequencies and their harmonics from appearing in the output.

[4] Stoner, "Stable Transistor V.F.O.," *Electronics World*, October, 1960.

» *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_5. 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 grounded-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.

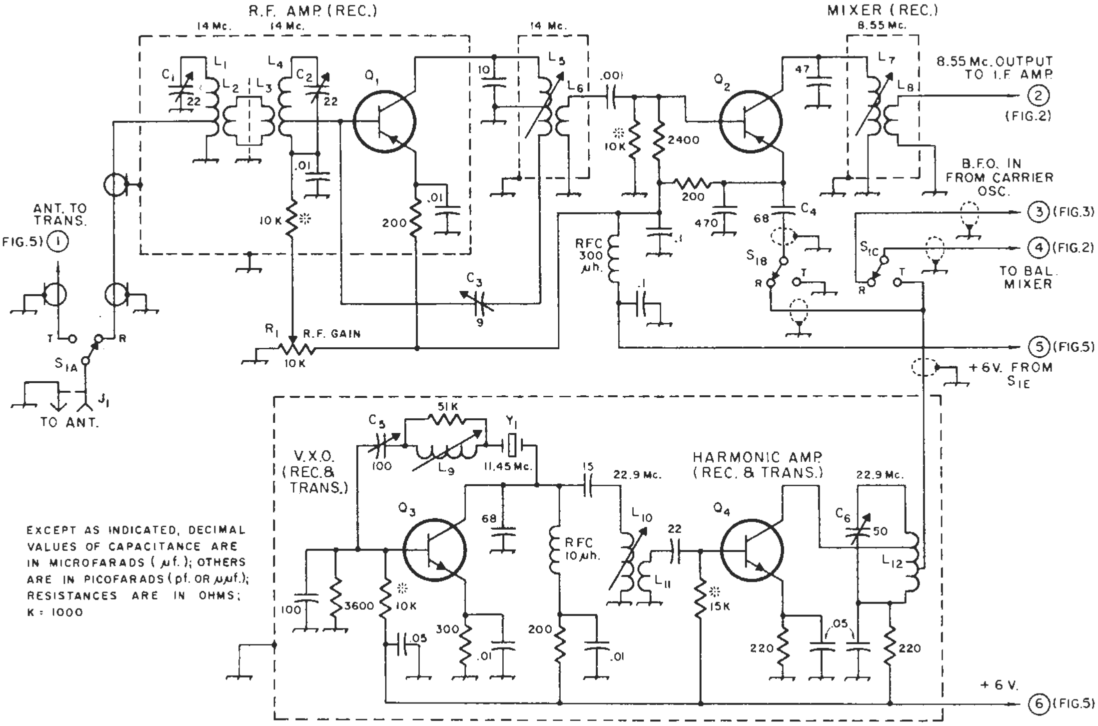


FIG. 1—VXO and high-frequency receiver circuits. The 22.9-Mc. VXO signal mixes with a 14.35-Mc. incoming signal to produce an 8.55-Mc. i.f. signal. On transmit, the 22.9-Mc. signal is transferred to the balanced diode mixer of Fig. 2. Fixed capacitors of decimal value are miniature ceramic or paper with a minimum rating of 6 volts. Others are NPO ceramic or dipped silver mica. Resistors are ¼ watt. Resistors marked with asterisks are bias resistors (see text).

C_1, C_2—Air trimmer from Command-set i.f. cans.

C_3—0.1–9-pf. trimmer, or "gimmick."

C_4—Nominal value; see text.

C_5—Air variable (Hammarlund APC-100B).

C_6—Ceramic trimmer.

J_1—Phono jack or chassis-mounting coax receptacle.

L_1—40 turns, tapped at 2 turns from ground end, on powdered-iron toroid (Stackpole D-1 iron) ½-inch outside diameter, ¼-inch inside diameter, circular cross section (Henry L. Crowley Co., West Orange, N. J., part No. C-2776).

L_2—Single turn on L_1; see text.

L_3—Same as L_2, wound on L_4.

L_4—Same as L_1, tapped at 4 turns.

L_5—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.

L_6—3 turns over L_5.

L_7—25 turns on CTC LS-9-4S shielded ferrite-slug coil form.

L_8—6 turns over ground end of L_7.

L_9—48 turns close-wound on 1-inch ceramic iron-slug form (National XR-60 form).

L_{10}—Inductance 3.5 µh., scramble-wound on CTC PLST-2C4L/N iron-slug form.

L_{11}—4 turns over ground end of L_{10}.

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

L_{12}—12 turns No. 24, ½-inch diam., 32 turns per inch (B & W 3004 Miniductor), tapped at 4 turns and 7 turns from ground end.

Q_1, Q_2—2N700, or similar u.h.f. p.n.p. transistor (see text).

Q_3, Q_4—2N706, or similar u.h.f. n.p.n. silicon transistor (see text).

R_1—Linear-taper control.

S_1—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.

Y_1—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, Q_4, 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.

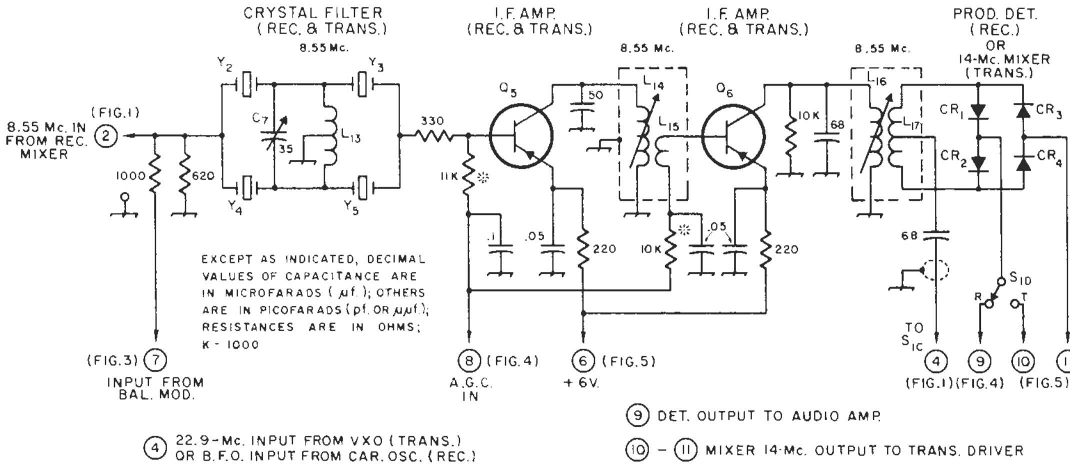


FIG. 2—8.55-Mc. i.f. circuit. On receive, the diodes in the output circuit operate as a product detector, the carrier oscillator (Fig. 3) serving as the b.f.o. On transmit, the 8.55-Mc. suppressed-carrier signal from the balanced modulator (also Fig. 3) passes through the crystal filter which strips 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. VXO 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 NPO ceramic or dipped silver mica. Resistors are ¼ watt. Asterisks identify biasing resistors (see text).

C_7—Ceramic trimmer (Centralab 827D).

CR_1, CR_2, CR_3, CR_4—Germanium diode (CK706, 1N34A or equivalent.)

L_{13}—20 μh., center-tapped, bifilar-wound on ¾-inch ferrite toroid core, and connected as described for L_5. Cores available from some source as L_1. See references 1, 2.

L_{14}—12 turns on CTC LS-9-4S, shielded ferrite-slug coil form.

L_{15}—4 turns wound over ground end of L_{14}.

L_{16}—24 turns on CTC LS-9-5S iron-slug form.

L_{17}—8 bifilar turns, wound over L_{16} and connected as described for L_5.

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

Q_5, Q_6—Same as Q_1.

S_1—See Fig. 1.

Y_2, Y_3—8550.3-kc. crystal.

Y_4, Y_5—8551.5-kc. 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 VXO depends on the inductance of L_9. I put on just enough turns to make the VXO 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 VXO 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 C_7 and coil L_{13} are chosen to resonate approximately (disconnected from the circuit) at the passband frequency of the filter. Adjustment of C_7 and the slug of L_7 (Fig. 1) can then be made to optimize the filter passband.

The i.f. amplifiers, Q_5 and Q_6, 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 Q_6 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. VXO signal are mixed to produce 14-Mc. output. The diodes are in a balanced arrangement so that both the VXO 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. L_{18} and L_{19} 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 Y_6 is similarly supported on the other two posts.

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

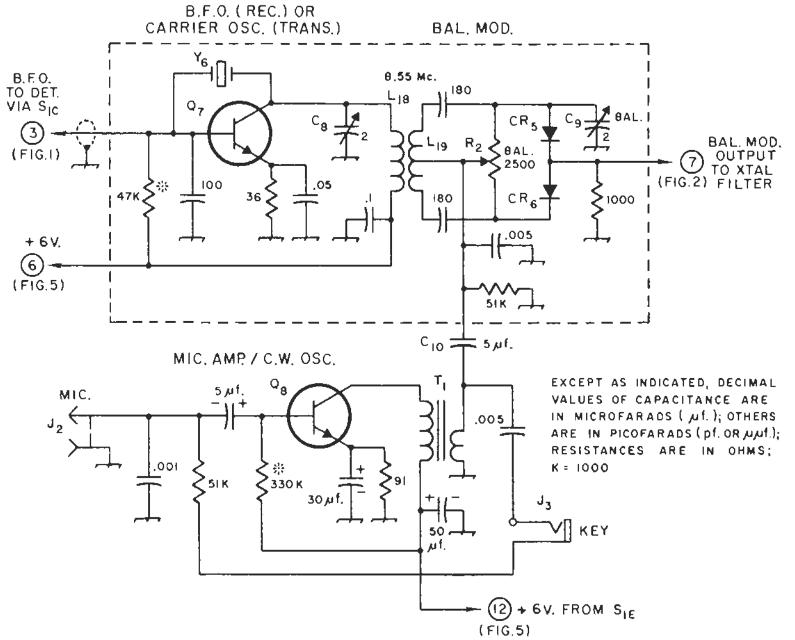


FIG. 3—Carrier-oscillator (b.f.o.), balanced-modulator, and transmitter audio circuits. A feedback circuit is provided to cause the microphone amplifier to oscillate for c.w. operation. Fixed capacitors of decimal value are miniature ceramic or paper. Except as listed below, others are NPO ceramic or dipped silver mica, except where polarity indicates electrolytic. All capacitors have a minimum rating of 6 volts. Fixed resistors are ¼ watt. Biasing resistors are identified by asterisks; see text.

C_8, C_9—22-pf. air trimmer from Command-set i.f. can.
C_{10}—Ceramic or paper.
CR_5, CR_6—Same as CR_1.
J_2—Microphone connector.
J_3—Miniature open-circuit phone jack; both sides must be insulated.
L_{18}—60 turns No. 34 enameled, close-wound on toroid form same as described for L_1.
L_{19}—10 bifilar turns over L_{18}, wound and connected as described for L_5.
Q_7—Same as Q_3.
Q_8—2N170 or similar.
R_2—Linear control.
T_1—Subminiature interstage audio transformer, 4:1 turns ratio, low-impedance winding in output.
Y_6—8553.0-kc. crystal.

C_9 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.

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

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, Q_{13}, in Fig. 5, is a straightforward Class A stage, The "final," Q_{14}, 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

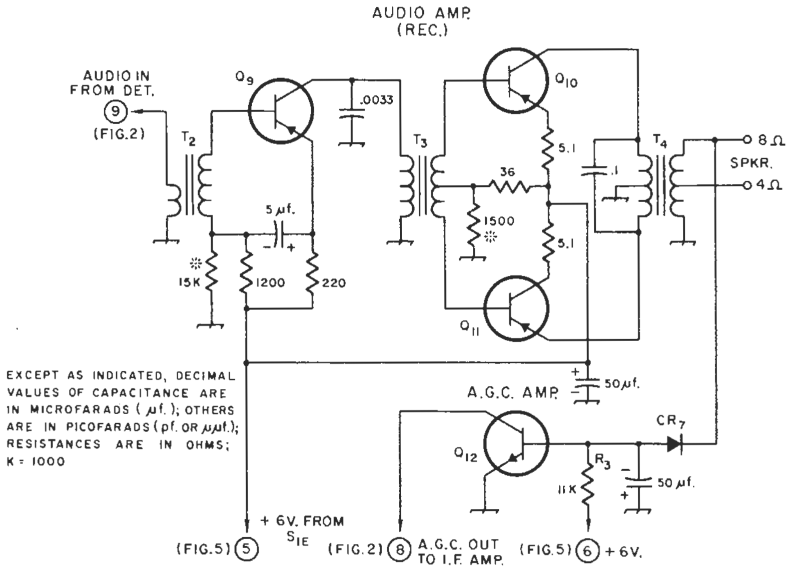


FIG. 4—Receiving audio circuit. An a.g.c. signal is obtained by rectifying and amplifying a signal taken from the audio output. Capacitors of decimal value are miniature ceramic or paper. Others are electrolytic. Both types have a minimum rating of 6 volts. Resistors are ¼ watt. Bias resistors are identified by asterisks; see text.

CR_7—Silicon junction diode, 50 p.i.v., 1N599 or equivalent.

Q_9—2N653 or similar.

Q_{10}, Q_{11}—2N586 or similar.

Q_{12}—Same as Q_8.

R_3—See text.

T_2—Same as T_1; low-impedance winding in output circuit.

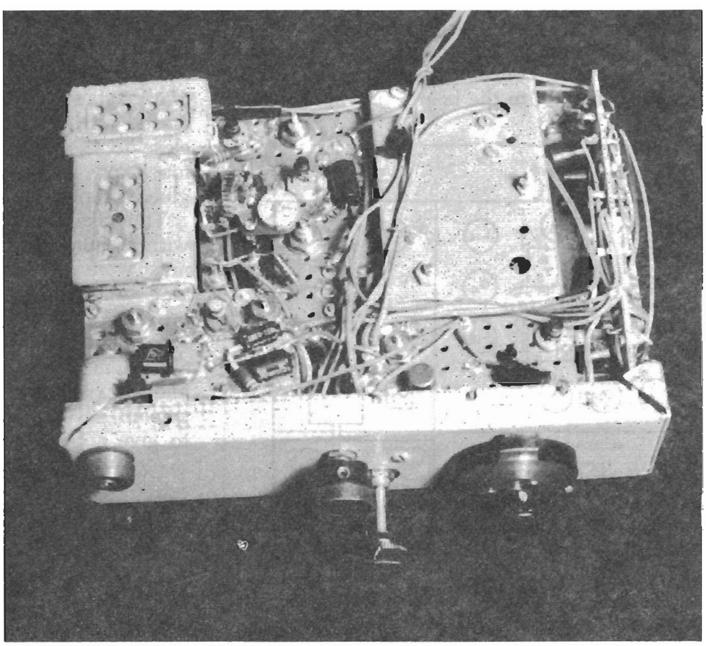
T_3—Subminiature interstage transformer, 4:1 turns ratio, secondary center-tapped.

T_4—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 × 7 × 2-inch unit with back apron sawed off.

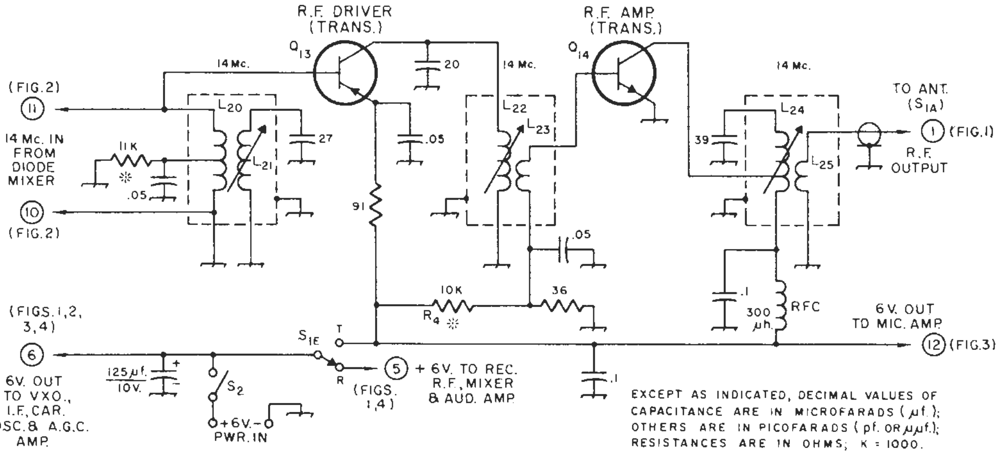


FIG. 5—Transmitter output circuits. This section receives 14-Mc. drive from the diode balanced mixer of Fig. 2. Capacitors of decimal value are miniature ceramic; others are NPO ceramic or dipped silver mica, except polarity indicates electrolytic. Capacitors have a minimum rating of 6 volts. Resistors are ¼ watt; asterisk indicates bias resistor; see text.

L_{20}—4 bifilar turns, center-tapped, wound over L_{21} and connected as described for L_5.
L_{21}—21 turns on CTC LS-9-5S iron-slug form.
L_{22}—21 turns on CTC LS-9-4S iron-slug form.
L_{23}—6 turns wound over ground end of L_{22}.
L_{24}—16 turns on CTC LS-9-5S iron-slug form, tapped at 4 turns from low-potential end.
L_{25}—6 turns wound over low-potential end of L_{24}.

Note: Above coils are close-wound with No. 34 enameled wire.
Q_{13}—Same as Q_1.
Q_{14}—Same as Q_3.
R_1—Nominal value, see text.
S_1—See Fig. 1.
S_2—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. C_7 is a Centralab type 827 ceramic trimmer capacitor. When Y_2 and Y_3 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. L_{13} 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 epoxyed together. After filing the corners smooth, several coats of spray paint were added to give a fairly professional-looking cover.

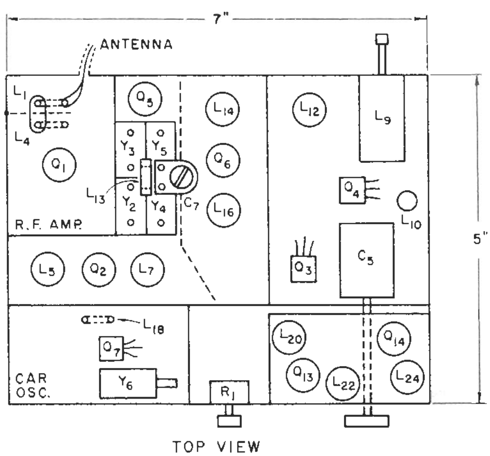
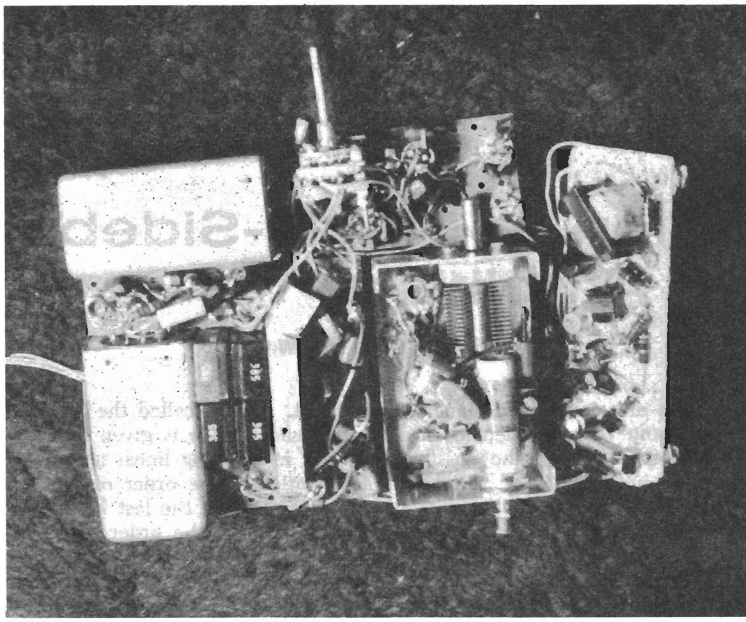


FIG. 6—Sketch showing layout of principal components.

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 Q_1, Q_2, Q_5, Q_6 and Q_{13}, I



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 principal causes of distortion in linear r.f. amplifiers, and what to do about them, are discussed in this article. Methods of measuring distortion also are described, and the operating principles of the "linearity tracer"—a checking device that also can be used as a continuous monitor of linearity with any voice waveform—are outlined.

Distortion In Single-Sideband Linear Amplifiers

WARREN B. BRUENE, WØTTK

When the envelope of a modulated signal is distorted, a great many new frequencies are generated. These represent all of the possible sum and difference combinations of the harmonics of the original radio frequencies. Since r.f. amplifiers use tank circuits, all distortion products are filtered out except those which lie close to the desired frequencies. These are all "odd order" products: third order, fifth order, and so on.

The third-order product frequencies are $2p-q$ and $2q-p$, where p and q represent any two *radio* frequencies present in the desired transmission. The fifth-order product frequencies are $3p-2q$ and $3q-2p$. These and some higher order products, such as might be produced by distortion in a single-sideband linear amplifier transmitting a two-tone signal,

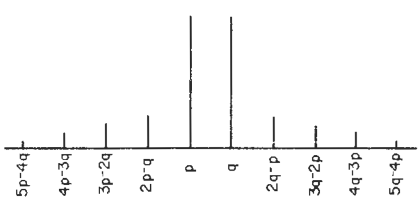


FIG. 1—Single-sideband distortion products.

are shown in Fig. 1. Note that the frequency spacing of the distortion products is always equal to the frequency difference between the two original tones, or legitimate sideband frequencies.

When a linear amplifier is badly overloaded these spurious frequencies can extend far outside the original channel and will cause unintelligible splatter interference in adjacent channels. Splatter of this type is usually of far more importance than the effect on intelligibility or fidelity of the distortion of the original signal. To minimize unnecessary interference, the distortion products falling in adjacent channels should be down as far as we can get them below the signal itself.

Using a two-tone test, the distortion is defined as the ratio of the amplitude of one test tone to the amplitude of the third-order product. This is called the "signal-to-distortion ratio" and usually is given in db. The state of the art in building linear amplifiers has limited S/D ratios to the order of 25 to 30 db. until recently. Within the last few years commercial performance of the order of 30 to 35 db. has been achieved. Recent developments indicate that even 40 db. is possible and practical.

In amateur transmitters where only one voice channel is used, the distortion requirements depend upon the allowable interference to others operating on near-by channels. Factors such as the relative amplitude of the signal with distortion to the amplitude of a near-by signal another amateur is trying to receive enter in. Common courtesy on the crowded amateur bands dictates the use of transmitters with as little distortion as the state of the art reasonably permits.

Causes of Distortion and Methods of Reduction

The principal causes of distortion are nonlinear characteristics of the amplifier tubes and grid-current loading. In order to confine the

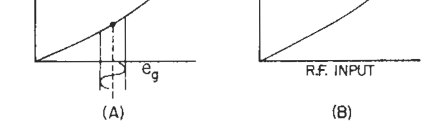


FIG. 2—Effect of nonlinear plate characteristics.

generation of distortion substantially to the last stage or two, all other stages are usually operated Class A. The plate current curve of Class A amplifier tubes in general can be represented by a simple exponential curve as shown in Fig. 2A. The distortion is kept low by operating the tube in the most linear portion of its plate current characteristic and by keeping the signal level low. Fig. 2B shows the nature of the linearity curve of a typical

From *QST*, November, 1954.

Class A amplifier. The curvature is greatly exaggerated since for S/D ratios of the order of 50 db., it cannot be detected visibly.

Class AB amplifiers usually have a very similar curvature. When the linearity characteristics of a series of cascaded amplifiers have similar curvatures, the distortion products generated by each add together in phase.

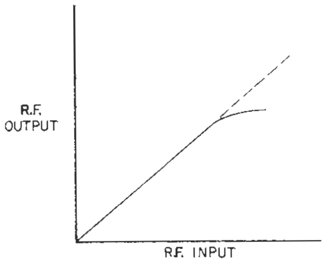


FIG. 3—Effect of grid loading on linearity.

When amplifier tubes are driven into the grid-current region, the resulting grid-circuit loading causes the linearity curve to droop at large signal levels as shown in Fig. 3. The distortion products from this type of curvature are 180 degrees out of phase with those previously discussed. When both types of curvature exist, the distortion products tend to cancel as shown in Fig. 4. When this happens, the fifth order product is usually stronger than the resulting third in the region of cancellation. For this reason, the value of distortion cancellation is not as great as it might seem.

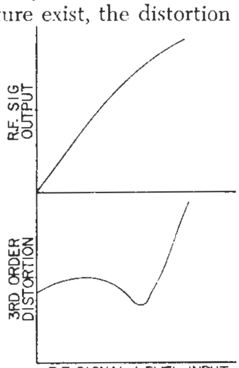


FIG. 4—Distortion concellation.

The nonlinearity caused by grid current loading is a function of the regulation of the grid driving source. The regulation of linear amplifiers with a varying load is poor in general. It is common practice to use a swamping resistor in parallel with the varying grid load, and to obtain satisfactory regulation it is usually necessary to absorb about ten times as much power in this swamping resistor as the grid consumes.

Another way of providing a low driving impedance is to use a very high resistance driver tube, such as a tetrode or pentode, and an impedance-inverting network.[1] The impedance-inverting network can be a quarter-wave or 90-degree network coupling the driver plate and power-amplifier grid tank circuits. Inductively-coupled tank circuits also have this property. Fig. 5 shows these two circuits. The disadvantage is that it is difficult to maintain proper coupling without special adjustment, and these circuits are seldom used in commercial general frequency coverage transmitters for this reason. Link coupling as used between exciter and final amplifier in many transmitters has this property also, if the line between the links is a small fraction of a quarter wavelength long. (This may explain why some rigs work as well as they do!)

It is apparent that it is best to choose tubes and operating conditions for low grid driving power. Tubes are available that will operate Class AB_1 at power levels up to 500 watts, and their use greatly simplifies the driver and bias regulation requirements.

In cathode-driven amplifiers the total grid and screen driving power should not exceed 10 per cent of the fed-through power at maximum signal level. For S/D ratios better than 30 db., it should be correspondingly less.

The plate current of all tubes drops off when the instantaneous plate voltage is low. Fig. 6 shows a typical plate-current curve taken along a straight load line on constant-current curves. The grid and screen currents are also shown. Two effects seem to cause the drop in plate current; the principal one is that current taken by the grid and screen is "robbed" from the plate, and it can be observed on tube curves

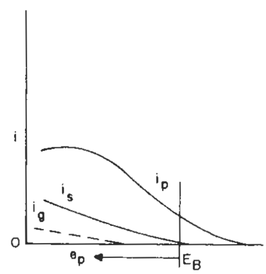
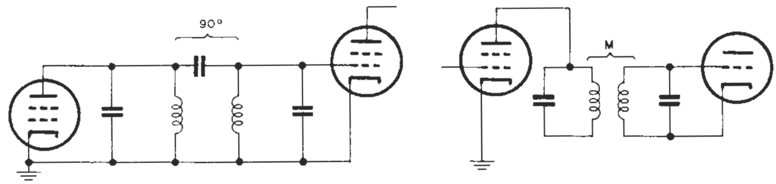


FIG. 6—Instantaneous plate current characteristic.

[1] Green, "Design of Linear Amplifiers for Single-Sideband Transmitters," *Marconi Review*, Vol. 10, pp. 11-16, January and March, 1947.

FIG. 5—High-resistance driver and impedance-inverting network.



that the plate-current lines depart from straight lines by approximately the amount of the grid and screen current. The amount of screen current and drop-off of plate current also depend upon the tube geometry. In all but a few transmitting tubes the plate can swing well below the screen voltage before plate saturation takes place, and when the plate swings down in this region the plate current drops off quite a bit. If the distortion requirements are not too high, the high plate efficiency realized by using large plate swings can be utilized. Fig. 7 shows a typical linearity curve of a

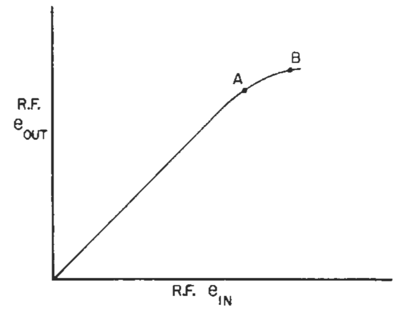


FIG. 7—Linearity curve of typical tetrode amplifier.

tetrode linear amplifier. At point "A", the plate is swinging down to the screen voltage. At point "B", it is swinging well below the screen and is approaching the grid voltage to the point where saturation or plate-current limiting takes place.

Estimating Distortion

A means of estimating distortion in a power amplifier is quite useful, and the approximate signal-to-distortion ratio of a two-tone test signal can be obtained from the linearity curve. Equations have been developed for calculating this, and are used to plot the curve in Fig. 8. This curve shows the distortion resulting from flattening of the envelope peak.

Distortion in the lower part of the linearity curve is due to incorrect voltages on the tube

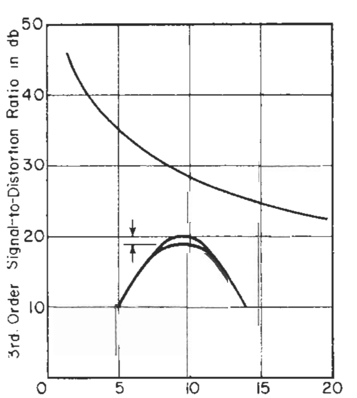


FIG. 8—Relationship between third-order distortion and envelope peak flattening.

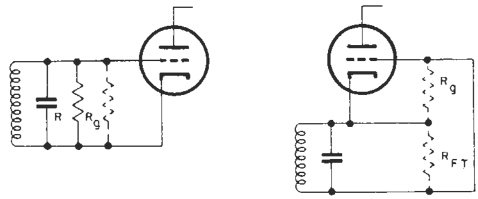


FIG. 9—Grid loading. A—Grounded-cathode input circuit; B—Grounded grid.

elements. It can be substantially eliminated by proper adjustment of bias, screen and plate voltages, so means of estimating distortion from this cause will not be discussed.

Envelope peak flattening which is due to grid current loading and plate current nonlinearity at large plate swings is often the major cause of distortion. The amount of envelope peak flattening due to grid current loading may be easily calculated. See Fig. 9. The equivalent grid load resistance R_g in Fig. 9 is calculated from the grid driving power and the r.f. grid swing.

$$R_g = \frac{e^2{}_g}{2P_g}$$

where e_g = peak r.f. grid voltage, and
P_g = grid driving power = $e_g I_c$,
where I_c = d.c. grid current in amperes.

The resistance of the swamping resistor, R, is known or can be chosen for the calculation. The equivalent resistance of R and R_g in parallel is then calculated by:

$$R_{eq} = \frac{RR_g}{R + R_g}$$

If the source of impedance looking back at the driver stage is very high compared with R, it will contribute little toward improving the driving voltage regulation. In this case, the grid voltage will be reduced on the envelope peak by the amount of reduction from R to R_{eq}.

$$\text{Peak flattening} = \frac{R - R_{eq}}{R} \times 100 \text{ (per cent).}$$

The resulting distortion can then be found using Fig. 8.

The calculation is made in a similar manner for cathode-driven amplifiers. Use the equivalent resistance, R_{ft}, of the fed-through power at the cathode in place of R in the above equations. In tetrode cathode-driven amplifiers the grid and screen driving power should both be considered in calculating R_g.

Usually the third-order distortion component is at least 6 db. greater than the fifth- or

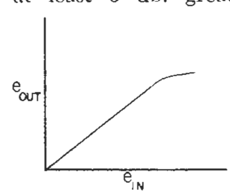


FIG. 10—Linearity curve with plate saturation.

higher-order components, but a sharp break in the linearity such as might be caused by plate-voltage-swing saturation, as shown in Fig. 10, will contain more fifth- and higher-order components than if it were a smooth curve. This type of nonlinearity is particularly objectionable because of the wide band over which the distortion products appear.

The other principal type of nonlinearity is caused by the exponential plate-current characteristics of the tube. Fig. 11 shows such a curve. As stated earlier, this type of curve

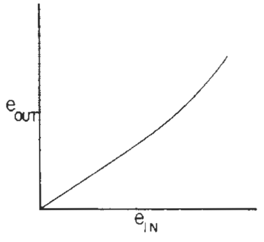


FIG. 11—Nonlinearity due to exponential plate current characteristic.

is obtained with Class A amplifiers. The distortion is kept low by proper tube choice and by operating at a low signal level over the most linear portion of the curve. In class AB amplifiers, the use of the optimum value of static plate current will do most toward reducing this type of nonlinearity. A smooth curve of this type usually contains mostly third-order distortion products. Even though the third-order products may be high, the bandwidth over which significant higher order products appear may be relatively narrow. Compound curves such as the one shown in Fig. 12 have

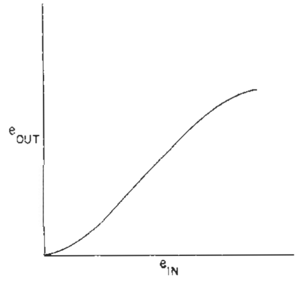


FIG. 12—Linearity curve with compound curvature.

relatively stronger fifth- and higher-order distortion components because the third tends to be cancelled as previously shown in Fig. 4.

Distortion Measurements

Distortion measurements are of particular importance in single sideband. The power output is often defined as the maximum peak envelope power output obtainable with a specified signal-to-distortion ratio. The distortion rises rapidly when the power amplifier is overloaded, and so has a considerable bearing on the power rating. A plot of the S/D ratio *vs.* peak envelope power is an excellent way of showing a transmitter's distortion and power capabilities. A typical curve is shown in Fig. 13. Two tones of equal amplitude are used for nearly all measurements in order to provide a "modulation envelope."

There are several different methods of indicating or measuring distortion, and each has a separate field of usefulness. The "Linearity Tracer" described below is especially useful for quick observation of amplifier operation as the effect of various adjustments can be instantly observed. This instrument consists of two s.s.b. envelope detectors with the output of one connected to the horizontal input of an

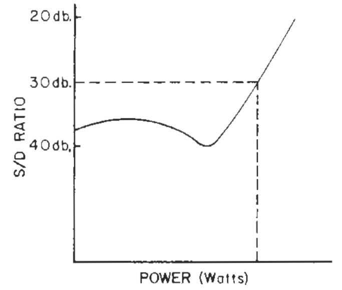


FIG. 13—Signal-to-distortion ratio *vs.* power output.

oscilloscope and the output of the second connected to the vertical input.

Fig. 14 shows a block diagram of this instrument connected to a power amplifier. A two-tone test signal is normally used to supply a single-sideband modulation envelope but any modulating signal that provides an envelope varying from zero to full amplitude can be used. Even speech modulation gives a satisfactory trace, so this instrument is unique in that it is an excellent visual monitor. It is particularly useful for monitoring the signal level, and clearly shows when the amplifier under observation is overloaded. The linearity trace will be a straight line regardless of the envelope shape if the linear amplifier has no distortion. Overloading causes a sharp break in the linearity curve. Distortion caused by too

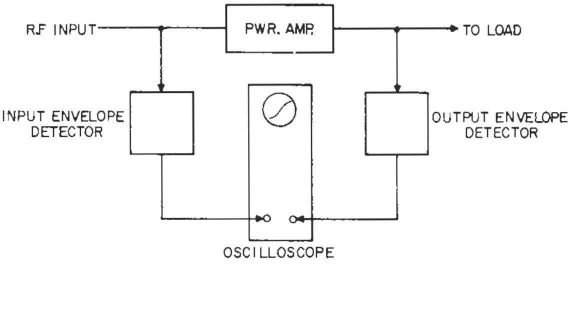


FIG. 14—Block diagram of linearity tracer.

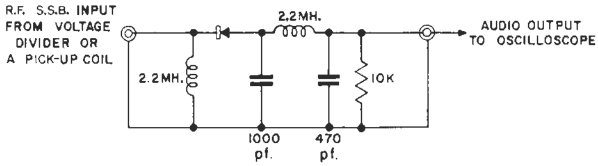


FIG. 15—Envelope detector.

much bias is also easily observed and the adjustment for low distortion can easily be made.

Another unique feature is that the distortion of each individual stage can be observed. This is helpful in troubleshooting. By connecting the input envelope detector to the output of the s.s.b. generator, the over-all distortion of the entire r.f. circuit beyond this point, including any mixer stages, is observed. It can also serve as a voltage indicator which is useful in making tuning adjustments.

Fig. 15 shows the circuit of an envelope detector. A germanium diode is used as the rectifier. Any type can be used, but the one used in the input detector must be fairly well matched to the one in the output detector. The detectors are not linear at low signal levels, but if the nonlinearities of the two detectors are matched the effects of their nonlinearities on the scope trace are canceled. Diode differences are minimized by using a diode load of 5000 to 10,000 ohms, as shown in the schematic. It is important that both detectors be operated at approximately the same signal level so their differences will cancel more exactly. Although they will operate well on r.f. voltages below 0.1 volt it is desirable to operate them on voltages above 1 volt, which further minimizes diode differences.

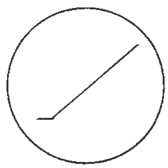


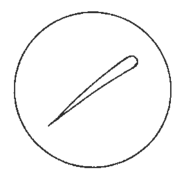
FIG. 16—Effect of inadequate response of vertical amplifier.

It is convenient to build the detector in a small shielded enclosure, such as an i.f. transformer can fitted with coax input and output connectors. Voltage dividers can be similarly constructed so that it is easy to patch in the desired amount of voltage stepdown from the voltage sources. In some cases it is more convenient to use a pick-up loop on the end of a short length of coaxial cable.

The frequency-response and phase-shift characteristics of the amplifiers in the oscilloscope should be the same and flat out to at least 20 times the frequency difference of the two test tones. An oscilloscope such as the DuMont type 304H is excellent for this purpose. It has d.c. amplifiers, which are best when monitoring speech because axis shift is avoided. Good high-frequency characteristics are necessary because the rectified s.s.b. envelope contains harmonics extending to the limit of the envelope detector's ability to detect them. Inadequate frequency response of the vertical amplifier may cause a little "foot" to appear at the lower end of the trace as shown in Fig. 16. If it is small, it may be safely neglected.

Another effect often encountered is a double trace as shown in Fig. 17. This can usually be corrected with an *RC* network between one detector and the oscilloscope. Such effects are easily remedied and an accurate linearity trace is not difficult to obtain.

FIG. 17—Double trace caused by audio phase shift.



The best method of checking the test set-up is to connect the inputs of the envelope detectors in parallel. A perfectly straight-line trace will result when everything is working properly. One detector is then connected to the other source through a voltage divider chosen to deliver an r.f. voltage amplitude such that an appreciable change in the setting of the oscilloscope amplifier gain controls will not be required. Fig. 18 shows some typical linearity traces. The probable causes and remedies follow:

Fig. 18A: Inadequate static plate current in Class A or Class AB amplifiers or a mixer. Reduce the grid bias, raise the screen voltage, or lower the signal level through mixers and Class A amplifiers.

Fig. 18B: Caused by poor grid-circuit regulation when grid current is drawn or by nonlinear plate characteristics of the tube at large plate swings. Use more grid swamping, lower the grid drive, or change plate loading.

Fig. 18C: Effect of (A) and (B) combined.

Fig. 18D: Overloading the amplifier. Lower the signal level.

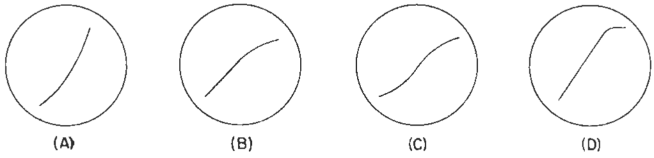


FIG. 18—Typical linearity traces.

Distortion Checking with a Selective Receiver

A fair idea of the S/D ratio of the transmitter can be obtained without requiring any equipment beyond what many amateurs already have. The method uses a receiver, such as the 75A-3 with the 800-cycle mechanical filter, that has sufficient selectivity to separate the frequency components of a two-tone test signal.

The transmitter should be modulated to produce a two-tone signal with a frequency separation of about 2000 cycles, and the amplitude of the third-order distortion can be compared with the amplitude of one of the tones simply by reading the difference on the S-meter as the receiver is tuned from one to another of the frequency components in the transmitter output. To avoid generating distortion in the front end of the receiver the r.f. gain control should be operated nearly wide open and the receiver input decoupled from the transmitter output to keep the maximum S-meter reading a little below full scale. Care must be taken to insure that the signal is getting into the receiver only through the antenna input terminals and not through the a.c. line, and also that the signal is coming from the output circuit of the stage being checked and is not a composite of stray radiation from several circuits and stages.

The accuracy of distortion measurements by this method depends on the care used in observing the precautions listed above and on the accuracy of the S-meter calibration. Even though the S-meter calibration is "off," the method is useful for adjustment purposes if the precautions are observed, since it will show qualitatively the effect of changes in operating conditions or tuning.

A REGULATED SCREEN SUPPLY

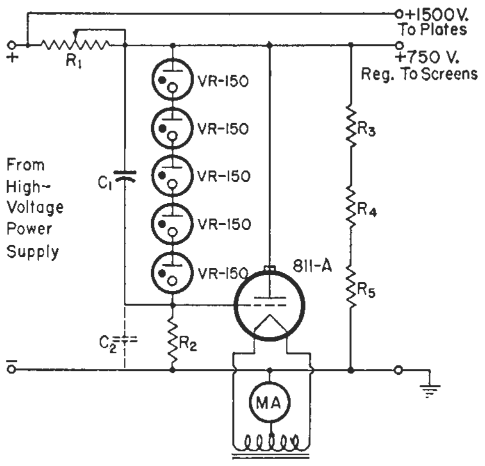
As everyone knows, or soon finds out, tetrode linear amplifiers require "stiff" screen-voltage supplies for lowest distortion. Earl Weaver, W2AZW, used a pair of 813s in his output amplifier, and devised the circuit shown here to stabilize the screen voltage. It is a shunt-type regulator that derives a regulated voltage from the high-voltage supply. Since the high-voltage supply will usually need a bleeder resistance for regulation purposes, the shunt regulator also takes care of that requirement.

A zero-bias tube is used, and the grid is always conducting, unless the source voltage drops so low that the VR tubes extinguish. The output voltage is equal to the sum of the VR drops plus the grid-to-ground voltage of the 811-A. This grid-to-ground voltage is the regulating potential, of course, and varies from 5 to 20 volts between full load and no load.

The initial adjustment is made by placing a milliammeter in the circuit as shown and adjusting R_1 for 15 to 20 ma. higher than the normal peak screen current. This adjustment should be made with the amplifier connected but with no excitation, so that the idling plate current will be drawn. After the adjustment is completed, the meter can be removed from the circuit and the filament center-tap wired directly to ground. Since R_1 is in a high-voltage circuit, it must be treated with full safety precautions, and all adjustments should be made only after the power is turned off and the high-voltage terminal has been grounded.

Any number of VR tubes may be used to provide a regulated voltage near the desired value. VR tubes with various operating voltages can be connected in series, if the current ratings are the same. Two 811-As can be connected in parallel if higher current capacity is required. The maximum current through the 811-A should be such that the manufacturer's plate-dissipation rating is not exceeded. It may be necessary to adjust R_1 for a slightly higher current under minimum load than is first expected, to compensate for full-load voltage drops in the high-voltage supply.

At W2AZW, the 813 screen current varied from about 5 to 60 ma., and the shunt regulator held the screen voltage constant to within 10 or 15 volts.



The regulated screen supply used with a pair of 813s at W2AZW.

C_1—0.01 μf., 2000 volts.
C_2—0.01 μf., 400 v., if needed to prevent oscillation.
R_1—Adjustable wire-wound, resistance and wattage as required.
R_2—22,000 ohms, 2 watts.
R_3, R_4, R_5—0.1 megohm, 2 watts.
MA—Milliammeter required for original adjustment.

» *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.

Condensed from "Intermodulation Distortion in Linear Amplifiers," *QST,* September 1963.

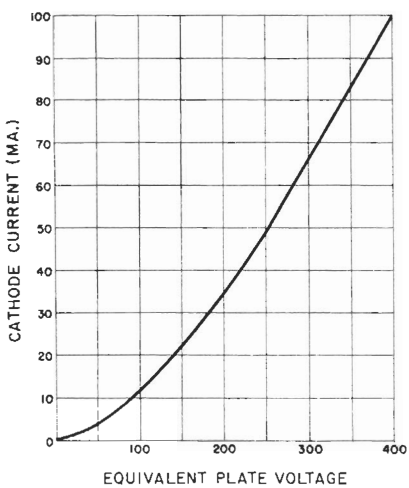


FIG. 1—The electron flow (cathode current) in a vacuum tube is a nonlinear function of the equivalent plate (or plate and screen) voltage and is described by the 3/2-power law. This curve illustrates typical electron flow, which plays an important part in establishment of tube linearity.

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

[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.

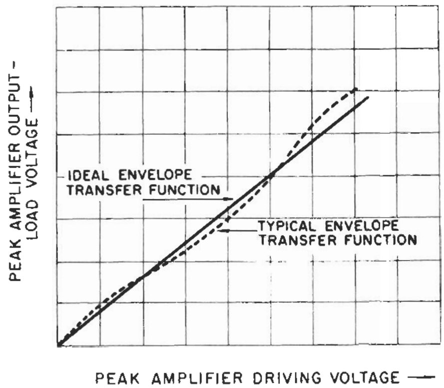


FIG. 2—Amplifier linearity is defined by the envelope transfer function. Departure from linearity is illustrated by curvature of the function (dotted curve) and may be traced directly to the nonlinear relationship between cathode current and electrode voltage shown in Fig. 1.

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 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

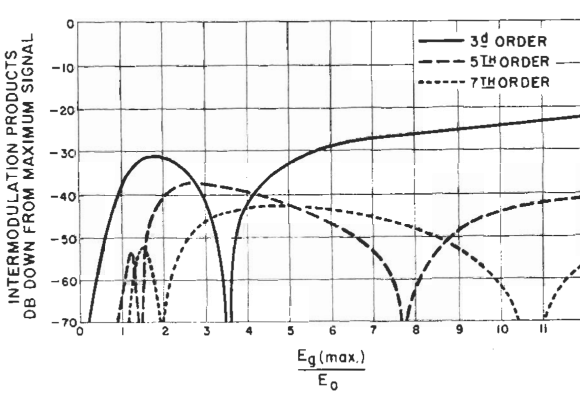
IMD products may be calculated by several methods.[3] The results of different valid mathematical techniques are in good agreement with each other, and also agree in general with data obtained from two-tone tests conducted with the IMD analyzer. A theoretical family of IMD curves of a perfect tube obeying 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)

[2] Cutoff 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.



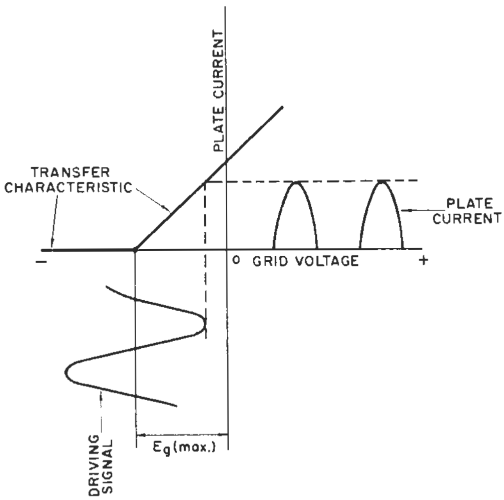


FIG. 4—An ideal tube transfer characteristic departs from the 3/2-power law. The ideal characteristic shown here consists of two linear portions, with the operating point set at the intersection. Half-wave plate current pulses are converted to sine waves by the flywheel effect of the plate tank circuit. 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 implicity 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

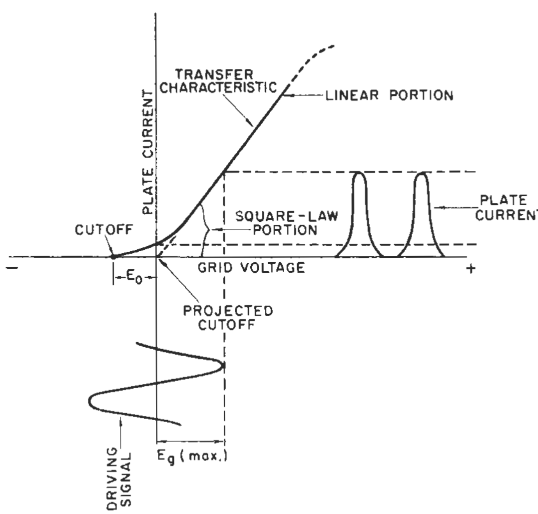


FIG. 5—Another ideal transfer characteristic for a linear tube consists of this form of curve, where the central portion is straight and the lower portion resembles a parabola. Practical tubes exhibit transfer characteristics of this general class, the upper portion of the curve showing additional curvature 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.

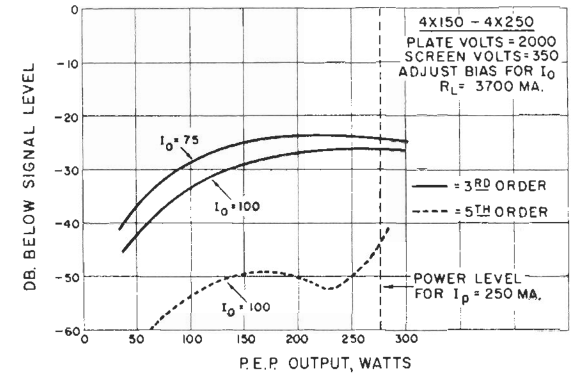


FIG. 6—A family of IMD curves for the 4X150–4X250 external-anode tubes. These curves are representative of this type of tube, and are typical for tubes made by different manufacturers. Intermodulation distortion products average about −25 decibels below peak signal, for 3rd-order products, while 5th-order 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 $E_{g(max)}/E_o$ 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.

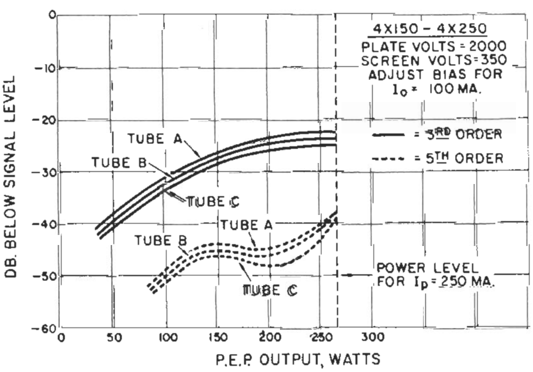


FIG. 7—Intermodulation distortion products vary from tube to tube of the same type, and also vary tube to tube as operating conditions are changed. Small "receiving-type" transmitting tubes are usually poorer than these curves by five to ten decibels.

» 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, W6KEV, 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-fidelty 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.

From August, 1961, *QST*.

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

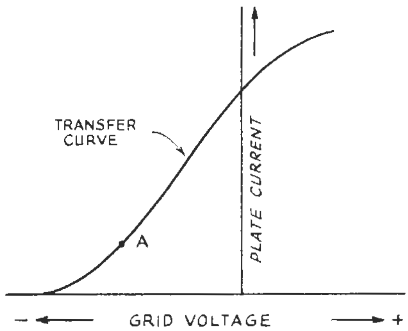


FIG. 1—Plate current vs. grid voltage curve (dynamic characteristic) of a vacuum tube. This curve is linear in the center portion and exhibits deviations at either extremity. The shape of the curve and the choice of the zero-signal operating point (A) will determine the distortion produced by the tube. 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.

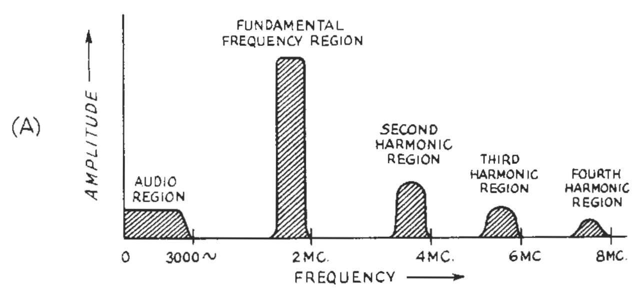
A standard test to determine the degree of mixing for a given circuit or tube is the *two-tone* test, in which two radio frequencies of equal amplitude are applied to the amplifier and the output signal is examined for spurious products (Fig. 2). These products, or "garbage," fall in the fundamental 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.

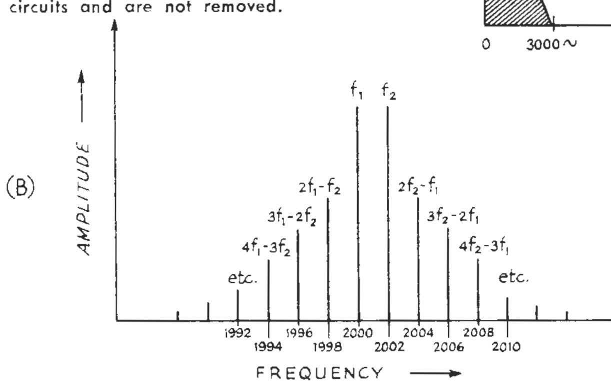
Shown in the illustration are two frequencies that make up a typical two-tone test signal. In this example, they are 2000 kc. and 2002 kc. Now, if the amplifier is perfect, these two signals will be the only ones appearing in the output circuit. An imperfect (but practical!) amplifier will have various 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

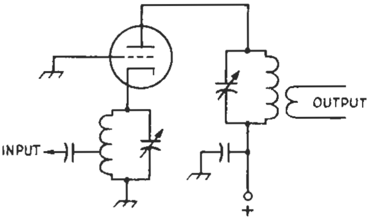


FIG. 3—The grounded-grid amplifier has the input circuit between cathode and ground. The control grid acts as a screen between the plate and the cathode, making neutralization unnecessary in most circuits. The input and output circuits are in series and a portion of the input power appears in the output 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-

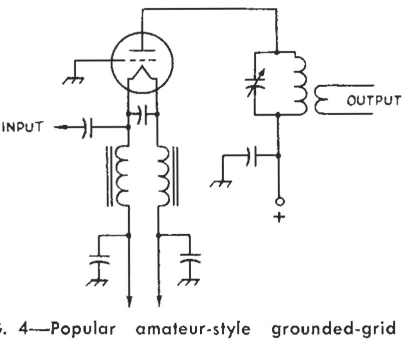


FIG. 4—Popular amateur-style grounded-grid amplifier used untuned filament choke in place of cathode tuned circuit. Laboratory tests showed that this simplified configuration produced higher intermodulation distortion products and had less power output than the "classic" circuit of Fig. 3, regardless of the type of tube used. In addition, the untuned input circuit proved hard to match and drive with pi-network-type sideband exciter.

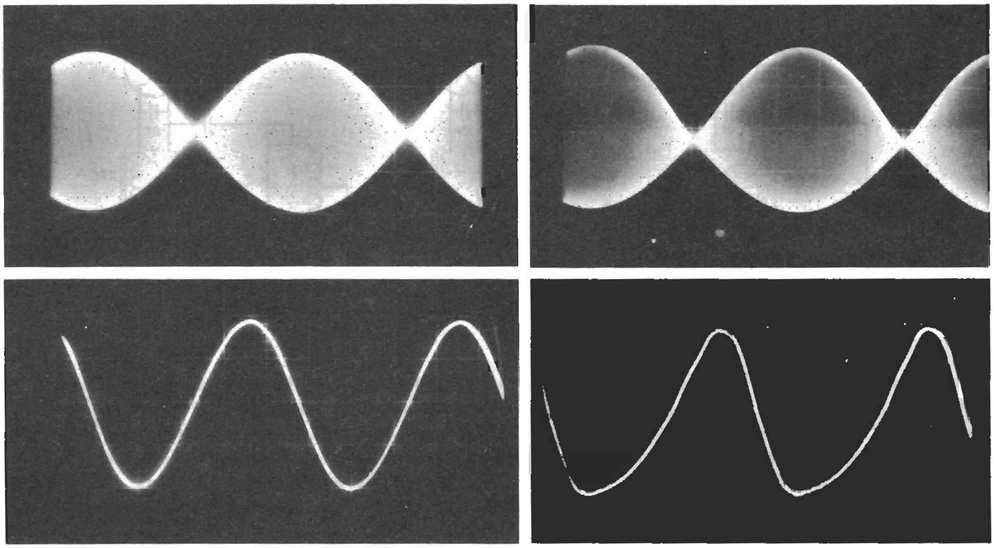


FIG. 5—Waveform distortion caused by half-cycle loading at cathode of grounded-grid amplifier can be observed in oscilloscope studies. *Upper left:* Two-tone test signal when tuned cathode circuit is used. *Lower left:* 3.5-Mc. waveform (single tone) from sideband exciter as seen at cathode tank. *Upper 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-

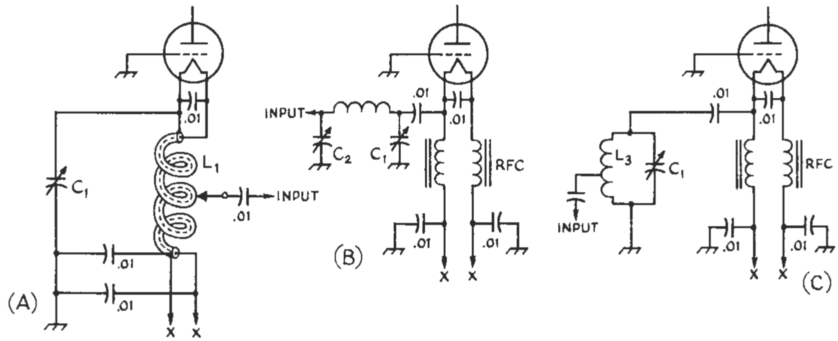


FIG. 6—Tuned cathode network for zero-bias tube may take the form of bifilar circuit (A), pi network (B), or a shunt *LC* circuit (C). A Q of 5 is recommended for optimum results. However, as this leads to rather bulky circuits at the lower frequencies, the Q may be decreased to 2 or 3 without serious effects. Capacitor C_1 is a 3-gong 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.

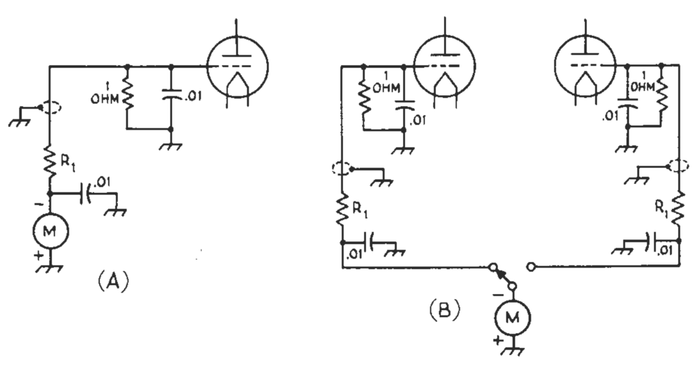
sistor, bypassed by a 0.01-μf. disk capacitor. The voltage drop generated by the flow of grid current across the resistor can easily be measured by a millivoltmeter calibrated to read in terms of grid current. Individual grid currents for each of a parallel pair of tubes may be measured by the circuit 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 milliampere flows through it, but at one milliampere 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-μf. 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.



» This article points out some of the problems involved in adjusting and tuning large linear amplifiers, and it goes on to show how it can be done easily with a minimum of extra gear in the shack. The s.s.b. man with a small linear amplifier and large ambitions can also benefit by using the same methods.

Linear Amplifiers and Power Ratings

BYRON GOODMAN, W1DX

Ever since the use of single sideband and linear amplifiers, there has been considerable misunderstanding of the power levels involved. *QST* advertisements for sideband gear carry references to "p.e.p.," but there are probably many readers who interpret this as an abundance of vitamins instead of "peak envelope power." On the air there have been instances of amateurs blithely announcing that they were "tuning up with the two-tone test and running 2 kilowatts peak input," a strictly illegal operation and a loud confession to not knowing the facts about linears and the FCC regulations. It is the intent of this article to clarify some of the points of linear-amplifier ratings and to show how a legal kilowatt linear can be tuned without breaking the law.

(What's that? Break the law with a legal kilowatt? This guy's off his rocker!)

The FCC says that the d.c. input to the stage or stages delivering power to the antenna shall not exceed 1 kw. On a.m. or c.w. this is easy to measure; read the meters while the carrier is on or while the key is down and you have it. If the constant-carrier a.m. rig is any good the meter readings will be the same when you modulate as when you aren't talking, but on c.w. the plate milliammeter reading never gets up to the kilowatt unless you hold a long dash. When a c.w. rig is loaded to a kilowatt, a string of fast dots will indicate about a half kw. input and a string of fast dashes will show about ¾-kw. input. Any c.w. operator who has watched the plate milliammeter while he's breezing along on a bug will recognize how difficult it would be to measure his power input without holding the key down until the plate meter stood still.

If you're wondering why c.w. is mentioned at all in an s.s.b. article, it's because it helps to understand the problems involved in the power considerations of linears used on s.s.b. If you load a c.w. rig, couple the output to an oscilloscope and hold the key down, the scope pattern will look like that of Fig. 1-A. The height h on the scope face is a measure of the peak r.f. voltage. The plate d.c. voltmeter and the plate d.c. milliammeter will show steady readings as long as the key is held down (and nothing burns up in the rig!). If the tuning and coupling are left as they were and a string of fast dots is sent, the scope picture may look like Fig. 1-B. The height h is the same as before. The plate d.c. voltmeter will show the same reading with an excellent power supply or a slightly higher reading with an average power supply. However, the plate milliammeter will hover around half of the key-down reading, because half the time the plate current is zero and about half the time the plate current is at the key-down value. The needle can't move back and forth fast enough to follow, so it shows a value halfway between.

In the c.w. case illustrated in Fig. 1, the

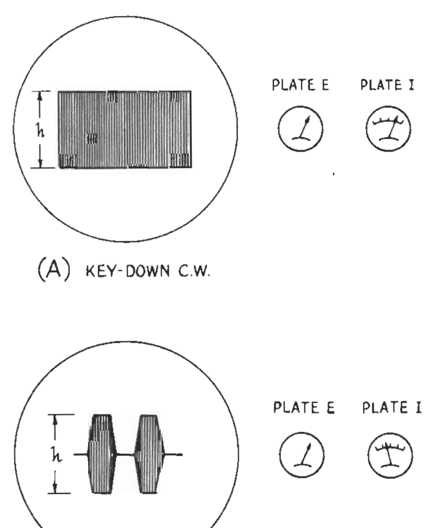


FIG. 1—(A) An unmodulated carrier looks like this on an oscilloscope. (B) Sending a string of dots, the peak amplitude of the signal remains the same, but the plate power input will show half the steady key-down value.

From *QST*, August 1957.

peak envelope input power is the key-down input power, and the peak envelope output power is the key-down output power. We can get the peak envelope input power easily from the product of plate voltage times plate current with the key held down, or we can get it the hard way by reading the input power when running a string of dots and dividing the indicated input power by 0.5 (Both answers will be the same when the bug is adjusted to give correct dots; "heavy" dots would require dividing by something greater than 0.5, and "light" dots would call for a divisor smaller than 0.5.) The string-of-dots case is a simple one; we divide by 0.5 because correct dots have a *duty factor* (pulse duration times pulse frequency) of 0.5. Dividing the indicated input power by the duty factor gives the actual power input during the "on" time.

Linear Amplifiers and Voice

Anyone should be able to see that it is easy to measure the input on c.w. with the key down, and not extremely difficult when running a steady string of dots (or dashes). But how would you like to be held responsible for knowing the input when all you were permitted to send was straight text? What would you use for a plate milliammeter reading? What divisor would you use?

The s.s.b. linear case is in the same class. A typical s.s.b. signal might show up on a scope as the sketch in Fig. 2. Most of the

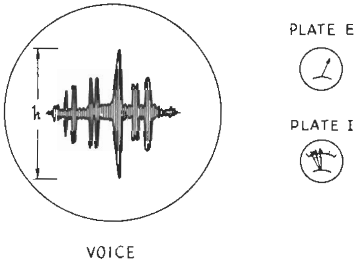


FIG. 2—Voice-modulated s.s.b. signals look something like this on a scope. Here the height h represents the maximum amplitude that the amplifier can handle without distortion. Since the signal hits these high peaks only occasionally, the indicated power input will be low. The syllabic nature of the signal is indicated by the jumpy plate milliammeter.

time the signal amplitude is at a relatively low level, but it rises to peaks during the loud syllables. If h represents the maximum amplitude the amplifier will handle without distortion (and consequent splatter), the intelligent s.s.b. operator will hold his voice at a level where these peaks are hit occasionally but never exceeded. The plate milliammeter will kick around a bit, so how do you correlate its reading with the peak envelope power (the power during that maximum peak)?

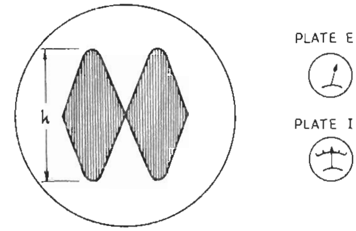


FIG. 3—The two-tone test signal applied to the linear of Fig. 2. The maximum permissible amplitude h is the same, but now the indicated power input bears a definite relationship to the peak envelope input.

The FCC recognizes the problem, of course, and consequently they give us a simple requirement to meet: the d.c. input shall not at any time exceed 1 kw. as indicated by meters with time constants not exceeding 0.25 second. (Time constant relates to the time it takes a meter to rise to the true value of current; a longer time constant means a more sluggish meter.)

But how do you tune up an amplifier that runs the legal limit? If a linear capable of handling the legal limit on s.s.b. is driven by a single r.f. signal to its full capability, the d.c. input will be more than a kilowatt. And, of course, while a big linear can be *resonated* with a single r.f. signal it can't be *loaded* correctly except under special circumstances (having prior knowledge of the plate current for a given level of signal).

Perhaps you have heard of the "two-tone test" and you suggest that next. No go. The two-tone test is a good method for checking the linearity of an amplifier [1]; it involves using two equal-amplitude r.f. signals through the amplifier, which gives a pattern on the scope as in Fig. 3. If the pattern just starts to flatten at an amplitude of h, indicating that this is the maximum signal the amplifier can handle without generating unwanted spurious signals, the peak envelope input is the indicated d.c. power input divided by 0.636. (The two-tone signal is used because it is easy to generate and the relationship between peak envelope input and indicated d.c. power input is accurately known.[2]) But a Class-B amplifier that runs up to the legal input on voice will run about 2 kw. p.e.p. input, and an amplifier indicating 1 kw. input with a two-tone test signal has a p.e.p. input of less than 1.6 kw. ($1 \div 0.636$). The two-tone test is fine for amplifiers that indicate up to perhaps 750 watts input on voice peaks, but beyond that the two-tone test *on the air* will result in a d.c. input of more than a kilowatt.

[1] Ehrlich, "How To Test and Align a Linear Amplifier," page 215.

[2] Strictly, the factor 0.636 applies only in the case of an ideal Class B amplifier—i.e., biased to exact cutoff and perfectly linear. Where there is appreciable idling current a different factor should be used. See note on page 134.—*Editor.*

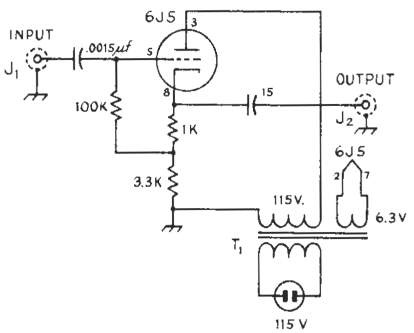


FIG. 4—Schematic diagram of a simple "pulser" for keying the audio tone used in the two-tone test. The capacitances are in pf. and the resistors are ½-watt. T_1 is a small TV-booster transformer (Merit P-3045 or equiv.).

Tuning a Big Amplifier

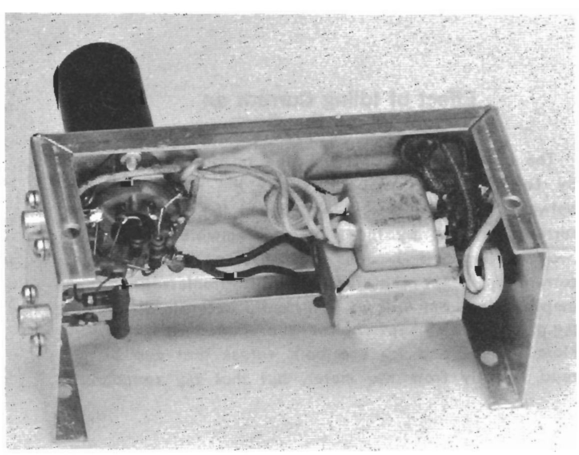
Obviously, one way to adjust a big linear is to use a dummy load and the two-tone test signal. This is certainly the only way to work with the amplifier in the early testing stages. It will enable you to find the proper loading for good linearity and efficiency, and you can then substitute the antenna for the dummy load at reduced signal levels. But suppose you want to make a quick check on the air, or suppose you hanker to see how much you can crowd your amplifier without distortion.

We can't use the two-tone test continuously, so the obvious solution is to key it. This is a cinch for anyone with an electronic bug key; all he has to do is to lock the key in the dot position and key the audio tone being fed to the sideband generator. Anyone with such a bug key available can forget about the electronic pulser to be described. For those without an electronic bug, the circuit in Fig. 4 can be used to key the audio tone. It is a simple method (although the electronic bug key is even simpler, if you have one) for testing an amplifier where the ordinary two-tone test would result in exceeding the legal d.c. input limit or possible failure of the tubes.

The gadget in Fig. 4 and the photograph is not offered as a refined device for exact measurement but merely a very simple means for testing a sideband system in which the final runs at the legal limit or where the tubes are run too hard on peaks for a sustained two-tone test. It consists of a cathode follower with a.c. instead of d.c. on the plate. A single tone (1000 cycles or so) is fed from an audio oscillator to J_1 and the output is run from J_2 to the microphone jack of the transmitter under test.[3] The follower conducts during the positive half cycles of the plate voltage and not during the negative half cycles, resulting in a test signal that is on about half the time. Since there is a 60-cycle component in the plate current, the output is taken off through a small (15-pf.) capacitor to reduce the 60-cycle component reaching the output. Most sideband transmitters will, or should, have poor 60-cycle response in their audio amplifiers, and a further reduction in 60-cycle component will be obtained.

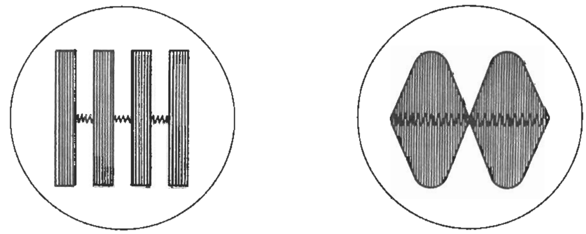
Using the pulser with a typical s.s.b. exciter and amplifier, about 2 volts of 1000-cycle audio at the input gave sufficient signal at the microphone jack to operate the transmitter. The amplifier ran 175 ma. when properly loaded for maximum p.e.p. with the two-tone test signal and 120 ma. with the keyed two-tone test. The plate voltage on the amplifier was 1000 for the tests, so the maximum p.e.p., as indicated by the two-tone test, was 275 watts [1000 (0.175 ÷ 0.636)]. With the pulsed

[3] The single tone is sufficient for producing the two-tone test signal if the sideband generator is of the phasing type and is temporarily set up for transmitting both sidebands without carrier—i.e., with one audio channel disconnected from the balanced modulator. With most filter-type generators it will be necessary either to supply a *two*-tone signal to the pulser, or to unbalance the modulator so that there will be enough carrier to supply the other (equal-amplitude) tone. In the latter case the factor 0.44 mentioned subsequently will not apply.—*Editor.*



This simple "pulser" can be used to "key" an audio tone being used in the two-tone testing of linear amplifiers. A triode, small power transformer and a few capacitors and resistors are all the circuit requires.

FIG. 5—(A) Pulsed two-tone test observed with slow sweep and 60-cycle sinc signal (B) Pulsed two-tone test observed with fast sweep and audio-tone sync signol.



(A) (B)

two-tone test the indicated power input was 120 watts, and the divisor works out to be $120/275 = 0.44$. This is low enough to give less than a kilowatt d.c. input when you're hitting 2-kilowatt input peaks.

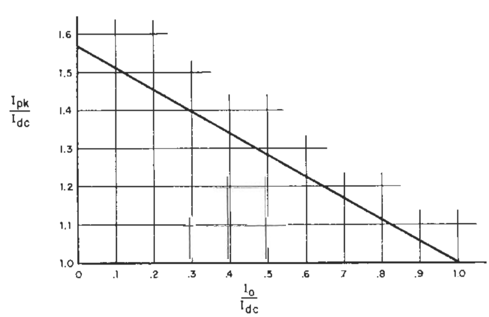
We make no claim that this is a highly refined way to check a linear. Actually it is the simplest way we could think of (other than the bug key), and anyone who wants to derive the factor 0.44 mathematically is welcome to try it. We do claim that this is a simple gadget for pulsing the two-tone test through your amplifier and permitting maximum peaks with little strain on the tube. When you try a slow sweep on the scope, the pattern will be as in Fig. 5-A, which shows the keyed character of the signal. Some audio leaks through the pulser even when the plate voltage is negative, and that's why a small signal appears between the large ones. When a fast sweep is used, synchronized with the audio input tone, the familiar two-tone test pattern is obtained. The "garbage" around the base line is merely the signal that leaks through during the "off" time of the pulser; it wouldn't be there if the electronic bug key were used for pulsing the audio tone.

"Crowding" a Small Amplifier

The fact that a speech s.s.b. signal has a rather low duty factor permits running tubes at a peak envelope power that would burn them up under steady operation at this power. To visualize it, think of the c.w. case. You might have a tube rated at 50 watts plate dissipation, and under c.w. conditions you might run 200 watts input in Class C and never have any trouble. The tube could be operated to handle perhaps 300 watts or so, *if* you confined yourself to sending a short "dit" every second or so. This is obviously impractical in c.w. work, but it is closely parallel to the case we have in s.s.b. linear operation. We can "pour the coal" to an amplifier for short periods of time if these periods are so short that the tube doesn't get a chance to overheat. The problem in s.s.b. work is to tune up the amplifier. You have to check the peak inputs to be sure the amplifier isn't going out of linearity, and you have to hold the long-time d.c. input down to avoid overheating and tube failure. Pulsing is the answer. Using the technique described earlier, either with an electronic bug or the simple tube pulser to key the audio tone input, will enable you to get the most out of any given amplifier. Obviously an oscilloscope and a source of audio tone are required, but you really need these anyway if you are to be sure of what's happening. With small tubes and higher plate (and screen) voltage than the book says, pulse tuning of the tubes will allow the maximum output to be obtained on voice peaks without distortion. Just don't forget and whistle or sustain a tone into the mike, unless you're prepared to replace the tubes!

Effect of Idling Current on Plate-Meter Readings

The accompanying graph, based on data supplied by Leigh Norton, W6CEM, shows the relationship between no-signal d.c. idling current (I_0), d.c. plate current as read by the plate milliammeter at full two-tone input (I_{dc}), and peak-envelope d.c. plate current (I_{pk}). The chart assumes distortionless linear amplification, but will be sufficiently accurate if the p.e.p. input is kept below the flattening point. Note that I_0 and I_{dc} are both values that can be read directly from the meter, but that I_{pk} cannot.



» In operating power tetrodes as linear amplifiers it is important to measure screen current, not only because of the danger of excessive power input to the screen, but because the screen current is one of the best indicators of proper loading.

Tetrode Screen Current

DAVID D. MEACHAM, W6EMD

Perplexing screen-current behavior has probably disturbed many amateurs, particularly single-sideband operators. The need for a thorough discussion of the subject has prompted this article. Class AB_1 operation has been chosen for discussion because of its current popularity as a means of achieving good linearity and TVI-free operation. The information given herein assumes grid-driven conditions, but it applies equally well to cathode-driven tetrodes operated Class AB_1 with normal d.c. voltages on the grid and screen, provided that grounded-grid characteristic curves are used for computations.

Screen Characteristics

Fig. 1 shows a set of *constant-current characteristics* for a typical 4CX300A. The term "constant current" is used because the lines plotted are lines of constant plate, screen, or grid current. The grid-voltage scale appears on the left axis and plate voltage is shown horizontally. These curves depict *instantaneous values* of plate and screen current for any given grid- and plate-voltage condition. In this reproduction, the grid-current lines are omitted because grid current is not drawn in Class AB_1 operation. The curves are valid only for a *fixed* screen voltage (350 volts in this case).

Inspection of Fig. 1 will reveal that the lines of constant plate current are nearly horizontal, whereas the constant-screen-current lines are tilted upward from left to right and are concentrated in the left-hand region of the plot. This is generally true for all tetrodes and accounts for the fact that the screen-current meter is *the most sensitive indicator of resonance.* This important fact will be explained subsequently.

Let us plot a typical operating line [1] on our set of curves, as in Fig. 1. Point O (at −55 volts on the grid in this case) is the operating point at which the tube rests with zero r.f. grid drive. Straight line OA represents a tuned r.f. circuit load (a pure resistance at the operating frequency).[2] As 100 volts peak-to-peak grid drive is applied, the first positive half cycle can be represented by a point moving along the operating line from O to A and back to O again. During this half cycle, the grid-voltage swing from −55 volts up to −5 volts and back to −55 volts has caused the plate current to swing from the value at point O (100 ma.) up to the value at point A (850 ma.) and back to 100 ma. again. At the same time, the plate voltage swings from 2000 volts down to 500 volts. The a.c. plate current is made up of all the instantaneous values intercepted by the point traveling along the operating line. The same is true of screen current. During the other 180 degrees of the driving cycle, our point merely travels from O down the slope through cutoff to a point oppo-

From "Understanding Tetrode Screen Current," July, 1961, *QST*.

[1] This is different from the usual load line associated with audio calculations using plate characteristic curves.

[2] OA is actually only half the operating line length. The other half continues from O out beyond the right-hand edge of the chart for an equal distance and represents the effect of the negative half-cycle of grid driving voltage as it swings down to −105 volts and back to O again. This half of the operating line is not important since the tube does not "work" during the negative half cycle.

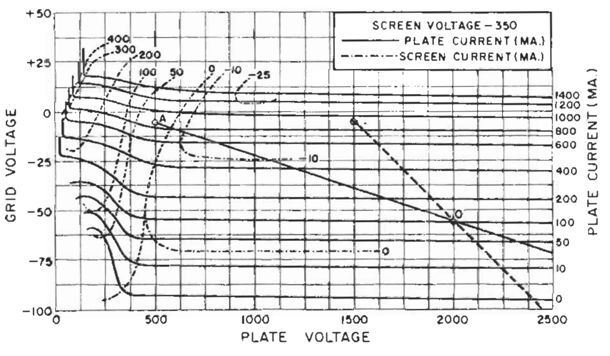


FIG. 1—Typical constant-current characteristics for the Eimac 4CX300A tetrode.

site −105 volts on the grid-voltage scale and back to point O again along the operating line. Thus, the negative-going grid voltage swings the plate current down to cutoff (for a small portion of the cycle). Plate voltage continues on up to 3500 volts and back down again due to the fly-wheel action of the plate tank circuit.

Drive and Tuning

Now that we can predict exactly what the screen and plate current will be for any *instantaneous* point during the grid-voltage cycle, let us ask some more probing questions. What happens when we cut our grid-driving voltage in half? The answer is simple. The *length* of our operating line is merely cut in half! The grid voltage swings to only one half the original peak-to-peak amplitude and the operating point O is still the center of the new operating line length. Now what happens if we detune the plate tank circuit? Detuning the plate circuit actually changes the plate load impedance. How does this appear on our set of curves? It tilts or rotates the operating line *about the operating point* O. As the load impedance is lowered (detuning from resonance), the operating line[3] assumes a steeper angle (a zero-impedance load would be represented by a vertical operating line).

As "seen" by the tube, the act of tuning to resonance amounts to increasing the load impedance to a maximum value consistent with the degree of antenna loading selected. Thus, the operating line will have *minimum slope* at resonance. Notice the angle at which our typical operating line in Fig. 1 cuts the constant-plate-current lines. It's a small angle. As the plate tank circuit is tuned to a point out of resonance, the operating line might assume the position indicated by the dashed line[3] (lower impedance). Note that the angle between the dashed line and the plate-current lines has not changed radically, and that our moving point will still intercept *essentially the same plate-current values*. This is precisely the reason that plate current in a tetrode is not a good indicator of resonance (very little dip). Look at the screen current. It consists of zero or even negative values in the out-of-resonance position. At resonance, though, it is positive. Thus, a peak in screen current *indicates resonance*.

During the rotation of the operating line while tuning, its length actually changes, since it is confined vertically only by the constant peak-to-peak amplitude of the grid-driving voltage (two imaginary horizontal lines, one at −5 volts and one at −105 volts). The length increases as resonance is approached and reaches a maximum at resonance. As the length increases, point A penetrates the heavy-screen-current region and the d.c. screen current reaches a sharp peak at resonance.

Loading

What happens if we change the antenna loading? This merely changes the plate-load impedance (still resistive). Again, the effect is to tilt the operating line about the operating point. As the load impedance is lowered (more coupling), the operating line assumes a steeper angle (such as the dashed line). It is easy to see that as loading increases, screen current decreases. Thus, screen current is *also an indicator of loading*. Screen current varies somewhat from tube-to-tube of a given type, but if each tube is loaded to the same value of screen current at resonance (with the same drive) power output differences will be small, and loading and linearity will be essentially the same.

D.C. Meter Readings

During the r.f. cycle, our point traverses the operating line and intercepts many different instantaneous values of screen current and plate current. The *average* of all these values is what the d.c. meter in the circuit reads. The *fundamental frequency component* of plate current is utilized in the plate circuit to produce output (except in a multiplier where use is made of a harmonic component of plate current). For a given operating line, both of these values can be calculated.[4] Suffice it to say that for Class AB_1 operation, the d.c. meter reading is approximately one third the peak value of current at the top of the operating line, and the fundamental component of plate current is approximately one half the peak value.

Tune-Up Procedure

Contrary to somewhat popular opinion, a linear amplifier should *never* be loaded for maximum power output. Loading should be set to obtain a predetermined value of screen current under single-tone or inserted-carrier driving conditions. Ideally, loading should be set for minimum distortion—a rather difficult feat in practice. It is recommended that the amateur try to duplicate as nearly as possible a given set of data-sheet conditions as presented by the tube manufacturer. These typical operating conditions are usually given for peak-envelope operation (single-tone or inserted-carrier) and represent the maximum input on c.w. or the peak-envelope-power input (*not* meter peaks) on single sideband. After adjusting drive, tuning, and loading to duplicate a given set of conditions, the single tone (or carrier) is removed and the single-sideband audio gain is adjusted so that grid current is

[3] The tank-circuit impedance would no longer appear resistive at the operating frequency, but would contain a reactive component. Under these conditions, the operating line becomes an ellipse whose center is point O and whose major axis is represented by the dashed line.

[4] By the use of the Eimac Tube Performance Computor, Application Bulletin No. 5, which is based on the method presented by Chaffee in the *Review of Scientific Instruments*, October, 1936.

never drawn and the condition adjusted for above is never exceeded on peaks. The peak-to-average ratio of d.c. plate current (as read on a fluctuating meter) varies, with the individual voice, from about 2:1 to over 3:1. Thus it is normal on voice peaks for the plate-current meter to read no more than *half* the value of current obtained in the maximum static single-tone condition.

A straightforward tune-up procedure consists of the following steps:

1) Insure that the tetrode amplifier is neutralized and free of parasitics.

2) With recommended heater, plate, and screen voltages applied, adjust the d.c. grid bias to obtain the recommended zero-signal value of plate current. This value affects linearity and plate dissipation.

3) Connect a suitable dummy load and set the loading control for rather heavy loading.

4) With a single-tone source, gradually increase the drive from zero to a value that produces a significant though small change in screen current.

5) Resonate the plate tank circuit by tuning for a *peak* (in the positive direction) *in screen current.*

6) Resonate the grid tank circuit (if any) by watching for a *peak in plate current.*

7) Now increase the drive until either the desired value of a single-tone screen or plate current is reached (whichever is reached first).

8) Without drawing grid current, adjust loading, plate-tank tuning, and drive level to duplicate as nearly as possible a given set of data-sheet peak-envelope conditions. Remember that plate current increases with drive, whereas screen current peaks at resonance and decreases with heavier loading.

After matching a set of data-sheet conditions, the amplifier is ready to connect to an antenna. With a suitable antenna connected, it should be easy to repeat the operation obtained in Step 8 above by merely adjusting plate-tank tuning and loading with the same drive level as before. Now set up for voice single-sideband drive and adjust the audio gain for the highest level possible without drawing grid current on voice peaks or flat-topping (check this with a scope).

Reverse Screen Current

Most transmitting tetrodes employing oxide-coated cathodes exhibit negative screen current under certain conditions of operation. This is nothing to get alarmed about—it merely means that on the average, more electrons are leaving the screen than are being intercepted by the screen. This results because of secondary electron emission at the screen grid. Small values of negative screen current are not detrimental to tube operation and are quite normal for some tetrodes. Such values usually appear under heavily-loaded conditions or during the idling condition.

Large values of negative screen current are abnormal and should be avoided. Excessive secondary emission usually results in higher values of intermodulation distortion. This condition also prevents an accurate determination of screen dissipation.

Use A Screen-Current Meter!

In conclusion, it should be obvious to the amateur that a screen-current meter is a vital necessity in modern transmitters employing tetrodes. By proper interpretation of screen-current readings, one can easily tune to resonance and properly load the tetrode amplifier. The plate-current meter is useful *only* as an indicator of drive level and *average* plate-input power (knowing the plate voltage). One more meter—for grid current—is useful but not absolutely necessary. A one-milliampere meter in the grid circuit will warn the operator by a slight kick when grid current is being drawn on voice peaks.

» An amplifier-cum-power supply on one chassis is a convenient package to have, especially when it fits in a receiver-size table-type cabinet and can be run at a half kilowatt input. Add constructional simplicity, a minimum of operating controls, and pleasing appearance and you have the assembly described here.

This amplifier uses a pair of 811As in grounded-grid, and is complete with power supply on a 13 × 17 × 4-inch chassis. The rack panel is 10½ by 19 inches. Front-panel controls, arranged to give a balanced appearance, include the plate tuning capacitor and band switch in the center, filament and plate power switches with their pilot lights at the lower left, sensitivity control and forward-reflected power switch for the directional coupler at the lower right, variable loading capacitor and auxiliary loading-capacitor switch underneath the 0-1 milliammeter at the right, and the grid-cathode milliammeter with its switch at the upper left.

A Table-Top Half Kilowatt

ERNEST A. COONS, W1JLN/FOE

The amplifier shown here will run at about 500 watts input on c.w.—or p.e.p. input as an s.s.b. linear—on all bands from 80 through 10 meters. I wanted a simple amplifier that would be small and neat, and with which I could change bands quickly. The result is small enough to sit on the operating table right along with the rest of the station equipment; no need for big racks here!

Using a pair of 811As in parallel in the grounded-grid circuit, this rig is a good one to use following a low-power c.w. transmitter for a worth-while increase in power output. As a linear amplifier following an s.s.b. exciter it requires no swamping, not only because the 811A grids provide a fairly constant load in themselves, but also because the fed-through power with ground-grid presents an additional constant load to the driver. The total driving power needed on any band is less than 20 watts.

An additional useful feature is a built-in directional coupler using a version of the "Mickey Match."[1] Besides its obvious application for checking the s.w.r. on the transmission line to the antenna or for help in tuning up a coax-coupled feeder-matching circuit or "antenna coupler," it is practically indispensable as an indicator of relative power output in tuning the amplifier, for reasons which will be discussed later.

The Circuit

A number of tube types could be used in an amplifier of this power class, but I decided on the 811As because they do not need a bias supply and are not expensive. (Surplus 811s can be used if you don't want to buy new tubes; the ratings are not quite as high but they can be pushed a bit in intermittent service such as c.w. and s.s.b.)

The complete circuit is shown in Fig. 1. Don't expect to find anything startling—the whole thing was taken from parts of proven circuits here and there in the *Handbook* and put together to meet my needs. To save trouble and work, standard components were used throughout—the only special construction is the shielding and a few simple r.f. chokes. The tube filaments are driven directly from coax input from the driver; no tuning is used in this circuit. The filaments are kept above ground by the B & W type FC15 filament choke.

The plate tank is the familiar pi network, using a B & W type 851 tapped coil and band-switch assembly. This assembly has been modified slightly in two respects: First, the copper-strip 10-meter coil normally mounted at the top of the rear plate was taken off and moved

From January, 1960, *QST*

[1] Bunce, "The 'Mickey Match'," *QST*, November, 1958.

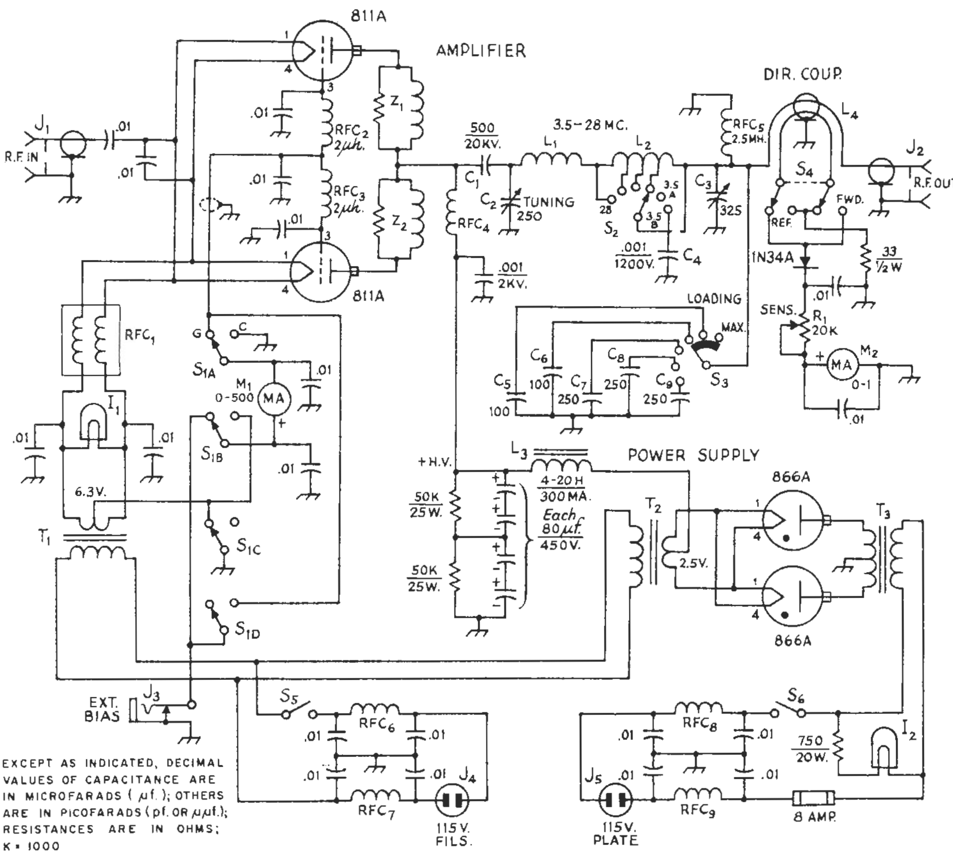


FIG. 1—Circuit diagram of the parallel-811A grounded-grid amplifier. Unless otherwise specified, fixed capacitors are disk ceramic, 600-volt or higher rating.

C_1—500 pf., 20,000 volts (TV "doorknob" type).

C_2—250-pf. variable, 2000 volts (Johnson 250E20).

C_3—325-pf. variable, receiving type (Hammarlund MC-325-M).

C_4–C_9, inc.—1200-volt mica, case style CM-45.

I_1, I_2—6.3-volt dial lamp, 150-ma. (No. 47).

J_1, J_2—Coax connector, chassis mounting.

J_3—Closed-circuit phone jack.

J_4, J_5—115-volt male connector, chassis mounting (Amphenol 61-M1).

L_1, L_2, S_2—5-band pi-network coil-switch assembly; see text (B & W 851).

L_3—Swinging choke, 4-20 henrys, 300 ma. (UTC S-34).

L_4—Section of coax line with extra conductor inserted; see Footnote 1 for construction references.

M_1, M_2—Milliammeter, 3½-inch plastic case (Triplett 327-PL).

R_1—20,000-ohm composition control, linear taper.

RFC_1—Filament-choke assembly, to carry 8 amp. (B & W FC15).

RFC_2, RFC_3—2 μh. (National R-60).

RFC_4—90 μh.; 4⅜-inch winding of No. 26, 40 t.p.i., on ¾-inch ceramic form (B & W 800).

RFC_5—2.5 mh., any type.

RFC_6–RFC_9, incl.—18 turns No. 14 enam., closewound, ½-inch diam., self-supporting.

S_1—4-pole 2-position rotary, nonshorting (Mallory 3242J or Centralab 1450).

S_2—Part of tank assembly; see L_1L_2.

S_3—Miniature ceramic rotary, 1 section, 1 pole, 6 positions used, progressive shorting (Centralab 2042).

S_4—Miniature ceramic rotary, 1 section, 2 poles, 2 positions used, nonshorting (Centralab 2003).

S_5, S_6—S.p.s.t. toggle.

T_1—Filament transformer, 6.3 volts, 8 amp. min. (UTC S-61).

T_2—Filament transformer, 2.5 volts, 10 amp. (UTC S-57).

T_3—Plate transformer, 3000 volts center-tapped, 300 ma. d.c. (UTC S-47).

Z_1, Z_2—Parasitic suppressor, 100-ohm 2-watt carbon resistor assembled inside 2½-turn coil of No. 16 tinned, ½-inch diameter, ¾-inch long.

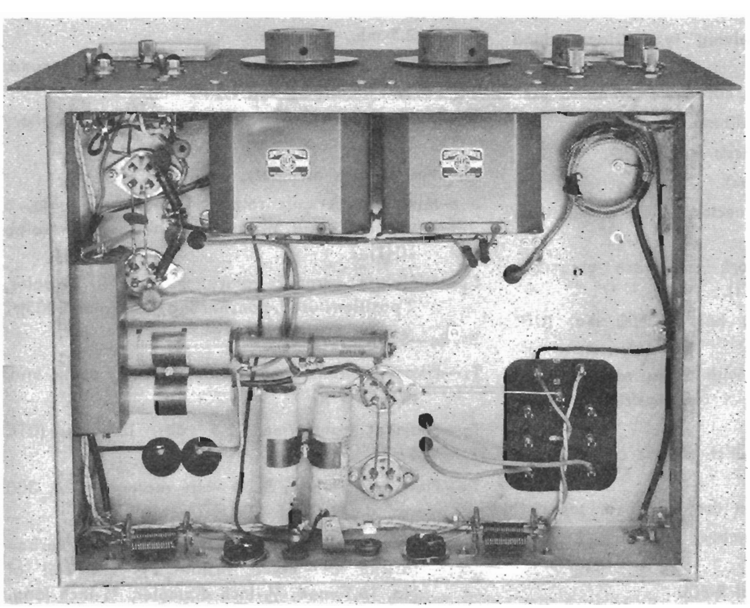
so that it is supported between the tank assembly and the stator of the tank tuning capacitor as shown in the top view. A short length of copper strip is bolted between the free end of the coil and the right-hand stator connection of the tuning capacitor, to support the free end. This change was made in order to avoid the long lead that would have had to be run from the capacitor to the regular input terminal on the tank assembly, since this terminal is at the right-hand side of the assembly as viewed from the top. The turns of the 10-meter coil were also squeezed together a bit to increase the inductance, because it was found that a rather large amount of capacitance had to be used to tune the circuit to the band with the coil at its original length. The length is now 1⅝ inches between mounting holes.

The second modification was the addition of a pair of switch contacts on the rear switch plate of the tank assembly. There is an extra position on this plate with holes already provided for contacts, and it seemed like a good idea to use a set of contacts here to switch in additional output loading capacitance on 80 meters, where a large output capacitance is needed. The variable loading capacitor, C_3, and the five fixed mica capacitors, C_5 to C_9 inclusive, give continuous variation of capacitance up to 1275 pf. on all bands, including the regular band-switch position for the 80-meter band. However, if the switch is turned to the extra position an additional 100-pf. mica capacitor is connected in parallel, so that continuous variation of capacitance to over 2200 pf. is possible on 80. This takes care of cases where the load resistance happens to be unusually low or reactive.[2]

A 500-ma. d.c. meter is used for reading either total cathode current or grid current alone. The cathode current is read in preference to plate current because of safety considerations. Putting the meter in the hot d.c. plate lead leaves nothing but a little plastic insulation between the high voltage and the meter adjusting screw. Although the meter could have been connected in the negative plate supply lead since the power supply is self-contained, I prefer to have the negative firmly grounded to the chassis. It is a bit of a nuisance to have to subtract the grid current from the cathode current in order to find the plate current, but it isn't serious. The d.c. grid circuit has a jack, J_3, for introducing external bias either for blocked-grid keying or for cutting off the plate current during receiving, and a four-pole switch, S_1, is therefore needed for handling the meter switching while keeping all circuits functioning normally.

The power supply uses 866As with a plate transformer giving 1500 volts each side of the center tap, and working into a single-section choke-input filter. The filter capacitor consists of four 80-μf. electrolytics connected in series to handle the voltage, giving an effective filter capacitance of 20 μf. This supply is running well below its capabilities in the intermittent type of operation represented by c.w. and s.s.b., and the amplifier is somewhat "overpowered" in this respect. A lighter plate transformer can be used since the average current in regular operation is only about half the maximum tube rating of 350 ma. for the two tubes.

[2] These contacts can be obtained directly from the manufacturer of the tank assembly. To secure a set of contacts with mounting hardware, send one dollar to Barker & Williamson, Beaver Dam Road, Bristol, Penna., specifying the type of tank assembly for which they are wanted. The contacts are not catalog items and are not available through dealers.—*Editor.*



In this below-chassis view, the two filament transformers are at the top, mounted on the chassis wall. The 811A sockets are at the upper left. The rectangular box on the left-hand wall contains the FC15 filament-choke assembly. The "Mickey Match" directional coupler is at the upper right. Filter capacitors and the bleeder resistors are in the lower section. A.c. inlets, fuse holder, bias jack, and the 115-volt line TVI filters are on the bottom chassis wall.

The a.c. inputs to both filaments and plates have TVI filters installed right at the a.c. connectors. The chokes in these filters, RFC_6 to RFC_9 inclusive, are homemade by winding 18 turns of No. 14 enameled wire close-wound on a half-inch dowel or drill.

Construction

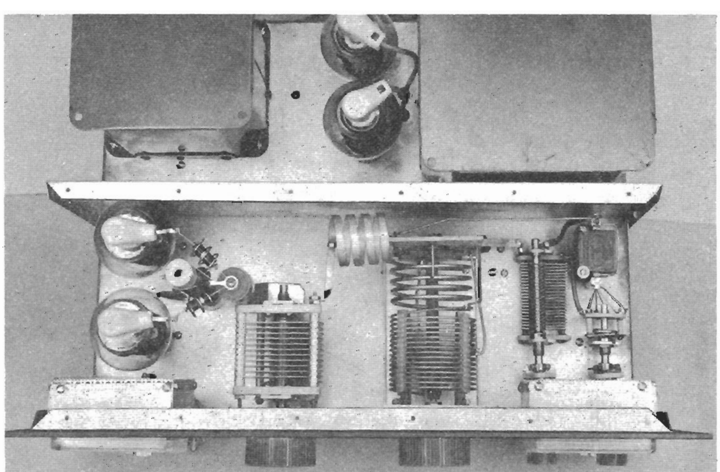
The r.f. layout shown in the top view is almost an exact copy of the circuit layout as given in Fig. 1. The plate blocking capacitor, C_1, is mounted on a small right-angle bracket fastened to the left-hand stator connection of the tank capacitor, C_2. The tube plates are connected to C_1 through individual parasitic-suppressor assemblies, Z_1 and Z_2. The hot end of the plate choke, RFC_4, also connects to this same point. The tank capacitor is mounted on ¾-inch ceramic pillars to bring its shaft to the same height as the switch shaft on the tank-coil assembly. The capacitor is grounded by connecting the bottom of its frame through a half-inch wide strip of aluminum to essentially the same point at which the plate-choke bypass capacitor, a 0.001-μf. 2000-volt disk, is grounded. The ground end of the aluminum strip actually is under the bottom of the plate choke, and the ground lug for the bypass capacitor is just to the left. This strip, plus short leads in the circuit from the tube plates through the tank capacitor to ground, keeps the resonant frequency of the loop thus formed well up in the v.h.f. region; this is important because it permits using low-inductance parasitic chokes in shunt with the suppressor resistors, and thus tends to keep the r.f. plate current at the regular operating frequencies out of the resistors. With other tank grounding arrangements originally tried, larger parasitic chokes had to be used and it was impossible to prevent the resistors from burning up when operating on 10, 15 and even 20 meters. Now they do not overheat on any frequency, and v.h.f. parasitics are nonexistent—although without the suppressors the parasitics are only too much in evidence.

The output loading capacitors, C_3 through C_9, are mounted toward the rear so the leads from the tank coil can be kept as short as possible. A length of copper strip is used between the coil and the stator of C_3; originally this lead was No. 14 wire but on 10 meters the tank current was enough to heat it to the point of discoloration. The ground lead from the fixed units, made to the rear bearing connection of C_3, is also copper strip. C_3 and S_3 are operated through extension shafts, using Millen flexible couplings to simplify the alignment problem.

Underneath the chassis, each 811A grid is bypassed directly to the socket-mounting screw nearest the plate choke (right-hand side of the socket in the bottom view). The d.c. leads have small chokes, RFC_2 and RFC_3, with additional bypasses for good r.f. filtering, particularly at v.h.f. since grid rectification generates harmonics in the TV bands. The filament choke, RFC_1, is mounted so that the filament side is close to the filament terminals on the tube sockets; the other end is bypassed directly to the chassis.

The shielding around the amplifier consists of two pieces of sheet aluminum and a perforated aluminum ("do-it-yourself" type) cover having the shape of an inverted U. The top view shows how the rear wall is made. Its edges are bent to provide flanges for fastening the cover with sheet-metal screws, and there is a similar flange projecting to the rear at the bottom for fastening the wall to the chassis. The front piece extends the full height of the panel and is identically drilled and cut out for meters and controls. It has flanges at the top and extending down the sides from the top to the chassis. The cover itself extends down over the sides of the chassis for about one inch. Numerous screws are used for fastening the cover, to prevent leakage of harmonics.

The r.f. section with the shield cover removed. Components here are readily identifiable by reference to the circuit. The meters are enclosed in rectangular boxes made from thin aluminum sheet, formed to be fastened by the meter mounting screws. The back covers on these boxes are made from perforated aluminum, folded over at the edges and held on the boxes by sheet-metal screws. The switch for shifting the 0-500 milliammeter (left) from grid to cathode is concealed by the box which encloses the meter.



The shields over the meters are made as described in the caption for the inside top view. Meter leads are bypassed to the shield boxes where they emerge.

The *Handbook* should be consulted for methods of checking the directional coupler.

Operating Conditions and Tuning

The voltage delivered by the power supply is approximately 1500 volts with no drive and with the tubes taking only the no-bias static plate current, which is about 60 ma. At the full load of 350 ma. the voltage is slightly under 1400. Optimum operating conditions for 1400 volts at 350 ma. peak-envelope power input as an s.s.b. linear call for a peak-envelope grid current of 60 ma. The peak-envelope tube power output is close to 350 watts under these conditions. The same operating conditions are also about optimum for c.w.

The behavior of the cathode current when tuning a grounded-grid triode amplifier is extremely confusing, and the meter is principally useful as a check on operating conditions rather than as a tuning indicator. The best indicator of proper tuning of the plate tank capacitor is the forward-power reading of the directional coupler. For any trial setting of the loading controls and driving power, *always* set the plate tank capacitor control at the point which results in a maximum reading on the power-output indicator. The power indications are only relative, of course, and the sensitivity control should be set to give a reading in the upper half of the scale.

The objective in adjusting loading and drive is to arrive at maximum power output simultaneously with a plate current of 350 ma. and a grid current of 60 ma.—that is, a total cathode current of 410 ma. when the grid current reading is 60 ma. The loading is critical. If the amplifier is not loaded heavily enough the grid current will be too high and the right value of total cathode current either will not be reached or, if reached, the amplifier will be operating in the "flattening" region as an s.s.b. linear. (It can be operated this way on c.w., however, since linearity is unimportant here.) If the loading is too heavy, the grid current will be low when the cathode current reaches the proper value, but the efficiency will be low and the tubes will overheat.

Getting the knack of it takes a little practice, but when the job is done right the tubes will run cool on all bands in regular operation. Running key-down over a period of time may show just a trace of dark red color on the plates since the input and dissipation are somewhat over ratings under these operating conditions, although perfectly satisfactory with ordinary keying or s.s.b. voice.

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 Compact 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 ⅛-inch aluminum suspended 1¼ inches below the chassis. Eight ¼-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, C_1, 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,

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[1] See Peck, "Homebrew Custom Designing," *QST*, April, 1961.

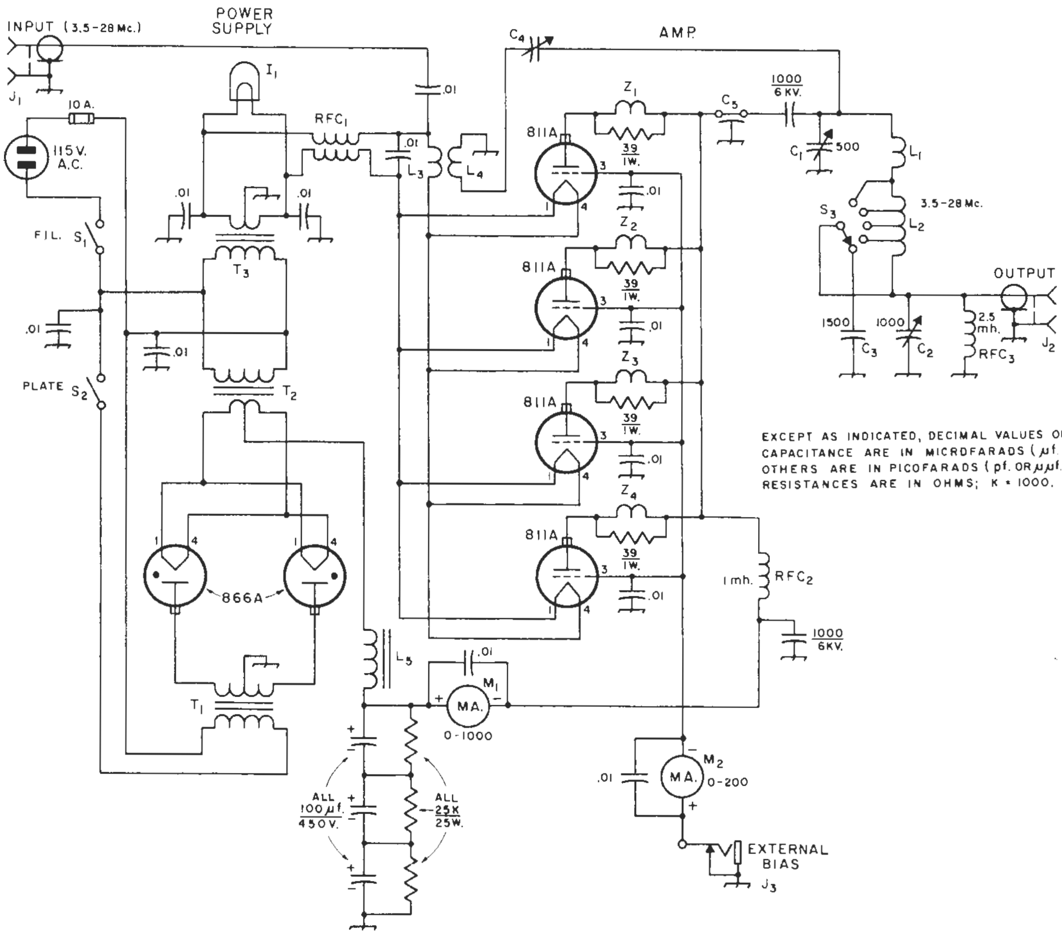


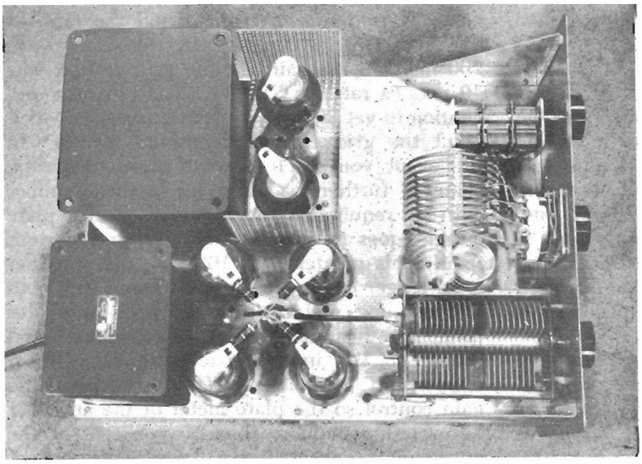
FIG. 1—Circuit of the high-power grounded-grid linear and its built-in power supply. Capacitors not listed below are disk ceramic, except those marked with polarity which are electrolytic. Resistances are in ohms.

C_1—500-pf. 2000-volt variable (Johnson 154-3/500E20 or similar—see text).
C_2—Triple section broadcast replacement variable, 365 pf. per section, sections in parallel.
C_3—2500-volt mica.
C_4—Neutralizing capacitor—approx. 6 pf. 0.06-inch spacing or greater (Bud CE-2028).
C_5—V.h.f. bypass (4-inch length of RG-58/U as connecting lead).
I_1—6.3-volt panel lamp.
J_1, J_2—Coaxial receptacle (SO-239).
J_1—Closed-circuit jack.
L_3, L_2—Pi-network inductor (Illumitronic PiDux No. 195-1) approx. inductances in use: 0.4, 0.7, 1, 2.2 and 4.5 µh., respectively, for 10–80 meter, L_2 wound with No. 8 wire, L_1 wound with ½-inch copper strap (see text).
L_3—6 turns No. 14, ½-inch diam., close-wound.
L_4—5 turns insulated hookup wire wound over L_3.
L_5—Filter choke: 5–8 h., 300 ma. (Stancor C-1722 or similar).
M_1—0–1000-ma. d.c. meter.
M_2—0–200-ma. d.c. meter.
RFC_1—Bifilar filament choke (B & W FC-15).
RFC_2—R.f. choke: 1 mh. 600 ma. (National R154-U).
RFC_3—2.5-mh. r.f. choke, 50–100 ma.
S_1, S_2—S.p.s.t. toggle switch.
S_3—Band switch (see text).
T_1—1250-volt (d.c.) 300-ma. plate transformer (Stancor PT-8313 or similar).
T_2—Filament transformer: 2.5 volts, 10 amp. (Stancor P-3024 or similar).
T_3—Filament transformer: 6.3 volts, 16 amp. (Triad F-22A or similar—see text).
Z_1–Z_4 incl.—Parasitic suppressor; 7 turns No. 18 wire, wound on and connected across a 39-ohm 1-watt resistor.

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.

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



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 reactifiers. 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.)*

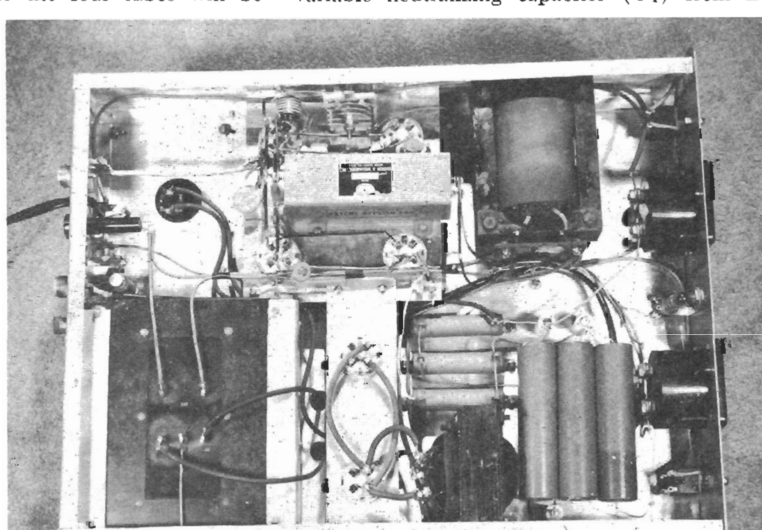
single pair of 811As, proved incapable of supplying the required voltage through the filament chokes with four tubes in the circuit. The secondary, which turned out to be wound with No. 16 wire, was removed, the turns being carefully counted as they were unwound. A new secondary was wound with No. 14 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), ½-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-μf. 450-volt electrolytic capacitor and 25K 25-watt bleeder resistor should be put in series with the three shown for the 1250-volt supply.

» *If you do all of your operating on one band, there isn't much point in building a multiband transmitter. On the other hand, if you are a band-hopper, individual finals requiring little if any adjustment when bands are changed are the ultimate in convenience. Ergo, these grounded-grid units described by K9LKA should have a universal appeal.*

Single-Band Grounded-Grid Linears

LARRY KLEBER, K9LKA

A great many amateurs using transmitters in the 75- to 150-watt class have one favorite band. Most of these operators would like more output, but hesitate to buy or build a multiband amplifier for several reasons. Aside from the cost, it just doesn't seem sensible to use an amplifier that will operate on five bands when operation on only one is desired. Even the multiband operator will find plenty of argument in favor of the single-band unit plan. Construction is simplified, usually resulting in less-frequent need for servicing, and servicing when required is much easier to handle. No single unit represents a major construction project, and handswitching can be much less complicated.

Each of the single-band grounded-grid linears shown in the photographs uses a pair of 813s in parallel to provide a one-kilowatt power capability. The tubes, with the screens grounded, operate as high-μ triodes, thereby eliminating the need for a screen supply. Operating Class B, the efficiency of the tubes will run between 65 and 70 per cent in s.s.b. or c.w. service.

Costwise there is quite a spread. If you're willing to scrounge around, raid the junk box and do some horse trading, you can build each unit complete for less than \$30. If you buy all the parts new, the cost will be approximately \$60, excluding tubes.

The Circuit

The r.f. driving power is fed to the filaments of the two 813s through a 0.01-μf. ceramic capacitor, as shown in Fig. 1. The filament transformer is isolated from r.f. by the bifilar filament choke RFC_2.

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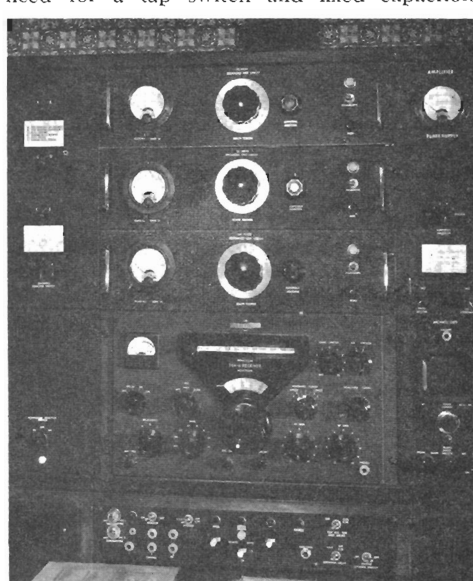
A built-in supply delivers 0 to 37 volts of bias to the control grids of the 813s, the value being determined by the setting of R_1. With the terminals of J_3 open, the voltage rises to –168, biasing the tubes beyond cutoff, and no plate current will flow. Shorting J_3 reduces the bias to the value selected by adjustment of R_1. Leads from J_3 should be run to relay contacts, such as auxiliary contacts on an antenna relay which close while transmitting. Cutoff bias on stand-by eliminates the "hash" which often bothers reception, especially when using a t.r. switch.

High voltage is fed to the 813 plates through RFC_3 and RFC_4. A 500-pf. 20-kv. doorknob capacitor, C_3, is used to isolate the high-voltage supply from the pi-network circuit. The rating of RFC_4 is only 300 ma. but, since the plate current swings up to 400 ma. only on peaks, the rating of this choke is satisfactory.

The two-section variable output capacitor C_2, with a total maximum of 730 pf., eliminates the need for a tap switch and fixed capacitors.



The operating position at K9LKA showing linear amplifiers for 10, 15 and 20 meters mounted in a rack above the receiver.

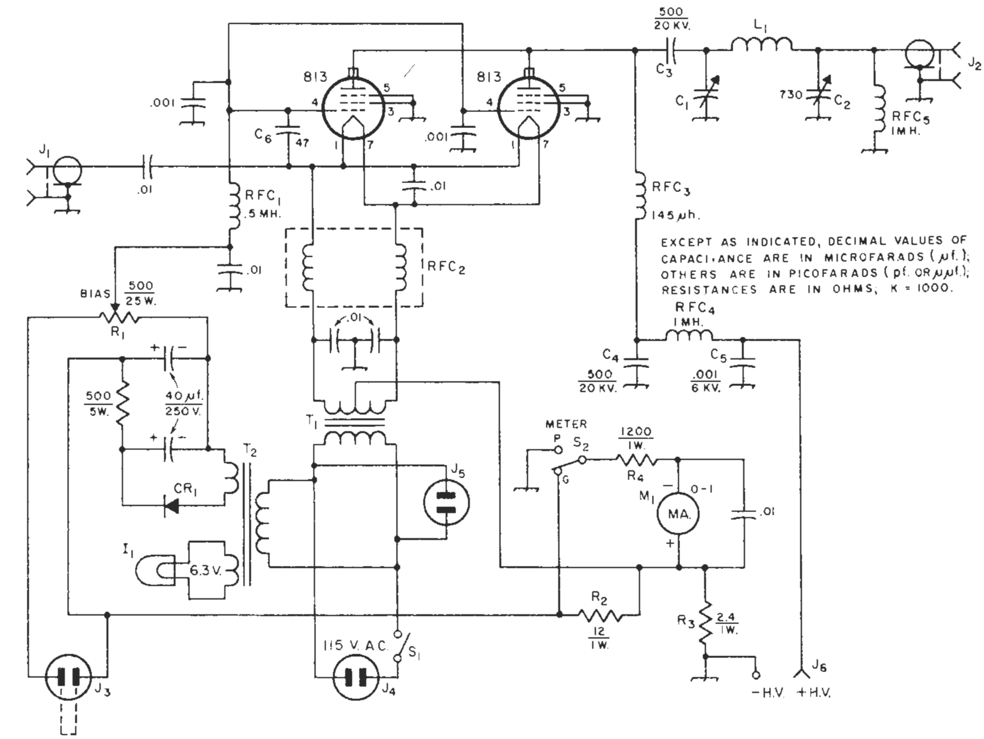


FIG. 1—Circuit used in the single-band high-power linear amplifiers. Capacitors marked with polarity are electrolytic. Resistances are in ohms and resistors are ½ watt unless indicated otherwise.

C_1—Transmitting variable, 0.075-inch minimum plate spacing; 100 pf. for 21, 28 Mc., and 14 Mc., 150 pf. for 7 Mc., 350 pf. for 3.5 Mc. (Johnson, 100E30/154–14, 150E30/154–8, 350E30/154–10, or similar, respectively).

C_2—Dual 365-pf. variable, broadcast-replacement type, sections in parallel (may be necessary to add 330-pf. transmitting mica capacitor in parallel for 3.5 Mc.)

C_3, C_4—Ceramic, TV doorknob type (Sprague 20DK-T5 or similar).

C_5—Disk ceramic (Centralab DD60-102 or similar).

C_6—Ceramic (used for stabilizing in 21- and 28-Mc. amplifiers).

CR_1—130-volt 75-ma. selenium rectifier (Sarkes-Tarzian type 75).

I_1—6.3-volt dial lamp.

J_1, J_2—Chassis-mounting coaxial receptacle (SO-239).

J_3—Recessed a.c. connector, male (Hart & Hegeman 80329; takes type 80325 female cable connector. Standard female outlet with male plug may also be used).

J_4—Chassis-mounting a.c. plug.

J_5—Miniature a.c. receptacle (Cinch-Jones S-302-AB or similar).

J_6—High-voltage connector (Millen 37001).

L_1—3.5 Mc.—16 turns No. 12 2½-inch diam., 6 t.p.i. (B & W 3905–1 stock).
7 MC.—Same as above, 9 turns.
14 Mc.—10 turns ¼-inch copper tubing, 1½ inches i.d., turns spaced ⅛ inch.
21 Mc.—Same, 7 turns spaced 3/16 inch.
28 Mc.—Same, 4 turns spaced ¼ inch.

M_1—D.c. milliammeter, 3-inch.

R_1—Wire-wound control (Ohmite H-0156).

R_2, R_3, R_4—Meter multiplier resistors, wire-wound, 5 per cent.

RFC_1—0.5-mh. r.f. choke (National R-300).

RFC_2—Bifilar filament choke (B & W FC-15 or similar).

RFC_3—Plate choke (National R-175-A).

RFC_4, RFC_5—1-mh. 300-ma. r.f. choke (National R-300).

S_1—S.p.s.t. toggle switch.

S_2—S.p.d.t. slide switch.

T_1—10-volt 10-amp. filament transformer (Merit P-3146, Stancor P-6461 or similar).

T_2—Power transformer: 125 volts, r.m.s., 50 ma.; 6.3 volts, 2 amp. (Thordarson 26R38, Stancor PA-8421).

The pi-network output of these linears is designed for 50- to 70-ohm unbalanced loads.

To obtain separate grid- and plate-current readings, meter M_1 is switched across multiplier resistors R_2 and R_3, respectively. Since the grid circuit is returned to the center tap on the filament transformer, only plate current is read in the plate position of S_2.

Chassis Assembly

The panel is a standard 5¼ × 19 × ⅛-inch aluminum rack-style unit, while the chassis is

All amplifier units have the same panel design. This unit is the one used for 20 meters. Tuning and loading controls are at the center. The small knob in the lower right-hand corner is for adjusting bias.

made up of a pair of See-Zak[1] R45 rails (4 by 5 inches), a pair of R417 rails (4 by 17 inches), and a P517 panel (5 by 17 inches).

First, lay out the P517 panel according to the drilling template of Fig. 2. The rear-view photo will be useful as a check. After locating all holes with a prick punch, drill pilot holes at I and J with a small drill (No. 35 or 36). At this point, mark the outer or mounting side of the P517 panel with a permanent reference mark, such as a file or scribe mark, so that there will be no confusion. Next, place the P517 panel on top of the *rear* of the rack panel and, after centering it on the rack panel, clamp the two together and transfer the pilot holes at I and J. These are the shaft holes for C_1 and C_2, so they must match perfectly. Enlarge the two holes in both panels to 9/32 inch.

Drill all remaining holes whose sizes are indicated in Fig. 2. Exact sizes are not given for holes at H and K. These are for feedthrough insulators and should be drilled to fit the ones you have on hand. Mount the 2-contact Jones socket J_5 (the a.c. outlet for the ventilating fan) at A.

Cut a piece of ½×½-inch aluminum angle 4⅞ inches long. Using a No. 25 drill, place a hole ⅜ inch from each end and ¼ inch from the outer edge of the angle. Transfer these holes to the chassis panel at points L and M. Submount the 813 sockets at B and C, using ½-inch spacers. This submounting helps to keep the over-all depth of the amplifier, including the shielding enclosure, to 15 inches or less so that it can be mounted in a standard rack cabinet.

Remove the short ceramic insulator from the mounting bracket of the National R-175A choke, RFC_3, and mount a 500-pf. 20-kv. capacitor, C_4, in its place. Place two solder lugs on the top terminal of this capacitor, and then thread the ceramic insulator onto the capacitor stud. One of the solder lugs is connected to the h.v. feedthrough insulator alongside, while the bottom lead of the choke is connected to the other lug. The 500-pf. 20-kv. blocking capacitor, C_3, is mounted on the top of the short insulator of RFC_3. This conversion may be seen in the rear-view photo. Position RFC_3 on a line midway between the 813s, and close to the bottom edge of the chassis panel. Scribe points E, F and G (corresponding to the choke mounting holes) and drill with a No. 25 drill. The plate tuning capacitor C_1 mounts at I and the loading capacitor C_2 at J. Cut the shafts of both capacitors so that they extend through the *chassis* panel ½ inch.

The feedthrough insulator at the top of the chassis, between the 813s (visible in the rear-view photo), was included in the original 10-

[1] SeeZak products are available from Radio Shack Corp., 730 Commonwealth Ave., Boston 17, Mass., Syracuse Radio Supply, Syracuse, New York, and California Electronic Supply, Los Angeles, Calif., among others.

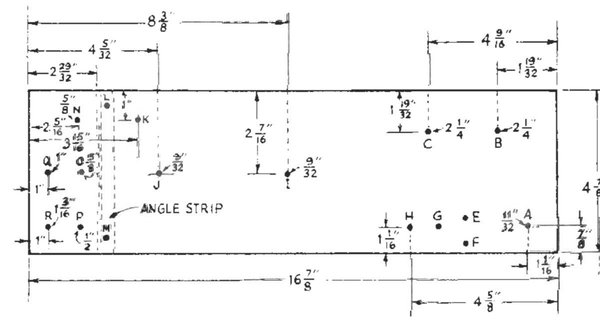
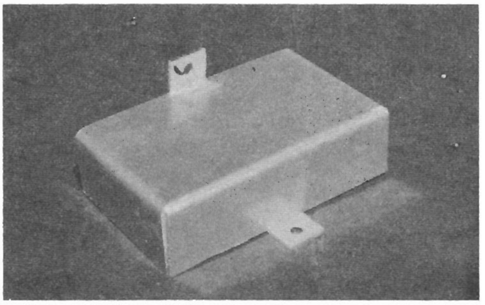


FIG. 2—Sketch showing dimensions and layout of the chassis panel. Lettered points are identified in the text.



The feedback shield described in the text and Fig. 3. The rear side of the shield is open.

meter amplifier to bring out a lead from a neutralizing coil on RFC_2. After completing the amplifier, it was found that neutralization was not required, so the insulator was not put to use.

Tap two diagonally-opposite holes in the SO-239 coax chassis connectors for ¼-inch 6-32 screws, and submount them at N (output) and O (input). The Millen high-voltage connector J_6 is mounted at P, with the male a.c. input connector J_4 at R, and the flush bias-control receptacle J_3 at Q.

Remove the other shell of the filament transformer, T_1. Cut two pieces of ½ × ½-inch aluminum angle to a length of 4 inches, and drill holes to match the two holes that go through the bottom edge of the core. Next, drill a No. 25 hole ⅜ inch from each end of both pieces for mounting. Fasten these mounting strips to the transformer core, using the original bolts.

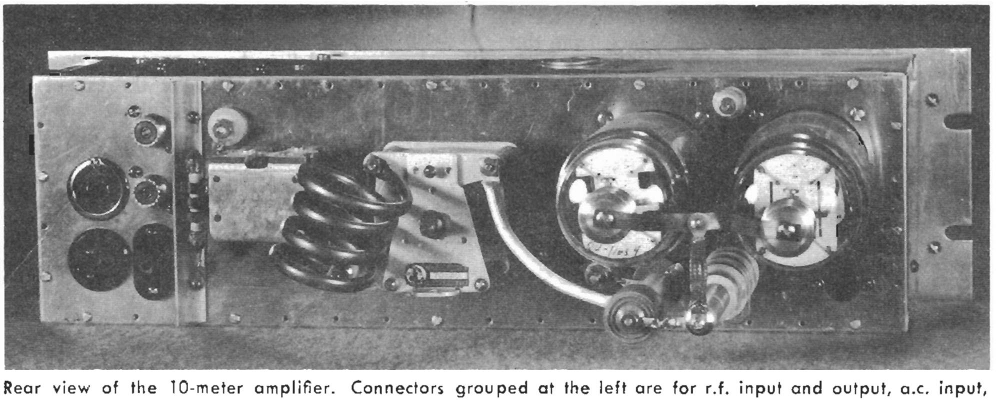
Place flexible couplings on the shafts of C_1 and C_2, mount the chassis panel on the rails with at least two sheet-metal screws (furnished with the rails) on each side, and one at each end. Before tightening the screws, use a mechanic's or carpenter's square to check the corners. Place extension shafts in the flexible couplings of C_1 and C_2 and then place T_1 in position against the front lip of the chassis and between the extension shafts. Check the clearance carefully, then scribe and drill the mounting holes for T_1. Place a 1-inch screened vent plug above and to the left of T_1, as shown in the photographs. Drill four or five ¼-inch vent holes in the bottom side of the chassis, near the front lip, between RFC_2 and T_1, and five or six directly above the pilot lamp.

A slight amount of feedback was encountered in the 15- and 20-meter amplifiers. This was eliminated by placing a small shield over the output coax connector and the feedthrough insulator connected to C_2. The shield is cut from sheet aluminum as shown in Fig. 3, and a photo shows the finished product after bending. Notice the ½ × 1/16-inch notches. These are made to clear the lip of the chassis rail. Use a ¼-inch 6-32 binder-head machine screw through the rail and Tab A, and a ¼-inch No. 6 sheet-metal screw through Tab B into the bottom of the chassis panel. One end of this shield is visible just to the right of the bias transformer in the interior view of the amplifier.

Support the front panel, face down, an inch or two off the work bench. Insert the extension shafts of C_1 and C_2 through the front panel holes and carefully center the chassis on the panel as before. Scribe on the panel an easily-seen mark all the way round the chassis. Remove the chassis panel from the rails and carefully reposition the rails inside the scribed mark on the back of the front panel. Holding the rails in position, use a long scriber or pencil to transfer to the panel the two outside holes on the lips of each end piece. Similarly, transfer the second hole from each end on the long side rails, also the eighth hole from the left-hand end of the bottom rail and the ninth hole on the top rail. Prick punch and drill clearance holes for the ½-inch No. 6 sheet-metal screws used to hold the front panel to the chassis. After checking the alignment of these holes, set the front panel aside.

Wiring

Mount a three-terminal ungrounded tie-point strip midway between T_1 and R_1 and one inch



Rear view of the 10-meter amplifier. Connectors grouped at the left are for r.f. input and output, a.c. input, stand-by bias control and high-voltage input. The small connector below and to the right of the 813s is for blower-motor power.

back from the front lip of the chassis. The primary leads from T_1 and T_2, as well as the leads from J_5, will be attached to the center and left-hand terminal. One of the 115-volt a.c. leads from J_4 is also attached to the left-hand terminal while the other a.c. lead goes to the right-hand terminal of the tie-point strip. When the front panel is mounted on the chassis, flexible leads will be run from the center and right-hand terminals to the power switch.

The location of most of the remaining components can be determined from the interior-view photo.

Much of the wiring can be done before mounting the P517 panel permanently on the rails. Use No. 12 wire for the filament circuit. Insulated hookup wire may be used for the bias-supply connections. Attach leads to J_5 that will reach the tie-point strip near the filament transformer. Use bent solder lugs under the heads of mounting screws of C_1 and C_2 to hold the wires in place and keep them from contact with high-voltage of r.f. wiring. Attach 5-inch leads of flexible wire to J_3 and J_4. High-voltage supply leads should be made with high-tension cable, or with rigid wire well spaced from the chassis and other metal. Attach the chassis panel to the rails with at least 12 sheet-metal screws. Complete the wiring and set the chassis aside.

The Front Panel

The chrome handles at each end of the panel are Bud No. H9168. Mount them ⅞ inch from each end and equidistant from top and bottom. You will find these handles to be the perfect answer for lifting the amplifier in and out of its rack mounting. They will also support the full weight of the amplifier, when you have it face down on your workbench for service, thus protecting the controls.

With the panel face up, locate three holes on a vertical line 2½ inches from the right end as follows: the pilot-lamp hole is 1½ inches down from the top, the hole for the filament switch is 2¾ inches from the top, and a 5/16-inch hole for the shaft of R_1 is 1½ inches up from the bottom of the panel. Drill a No. 25 hole 2 inches to the left of the pilot light and 1½ inches down from the top of the panel. Mount a one-terminal ungrounded tie point on the rear of the panel. Mount the meter with

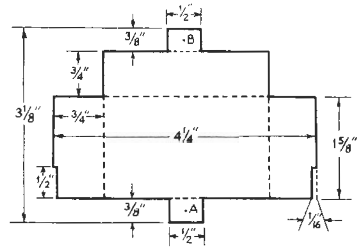


FIG. 3—Sketch showing dimensions of the feedback shield. Bends are made along the dotted lines. See detail photo.

its center 3¼ inches from the left-hand end of the panel and 2¼ inches from the top. The s.p.d.t. slide switch, S_2, is centered directly below the meter. Place a solder lug on the left-hand mounting screw of S_2.

The bracket for R_1 is made from a piece of ⅛ × 1 × 2½-inch aluminum or brass. Bend a 1-inch leg for attaching to the chassis and, after drilling two No. 25 holes, mount with the center line of the bracket in line with the 5/16-inch hole in the panel. Leave a ⅜-inch space between the bracket and the panel. Transfer the panel hole to the bracket, drill a 7/16-inch hole and elongate it with a round file to simplify lining up the shaft of R_1 in the panel hole.

After placing the 0.01-μf. capacitor across the meter terminals, wire R_3 from the positive terminal of M_1 to the ground lug on S_2, and ground the terminals of S_2 closest to the lug. Wire R_4 from the negative post of M_1 to the center contact of S_2. Connect R_2 from the positive post of M_1 to the other terminal of S_2, and run a piece of No. 18 solid insulated hook-up wire from this switch terminal to the tie point near the pilot light.

The 6.3-volt winding on T_2 can be used for the pilot light. Pass the center-tap lead from T_1 over the top of the extension shaft of C_1 before connecting it to the positive post of M_1. This will prevent it from coming in contact with the high-voltage lead-or the plate choke. Leave enough slack in the leads to the pilot light, power switch and bias supply, so that the front panel can be easily lifted on or off the shafts of C_1, C_2 and R_1.

After soldering these leads, position the front panel and insert the ten No. 6 ½-inch self-

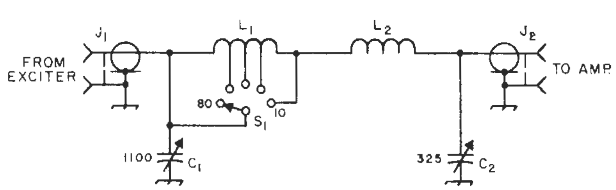


FIG. 4—Pi network for coupling fixed-impedance exciters to the grounded-grid amplifiers. Capacitances are in pf.

C_1—Miniature triple-section variable, 365 pf. per section, sections in parallel.

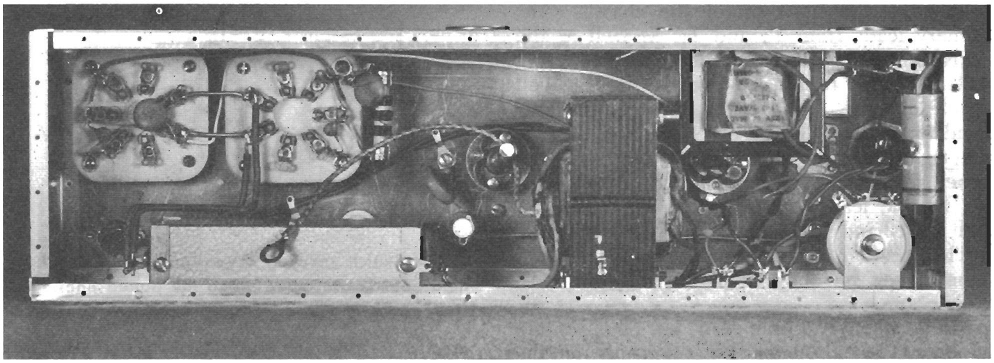
C_2—Miniature receiving-type variable (Hammarlund MC-325-M).

J_1, J_2—Chassis-mounting coaxial receptacle (SO-239).

L_1—17 turns No. 16, 1¼-inch diam., 2 inches long, tapped at 10, 4 and 2 turns from 10-meter end.

L_2—4 turns No. 12, 1-inch diam., 1 inch long.

S_1—Single-pole 5-position ceramic rotary switch.



Interior view of the chassis. The bifilar filament choke is below the 813 sockets. The bias-supply transformer is to the right of the filament transformer, suspended from the top of the chassis. The bias-control potentiometer is in the lower right-hand corner.

tapping metal screws. Position a piece of ¼-inch tubing from the bottom of the 500-pf. blocking capacitor, around the nearest 813 to a stator terminal of C_1. The mounting of L_1 will depend upon the size of the coil which, of course, will vary with the frequency for which the amplifier is being built. The rear-view photo shows the 10-meter amplifier with one end of L_1 attached to C_1 and the other end supported by a stand off insulator. Bud heat-dissipating plate caps are used on the 813s. Copper strap, ⅜ inch wide, is used to connect the tube caps to RFC_3.

Shielding

The shielding enclosure is made of sections of perforated aluminum sheet supported on a framework of ½×½-inch aluminum angle stock. The front edges of the shield overlap the chassis on top, bottom and the left-hand side. The right-hand end of the enclosure is fastened to the angle piece attached to the chassis panel.

Use ¼-inch No. 6 sheet-metal screws for assembly and space them approximately 2 inches all around. The ventilating fan is obtainable from Allied Radio (Cat. No. 72P715.) It is mounted against the inside of the rear wall of the shielding enclosure with the axis of the fan exactly opposite the plate caps of the 813s. Before attaching the top of the enclosure, run the a.c. leads from the fan motor along the bottom to the 2-prong Jones socket, J_5.

Ajustment

Check out the bias voltage and filament circuit before applying high voltage. A variable h.v. power supply is definitely recommended. If not available, arrange to insert a 100-watt lamp in series with the primary of the plate transformer while testing. A power supply delivering from 1800 to 2250 volts d.c. at 400 to 500 ma. is ideal. Before applying high voltage, connect a dummy load to J_2. With a plate voltage of 2000 volts, S_2 in the PLATE position, and the terminals of J_3 shorted, adjust R_1 for 40 ma. of plate current. With carrier injected in the s.s.b. exciter and S_2 in the GRID position, adjust the exciter loading for a full-scale reading on M_1.

Turn S_2 to PLATE, C_2 to maximum, and adjust C_1 for minimum plate current. With reduced plate voltage, decrease the capacitance of C_2 for 200 ma. of plate current, maintaining resonance with C_1. With plate voltage increased to 2000, adjust C_1 and C_2 for approximately 400 ma. Grid current should be 100 ma.

With the exciter adjusted for normal s.s.b. r.f. output, the linear amplifier, with voice, should swing to approximately 150 ma. of plate current. A steady whistle will increase the plate current to 400 ma. The output should be checked for linearity with an oscilloscope during initial adjustment and at regular intervals thereafter.

For c.w. operation, use the same procedure as for s.s.b. operation, adjusting R_1 for approximately zero plate current and the exciter for 100 ma. of grid current without plate voltage on the amplifier. Load the amplifier to 175 ma. with reduced plate voltage and then to 350 ma. with a full plate voltage of 2000.

Exciter Matching

Most exciters presently in use have enough range in output impedance to provide a match to the cathode circuit of the 813s. This impedance runs from a nominal 140 to slightly more than 200 ohms, depending upon the frequency. In the event that your exciter output impedance is fixed at 50 or 70 ohms, it may not be possible to obtain sufficient drive for the 813s. In such a case, a pi-network such as shown in Fig. 4 may be used for matching. In this particular instance, the relative values of C_1 and C_2 near the correct-adjustment condition are such that the output capacitor, C_2, has a greater effect on tuning than C_1. Therefore, the input capacitor, C_1, rather than the output capacitor, C_2, is used as the "loading" control.

» Single-tube amplifier runs 1000 watts input on s.s.b., c.w. or a.m. as a linear amplifier with no grid current. A high-power tube designed specifically for AB_1 operation makes it possible.

Compact AB_1 Kilowatt

RAYMOND F. RINAUDO, W6KEV

Because it is the almost universal practice to generate an s.s.b. signal at a low level and then amplify it to the required output with one or more linear amplifiers operating Class A, AB_1, AB_2 or B, the linearity of the amplifying stages is all important. The stages following the best s.s.b. generator can turn a clean signal into one which is distorted and unnecessarily broad. Thus the need for truly linear amplifiers.

While the individual designer has his choice as to the class of operation in which the amplifier will run, Class AB_1 has several desirable characteristics. Because the control grid is never driven positive the very serious problem of adequate driver regulation never has to be faced, as it does if the mode of operation is AB_2 or Class B. In addition, no driving power is required for the tube: only the grid circuit losses must be supplied.

It should be pointed out that most tetrodes and many triodes appear as a resistance of 200 to 500 ohms from grid to cathode when the grid is positive. During the part of the r.f. cycle when the grid is negative the resistance is infinite. A driving source that can supply either an infinite resistance or a load of a few hundred ohms, without distortion of the voltage wave form in either case, would have to have very low internal resistance. A working approximation is usually achieved by making the tuned grid circuits of r.f. amplifiers extremely high C. In audio amplifiers it is obtained by using low-plate-resistance driver tubes plus a step-down transformer.

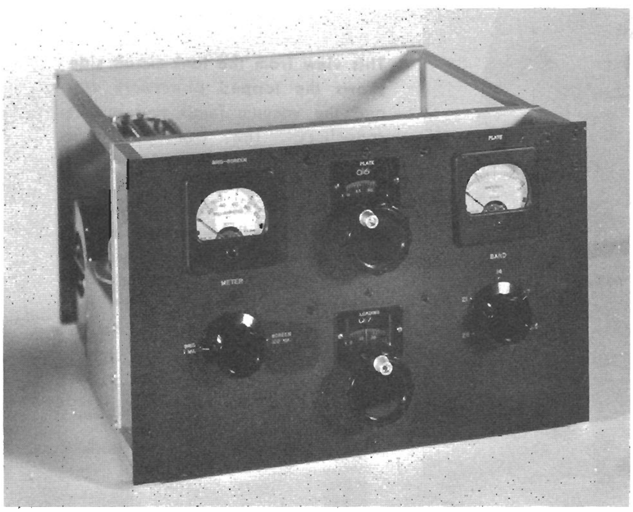
Class AB_1 amplifiers compare very favorably in efficiency with AB_2 and Class B. In fact, over-all amplifier efficiencies, which take into account the losses in the tube and the circuit, are usually of the order of 55 to 65 per cent. It is only when compared with Class C operation that AB_1 represents a significant lowering of efficiency.

It is for this reason that some of the older tube types do not look particularly attractive in s.s.b. service. In the past almost all transmitting types were designed for optimum service in Class C amplifiers. This optimum provided a balance between plate current and plate dissipation; the higher efficiency realized required less plate dissipation capability for a given input power. In contrast, a tube designed especially for AB_1 application would be expected, for a given output, to have a higher plate-dissipation rating than we have become accustomed to.

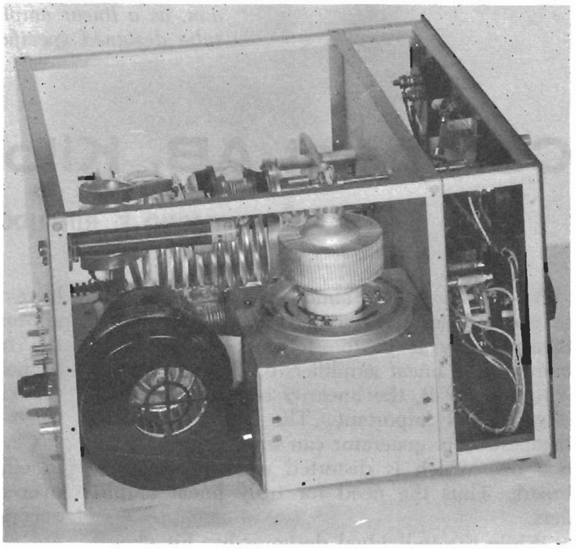
The 4CX1000A Tetrode

A tube designed to have exceptionally good linearity in Class AB_1 r.f. amplifiers is the

From *QST*, November, 1957, and subsequently modified.



There is a pleasing symmetry to the control layout on the $10 \times 15\frac{1}{2}$-inch panel. The grid circuit is untuned, so the only r.f. controls are the band switch, plate tuning, and loading. Separate meters are provided for plate and screen currents, with the screen meter also used as a grid-current monitor. The amplifier, 15 inches deep, contains filament transformer and cooling fan in addition to the r.f. circuits.

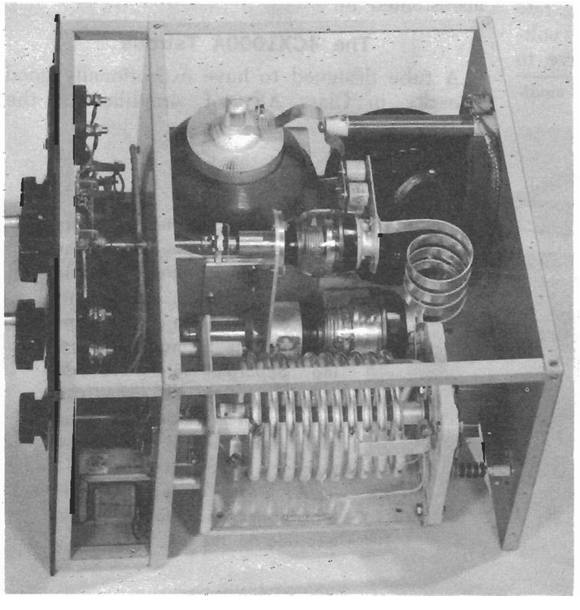


Vertical chassis construction is used, as this view from the tube side shows. The air-system socket is mounted on the 6 by 6-inch top of an aluminum enclosure 4 inches high, with the chassis pan forming one wall. When the bottom plate is in place this forms a pressurized area for forcing air from the blower through the socket. The socket chimney has been removed in this photograph to show the 4CX1000A tube.

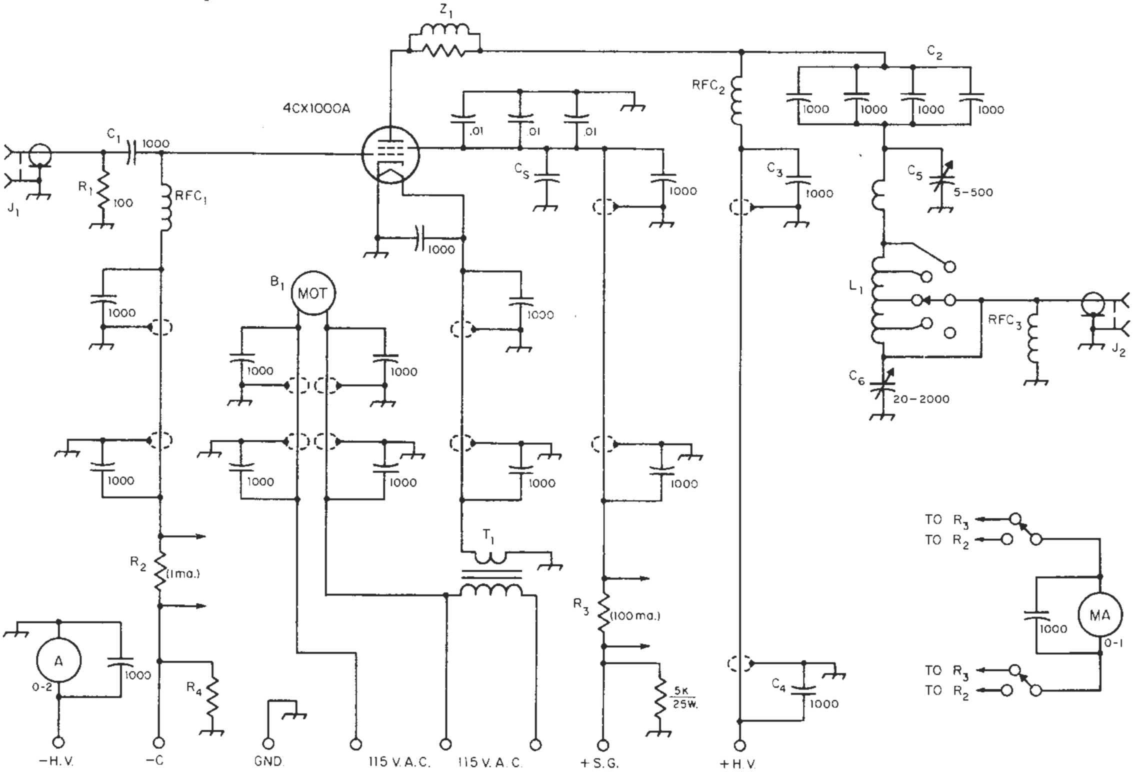
Eimac 4CX1000A. It is a power tetrode of all ceramic and metal construction having an external anode capable of dissipating 1000 watts with 35 cubic feet of air per minute blown through the cooling fins. The filament requires 6.0 volts at 12.5 amperes to heat the oxide coated cathode, although at amateur power levels the tube will have a longer life if the filament voltage is held to about 5.7.

The power output will vary with the type of service for which the tube is used. For single sideband suppressed-carrier single tone, the output is 1680 watts for 2700 watts input at the maximum plate voltage of 3000. If the driving signal is an amplitude-modulated carrier, either single or double sideband, a carrier output of about 300 watts can be expected from a kilowatt input. If a c.w. signal is being amplified then the output power would be approximately 600 watts. Since for a.m. phone or for c.w. the carrier or key-down conditions apply in measuring power input, it is the legal power-input limit that largely determines the output power.

The connection to each element is made by means of three metal tabs or ears which protrude through the side of the envelope at 120-degree intervals around the circumference. The screen tabs are nearest the anode; the control grid, cathode plus one side of the



This view from the tank-circuit side shows the tapped pi-network coil and the vacuum input and output capacitors. The capacitors are mounted on an aluminum bracket fastened to the tube compartment. The plate blocking capacitor—four units in parallel—mounts on a plate fastened to the hot terminal of the input tuning capacitor. The plate chake is mounted on the rear wall. The chimney is around the tube in this photograph.



Note: Power lead for blower motor is brought out separately for resistance control of speed during stand-by.

FIG. 1—Circuit diagram of the amplifier. Unless otherwise indicated, capacitances are in pf., resistances are in ohms. Capacitors not listed below are 600-volt disk ceramic.

B_1—Blower motor.

C_1—1000-pf. mica.

C_2—Four 1000-pf. ceramic in parallel, 5000-volt rating (Centralab 858).

C_3, C_4—1000-pf. ceramic, 5000 volts (Centralab 858).

C_5—5-500-pf. vacuum variable (Jennings UCSL 500 3KV).

C_6—20-2000-pf. vacuum variable (Jennings UCSL 2000 2KV).

C_8—Built-in socket bypass, 1450 pf.

J_1, J_2—Coax receptacles, chassis mounting.

L_1—Pi-network tank assembly (B & W 852).

R_1—100 ohms, noninductive, to dissipate at least 15 watts (see text). Can be assembled from 2-watt composition resistors in parallel or series-parallel.

R_2—Approx. 1000 ohms (should be 20 or more times meter resistance).

R_3—Adjusted to shunt 1-ma. meter for 100 ma. full scale; approx. 0.5 ohm in average case.

R_4—Adjust for 20-ma. bleed.

RFC_1, RFC_3—2.5-mh. r.f. choke.

RFC_2—Solenoid choke, 500 ma. (B & W 800).

T_1—6 V., 13 a., (Stancor P-6463).

Z_1—½-inch copper strip formed into U, 2 inches long, 1 inch wide, paralled by three 100-ohm, 2-watt composition resistors.

heater, and heater follow in order to the bottom. The Eimac SK-800 socket for the tube has a built-in screen bypass capacitor. The height of the tube is just under 4¾ inches, and the diameter approximately 3⅝ inches.

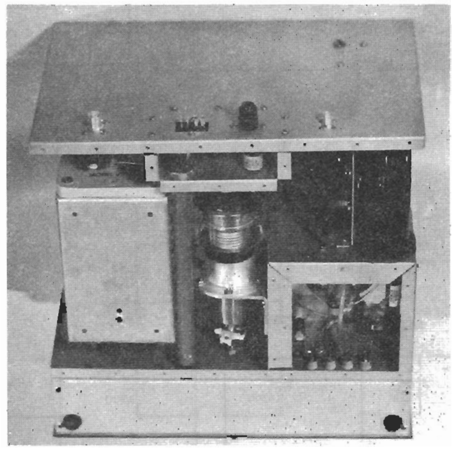
The control grid is rated at zero dissipation. In designing the tube for AB_1 operation the location and number of grid wires was not hampered by compromises such as would be necessary if the grid were called upon to handle power. Consequently, a large number of fine wires were closely spaced to the cathode to give an unusually sharp-cutoff grid voltage-plate current characteristic. Thus linearity is maintained near cutoff.

While the tube is capable of powers in excess of the legal amateur limit it is quite legal to have peak-envelope inputs in amateur service well in excess of a kilowatt if the average power does not exceed that figure. In such cases the tube cathode is asked to supply quite high currents and must be capable of such operation if linearity is to be maintained.

A Compact Amplifier

The amplifier shown in the photo is contained in a package measuring 10 inches high by 15½ inches wide by 15 inches deep. The r.f.-tight enclosure is 12 inches front to back, with a 3-inch space between the front panel and shielded box. Not shown in the photographs are the perforated aluminum U-shaped

This bottom view gives a glimpse inside the grid compartment, upper left. R.f. input is through the coax connector on the rear wall and a short length of coax into the shield box. Power leads come in through the socket and high-voltage connector at the center, where they are enclosed by a small aluminum shield mounted on the rear wall. All except the high-voltage lead and leads to the blower motor go through the conduit (running alongside the bottom of the tank coil assembly) to the front of the unit. The high-voltage lead goes through shielded wire to the plate choke. Those to the blower are also shielded.



cover, which forms the top and two sides, or the solid sheet of aluminum that completes the shielding on the bottom. The space between the front panel and the main shielded enclosure is out of the r.f. field and so was not made r.f. tight. In spite of the compactness there is no crowding of parts.

The plate circuit is a conventional pi network. The blocking capacitor is made up of four 1000-pf. ceramic units in parallel, resulting in a capacitance about double that normally used. This was done because of the low impedances involved in the low-voltage high-current application. The Jennings variable vacuum capacitors contributed immeasurably to the compact construction, and here again the 500-pf. input capacitor is higher in capacitance than usually expected. The high *C* is necessary at 3.5 Mc. to maintain the operating *Q* of the circuit. The low inductance of these capacitors helps considerably in the elimination of parasitic oscillations. Note that the pi capacitors are grounded to the box the 4CX1000A is mounted on, and are insulated from the front panel by shaft couplings. This is done to keep ground currents where they belong, and to prevent ground loops.

As was mentioned previously, AB_1 operation precludes driving the grid positive and so the voltage stabilizing influence of a high-*C* circuit is not needed. Instead, a 100-ohm resistor is used in the r.f. circuit between grid and cathode. This also represents very heavy loading of the grid and makes neutralization unnecessary. When using the grid bias indicated for typical operating conditions, −55 volts, the r.f. power lost in the resistor is 15.1 watts and is the total required driving power. For those who would like to terminate the transmission line from the driver in a 50-ohm resistor, the driving power would be 30.3 watts. The photograph of the under side of the unit shows two noninductive wire-wound resistors which make up the 100-ohm load; these have since been replaced by a bank of carbon resistors.

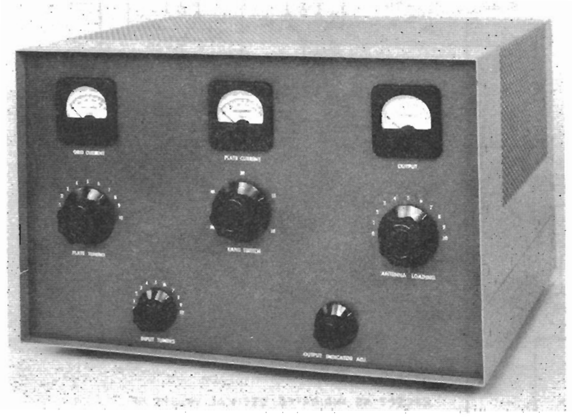
If the driving power requirement of this untuned arrangement can not be tolerated, a tuned circuit can be added. In such case the only power needed is that required to supply the tuned-circuit loss. Neutralization, of course, would become necessary, and the usual bridge circuit is the logical choice.

The front panel shows that two meters are used, though one is dual purpose. The plate current meter has a full-scale reading of two amperes; however, the maximum plate current that can be drawn is 1 ampere using the single tone test (into a dummy load). The dual-purpose meter is one milliampere full scale and is used in combination with a switch and shunts to read grid current at 1 ma. full scale, or screen current at 100 ma. full scale. The one-milliampere scale is used to monitor s.s.b. AB_1 operation so as never to drive into grid current. From ½ to 1 watt of grid dissipation can be handled, but this leaves no margin of safety. The rating of zero dissipation still stands.

Although AB_1 operation minimizes the generation of harmonics, standard TVI-proofing techniques are used throughout. All leads leaving the shield enclosure not normally carrying r.f. are shielded and bypassed at both ends. Leads to the front panel from the compartment that shields the power-input socket are carried through the r.f. enclosure in a length of ½-inch conduit.

In operation the amplifier has proved to be quite stable. The 100-ohm resistor between grid and ground undoubtedly contributes a great deal to this stability. However, stability can be improved further by placing a copper wire screen over the top of the tube socket, soldering it to the flange atop the socket, with a suitable cutout for the tube. This will permit air to pass through but will provide r.f. isolation.

» *This amplifier takes a kilowatt d.c. input in stride. On the customary assumption that the peak-envelope to average power ratio is 2 to 1 for voice signals, a d.c. kilowatt calls for a p.e.p. input of about 2000 watts.*



One kilowatt d.c., two kilowatts p.e.p., is the input this receiver-size amplifier will handle. The "no-screws" front panel is the result of laminated construction. Wrap-around cabinet is homemade.

A Two-Kilowatt P.E.P. Amplifier Using the 3-1000Z

ROBERT I. SUTHERLAND, W6UOV AND HAROLD C. BARGER, W6GQK

The 3–1000Z grounded-grid amplifier shown in the photographs is designed for 2 kilowatts p.e.p. input for single-sideband service on amateur bands between 3.5 and 29.7 Mc. The plate circuit is a pi-L tank presenting a tube load impedance of 2650 ohms, at a plate potential of 3000 volts, and capable of matching an antenna load impedance of 50–75 ohms. The pi part of the network has a Q of 20 and the L section has a Q of 10. With 2500 volts and a plate current of 800 milliamperes, the plate load impedance is 1700 ohms. The pi part of the tank circuit will then have a Q of about 12. When the amplifier is tuned for the maximum legal conditions for 1 kilowatt c.w. the network Q is somewhat higher, but this presents no problems.

A plate potential of about 2500 volts will provide a signal with somewhat less intermodulation distortion than is generated at 3000 volts, but any voltage from 2500 to 3000 will provide efficient operation.

The pi-L[1] design was chosen for use in this amplifier because of two inherent advantages. First, the pi-L network provides an improvement in harmonic suppression of about 15 decibels over the simple pi configuration. Second, the loading capacitance can be chosen to allow the use of inexpensive variable air capacitors, without the usual selector switch for adding fixed capacitors in parallel for operation at the lower frequencies. In this amplifier it was decided that loading capacitor C_5 (see Fig. 1) would be a single 2000-volt 1000-pf. variable unit, or two 500-pf. units in parallel. In the design, the r.f. voltage across capacitor C_5 was arbitrarily limited to 1000 volts. The calculations indicated that a loading capacitance of 891 pf. would be required for 3.5 megacycles. Two E. F. Johnson 500-pf., 2000-volt air capacitors were then chosen to be driven in parallel by a set of 2⅝-inch diameter surplus gears.

A simple semiconductor voltmeter is incorporated in the output portion of the tank circuit to indicate relative power output. The sensitivity is adjusted by means of a rheostat.

The only disadvantage of the pi-L network (if it can be considered a disadvantage) is the requirement for two sets of coils and two bandchange switch decks for the plate tank. In addition, the operator may be surprised at the fast change in loading when the loading capacitor is placed across a relatively high-impedance part of the circuit. This effect could be eliminated by using a reduction gear arrangement between the loading capacitor and the dial.

From December, 1962, *QST*.

[1] Rinaudo, "The Pi-L Plate Circuit in Kilowatt Amplifiers," *QST*, July 1962.

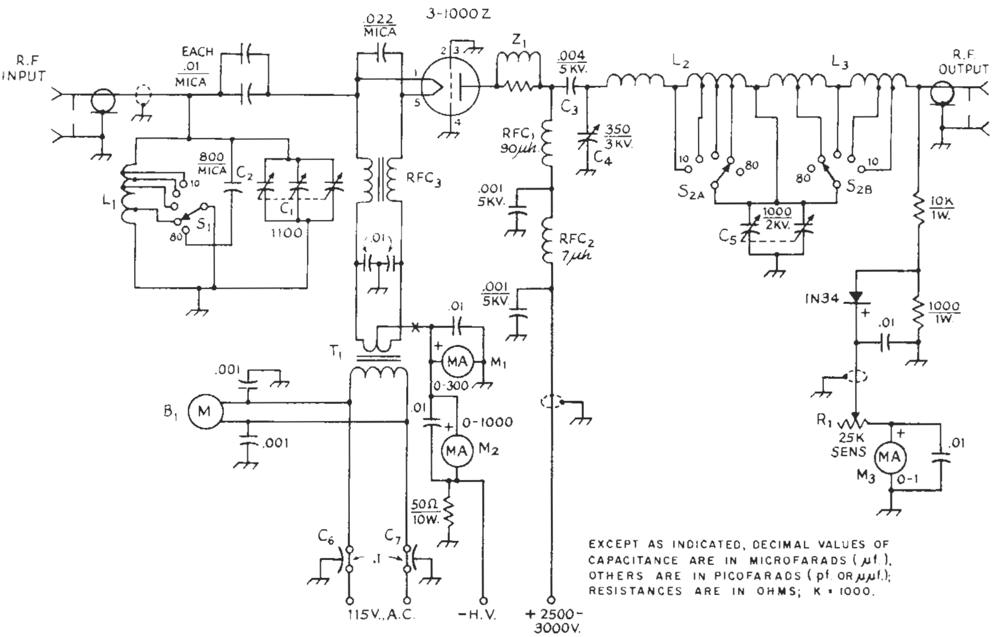


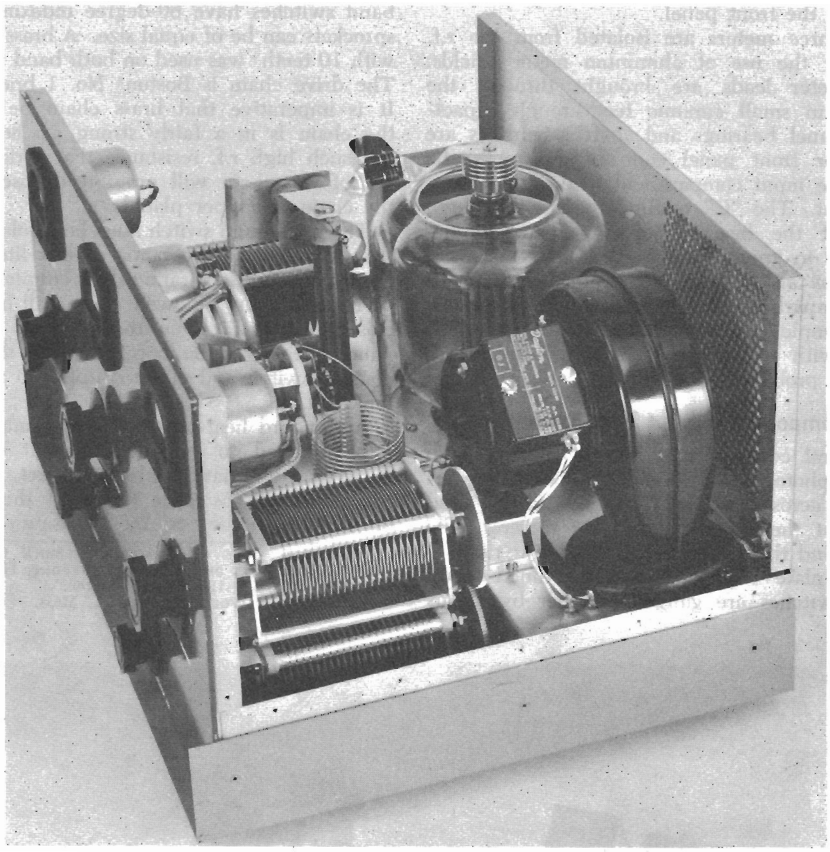
FIG. 1—The amplifier circuit diagram. Capacitors are 600-volt disk ceramic except as indicated. Resistances are in ohms. S_1 and S_2 are ganged as described in text.

B_1—Blower, 20 cu. ft./min. at socket (Dayton IC-180).
C_1—3-section variable, 365 pf. per section, broadcast-replacement type (Miller 2113).
C_2—800 pf. mica (500 pf. and 300 pf. in parallel); (C-D type 4 or 9, 1200 volts d.c. test).
C_3—0.004-µf. ceramic (four 0.001-µf. in parallel); (Centralab 858-S).
C_4—350-pf. 3000-volt variable (Johnson 350E30, catalog No. 154-10).
C_5—0.001-µf. variable, 2000 volts (two Johnson 500E20, catalog No. 154-3, in parallel).
C_6, C_7—0.1-µf. feedthrough (Sprague Hypass 80P3).
L_1—9 turns No. 10 enam., diameter 1 inch, length 1½ inches. Tapped as follows, measured from "hot" end: 7 Mc.: 4½ turns; 14 Mc.: 2½ turns; 21 Mc. 1¾ turns; 28 Mc.: ½ turn plus 2-inch lead.
L_2, L_3—See Fig. 2.
M_1—0-300 d.c. milliammeter.
M_2—0-1000 d.c. milliammeter.
M_3—0-1 d.c. milliammeter.
R_1—25,000-ohm control, linear taper.
RFC$_1$—90 µh., 500 ma. (B & W 800).
RFC$_2$—7-µh., 1000 ma. (Ohmite Z-50).
RFC$_3$—Four windings, 25 turns each, No. 12 enam., on ½-inch diameter ferrite rod (Lafayette MS-333), winding length 4 inches. Bifilar wound in two layers, with two windings paralleled in each leg. (B & W type FC-30 may be substituted.)
S_1—1-section, 1-pole, 5-position ceramic rotary (Centralab 2501).
S_2—2-section, 2-pole, 5 position ceramic rotary (Radio Switch Corp. Model 86; two type H wafers, 60-degree detent).
T_1—Filament transformer, 7.5 volts, 21 amp. (Stancor P-6457).
Z_1—½-inch copper strip formed into U, 2 inches long, 1 inch wide, paralleled by three 100-ohm, 2-watt composition resistors.

The tuned input circuit is a conventional high-*C* tank, with a *Q* of about 2, shunted across the filament circuit. The filament is above r.f. ground by virtue of a homemade bifilar choke wound on a ferrite core. The tuned cathode[2] circuit aids in matching the exciter to the final and reduces the magnitude of the intermodulation distortion products. As the input impedance of the 3–1000Z is close to 50 ohms, it is not necessary to provide driving-point taps on the input coil for each band. The input tuned circuit is switched simultaneously with the plate tank by ganging the plate and input band switches with a sprocket and chain-drive scheme. The drive power requirement is compatible with the power output from a modern exciter using a pair of 6146s or similar type tubes.

Meters are provided for measuring plate and grid currents. These currents are independently monitored by inserting the meters in the grid and plate return leads to the filament center tap. This method allows the three control-grid terminals of the socket to be directly strapped to the chassis. There are slots in the SK–510 socket especially provided to allow low induct-

[2] Orr, Rinaudo, Sutherland, "The Grounded-Grid Linear Amplifier," page 126.



Looking in from the right-hand side. Both loading capacitors are visible in this view. The small aluminum box immediately below the blower motor houses the r.f. voltmeter circuit. Note perforations in the cabinet behind the blower to admit cooling air.

ance ground terminations to be made to each of the three grid terminals. This strapping technique is far superior to the use of grid by-pass capacitors, and the low inductance of the straps contributes significantly to the stability of the amplifier. Each meter responds to one current; i.e., no subtraction of the meter readings is necessary. A 50-ohm, 10-watt resistor is placed between the "floating" negative side of the power supply and ground. This is to protect the operator in case the meters should become open-circuited. Without this resistor, an open circuit could cause the negative side of the power supply to rise dangerously above ground potential.

Shielded Enclosure and Chassis

This compact table-top amplifier is packaged in a homemade aluminum cabinet measuring 10½ inches high, 17 inches wide, and 14 inches deep. The small chassis at the back of the cabinet was fabricated after a layout satisfying the electrical requirements was determined. This chassis measures 2½ inches high, 17 inches wide, and 6½ inches deep, and is made of 0.063-inch sheet aluminum. It shields the input circuit and provides the mounting deck for the filament transformer, blower and tube socket.

The front panel is made of two pieces of aluminum sandwiched against one another. The "rear" panel, of 0.063-inch aluminum, is used for mounting all of the parts attached to the front panel as seen in the photographs. Flat-head machine screws in countersunk holes support the various components and provide a smooth surface for the "outer" front panel to seat against. The "outer" front panel is made of ⅛-inch Dural and is painted and labeled to provide a neat looking appearance with no screw heads visible. The cabinet enclosing the amplifier is formed from two pieces of 0.063-inch aluminum sheet. The bottom is formed in the shape of a U from a solid piece of aluminum. The top piece was perforated to aid in cooling the tube and components. An edge of solid aluminum was left around the top piece for esthetic reasons. Both the top and bottom covers extent ½ inch beyond the edge of the front panel.

All three meters are isolated from the r.f. field by the use of aluminum meter shields. The meter leads are brought through the shields in small ceramic feedthrough capacitors. Panel bearings and shaft assemblies are used for front panel controls and on shafts from the input compartment to the plate compartment. The "C" retaining washers on both sides of the bearing serve as good wiping contacts to ground and prevent the shaft from acting as an "antenna", coupling power from one compartment to the other, as can happen with simple panel bushings. A control shaft "hot" with r.f. can lead to circuit instability or TVI problems.

Component Layout and Assembly

General component placement may be seen in the photographs. The three meters are arranged across the front panel with the grid meter at the far left, the plate meter in the center and the r.f. voltmeter at the far right.

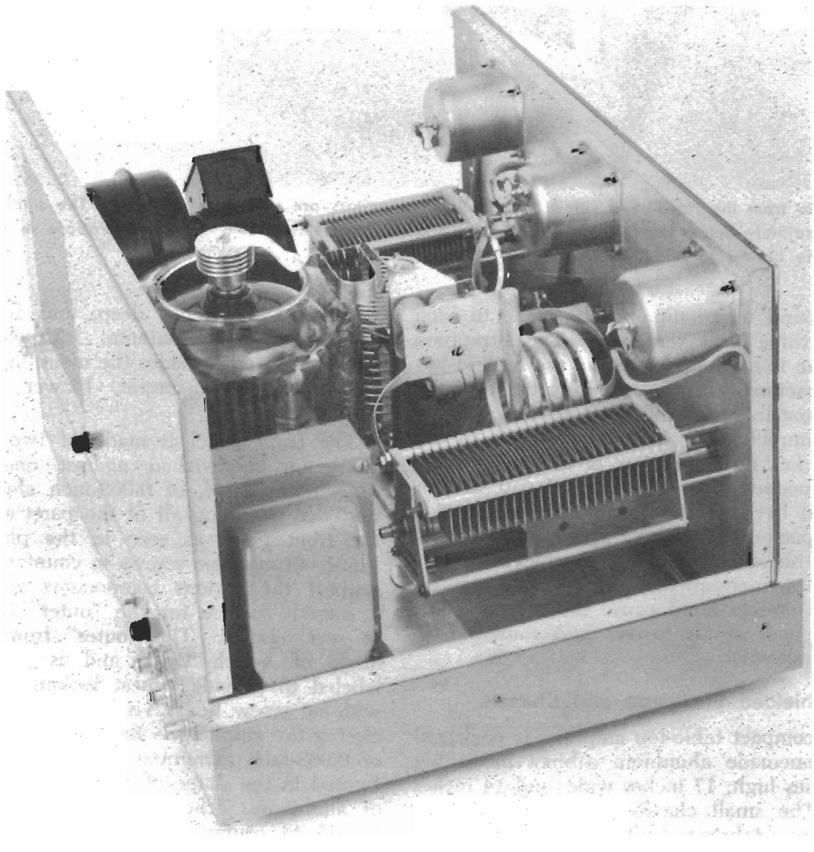
The plate tank and cathode tuned-circuit band switches are ganged together by means of a chain and sprocket drive system. Both band switches have 60-degree indexing so the sprockets can be of equal size. A brass sprocket with 10 teeth[3] was used on both band switches. The drive chain is Boston[4] No. 1 brass chain. It is imperative that brass chain be used as the chain is in a fairly strong r.f. field. Iron has such high r.f. resistance that serious heating of the chain will prohibit its use even if the chain is copper plated.

The plate band switch must be insulated from the front panel by ceramic or bakelite mounting spacers and a ceramic shaft coupler. If this precaution is not followed, there will be serious closed-loop loading on the 80- and 40-meter bands because of the loop formed by the panel, chassis, switch and chain coupled to the plate tank coil.

A standard heat-radiator cap (Eimac HR-8) was modified to provide sufficient clearance between the radiator and the cabinet. The two top radiating fins were removed; this can be done in a lathe or by hacksaw and file.

[3] Perfection Gear Co. (American Stock Gear Division), 152nd Street and Vincennes Avenue, Harvey, Ill.; Sprocket #C-10, 1.125 inch diameter.

[4] Boston Gear Works, Quincy 71, Mass.



View from the left-hand side, showing the filament transformer, plate tank capacitor, and mounting for the plate blocking capacitor.

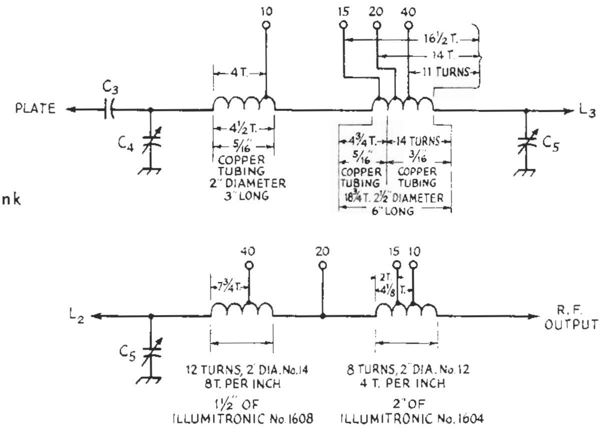
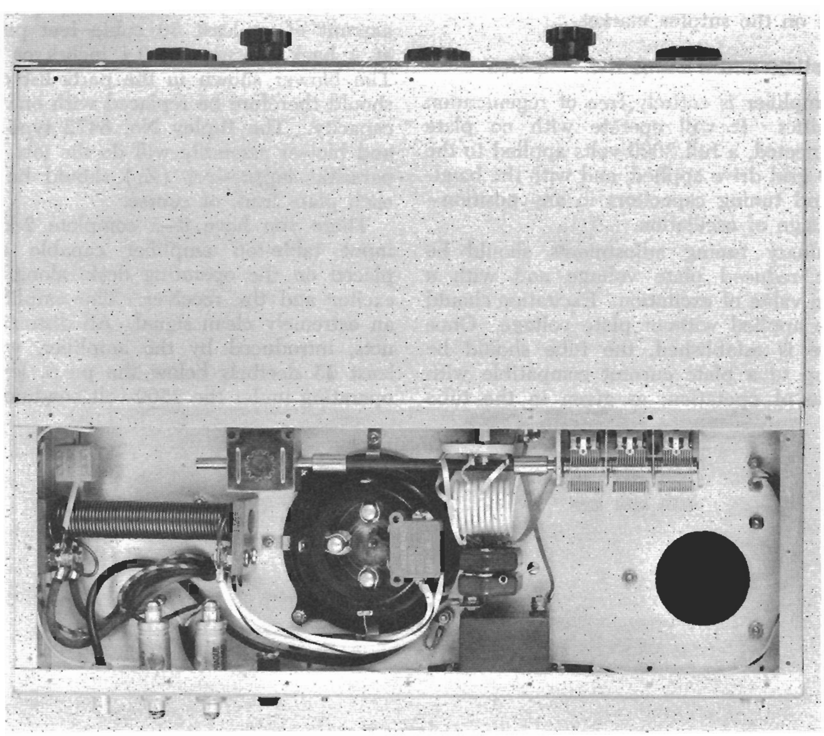


FIG. 2—Winding details of the pi-L tank inductors. L_2 above, L_3 below.

Ordinarily this is not a recommended procedure. However, after the modifications were made, the plate seal temperature was measured by a thermocouple and was found to be well within the manufacturer's ratings. The chimney and modified heat-radiator cap were in place and cooling air was supplied by the fan described in the parts list.

The SK-510 socket was modified to allow a shallower chassis to be used. The SK-510 has a short plastic cylindrical protrusion on the bottom to couple to an air hose for cooling purposes. This cylinder is not used when the chassis is pressurized as in this amplifier. The plastic cylinder can be cut off with a hacksaw and file after the contact pins have been removed. The pins are snapped back into the socket after the modification.

The semiconductor r.f. voltmeter is mounted in a small aluminum box bolted to the chassis just below the blower motor. A length of hook-up wire is attached between the volt-



This is the input compartment as seen from the bottom. The homemade filament choke is at the left, supported on small phenolic blocks. The input tank circuit is to the right of the tube, with the capacitor turned through a right-angle drive. Grid terminals on the tube socket are grounded directly to the chassis with copper strap.

meter and the pi-L output to sample the r.f. voltage. The sensitivity control is mounted on the front panel and is connected to the voltmeter by means of a shielded wire. The indicating milliammeter is attached to the detector unit by means of a shielded wire.

Amplifier Wiring

Shielded wire is used on all low-voltage leads in the plate compartment except the blower power leads and r.f. sampling wire. Small ceramic feedthrough capacitors pass all leads from one compartment to another. Coaxial capacitors are employed as 115-volt a.c. terminals on the rear apron of the chassis.

Silver-plated ½-inch copper strap is used for all taps on the pi section of the plate tank and for the plate leads connecting the tuned circuit to the 3–1000Z plate cap. No. 14 tinned copper wire is used to connect the taps on the L part of the plate tank to the bandswitch.

Large transmitting-type mica capacitors must be used as the coupling capacitors in the cathode tuned circuit, as all the drive power flows through these capacitors. The capacitor shunting the input circuit on 80 meters will carry part of the circulating current in this tuned circuit and therefore must be a good transmitting type. All of the mica capacitors used for the cathode tuned circuit are widely available on the surplus market.

Testing and Tuning the Amplifier

The amplifier is entirely free of regeneration or parasitics. It will operate with no plate load connected, a full 3000 volts applied to the plate, no grid drive applied, and with the bandswitch and tuning capacitors in any position—with no sign of oscillation.

Preliminary tuning adjustments should be made at reduced plate voltage and with a minimum value of excitation. Excitation should never be applied without plate voltage. Once resonance is established, the tube should be loaded up to a plate current compatible with the mode of operation, as given in the tube data sheet. If the grid current is excessive the plate circuit loading is too light. Low grid current indicates that plate loading is too heavy. As a final check, the r.f. output (as observed on the output voltmeter) should increase in direct proportion to the excitation level. With an average voice, plate current should "kick" to about 360 or 400 milliamperes operating at a plate potential of 2500 volts. With 3000 volts on the plate, the voice current will read about 290 to 330 milliamperes. In either case, the grid-current reading will be approximately one-third the plate current reading.

The resting plate dissipation of the 3–1000Z runs between 400 watts and 720 watts, depending upon the plate voltage on the tube. It is possible to cut the quiescent plate current and accompanying plate dissipation to a negligible amount by inserting a 10,000-ohm, 10-watt wire-wound resistor in the filament center-tap return lead at point "X" in Fig. 1. The cathode bias developed by a small plate current flow through the resistor biases the tube to a few ma. of resting current. The resistor can be shorted out for proper amplifier operation by an extra set of relay contacts in the VOX circuit.

Finally, it should be noted that a pair of 3–400Z tubes connected in parallel may be substituted for one 3–1000Z with equivalent results. Additional air is required, in the amount of at least 30 cubic feet per minute at a back pressure of 0.4 inches or mercury. The blower shown in the parts list of Fig. 1 should therefore be replaced with one of larger capacity. The Ripley No. 8472 type 3 motor and blower assembly will do the job. Separate parasitic suppressors (Z_1) should be used in each plate lead, of course.

There you have it—a complete 2-kw. p.e.p. input table-top amplifier capable of being placed on the operating desk, along with the exciter and the receiver. The amplifier emits an extremely clean signal. All distortion products, introduced by the amplifier, will be at least 35 decibels below the p.e.p. level when operating under the 2500-volt condition.

» Getting on v.h.f. s.s.b. is easy if you already have equipment for the lower bands. All it takes is a simple converter such as this one designed by W2UTH. It converts a 14-Mc. s.s.b. signal to the 50-Mc. band.

A Simple Heterodyne Unit for 50-Mc. S.S.B.

HENRY A. BLODGETT, W2UTH

Nearly every v.h.f. sideband enthusiast has his own pet circuit for the transmitting converter, mixer, transverter, or whatever the chosen name for a device used to convert his h.f. sideband signal to a v.h.f. band. Being no exception, the writer set about arriving on 50-Mc. s.s.b. in the easiest possible manner. The mixer unit described herein is the result. Certain criteria exist in the design of a transmitting mixer for v.h.f. use. First, it must be completely stable. Second, the output should be free of spurious frequencies. The writer added a third: it had to be simple and inexpensive.

Assuming that you have a satisfactory s.s.b. signal to start with, stability is mainly a matter of the design of the heterodyning oscillator. Freedom from spurious-signal output is most readily attained by using a crystal oscillator on as high a frequency as practical. Overtone oscillators are notorious for instability, but mostly because amateurs tend to run them at too high a power level. The 36-Mc. crystal and simple triode oscillator shown provide very good stability if the input is kept low and the plate voltage is regulated.

From April, 1964, *QST*.

The old reliable 6J6 is used for the crystal oscillator, V_1. The tube is an excellent oscillator, and connecting the elements in parallel allows operation well below the point at which excessive heating and resultant frequency drift would occur. Drift that might result from heating cycles can be eliminated by allowing the oscillator to run all the time, instead of removing plate voltage during standby periods. The 36-Mc. output may, in fact, be used for injection in the 50-Mc. receiving converter, if desired.

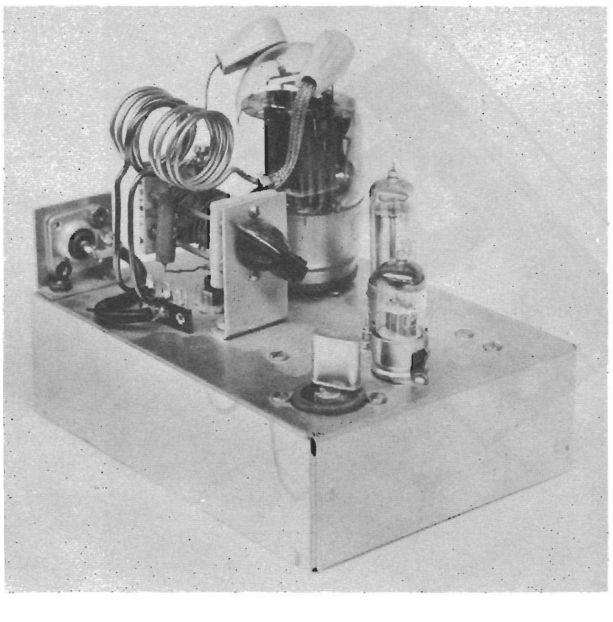
An 815, V_2, is used for the mixer. This dual tetrode may not be familiar to some newer hams, but it can be described as basically two 2E26s in one envelope, with a common screen and a common cathode. Push-pull mixer operation is desirable, as it tends to balance out unwanted frequencies that might otherwise appear in the output. Other dual tetrodes, from the 6360 to the 5894, should work equally well.

The 36-Mc. oscillator signal is fed to the grids in push-pull. Injection of the 14-Mc. sideband signal was tried at both cathode and screen of the 815, the latter giving the better results. Very little drive is required, and no mixer instability of any kind was ever encountered.

Construction and Use

The heterodyne unit is constructed on a 5×7×2-inch aluminum chassis. Layout of parts is not critical. Power can be supplied from any source giving 270 to 400 volts d.c. at 125 ma., and 6.3 volts a.c. at 2 amp.

Upon completion of construction, the 36-Mc. oscillator should be checked first, and the grid circuit of the mixer resonated at 36 Mc. Oscillator output should be sufficient to provide about 18 to 21 volts across the mixer grid resistor. Now apply plate and screen voltage to the 815, and set the screen control so that the plate input is no more than 20 watts, with no s.s.b. drive. A 14-Mc. s.s.b. signal of 5 to 10 watts is then coupled into the screen circuit



This unit converts a 14-Mc. s.s.b. signal to 50-Mc. with an output of approximately 10 watts. The crystal oscillator is nearest the camera. Chassis is standard 5×7×2-inch aluminum.

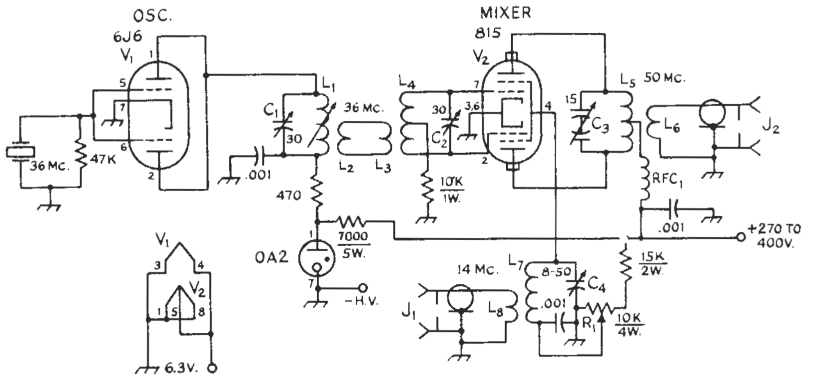


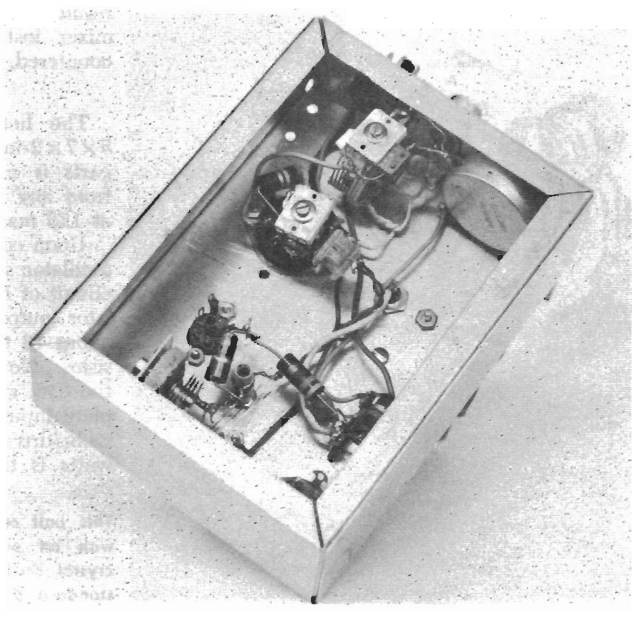
FIG. 1—Schematic diagram and parts information for the 50-Mc. sideband unit. Capacitors not described are 0.001-μf. disk ceramic or mica. Resistors are ½-watt composition unless specified otherwise.

C_1, C_2—30-pf. trimmer, any type.
C_3—15-pf. per section split-stator.
C_4—50-pf. trimmer.
J_1, J_2—Coaxial receptacle.
L_1—8 turns No. 32 enam., close-wound on ¼-inch iron-slug form.
L_2—2 turns insulated hookup wire at cold end of L_1.
L_3—2 turns insulated hookup wire around center of L_4.
L_4—12 turns No. 22 tinned, ½-inch diam., 32 t.p.i., c.t. (B & W No. 3004).
L_5—8 turns No. 12 enam., 1-inch diam., center-tapped, spaced wire diam. Spread out center turns ½ inch for L_6.
L_6—2 turns No. 12 enam., 1-inch diam., at center of L_5.
L_7—12 turns No. 20 tinned, ¾-inch diam., 16 t.p.i. (B & W No. 3011).
L_8—3 turns insulated hookup wire on cold end of L_7.
R_1—10,000-ohm 4-watt control.
RFC_1—7-μh. r.f. choke (Ohmite Z-50).

through J_1. The level of injection is controlled by R_4, but the operator should observe the 20-watt plate-input limit when no drive is applied. With 270 volts on the 815 plates, maximum mixer output occurs with 50 to 70 volts on the screen. Use the least amount of s.s.b. drive that will give maximum mixer output.

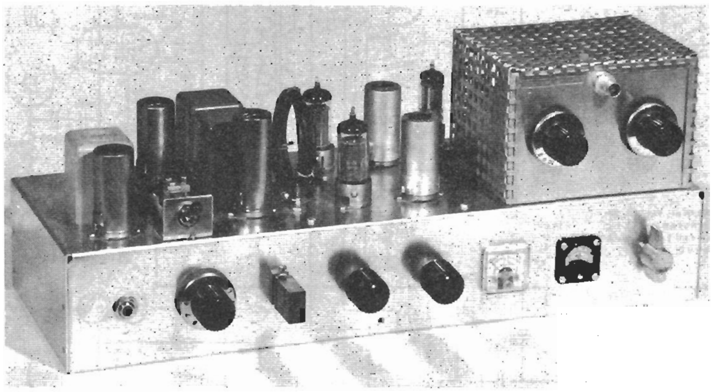
It should be remembered that the 50-Mc. s.s.b. signal quality will never be better than that of the 14-Mc. signal. Properly operated, the mixer will deliver a useful 50-Mc. s.s.b. output of more than 10 watts. This is sufficient for use by itself, or it can be used to drive a high-powered linear amplifier. The setup has been in use at W2UTH both alone as a driver for a 500-watt linear. All reports have been excellent.

The writer thanks W2YPW for his photography.



Bottom view of the 50-Mc. s.s.b. mixer. Capacitor C_4, here shown as a mica trimmer, was later changed to an 8–50-pf. ceramic for greater range.

» *The applications of the unique balanced-modulator circuit used in this exciter aren't confined to 6-meters—the circuit is usable at any frequency. Combining beam-deflection mixing and cascode output coupling, it offers a high degree of stability in balance, along with enough output for driving small power tubes directly.*



The complete 12-watt, 50-Mc. s.s.b. transmitter is shown on the 17×6×3-inch chassis. The 2E26 linear amplifier is enclosed in the shielding box on the right, with its tuning and loading controls on the upper right. The 6JH8 balanced-modulotor tubes are to the right of the loop of coax, and are shown with tube shields removed. The small coil of coax provides 90-degree r.f. phase shift at 50 Mc. The left knob is the audio gain. The sideband-selection switch is directly above. Next to the audio gain control is the 8-Mc. crystal, and on its right are the two carrier-balance controls. On the extreme right is a meter switch that was little used and has been omitted from the schematic. Power requirements are 300 volts at 60 ma. and 600 volts at 90 ma. The —75-volt bias supply and the regulated 150-volt screen supply are included in this chassis.

The Single-Sideband Sixer

JAY GOOCH, W9YRV AND ESTIL CARTER, WA9DNF

A simple, straight-through phasing type 6-meter s.s.b. exciter has been constructed and shows smooth, stable operation. Type 6JH8 beam deflection tubes,[1] V_4 and V_5, Fig. 1, are used as balanced modulators in a novel cascode circuit. Sufficient power is obtained from the balanced modulators along with two 12AT7s, V_6 and V_7, to drive directly a 2E26 linear amplifier which provides 12 watts p.e.p. output measured at the 50-ohm load.

The 50-Mc. suppressed carrier frequency, which provides the r.f. drive to the balanced modulators, is furnished from an 8-Mc. crystal, a triode third-overtone oscillator, V_{3A}, and a pentode frequency doubler, V_{3B}. The two sections of a 6AW8 are used. The 90-degree r.f. phase difference between the control grids of the two balanced modulators is obtained by a length of 75-ohm coax cable, W_1.

A speech amplifier, V_1, a 90-degree audio phase-difference network, Z_1, and two split-secondary audio transformers, T_2 and T_3, provide the required push-pull audio drive for the balanced modulators.

From October, 1963, *QST*.

[1] *Description and Rating Sheet, 6JH8, ET-T3029.* General Electric Company, Receiving Tube Department, Owensboro, Kentucky.

D.c. feedback, obtained from the balanced-modulator cascode-tube outputs, coupled through neon bulbs and the audio transformer split secondaries to the deflector elements, gives unusually good carrier-null balance stability.

The buffer amplifiers in the plates of the balanced modulators are cascode-connected, direct-coupled, and grounded-grid. These have advantages which include:

1. Some r.f. power gain.
2. Increased d.c. feedback gain, resulting in improved carrier-null balance stability.
3. Isolation between plates of the quadrature-driven beam-deflection tubes, resulting in less distortion due to intercoupling.
4. High-impedance output from the cathode driven amplifiers following the balanced modulators, which makes possible more ideal current addition of the two balanced-modulator channels.
5. Elimination of the need for special quadrature balance of the carrier null because of a large increase in the plate resistances shunting the two halves of the output tank circuit.

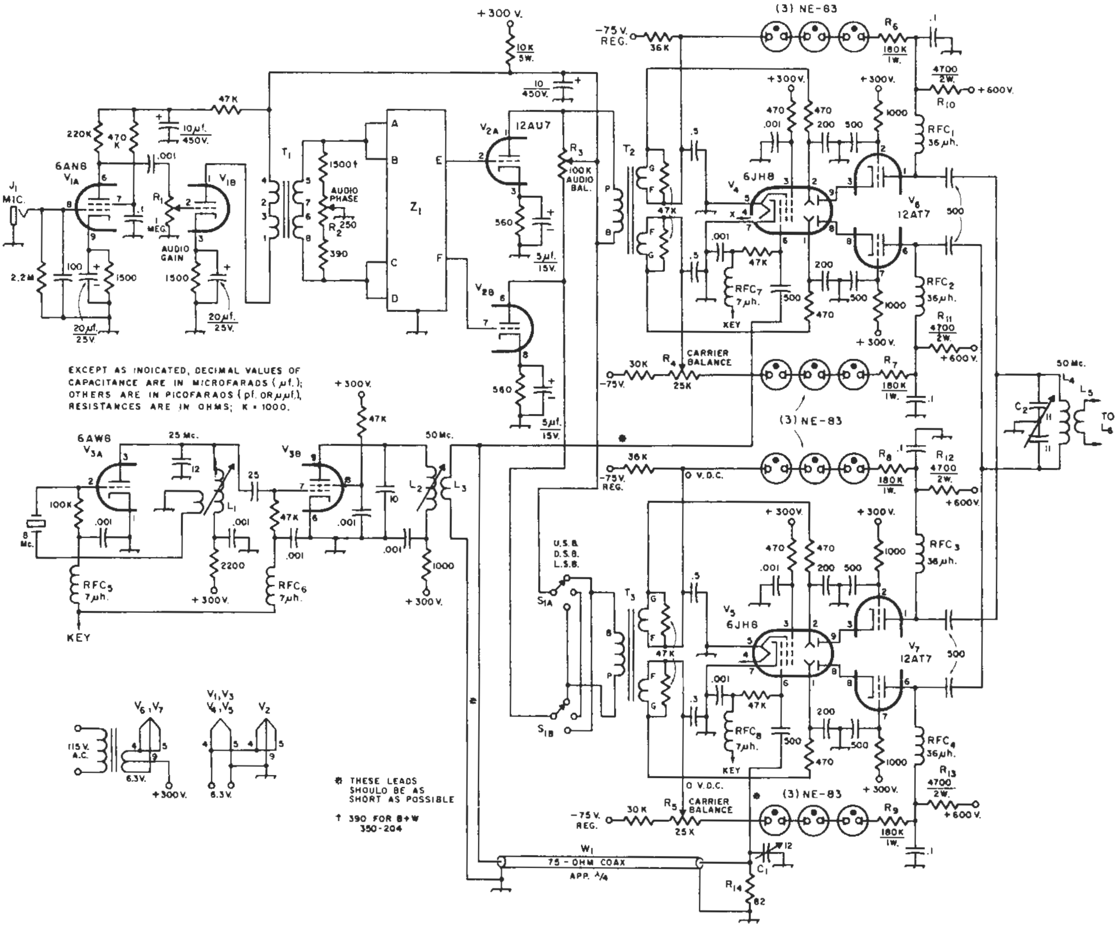


FIG. 1—Circuit diagram of the s.s.b. exciter for 6-meter operation. Except as indicated, resistors are ½-watt composition; fixed capacitors above 0.001-μf. are paper, 0.001-μf. capacitors are ceramic, those with polarity marked are electrolytic, others are mica. See text on heater supply for V_6 and V_7.

C_1—Ceramic piston, 1.5-12 pf. (Cambion CST-50 or equivalent).

C_2—10.8 pf. per section, butterfly (Johnson 11MB11 or equivalent).

J_1—Open-circuit jack (or microphone connector).

L_1—3-5.5 μh., slug-tuned (North Hills 110B, Miller 4504 or equivalent). Feedback coil 5 turns No. 24 close-wound at cold end.

L_2—7 turns No. 24 close-wound on ⅜-inch diam. slug-tuned form (CTC PLS5-2C4L/N or Miller 4400 form).

L_3—3 turns No. 24, diam. ¾ inch, close-wound at cold end of L_2.

L_4—8 turns No. 20, diam. ¾ inch, 16 turns/inch (B & W Miniductor 3011); see L_5.

L_5—Center turn of L_4, cut to form link; inner ends of remaining two sections of L_4 connected together.

R_1—Audio-taper control.

R_2, R_3, R_4, R_5—Linear-taper control.

R_6-R_{14}, incl.—For text reference.

RFC_1-RFC_4, incl.—Single-pie choke (Miller 6176-TV peaking coil or equivalent).

RFC_5-RFC_8, incl.—V.h.f. choke (Ohmite Z-50 or equivalent).

T_1, T_2, T_3—See text and photo captions.

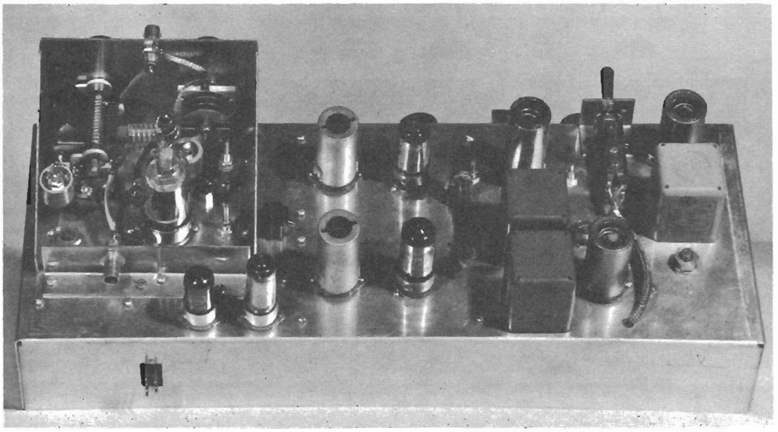
W_1—See text.

Z_1—Audio phase-shift network; see Table 1.

These advantages result *without requiring* additional tuned circuits.

The unit is operated from a TV transformer and bridge-rectifier power supply which furnishes +300 and +600 volts. A −75-volt d.c. bias supply is built into the transmitter.

The beam-deflection tube is attractive as a balanced modulator since it (1) operates well at frequencies as high as 50 Mc., (2) gives moderate power output without excessive distortion, (3) has good inherent carrier-null balance stability due to its single cathode and



The 2E26 compartment can be seen on the left, with perforated metal cover removed. On the extreme left is the pi-network loading capacitor, and to its right is the 2E26 plate tuning capacitor. The piston-type 2E26 neutralizing capacitor is near the rear right corner of the tuning capacitor. It is mounted on top of a 500-pf. feedthrough capacitor which is the partial bypass for the bottom end of L_7, the 2E26 grid coil. T_2 and T_3, the audio transformers for driving the deflectors, are UTC type A-19 but can be replaced by less expensive Chicago-Stancor type A-4774. T_2 and T_3 furnish 12 volts of peak-to-peak audio (4.2 volts r.m.s.) to each deflector element of each 6JH8.

control grid associated with two anodes, and (4) has separate high-impedance inputs for r.f. and for audio drive.

Several phasing-type beam-deflection-tube exciters have been described.[2,3,4] These generally have been lower-frequency, lower-power units which require additional amplifier or heterodyne stages for v.h.f. This unit, although simple, is attractive as a complete transmitter for a 6-meter station.

50-Mc. Carrier Generation

An 8-Mc. crystal is used in a third-overtone oscillator with its output on 25 Mc. This is followed by a pentode frequency doubler. The two sections of a 6AW8 are used in a circuit similar to that in *Handbook* v.h.f. transmitters.[5] A link, L_3, on the output coil, L_2, of the doubler is connected to two paralleled loads. One is the control grid of the first balanced modulator. The other is a length of 75-ohm coax cable which provides 90-degree phase shift to the 50-Mc. signal and feeds the grid of the second balanced modulator.

R.F. Phase Shift

This phase shift is accomplished by an approximate quarter wavelength of 75-ohm coax cable. Both subminax, 21-579, and RG-59/U were tried with no noticeable difference.

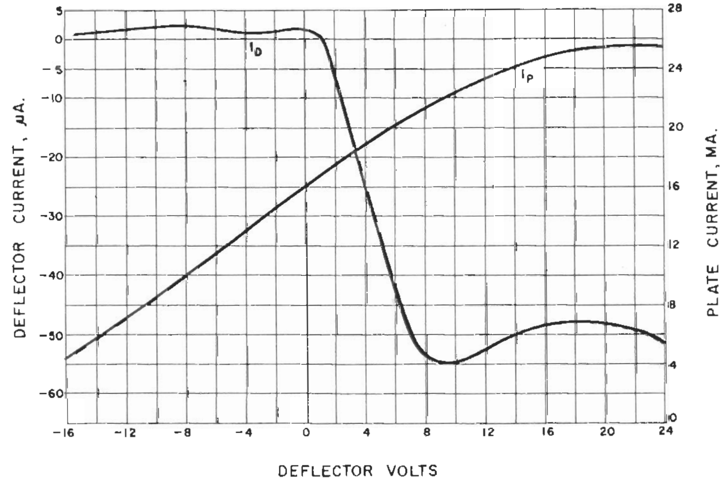
[2] Vance, "S.S.B. Circuits Using the 7360," page 29.

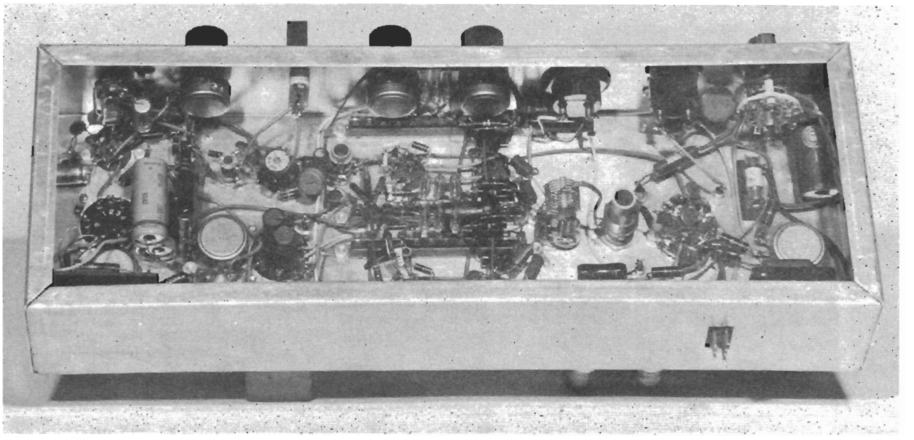
[3] "A Phased Single-Sideband Exciter," *The Radio Amateur's Handbook*, ARRL, 38th edition (1961), pp. 307-312.

[4] Evans, "Another Phasing-Type S.S.B. Exciter" page 87.

[5] "Simple Transmitters for 50 and 144 Mc.," *The Radio Amateur's Handbook*, 38th edition (1961), p. 437, and 39th edition (1962), p. 442.

FIG. 2 — Deflector current and plate current as a function of deflector voltage, 6JH8 beam-deflection tube.





The 6AN8 speech amplifier socket is in the upper left corner. The overtone crystal oscillator and doubler are adjacent to the crystal socket on the front panel. Two carrier-balance pots occupy the center position of the front panel. The left-hand miniature meter reads r.f. output voltage and the right-hand meter reads 2E26 plate current. The wafer switch at upper right permitted reading 2E26 plate, screen, or control grid current, but proved of little use and was omitted from the schematic. At center right is the 2E26 socket and, on its left, the 2E26 grid tank coil, with link coupling from the balanced modulator plate coil. The silicon bias rectifier is at the far right center, adjacent to the bias-supply electrolytic filter capacitor. The 6JH8 balanced modulator sockets with their short leads connecting to the coax cable are below the left carrier-balance pot. This photo shows that all coils and tuning capacitors except the 2E26 plate components are mounted below the chassis. The 2E26 plate circuit components are kept above the chassis for isolation. The output connection to the antenna is also above the chassis.

The amount on phase shift furnished by the coax depends on its length, its characteristic impedance, and its terminating impedance. Here the 75-ohm cable is terminated in approximately 8 pf., the capacitance of the 6JH8 control grid input. An 82-ohm resistor, R_{14}, is added to lower the v.s.w.r. on the coax. This avoids the large change in phase shift which accompanies a very small length change in a cable operated at a high v.s.w.r.

Simplest construction if operation near 50.2 Mc. is intended is to omit C_1, the 1.5- to 12-pf. trimmer capacitor, and to cut the total length of coax to 31¾ inches. This is estimated to give phase shift of 90 ± 1 degrees over about a ½-Mc. frequency range. Lengths for other spot frequencies in the 6-meter band, where the cable is terminated in the 6JH8 control grid capacitance, can be calculated from

$$Length,\ inches = \frac{1592}{Freq.\ in\ Mc.}$$

This length was verified both for RG-59/U and Amphenol subminax 21-579, both of which have a velocity factor of 0.66. Alternate construction is to incorporate a 1.5-12 pf. piston capacitor (C_1) as an r.f. phase adjustment and cut the coax to 29 inches. This allows adjustment that will give a 90-degree phase shift on any frequency in the 50 to 54-Mc. band.

The r.f. voltages measured at the control grids of the two 6JH8 tubes are nearly equal. They read 5.1 volts r.m.s. on the link and 5.3 volts r.m.s. at the terminated end of the coax.

Cascoded Balanced Modulators

The control grids of the two 6JH8 balanced-modulator tubes, V_4 and V_5, are fed 90-degree phase difference r.f. Likewise, each tube has its set of deflector elements fed 90-degree phased push-pull audio. Each tube has a push-pull double-sideband signal appear at its two anodes. When equal outputs from the two channels are added, single sideband results. Previous experiments, without the cascode amplifiers, indicated that a considerable amount intercoupling was occurring between the two beam-deflection tubes when the anodes were tied directly together to add their outputs. Only small output was possible before distortion became large. With outputs in parallel, but inputs consisting of quadrature signals, it appeared that distortion was being caused by the first balanced modulator plate modulating the second, and vice versa.

A better understanding of the 6JH8, or any of the other beam deflection tubes, can be had by considering the following:

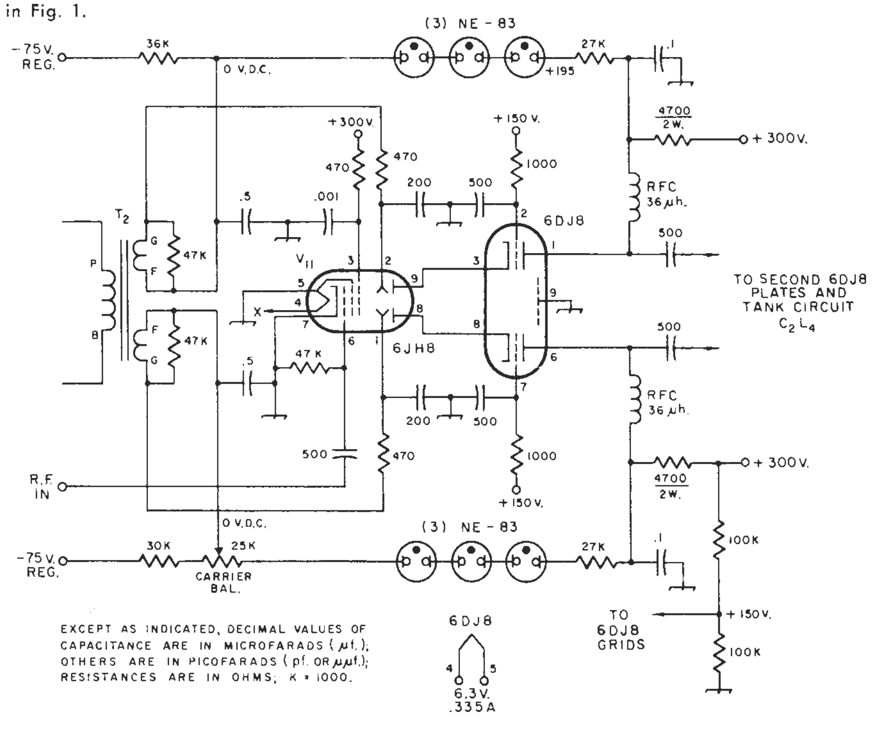
The 6JH8 deflector-anode characteristics, which are typical of most beam-deflection tubes, describe the portion of the tube beginning at the accelerator (screen grid) and including the deflectors and two anodes. If this part of the tube is considered to be analogous to a dual triode with common cathode, operating in class-A push-pull, the class-A characteristics of one triode section, which is analogous to half of the above portion of the 6JH8, would have an amplification factor (μ) of 6.3, a plate resistance of 9000 ohms, and a transconductance (g_m) of 700 micromhos.[6] The output circuit of the 6JH8 should be designed considering the equivalent of two of these triodes in class-A push-pull.

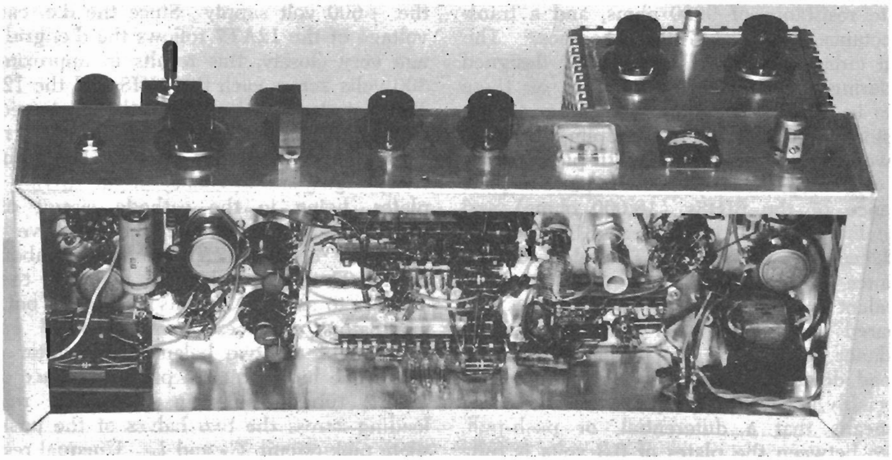
Distortion caused by intercoupling between the two balanced modulators with plates paralleled was attributed to the low plate-to-plate resistance of the 6JH8s (18,000 ohms) and to the low amplification factor (6.3) between the deflectors and anodes. The plate-to-plate resistance of the tube, when viewed as the equivalent generator internal resistance, is low compared with the usual tank-circuit load impedance. Therefore, the output tank does not load down the voltage swing of the plates appreciably. The low amplification factor of 6.3 means that a differential, or push-pull, voltage between the plates of 6.3 volts is fully as effective in deflecting the beams as would be a one-volt push-pull signal applied between the deflectors. A rather unfortunate situation exists whereby one tube can very effectively plate modulate the other. To avoid this intercoupling without undue circuit complexity, cascode 12AT7 triodes, V_6 and V_7, were added in the plate leads of the 6JH8 tubes. The grids of V_6 and V_7 were connected to +300 volts and grounded for a.c. through bypass capacitors. The 12AT7 plates were fed from the +600 volt supply. Since the d.c. cathode voltage of the 12AT7 follows the d.c. grid voltage very closely, this results in approximately 300 volts across each the 6JH8 and the 12AT7. The plate load of the 6JH8 is the low impedance looking into the 12AT7 cathodes. This makes a good high-frequency circuit. The comparatively high plate resistance of the driving 6JH8 plates, being in the cathode circuit of the 12AT7, acts to increase the effective plate resistance of the 12AT7 tubes by about 40 times. Thus paralleling the plates of the two 12AT7s to add the outputs of the two balanced modulators produces a nearly ideal current addition of the two balanced-modulator channels. The high effective plate resistance of the 12AT7 tubes results in very little resistive loading across the two halves of the push-pull plate tank circuit, C_2 and L_4. Unequal resistive loading of the two halves of the output tank circuit would cause out-of-phase currents in the two halves of the tank. This would make fairly critical "quadrature null balance" adjustment[7] necessary to achieve good carrier suppression. However, the high effective plate resistance of the grounded-grid cascode tubes, V_6 and V_7, which results from being cathode-driven by the 6JH8 plate resistance, makes a

[6] Private Correspondence: W. P. Kimker, Advanced Applications, General Electric Company, Owensboro, Kentucky.

[7] Quadrature balance is usually achieved by a differential capacitance or resistance connected across the plate tank circuit with the capacitor rotor or resistor arm grounded. Also see p. 285 of M. B. Knight, "A New Miniature Beam-Deflection Tube," *R.C.A. Review*, Vol. 21, No. 2 (June, 1960), p. 266.

FIG. 3—Alternative balanced-modulator circuit for 300-volt supply. Circuit designations and components same as in Fig. 1.





Another view inside the chassis. Mounted on the resistor board at left bottom is the homemade audio phase-shift network. A commercial audio phase network could plug into an octal socket mounted at this location. The NE-83 neon bulbs, identical in appearance to NE-2 types, are mounted on tie points near the center of the chassis. NE-2 bulbs can be substituted but have a shorter operating lifetime. The audio components, including the amplifier and deflector driving transformers, are located in the left one-fourth of the chassis. The balanced-modulator plate coil is mounted by its leads on the back of the vertical split-stator tuning capacitor, C_2, adjacent to the left wall of the 2E26 compartment. On the right rear is the backwards-operated filament transformer which is part of the —75-volt bias supply.

quadrature balance control completely unnecessary. The resulting effect of all these considerations is a balanced modulator circuit that has excellent performance.

It should be noted that an isolated 6.3-volt filament supply, with the filament line connected to + 300 volts, should be used to supply the filaments of the cascoded 12AT7 tubes, V_6 and V_7.

Audio System

The speech amplifier is conventional, largely taken from the *Handbook*.[8] The triode second stage is, for convenience, transformer-coupled to the audio phase-difference network. Any plate-to-line transformer should be satisfactory for T_1. Connections for commercial audio phase-shift networks are shown in Table I. A comprehensive description of how to build such a network has been given.[9] The network used in the model shown in the photograph was homemade. The audio phase network outputs are connected to the grids of amplifiers V_{2A} and V_{2B}, the two halves of a 12AU7, which have the primaries of T_2 and T_3 as their plate loads. Better differential gain control was obtained by using a plate-load-shunting audio balance control, R_3, than from the more usual cathode-bias control. To help realize maximum output with low distortion in the balanced modulators, push-pull audio from the secondaries of T_2 and T_3 was used to drive the deflectors. While single-ended audio drive to one deflector is simpler, lower distortion results from this push-pull drive.

Under conditions of high deflector d.c. return resistance, or high drive impedance, deflector secondary emission currents have been found to cause distortion.[10] D.c. return resistance through the transformers is low.

Deflector currents of the 6JH8 were measured under large drive conditions and were found to be lower than those of the 7360 or 6HW8. The measured 6JH8 deflector currents are shown in Fig. 2.

Deflector circuit d.c. return was made through the relatively low resistances of the secondaries of transformers T_2 and T_3, and the beam-centering d.c. voltages are series-fed through the appropriate split secondaries of the transformers. The sum of a d.c. and an audio voltage appears at each deflector.

D.C. Carrier-Null Feedback

Drift in the balanced-modulator tubes could necessitate frequent readjustment of the carrier-null balance controls. However, use is made of the fact that r.f. output from each half of the beam deflection tube is closely related to the amount of d.c. plate current drawn by that half tube. A sample of the plate current is obtained in each case by taking the proportional voltage drop across a plate decoupling resistor (R_{10}, R_{11}, R_{12}, R_{13}) and feeding this voltage back through neon bulbs where it is applied as a d.c. beam center stabilizing voltage on the deflector element. The burden of

[8] "A 25-Watt Modulator Using Push-Pull 6BQ6GTs," *The Radio Amateur's Handbook*, ARRL, 39th edition (1962), p. 272.

[9] "S.S.B. Jr.," *G.E. Ham News*, Vol. 5, No. 6 (November-December, 1950).

[10] Technical Correspondence "7360 Deflector Currents and Large-Signal Operation," *QST*, March, 1962, p. 41.

maintaining carrier balance is thus transferred from the active tube structure to the more stable d.c. divider chain. The direct-coupled 12AT7 tubes, neon bulbs, and divider return to the −75-volt supply result in excellent stabilization of carrier-null adjustment.

TABLE I

Connections for Commercial Audio Phase-Shift Networks

Terminal (Fig. 1)	Central Electronics PS–1	B & W 350–2Q4	Millen 75012
A	2	1	A–IN
B	6	5	B–IN
C	3	3	A–COM
D	7	7	B–COM
E	4	2	A–OUT
F	8	6	B–OUT
GND	1		Case

The nominal drop (constant 65 volts) across the NE-83 neon bulbs may vary from bulb to bulb and require trimming of resistors R_6, R_7, R_8 and R_9. This can be done by centering the carrier-balance pots and placing a d.c. voltmeter on the deflectors. The four resistors should be trimmed in value until each deflector is within a few volts of ground under no-signal conditions.

6JH8 Cascoded with 6DJ8

The alternative balanced-modulator circuit shown in Fig. 3 has been used with a 300-volt supply. Here, the plates of the cascode tubes are operated from 300 volts, and their grids are pegged at +150 volts. Resistor values in the neon bulb-resistor divider chain are changed, so that the deflectors remain within a few volts d.c. of ground, under no-signal conditions, with the +300-volt power-supply voltage.

The cascode-tube type is changed to a 6DJ8.[11] This tube performs well in this lower-voltage circuit because its characteristics are such that it will handle the required 30 ma. peak plate current with as little as 75 volts across plate to cathode—and without the grid going positive with respect to cathode.

The circuit will not provide sufficient drive for the 2E26 at 50 Mc., a fact made clear by the condition of no excess drive for the 2E26 from the balanced modulators operating from a 600-volt supply. However, at lower frequencies where drive is easier, or when driving a smaller tube such as a 6BQ5 or 6CL6, the 6JH8 and 6DJ8 balanced modulators work well.

Neon Bulbs

NE-83 neon bulbs were used.[12] These appear similar to NE-2 bulbs, but are an improved design, rated at 500 hours of life at 10 ma. current, and at greatly extended life at lower currents, such as the 2 ma. used here.

The NE-83 should prove useful in furnishing regulated voltages in multiples of 65 volts where currents under 10 ma. are required.

Survey of Beam Deflection Tubes

The 6JH8 beam deflection tube was picked for this unit after considering the 7360, 6AR8, 6JH8, 6HW8, and 7763. Its merits are a useful peak plate current which is approximately three times that of the 7360, and a higher plate dissipation. The large-signal deflector current is lower than that of the 7360 or 6HW8. Although the 6JH8 has a higher-current beam, it has a more elaborately-focused electron gun and the deflection linearity is better than that of the 7360 or 6HW8. The tube capacitances are about equal: More deflector drive voltage (audio in the present case) is required by the 6JH8. The 6AR8 is an early tube[13] and had poor internal anchoring of elements, causing a shift of carrier null under mechanical shock.

None of these has a screen grid between deflectors and anodes, and a cascode circuit can sometimes improve performance. The 7763 has a screen grid between deflectors and anodes, but it is an experimental high frequency limiter tube and has no control grid.

Balanced modulators operate at lower efficiency than do linear amplifiers. For this reason, balanced modulators usually are operated at low levels followed by large amplification. However, for simplicity in this unit, the balanced modulators are operated at a high enough power level to be able to drive the 2E26. Fortunately, the 6JH8 is available to do this at low distortion.

Single-sideband power generation at 50 Mc. is more difficult than at lower frequencies, and there is no reserve of drive for the 2E26. For this reason, low-loss coils and capacitors and high L-to-C ratios should be used in the balanced modulator output and the linear-amplifier grid tuned circuits, consisting of L_4, C_2 and L_5.

2E26 Linear Amplifier

Straightforward design and construction were used for the linear amplifier, Fig. 4. The tube is operated with +150 volts regulated on the screen grid and 600 volts on the plate. This screen voltage is slightly less than is usually specified, but it results in a negative bias closer to zero (about −16 volts) than would

[11] *Tube Data Sheet 6DJ8*, Amperex Electronic Company, 230 Duffy Avenue, Hicksville, Long Island, New York.

[12] *Glow Lamps as Circuit Control Components*, Miniature Lamp Department, General Electric Company, Cleveland 12, Ohio. Bulletin 3-1177.

[13] Adler and Heuer, "Color Decoder Simplifications Based on a Beam Deflection Tube," Trans I.R.E., PGBTR (January, 1954), p. 64.

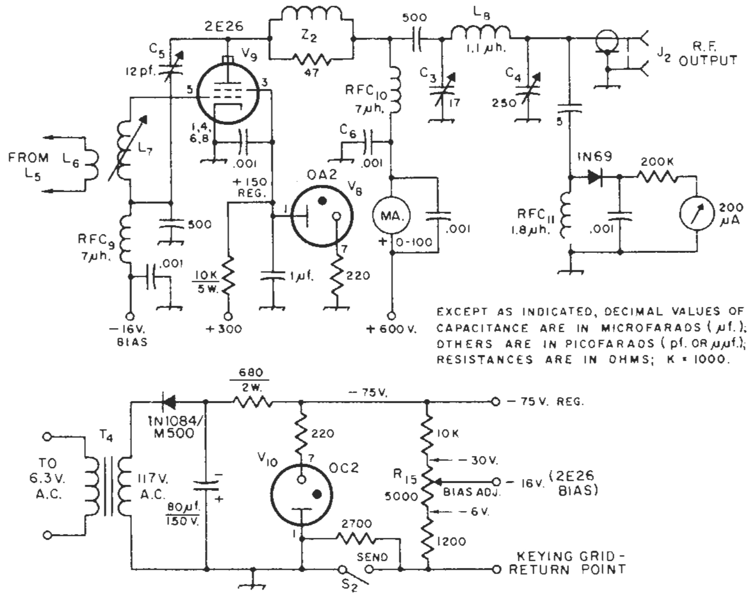


FIG. 4—Linear amplifier and bias supply. Except as indicated, resistors are ½-watt composition; 0.001-µf. fixed capacitors are ceramic, those with polarity marked are electrolytic, others are mica.

C_3—Approx. 17-pf. variable (Hammarlund HF-15 or equivalent).

C_4—250-pf. variable (Hammarlund MC-250-M or equivalent).

C_5—Ceramic piston, 1.5-12 pf. (Cambion CST-50 or equivalent).

J_2—Coax receptacle, chassis-mounting.

L_6—2 turns No. 20 wound on form for L_7 (½ inch diam.) at cold end.

L_7—6 turns No. 20, diam. ¾ inch, 16 turns/inch (B & W Miniductor 3011) cemented on ½-inch diam. slug-tuned form (CTC PLS7-2C4L/Q or Miller 43A000CBI).

R_{15}—Linear-taper control.

RFC_9, RFC_{10}—7 µh. (Ohmite Z-50 or equivalent).

RFC_{11}—1.8 µh. (Ohmite Z-144 or equivalent).

T_4—Filament transformer, 6.3 volts, 1 amp.

Z_2—3 turns No. 18 wound on 47-ohm, 2-watt composition resistor.

be required with higher voltage; consequently, less r.f. drive voltage is required to overcome the bias and drive the grid up to the required peak of zero volts.

Bias is adjusted, under no signal conditions, by R_{15}, until plate current is about 20 ma. This adjustment places about −16 volts on the grid. At a plate voltage of 600 volts, the tube is at rated plate dissipation. During voice peaks the dissipation rating is exceeded slightly, but no ill effects resulted from prolonged single tone testing; and operation at this screen and bias voltage minimizes linear amplifier crossover distortion.

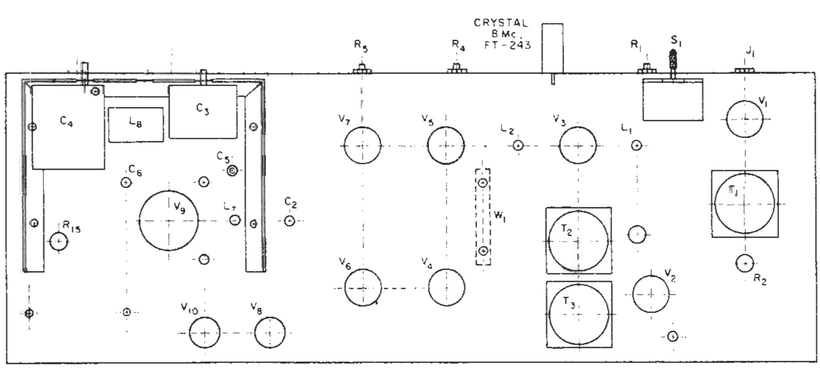


FIG. 5—Top-of-chassis layout of the 6-meter s.s.b. transmitter. This drawing is to scale and may be used for layout dimensioning; ¼ inch in the drawing above equals one inch of actual chassis size.

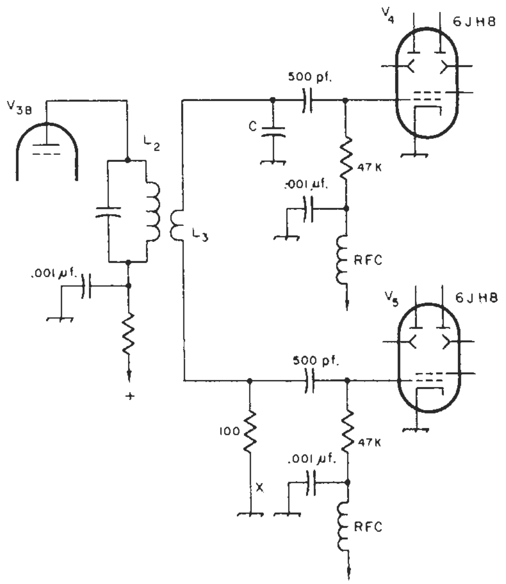


FIG. 6—Ninety-degree r.f.-phasing network for use on lower frequencies where transmission-line elements are impracticably large. L_2 and L_4, Fig. 1, must be replaced by tuned circuits appropriate for the frequency used. L_3 and its leads should have low capacitance to ground. Values for capacitor C are given below:

Freq. Mc.	Capacitance, pf.
3.9	400
7.2	210
14.3	100
21	65
29	46

A parasitic suppressor was incorporated only in the 2E26 plate lead. Precautions were taken in layout to keep the balanced-modulator output tank and the 2E26 grid coil under the chassis, while the 2E26 plate coil and capacitor were kept above the chassis.

The bias supply, regulated at −75 volts, also is shown in Fig. 4. This supply also furnishes the −75 volts for the deflector d.c.-voltage-divider return point.

A conventional pi network is used in the 2E26 plate circuit to match to a 50- or 75-ohm coax line to the antenna.

Adjustment for SSB Output

With an audio oscillator and oscilloscope, adjustment for sideband and carrier suppression is fairly routine. The procedure for phasing-type exciters has been well covered.[14,15] Adjustment of the 2E26 linear amplifier includes neutralizing and loading.

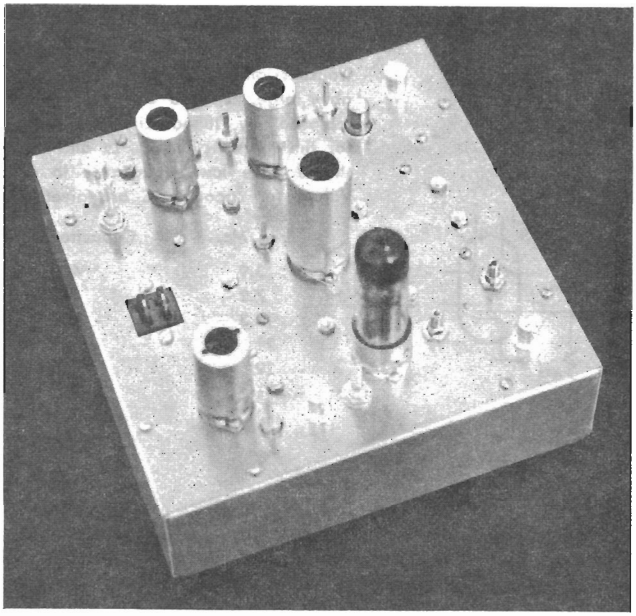
[14] Erlich, "How to Adjust Phasing-Type S.S.B. Exciters," page 207.
[15] *The Radio Amateur's Handbook*, ARRL, 38th edition (1961), p. 310.

Additional features and accessories can certainly be added, but the mission of this article is accomplished if it is an aid to you in generating an s.s.b. signal on v.h.f.

It has been suggested that this balanced-modulator circuit might also be useful on lower frequencies. This has indeed been the case in this area, where Jack Washburn, W9IVB, and Lou Goodman, W9DCZ, are using this circuit on 4 Mc. with excellent results. Drive power on 4 Mc. is more plentiful than at 50 Mc., and the balanced modulators are used to drive a 6146 in one case and a pair of 6DQ6Bs in the other.

The quarter-wave coax is impractical for r.f. phase shift at lower frequencies, and the circuit shown in Fig. 6 is used. The value of C is selected for the frequency band used from the table in Fig. 6. The link, L_3, should have a minimum of capacitance to ground. Circuit calculations show that an inductance of about 0.08 μh. should be used at point X in Fig. 6, but experimentally no improvement could be found, probably because the distributed inductance of the 100-ohm resistor is near this value.

» A "transverter" is a device for giving you v.h.f. transceiver operation from a low-band transceiver. This one opens up the 144-Mc. band for you if you have a 14-Mc. transceiver.



This looks like a typical v.h.f. converter, but it is more than that: a 144-Mc. transverter which will handle both transmitting and receiving conversions from a 14-Mc. station setup for s.s.b. or other modes.

Two-Meter Transverter

ERNEST P. MANLY, W7LHL

This 6-tube package will convert 144-Mc. signals down to 14 Mc., and a 14-Mc. transmitter signal up to 144 Mc. An s.s.b. transceiver for 20 will work beautifully with the transverter, or a communications receiver and a 20-meter transmitter may be used. The 20-meter band was picked for the conversion process because it was felt that there is more s.s.b. gear on 20 than on any other band that would be suitable for this purpose.

The block diagram, Fig. 1, shows how the system works. The top line of the diagram is a typical 2-meter converter, with 6CW4 and 6AK5 r.f. stages ahead of a 6AK5 mixer. Mixer output is 14 Mc. or higher. In the middle is the oscillator-multiplier, a 6AN8 with its triode working as a crystal oscillator on 43 Mc. and the pentode tripling to 130 Mc. This is standard v.h.f. converter practice so far. The difference is that the same injection stages are used for transmitting. In the bottom line, the 130-Mc. energy is fed to a 6AK5 amplifier, and then to a 6360 transmitting mixer. The 14-Mc. signal from the 20-meter sideband rig is also fed to this mixer, and output is on 144 Mc., with the same characteristics as the 20-meter signal. Though the transverter idea is associated with s.s.b. in most amateurs' minds, it can be used with other modes as well.

From September, 1963, *QST*.

Circuit Details

The first r.f. amplifier in the receiver side, V_1, Fig. 2, is a grounded-grid Nuvistor. Its cathode input impedance is matched by means of a quarter-wave section of 93-ohm coaxial line, L_{16}. An alternate method would be to tune the cathode coil, and tap the antenna line down on it. These matching systems give equal performance. Gain of the grounded-grid stage is about 10 db. The second stage, V_2, is a 6AK5 pentode, with a gain of 25 to 30 db. If cross-modulation is expected to be a problem, a gain control could be included readily in this stage. Both this and the 6AK5 mixer, V_3, follow conventional converter circuit practice.

The three parts of the transverter are separated by shields. This permits the injection to be adjusted to the desired level, and provides isolation, to keep down spurious responses. Output from the 6AN8, V_4, at 130 Mc., is link-coupled to the receiving mixer grid circuit, with light coupling at each end. Noise figure of the converter is under 4 db. Other tubes could be used equally well in the oscillator-multiplier stages, and early versions of this transverter used a 12AT7 for this purpose. The triode-pentode gives somewhat more output, which is helpful in the transmitting side.

The injection stages run at low input, for good stability, so an amplifier stage is needed to build up the 130-Mc. energy for the transmitting mixer. This is done with the 6AK5 stage, V_5, which also helps to keep down the energy injected into the transmitting mixer at frequencies other than 130 Mc. Its output is link-coupled to the cathode of the 6360 mixer, V_6. Energy from the 14-Mc. transmitter is fed push-pull to the mixer grid circuit, which is tuned to 14 Mc. The mixer plate circuit is also push-pull, and tuned to 144 Mc.

A 22-volt zener diode, CR_1, is used in the mixer cathode circuit for bias, eliminating the need for an external bias supply. Operating conditions are designed to keep the tubes below their maximum dissipation rating if the crystal is removed or if it drops out of oscillation.

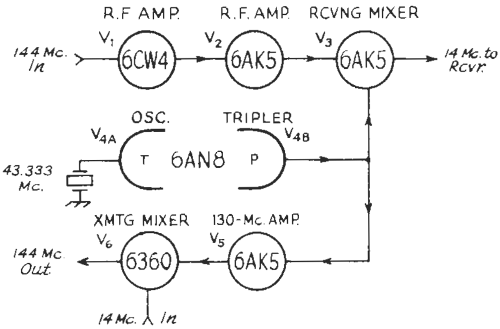


FIG. 1—Block diagram of the W7LHL 144-Mc. transverter. A single oscillator-multiplier section furnishes both transmitting and receiving injection, to separate mixers.

Construction and Adjustment

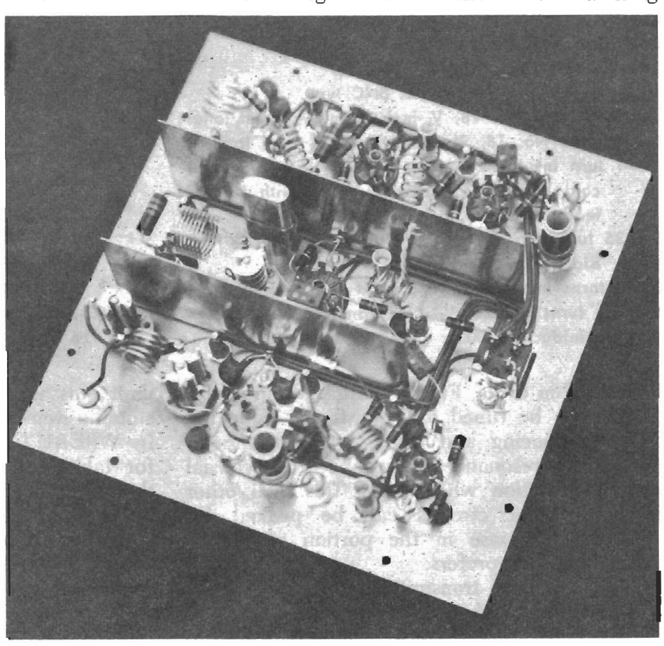
The transverter is built on a 7 × 7 × 0.032-inch brass plate. The shields are also made from 0.032-inch brass. Aluminum should work equally well. A 7 × 7 × 2-inch aluminum chassis is used to mount the transverter. The insulated terminals are CTC type 1581. Parts layout follows good v.h.f. practice, but is otherwise not particularly critical.

Tuning the oscillator, tripler and 130-Mc. amplifier is done by inserting a 50-ma. meter in the 6360 plate supply lead. The tube will draw approximately 1 ma. of plate current without drive, increasing as each stage is peaked. A grid-dip meter may be used to check each tuned circuit, to be sure that it is on the right frequency. Adjust the link between the 130-Mc. amplifier and 6360 cathode for maximum plate current, 10 to 20 ma. when 130-Mc. tuning is completed. A 14-Mc. carrier is fed in and the 6360 grid and plate circuits are tuned for maximum output at 144 Mc. The 14-Mc. carrier will drive the plate current to about 40 ma. before the 144-Mc. output starts to flatten out. Output at 144 Mc. will light a No. 47 pilot light to near full brightness.

There will be a little 130-Mc. and 116-Mc. energy appearing in the output, too. Some form of filter is desirable between the mixer output and the antenna or following linear-amplifier stages. A high-Q coaxial tank circuit with low-impedance coupling in and out will serve this purpose well. Adjustment of coupling into and out of such a filter should be made for maximum attenuation of unwanted frequencies, rather than maximum transfer of 144-Mc. energy.

A Z-144 r.f. choke was used originally in the 6360 plate circuit in place of the 100-ohm resistor shown in Fig. 4, but this resulted in the mixer operating as a doubler from 14 to 28 Mc., because of choke resonance near the latter frequency. A Z-235 r.f. choke or the 100-ohm resistor will correct this tendency.

When using the circuit shown for matching



Bottom view of the 2-meter transverter. Receiving stages are at the top, transmitting section of the bottom, and injection stages in the center section. Construction follows v.h.f. receiving converter practice.

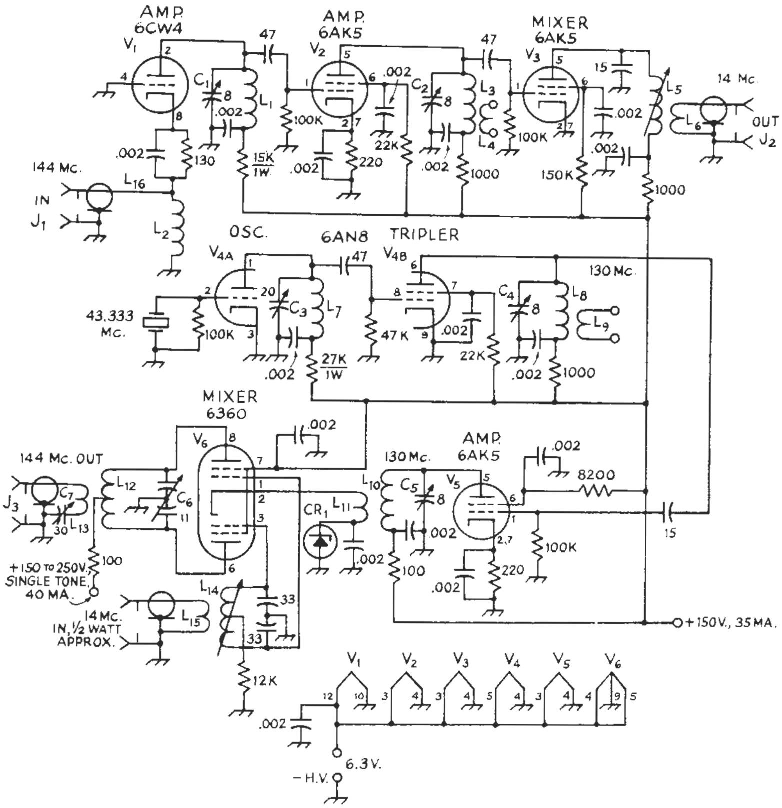


FIG. 2—Schematic diagram and parts information for the 144-Mc. transverter. Capacitors are ceramic unless specified. Decimal values of capacitance are in μf., others in pf. Resistors are ½-watt composition, unless specified.

C_1, C_2, C_4, C_5—8-pf. cylindrical trimmer.
C_3—20-pf. miniature trimmer.
C_6—11-pf. miniature butterfly variable.
C_7—30-pf. miniature trimmer.
CR_1—22-volt Zener diode. (1N1527).
J_1, J_2, J_3—Coaxial receptacle, BNC type.
L_1, L_3—5 turns No. 18, ¼-inch diam., ¾ inch long.
L_2—4 turns No. 18, ⅜-inch diam., ⅜ inch long.
L_4—1 turn insulated hookup wire, ¼-inch diam., at cold end of L_3. Connect to L_9 with twisted leads 2⅜ inch long.
L_5—35 turns No. 26 enam., close-wound on ⅜-inch diam. iron-slug form.
L_6—3 turns insulated wire around cold end of L_5.
L_7—7¼ turns No. 20, ½-inch diam., 16 t.p.i. (B & W Miniductor 3003).
L_8—5 turns No. 18, 5⁄16-inch diam., ½ inch long.
L_9—1 turn insulated hookup wire around cold end of L_8. Connect to L_4 with twisted leads. L_4 and L_9, plus leads, can be made from single piece of wire.
L_{10}—5 turns No. 18, ⅜-inch diam., ½ inch long.
L_{11}—1 turn insulated wire around cold end of L_{10}
L_{12}—5 turns No. 16, ⅜-inch diam., ½ inch long, center-tapped.
L_{13}—1 turn insulated hookup wire at center of L_{12}.
L_{14}—33 turns No. 26 enam., close-wound on ⅜-inch diam. iron-slug form, center-tapped.
L_{15}—2 turns insulated wire around center of L_{14}.
L_{16}—93-ohm coaxial matching section; RG-62/U, 16 inches long. See text.

with 93-ohm cable, Fig. 2, the input coil of the 6CW4 can be tuned to 144 Mc. by adjusting the turn spacing and using a grid-dip meter to indicate resonance. Disconnect the coaxial matching section when doing this. All other r.f. and mixer circuits can be peaked for maximum response in the portion of the 2-meter band one prefers.

Most 20-meter transmitters and transceivers will have more than enough power to drive the transverter mixer. A way must be found to limit this power. Construction and use of a suitable step-type attenuator has been described by W9ERU.[1] Some transmitters have provision for taking off power at levels below that of their output amplifiers, and such a tap should be used, where available.

Output from the 6360 mixer is sufficient to drive a linear amplifier with a pair of 4X250Bs or similar tubes to several hundred watts input on s.s.b. or c.w.

[1] Hubbell, "A Step-Type Attenuator," p. 00.

» *Single-sideband reception by the phasing system uses principles similar to those of the phasing method of generating a single-sideband signal. The same audio phase-shift networks can be used in both applications. This article explains the operation, and also discusses the advantages of exalted-carrier reception.*

Single-Sideband Reception by the Phasing Method

DONALD E. NORGAARD, W6VMH, ex-W2KUJ

The fact that sidebands—and sidebands alone—provide transmission of intelligence makes single-sideband systems possible. In the case of amplitude modulation or phase modulation, the carrier (by definition of these modes of transmission) must be transmitted along with the sidebands that appear in symmetrical pairs about the carrier. The carrier plays no part in the transmission of intelligence, but it is used in normal reception to act as a "key" for the demodulation (detection) process. Sometimes this key fails to work because of selective propagation or because of interfering signals that reach the detector along with the desired signal. The result is either partial or complete loss of the desired transmission.

Extreme selectivity ahead of the detector in a receiver can help prevent blanketing effects from strong adjacent-channel signals, but in itself is not a complete solution to the problem of amateur phone reception. To carry the analogy of the key a little further, it might be said that every transmission must be "unlocked" by a key—the right key—in order to be received. The transmission can be jammed by other keys that fail to work or prevent the right key from being used. The obvious solution to this situation is to keep the right key in the *receiver* all the time, so that other keys cannot jam the detector. This is basically the idea of "exalted-carrier" demodulation, in which a strong "synthetic" carrier is supplied to the detector to demoduate the sidebands of the desired transmission and to make other signals subsidiary to this one key signal. Reception of c.w. signals has always employed this principle, but it can be applied to phone reception, too.

Single-sideband reception can be employed on either single-sideband transmissions (s.s.b. or c.w.) or on double-sideband transmissions (a.m. or p.m.), since in the latter the upper and lower sidebands contain identical information in duplicate. It has been found that the combination of exalted-carrier operation and single-sideband reception is of great benefit in overcoming the vast devastation caused by QRM and selective fading. Another feature of this mode of reception is that with a receiver so equipped one may listen to either of the two sidebands characteristic of a.m. or p.m. transmissions (and dodge some QRM), receive s.s.b. phone transmissions of either upper or lower sideband, or receive c.w. signals in real single-signal fashion.

From "Practical Single-Sideband Reception," *QST*, July, 1948.

Dual Exalted-Carrier Demodulator

A brief explanation of Fig. 1 is in order, since an understanding of the characteristics of this form of exalted-carrier demodulator will make it easy to follow the explanation of the single-sideband receiving system as a whole. Two relatively strong signals of the same frequency but having 90° phase relationship are supplied to the two demodulators by the oscillator and the 90° r.f. phase-shift circuits as indicated. These signals may have a peak voltage of 100 to 150 volts each. If a one-volt signal having a frequency, for example, 1000 cycles different from that of the oscillator, is applied to these demodulators, the output of each will be a heterodyne tone of 1000-cycle frequency at one-volt amplitude. The interesting and useful thing about this otherwise commonplace result is that the two audio output signals will have a 90° phase relationship, and this will hold true regardless of the heterodyne frequency. This phase relationship *reverses,* however, when the one-volt signal causing the heterodyne is on the reverse side of zero beat. In other words, if the small signal has a frequency lower than that of the oscillator, a plus-90° phase relationship is produced; if the

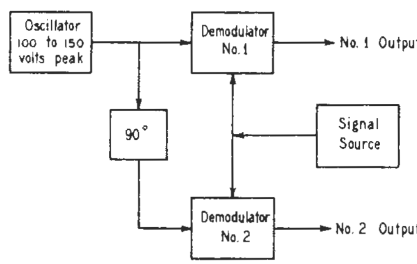


FIG. 1—The elements of a dual exalted-carrier demodulator.

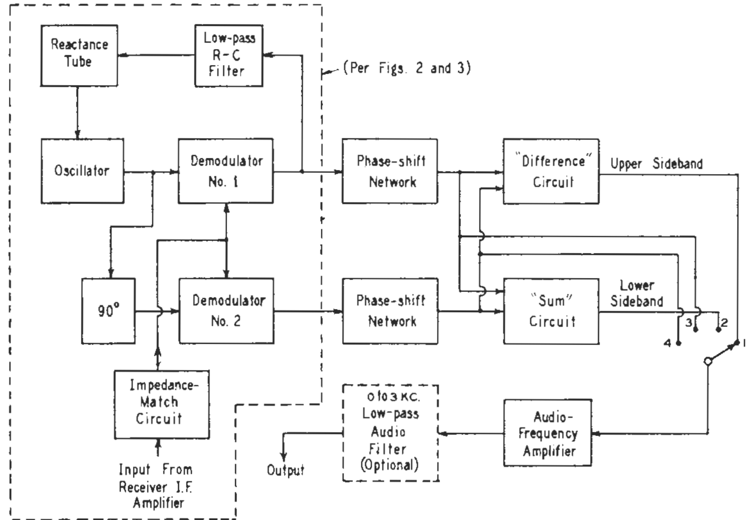


FIG. 2—A block diagram of a single-sideband receiving system incorporating exalted-carrier demodulation.

frequency is higher, a minus-90° phase relationship results.

Naturally, Fig. 1 has been simplified a little for purposes of explanation. The oscillator serves as a synthetic carrier that is so large compared to all other signals that it controls the action of each demodulator. The signal source may be the r.f. and i.f. portion of a receiver. The oscillator operates at intermediate frequency, replacing the b.f.o. of the conventional receiver setup.

Single-Sideband Receiving System

The block diagram of Fig. 2 illustrates a single-sideband receiving system that employs the exalted-carrier demodulator driving phase-shift networks that have the property of 90° differential phase shift over a wide range of audio frequencies. The operation of the system depends upon the transmission properties of the phase-shift networks to resolve the demodulator output signals into two groups, the upper- and lower-sideband responses. Fig. 3 will be helpful in understanding the action.

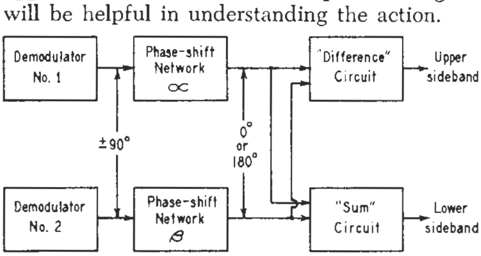


FIG. 3—The single-sideband selection is obtained by coöperative action of two demodulators and phase-shift networks.

Suppose a single incoming signal has a frequency lower than that of the synthetic-carrier oscillator. The output signals of the two demodulators are two audio tones of identical frequency and amplitude, but one signal (the one from the No. 1 demodulator, for example) leads the other by 90°. If the ∝ network has 90° more phase delay than the β network, the signals at the output terminals of these two networks are *in phase,* so that the vector sum of these two signals appears at the output of the "sum" network. If, however, the incoming signal had a frequency higher than that of the oscillator, the No. 1 demodulator output would lag the No. 2 demodulator output by 90°. The signals at the output terminals of the phase-shift networks would then be *out of phase,* and their vector sum would be zero. Thus, the sum circuit contains only signals created by incoming signals of frequency lower than that of the synthetic carrier, or lower sideband. In the same way, the "difference" circuit contains only upper-sideband signals. When both upper- and lower-sideband signals are applied to the demodulators at the same time, these actions take place independently, with the result that upper and lower sidebands are separated simultaneously. The dividing line between upper and lower sideband is the frequency of the synthetic-carrier oscillator. When this oscillator is synchronized with the carrier of an incoming a.m. signal, the sidebands thus defined coincide with the sidebands of that signal.

The sum and difference circuits are simply potentiometers which can be connected to the outputs of the phase-shift networks as shown in Fig. 4. These serve as balance controls that should be set for maximum attenuation of the unwanted sideband. The upper- and lower-sideband outputs from these sum and difference circuits can be used simultaneously to drive separate output channels if desired. A simple switching arrangement such as that indicated in Fig. 2 permits either sideband range to be used in a single channel. Positions *1* and *2* are the separate demodulated-sideband outputs, while positions *3* and *4* are demodulated double-sideband outputs. Since the signal level

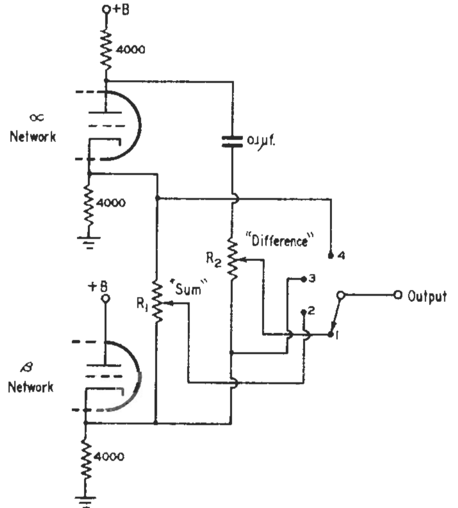


FIG. 4—The method of connecting the "sum" and "difference" circuits to the phase-shift networks. In positions *1* and *2*, the switch gives output from one sideband or the other, while positions *3* and *4* give demodulated double-sideband outputs.

at these points will be only about 1/10 volt, an amplifier is required to bring the signal to a level suitable for further use.

It is important to drive the demodulators of Fig. 2 from a source of low impedance. This is conveniently done by using a cathode follower acting as an impedance-reducing device to couple the signals from the i.f. amplifier of the receiver into the demodulators. Of course, any other means that accomplishes the same result should be equally satisfactory. Care should be taken to insure that the last i.f. stage of the receiver is not operated in such a way that distortion occurs, since distortion can cross-modulate signals and impair the otherwise good performance of the system. A diode detector left in the receiver is quite likely to cause considerable distortion. A suitable input level for the demodulators is about 0.3 volt (r.m.s.) or 1 volt, peak-to-peak.

As in the case of generation of single-sideband signals using 90° phase-shift networks, the attenuation of the undesired sideband depends in part on how nearly the phase-shift networks hold 90° phase shift over the band of audio frequencies. Similarly, there are other factors that can prevent realization of ideal operation. Distortion in the receiver ahead of the demodulators has already been mentioned as one cause for imperfect performance. Distortion occurring in amplifiers, if any, associated with the phase-shift networks, because of operation at too high a signal level, is another. Serious amounts of these effects can be avoided by the choice of operating conditions, so that the performance is not greatly poorer than the limit set by the phase-shift networks themselves.

Receiving-System Characteristics

The design of phase-shift networks permits rather good attenuation of an undesired sideband in a sideband range as great as 60 to 7000 c.p.s. The characteristics of networks of this type are such that the response of the entire system of Fig. 2 in the non-rejected sideband range is usually limited only by the bandwidth of the intermediate-frequency amplifier of the receiver used as a source of signals. An example of the type of operation that may be expected with the system described in this article is illustrated in Figs. 5A and 5B, which are plots of attenuation *versus* frequency. An overly-generous bandwidth of 12 kilocycles is assumed for the i.f. system of the receiver used as a signal source. The actual response is indicated as curve *1*, which might be measured at the output of either demodulator of Fig. 2. When the synthetic carrier is set at the center of the band as indicated, the apparent i.f. response measured at switch position *1* would have the appearance of curve *1-U*, while at

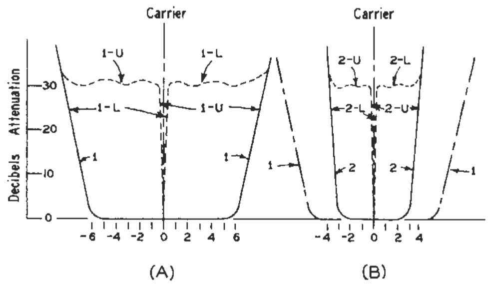


FIG. 5—Over the range for which the audio phase-shift networks hold close to 90° difference, the apparent i.f. response of the system is determined (in combination) by the i.f. and audio bandwidths. The characteristic with no audio filter is shown at (A), and (B) demonstrates how greater effective selectivity is obtained with an audio filter.

switch position *2* the response would be as indicated by curve *1-L*. This certainly is single-sideband performance, since curves *1-U* and *1-L* overlap one another only an extremely small amount near the carrier. With the i.f. bandwidth of 12 kc., each sideband is about an octave wider than is desirable for reliable phone communication. A narrower i.f. amplifier will reduce the bandwidth, but somewhat more satisfactory results can be obtained by using a 3-kc. low-pass audio filter to limit the *apparent* i.f. bandwidth to the value desired. The response characteristic obtainable in this manner is shown in Fig. 5B. The sideband-rejection performance near the carrier is unaffected, but the bandwidth is effectively limited to 3 kc. in each sideband position (curves *2-U* and *2-L*). Double-sideband reception in positions *3* and *4* doubles the apparent i.f. bandwidth for the same audio response.

» *The requirements of a good a.g.c. system for s.s.b. are a fast "attack" and a slow decay. Merely using a long time constant in the conventional a.g.c. circuit is not adequate.*

Improved A.G.C. For S.S.B. Reception

GEORGE W. LUICK, WØBFL

Having acquired a Collins mechanical filter, I set out to build an i.f. strip around it. The "hang a.g.c." [1] seemed like a good idea, but I remembered that Luther Couillard, of Collins Radio, writing in the December, 1956, issue of the *I.R.E. Proceedings*, suggested that receivers for s.s.b. and c.w. should derive their a.g.c. voltage from the audio, which would eliminate isolation problems and give extra gain for a flatter a.g.c. characteristic. As a consequence I revised the hang a.g.c. circuit for audio rectification and installed it in the new i.f. strip. It works so well that I want to pass it along. I've never seen a flatter a.g.c. characteristic on any receiver, there is no problem with b.f.o. leakage into the rectifier as there is with the i.f. type, and it is very simple to set the threshold of compression so that a product detector can be run at the level that is the optimum compromise between detector overload and available audio gain.

Those familiar with the i.f. hang-a.g.c. circuit will see that the audio-a.g.c. circuit, shown in Fig. 1, bears a family resemblance. Audio from the receiver is amplified in the a.g.c. amplifier, and rectified in the attack diode. The resultant voltage is applied to the a.g.c. line through the attack gate diode. The capacitor C_1 charges quickly and will remain charged until discharged by the recovery gate V_{1B}. This will occur some time after the signal has disappeared, because the audio was stepped up through T_1 and rectified in the recovery diode, and the resultant used to charge C_2. This voltage holds V_{1B} cut off for an appreciable time, until C_2 discharges through the 4.7-megohm resistor.

A point of difference between this and the i.f.-type circuit, other than the frequencies involved, is the use of bias on both the recovery diode and the attack diode. If bias is applied only to the attack diode, noise and such can keep the recovery gate biased to cut-off and the a.g.c. bus won't discharge. The threshold of compression is set by adjusting the bias on the diodes (changing the value of the 3.3K or 100K resistors).

Before I tried the circuit, I wondered if the attack would be as rapid as with the i.f. type, but it appears to be instantaneous. Once in a while I get a strong noise pulse that will cause the a.g.c. to hang until C_2 discharges, but most of the time the gain returns very rapidly to that set by the signal. For an S-meter circuit I use a triode and a 0-1 milliammeter in the conventional bridge, as shown in *The Radio Amateur's Handbook.* It holds so still on a steady s.s.b. c.w. signal that you would think the meter was stuck.

Some users might prefer to bypass the cathode resistor of the a.g.c. amplifier with a large electrolytic capacitor, to increase the gain of that stage. This necessitates raising the threshold bias if the audio output of the detector is to remain the same level as before. The additional gain should give a still flatter a.g.c. action, but I can notice no practical difference.

In my i.f. strip I feed manual gain-control bias to the a.g.c. bus through a diode and it works fine that way, but the a.g.c. works so well even on weak signals that I never use manual gain control.

From *QST*, October, 1957.

[1] Goodman, "Better A.V.C. for S.S.B. and Code Reception," *QST*, January, 1957.

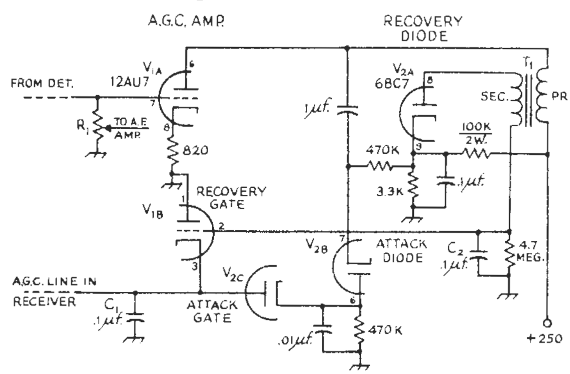


FIG. 1—Schematic diagram of the improved hang a.g.c. system. Resistors are ½-watt unless specified otherwise.
R_1—Normal audio volume control in receiver.
T_1—1:3 step-up audio transformer (Stancor A-53 or equiv.)
The hang time can be adjusted by changing the value of the recovery diode load resistor (4.7 megohms shown here). The a.g.c. line in the receiver must have no d.c. return to ground and the receiver should have good skirt selectivity for maximum effectiveness of the system.

» This is an "idea" article rather than a blow-by-blow description of construction; nevertheless, there is ample detail for the reasonably-savvy ham who might want to copy it. Besides ideas, the accent is on design and adjustment of the less familiar circuits incorporated in the receiver.

Some Ideas in a Ham-Band Receiver

PITT W. ARNOLD, W9BIY and CRAIG R. ALLEN, W9IHT

More and more hams appear to be discovering that they can build better receivers than they can buy, and for less money. But even if you have no intention of building a complete receiver, you may find a few points of interest in this receiver description. For example, if you have considered making a high-frequency lattice crystal filter for a receiver or sideband exciter, you will find some dope here on building and aligning it, and a circuit with an extra adjustment for extremely flat response in the passband. The h.f. oscillator is a good deal more stable than receiver oscillators usually are. Finally, the product detector has more than 300 times the gain of the double- or triple-triode circuits, and its linearity is at least as good.

Design of the receiver follows Goodman's philosophy [1] of keeping gain low before the "knothole" to reduce overload problems. Plug-in coils cover the amateur bands from 80 through 6 meters. The home-brew crystal filter at 4.5 Mc. gives the maximum usable selectivity for s.s.b. The a.g.c. system is very flat and works on c.w., s.s.b. and a.m. A noise limiter and a sharp c.w. filter are included in the audio circuitry.

Front End

As shown in Fig. 1, the r.f. stage uses a 6AK5, which gave better sensitivity on 6 meters than any other pentode tried. It was even superior to a cascode circuit that was used for a while. The 6AK5 is contact-potential biased to permit grounding its cathode pins directly to chassis as an aid to stability.

The mixer is one section of a 6J6, cathode-biased, driven by the other section as a cathode follower. R.f. and mixer tuning capacitors are ganged and tuned by an "R.F. Peak" control on the panel.

H. F. Oscillator

The art of making oscillators stable has made great strides in the last decade. V.f.o.'s for transmitters are much better than they used to be, largely because the Clapp and Vackar circuits have become popular. Receivers, though, continue to use the ancient and mostly inferior plate-tickler, grid-tickler, and Hartley circuits,[2] usually with a low g_m tube such as the 6C4. This seems strange, because oscillator stability is just as necessary in a receiver as in a transmitter.

This receiver uses the Vackar oscillator,[3] which has several advantages over other configurations. Like the Clapp, it is a variation of the Colpitts which steps down the tuned-circuit impedance by a capacitive voltage divider, so that variations in load or in tube capacitances are swamped by the low impedances presented to the tube. A change in heater voltage or plate voltage thus has little effect on frequency. The Vackar, unlike the Clapp, permits the oscillator cathode to be grounded to avoid 60-cycle f.m. caused by heater-cathode capacitance. Its output is more constant over a band than the Clapp's, and it does not require such a large coil on the lower-frequency bands.

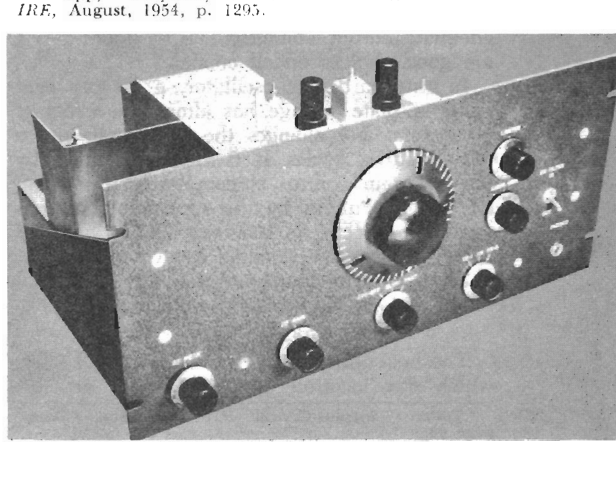
The choice of tube for a Vackar or Clapp oscillator is important. A suitable tube will have high transconductance so that the impedances presented to the tube by the tuned

[2] It would be perhaps fairer to say that these three circuits are often inferior, *in practice*, although not necessarily so in theory. It seems quite well established by now that all circuits are capable of equal stability if the same tube and operating parameters are used. However, component characteristics are generally more favorable for realization of optimum operating conditions in the case of the Clapp and Vackar.—*Editor.*

[3] Clapp, "Frequency Stable *LC* Oscillators." *Proc. IRE*, August, 1954, p. 1295.

From May, 1960, *QST*.

[1] Goodman, "What's Wrong with Our Present Receivers?," *QST*, January, 1957.



This thirteen-tube receiver covers 3.5 to 54 Mc., includes a ham-built lattice crystal filter, "hang" a.g.c., high-stability oscillator, and a novel product detector.

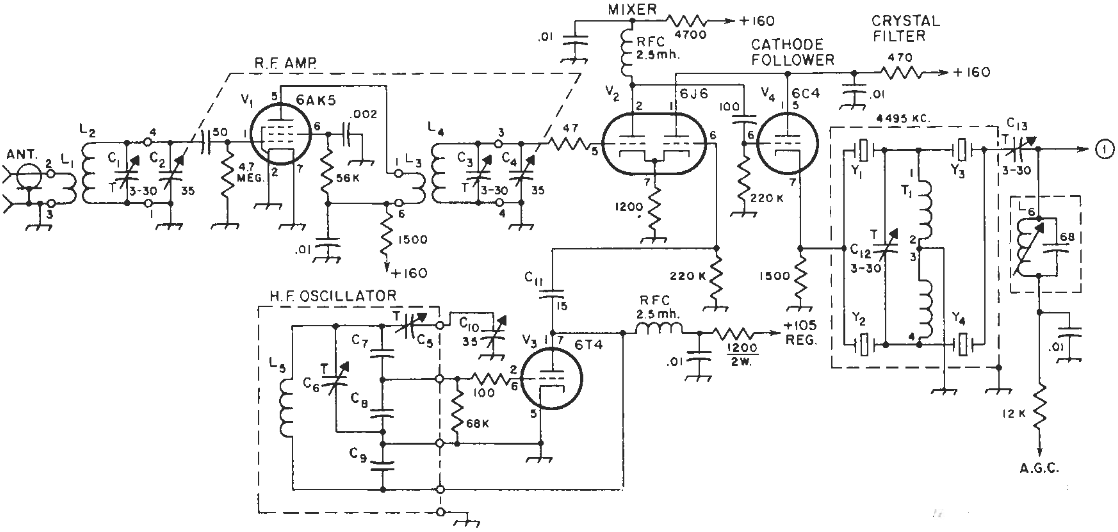


FIG. 1—Schematic diagram of the receiver. Unless indicated otherwise, resistances are in ohms, fixed resistors are ½ watt; fixed capacitors marked with polarity are electrolytic, those having values over 0.01 μf. are paper, others not listed below are disk ceramic.

C_1, C_3, C_{12}, C_{13}—3–30-pf. mica compression trimmers.

C_2, C_4, C_{10}—35-pf. double-bearing variable (Bud MC-1835).

C_5, C_9, inc.—See coil table.

C_{11}—15-pf. zero-temp. ceramic.

L_1-L_5, inc.—See coil table.

L_6, L_7, L_8—15 to 25 μh.; 45 turns No. 32 enam. close-wound at bottom of ⅜-inch slug-tuned form (CTC PLS-5), mounted in shield can (Bud SH-294).

L_9—2-hy. high-Q audio toroidal inductor (UTC HQA-13); see text.

R_1—10,000-ohm control, linear taper.

R_2—15,000-ohm control, linear taper.

R_3—0.5-megohm control, audio taper.

R_4—500-ohm control, screwdriver adjusted.

S_1—Rotary, 1 section, 2 poles, 3 positions.

S_2—S.p.s.t. toggle.

S_3—Rotary, 1 section, 1 pole, 3 positions.

T_1—Bifilar winding on ferrite toroid; see text.

T_2—Interstage audio, 2:1 or 3:1, secondary to primary.

T_3—Ouput, 10,000 ohms to voice coil (Thordarson 24S52).

Y_1-Y_6, inc.—4495-kc. FT-243 surplus crystals, etched to frequency; see text. Y_1 and Y_4 have the same frequency; Y_2 and Y_3 are 1800 cycles higher.

Note: Numbers on r.f. and mixer coil terminals are standard pin numbers on the coil forms and sockets. R.f. coils are on 4-prong forms (Amphenol 24-4P) and mixer coils are on 6-prong forms (Amphenol 24-6P). Coils for 50-Mc. band are mounted inside coil forms.

circuit can be made lower without stalling the oscillator. Interelectrode capacitances should be small so that any changes in capacitance within the tube will also be small. Finally, the amplification factor must be fairly low to ensure adequate output voltage. The 6T4 and 6AF4 are good choices on all counts.

In this particular oscillator, a 50 per cent jump in plate voltage has almost no effect at 40 meters, and changes the beat note only a few hundred cycles at 6 meters. Pulling by the r.f. gain control is completely absent on all bands. Pulling by the r.f. tuning knob is negligible except on 6 meters and is not bad enough to be objectionable even there.

The oscillator frequency is above the signal on 80 and 40 and below it on 20, 15, 10 and 6.

Crystal Filter

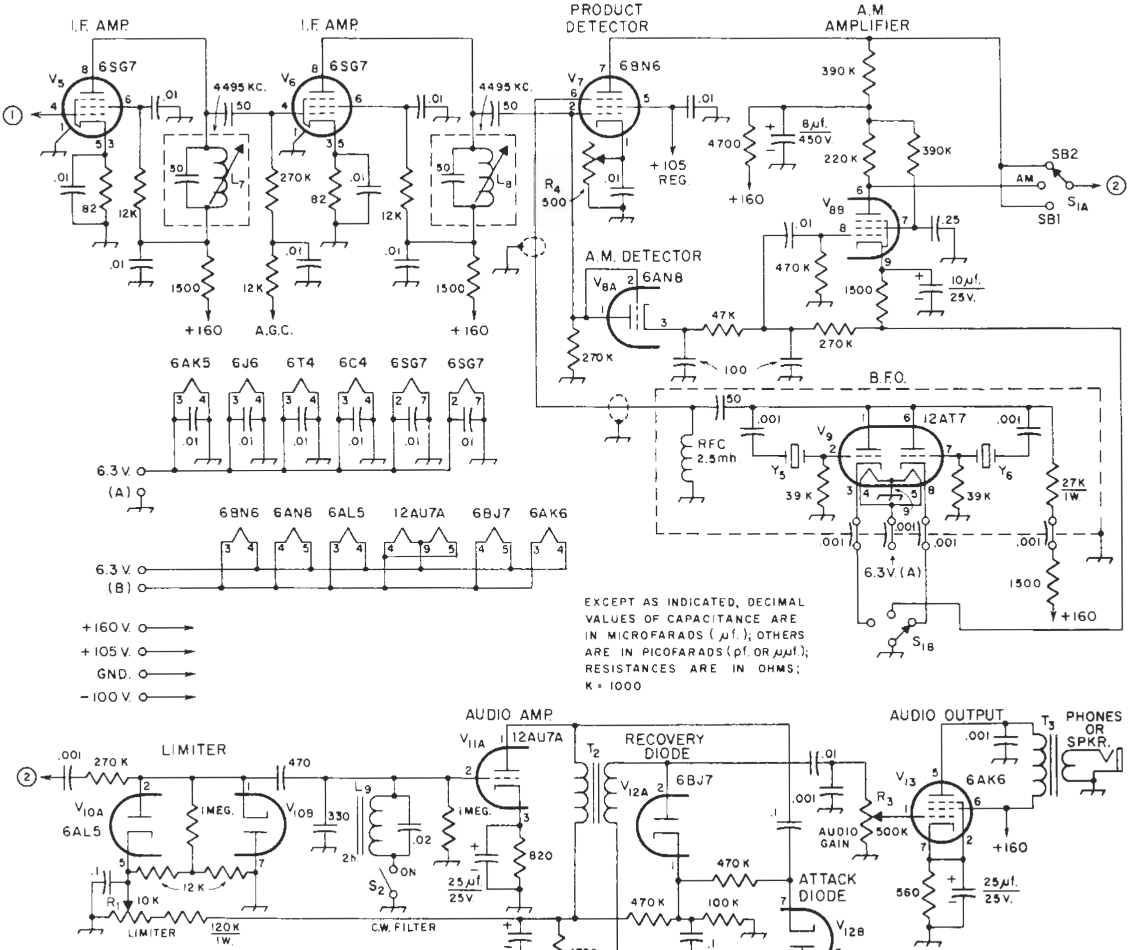
The heart of the receiver is the crystal filter, which was inspired by Ben Vester's article.[4] Its bandwidth is 2500 cycles between 6-db. points; final attenuation in the stop band is about 60 db. Insertion loss is negligible—less than a decibel.

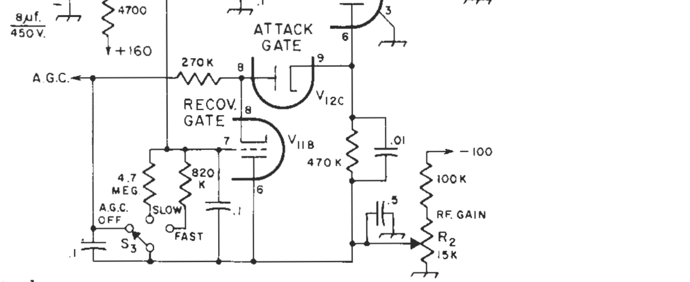
The secret of really flat passband response lies in resonating the toroid T_1 (Fig. 1) with trimmer C_{12}. Without the trimmer there was a dip of several db. in the middle of the passband. With C_{12} properly adjusted the response is flat within a few tenths of a decibel.

Building and aligning a crystal filter is really not so tough. It's a good idea to buy ten or twelve of the surplus crystals. The next requirement is some means of measuring the pole-zero spacing [5] of each crystal and checking it for spurious resonances for 50 kc. or so above the main response. Vester [4] outlines one method using a signal generator and the station receiver.

[4] Vester, "Surplus-Crystal High-Frequency Filters," page 48.

[5] A "pole" of impedance is the parallel-resonant frequency of the crystal; a "zero" is the series-resonant frequency. The zero is lower in frequency, with the pole a kilocycle or two above it.





We didn't have a stable enough signal generator, so we haywired together a little three-tube test chassis using the circuit shown in Fig. 3A. The tunable 4.5-Mc. output of the test chassis is fed to the crystal as in Fig. 4 and the v.f.o. adjusted for a peak (at the zero) or a null (at the pole) on the v.t.v.m. Since relative frequency is all we need to know, the v.f.o. is heterodyned with a crystal oscillator and the resulting audio beat measured by Lissajous figures with a scope and a calibrated audio oscillator, set up as in Fig. 3B.

Four crystals with pole-zero spacings of 1600 cycles or more and a minimum of spurious peaks should be selected for the filter. Set aside two of the remaining crystals for use in the b.f.o. The filter crystals may then be etched [6] with ammonium bifluoride solution until two of them have zero frequencies about 1800 cycles above the zeros of the other two. All four crystals should be etched high enough so that the zero of the lower pair is at least a kilocycle above the pole of the lower b.f.o. crystal. The b.f.o. crystals will be etched to exact frequency after the receiver is completed.

The filter may next be assembled in a Minibox of convenient size. In the filter assembly used in this receiver a Plexiglas plate, with holes cut in it for two octal sockets to hold the crystals, is mounted horizontally between the two long sides of the box. The number of turns on the toroid T_1 should be chosen so that it resonates at 4.5 Mc. with 20 to 25 pf. when the two sections of the bifilar winding are connected series aiding (see Fig. 5). A Q

[6] Newland, "A Safe Method for Etching Crystals," page 52.

	Oscillator Coil					R.F. Coil				Mixer Coil			
Band	L_5	C_5	C_6	C_7	C_8, C_9	L_1 (Note 4)	L_2	Diam.	Length	L_3 (Note 5)	L_4	Diam.	Length
80	31 turns B&W 3007 (Note 1)	100 APC	100 APC	25 zero-temp. ceramic	470 silver mica	5¾ turns No. 32 enam.	45¾ turns No. 32 enam.	1¼	1½	16¾ turns No. 32 enam.	45¾ turns No. 32 enam.	1¼	1½
40	22 turns B&W 3007	50 APC	50 APC	25 NPO	470 silver mica	4¾ turns No. 24 enam.	23¾ turns No. 24 enam.	1¼	1½	8¾ turns No. 24 enam.	23¾ turns No. 24 enam.	1¼	1½
20	28 turns B&W 3007	50 APC	50 APC	25 NPO	470 silver mica	3¾ turns No. 24 enam.	11¾ turns No. 24 enam.	1¼	1½	5¾ turns No. 24 enam.	11¾ turns No. 24 enam.	1¼	1½
15	12 turns B&W 3007	50 APC	50 APC	25 NPO	390 silver mica	2¾ turns No. 24 enam.	8¾ turns No. 24 enam.	1¼	1½	4¾ turns No. 24 enam.	8¾ turns No. 20 enam.	1¼	1½
10	10 turns No. 8 bare (Note 2)	50 APC	50 APC	25 NPO	300 silver mico	1¾ turns No. 24 enam.	6¾ turns No. 20 enam.	1¼	1¼	3¾ turns No. 24 enam.	6¾ turns No. 20 enam. (Note 6)	1¼	1¼
6	8 turns No. 8 bare (Note 3)	56 zero-temp. ceramic fixed	50 APC	25 NPO	150 zero-temp. ceramic	tapped 2 turns from ground end of L_2	6 turns No. 18	⅝ (B&W 3006)	⅝	6 turns No. 20 inserted in L_4	6 turns No. 18 (Note 6)	⅝ (B&W 3006)	⅝

1 B&W 3007 Miniductor coil stock, ⅝-inch diam., 16 turns per inch.

2 Length 1¼ inches. Self-supporting. Wound on ½-inch diam. form and allowed to spring out to ⅝-inch diam., center-to-center.

3 Length 1¼ inches. Self-supporting. Wound on ⅜-inch form and allowed to spring out to ½-inch diam., center-to-center.

4 L_1 interwound at cold end of L_2.

5 L_3 interwound at cold end of L_4.

6 3–30 pf. trimmers, C_3, across L_4 are omitted in the 10 meter and 6-meter mixer coils.

meter or grid-dip meter is a big help here.

Preliminary adjustment of the completed filter box is made using the setup of Fig. 6. Tuning the test chassis v.f.o. through the passband will show two peaks, at the upper and lower ends, respectively, of the passband. These peaks will not necessarily be of equal amplitude. Set the v.f.o. halfway between the peaks and adjust C_{12} for maximum reading on the v.t.v.m. or scope. Don't expect the passband response to be absolutely flat at this stage. It will look better later on when the filter has been mounted in the receiver and terminated in a properly adjusted L network.

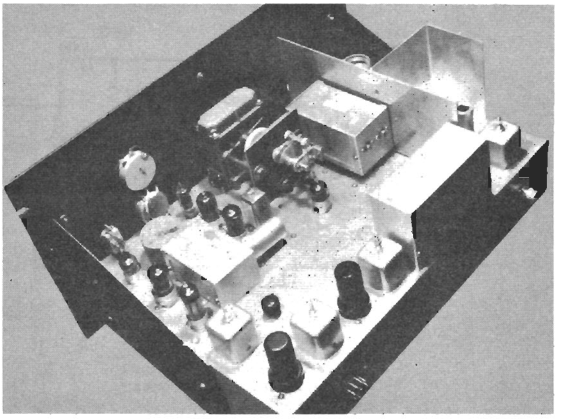
I.F. Circuits

The 6C4 cathode follower after the mixer has about the right output impedance to drive the filter, which has a characteristic impedance of approximately 500 ohms. The L network (C_{13}, L_6 and the 68-pf. capacitor) can be adjusted to terminate the filter properly for flat response.

The b.f.o. is crystal-controlled to eliminate the drift problem and ensure that b.f.o. frequency is set correctly with respect to the filter passband. The entire b.f.o., crystals and all, is built in a 3¼ × 2⅛ × 1⅝-inch Minibox, and all power leads entering the box are filtered by 0.001-μf. feed through capacitors. The output lead is made of miniature coaxial cable. These precautions proved to be necessary because a very little b.f.o. signal leaking into the i.f. circuits can block the product detector.

Product Detector

We believe that the product detector is a significant improvement over many of the circuits which have been published. It uses the 6BN6 gated-beam tube, a type originally developed



Shielding encloses the r.f. stage and mixer, along the right-hand edge of the chassis in this view. The small shield can in the far right corner has been replaced by the 6C4 cathode follower, V_4, since the photo was taken. Crystal-filter box and i.f. components occupy the rear edge of the chassis, with detectors and audio stages along the left-hand edge. The 12AT7 projects horizontally from the b.f.o. shield box. The plug-in oscillator coil box is to the right of the main tuning capacitor.

for service as limiter and phase detector in f.m. receivers. The signal grid of a good product detector must be very linear so that there is no intermodulation among components of the signal. A glance at the 6BN6 curves show that grid 1 is almost perfectly linear over a range of 2 volts peak-to-peak (0.7 volts r.m.s.), while outside this range the tube limits sharply. Grid 3 has similar characteristics except that its gain is lower.

Tests have shown that the linearity of the 6BN6 as a product detector is excellent. At 0.3 volt r.m.s. input to grid 1, the modulation recovered from a 50-percent modulated signal, measured with b.f.o. off, was 40 db. below the normal beat note obtained with the b.f.o. on. At an input of 0.7 volt the distortion products were still 35 db. down. Above 0.7 volt grid 1 was driven into the limiting region and distortion increased rapidly. Signal input in this receiver is 50 to 100 millivolts, well below the limiting threshold.

With 3 or 4 volts of b.f.o. injection on grid 3, the 6BN6 has a conversion gain of 50—that is, 100 millivolts of i.f. signal at grid 1 produces 5 volts of audio at the plate. By contrast, a 12AU7 in the double-triode detector circuit showed a conversion gain of 0.15 with similar input levels. Noise peaks, incidentally, are clipped by the 6BN6, leaving less work for the regular noise limiter.

The 6BN6 has one drawback—it is slightly microphonic. Trouble from this source can be avoided by mounting the tube socket on a

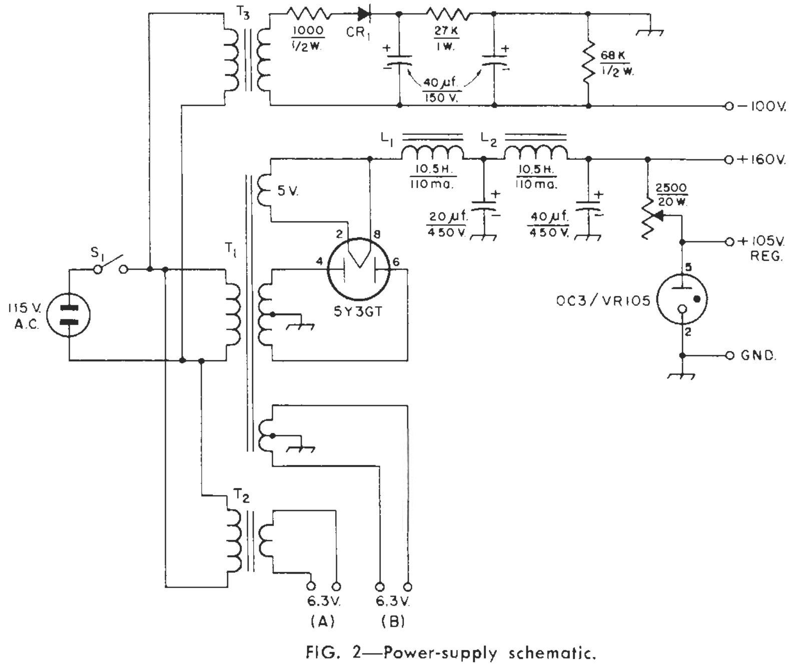


FIG. 2—Power-supply schematic.

CR_1—Silicon rectifier, 130 volts r.m.s., 150 ma. (Sarkes-Tarzian M150).

L_1, L_2—10.5 hy., 110 ma. (Stancor C-1001).

S_1—S.p.s.t. toggle.

T_1—Power, 540 volts c.t., 120 ma.; 5 volts, 3 amp.; 6.3 volts, 3.5 amp. (Stancor PC-8405).

T_2—Filament, 6.3 volts, 3 amp. (Thordarson 21F10).

T_3—Power, 117 volts, 20 ma. (Thordarson 26R32) heater winding not used.

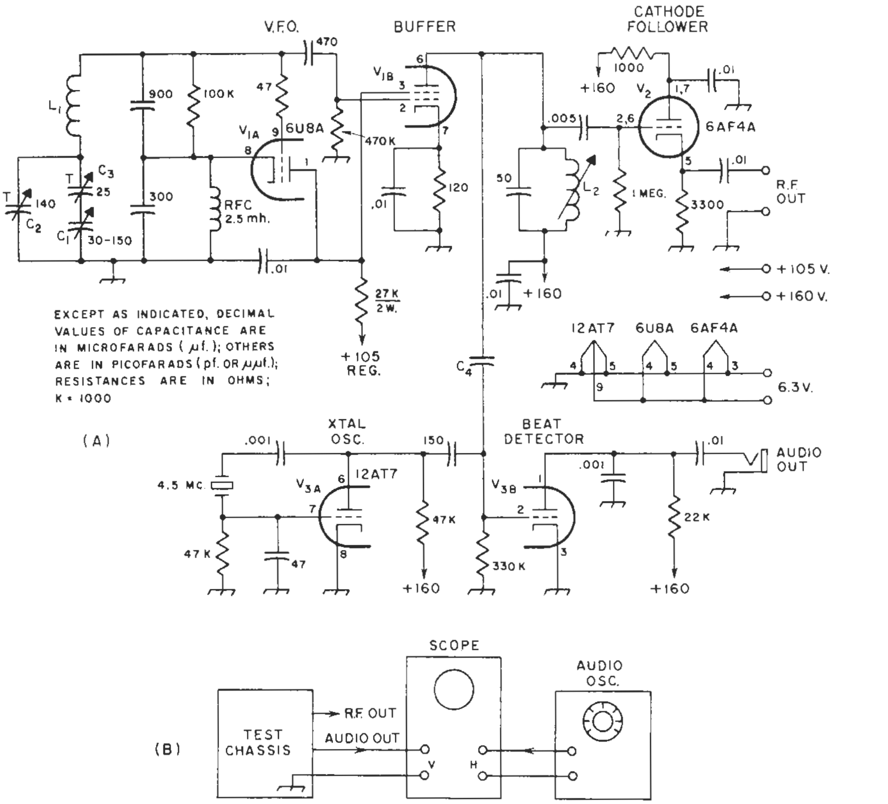


FIG. 3—(A) Circuit of test chassis. (B) Setup for measuring relative frequency of v.f.o. Resistances are in ohms; resistors are ½ watt. Fixed capacitors are ceramic.

C_1—30–150-pf. variable with worm drive, taken from ARC-5 transmitter.
C_2—140-pf. air trimmer (Hammarlund APC-140).
C_3—25-pf. air trimmer (Hammarlund APC-25).
C_4—Two insulated wires twisted together for an inch.
L_1—27 turns No. 20 enam. close-wound on 1-inch diam. form, 1 inch long.
L_2—Same as L_6 in Fig. 1.

small metal plate and bolting the plate to the chassis through rubber grommets.

Detection of a.m. signals is accomplished by an ordinary diode using one section of a 6AN8. The pentode half of the 6AN8 supplies enough gain following the diode so that one can switch from s.s.b. to a.m. without readjusting the audio gain control.

The 6AL5 noise limiter is a double-ended shunt type. It is located ahead of the a.g.c. circuits to keep noise pulses from operating the a.g.c. It also precedes the c.w. filter so that the filter will not ring on noise peaks. A shunt limiter does not clip quite as sharply as the series-diode type, but neither does it distort signals below its limiting threshold; a series limiter produces a noticeable amount of distortion on all signals.

A simple audio filter is a good way to get c.w. selectivity in a receiver of this kind, since the crystal filter knocks out the audio image, or "other side of zero beat." The tuned circuit made up of L_9 and the 0.02-μf. capacitor resonates at 850 cycles. The coil specified for L_9 has a Q of almost 100 and tunes as sharply as anyone could want. In fact, it rings a little on signals; many operators might prefer a bit less Q. There are toroids available from Arrow Sales [7] with a Q of about 24, at prices considerably lower than the eleven-dollar tag on the UTC HQA-13.

A.G.C.

The "hang" a.g.c. system was taken from WØBFL's article [8] with minor modifications. The 270K resistor in series with the a.g.c. line slows down attack time enough to prevent noise peaks from operating the a.g.c. A very slight "burst" can be noticed now on the first syllable of a transmission, but it is not bothersome at all. A choice of two recovery time constants is provided. The "fast" position is occasionally useful on rapidly fading signals, but the "slow" position is used most of the time. Delay bias on the attack and recovery

[7] Arrow Sales, Inc., 2534 So. Michigan Ave., Chicago 16, Ill., and 7035 Laurel Canyon Blvd., North Hollywood, Calif.

[8] Luick, "Improved A.V.C. for Sideband and C.W.," page 180.

diodes (determined by the 470K-100K divider) is set so that the i.f. signal at the detectors is about 50 to 100 millivolts, as already noted.

The principal change from Luick's a.g.c. circuit is the method of applying manual r.f. gain control. The r.f. gain knob controls a variable negative bias which is fed to the a.g.c. line in such a way that it controls receiver gain and at the same time acts as additional delay bias on the a.g.c. diodes. Thus the r.f. gain knob can be set to prevent background signals and noise from booming in during pauses, while full a.v.c. remains available to handle normal fading.

Shunt capacitors in the audio circuits are chosen so that high-frequency response drops off above 2500 cycles to reduce fatigue from high-pitched hiss. Low frequencies are cut below 300 cycles to restore balance on voice signals. The resulting audio quality is crisp and intelligible.

Construction

The receiver is built on a 12×17×3-inch aluminum chassis with an 8¾-inch aluminum rack panel. The top-view photo shows the layout.

The oscillator tuning capacitor, C_{10}, is driven by a National NPW-0 dial and gear unit, through an insulated coupling. The capacitor is mounted on a ⅛-inch sheet of mica-filled bakelite, and its mounting feet are bolted to an aluminum L-bracket which is fastened to the bakelite sheet. This arrangement provides two-point support to prevent the stator from twisting. The bakelite sheet is held away from the gear box by three metal spacers and 12-24 threaded rods. The only electrical ground on the rotor of C_{10} is a heavy wire lead passing through a hole in the chassis and connected to a solder lug on the underside. Thus the circulating current through C_{10} has a single definite path so it can't wander all over the chassis looking for a route to the under surface.

The oscillator coil and associated capacitors for each band are assembled in a 4×2¼×2¼-inch Minibox (see close-up photo). A piece of mica-filled bakelite in the bottom of the box supports a row of four banana plugs which project through a rectangular cutout in the 4×2¼-inch surface of the box section. A fifth banana plug grounds the shield box to the chassis.

A small copper shield is soldered across the 6BN6 socket to isolate the signal grid (pin 2) from the b.f.o. injection grid (pin 6). All power wiring is done with shielded wire to eliminate one source of feedback.

The power supply is built on a separate

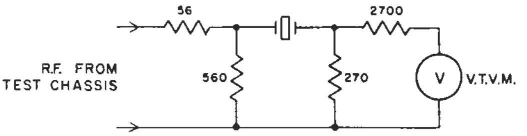


FIG. 4—Isolating network used for measuring crystal pole-zero spacing and spurious resonances. The crystal is plugged into an octal socket and the remaining socket contrcts used as tie points for the ½-watt resistors. Indicator can be a v.t.v.m. with r.f. probe or a wide-band scope.

5×10×3-inch chassis. Its schematic is shown in Fig. 2.

Alignment

Alignment of the front end is easy, since the receiver is not gang-tuned. The i.f. stages and the L network terminating the crystal filter can be aligned with the aid of the test chassis of Fig. 3. Pull out the 6AK5 and wrap a wire from the test chassis r.f. output around the mixer coil. Set the sideband switch to the a.m. position, a.g.c. off, and connect a d.c. v.t.v.m.

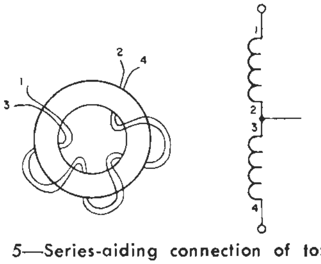


FIG. 5—Series-aiding connection of toroid T_1.

or high-resistance voltmeter across the 270K load resistor in the cathode of V_{8A}. Pull out the 6BN6 to avoid loading by its grid current. Tune the test chassis v.f.o. within the filter passband, set C_{13} near maximum capacitance, and peak L_6, L_7 and L_8. Now tune the v.f.o. carefully through the passband and observe the flatness of the filter response. Adjust C_{13} to a slightly different value and repeak L_6. Repeat this process until the passband response is as flat as possible. Set the v.f.o. to the exact center of the passband and recheck C_{12} for maximum signal. The filter response should be flat within about 5 per cent. Reinsert the 6BN6 and repeak L_8 with the r.f. gain control

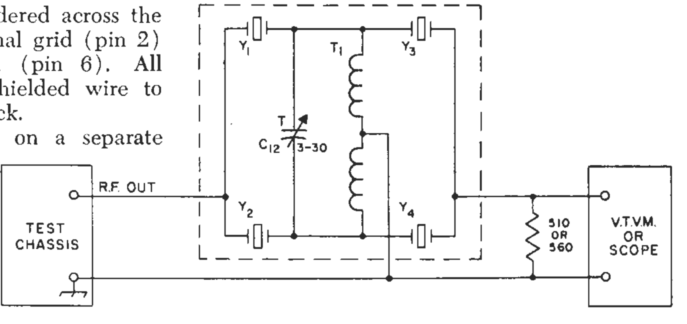
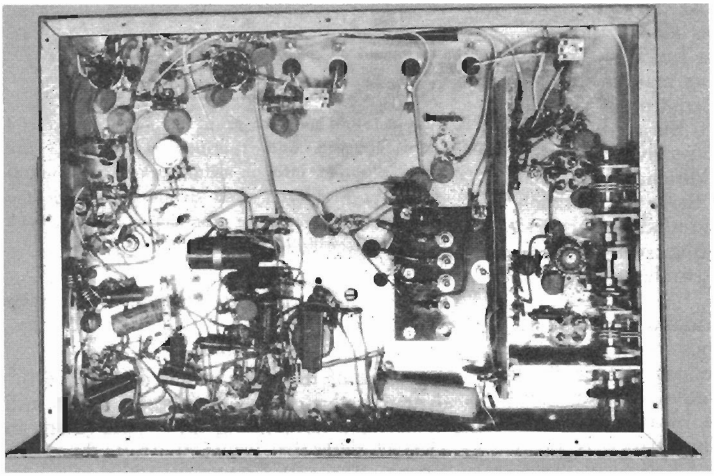


FIG. 6—Setup for preliminary adjustment of C_{12}. Indicator can be a v.t.v.m. with r.f. probe or a wide-band scope.



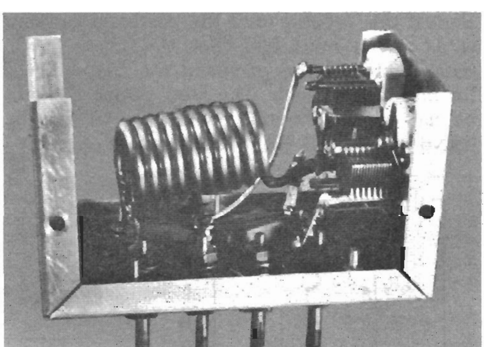
The r.f.-section tuning capacitors are along the left-hand edge of the chassis in this view of the bottom. Leads to the crystal filter go through the holes along the lower center edge of the chassis.

set to give the smallest observable deflection on the voltmeter.

When the filter alignment is complete, the b.f.o. crystals may be etched to frequency. Set the test chassis v.f.o. about 10 db. down one skirt of the filter response curve (voltage one-third of maximum). Etch the b.f.o. crystal until it is in zero-beat with the v.f.o. Do the same thing with the other b.f.o. crystal on the other filter skirt.

To check for spurious filter responses, restore the receiver to normal operation and tune in a strong modulated signal on a dead band, such as 10 meters in the evening. Tune the main dial through about 50 kc., tuning above the signal on 80 or 40 or below the signal on 20, 15, 10 or 6. Listen carefully to see if the signal appears at another dial setting. If so, the spurious response can sometimes be reduced by interchanging Y_1 and Y_4. It is then necessary to readjust C_{12}.

The product detector is adjusted for best linearity by tuning in a modulated signal and setting the r.f. gain to give about 15 volts of audio at the 6BN6 plate with b.f.o. on. A modulated signal generator is best, but a voice signal will do. Disable the b.f.o. by pulling out the 12AT7 and adjust the 500-ohm resistor, R_1, in the 6BN6 cathode for minimum recovered audio. There should be a sharp null near mid-range on the resistor. If the null is broad, try changing the r.f. gain till you find a definite setting of the 500-ohm resistor where the signal almost disappears. A setting near maximum or near zero resistance is not correct; the tube is in the limiting region here.



Interior of the 10-meter oscillator coil box. Layout of components in the other coil boxes is similar.

Toroids

The only unusual item in the parts list is the ferrite toroid T_1. A readily-available source of toroids, suggested by Brian Voth, W9ARZ, is the hollow ferrite core of a b.c. set antenna made by Grayburne and sold under the name Superex Ferri-Loopstick, net price 44 cents. A machinist who owns a small diamond cutting wheel can slice off a few toroids for you. It is also possible, with a little luck, to break the ferrite like a piece of glass tubing after filing a notch in it. The break may require smoothing with a file or grinder. No. 34 or 36 enamel wire is about right for the winding. The Ferri-Loopstick toroid takes about 9 bifilar turns to resonate with the 3–30-pf. trimmer.

Apparently almost any kind of ferrite will do for the toroid. The transformer is connected at a low-impedance point, so losses have little effect. Some rather high-loss material has been used experimentally with good results.

The completed receiver has proved very satisfying to operate on the crowded bands. A.g.c. is remarkably flat and works equally well on sideband, a.m. or c.w. Noise figure has not been measured, but it appears to be as low as necessary even on 6 meters.

» Receivers that lack real s.s.b. selectivity can be converted to selectable-sideband reception by the addition of a "Signal Slicer" such as is described here. It uses the phasing method for rejecting the unwanted sideband.

An S.S.B. Product-Detector Adapter

CARL F. BUHRER, K2OHF

The proper demodulation of a sideband signal can be accomplished by combining it with an excess of carrier injection voltage in a diode detector or by means of a product detector. A product detector alone, however, will not reject signals on the undesired side of the injected carrier; these must be eliminated either by a selective i.f. strip or by means of a suitable phasing system and double product-detector combination. The adapter described plugs into the receiver in place of a 6H6 a.m. detector. A minor modification of the circuitry was necessary that does not affect a.m. operation.

Circuit Description

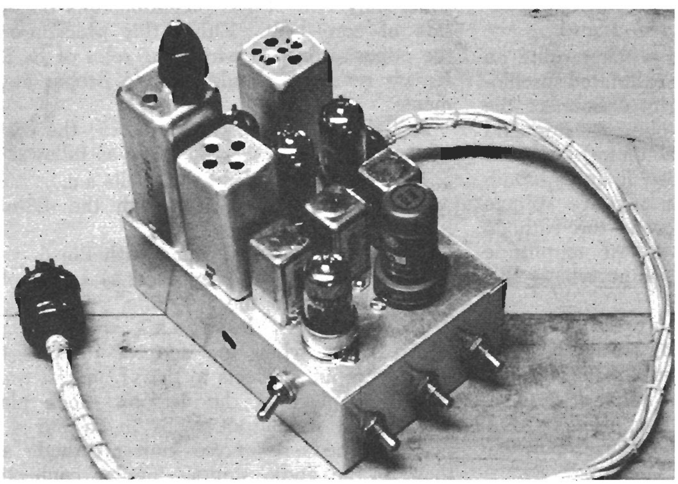
Two type 7360 tubes are used as product detectors, as shown in Fig. 1. Their design and application have been previously described.[1,2] The 6AG5 injection oscillator is used to produce two 10-volt peak-to-peak 455-kc. voltages on the grid of each of the 7360s. The resistor and capacitor across the secondary of T_2, with resistance and reactance equal, provide equal-amplitude injection signals 90 degrees out of phase. The deflecting electrodes are fed with push-pull i.f. signals biased at +25 volts. The audio output from the plates is reduced by means of T_3 and T_4 to a lower impedance necessary to feed the phase-splitting potentiometers, R_1 and R_2, at the inputs of the phase-shift networks. Here the audio signals, which were 90 degrees out of phase, are given a 90-degree relative phase shift so they are now either in phase or 180 degrees out of phase. Those in phase cancel in T_5, while the others add to give useful output. The phase relations are such that only one sideband is heard. Switch S_1 determines the sideband selection, the center off position resulting in double-sideband response. The low-pass filter at the output restricts the audio range to where the networks perform well but the filter could probably be omitted.

From August, 1961, *QST*.

[1] Vance, "S.S.B. Circuits Using the 7360," page 29.
[2] Filipczak, "The 7360 as a Product Detector," page 36.

Construction

The adapter was built on a 2×4×6-inch aluminum chassis (which is a bit too small), and is connected to the receiver by a cable with octal connectors. Most of the parts are standard, but some are modified or surplus. The 455-kc. transformers are of the capacitor-tuned variety. In T_2, the tuning capacitor and half of the secondary winding were removed. The small coil was then slid to within ¼ inch of the primary winding. The three audio transformers were obtained in surplus. Each has three center-tapped windings of 22,000 ohms, 5200 ohms, and 600 ohms, but any center-tapped 20,000-ohm to 500-ohm (or 600-ohm)



Plan view of the productor-detector adapter. The i.f. transformers are at the left-hand end of the chassis, the three audio transformers at approximate center and the phase-shift unit to the right. The b.f.o. unit at the extreme left was eliminated later in favor of a crystal oscillator.

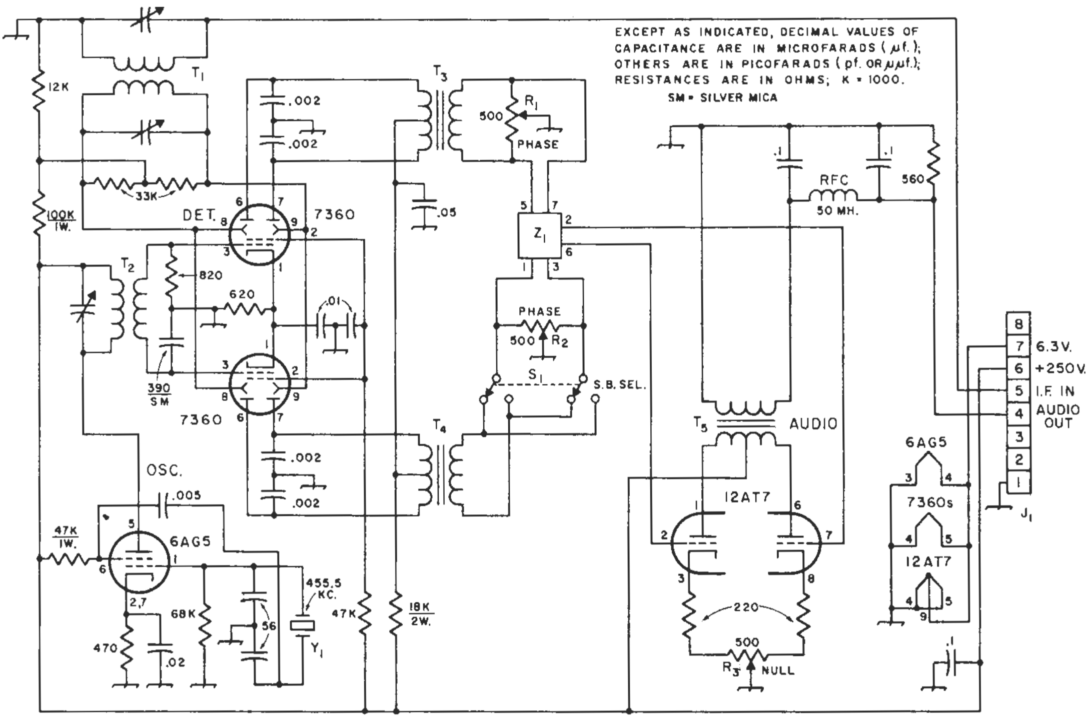


FIG. 1—Circuit of the s.s.b. adapter. Resistances are in ohms and resistors are ½ watt unless indicated otherwise.

J_1—Octal plug.
R_1, R_2, R_3—500-ohm control, linear taper.
S_1—D.p.d.t. center-off toggle switch.
T_1, T_2—455-kc. interstage i.f. transformer (Miller 112-C2). See text for modification of T_2.
T_3, T_4, T_5—Audio transformer: 20,000 ohms center-tapped to 500 ohms (see text).
Z_1—Phase-shift network (B & W 2Q4).
Y_1—Surplus crystal, 455.55 kc. (FT-241, Channel 46).

audio transformer should be satisfactory provided that T_3 and T_4 are identical.[3]

Alignment and Operation

After connecting the detector to a power source, check to see that the oscillator is working. Set T_2 such that 10 volts peak-to-peak appears on each 7360 No. 1 grid.

A possible way to do this without using an r.f. probe or a wide-band calibrated oscilloscope would be to temporarily disconnect the grounded side of the 820-ohm resistor and insert a v.t.v.m. (or 0-1 milliammeter) bypassed by a 0.1-μf. capacitor between it and ground. Grid current will produce a negative voltage reading if the peak r.f. exceeds the cathode bias of about 5 volts, and the r.f. amplitude should be held below the point where this occurs.

Move the receiver tuning upward to give a 1000-cycle beat note and using an a.c. v.t.v.m., adjust R_1 so that the voltages on Pin 7 and Pin 5 of the 2Q4 are in the ratio of 2 to 7. Do the same for Pin 3 and Pin 1, using R_2. Set R_3 to its center, flip S_1 to whichever position rejects the audio best, and adjust R_2 and R_3 to null out any remaining 1000-cycle tone, disregarding audio harmonics. Now tune the receiver down to 1000 cycles below zero beat, throw S_1 to the opposite side, and adjust R_1 and R_3 for a null. Repeat this several times, using only two of the potentiometers on each side of zero beat. Either this procedure or the opposite one, in which the roles of R_1 and R_2 are reversed, should lead to proper alignment.

In the receiver here (an NC-125) the wiring of the 6H6 socket was modified as follows:

Pin 7 was connected to 6.3 volts a.c.

Pin 2 was grounded through the 4.3-ohm heater dropping resistor.

Pin 1 was grounded along with Pin 8.

Pin 6, which had been used as a tie point, was connected to the B+ line.

These changes leave a.m. operation unaffected, but when the adapter replaces the 6H6, its input connects to the i.f. output transformer in place of the detector diode plate, Pin 5, and its output feeds in to Pin 4 in place of the series noise limiter cathode and then to the audio stages. Similar connection should be possible with many other receivers.

[3] The Lafayette AR-151 is suitable; use only half the secondary for the low-impedance output.—*Editor.*

» *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

Based on "Looking at Phone Signals," December, 1962, *QST*, and "Checking Signal Quality With the Receiver," March, 1963, *QST*.

[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.

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 overmodulation 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.

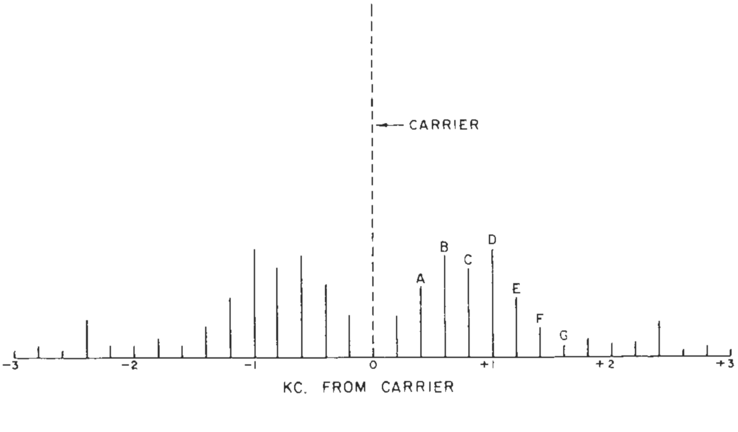
Analyzing the Process

If you aren't wholly familiar with receiver operation a diagram of this process may help. Fig. 1 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. 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.



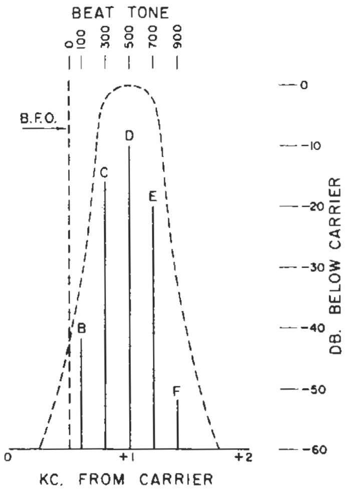


FIG. 2—With high selectivity, only those sideband frequency components to which the receiver is actually tuned will give appreciable response. This drawing shows the relative response a selective receiver would give on the lettered components in Fig. 1. The scale at the top shows the beat tone each component 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 pick-up. 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.

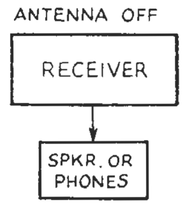


FIG. 3 — Setup for using the station receiver for transmitter checking.

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 appreciable 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.

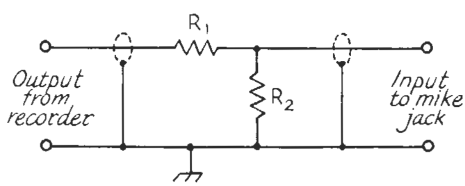


FIG. 4—Simple voltage divider for reducing audio voltage to a manageable level for the transmitter's speech amplifier. Ordinarily R_1 will have about ten times the resistance of R_2. A variable control having the same over-all resistance can be substituted for the two resistors.

[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.

» Before an oscilloscope can give you the information you want, it has to be hooked up right. This article discusses typical setups, including a simple r.f. coupling device suitable for any power level.

Oscilloscope Setups

GEORGE GRAMMER, W1DF

The primary necessity in using a scope for transmitter checking is to put a sample of the r.f. output of the transmitter directly on one set of c.r. tube deflection plates. Customarily the vertical plates are used.

By "directly on the plates" we mean either an actual direct connection to the plates, or a connection through blocking capacitors to isolate any d.c. voltages that may be in the circuits. It is rarely possible to use a scope's built-in vertical amplifier, except possibly at relatively low radio frequencies.

Unfortunately, not all of the low-cost scope kits have provisions for direct deflection-plate connections. If your scope does not have such provisions it will be necessary to modify it. The details of such a modification will vary with the actual scope circuit. In principle, it is quite simple. The deflection plates are merely disconnected from the vertical amplifier and brought out externally. The catch is that this must be done without sacrificing the spot-centering control.

Fig. 1A is more-or-less representative of the way in which the centering bias is applied to the deflection plates in the regular amplifier setup. In some circuits a parallel feed arrangement may be used, the centering voltage being fed to the deflection plates through high resistances. In such cases blocking capacitors will be placed between the amplifier-tube plates and the deflection plates.

Fig. 1B shows how the series-feed circuit of Fig. 1A can be modified. Additional resistors, R_1 and R_2, are added to isolate the deflection plates from the scope circuit so the signal input will see a high impedance. The ability to use the centering control is retained because the high resistances carry no current, and thus the centering bias continues to be applied to the deflection plates, through the dashed-line connection indicated between the two pairs of terminals. The original circuit can be restored by shifting the dashed jumpers to tie the c.r.t. to the amplifier. To isolate the source of input signal from the direct voltage on the deflection plates the signal is introduced through the blocking capacitors C_1 and C_2.

In practical scopes the circuit-changing connections frequently are made through wire jumpers between screw terminals on an insulating board. This is somewhat preferable to using a switch, since the shunt capacitance is lower. The mechanical details of mounting and wiring have to be worked out for the individual scope, because different models vary in internal construction. The principal point to be observed is that the leads to the deflection plates should be as short and as well separated as possible from other wiring. This helps to keep

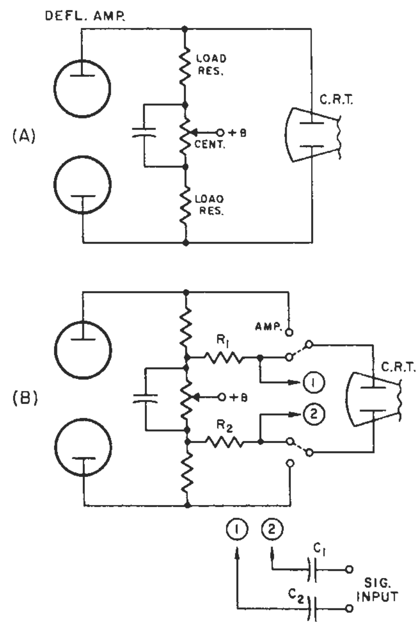


FIG. 1—Oscilloscope modification for making r.f. connections to deflection plates without going through internal amplifiers. A—Representative amplifier-c.r.t. coupling. B—Modified for alternative connection to external source. R_1 and R_2 should be 1 megohm or more, 1-watt rating. For vertical r.f. deflection C_1 and C_2 should be mica or ceramic, voltage rating according to plate voltage on vertical amplifiers; capacitance may be 500 pf. or more. For audio input to horizontal plates, C_1 and C_2 should be 0.1-μf. paper, 600 volts.

From "Oscilloscope Setups for Transmitter Testing," October, 1964, *QST*.

r.f. from straying into other parts of the scope circuit.

If the centering circuit in the scope uses parallel feed, R_1 and R_2 are already built in. In that case, simply add the terminal board as shown in Fig. 1B, leaving the equivalents of R_1 and R_2 connected to the deflection plates. Connect the amplifier's blocking capacitors to the "Amp." terminals.

C_1 and C_2 can be of the order of 0.001 μf. for r.f. work. The value is not at all critical.

Coupling the Transmitter to the Scope

With the scope itself in readiness for direct introduction of r.f. to the deflection plates, some method of sampling the r.f. output of the transmitter is the next order of business. Only a microscopic amount of r.f. power is needed for scope deflection—mainly, to supply whatever losses there may be in the system used for transferring the signal from the transmitter to the scope. However, a fairly large voltage—of the order of 50 to 100 volts r.m.s. —will be needed. The voltage required depends on the deflection sensitivity of the oscilloscope tube actually used, but in practically any case about 100 volts r.m.s. will be sufficient.

Getting such a voltage is not difficult, even from transmitters of very low power. The best way is to use a tuned circuit loosely coupled to the transmitter's output circuit. Besides providing ample deflection voltage, this method also eliminates transmitter harmonics from the deflection voltage and offers a very convenient means for adjusting the pattern height.

Fig. 2 is a useful circuit for this purpose. The "pick-up unit" is just a pair of coax sockets mounted in a small Minibox, with a straight-through connection having a 1-inch loop at its center. To this is coupled a 2-turn link of insulated wire that connects to a section of small coaxial line of any convenient length. The other end of the line connects to a similar link tightly coupled to an LC circuit

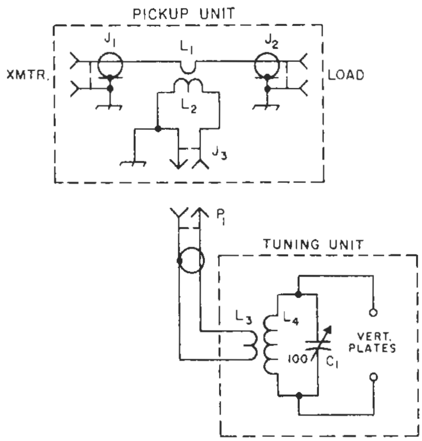


FIG. 2—R.f. pickup and tuning units for vertical plates of oscilloscope. See text for discussion of circuit constants.

C_1—100-pf. variable.

J_1, J_2—Coaxial receptacle, chassis-mounting.

J_3—Phono connector.

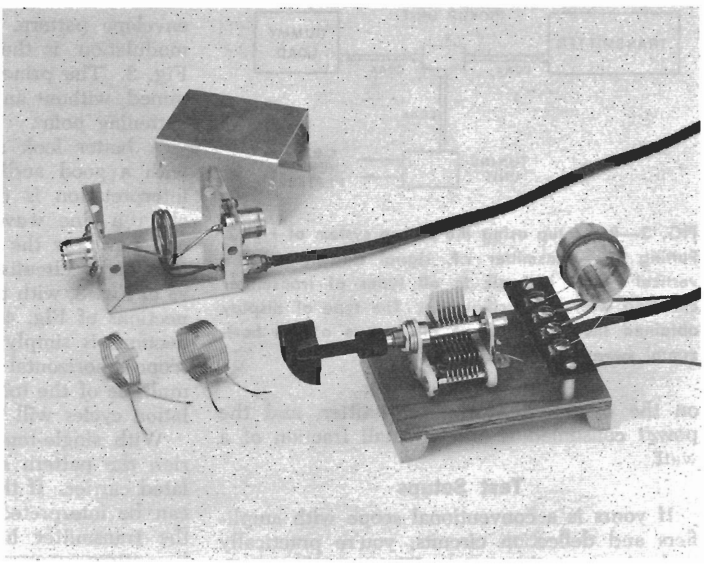
L_1—1 turn No. 14, 1-inch diameter (see photo).

L_2—2 turns insulated wire tightly coupled to L_1.

L_3—2 turns insulated wire, 1 1/8 inch inside dia.

L_4—3.5-7 Mc.: 32 turns No. 24, diameter 1 inch, length 1 inch (B & W Miniductor 3016).
14-21 Mc.: 8 turns No. 20, 1-inch dia., 16 turns/inch (Miniductor 3015).
21-28 Mc.: 4 turns No. 20, 1-inch dia., 16 turns/inch (Miniductor 3015).

P_1—Phono plug.



R.f. of adjustable amplitude can be obtained easily and inexpensively from the output of the transmitter, using simple circuits of the type shown in this photograph. The No. 14 wire joining the two coaxial connectors is formed into a loop to which a link is coupled for transferring r.f. to the tuned circuit on the separate board. The tuned circuit can sit on the oscilloscope near the deflection plate connections.

that is capable of being tuned to the operating frequency. Suggested constants are given in Fig. 2, and a simple type of construction is shown in the photograph. The tuned circuit here is mounted on a 2×3 inch piece of plywood, which can be set on top of the scope with short leads to the scope input terminals. To reduce stray capacitance to the scope case the wood base is elevated a little by wooden spacers. Since the tuning capacitor is single-ended, an insulating extension shaft is used for the control knob in order to reduce hand-capacitance effects. A refinement would be to use a split-stator capacitor, which would avoid hand capacitance. However, the tuning range will be narrowed, unless a physically large capacitor is used. With the 100-pf. capacitor shown it is possible to cover two adjacent bands with a single coil.

With the constants given, 100 watts in a 50-ohm matched load will easily develop a pattern that goes off the top and bottom of the c.r. tube screen. An output of as little as one watt will give a *usable* pattern height. For high power one-turn links at each end of the line connecting the pickup and tuning units should suffice. In fact, the circuit values are not at all critical. The principal requirement is that L_4C_1 tune through the operating frequency. A high-Q coil at L_4 is desirable for maximum sensitivity. The small Miniductor type coil material is quite satisfactory.

For transmitter testing this pickup assembly should be connected between the transmitter and a dummy load as shown in Fig. 3. The pickup unit can be left in the line when the antenna system is substituted for the dummy, for continuous scope monitoring. The small loop in the pickup unit will have no effect

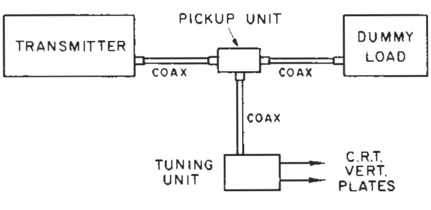


FIG. 3—R.f. setup using the pickup system of Fig. 2. Putting the transmitter r.f. signal directly on the vertical plates is basic to all types of transmitter checking with the oscilloscope. The type of display obtained then depends on the nature of the horizontal sweep signal.

on the operation of the transmitter, and the power consumed is only a small fraction of a watt.

Test Setups

If yours is a conventional scope with amplifiers and deflection circuits, you're practically in business at this point. Wave-envelope patterns of all types can be inspected with no further equipment. For looking at voice-modulated patterns, simply give the tube a horizontal sweep of some kind—either a.c. or a low-frequency linear sweep—and inspect the pattern. The main thing is that the horizontal sweep frequency should be low compared with the lowest audio frequency in the modulation. If it is too high, the pattern will be displayed just a bit at a time spread out over the entire sweep width, giving the impression of a jumble of traces sweeping across the tube at various heights. Such a pattern is meaningless.

S.S.B. Patterns

Most sideband transmitters lend themselves well to checking only by the use of the wave-

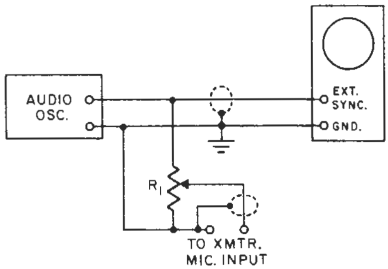


FIG. 4—Audio oscillator to scope connections for synchronizing a linear sweep with single-tone modulation. R_1 is for independent adjustment of the audio level at the transmitter's microphone input jack. An audio-taper control having a value of 5000 or 10,000 ohms will be satisfactory for use with audio oscillators having moderately-low output impedance. The control may not be necessary in all cases. Shielded wiring is desirable to prevent hum pickup. The resistor may also need to be shielded if hum shows up in the transmitter's output.

envelope pattern. All this requires, for voice modulation, is the scope and the r.f. setup of Fig. 3. The principal thing that can be determined, without an audio oscillator, is the peak-flattening point.

A better look at linearity can be obtained with a good audio oscillator although critical interpretation is difficult; it depends on how clean a sine wave the audio generator gives and how low the distortion is in the transmitter's audio circuits. The r.f. setup is the same as in Fig. 3, with the addition of the audio connections of Fig. 4. The purpose of these connections is simply to permit synchronizing the scope's horizontal sweep circuit at some submultiple of the tone frequency so a few modulation cycles will be displayed.

With single-tone input and suppressed carrier, the pattern should resemble an unmodulated carrier. If there is ripple on the carrier it can be interpreted as described elsewhere.[1] If the transmitter has provision for inserting a

[1] "Sideband Scope Patterns," page 200.

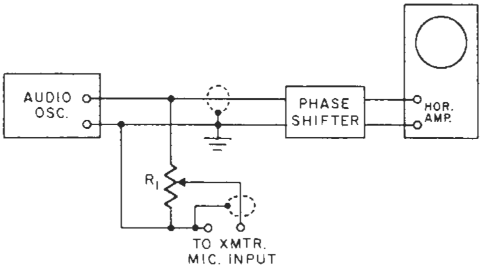


FIG. 5—Audio connections for generating a bow-tie pattern with phasing-type s.s.b. transmitters capable of producing a double-sideband suppressed-carrier signal from single-tone modulation. Except for the addition of the phase-adjustment circuit (Fig. 5) the setup is the same as in Fig. 6 and the some considerations hold.

fairly husky carrier, as most do, the two-tone pattern can be obtained. To get it, watch the pattern while increasing the carrier insertion, after having set the single-tone s.s.b. pattern to about half the maximum height that the transmitter is capable of producing by adjustment of the audio gain. As the carrier insertion is increased, the modulation pattern will form. Eventually, when the carrier and sideband have equal amplitudes, the typical two-tone pattern of Fig. 2D in the referenced article will be displayed. If there is flattening at the tops of the waveform the transmitter is being overdriven; the audio gain should be reduced and the carrier insertion reduced to correspond, until the flattening disappears. The transmitter is then being driven to its maximum linear level. Linearity throughout the modulation cycle can be estimated by observing how closely the waveform resembles half sine waves on each side of the horizontal axis.

The bow-tie pattern can be obtained with phasing-type transmitters that are capable of transmitting a double-sideband suppressed-carrier signal, but not with a filter-type transmitter that has no provision for bypassing the filter. For a phasing transmitter, use Fig. 3 in conjunction with the audio connections of Fig. 4, and set the transmitter controls for double sideband with suppressed carrier. The phase shifter circuit of Fig. 5 may be necessary. It should be adjusted to give clean line edges, rather than ellipses, on the bow-tie pattern. The phase-shift network goes between the audio source (the modulator) and the horizontal-amplifier input terminals of the scope. Which of the two circuits to use must be

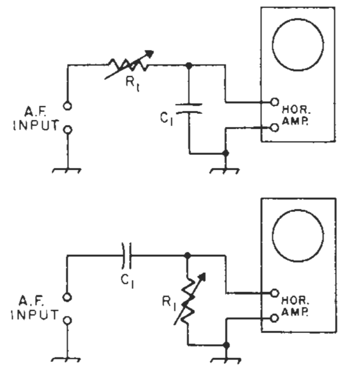


FIG. 6—Simple phase-correction circuits. For a frequency of approximately 1000 c.p.s. C_1 may be 0.002 μf. in both circuits. R_1 is a 1-megohm variable resistor, audio taper preferred. To prevent hum pickup, shielded wire should be used between the audio source and the network, and the latter should be connected to the scope by very short leads.

determined by trial; if it doesn't compensate the first way tried, interchange the resistor and capacitor. Adjustment of R_1 will shift the phase enough, in nearly all cases, to make the pattern edges go from an oval to a line. In addition to being frequency-sensitive—which is why the circuit is usable only with a single tone—the circuit also suffers from the fact that the amplitude of the output from the network changes with adjustment of R_1. Thus it is necessary to adjust the horizontal-amplifier gain, or the input-signal amplitude, concurrently with the phasing adjustment in order to maintain a desired pattern width. These disadvantages are minor compared with the desirability of single-line pattern edges.

» *Understanding how the basic modulation patterns of s.s.b. are formed will be of help in interpreting the patterns actually displayed on the screen. Photographs of some of the latter are presented in other articles in this book.*

Sideband Scope Patterns

GEORGE GRAMMER, W1DF

When the pattern formed by an amplitude-modulated r.f. wave is viewed on the scope, using a linear horizontal sweep, the top and bottom edges of the pattern outline the modulation envelope, the pattern of the modulating signal being traced by either edge. The individual r.f. cycles cannot be distinguished, in the ordinary case, because the radio frequency is so great compared with the modulating frequency.

Sideband techniques introduce other variations into the modulated pattern. A good starting point for considering these is the balanced modulator.

The Balanced Modulator

While the balanced modulator appears in many circuit guises, the principle is always the same: the carrier is made to disappear from the output, leaving only the two sidebands that are generated by amplitude modulation. The circuit of Fig. 1 is representative. Although perhaps more complicated-looking than some, it is easy to understand. R.f. drive is applied to the grids of the two tetrodes in push-pull, so the r.f. voltages at the grids are out of phase. That is, one grid is positive with respect to cathode when the other is negative, and vice versa. The amplified r.f. voltages at the plates also are out of phase, and if the amplification is the same in both tubes the two r.f. voltages have equal amplitudes. Since the plates are connected in parallel, the two voltages buck each other out, and there is no output.

In this circuit the amplification in each tube depends on its screen-grid voltage. If the two tubes are well matched and the r.f. grid circuit is carefully balanced, the amplification will be the same in both tubes when both have the same d.c. screen voltage. The audio modulating signal is superimposed on the d.c. screen voltage to drive the screens in push-pull, so if the bottom end of the audio transformer secondary is instantaneously positive with respect to the center tap, the upper end is simultaneously negative. Thus the instantaneous voltage will be increased on tube 2's screen while the voltage on tube 1's screen will be decreased. (In a later part of the audio cycle this will be reversed.) The unequal screen voltages mean that one tube amplifies more while the other amplifies less, resulting in an output from the modulator that is proportional to the difference in amplification. Thus there is output only when audio voltage is applied to the screens, and if the operating conditions are properly chosen the r.f. output voltage is proportional to the modulating voltage.

The effect is similar if different d.c. voltages are applied to the screens of the two tubes. If the voltage on tube 1's screen is fixed while that on tube 2's screen is adjustable, varying the latter voltage will cause the r.f. output to vary. This screen-voltage adjustment can be used either for obtaining exact balance and zero carrier output or for allowing any desired amount of carrier to get through.

Fig. 2 illustrates a number of possible conditions. The audio signal applied to the modulator is a sine wave having the amplitude shown at the top. In A the modulator is unbalanced sufficiently to allow a carrier of amplitude *X* to get through. The resulting signal is simple amplitude modulation of about 60 per cent. If the modulating frequency is 1000 cycles the signal has the r.f. spectrum shown in the third column, consisting of the carrier and two sidebands spaced 1000 cycles above and below the carrier.

If we now bring the modulator nearer to

From "Sideband Scope Patterns," August, 1964, *QST*.

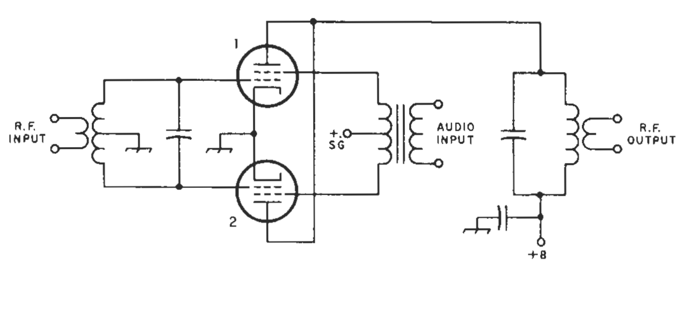


FIG. 1 — A representative balanced-modulator circuit.

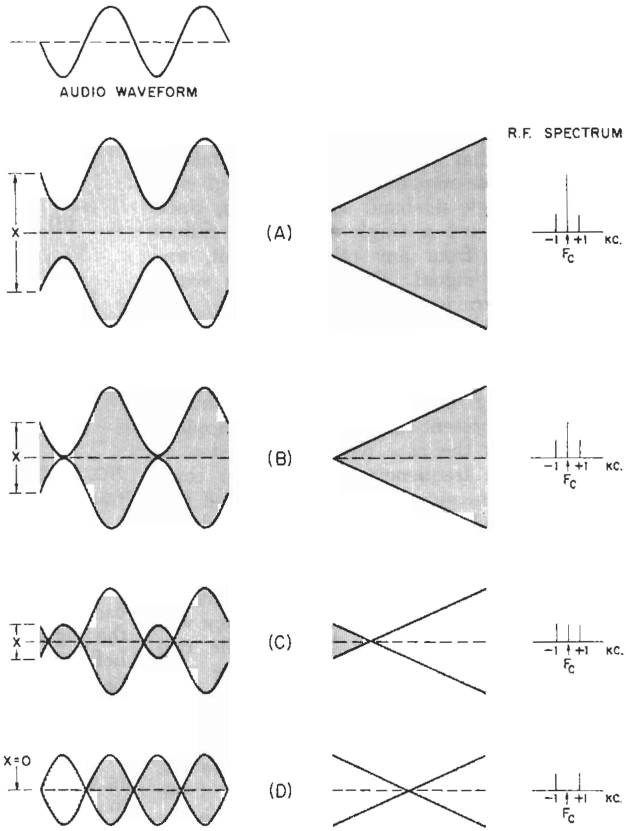


FIG. 2—Patterns obtained with a balanced modulator having varying amounts of carrier feedthrough. First column: Wave-envelope patterns using an ideal linear sweep locked to show two cycles at the audio frequency. Second column: Corresponding trapezoidal patterns using the audio frequency as the horizontal sweep. Third column: Radio-frequency spectrum corresponding to the oscilloscope pattern. F_C is the carrier frequency.

balance by adjustment of the d.c. voltage on one of the screens, only the carrier output is affected, since the modulating audio signal is still the same and the *variations* in output voltage are unchanged. The sideband amplitudes therefore remain constant, and bringing the modulator nearer to balance simply reduces the carrier. This has the effect of increasing the percentage of modulation, as shown at B. Here the carrier output has been adjusted so the wave is just 100 per cent modulated. As shown by the r.f. spectrum at the right, the relationship between the carrier and sideband amplitudes differs from that at A, but the same frequency components are present.

If we reduce the carrier amplitude still more, as in C, the sideband amplitudes are still unaffected. The lower and upper edges of the pattern now cross over for part of the time, the result being a picture that *appears* to have had something added to the modulation. What the scope does not show is that during this crossed-over period the phase of the r.f. output is reversed: In the patterns shown at A and B one of the two tubes—tube 1, perhaps—is actually supplying the output power, with tube 2 doing some bucking, but not enough to cancel all of tube 1's output since tube 2's d.c. screen voltage was deliberately reduced to let some carrier through. In C, the d.c. screen voltages are nearly enough balanced to allow tube 2 to take over during the part of the time where the crossover shows. Since the r.f. output voltages of the two tubes are out of phase, the resultant r.f. phase is determined by the tube which happens to predominate at any given time. When this is carried to the extreme and the carrier is balanced out completely, tube 1 predominates during exactly one half of the audio modulating cycle and tube 2 is top man during the other half. This is shown at D.

Note again that in all these examples the amplitudes of the two sidebands have remained exactly the same, as shown by the spectrum drawings at the right. Only the carrier amplitude has changed. Furthermore, in spite of the highly-distorted appearance of the modulation envelopes in C and D, there is actually no distortion. No new sidebands appear when the carrier is reduced beyond the limit of 100-percent pure amplitude modulation. The carrierless signal occupies no more bandwidth than the original a.m. signal, A, which had a low percentage of modulation.

Single Sideband

The balanced modulator represents the first step in the process of generating a single-sideband signal, in most circuit designs. However, regardless of the method, the ultimate result is the more-or-less complete elimination of the carrier and one of the sidebands from the final output. In Fig. 2D, for example, the spectrum consists of only two side frequencies, since the carrier has been suppressed and the modulating signal was of sine form. Eliminating one of these remaining frequencies gives a single-tone single-sideband signal which, since it consists of just a single frequency, differs in no respect from any other constant, unmodulated r.f. signal. Displayed on a scope tube, the pattern is a simple rectangle.

That, at least, is the *ideal* single-tone s.s.b. signal. The top and bottom edges of the pattern would be perfectly smooth, straight lines, as at A in Fig. 3. If more than the wanted single tone is present the edges will be rippled. Fig. 3B shows the case where suppression of the other side frequency has not been complete. This remnant modulates the desired frequency to produce a ripple that resembles a low percentage of amplitude modulation by an audio frequency equal to the *difference* in frequency between the two sidebands. That is, if the original audio modulating frequency was 1000 cycles the difference between the two side frequencies is 2000 cycles. In the case shown, the amplitude of the undesired side frequency is 10 per cent of that of the desired frequency, or 20 db. down.

If the sideband suppression is complete but the carrier suppression is not, the edges of the pattern are again rippled, but in this case the difference in frequency between the two components is equal to the frequency of the original modulating tone, 1000 cycles in this example. This is shown in Fig. 3C, where the width of one ripple cycle is just twice that in B. This assumes that the horizontal linear sweep frequency is the same in both cases, of course.

When carrier and undesired side frequencies both are present, the ripple pattern becomes more complex, and the shape depends on the relative amplitudes and phases of the "spurious" signals that, together with the desired single tone, make up the composite pattern. These spurious signals may include r.f. components resulting from harmonics in the original audio modulating tone, since only a really pure sine wave will be free of such harmonics. At this stage it is better to resort to using a highly-selective receiver to identify the components in the complete signal, since anything more than a rough analysis of the pattern (based on the cycle width, as in C and D) becomes difficult.

The single-tone pattern is a useful one for showing carrier and sideband suppression up to the limit of resolution of the c.r. tube. With a tube in good focus and with the brightness adjusted to give the finest possible trace, it is possible to see the effects of unwanted components that are 35 to 40 db. below the desired output. This assumes a 5-inch tube with the pattern height adjusted to make the maximum use of the screen area, or about 3 inches. If the signal has gone through linear amplifiers before being applied to the scope, care must be taken to see that the drive level is kept well within the limits of amplifier linearity. There is no easier way to clean up the edges of a ripply pattern than by increasing the drive to the saturation point! While the pattern may look fine, tuning through the signal with a selective receiver will quickly dispel any illusions about its quality—just another instance of how one can be misled in interpreting a scope pattern without knowing all the factors that entered into its makeup.

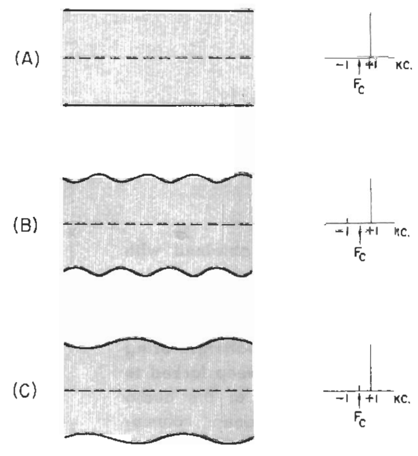


FIG. 3—Single-tone s.s.b. patterns, using the same type of horizontal sweep as in Fig. 2.

The Two-Tone Pattern

The single-tone pattern just discussed is excellent for checking carrier and sideband suppression, but it tells little or nothing about the linearity of the single-sideband system. To get a check on linearity it is necessary that the output signal vary in amplitude (which it should not do in the case of a single tone), and this requires that at least two frequencies be present in the modulation. Hence the utility of the "two-tone test."

Ideally, the two tones should be quite pure (i.e., sine waves) and should have the same amplitude, since tones meeting this specification will give patterns that are easiest to interpret. In s.s.b. such a pair of tones will generate just two radio frequencies, each removed by its own frequency from the suppressed carrier. The pattern formed by two such tones will be identical with the pattern formed by the two side frequencies in a balanced modulator as given in Fig. 2D, insofar as external appearances go. But there is a difference. In the two-tone s.s.b. signal the relationship between the sideband frequencies and the suppressed-carrier frequency is not the same as it is in the double-sideband suppressed-carrier signal created by the balanced modulator alone. This does not show in the scope pattern, but it has an important bearing on the setup for *creating* the pattern.

The wave-envelope pattern of Fig. 2D was formed by using a linear horizontal sweep synchronized at one-half the modulating frequency so that two modulation cycles were displayed. The modulating frequency is equal to one-half the difference between the two side frequencies shown in the spectrum diagram. That is, a modulation frequency of 1000 cycles creates side frequencies that are 1000 cycles each side of the carrier, or 2 kc. apart. Thus when the 1000-cycle voltage is used for the sweep, to give the trapezoidal pattern in the second column, the picture has the bow-tie form shown, indicating that the modulation envelope has a double-frequency component. The two side frequencies automatically have the same amplitude, so the pattern is symmetrical when the carrier is completely suppressed.

To get the same type of pattern from a two-tone s.s.b. signal the sweep frequency has to have this same relationship to the tone frequencies. If the audio tones are, for example, 1000 and 1500 cycles, their difference is 500 cycles and the basic sweep frequency has to be 250 cycles in order to generate the bow-tie pattern. It is possible to derive such a frequency from the two original tones, but not by simple means. In general, the bow-tie pattern must be obtained directly from a balanced modulator.

The linear sweep is a little different. It is possible to use an adjustable free-running (not synchronized) linear sweep and set it to a frequency that will meet the requirements for displaying one or more cycles of the modulation envelope. Often this is not too satisfactory because the sweep oscillator is not stable enough to hold the pattern stationary more than momentarily. An alternative is to use a pair of tones having a frequency difference equal to some submultiple of the lower one, and then use the lower tone alone to lock the sweep sync circuit.

Whatever the synchronizing method, the linear-sweep two-tone test pattern is useful only for detecting fairly gross misbehavior in a linear amplifier. Accurate analysis tends to become impossible with s.s.b., since the modulation envelope does not resemble the original audio signal at all. Here the best method is to observe the actual radio frequencies present in the signal (by using a receiver[1]) and measure their relative amplitudes.

The bow-tie pattern, when it can be used, will provide more information than the envelope pattern. The linearity of the system can be judged by the straightness of the edges of the bow tie, just as linearity with regular amplitude modulation is judged from the trapezoidal pattern. The limitations in generating the bow tie, mentioned above, occur particularly in filter-type sideband generators. Most of these have no provision for bypassing the sideband filter to allow the entire balanced-modulator output to be used for exciting the following stages. This limitation is not so prevalent in phasing-type generators.

In any case, *if* the balanced-modulator output can be used to drive the rest of the transmitter a single audio tone will suffice, and it can be used to synchronize the scope to get a stationary pattern. The tone need not be very pure for the bow tie. However, for single-tone tests—which supplement the two-tone test as described earlier—the audio tone should be as pure as possible.

In general, the most practical linearity test for an amplifier, regardless of the type of s.s.b. generator, is one in which the r.f. input to a linear stage is compared directly with its output. In this way the signal is compared with its amplified reproduction. If the two are exactly in or out of phase, the pattern will be a sloping straight line.[2] Making this test requires a two-tone signal, since the instantaneous amplitude must vary over the entire amplifier characteristic in order to show its total operation. Provided no appreciable r.f. harmonics are present in either the input or output signals, the presence of curvature in the line pattern will indicate amplifier nonlinearity. R.f. harmonics will be inconsequential if the input and output signals come from tank circuits having ordinary Q values.

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[1] See "Checking Signal Quality With the Receiver," page 191.

[2] See "Distortion in Single-Sideband Linear Amplifiers," page 116.

» Although the two-tone test pattern so widely used in s.s.b. transmitter testing can be generated with only one audio tone, it is more convenient, especially with filter-type sideband generators, to have a signal generator that supplies both tones. Good waveform is important.

Two-Tone Test Generator

ROBERT F. TSCHANNEN, W9LUO

The "Two-Tone Test Generator" described here is designed to provide two independent low-distortion audio test signals. The unit is compact and uses only two tubes. No special components are used in the construction. If the generator is carefully made and adjusted, the total harmonic distortion can be as low as 0.1 per cent.

The Basic Circuit

The basic circuit of a phase-shift oscillator is shown in Fig. 1. Operation depends upon producing 180-degree phase shift in the *RC* network consisting of three capacitors and three resistors; sufficient gain must be produced by the oscillator tube to make up for the loss in the network. For the circuit shown, a gain of 29 times is required to sustain oscillation.[1] The frequency of oscillation is determined by the equation

$$f = \frac{10^{12}}{2\pi\sqrt{6}\,RC} = \frac{10^{12}}{15.4\,RC}$$

where *R* is in ohms and *C* is in micromicrofarads. If the oscillator tube has a gain less than 29, oscillation will not begin; if excessive gain is obtained, appreciable distortion may be produced.

The phase shift through the network at harmonic frequencies is always less than 180 degrees and in some cases approaches zero. This gives rise to negative feedback which reduces the gain at harmonic frequencies; therefore, essentially a pure sine wave results. Maximum harmonic reduction occurs at the point where the system is just able to sustain oscillation.

General Circuit Description

A single 6AN8 tube is used as oscillator and output section for each channel of the generator. The pentode section functions as the oscillator proper; the triode section operates as a cathode follower output. A half-wave selenium rectifier followed by considerable filtering provides good d.c. for the oscillators. The complete schematic is shown in Fig. 2.

The 1000-ohm controls in the cathode circuits of the pentode stages are used for controlling distortion. The controls permit adjustment of the oscillator tube gain to the point where oscillation will just be sustained. This also corresponds to the point where minimum distortion occurs. Two additional 1000-ohm controls in the cathodes of the triode cathode followers provide control of outputs from either channel.

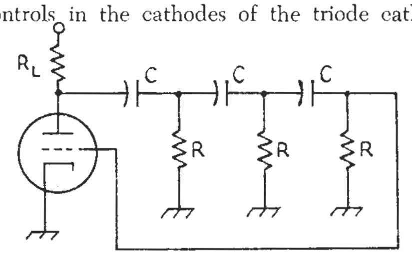


FIG. 1—The basic phase-shift oscillator circuit.

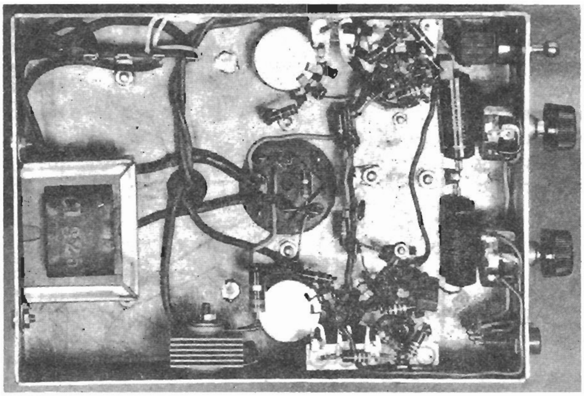
The *R* and *C* component values for the networks shown in the schematic of Fig. 2 are approximately correct for the generation of 400- and 1000-cycle tones. Other values are given in Table I.

It is important that the linearity of the cathode follower be good since otherwise distortion may be added by this stage. The use of a low-μ triode tube such as the triode portion of the 6AN8 permits the handling of higher grid swings without distortion. Since the signal handled is small, the possibility of distortion becomes negligible. The effective

From "A Compact Two-Tone Test Generator," May, 1955, *QST*.

[1] Ginzton and Hollingsworth, "Phase-Shift Oscillators," *Proc. of the IRE*, Feb., 1941.

The two-tone test generator is a compact and inexpensive unit and provides two audio signals of different frequencies and equal amplitudes for testing any type of s.s.b. generator. Distortion is extremely low if proper care is used in adjustment.



Arrangement of parts below chassis. The two oscillator-buffer circuits are identical in circuit but not in component values. The three electrolytic condensers in the power supply are contained in a single can-type unit (Mallory 311.9) thus conserving space underneath.

output impedance of the cathode follower is approximately equal to

$$\frac{10^6}{g_m}$$

in shunt with the cathode resistance to ground (where g_m is the transconductance in micromhos). The output impedance is therefore of the order of only several hundred ohms. This is desirable since output signals may readily be coupled into a combining network without appreciable interaction. The tapped-down take-off point from the plate of each oscillator tube reduces external loading on the oscillator and also reduces the output level to the point where the cathode-follower grid circuit can handle the signal without distortion.

When used for lowest distortion, the output of either channel is of the order of 1 to 1.5 volts r.m.s. Output levels of 8-10 volts r.m.s. are obtainable if a few per cent distortion is tolerable. The increased output capability is obtained by readjusting the oscillator cathode resistance.

The total "B" current drain of both oscillators and output stages is about 16 ma. Line-voltage variations do not greatly influence the oscillator frequency; therefore voltage stabilization is not required. Larger screen bypass and coupling capacitors do not add particularly to the performance of the unit since fixed-frequency operation is used.

Construction

The chassis layout of the phase-shift oscillator is not critical. The entire unit is constructed on a 5×7×1½-inch chassis. The grid leads of the oscillator tubes are preferably kept short and dressed away from a.c. supply and filament leads. One side of each filament of the 6AN8 tubes is grounded. The photographs of the chassis will assist the builder in making a suitable layout. The

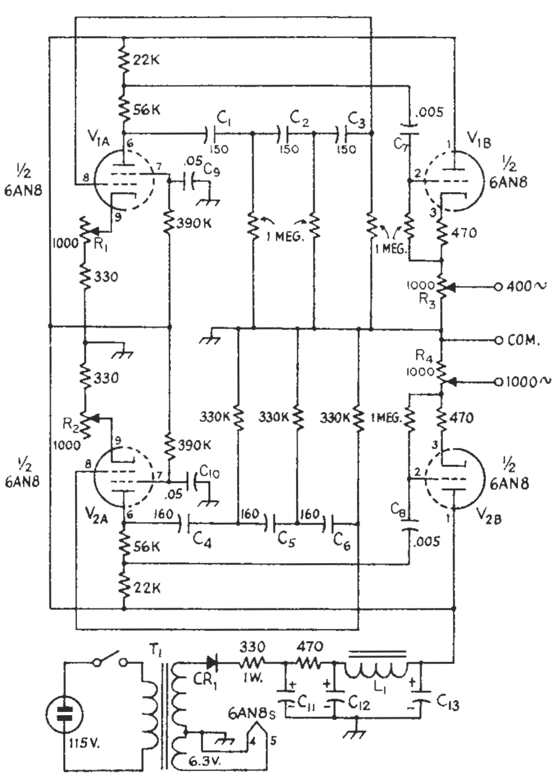


FIG. 2—Circuit of the dual a.f. test oscillator. Resistors are ½ watt, 10 per cent tolerance, unless otherwise specified. Capacitance values below 0.001 μf. are in pf. Potentiometers are linear-taper 1-watt composition.

C_1-C_6, inc.—Silver mica, 5 per cent tolerance.
C_{11}, C_{12}—120-μf. 250-volt electrolytic.
C_{13}—40-μf. 250-volt electrolytic.
L_1—5 henrys, 50 ma. (Stancor C-1325).
CR_1—75-ma. selenium rectifier.
T_1—125 volts, 50 ma.; 6.3 volts, 2 amp. (Stancor PA-8421).

small power transformer is capable of supplying as many as four individual oscillators. If desired, a 6X4 rectifier may be substituted in place of the selenium rectifier; in this case the 330-ohm 1-watt current-limiting resistor in series with the rectifier may be removed.

Miniature silver mica capacitors were used in the phase-shift networks for compactness; however, conventional micas may be used successfully if space is available. The coupling capacitors C_7 and C_8 may be Hi-K disk ceramic

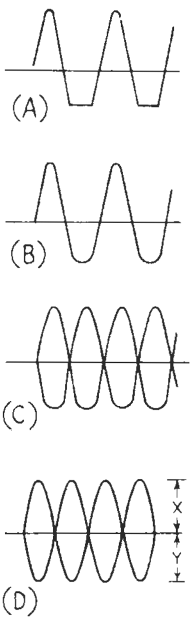


FIG. 3—Improper operating conditions are shown by 'scope traces. A—Excessive oscillator tube gain. B—Excessive oscillator tube gain, but not as much as in A. C—Same as B except with change in 'scope sweep speed to facilitate estimating second-harmonic distortion by the degree of asymmetry. D—Optimum symmetry ($X = Y$), minimum even-order harmonics, low distortion in output.

or paper types. Components for the phase-shift network are mounted on terminal strips or boards for rigidity and neatness. The capacitors C_1 through C_6 are not visible in the bottom view since they are beneath the terminal strips which are located on each side of the chassis. The controls R_1 and R_2 are located on each side of the electrolytic filter capacitor. The output controls, a.c. switch, and output tip jacks are on the front flange of the chassis. The layout shown will provide good accessibility to nearly all components.

Adjustment & Checking

After the wiring has been completed and checked the unit may be turned on and each output observed on a 'scope. If no output appears, adjust the cathode resistor of the oscillator to just slightly beyond the point where oscillation starts.

With the values of cathode resistances shown on the schematic, it should normally be possible to stop oscillation near one end of the control and produce high (but slightly distorted) output near the other end of the control. At the point where the distortion

TABLE I

Freq. (c.p.s.)	R	C pf.
250	1 meg.	260
300	"	216
350	"	186
400	"	162
500	680 K	191
600	"	159
800	"	119
1000	390 K	166
1250	"	133
1500	"	111
2000	270 K	120

becomes noticeable, the wave will usually have an appearance similar to that shown in Fig. 3A or 3B, which indicates even-harmonic distortion (principally second). If a distortion meter or wave analyzer is available it will be simple to adjust each cathode control to the point where lowest distortion is obtained. Since such equipment is seldom available to the ham or experimenter, a reasonable means of minimizing the distortion is to apply the signal under test to the vertical plates of a 'scope and adjust the horizontal sweep speed until a pattern similar to Fig. 3C is obtained. The distortion control can now be rotated until dimensions X and Y are as nearly equal as possible (see Fig. 3D). In other words if X and Y are made equal, any asymmetry due to second harmonic distortion is negligible.

Using the Two-Tone Generator

If the generator is used to test an s.s.b. exciter equipped with a high-impedance microphone input circuit, it will be desirable to divide down the output signal by means of a circuit such as shown in Fig. 4. If an input terminal or jack for audio input at higher levels is provided on the unit, the output of the generator need not be divided down. Since a few volts of d.c. exists from the output of the generator to ground, a blocking capacitor should be used if one is not employed in the equipment under test.

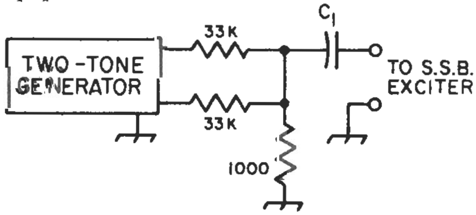


FIG. 4—Method of connecting the two-tone generator to the microphone input terminals of a speech amplifier. The 33K resistors provide good isolation between the sources of the two output frequencies. C_1 may be 0.01 μf. for high-impedance circuits.

» The step-by-step adjustment procedure described here should lead to optimum performance of your phasing-type exciter. You'll be able to recognize that "optimum" stage when you reach it, and will get some ideas of where to look for trouble if you don't achieve it at first trial.

How To Adjust Phasing-Type S.S.B. Exciters

ROBERT W. EHRLICH, W2NJR

A well-adjusted phasing-type s.s.b. exciter usually holds its adjustments over a reasonably long period, but it seems there comes a time in the life of every piece of equipment when the OM says to himself, "I'd just like to get in there and give the thing a good going over." Because a good adjustment job is so important in determining performance of this type of equipment, it behooves the owner to avail himself of the best techniques available. Let's review these techniques and point up certain of the tricks that will help (and pitfalls to be avoided) to achieve the utmost in performance.

Principles

To start off, let's take a look at the exciter itself—the circuits that need to be adjusted and what they are supposed to do. Fig. 1 is a block diagram of a typical exciter. It should be possible to identify the corresponding circuit sections in any phasing exciter.

First, there are two balanced modulators, each of which consists of a *pair* of crystal diodes or tube sections. Associated with each pair is a balancing control, usually on the front panel, and usually called "carrier balance" or just "carrier." The two controls must be adjusted alternately until there is no output from the exciter when there is no modulation. This is a very simple thing to do; any simple r.f. indicator or even the plate meter of the final can be used to tell when the carrier is balanced out.

The remainder of the circuitry in Fig. 1 is involved in achieving sideband suppression. Keep in mind that carrier suppression and sideband suppression are different things, and either can be out of adjustment without affecting the other. It is the sideband suppression adjustment which is the more difficult of the two and will be the principal subject of the remainder of this article.

The function of the r.f. phase-shift network (in the upper part of Fig. 1) is to accept a signal from the local oscillator and divide it into two r.f. driving signals for the balanced modulators. The two driving signals must be *about* equal in amplitude but *precisely* 90 degrees out of phase. Fig. 2 shows two typical phase-shift-network circuits. In the transformer type (Fig. 2A), tuning the primary simply resonates it to the oscillator frequency, while tuning the secondary will cause a phase change

From *QST*, November, 1956.

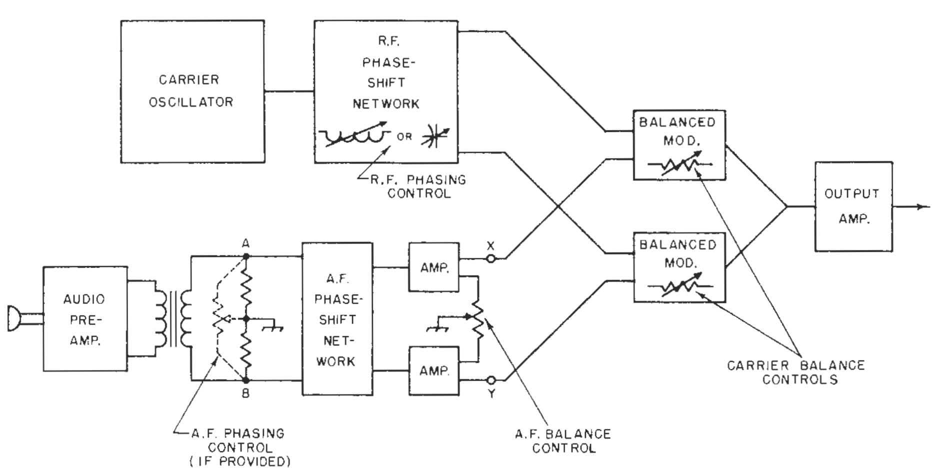
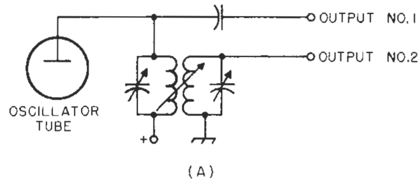


FIG. 1—Block diagram of typical phasing-type exciter, showing locations of significant controls and adjustments.



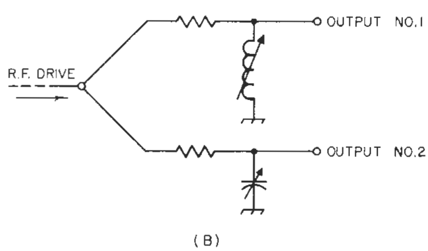


FIG. 2—Two of the many possible types of r.f. phase-shift networks. Adjustments are described in the text.

between the two outputs; hence the tuning element in the secondary becomes what we will call the "r.f. phasing control." In the other type of network (Fig. 2B), either of the reactive elements may affect phase shift, so it is customary to leave one of them at about mid-position and use the other for the "r.f. phasing control."

The second control involved in getting sideband suppression is called "audio balance." As shown in Fig. 1, it is connected between the two sides of the audio amplifier driving the modulators. Its purpose is to make the audio drive to the two balanced modulators effectively equal; that is, to make the sideband output from each balanced modulator equal for equal audio voltages applied to the audio-amplifier grids.

The remaining element in obtaining sideband suppression is the audio phase-shift network (Fig. 1). Its function is to take the audio speech signal and produce two audio output signals that are exactly equal in amplitude and precisely 90 degrees out of phase. To do this job over the whole range of speech frequencies is quite a trick, and it requires precise values of resistance and capacitance within the network. Fortunately, several maunfacturers make prealigned plug-in phase-shift networks at very reasonable cost, so for the purpose of this article we shall assume that the network has been factory wired and is OK. (The only remaining stumbling block here is to make very sure that the socket connections to the network are wired correctly!)

There is one additional point about the audio phase-shift network. Most present-day networks require an audio driving signal that is push-pull in nature (180 degrees out of phase) and unbalanced in amplitude in the ratio of 2 to 7. Referring to points A and B in Fig. 1, the required voltage division is usually accomplished by wiring a pair of precision resistors, with a resistance ratio of 2 to 7, between the input terminals and ground. In some exciters, however, an adjustable potentiometer (shown by dotted lines) is used in place of the precision resistors. If used, this potentiometer becomes a third control which must be adjusted to obtain sideband suppression. It may be called the "audio phasing control."

Test and Measurement Methods

Now, having identified the controls in the exciter to work with, let's take a few minutes to look at the means for testing and measuring sideband suppression. Fig. 3 illustrates the features of a typical test setup.

The audio oscillator should put out a good sine wave. Most commercial designs (and kit designs) will be satisfactory. In addition, the frequency of the oscillator should be capable of being set within about 100 cycles of the specified test frequency for the audio phase-shift network being used. In particular, if your exciter uses a Millen 75012 network, the frequency should be 1225 cycles, while the B & W 2Q4 network requires 1000 cycles. For other

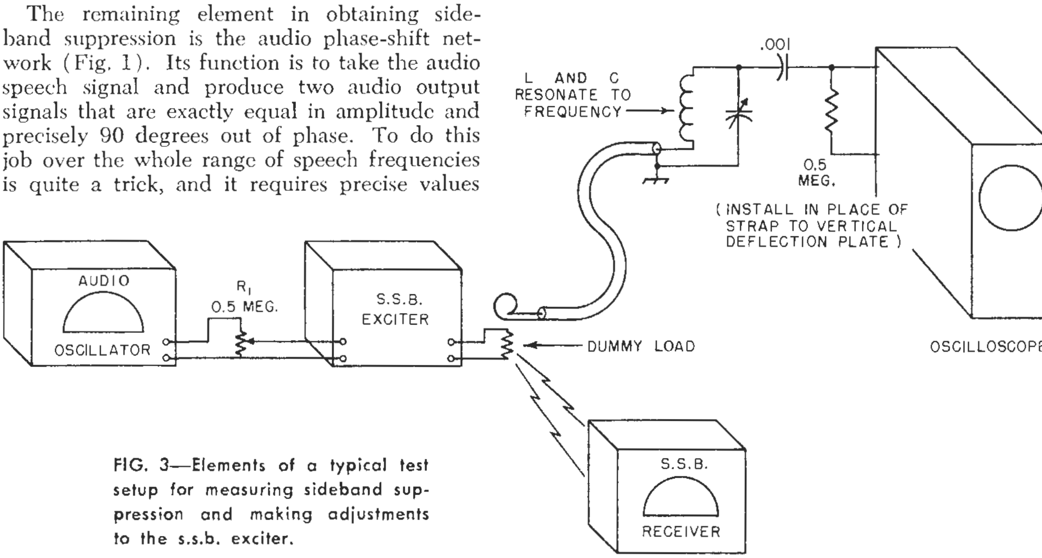


FIG. 3—Elements of a typical test setup for measuring sideband suppression and making adjustments to the s.s.b. exciter.

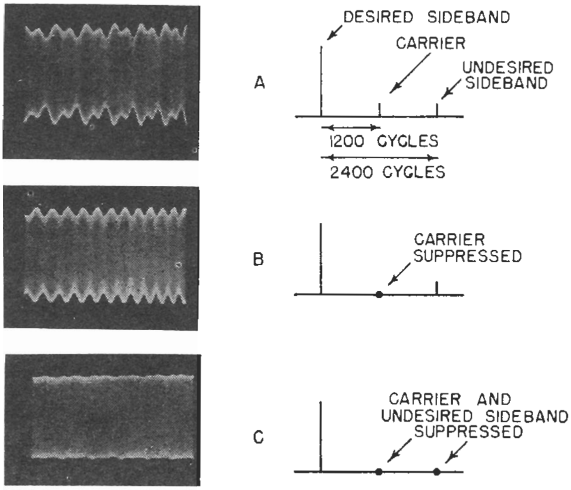
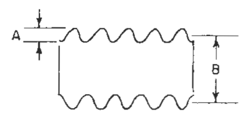


FIG. 4 — Typical oscilloscope patterns and corresponding signal-spectrum diagrams. For these photos the audio test frequency was 1200 cycles and the scope sweep rate was 200 cycles.

TABLE I

RATIO A TO B	Approx. Suppression
1:10	20 db.
1:15	24 db.
1:20	26 db.
1:30	30 db.
1:50	34 db.
1:100	40 db.



networks, refer to the manufacturer's instructions.

Potentiometer R_1 is introduced in Fig. 3 to avoid overloading of the audio stages in the exciter. Such overload is frequently a source of trouble, as it distorts the sine wave and gives erroneous readings. The recommended procedure is to keep the audio gain control on the s.s.b. exciter up at its normal operating level—say, 12 o'clock—and use the external potentiometer, R_1, to control the audio level going through the exciter during tests.

Over on the output side, Fig. 3 shows two optional methods of observing sideband suppression, the oscilloscope and the s.s.b. receiver. Each has its merits, and some discussion will be in order.

The oscilloscope method is based on the principle that if a pure audio tone is sent through a perfect s.s.b. exciter, only one r.f. signal will come out: the desired sideband. On the scope, this will look like a c.w. signal —that is, a smooth stripe. If there are any other signals present, unwanted sideband or carrier, they will beat with the desired signal and produce ripples on the scope pattern. The object, is to adjust the exciter until the pattern is a smooth stripe with no ripple.

Ripple Patterns

Fig. 4 shows some of the typical scope patterns. Note particularly that ripples due to the carrier are twice as wide (half the frequency) as those due to the unwanted sideband. In this way it is possible to tell what kind of adjustment is needed. Another feature is that the amount (height) of ripples gives a convenient measure of the amount by which the unwanted signal is suppressed. Table I gives some guide figures for evaluating suppression in db.

The oscilloscope method has its drawbacks when getting down to high degrees of suppression. To begin with, it's difficult to see the ripple when it's only 1/100 of the total pattern height. (This would be 40 db. suppression; see Table I.) Something a little more tricky is the effect of the audio third harmonic. This harmonic falls at a frequency that is two times the fundamental audio frequency away from the wanted sideband, but the unwanted sideband is separated from the wanted sideband by precisely the same frequency in the opposite direction! The scope can't distinguish the two signals, so it is entirely possible to find yourself making adjustments to introduce just enough unwanted sideband in the correct phase to cancel out the insidious third harmonic. As can be imagined, this condition becomes important when there is some distortion in the sine wave from the audio oscillator or in the exciter audio system. Generally speaking, with practical equipment, it may be said that the oscilloscope method breaks down in the vicinity of 30- or 40-db. suppression, depending on the degree of audio distortion.

Using a Receiver

The second method for observing sideband suppression, as shown in Fig. 3, is through the use of an s.s.b. receiver. The idea, of course, is to listen to the signal on the unwanted sideband while making adjustments to null it out. The receiver method overcomes the disadvantages discussed above for the oscilloscope method, but it has some difficulties and tricks of its own.

First of all, it is almost impossible to use a receiver successfully for aligning an exciter of the type which uses a v.f.o. directly on the operating frequency. Direct radiation from the oscillator comes into the receiver and tends to block it. Use of the receiver method, therefore, is generally restricted to heterodyne-type exciters.

Next, it is imperative not to overload the receiver. The strongest signals should not exceed S9. This may require shorting the antenna terminals and similar precautions.

It is also important to keep in mind that the output spectrum of the exciter contains many signals, as depicted in Fig. 5, and this can lead to confusion. One technique to help identify which signals are which is to unbalance the carrier temporarily or to remove the audio. Another is to set the exciter temporarily to transmit the opposite sideband so that the unwanted signal will become strong and can be tuned in carefully, after which the exciter can be returned to the first sideband while making adjustments.

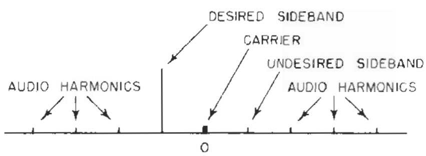


FIG. 5—Typical output spectrum of s.s.b. exciter driven by an audio oscillator.

Once the undesired sideband signal has been identified, it may be helpful to detune the s.s.b. receiver away from its normal setting at zero beat with the carrier, and toward the unwanted sideband signal. This will bring the signal you're concerned with through the receiver at a low pitch, which makes it easier to identify it by ear apart from any other signals that may be present. One-time c.w. operators will recognize this technique as being similar to the matter of digging a weak signal out from under some strong QRM.[1]

Adjustments

Let's get down to the job. Just how should the adjustments be made in order to realize the best in sideband suppression? We'll take it on a step-by-step basis.

Initially, we'll confine the discussion to the type of exciter that has fixed precision resistors at the input to the audio phase-shift network.

Let's assume that everything has been set up to measure sideband suppression (Fig. 3). Set the exciter to transmit either of the two sidebands—we'll call it Sideband 1. Now make alternate adjustments of the "r.f. phasing" and "audio balance" controls until the opposite side band is completely suppressed. This should be fairly easy to do if the exciter is working properly, but there are certain points that should be watched in order to stay out of trouble:

1) Be sure the audio system is not being overloaded. Keep the audio gain control of the exciter up, and use minimum signal from the oscillator, under control of R_1 (Fig. 3).

2) Be sure the r.f. portion of the exciter is not being overloaded. Select an audio driving level that is considerably below that required to produce any noticeable saturation or flattening of r.f. output. If a final amplifier is available, another guide is to select a driving level that will cause the amplifier to operate at about half its peak input.

3) If the oscilloscope method is being used, there may appear to be some carrier unbalance (Fig. 4A) even though the carrier was previously balanced out with no modulation. It happens that most phasing exciters exhibit this characteristic to some degree. Don't worry about it; feel free to make carrier adjustments at any time in order to get the desired pattern (Figs. 4B and C).

Having made the adjustment for Sideband 1, don't be misled into thinking the job is done. It will be necessary to do some checking on the other sideband, too. Perhaps a few words on this important point will be in order.

Suppression of one audio frequency on one sideband is a good indicator but not a final check of exciter performance. It is possible, for example, to have both audio and r.f. phase shifts something like 85 degrees instead of 90 degrees and still obtain what appears to be good sideband suppression for *one* frequency on *one* sideband. Under these conditions, suppression will fail for all *other* audio frequencies and on the opposite sideband, too. It is necessary, therefore, to check suppression under at least one other set of operating conditions. The most convenient "other operating condition" is simply to switch the exciter to its opposite sideband.

From the foregoing, it follows that it is not too good to try to "favor" one sideband, even if operation on the other is never contemplated.

Let's get back to adjustments. Switch the exciter to Sideband 2. The usual condition will be that suppression on the new sideband is good but not perfect. It should then be possible to regain almost perfect suppression by finding a new setting of the "r.f. phasing" control. Note how much the adjusting screw has to be turned and finally set it midway between the first and second positions.

At this point, a typical exciter in good condition will exhibit about 25-30-db. suppression on either sideband. This can be checked easily with the oscilloscope by reference to Table 1. In an s.s.b. receiver, the unwanted sideband will be barely audible when the wanted sideband comes through at moderate level.

If the suppression so obtained is not good

[1] By "s.s.b. receiver" the author means just that—a receiver that has the type of selectivity necessary to reject one sideband while receiving the other. This is not the same thing as setting up a conventional receiver for s.s.b. reception, by turning off the a.v.c. and turning on the b.f.o. The conventional receiver can be used, however, if it has a variable-selectivity crystal filter. With the crystal filter in the sharpest position, tune in the signal component being inspected and use the phasing control to notch out the desired sideband signal. The b.f.o. can be set to a tone convenient for checking purposes, as described by the author.—*Ed.*

enough, or if you want to try to go for more than the typical 25-30 db., the next step will be to go on into some refinements of the voltage division at the input to the audio phase-shift network. To do this, it will be necessary to connect a high-resistance potentiometer (50 or 100 thousand ohms or so) across the fixed resistors in the network as indicated by the dotted lines in Fig. 1. This will become an "audio phasing" control. It will also be desirable to go over to the use of the s.s.b. receiver exclusively for adjusting sideband suppression, inasmuch as the scope has its limitations beyond 30 db., as discussed previously.

Start with the new potentiometer at about mid-position. Go back and try adjustments on both sidebands as described earlier, using different arbitrary settings of the new "audio phasing" control. Eventually a setting should be found which gives good suppression on either sideband and requires almost no readjustment of "r.f. phasing" as between one sideband and the other. The results should be considerably better, such that the unwanted signals can hardly be heard on the s.s.b. receiver when the wanted signals are quite loud. Meanwhile, on the scope there should be only a trace of ripple when transmitting either sideband.

Coming back to the type of exciter that already has a potentiometer for "audio phasing," the technique is, as you might suspect, much like that described in the preceding paragraph. The only difference is that, at least in most of the circuits observed by the writer, the setting of this potentiometer will be very critical. Initially, some rather widely-separated settings may be required to come even close to normal behavior, but once the exciter begins to act normally, as described above, only the smallest shifts in shaft positions should be made between one trial and the next. The final setting may involve as little as a degree or so of shaft rotation.

To summarize, it may be well to consider some objectives and possible practical results. Most commercially-available audio phase-shift networks are rated at about 40-db. suppression over a specified audio band. This ideal condition will be affected by such practical considerations as:

At the specified test frequency, the performance on *both* sidebands will have to be much better—in the order of 45-50 db.

The input impedance of the audio amplifier following the phase-shift network should theoretically be infinite; usually there is a certain amount of input capacitance.

There can be no distortion in the audio transformers and tubes following the audio phase-shift network.

There can be no difference in audio phase shift through the amplifiers following this network.

There should be no shift in stray capacitance in the r.f. circuit when shifting from one sideband to the other; usually the wiring of the sideband switch introduces some of this.

All components must be stable from the standpoints of age, temperature and humidity.

This imposing list of requirements is not intended to be discouraging but rather to point up the challenge involved, both in design and adjustment. As a practical matter, it should be no trick to obtain 30-db. sideband suppression with any good design, and with care and patience it is not at all impossible to get 35-db. suppression. The former appears to be quite acceptable judging from the average performance of signals heard on the air.

Trouble Shooting

The foregoing procedures will work fine provided everything is functioning properly and is wired correctly. Sometimes, however, things just don't seem to come around the way they should, particularly on a brand-new exciter. The problem then is to determine whether the trouble lies in the audio circuits, in the r.f. circuits, or in the balanced modulators themselves. The following paragraphs offer a few suggestions for isolating the trouble, using simple tests.

The audio system is easiest to check. Apply a steady audio tone to the input of the exciter by means of an audio oscillator. Using a high impendance a.c. voltmeter or v.t.v.m. (the usual multitester set on its output scale will usually do), measure the two a.c. voltages at the plates of the audio output tubes, points *X* and *Y* in Fig. 1. These voltages should be about the same, or it should be possible to make them equal by adjusting the "audio balance" control.

Another check point in the audio system is at the input to the phase-shift network, points *A* and *B* in Fig. 1. For most modern networks, the two a.c. voltages measured to ground should have the precise ratio of 2 to 7.

Assuming the audio system looks OK, any remaining difficulty is probably in the r.f. phase-shift network or in the balanced modulators. First, check the action of the "carrier" controls. The point of minimum carrier output should occur when each of the two controls is within about the center third of its rotation range. If one of them has to be set 'way off to one side in order to achieve carrier balance, you can suspect that one of the modulator tubes or crystals is defective.

If the balance points look OK, try unbalancing each of the two "carrier" controls one at a time by the same amount, say, 10 degrees of shaft rotation. The r.f. output produced by unbalancing each control should be about the same. If the action of one of them seems to be sluggish, a further check may be made by temporarily interchanging the leads carrying r.f. drive to the two balanced-modulator pairs. Now, if the inconsistent action of

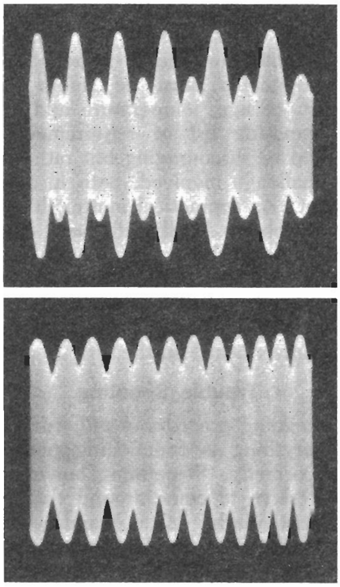


FIG. 6—R.f. phasing can be checked by observing the r.f. output envelope under phase-modulation conditions. Correct adjustment is shown, lower photo, where the ripple peaks are all evenly aligned.

the "carrier" controls seems to follow this change, it is an indication that something is wrong with the r.f. phase-shift network, whereas if the same control continues to perform poorly, the trouble probably lies in that balanced modulator.

Even when there is no actual component trouble anywhere in the exciter, it is possible for the r.f. phase-shift network to be so far off that the desired sideband adjustments just can't seem to be made. This is especially true of networks that have two variable elements, such as Fig. 2B. If you are suspicious of the r.f. phase-shift network, the scheme outlined below can be used to set this network independently of everything else in the exciter.

It is based on the principle that phase modulation (p.m.) requires the two modulators to be supplied with carriers that are 90 degrees apart, and also that pure p.m. is characterized by the absence of residual a.m. at the fundamental frequency. It should be possible, therefore, to set up the required 90-degree phase relationship by watching the p.m. signal on a scope.

The test setup is as shown in Fig. 3, with the oscilloscope. Set up the exciter for phase modulation (p.m.) according to its instructions, and introduce a moderate amount of carrier. Now apply audio drive until 20-30 per cent medulation appears on the scope. Referring to Fig. 6, adjust the r.f. phase-shift network until the ripple peaks are even with each other as in Fig. 6B. The "r.f. phasing" adjustment made this way should be very close to the final setting required for best sideband suppression.

» The necessity for care in preventing audio distortion following the audio phasing network has been emphasized in other articles on the phasing system. The reason why amplitude distortion causes spurious signals to be generated both inside and outside the wanted sideband is discussed here.

Post-Phasing Distortion

GEORGE GRAMMER, W1DF

A source of spurious emission that is peculiar to phasing-type single-sideband transmitters results from harmonic distortion in the audio amplifiers following the audio phase-shift network, and from nonlinearity in the balanced modulator. In general, these amplitude distortion products are not in the proper phase to cancel out on the unwanted side of the suppressed carrier.

Fig. 1 is typical of second-harmonic distortion that occurs in vacuum-tube circuits. The dashed curve is the output waveshape and the solid curves show the fundamental and second-harmonic frequencies of which it is composed. Two different frequencies cannot be shown on a single vector diagram but their behavior can be depicted by a series of "snapshots" at consecutive time intervals. Fig. 2 shows a series at 90-degree intervals over one cycle of the fundamental frequency. The second harmonic, having twice the frequency, changes phase twice as rapidly as the fundamental.

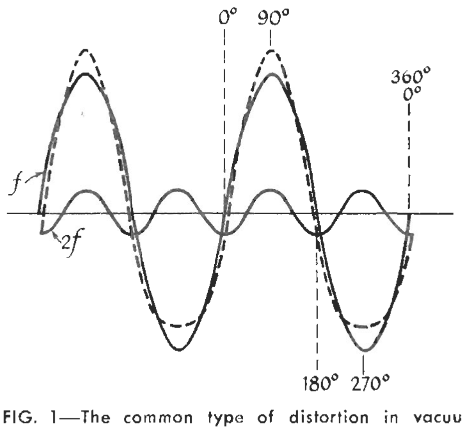


FIG. 1—The common type of distortion in vacuum tubes can be resolved into a fundamental and second harmonic having the time relationships shown. The fundamental, f, is the amplified grid signal, and 2f is the generated harmonic.

Now suppose we have two signals such as shown in Fig. 1, identical except for the fact that the fundamental component of one lags 90 degrees behind the fundamental component of the other. This situation can arise when pure fundamentals, 90 degrees apart out of an audio phase-shift network, are applied to the

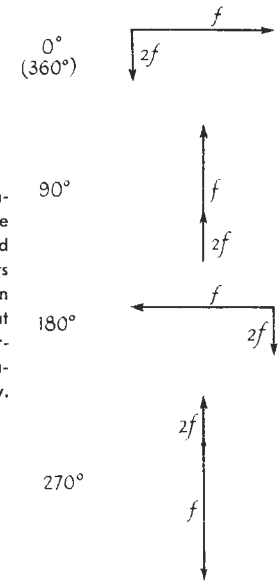


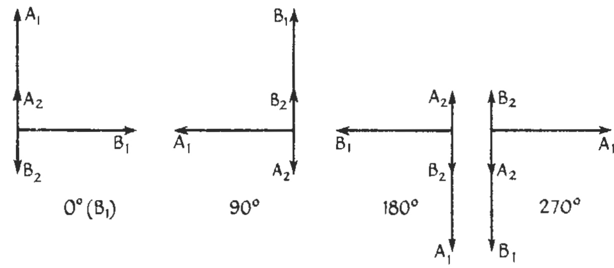
FIG. 2—Relationship between the fundamental and harmonic vectors of the wave in Fig. 1, taken at 90-degree intervals of the fundamental frequency.

grids of separate amplifiers that distort each signal in exactly the same way. The vector diagrams of Fig. 2 apply to each individually, so when the 90-degree fundamental shift of one channel is taken into account and the two channels are viewed simultaneously, the behavior with time is as shown in Fig. 3. Each harmonic component retains its same relative phase with respect to its fundamental. These harmonic components, it will be observed, are 180 degrees apart, not 90 degrees.

If the signals from the two channels in Fig. 3 are applied to balanced modulators in the usual way, there is no sideband cancellation at the harmonic frequency. Fig. 4 shows what happens. To avoid confusion in the drawings, the fundamental vectors have been advanced 30 degrees from the position at the left end in Fig. 3. Consequently, the harmonics advance 60 degrees in the same time interval. The audio relationships at this instant are shown at the top. When the outputs of the two channels are applied separately to modula-

From *QST*, February, 1954.

FIG. 3—Phase relationships in two channels when the fundamental in one (A_1) leads the fundamental in the other (B_1) by 90 degrees. A_2 is the second harmonic associated with A_1, and B_2 the second harmonic associated with B_1.

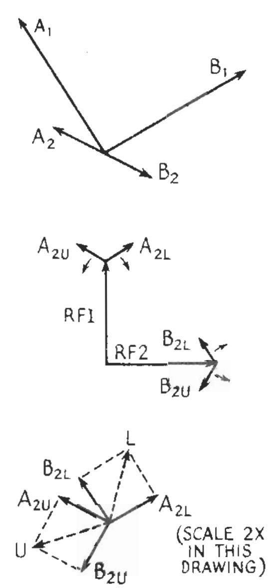


tors with 90-degree r.f. phasing, the composite diagram is as shown in the middle drawing. The fundamentals, A_1 and B_1, have been omitted in this drawing since they are no longer needed and merely follow the usual pattern for single-sideband generation. The second harmonic, A_2, of the first audio channel, when applied to the first r.f. channel generates side frequencies A_{2U} and A_{2L}, upper and lower respectively, spaced twice the fundamental frequency from the carrier. Similarly for the harmonic B_2 generated in the second audio channel when applied to the second r.f. channel. When the carrierless r.f. components are combined in the balanced modulator the two lower side frequencies add together as shown at L in the lowest drawing, and the two upper side frequencies combine into U. These are identical with the upper and lower side frequencies produced by single-tone modulation in one balanced modulator—that is, double-sideband suppressed-carrier transmission.

All harmonics generated in the circuits under discussion will be shifted in phase in proportion to frequency. Thus, if the fundamentals in the two channels are 90 degrees apart the second harmonics will be 180 degrees apart, the third harmonics 270 degrees, and so on. It has been shown that the second harmonic is transmitted as a carrierless double-sideband signal; this is true of all even harmonics. The case of the third harmonic is interesting because a phase difference of 270 degrees between the two audio channels is equivalent to a 90-degree difference, but with the lead and lag reversed as compared with the fundamentals. Hence, identical third-harmonic distortions in the two channels will give single-sideband transmission again, but with the output in the *unwanted* sideband. On the other hand, the shift at the fifth harmonic is 450 degrees, which is identical with the 90-degree shift at the fundamental; hence, fifth-harmonic distortion gives rise to a spurious component in the desired sideband but not in the other. Odd harmonics give single side frequencies, alternating between the undesired and desired sidebands.

This discussion has been idealized to the extent that it has assumed identical distortions in each channel, exact 90-degree r.f. phasing, and complete carrier suppression. For even harmonics, departure from exact audio phase and amplitude balance will affect the amplitudes of the side frequencies to some extent; so will inaccurate r.f. phasing. If some carrier gets through, there will be phase modulation along with the amplitude modulation, and under special conditions pure phase modulation would be possible. Similar departures in the case of odd harmonics will cause components to appear in both sidebands (although generally of unequal amplitudes), and incomplete carrier suppression also will cause phase modulation along with amplitude.

FIG. 4 — Sideband output when the harmonics (A_2 and B_2) in the upper drawing are applied to balanced modulators having 90-degree r.f. phasing. The r.f. carriers are shown for reference in the middle drawing. When carriers are eliminated the upper and lower side components, subscripts U and L respectively, combine as shown in the lower figure.



As a first approximation the spurious components can be taken to be of the same order as the original distortion although, as is evident from Fig. 4, they may be somewhat smaller. In a transmitter in which, in the absence of such distortion, the sideband suppression actually is close to the common target figure of 40 db., the distortion in the post-phasing audio amplifiers and in the balanced modulators would have to be kept to the order of 1 per cent to avoid degrading performance.

» *Articles earlier in this book have dealt with the linearity of amplifiers as a matter of tube design and operation. Now, given an oscilloscope for checking, how should it be hooked up and how should the resulting patterns be interpreted? Here are the answers.*

How To Test and Align a Linear Amplifier

ROBERT W. EHRLICH, W2NJR

It can generally be said that a transmitter is no better than its final amplifier, and this statement applies as much to a single-sideband transmitter as to any other kind—perhaps a little more so. If the linear final in an s.s.b. rig is out of adjustment, it not only can cause roughness, splatter and TVI but also will put signals right back in the suppressed-sideband space from which the exciter is working so hard to eliminate them. In other words, it can make the best exciter in the world sound pretty sick. When the linear is properly adjusted, however, the distortion or splatter components will generally represent less than one thousandth of the total power (30 db. down), effectively confining the whole signal to just the passband of the exciter.

One of the more important features of the linear amplifier is that the ordinary plate and grid meters are at best only poor indicators of what is going on. As the meters bounce back and forth, even a person who is thoroughly familiar with this kind of amplifier would be hard put to sense whether the input power registered is attributable to (a) overdrive and underload, which yield distortion, splatter, TVI, etc., or (b) underdrive and too-heavy loading, resulting in inefficiency and loss of output.

The simplest and best way to get the whole story is to make a linearity test; that is, to send through the amplifier a signal whose amplitude varies from zero up to the peak level in a certain known manner, and then observe, by means of an oscilloscope, whether this same waveform comes out of the amplifier at maximum ratings.

Test Equipment

Even the simplest type of cathode-ray oscilloscope can be used for linearity tests, so long as it has the regular internal sweep circuit. If this instrument is not already part of the regular station equipment, it might be well to purchase one of the several inexpensive kits now on the market, so that it will be on hand not only to make initial tests but also as a permanent monitor during all operation. Barring a purchase, it is recommended at least that a scope be borrowed to make the line-up checks, whereupon the regular plate and grid meters can serve thereafter to indicate roughly changes in operating conditions.

From *QST*, May, 1952.

All linearity tests require that the vertical plates of the scope be supplied with r.f. from the amplifier output. To avoid interaction within the instrument, it is usually best to connect directly to the cathode-ray tube terminals at the back of the cabinet. A pick-up device and its connections to the oscilloscope are shown in Fig. 1. Normally, the pick-up loop

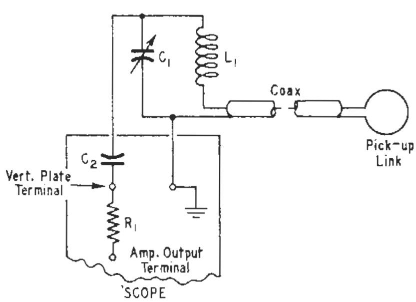


FIG. 1—The recommended method for sampling r.f. and applying it to the vertical plates of a scope. The pattern height can be varied by changing the location of the pick-up loop or by varying C_1.

C_1, L_1—Resonate to operating frequency.

C_2—0.01-μf. mica or ceramic, 500 volts.

R_1—0.47 megohm. Replaces normal direct connection.

should be coupled to the dummy load, antenna tuner, or transmission line—in other words, to a point in the system beyond where any tuning adjustments are to be made.

The only other piece of test equipment will be an audio oscillator. Since only one frequency is needed, the simple circuit of Fig. 2 works quite well. In fact, many stations have a circuit similar to this one built right into the exciter audio system.

Two-Tone Test

The two-tone test involves sending through the amplifier or the system a pair of r.f. signals

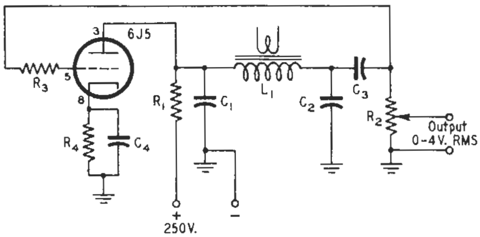


FIG. 2—Fixed-frequency audio oscillator having good output waveform. The frequency can be varied by changing the values of C_1 and C_2.

C_1, C_2—0.02 μf., 600 volts.
C_4—10-μf. 25-volt electrolytic.
R_1—47,000 ohms, 1 watt.
R_2—0.5-megohm potentiometer.
R_3—2.2 megohms, ½ watt.
R_4—1000 ohms, ½ watt.
L_1—Small output transformer, secondary not used.

of equal amplitude and a thousand cycles or so apart in frequency. The combined envelope of two such signals looks like two sine waves folded on one another. If this waveform comes out of the final, well and good; if not, there is work to do. More about that later.

There are two commonly-used ways to generate the two-tone signal, and the choice of which to use depends on the particular exciter. For purposes of this article, the two procedures are designated Method A and Method B, and they are outlined below:

Method A–for Filter or Phasing Exciters:

1) Turn up the carrier insertion until a carrier is obtained at about half the expected output amplitude.

2) Connect an audio oscillator to the microphone input and advance audio gain until (when the carrier and the one sideband are equal) the scope pattern takes on the appearance of full modulation; i.e., the cusps just meet at the center line. See Chart I, photo No. 1.

3) To change the drive through the system, increase or decrease the carrier and audio settings together, maintaining equality of the two signals.

Method B–for Phasing Exciters:

1) Disable the audio input to *one* balanced modulator.

2) Connect the audio oscillator and advance audio gain to get the desired drive. Note that with one balanced modulator cut out, the resultant signal will be double-sideband with no carrier, hence two equal r.f. signals.

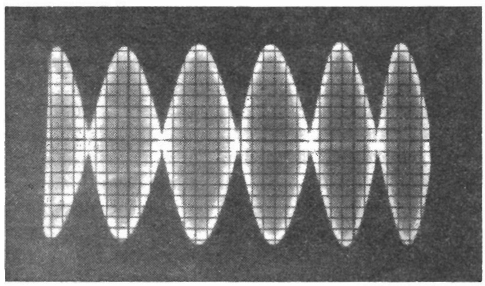
Double-Trapezoid Test

When Method B can be used with phasing exciters, it is possible to derive a somewhat more informative pattern by making a connection from the exciter audio system to the horizontal signal input of the oscilloscope and using this audio signal, instead of the regular internal sweep, to cause the horizontal deflection. Those who are familiar with the regular trapezoid test for a.m. tarnsmitters will recognize this setup as being the same, except that instead of one trapezoid, this test produces two triangles pointing toward each other.

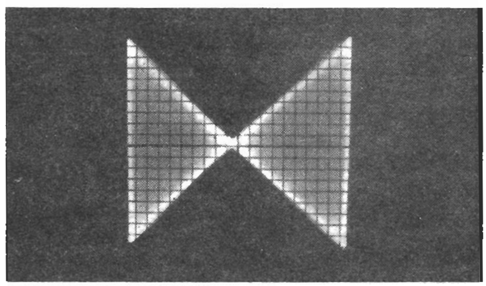
Each individual triangle is subject to the same analysis as the regular trapezoid pattern; i.e., the sloping sides of the pattern should be straight lines for proper operation. Since it is much easier to tell whether a line is straight or not than to judge the correctness of a sine curve, the double trapezoid has the advantage of being somewhat more positive and sensitive to sight departures from linearity than is the regular two-tone pattern.

If the audio can be picked off at the plate of the audio modulator tube that is still working, the input signal need not be a pure sine

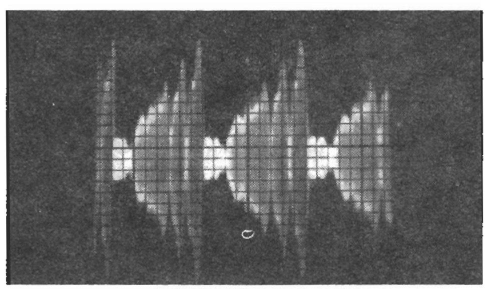
CHART 1—CORRECT PATTERNS



(1) Desired two-tone test pattern.



(2) Desired double-trapezoid test pattern.



(3) Typical voice pattern in a correctly adjusted amplifier, scope set for 30-cycle sweep. Note that peaks are clean and sharp.

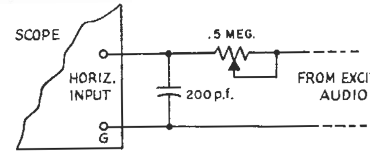


FIG. 3—"Phaser" circuit for the oscilloscope.

C_1—200 pf. or as required.
R_1—0.5-megohm potentiometer.

wave; merely whistling or talking into the microphone should produce the appropriate pattern. If, because of the exciter layout, it is necessary to pick up the audio signal ahead of the phase-shift network, it will then be necessary to use a good sine-wave audio oscillator as before. Also, with the latter setup, the pattern will probably have a loopy appearance at first, and phase correction will be needed to make the figure close up. This can be done either by varying the audio frequency or by putting a phaser in series with the horizontal input to the scope, as shown in Fig. 3.

Ratings

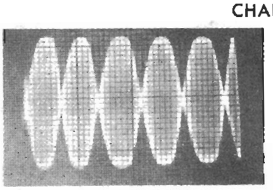
Before proceeding with linearity tests, it is well to have in mind the current and power levels to expect. A suppressed-carrier signal is exactly like an audio signal, except for its frequency, so the audio ratings for any tube are perfectly applicable for linear r.f. service where no carrier is involved. On the other hand, the ratings sometimes shown for Class B r.f. telephony are *not* what is wanted, because they are for conventional a.m. transmission with carrier.

Class B, AB or A can be used. Audio ratings are frequently given for two tubes in push-pull but, unlike audio service, a Class B r.f. amplifier works quite well in a single-ended circuit. Therefore, if the amplifier is to be a single-tube stage, one-half the power and current ratings given for two tubes should be used.

If audio ratings are not given for the desired tube type, it will be safe to assume that the maximum-signal input for Class B or AB_2 service is about 10 per cent less than the key-down Class C c.w. conditions. The input will have to be held somewhat lower in Class AB_1 operation because the average efficiency is lower and, also, the tube can draw only a limited amount of current at zero grid voltage.

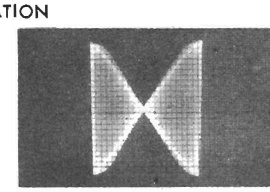
The maximum-signal conditions determined from tube data correspond in s.s.b. work to

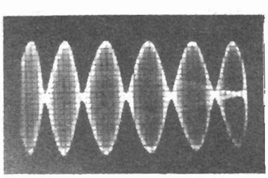
CHART II—IMPROPER AMPLIFIER OPERATION



« (4) Overdrive, indicated by flattening of peaks.

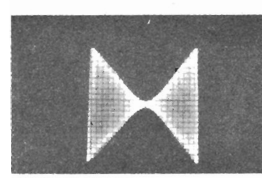
» (5) Same as (4), double-trapezoid test.

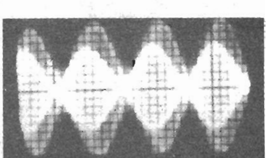




« (6) Too much bias, causing crossover to become pinched together rather than cutting straight across center line.

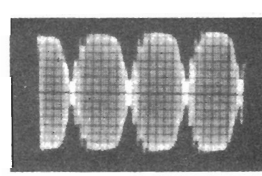
» (7) Same as (6), double-trapezoid test.

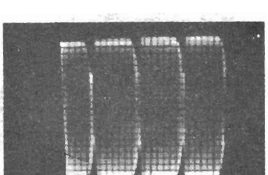




« (8) Two-tone test with v.h.f. parasitics. Note fuzzy halo or fringe. In milder cases the fuzziness will appear just at the peaks.

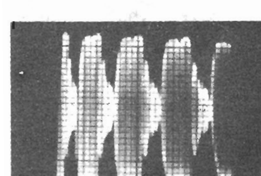
» (9) Two-tone test with fundamental frequency parasitics, accompanied by overdrive.





« (10) Severe overdrive and parasitics.

» (11) Voice pattern showing flattening of peaks due to overdrive. When flattening is apparent on the voice pattern, the case is a severe one.



the very peak of the r.f. envelope. In a correctly-adjusted amplifier, the rated peak input would register on the meters only if one were to whistle into the microphone, otherwise the meters will always read less. In particular, the average input under two-tone linearity-test conditions is close to 65 per cent of the actual peak input for a Class B amplifier, about 75 per cent for a Class AB_2 stage, and 80 to 90 per cent for Class AB_1. With typical voice operation, the meters will kick up only to a smaller fraction of the same peak input—around 30 to 60 per cent for Class B, 50 to 70 per cent for Class AB_2, and approximately 70 to 80 per cent for Class AB_1.

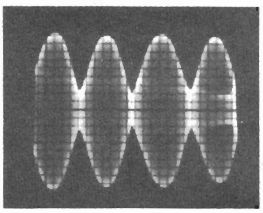
To take a typical example, two 811As are rated for a maximum Class B input of 470 watts. If a single 811A is used in the r.f. final amplifier, its maximum signal input should be 235 watts and, to operate up to this rating, it should be lined up with a linearity test to about 150 watts input. Under normal voice operation, the meter will then read up to around 100 watts.

Using the Linearity Tests

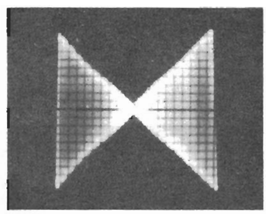
The photos accompanying this article have been taken to show many of the typical patterns that may be encountered with either of the test arrangements described previously. They are classified separately as to those representing correct conditions (Chart I), faulty operation of the r.f. amplifier (Chart II), and various other patterns that look irregular but which really represent a peculiarity in the test setup or the exciter but not in the final (Chart III).

Aside from the problem of parasitics, which may or may not be a tough one, it should be possible without much difficulty to achieve the correct linearity pattern by taking action as indicated by the captions on the photos. It

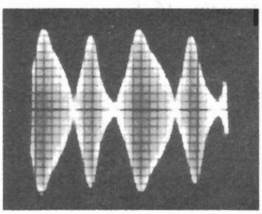
CHART III—IMPROPER TEST SET-UP



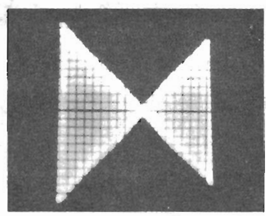
« (12) Two r.f. signals unequal. In Method A, caused by improper setting of either carrier or audio control. Method B, either carrier leakage through disabled modulator or unequal sidebands due to selective action of some high-Q circuit off resonance.



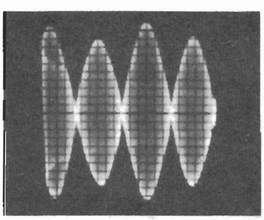
» (13) Same as (12), double-trapezoid test (Method B).



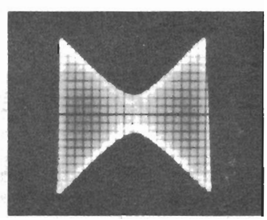
« (14) Distorted audio. A clue to this defect is that successive waves are not identical.



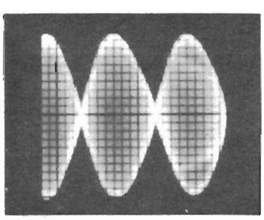
» (15) Same distortion as (14), but switched to double-trapezoid test pattern. Note that correct pattern prevails regardless of poor audio signal.



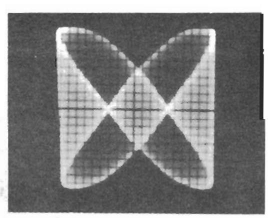
« (16) Carrier leakage through working modulator (Method B only).



» (17) Same as (16), double-trapezoid.



« (18) (Note tilt to left.) Caused by incomplete suppression of unwanted sideband (Method A) or by r.f. leakage into horizontal circuits of scope.



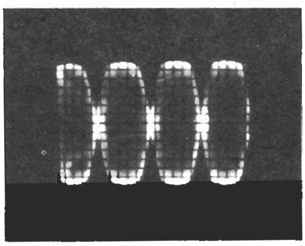
» (19) Double trapezoid with audio phase shift in test setup.

can then be assumed that the amplifier is not contributing any distortion to the signal so long as the peak power level indicated by the test is not exceeded. It is entirely possible, however, that good linearity will be obtained only by holding the power down to a level considerably below what is expected, or conversely that there will be signs of excessive plate dissipation at a level that the tubes should handle quite easily. In such cases, some attention should be given to the plate loading, as discussed below.

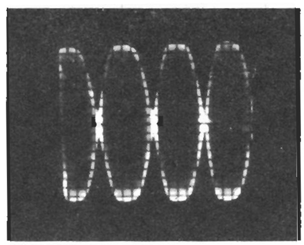
3) If the power level obtained above is less than should be expected, use more driving power.

There are several ways to tell whether or not the final is being driven to its limit. One way is to advance the drive until peak limiting is apparent in the output, then move the oscilloscope coupling link over to the driver plate tank and see whether or not the same limiting appears there. Another way is to decrease or increase the final loading slightly and note whether the limiting output level increases

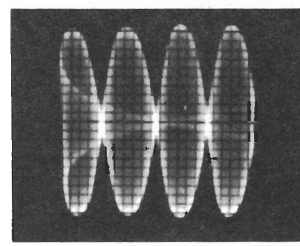
CHART IV—AMPLIFIER LOADING CHARACTERISTICS



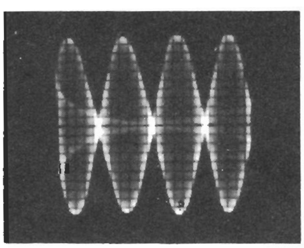
(20)



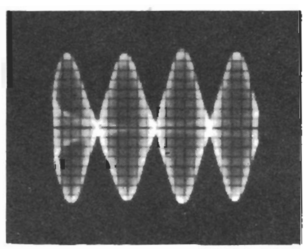
(21)



(22)



(23)



(24)

Two-tone patterns taken at the output of a Class B linear amplifier with constant drive and successively heavier loading. Measured input power is
(20) 90 watts.
(21) 135 watts.
(22) 250 watts.
(23) 330 watts.
(24) 400 watts.

or decreases correspondingly. If it does not, the final is not controlling the system. Still another but similar method is to detune the final slightly while limiting is apparent, and if proper drive conditions prevail the pattern will improve when the amplifier plate is detuned.

The several patterns of Chart IV were made to show how loading affects the output and efficiency of a linear amplifier. In the first two, loading is relatively light and limiting takes place in the final plate circuit. Reserve power is still available in the driver, evidenced by the fact that heavier loading on the final allows the peak output to increase up to the optimum level of the third pattern. With still heavier loading the output ceases to increase but in fact drops somewhat; even though the input power goes up all the time, the efficiency goes down rapidly. In the last two patterns, the driver is the limiting element in the system, and the extra power-handling *capability* of the final, due to heavier loading, is wasted by inability of the driver to do it justice. The following conclusions may be taken:

1) For good efficiency, the final itself must be the limiting element in the power-handling capability of the system.

2) If the final is not being driven to its limit, it should be loaded less heavily until such is the case.

The intermediate and driver stages will follow the same laws, except that the thing called "loading" on a final is often referred to as "impedance matching" when going between tubes. More often than not, an apparent lack of power transfer from a driver to its succeeding stage is due to a poor match. Just as in Class B audio service, a step-down type of coupling is required between power stages, and the person who is accustomed to the conventional plate-to-grid coupling-capacitor technique will be surprised to find how effective it is to tap the driven stage down on its tank—or otherwise to decouple the system. For example, an 807 driving a pair of 811s requires a voltage step-down of about 3 or 4 to 1 from plate to each grid.

Dummy Load

For the sake of everyone concerned, linearity tests should be kept off the air as much as possible. They make quite a racket and spurious signals are plentiful in earlier stages of misadjustment. Ordinary lamp bulbs make a fine dummy load so long as it is recognized that their impedance is not exactly the same as the antenna and that this impedance changes somewhat as the bulbs light up. These factors can be taken into account by making careful note of plate and grid currents after the transmitter has been adjusted and is operating with a linearity test signal at maximum linear output into the lamp load. Then, having reconnected the regular antenna, the same loading conditions for the final will be reproduced by adjusting its tuning and loading until the identical combination of plate and grid currents can be obtained. This process will require only a few moments of on-the-air operation.

Conclusions

When the final on-the-air checks are made, it will be convenient to make a few reference marks on the oscilloscope screen to indicate the peak height of the pattern. The scope will then serve as a permanent output monitor for all operations. For best results the sweep adjustment should be set for about 30 cycles, in which case the voice patterns will stand out clearly and can easily be kept just within the reference lines. Incidentally, the scope pattern is really fascinating to watch.

The writer wishes to acknowledge with thanks the kind assistance and suggestions offered by C. B. Grady, W2SNQ, in making the photos for this article.

» 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: AB_1, AB_2 or B.

From *QST*, August, 1955.

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 gratifynng number of amateurs are now using scopes, it would be unrealistic to think that all s.s.b.

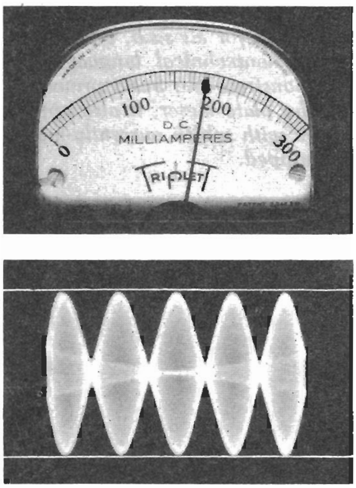


FIG. 1

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.

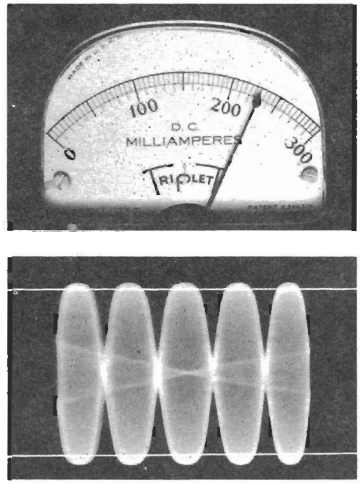


FIG. 2

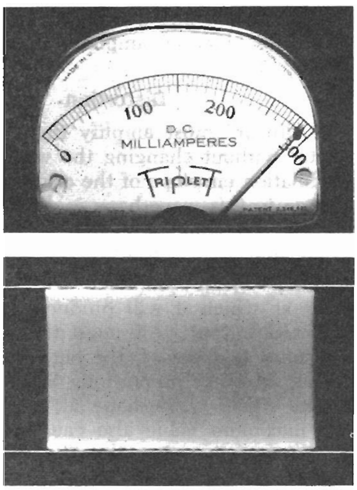


FIG. 3

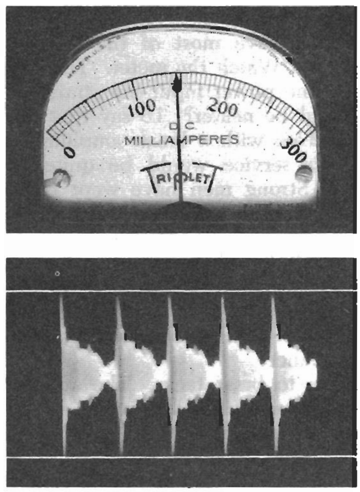


FIG. 4

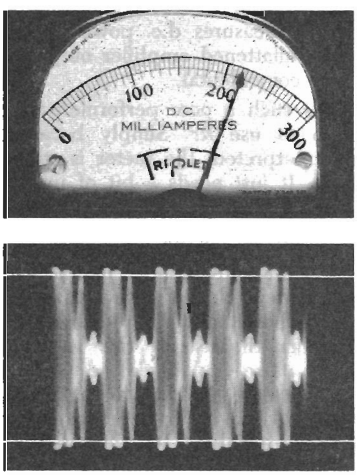


FIG. 5

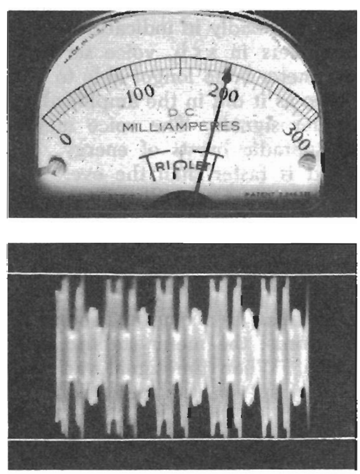


FIG. 6

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.

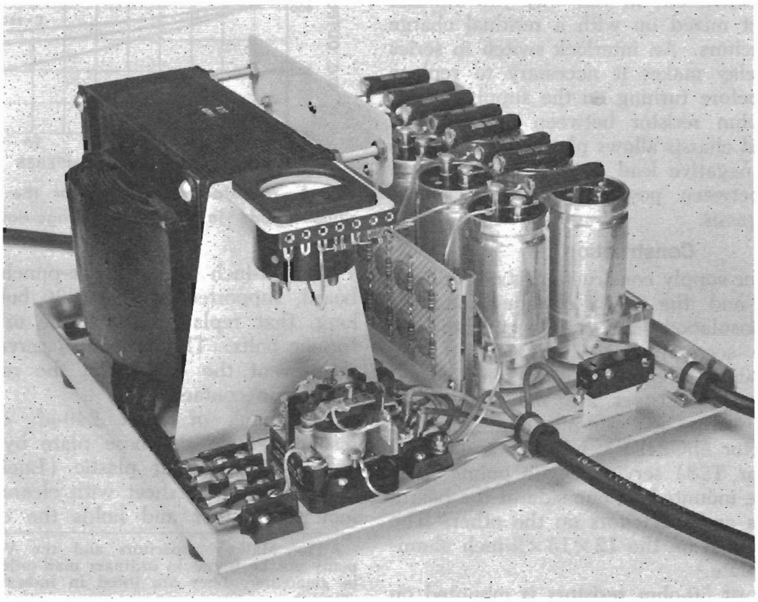
Now, regardless of the power I believe my amplifier should handle, I make a crude check to determine at what point it actually flattens. I vary the amount of carrier insertion, watching mainly the output indication, and note the point at which increasing the carrier no longer results in a rapid increase 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 ma. × 1500 volts). My maximum signal or peak envelope power is 375 watts (250 ma. × 1500 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.

» Silicon rectifiers have changed the complexion of power supplies in amateur equipment—at all voltage levels. Here's a 3000-volt supply that will easily deliver a d.c. kilowatt, using the voltage-doubler circuit with a string of electrolytic filter capacitors.



The semiconductor power supply uses a string of silicon diodes and a bank of 450-volt electrolytic capacitors. The diodes are mounted on the pre-punched terminal board mounted between the transformer and the capacitors; in this view only the equalizing capacitors and resistors can be seen. The small switch at the right foreground is the interlock.

Power Supply for a Kilowatt Linear

BYRON GOODMAN, W1DX

The power supply shown in the photographs does a beautiful job of supplying the necessary plate power for a 3-1000Z grounded-grid linear amplifier (3000 volts at 330 ma. indicated peaks, 600 ma. or so actual). The power supply could have been built with a couple of silicon rectifier units rated at 4000 volts p.i.v. (and costing $18 each), but instead it was built with 16 type 1N1764 500-p.i.v. units costing 84 cents each (a total of $13.50). The 1N1764 is rated at 500 ma. d.c. up to 75° C.

Equalizing resistors are used to wash out minor variations in back resistances, and bypass capacitors to protect against voltage spikes.[1] Half-megohm half-watt resistors and 0.01-μf. disk ceramic capacitors are connected across each silicon diode.

Filter Capacitors

The filter has an effective capacitance of 30 μf., obtained by connecting eight 240-μf. 450-volt electrolytic capacitors in series. This filter is quite adequate for use with a sideband amplifier. The indicated voltage drops from 3000 at idle (180 ma.) to not less than about 2800 on voice peaks (350 ma.).

The Practical Circuit

The circuit for the actual power supply is shown in Fig. 2. The transformer is made by the Berkshire Transformer Co., Kent, Conn., and is available directly from them for $50. In service with the 3–1000Z amplifier it is hardly warm to the touch, testifying to its efficiency and conservative ratings. Dual primaries are included, to permit operation from either a 115- or a 230-volt line. The higher voltage is recommended.

The filter capacitors are called "computer grade" capacitors; the 25K resistors across them serve both as the bleeder resistor and the equalizing resistors. In operation, the idling

From "Inexpensive Power Supply for a Kilowatt Linear," August, 1963, *QST*.

[1] Geiser, "Semiconductor Rectifiers," *QST*, July, 1961.

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 chassis and amplifier chassis.

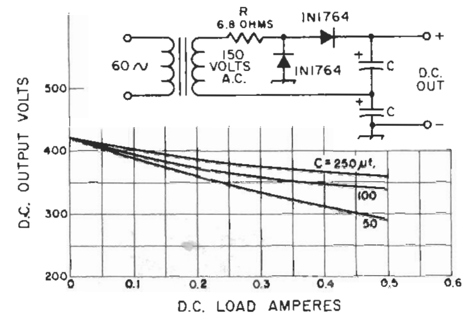


FIG. 1—Manufacturer's curve for the RCA 1N1764 silicon diode in full-wave voltage-doubling circuit.

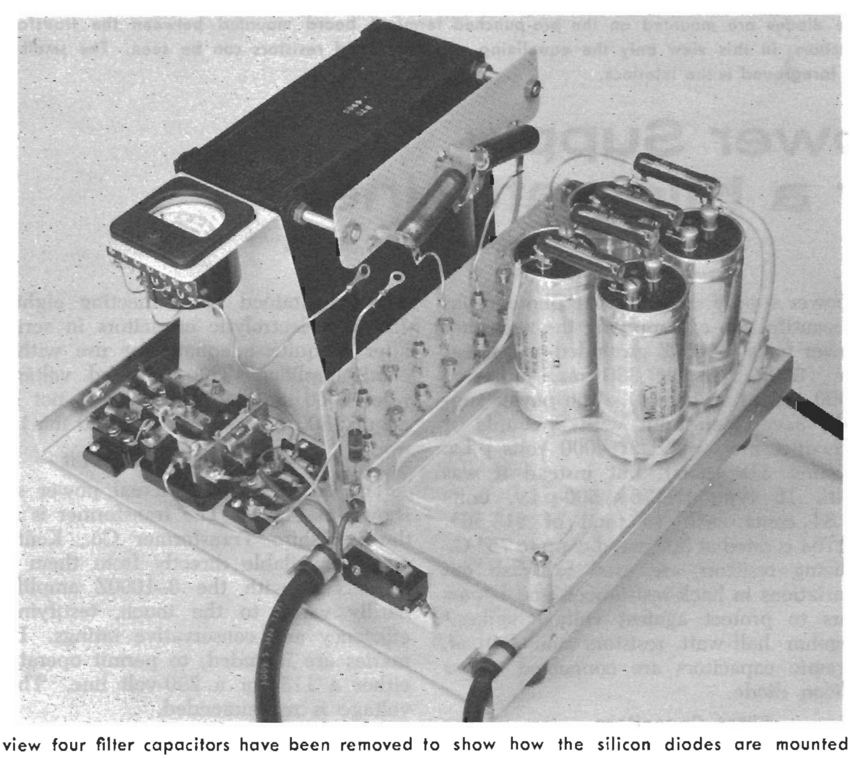
Construction

The power supply construction is not critical, of course, and the main considerations are adequate insulation and safety precautions. As can be seen from the photographs, the string of silicon diodes and their associated capacitors and resistors are mounted on a 3 × 9¼-inch strip of pre-punched 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 × 13 × ⅛-inch aluminum base plate.[2]

The pair of 50-ohm resistors is mounted on a 7½ × 1¾-inch strip of pre-punched terminal board, supported by two ¼-20 bolts, 5 inches long, that replace two of the original transformer bolts. This strip also serves as a stop to prevent the cover and the resistors from coming in contact.

The bank of eight 240-μf. capacitors is insulated from the base plate by a sheet of 4¾ × 9 × ¼-inch clear plastic (Lucite or Plexiglas). A similar sheet with clearance holes is mounted higher and holds the capacitors in

[2] The Mallory capacitors and the Vector products many not be found in ordinary mail-order catalogs sent to amateurs. They are listed in industrial electronics catalogs.



In this view four filter capacitors have been removed to show how the silicon diodes are mounted on the terminal board. The meter mounting bracket is held to the base plate by two of the bolts that run through the feet of the transformer. Normally a perforated metal cover protects the power supply; there is a hole cut for observation of the voltmeter.

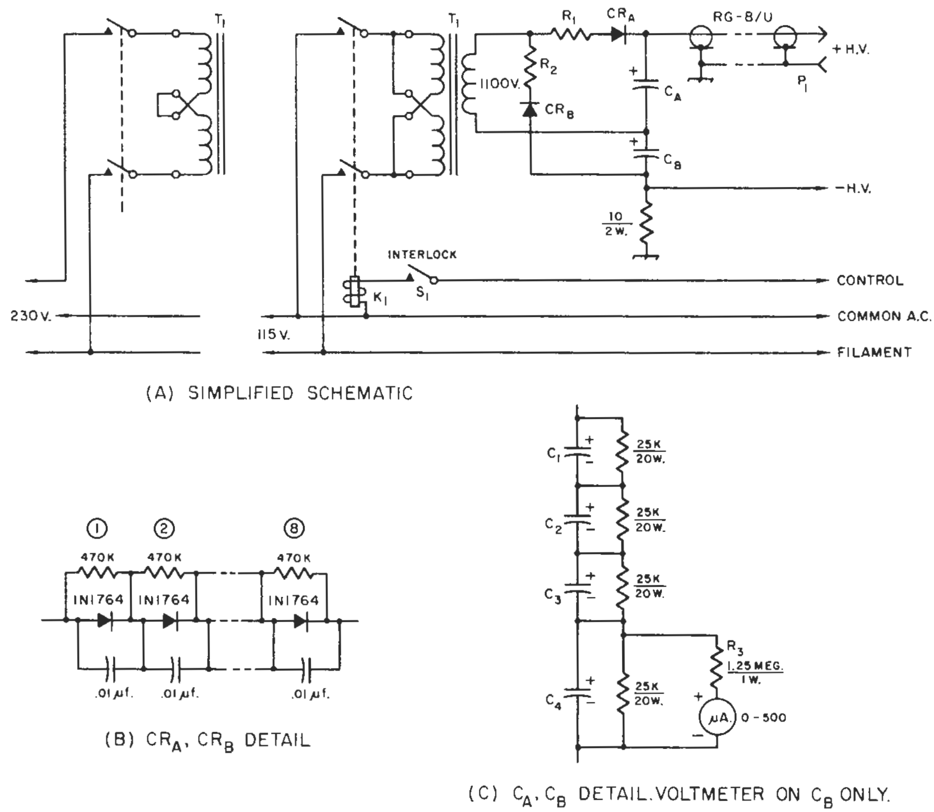


FIG. 2—Schematic diagram of the 3000-volt power supply.

C_1-C_4—240-μf. 450-volt electrolytic (Mallory CG41T450D1).
K_1—D.p.s.t. relay, 25-ampere contacts (Potter & Brumfield PR7AY, 115-v.a.c. coil).
P_1—Coaxial plug, UG-59B/U (Amphenol 82-804).
R_1, R_2—50-ohm 25-watt wirewound (Ohmite 0200D).
R_3—Selected 0.47- and 0.68-megohm, ½ watt, in series.
S_1—S.p.s.t. miniature switch (Acro BRD2-5L).
T_1—1100-v., 0.3-amp. transformer, dual primary (Berkshire BTC-4905). See text.
25K, 20-watt resistors are Ohmite Brown Devil 1845, 470K resistors are ½-watt, 0.01-μf. capacitors are 1000-volt disk ceramic.

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 (we used a cable marked "14-3 Type SJ 300 V), and No. 16 wire will suffice for the control wiring.

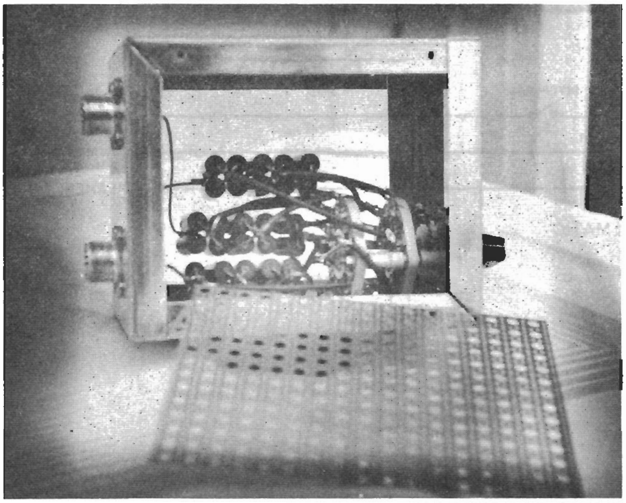
If desired, a precision resistor can be used for R_3, the voltmeter multiplier. However, we merely selected standard 20-percent resistors until we had the value that gave a correct reading.

Initially we had some slight fear that turning on the power supply would result in surge currents that would exceed the ratings of the diode rectifiers. However, measuring the surge current with an oscilloscope indicates that the surge current runs about 12 to 15 amperes, well below the ratings. The 50-ohm resistors and the reactances of the transformer windings do a good job of limiting the surge without causing excessive regulation.

Safety Precautions

It should be unnecessary to point out that a 3000-volt power supply with a 30-μf. 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.

» *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 AB_1, 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 was indicated. A continuously-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.

From December, 1959, *QST*.

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

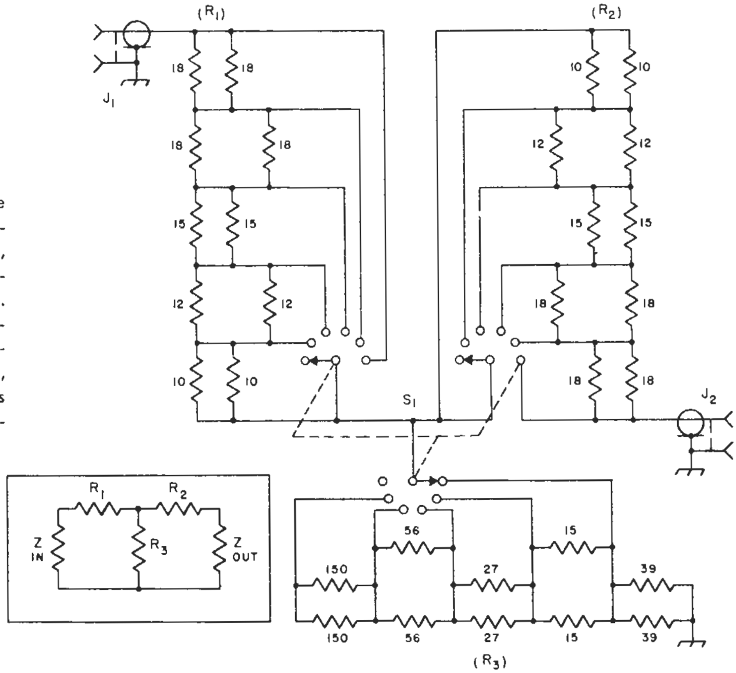


FIG. 1—Circuit of the T-network attenuator. Resistances are in ohms, and resistors are 10-percent, 2-watt composition. J_1 and J_2 are chassis-mounting coax receptacles (SO-239). The switch, described in the text, is in the maximum-attenuation position.

legs of the pad for various levels of attenuation are given in the accompanying table. The table includes both the calculated values and the actual values of standard resistors that were used.

The resistors are 2-watt types made by Ohmite or Allen-Bradley. Since the calculated values were below 10 ohms in many cases, the required resistance was obtained by paralleling two resistors. Ten ohms is the lower limit for these carbon resistors. Be careful not to use wire-wound units since they may have sufficient inductance to make the attenuator useless at radio frequencies. The paralleled values also result in sufficient dissipation rating to handle the full output of the Pacemaker over the short duty cycle of s.s.b.

Construction

The box in which the resistors and switch are mounted is a Premier AC-453, 3 by 4 by 5 inches in size, with two removable 4 × 5-inch sides. The original sides are replaced by perforated aluminum sheet or aluminum screen, for ventilating purposes. Two coax connectors are mounted in one 3 × 4-inch end, about 2 inches apart, and the switch is mounted in the opposite end, as shown in the photograph.

The switch is special, made up of a Centralab index kit P-270 and two ceramic sections, one having two-poles and five positions, while the other has a single pole and six positions. The two-pole section actually has six positions, one of which is "off," where the rotor contact is made. The index stop is adjusted to use this sixth position in the maximum-attenuation position. Hardware furnished with the index kit includes sufficient ½-inch spacers so that the two switch sections may be assembled on the index. Place the single-pole section next to the index and the double-pole section away from it. The sections used may be either types TD and RD (shorting) or XD and RRD (nonshorting).

Atten. (db.)	R_1, R_2 *Cal.*	R_1, R_2 *Actual*	R_3 *Cal.*	R_3 *Actual*
3	8.5	9	144	143.5
6	16.7	18	67	68.5
9	24	25.5	40.5	40.5
12	30	31.5	27	27
15	35	36.5	18.5	19.5

Assembly of the resistor bundles shown in the photograph is best accomplished with the aid of the wooden jig illustrated in Fig. 2. In a small piece of 1-inch soft pine (actually about 13/16 inch thick), 1½ by 8 inches, drill ten 5/16-inch holes in a two-hole by five-hole pattern. Put the holes close enough together so

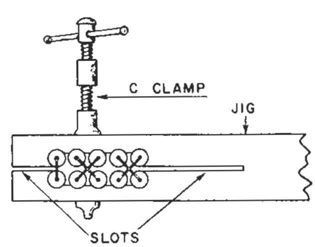
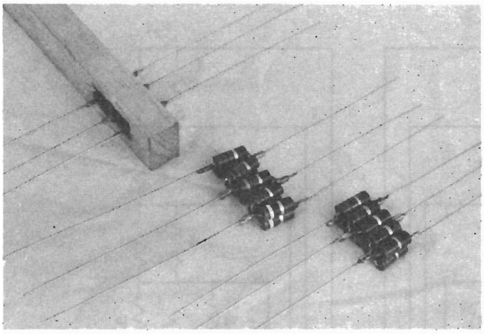


FIG. 2 — Sketch showing wood jig and clamping arrangement for assembly of resistor bundles.



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 ½ 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 AB_1, 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!

» There is no need to spell out the advantages of voice-controlled break-in operation to any active s.s.b. man. Here is a good circuit for sure-fire anti-trip and VOX operation.

"Universal" Voice-Control Circuit

L. O. LEIGH, KT1LS

In the quest at KTILS for a solution to the VOX problem, we have been trying to find a simple circuit which would satisfy the following conditions:

1) The circuit should not require critical adjustments of relays or levels for proper operation.

2) Blocking biases to the transmitter and receiver should be electronically developed, have limits from absolute zero to a definite voltage and have reasonable power capabilities.

3) The operation of relays for auxiliary control of antenna change-over relays, etc., should be accomplished without extra tubes, and the relay coil currents should have limits from zero to a fixed value to eliminate the necessity for delicate relay adjustments.

4) The anti-trip part of the circuit for loudspeaker break-in should be positive and not affect or alter the time delay constant on the transmit side.

From *QST*, November, 1956.

5) It should do all this without resembling an electronic computer.

Circuit

The circuit finally evolved is shown in Fig. 1. In the static state, with no voice input, there is practically zero voltage between the grid and cathode of tube V_{3A} and this half of the 12AT7 draws a fairly heavy plate current of about 6½ ma. This current develops about 50 volts to ground across its plate resistor, which is used to block off the transmitter and, at the same time, drops about 6½ volts across the 1K common cathode resistor of V_3, so that V_{3B} is completely cut off and develops no voltage across its plate resistor, which supplies the receiver blocking bias.

When speech is applied, the condition changes abruptly. Voice voltages applied to the input of transformer T_1 are amplified by V_{1A} and rectified by diode V_{2A}, causing a voltage to be built up across the timing circuit, R_1C_1,

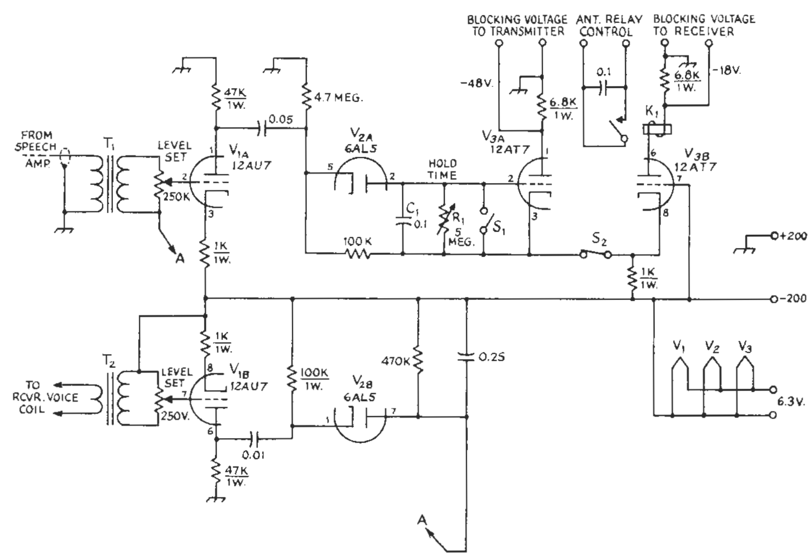


FIG. 1—Schematic diagram of the "universal" voice-control circuit. All capacitances are in μf. All resistors are ½ watt, unless otherwise specified.

S_1, S_2—S.p.s.t. toggle switches. See text for functions.
T_1—Interstage transformer.
T_2—Low-power tube-to-voice-coil transformer.
Values of T_1 and T_2 are not too important, as pointed out in the text.

which biases V_{3A} to complete cut-off. With V_{3A} cut off, the voltage across its plate resistor disappears, unblocking the transmitter and, at the same time, the voltage on the grid of V_{3B} falls below cut-off because of the reduced current in the common cathode resistor. In this condition, V_{3B} draws current developing about minus 20 volts to ground across its plate resistor, which can effectively squelch the receiver. The positive action of V_3 is an adaptation of the familiar flip-flop circuit.

Provided the gain of amplifier V_{1A} is great enough to give decisive cut-off on V_{3A} on voice pulses, a condition easily met, the hold-in on transmit is determined by the time constant of R_1C_1.

The anti-trip part of the circuit is equally simple. Receiver audio voltage, in our case taken from across the voice coil of the receiver, is supplied to the grid of amplifier V_{1B} via transformer T_2. Amplified, and then rectified by V_{2B}, this develops a negative bias which is applied to the grid return of the speech amplifier V_{1A} in the same manner as an a.v.c. voltage. Thus, in effect, the gain of V_{1A} is automatically regulated by the speaker audio volume, giving a surprisingly non-critical and reliable anti-trip action which is completely isolated from the transmit-receive time delay circuit.

Because the blocking voltages change fast but not instantaneously, the circuit flips from one condition to the other smoothly without the thumps or clicks often troublesome in relay systems.

Antenna Changeover

Although we have hopes at KT1LS of some day eliminating the mechanical antenna changeover relay, for the time being, as in many stations, it is still a necessary part of the system which must be voice-control operated.

This is accomplished by placing an auxiliary relay (K_1) in series with the plate of V_{3B}. Here the current goes from complete zero (on receive) to about 2½ ma. (transmit), which is enough to close a 15-mw. sensitive relay having a coil resistance of about 2000 ohms. The contacts of the sensitive relay operate the 110-volt a.c. antenna change-over relay. This combination has worked out to be completely satisfactory and considerably faster than it sounds. Another possibility, although not tried, would be to connect a relay having normally-closed contacts in series with V_{3A} to operate the antenna relay. This would give 6½-ma. relay-coil current. One point of caution: We had trouble originally with sparks across the contacts of the auxiliary relay K_1 causing false action of the voice amplifier and tripping the circuit. It was easily fixed by shunting the auxiliary contacts with a 0.1-μf. capacitor.

General

There are a few other details about the circuit which may be helpful to the prospective user. The 4.7-megohm resistor from the cathode of V_{2A} to ground provided a slight threshold voltage to overcome random noise from the preceding amplifier and external sources. In our case R_1 is a variable resistor, but a satisfactory time constant is also afforded by the use of a 3-megohm fixed resistor. For satisfactory operation of V_{1A}, the peak-to-peak voltage (by scope measurement) across its plate resistor should be a minimum of 100 volts on voice crests. This test point was chosen because one side is grounded and the requirement is not affected by difference in the turns ratio of T_1. Likewise, the peak-to-peak voltage across the plate resistor of the anti-trip amplifier V_{1B} will be approximately 150 volts minimum value for reliable anti-trip action.

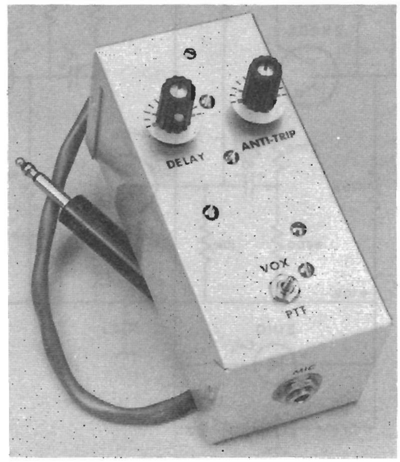
Transformers T_1 and T_2 have been found to be essential for the isolation of the grids of amplifiers V_{1A} and V_{1B} and the successful operation of the circuit. The transformers themselves are not critical; almost anything will do. For T_1 we are using the small output transformer of a surplus BC-453 unit. T_2 is an ordinary plate-to-speaker transformer. Because of the inverted power-supply operation of the voice-control circuit, resistor-capacitor coupling of the grids of V_{1A} and V_{1B} to external circuits results in excessive noise and hum pickup.

Toggle switches S_1 and S_2 are added to provide the possibility of manual operation of the control circuit when desired. To operate, S_1 is closed to disable the voice circuit, and S_2 is closed for receive and opened for transmit.

The power supply used here came from the junk box and is actually much larger than required. The circuit is entirely conventional except, of course, the plus side is grounded. At 200 volts, the maximum drain of the voice control unit is only 12 ma.

Putting the circuit in operation is very simple. The voice level-set control is turned up so that reliable voice operation results, the time-delay control R_1 is set for the desired hold-in and the anti-trip level-set control is adjusted to allow uninterrupted listening. From there on, it is mostly automatic. If the speaker is turned up very high or the microphone is very close, you may have trouble shouting down the receiver's audio but, once you do, the receiver goes dead and normal voice levels are in control. There are no false trips by the receiver's output.

» *Stuck with manual operation of your present mobile gear? This little voice-control unit lets you have both hands free for operating the car instead of the radio.*



The completed VOX unit goes between the microphone and the transmitter. The miniature knobs are Johnson Collet type 116-603.

Transistorized VOX in a Box

E. LAIRD CAMPBELL, W1CUT

The little gadget shown in the photographs is designed to give voice-operated break-in (VOX) capability to transmitters that now have only push-to-talk operation.

This VOX unit is not restricted to fixed-station use. In fact, its logical application is for mobile work—even for a.m.—especially from a safety point of view.

The Hookup

Operation of the VOX circuit is simple. Audio from a high-impedance microphone is amplified in several transistor stages, rectified, and applied to the base of a transistor that operates a relay. Contacts on the relay are connected to the push-to-talk circuit of the transmitter. Once the relay has closed, it will hold in for any desired amount of time, up to several seconds. In Fig. 1, transistor Q_1 is operated as an emitter follower to present a high impedance to the microphone and to act as a relatively low-impedance source for driving Q_2. Transistors Q_2 and Q_3 are audio amplifiers. Audio output from Q_3 feeds into the VOX rectifier, CR_2, which is part of a control circuit similar to that described by W3UWV several years ago.[1]

The negative bias developed at R_1 is applied to the base of Q_4 through CR_4. This increases Q_4's collector current and closes the relay, K_1. Diode CR_4 acts as a gate to prevent any positive-going signal from getting to the base of Q_4.

To prevent signals from the shack speaker from triggering the VOX, an anti-trip circuit is built in. Some of the output from the receiver (which can be taken from the speaker connection at the receiver) is rectified by CR_3, which is connected so that it produces a positive bias to buck the negative bias from CR_2, developed through the VOX stages.

Transistors used in this circuit can be most any of the available small-signal audio types. The ones shown here were chosen because they are all available for about 35 cents each.

Power for the VOX unit is a 15-volt battery, BT_1, regulated at 10 volts by a Zener diode, CR_1. It was found to be absolutely necessary to use the Zener diode, especially in mobile service, since the relay hold-in delay time will change with battery voltage. A ¼-watt unit can be used instead of the one specified. If the VOX device is to be used exclusively for mobile work, the car battery can be used instead of the dry-cell battery. The circuit is designed for voltages between 12 and 15 volts and for either positive or negative battery grounds.

Construction

The case is a Minibox that measures 2¼ × 2¼ × 5 inches (Bud 3004A). Close inspection of the photographs will show where most of

From March, 1964, *QST*.

[1] Packham, "A Transistorized Control Unit," *QST*, November 1955, p. 32.

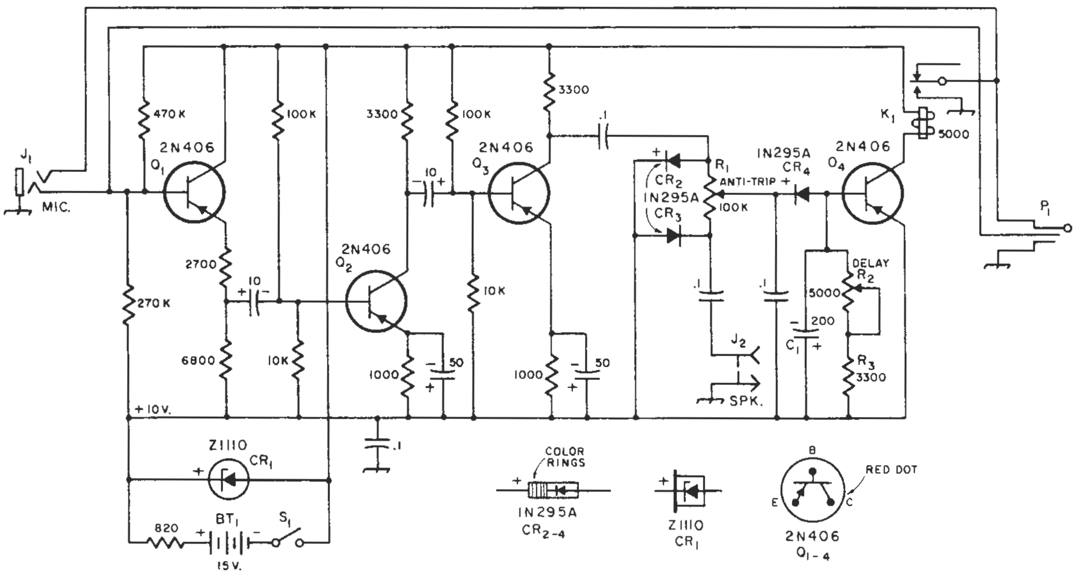


FIG. 1—Circuit diagram of the VOX unit. Capacitances are in μf.; resistances are in ohms; resistors are ½-watt.

BT_1—15-volt battery (Burgess K10).

C_1—200-μf. subminiature electrolytic capacitor (Sprague TE-1119.6).

CR_1—10-volt Zener diode (International Rectifier Z1110).

CR_{2-4}—1N295A crystal diodes.

J_1—3-conductor military type phone jack (Switchcraft C-12B).

J_2—Phono jack.

K_1—5000-ohm relay (Advance RC1C500D or Argonne AR-21).

P_1—3-cond. military type phone plug (Switchcraft 480).

Q_1-Q_4, inc.—2N406 transistors.

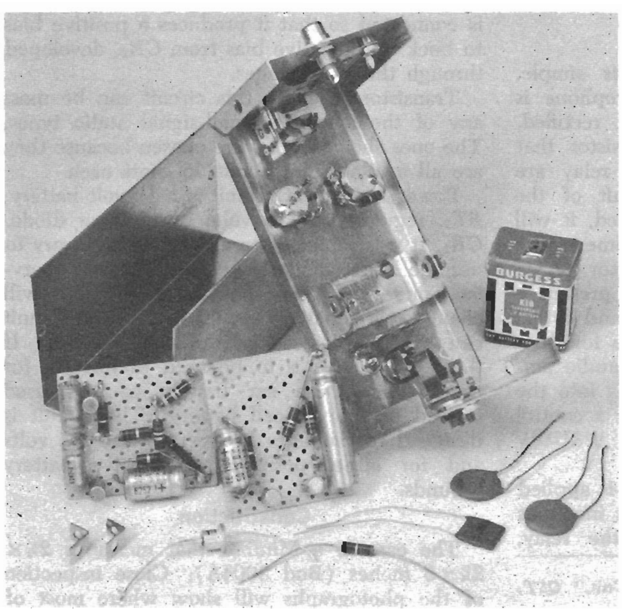
R_1—100,000-ohm miniature control (Mallory MLC-15L).

R_2—5000-ohm miniature control (Mallory MLC-53L).

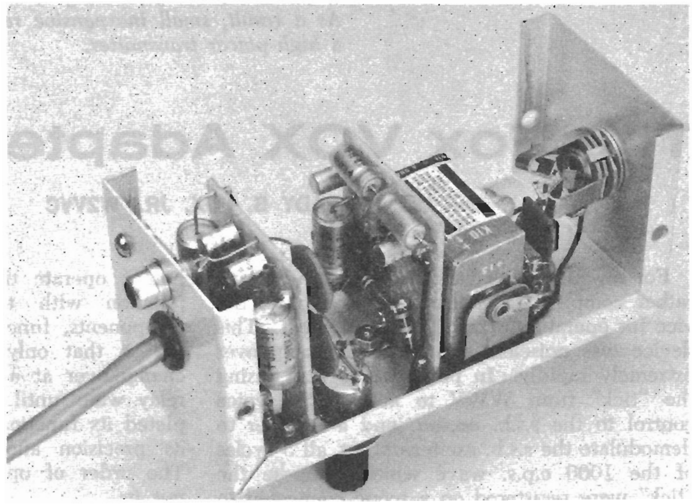
R_3—3300-ohm, ½-watt resistor.

S_1—Miniature toggle switch (Lafayette SW-76).

the components are mounted. Conventional construction and layout can be used in a larger chassis or box. Only two lug-type terminal strips (H. H. Smith 830) were necessary. One is a strain reliever for the output cable and the other is a tie point for mounting the Zener diode. The battery holder is a modified Keystone type 166. Originally, this holder had a



This view shows the VOX unit in the final stages of completion. Starting at the lower right of the chassis, the ports attached to the box are the phone jack, toggle switch, three-terminal tie point, battery holder, the two miniature controls, relay, three-terminal tie point, and phono connector.



The finished VOX unit with its cover removed takes on a compact look, although a large part of the space inside the chassis actually is taken up by the battery and its holder. This view also shows the phono connector and the output cable.

spring clip on both sides to help hold the battery in place. However, the battery used here is too wide for the holder and the side clips must be removed. The end clips with the terminals have sufficient holding power to keep the battery in place.

Most of the components—resistors, capacitors, transistors, and diodes—are mounted on 1¾×2-inch prepunched terminal boards (Vector 85G24EP). The boards are attached to one side of the Minibox case (see photo on page 235) with small angle brackets (General Cement H570-F). All of the electrolytic capacitors used here are Sprague type TE 10-volt subminiatures. Layout of the components on the terminal boards is not critical, except from a mechanical standpoint. That is, junctions and connections should be arranged so that it will be convenient to make board-to-board or board-to-external-component connections.

The two controls, DELAY and ANTI-TRIP, must be insulated from the Minibox chassis if their cases are used as tie points as shown in the "exploded" photo. This can be done easily by using extruded fibre washers with ¼-inch holes (General Cement 6528-C) and flat fibre washers with ¼-inch holes (General Cement 6516-C). Finally, the 5000-ohm relay is attached to the Minibox with its own mounting screw. The relay is designed for use in radio-controlled models and has a pull-in current of about 1.5 ma.

Just Plug It In . . .

Using the VOX gadget is a simple matter of plugging the microphone into the VOX unit and plugging the VOX cable into the microphone jack of the transmitter. There are no gain controls on the unit; it runs wide open all the time. We got our unit to close with as little as 3 millivolts input. Since most high-impedance microphones have at least 10 to 20 millivolts output, there should be no driving problem.

Delay between the time of the last word spoken into the mike and the time the relay opens can be adjusted from almost zero to several seconds with control R_2. The time constant is determined by the value of capacitor C_1 and the resistance, R_2R_3, across it. It may be necessary to juggle these values around somewhat to get the desired range of delay.

» By proper sequencing of the operation of transmitter-power and antenna relays, arcing at the antenna relays is avoided. As a result, small inexpensive relays may be used even with a high-power transmitter.

The Fox VOX Adapter

GRADY B. FOX, JR., W2VVC

For several years I have been using a unique, rather complex transmitter-receiver control device for operation of my s.s.b. equipment. This device was noiseless and did the change-over extremely rapidly. In performance tests, using the "tick" from WWV to actuate the voice control to the s.s.b. exciter, and a receiver to demodulate the s.s.b. exciter output, all 5 cycles of the 1000 c.p.s. wave which compose the "tick" were registered on a scope connected to the receiver output.

Electronic T.R. Systems

Despite the noiseless and rapid performance of this circuit, it lacked one feature necessary for satisfactory voice break-in operation. It did not provide for t.r. operation of the transmitting antenna. Many operators would have been satisfied with the operation of an electron-type t.r. device in this assignment. But the manner in which such devices are used makes the arrangement prone to unpredictable performance with regard to signal loss suffered when QSYing the receiver within a given band, with random final-amplifier tuning adjustments.[1] If one wants the best signal to receiver-noise ratio, it seems that the only really satisfactory antenna t.r. device to date is the old metallic-contact gadget, the s.p.d.t. relay.

Relay Change-Over

On-the-air listening to various s.s.b. signals showed that most relay-operated voice-control system perform with no noticeable clipping of the first syllable. However, there seems to have been little concern by designers about the timing precision desirable in the sequence of operation of the relays used to control the receiver, change over the antenna, and control the transmitter. Inattention to the relay-sequence problem can result in quite fat arcs at the antenna relay contacts from time to time.

Improving the Relay System

An adapter which will provide improved performance of the usual relay-type voice-control system has been developed and used for several months. It uses three carefully-sequenced relays for receiver control, antenna change-over and transmitter control. The inherent operate time of these relays, in conjunction with their associated time-control components, functions to cause them to operate so that only one relay is in process of change-over at a time. In other words, each relay waits until the preceding one has completed its function. The circuit is beautiful in its precision and electron-tubeless simplicity. The order of operation from listen to transmit is:

1) Turn off receiver.
2) Change antenna from receiver to transmitter.
3) Turn on transmitter.

From transmit to receive, the reverse sequence of events takes place. This is not the order of operation usually obtained when one relay is used to control one or more additional relays.

Causes of Arcing

In the conventional arrangement where relays are operated in tandem, the operator may wonder why he still gets arcs and contact burning at the antenna relay even though he may have interlocked the contacts and done some contact bending in an effort to get the sequence of operation correct. The trouble he is experiencing may be due to antenna relay-contact bounce on make. The antenna change-over contact has simply not stopped bouncing before the transmitter starts delivering r.f. energy to the still-bouncing relay contact. The proposed arrangement takes this contact bounce time into account and delays r.f. output from the transmitter until contact is firmly established between the transmitter and antenna.

Antenna Relays

One advantage of a properly-sequenced control-relay arrangement is the ability to use almost any small relay to do the antenna change-over job. Since the antenna change-over contacts do not switch under power, almost any but the frailest relay contacts will handle full amateur power, if the r.f. switching is done at a low-impedance point as it usually is. The principal restriction on relay selection is that the dielectric on which the relay contacts are mounted be able to withstand the applied r.f. voltage. This requirement is not one of arc-over but that of dielectric heating caused by high-frequency currents.

From November, 1960, *QST*.

[1] Campbell, "Variations in TR Switch Performance," *QST*, May, 1956.

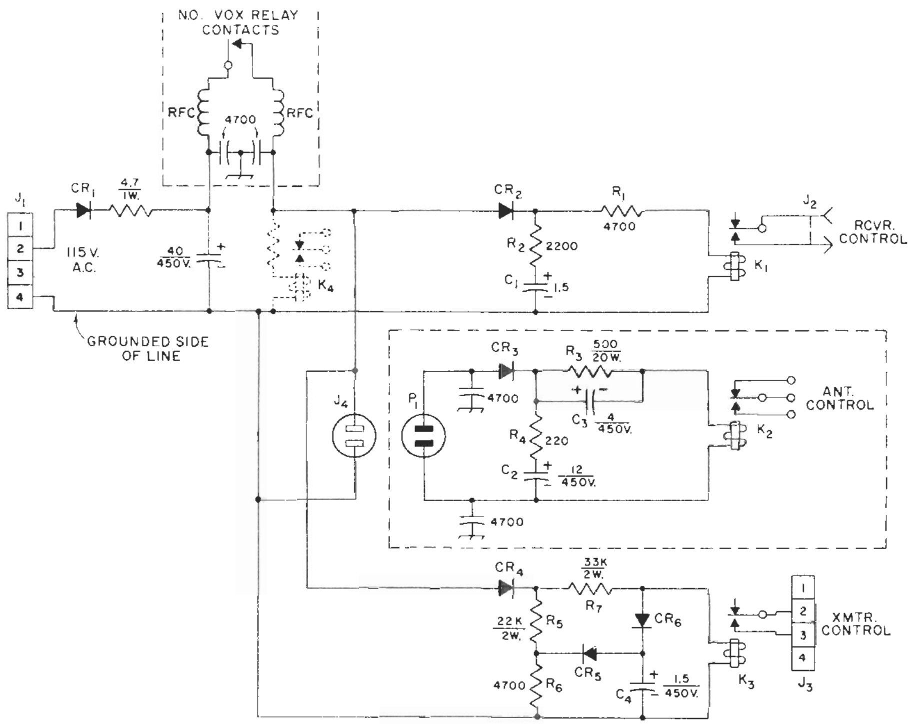


FIG. 1—Diagram of the sequenced control circuit. Resistances are in ohms. Capacitors marked with polarity are electrolytic and capacitances are in μf. (see text in regard to C_1 and C_4); others may be paper or ceramic and values are in pf. Resistors are 10-per-cent ½-watt unless indicated otherwise. Antenna-control portion (shown within dashed lines) should be installed at the transmitter. Note: pin 4 of J_1 must go to grounded side of line.

C_1-C_4 inc.—450-volt electrolytic (see text).
CR_1-CR_6, inc.—Silicon diode: 360 p.i.v., 200 ma. (Sarkes-Tarzian K-200).
J_1—Chassis-mounting 4-prong male connector (or other connector whose connections may be polarized) (Cinch-Jones P-304-AB).
J_2—Phono connector.
J_3—Chassis-mounting 4-prong female connector (Cinch-Jones S-304-AB).
J_4—Miniature polarized chassis-mounting a.c. outlet.
K_1, K_3—S.p.d.t. relay, 600-ohm coil (Sigma 11F-6000-G/SIL).
K_2—Antenna relay from ARC-5 antenna-tuning unit, or similar (see text).
K_4—See text.
P_1—Polarized a.c. plug.
R_1-R_7 inc.—See text.

Dielectric heating from the r.f. current can cause insulation breakdown where the same d.c. or 60-cycle voltage would cause no trouble. In particular, relays which have the movable contact connected to the relay frame are susceptible to dielectric heating. The construction promotes heating of the relay coil due to the effect of r.f. current flowing from the core piece of the coil through the coil insulation to the coil winding.

Two very commonly-available relays have been satisfactorily used as antenna changeover relays in a sequenced control system. Either one will handle any legal amateur power into the usual coaxial cable. My favorite of the two is a little gem which I have wanted to put to work for some time. It is the relay from the BC-442 antenna tuning unit of the ARC-5 and SCR-274 series. There must be thousands of these little beauties which have been relegated to the attic by amateurs who bought tuning units to get the vacuum capacitor and thermocouple meter. They are still available on the surplus market at a reasonable price. This relay is beautifully insulated for high r.f. voltage. It has a set of s.p.d.t. contacts for r.f. switching and a set of normally-open control contacts, one contact of which is at relay-frame potential. The coil of this relay is designed for 24/28-volt d.c. operation. The other relay which has been proven in the antenna changeover circuit is a commonly-available surplus d.p.d.t. relay. Some of these also have a set of s.p.d.t. control contacts. Both Advance and Leach have manufactured a relay of this type. The contacts are insulated with ceramic and the coil is intended for 115-volt 60-cycle operation.

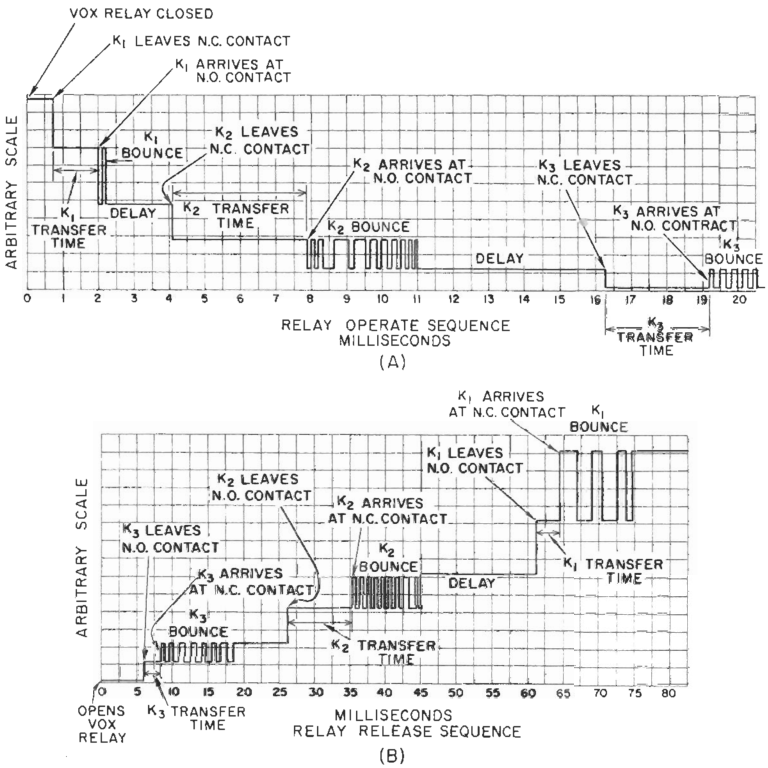


FIG. 2—Graph showing relay sequencing, including relay "bounce" characteristics. A shows the characteristics on "operate" (relays energized), while B follows the sequence on release (relays deenergized).

These relays will work well from the same d.c. supply voltage as the ARC-5 relay, although experimental readjustment of delay values may be necessary.

The relays used for receiver and transmitter control are miniature sensitive relays manufactured by Sigma and cost less than $2.00 each new. They operate quietly and rapidly.

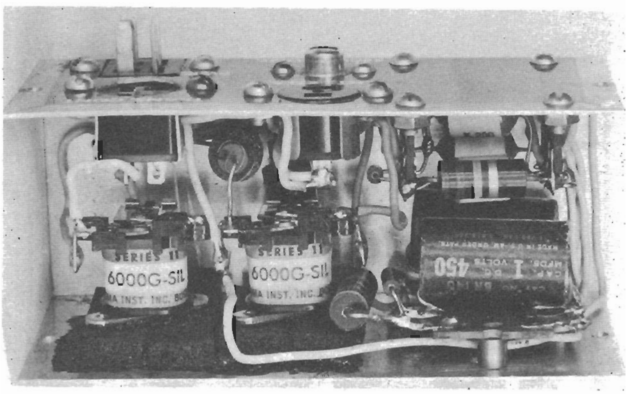
The circuit of the adapter is shown in Fig. 1. The system is controlled by a pair of normally-open contacts on the VOX relay. K_1 is for receiver muting, K_2 for antenna change-over and K_3 for transmitter power control. The contacts of K_1 and K_3 may be used in any desired manner to suit the individual control arrangement, since the adapter system provides proper sequencing, including allowance for bounce, in both directions of the armature travel. However, the relays of the adapter should not be used to control other relays; they should perform the intended function directly to preserve the desired sequencing. This sequencing is illustrated graphically in Fig. 2. Fig. 2A shows the progression in changing over from the receiving condition to the transmitting condition, while Fig. 2B shows the reverse progression in returning to the receiving condition.

No specific circuitry is shown for receiver and transmitter controls, since this will depend upon the equipment in use and the operator's preference. As suggestions, the receiver-control contacts may be used to raise the cathodes of the r.f. and i.f. stages far enough above ground to disable the receiver, or they can supply the required cutoff negative voltage to the a.g.c. line. The transmitter-control contacts may be used to open cathode circuits on stand-by. In cases where additional transmitter-control contacts are desired to provide a cutoff bias to the final amplifier for tube protection or shot-noise quieting, an additional relay, K_4 may be used without delay and connected as shown in dotted lines.

Delay Circuits

As previously stated, in going from receive to transmit, K_1 should close first and as quickly as possible to disable the receiver. When the normally-open contacts of the VOX relay close, approximately 140 volts is applied to K_1 through R_1 which serves to limit the power dissipated by the coil to a safe level. Capacitor C_1 produces an insignificant delay in the closing of K_1, since the series resistance common to C_1 and K_1 is negligible. The resistor R_1 tends to speed up the action of K_1.

K_2 must wait for K_1 to close and for contact



This unit includes K_1 (center) and K_3 (left). The components are assembled in a 2¼ × 2½ × 5-inch aluminum Minibox. Note the sponge-rubber relay mounting.

bounce to subside. This necessary delay is an inherent characteristic of the relays recommended for K_2. Therefore no components are needed to delay the operate time of K_2. R_3 limits the power dissipated by the coil of K_2. R_3 and C_3 in series with the coil of K_2 actually act to make the relay operate faster than it would if 24 volts was applied directly to its coil. Nevertheless, the resultant delay is adequate if the suggested relays are used. Again, C_2 has an insignificant effect on the *operate* time of K_2 both because of the negligible common series resistance.

K_3 is inherently fast-acting, so its action must be delayed to permit K_1 and K_2 to operate first. This delay is supplied by C_4 and the common series resistance of R_7. These latter are the only components affecting the operate time of K_2 since, on the receive-to-transmit cycle, the voltage divider R_5-R_6 biases CR_5 to nonconduction to prevent R_6 from affecting the time delay, although CR_6 will conduct. We now have all relays energized with the receiver off and the transmitter on.

From Transmit to Receive

To go from transmit to receive, we first must turn off the transmitter. To do this as quickly as possible after the VOX relay opens, we should open the K_3 coil circuit. This is done quite effectively by CR_6 which electronically disconnects C_4 from K_3. (CR_3 is so polarized that it will not conduct in the direction of the discharge current from C_4.) CR_5 is now no longer back-biased, so that C_4 is rapidly discharged through R_6. (Discharge of C_4 is necessary, of course, so that C_4 will be ready to provide an accurate delay on the next receive-to-transmit cycle.)

The release of K_2 is delayed by the charge on C_2, and K_1 is delayed on release by the charge on C_1. CR_2, CR_3, and CR_1 are isolators to prevent interaction between the timing circuits.

Receiver Noise Suppression

The r.f. chokes at the VOX relay contacts are used to suppress electrical noise generated by the small arc as the contacts open. Since the receiver is still sensitive for a few milliseconds after the VOX relay operates, any r.f. disturbance of this sort can be picked up by the receiver unless the shielding is very complete in the antenna circuit. The chokes may be omitted and added later if found to be required. In most instances, however, they will be found to be beneficial. They should be installed at the VOX relay contacts with the *shortest possible* leads between the chokes and the relay contacts.

The installation of chokes in the receiver-control contacts of K_1 may also prove desirable. Even the low current switched by these contacts can cause an r.f. disturbance which can be picked up by the receiver in some instances. While not shown in the schematic of Fig. 1, these chokes can be seen in the photograph of the adapter unit.

Capacitors

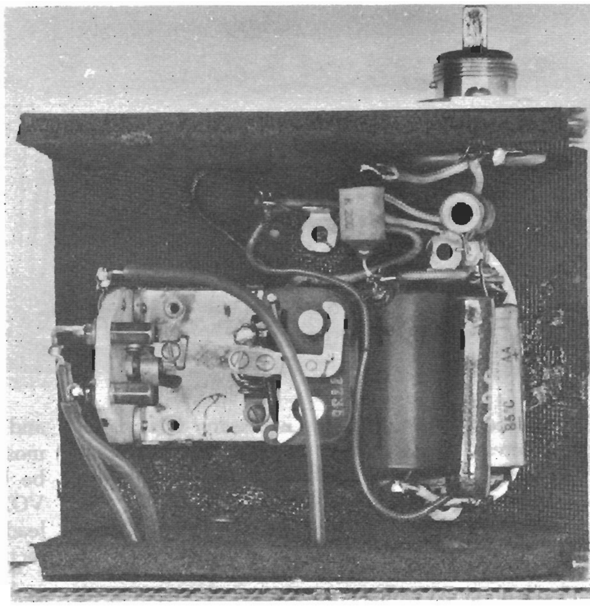
Electrolytic capacitors are used because of the size problem which paper capacitors would present. Electrolytic capacitors have proven to be entirely satisfactory in this low-impedance application. Because of the manufacturing tolerance associated with the relays and capacitors, it may be necessary to determine experimentally the value of capacitance to provide the correct timing. The final proof of performance of this gadget is its ability to operate without antenna arcing or "pops" in the receiver. The tolerance on electrolytics is none too good, of course, and in the case of the 1.5-μf. capacitors, C_1 and C_4, it will be necessary to make a selection from a group of 1- and 2-μf. units. For this reason, it may be more practical to use tubular paper capacitors in these two positions, if space can be found for them. The desired capacitance can be made up of a 1-μf. unit in parallel with a 0.5-μf. capacitor.

Relay Characteristics

The operate times shown in Fig. 2 are representative of what can be done with the relays used. The over-all design is pretty well dictated by the characteristics of K_2 in this instance, since it is inherently slower-acting than K_1 or K_3.

The total timing period of the release sequence of Fig. 2B is not particularly critical. As shown, it is faster than the normal release

[2] See Stein, "Some Hints on Relay Operation," *QST*, June, 1956.



The antenna relay, K_2, and associated delay components are mounted in a 4 × 5 × 3-inch Minibox fastened to the rear of the transmitter. The box is lined with sound-deadening material.

adjustment of most VOX systems. As long as the total release timing period does not cause undesirably long "hang-on" of the VOX, the release sequence of the relays can fit into this period in any manner which keeps them from coinciding in operate time. The circuit shown will follow c.w. up to about 15-20 w.p.m. and can be adapted for phone-c.w. operation by means of some additional switching.

It should be emphasized that the operating characteristics of no two relays, even of the same make and model, are identical, and that some adjustment of the delay-circuit values given should be expected.

Reducing ARC-5 Relay Bounce

The very desirable mechanical construction of the ARC-5 antenna relay, which gives it the large contact spacing so desirable for high r.f.-voltage use, brings about a problem in contact bounce. The long spring-like contact of this relay is very prone to bouncing on release as the movable contact meets the normally-closed contact.

A group of four of these relays was checked for contact bounce. The bounce of the movable contact when it reached the normally-open contact was not abnormal, but all four relays showed 35-40 milliseconds bounce time on release when the movable contact arrived at the normally-closed contact. One other ARC-5 relay had a piece of felt inserted between the turns of the springlike movable contact. Tests on this relay showed less than 10-15 milliseconds bounce time on release. Installing a similar piece of felt in each of the four relays showed that bounce time could be cut to less than 20 milliseconds on all relays.

Fig. 3 is a sketch of how the felt was inserted in the contact mounting. The felt is

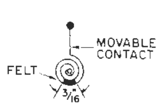


FIG. 3—Sketch showing method of damping bounce in ARC-5 antenna relays.

material which can be taken from underneath the cover flaps of the ARC-5 transmitter series and is about 1/16 inch thick. A piece about 3/16 by 1/4 inch is used.

The graphs of Fig. 2 were taken with the felt installed. If no felt were used, C_1 of Fig. 1 would have to be larger to delay K_1 enough on release to allow for the larger bounce time of K_2.

Acoustical Noise Reduction

The ever-present noise of relay operation has been a subject of discussion for years. Nothing new is advanced here toward solving this problem. Its existence is, however, acknowledged. The small Sigma receiver- and transmitter-control relays are mounted on sponge rubber to reduce their operational noise. This rubber serves another important purpose. The moveable contacts of these relays are connected to the relay frame, so that it is necessary to insulate the relays electrically. The sponge rubber does both the job of mechanical isolation and electrical insulation. Rubber cement of the type used to cement automobile weatherstripping does the job of cementing the sponge rubber to the chassis and the relay to the sponge rubber. The sponge rubber can be seen in the photograph of the adapter unit.

The antenna relay is much the noisiest of the three. It is mounted at the final amplifier in a Minibox which is lined with sponge rubber. The relay is cemented to the rubber without using any mounting screws which would conduct sound to the box.

These noise-reduction techniques are about 50 per cent effective. Enclosing each of the two chassis inside another rubber- or acoustic-tile-lined box is an idea for further experimentation.

Keying Filter

The graphs of Fig. 2 show that K_3, the transmitter-control relay, bounces quite a bit when the movable contact makes with either the normally-open or normally-closed contact. If this relay is used to key a transmitter for c.w. operation, clicks can result if a keying filters is not used.

» *Although there are arguments about the efficacy of speech clipping in s.s.b. transmission, there are none at all about the benefits to be realized through speech compression. By keeping the average modulation level near maximum, it makes you "talk up"—automatically.*

High-Quality Speech Compressor

NICHOLAS G. RICHARDS, W3ZVN, AND WALTER PAINTER

The speech compressor, like the speech clipper, provides a high average level of audio output, while at the same time keeping the peak percentage of modulation within the legal limit. The advantage of a speech compressor over a speech clipper is the extremely low distortion. This results from the use of automatic volume control in contrast to peak signal clipping in clipper circuits.

The use of speech compression with s.s.b. is particularly rewarding. By increasing the average audio input to the transmitter, the average r.f. output is proportionally increased within the limits of the transmitter.

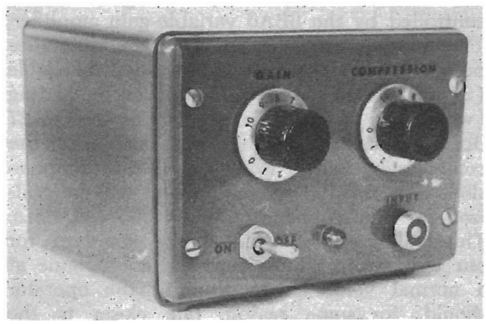
The speech compressor described in this article provides for adjustable compression to 27 db. for 4-db. output variation (see Fig. 1). Any high-impedance microphone with an r.m.s. output of 0.05 volt will drive this unit to full compression.

Circuit

The circuit of the compressor is shown in Fig. 2. The microphone input stage, V_1, is a resistance-coupled pentode amplifier. The 5879 was chosen because of its relatively high gain and low noise figure. To realize the full compression capabilities of this unit, an r.m.s. input voltage of 0.05 volt is required on the grid of the 5879. A microphone preamplifier is recommended when the microphone output is well below this figure.

The output of the microphone amplifier is split and fed into V_2, a 6AB4 a.g.c. amplifier,

From February, 1963, *QST*.



The compressor amplifier, complete with power supply, is housed in a cabinet approximately 4 by 5 by 6 inches in size. The power switch and microphone connector are under the gain and compression controls, respectively.

and V_3, a 6BA6 variable-compressor amplifier. Tube V_2 amplifies the signal, and its output is rectified as a negative d.c. voltage by diode CR_1. The filter network consisting of R_4, R_5, R_6, and C_5, C_6, C_7, smooths out this pulsating d.c., and it is then applied through R_7 to the grid of V_3. The d.c. level on the grid of V_3 is directly proportional to the audio level from V_1. This, in turn, is proportional to the audio level from the microphone.

The compression amplifier, V_3, receives audio from V_1 (by coupling capacitor C_4) in addition to the a.g.c. bias previously mentioned. The gain of V_3 depends on the d.c. level (a.g.c. bias) on its control grid, and varies inversely as this level.

Adjustment of R_1 controls the signal level into both V_2 and V_3. However, since the gain of V_3 varies inversely as the a.g.c. bias voltage, if the input signal to V_2 and V_3 is reduced by one half (by adjustment of R_1 or by decreasing the input to V_1), the gain of V_3 increases by a factor of 2 and the output remains the same.

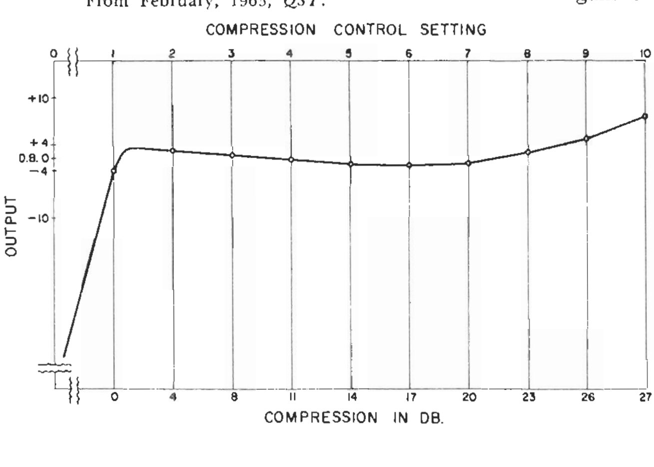
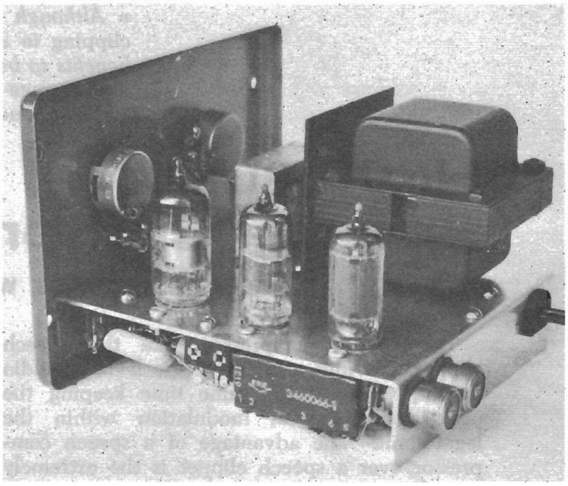


FIG. 1—Compressor characteristics. Curve shows compression in db. for a 4-db. output variation, with a 0.05-volt input signal at 1000 c.p.s.



Components are assembled on a small chassis. The power transformer is mounted on a bracket to make its mounting screws easily accessible. Tubes from left to right are V_1, V_2 and V_3. The extra output connector provides connection to a scope monitor.

While the output does not remain absolutely constant, it varies by only 4 db.

The audio distortion will not run above 3 per cent. The low value of distortion is possible in this circuit because there is no clipping action on the waveform as the signal level is controlled by the 6BA6 variable-gain amplifier. The limiting action is equal to the difference between the maximum compression available (27 db.) and the actual compression setting (which can be varied in ths unit from 0 to 27 db. by adjustment of R_1). For example, if the compression control is set for 14-db. compression, only 14 db. of compression will be realized for a microphone input of 0.05 volt r.m.s. This voltage is considered the standard for our tests when talking into the microphone at a normal voice level. Now, if the operator should, for some reason, talk closer to the microphone or raise his voice level, there is an additional 13 db. of compression available before the full limit of the unit is reached. Above the full limit (27 db. of compression) the output no longer remains constant, but increases with an increase in microphone output and eventually the 6BA6 becomes overdriven.

While this compression amplifier is basically a simple circuit, it has the advantage of a

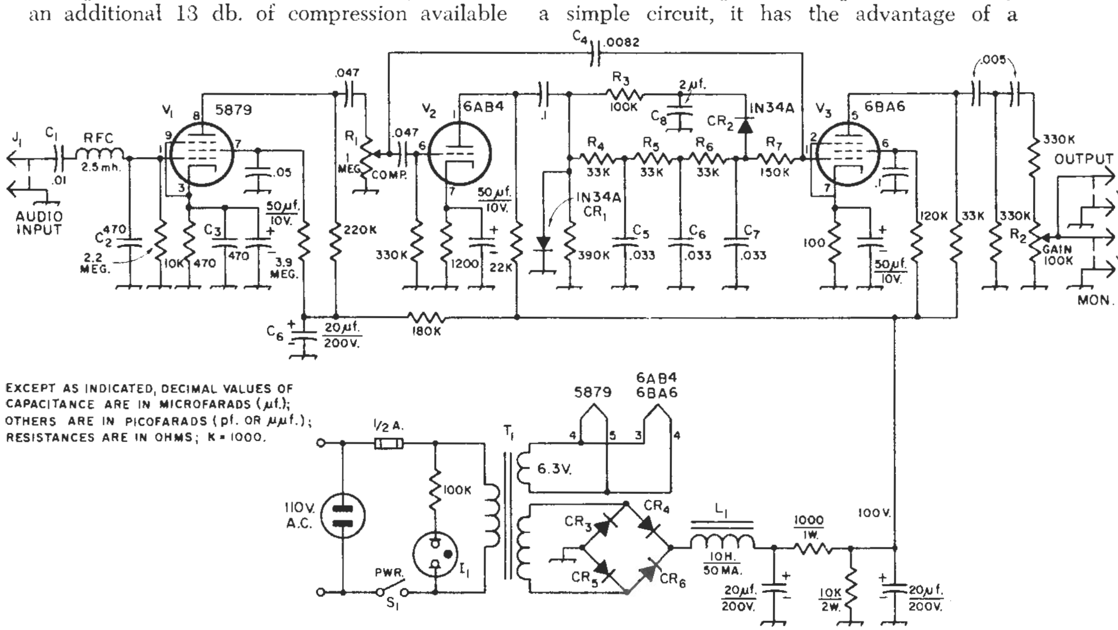


FIG. 2—Circuit of the W3ZVN speech compressor and power supply. Capacitors marked with polarity are electrolytic; others not listed below are 200-volt tubular paper, or mylar. Fixed resistors are ½ watt unless indicated otherwise. Components labeled, but not listed below, are identified for text-reference purposes.

C_1, C_2, C_3—Disk ceramic.

C_4—Tubular, 10-per-cent tolerance.

CR_3, CR_4, CR_5, CR_6—Silicon diode, 50 ma. 300 p.i.v. or higher.

I_1—NE-2 neon lamp.

J_1, J_2, J_3—Microphone connector (Amphenol 75-PC1M).

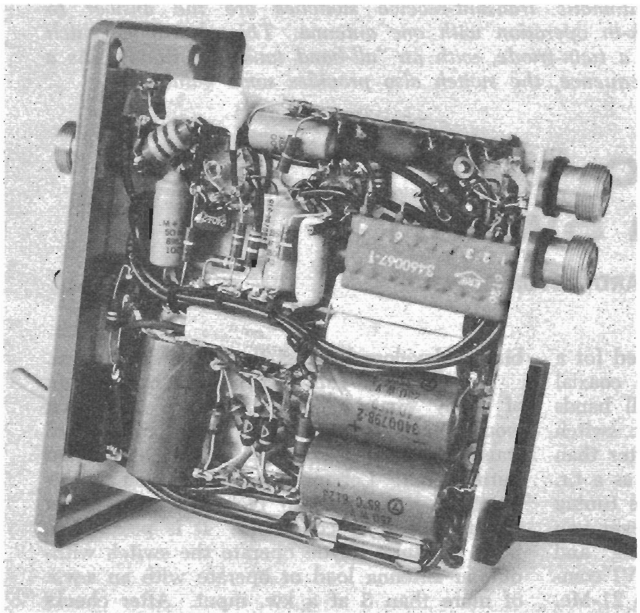
L_1—Filter choke (Triad C-3X).

R_1—Linear control.

R_2—Audio-taper control.

S_1—S.p.s.t. toggle switch.

T_1—Power transformer: 125 volts, 15 ma.; 6.3 volts, 0.6 amp.



Although close spacing of components is required, "layer" construction has been avoided so that all are accessible. As mentioned in the text, some of the components used in this model are of larger physical size than necessary.

relatively fast attack time of approximately two milliseconds. This eliminates the very objectional transient change in gain and high transient distortion that a slower attack time gives. The decay time, on the other hand, is relatively slow, approximately 1.5 seconds. This means that the gain of V_3 does not have to vary during short pauses. This also prevents any background noise from appearing during these pauses. The slow decay time is obtained by diode CR_2, resistor R_3, and capacitor C_8. If this network is eliminated, the attack and decay times will be approximately the same (2 milliseconds). No difficulty has been experienced with "thump" as a result of the fast attack time. This is probably because of the a.g.c. filtering, its associated roll-off, and the restricted frequency response of the compressor.

The frequency response of the circuit is 300-3000 c.p.s. The response could be extended, but this unit was intended for the best possible voice communication and any increase in response would detract from this purpose.

Power Supply

The power supply consists of a full-wave bridge circuit followed by two stages of filtering, and furnishes 100 volts at 15 ma. Any well-filtered power supply delivering 100 to 150 volts at 15 ma. can be used. Good regulation is absolutely necessary for proper circuit operation.

Construction

The unit is constructed in a $4\frac{7}{32} \times 6\frac{7}{32} \times 5\frac{1}{4}$-inch cabinet (Bud CU465). The chassis was made from a piece of aluminum sheet, and bent after punching. The front lip of the chassis is secured to the front panel by the toggle switch, neon lamp and microphone connector. The compression and gain controls are mounted on the front panel for easy access. An additional output connector was installed on the rear chassis flange for connecting a scope or 1-kc. oscillator for a two-tone test.

The general layout of the circuit is not critical, but lead lengths in the audio section should be kept to a minimum, especially those to the grids of V_2 and V_3. Shielded cable should be used for connections to and from controls R_1 and R_2.

Two module packs containing several of the necessary components were used to conserve space. Ample space will be available for individual components if the sizes of the components are kept to a minimum. Many capacitors larger than needed were used in this layout simply because they were on hand.

Adjustment

Adjustment of the unit is relatively simple, since it involves only two controls—for compression and output levels. Amateurs with access to laboratory test equipment can plot a compression curve for their unit similar to the one shown in Fig. 1. This is accomplished by feeding a 1-kc. signal at 0.05 volt into the input connections and measuring the output with a high-impedance audio voltmeter for all positions of the compression control. The output voltage scale is arbitrary, since it is dependent on the gain-control setting.

If test equipment is not available, the circuit can be checked with a tape recorder, or on the air. Turn the compression to maximum and adjust the gain control for full modulation while talking into the microphone. The modulation level should remain high as you back away from the microphone while talking at the same level. With the compression level turned down, the level of modulation will drop rapidly as you move away from the microphone, just as though the compressor were not used. During an on-the-air test with 14 db. of compression, good modulation was reported when talking 10 feet from the microphone. With the compression at zero, no modulation was reported under the same conditions.

Since the compression is adjustable, it should be possible to obtain the best operating conditions under any noise conditions. For example, the minimum compression would be normally used where the background noise is high.

» *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]

From *QST*, October, 1957.

[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.*

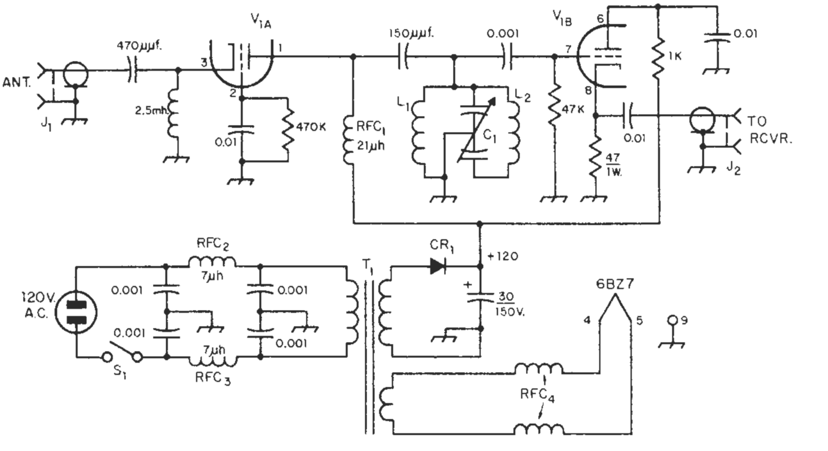
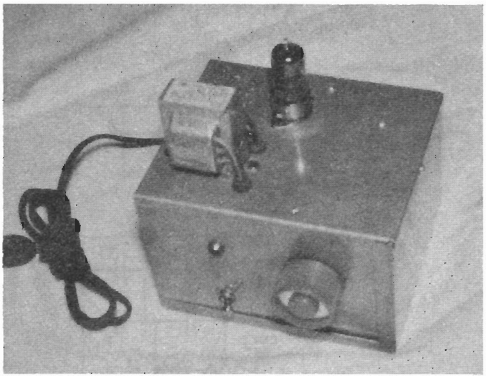


FIG. 1—Schematic diagram of the electronic t.r. switch. Capacitances are in μf. unless otherwise noted. Resistors are ½ watt unless otherwise specified.

C_1—450-pf. per-section, broadcast-receiver type.
CR_1—130-volt 65-ma. selenium rectifier (Federal 1002A or equiv.).
J_1, J_2—Cable connectors, SO-239.
L_1—19 turns, 1-inch diam., 32 t.p.i. (B & W 3016).
L_2—23 turns, ½-inch diam., 16 t.p.i. (B & W 3003).
RFC_1—Ohmite Z-28 or equiv.
RFC_2, RFC_3—Ohmite Z-50 or equiv.
RFC_4—Bifilar winding. See text.
V_1—6BZ7. See text.
T_1—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 ½-inch diameter form 1½ 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 4×5×3-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.

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-μf. capacitor is simply an r.f. bypass). A value of about 0.1 μf. 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

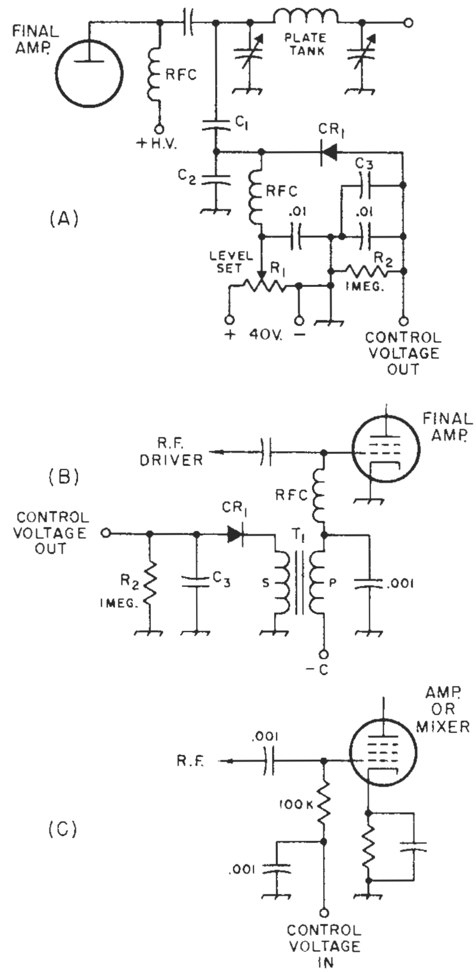


FIG. 1—Automatic Level control circuits.

(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 AB_1 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 AB_1. It makes use of the fact that a small amount of rectification occurs in the grid-cathode circuit of a tetrode AB_1 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.

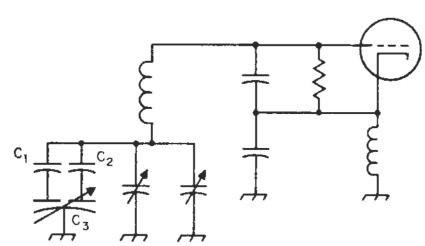


FIG. 1—The v.f.o. in the HT-32 can be set to the best condition of temperature compensation through the use of a differential capacitor of N1500 and NPO coefficients. Changing the rotor position of C_3 permits effective adjustment of the coefficient from an NPO characteristic to N1500.

Notes

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 AB_1 or AB_2, so that the plate

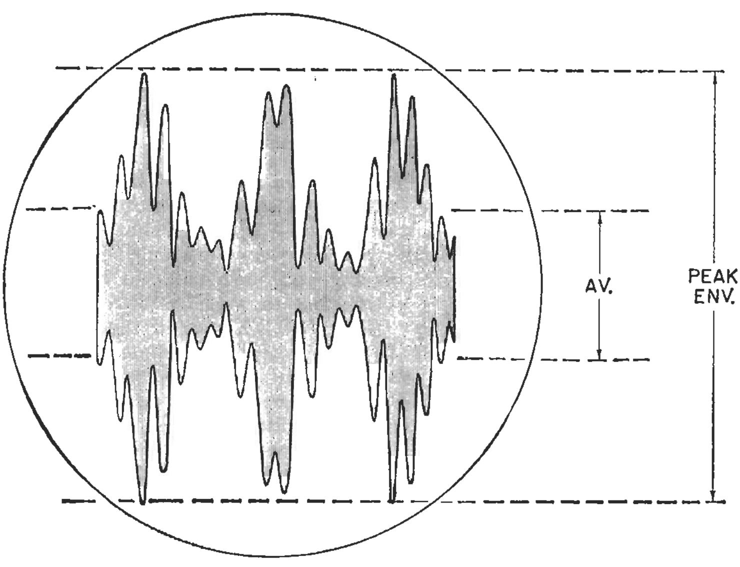


FIG. 1—A typical s.s.b. voice-modulated signal might have an envelope of the general nature shown, where the r.f. amplitude (current or voltage) is plotted as a function of time, which increases to the right horizontally.

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 transmission 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.

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.

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 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 ing 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. Grid Current[2]	Max.-Sig. Driving Power[3]	Max.-Rated 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		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		750	300	—35	15	70	.3	8	35	—	0	3.5	—	30	35
829B	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
4-65A	AB_1	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	—	0	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		2000	600	—115	25	175	0	7	112	0	0	10	—	125	210
7094	AB_1	1500	400	—65	30	200	—	35	65	0	0	20	—	—	185
		2000	400	—65	30	200	—	35	60	0	4	20	—	—	250
813	AB_1	2500	750[5]	—95	25	145	—	27	90	0	0	—	—	—	245
	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]	—	6.0 (1.5)[4]	95	0	0	20	—	—	200
	AB_2	1500	350	—41	44	200	0	17	141	9	1.25	20	5	125	175
		2000	350	—45	36	150	0	3	105	7	.7	20	5	125	175
		2500	350	—43	47	130	0	3	89	6	.5	20	5	122	200
7034/ 4X150A	AB_1	1000	300	—50	50	225	0	11	50	0	0	12	—	—	115
		1500	300	—50	50	225	0	11	50	0	0	12	—	—	200
		1800	300	—50	50	225	0	11	50	0	0	12	—	—	250
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
7203/ 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
4-250A/ 5D22	AB_1	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
	AB_2	1500	300	—48	50	243	0	17	96	11	1.1	35	10	150	214
		2000	300	—48	60	255	0	13	99	12	1.2	35	10	185	325
		2500	300	—51	60	250	0	12	100	11	1.1	35	10	205	420
		3000	300	—53	63	237	0	17	99	10	1	35	10	190	520

7580	AB1	2000	400	— 77	70	350	—	35	77	0	0	12	—	—	400
7580W/ 4CX250R	AB1	1500	350	— 62	133	350	—	—	62	0	0	12	—	—	262
		2000	**400**	— 80	70	350	—	—	80	0	0	12	—	—	470
PL-175A	AB1	2500	750	—143	100	350	1	35	143	0	0	25	—	265	570
		3000	750	—150	80	350	1	29	150	0	0	25	—	305	680
		3500	750	—160	75	350	1	24	160	0	0	25	—	345	790
PL-8295/172 PL-8432	AB1	2000	500[6]	—110	200	800	12	43	110	0	0	30	—	—	1040
		2500	500[6]	—115	200	800	11	40	115	0	0	30	—	—	1260
		3000	500[6]	—115	220	800	11	39	115	0	0	30	—	—	1590
4CX1000A	AB1	2000	325	— 60	250	1000	—2	35	60	—	0	12	0	—	1020
		3000	325	— 60	250	900	—2	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—GROUNDED-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
3-400Z	2000	0	62	400(265)[3]	—	148 (87)[3]	—	—	—	445[4]
	2500	0	73	400(274)[3]	—	142 (82)[3]	—	—	—	560[5]
	3000	0	100	333	—	120	—	—	32	655
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 at 1 k.w. p.e.p. input.

[6] Approximate; adjust to give stated zero-signal plate current.

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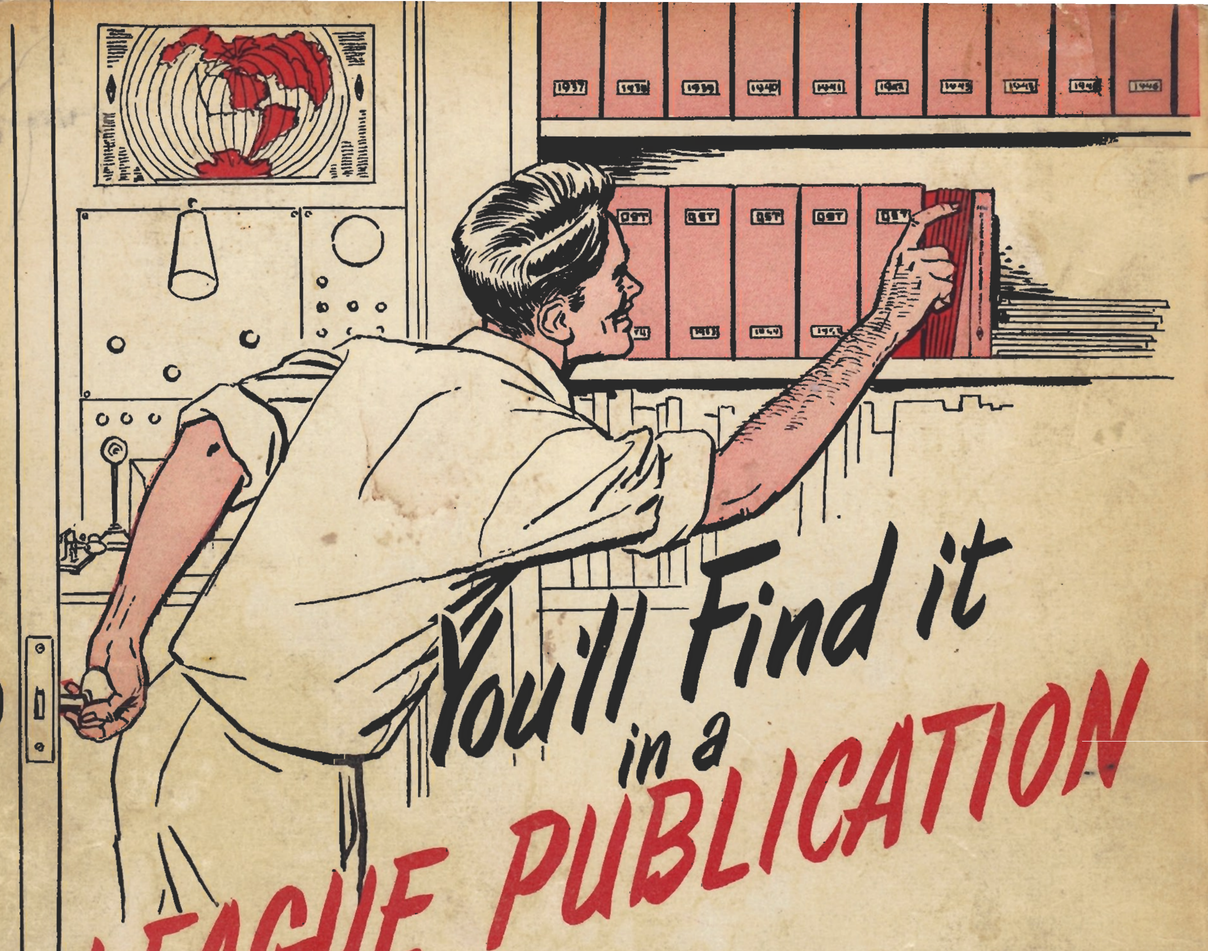

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