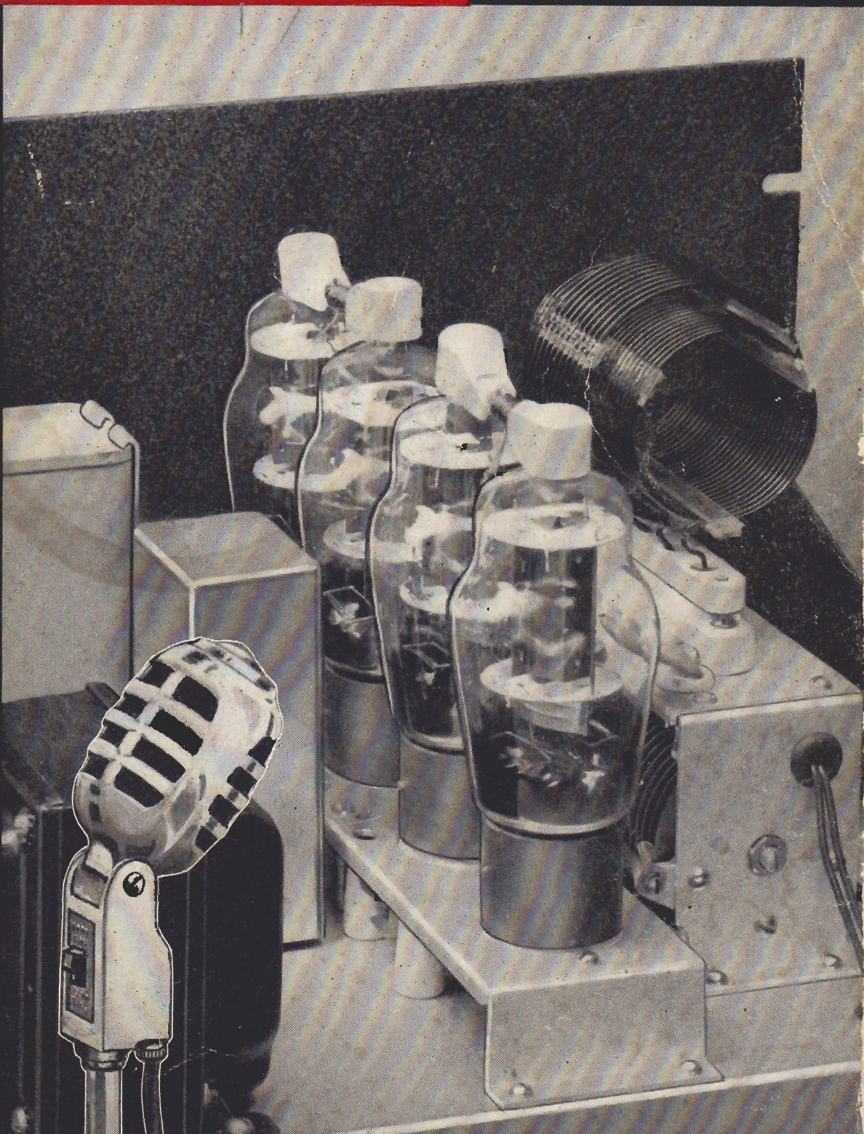
SINGLE SIDEBAND

for the Radio Amateur

$1.50

A DIGEST OF AUTHORITATIVE ARTICLES ON AMATEUR RADIO SINGLE SIDEBAND



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SINGLE SIDEBAND

for the

RADIO AMATEUR

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Foreword

It is some six years since single-sideband telephony was introduced through the pages of *QST* in a form practical for use by the radio amateur. Admittedly a more difficult technique than other forms of 'phone, its inherent advantages have continued to sell themselves to an ever-increasing number of amateurs. As interest in the subject has grown, so has the demand for a publication — a "single-sideband handbook" — that would put between a pair of covers all the currently useful information.

One recurring suggestion of the past few years has been that the *QST* articles on single sideband be printed and bound together in one volume. In debating the form which a book on s.s.b. should take, the editors became more and more intrigued by the virtues of this idea, and impressed by the scope and quantity of the material on single sideband that has appeared in *QST*. It was evident that practically everything needed to make a well-rounded reference book was there, most of it too good to be forgotten (as is human nature) by those who had seen it when it was first published, and much of it not accessible to newcomers.

This book is the answer to the question you, the reader, might ask if you were just getting interested in s.s.b.: "What do you recommend that I read in past issues of *QST* to get acquainted with the subject, and what should I think about building?" The book is not a simple reprint of a number of *QST* articles. It is first of all a selection, aimed at covering the subject from all sides and eliminating those methods and ideas that, in common with experience in any developing art, have for one reason or another not survived to the present. Second, the selected articles have been coördinated with each other to eliminate unnecessary duplication — in the early days, for instance, every author justly believed it necessary to outline the advantages of s.s.b. over a.m.; in a book such as this, these arguments would be merely redundant and one good general article suffices. In this respect, the book is a sort of "digest."

It is important to observe, however, that the "digest" idea has not been carried to the point of eliminating desirable material by condensation. If an article as printed here is shorter than it was originally in *QST*, the parts deleted either are subjects covered in another article, or do not represent current practice. *Everything that you could use today if you read the original articles has been retained.*

The book could not exist, of course, had it not been for the enthusiasm with which the pioneers and experimenters took up s.s.b., and their willingness to make known their findings through the pages of *QST*. The editors feel it is a privilege to preserve their works in a book such as this, and thus forestall the obscurity that so often overtakes a useful contribution as the older issues of a periodical recede into the background.

A. L. Budlong

West Hartford, Conn. *General Manager, A.R.R.L.*

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A HISTORY *of* AMATEUR SINGLE SIDEBAND

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

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

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

The basic s.s.b. techniques are almost as old as radiotelephony itself, and communications companies have used s.s.b. 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 s.s.b. 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 s.s.b., 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 s.s.b. 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 by the 50s that it is no longer a consideration. All radio is complex these days, and we are conditioned to expect it.

» *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 and continue through the following article, then skip to "How To Tune in a Single-Sideband Signal" and try it on your own receiver. You'll then be ready for "What Single Sideband Has To Offer" — and you're sold.*

How To Visualize a 'Phone Signal

BYRON GOODMAN, W1DX

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,

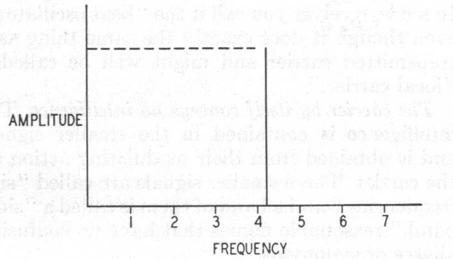


Fig. 1 — A representation of a single radio frequency, shown by plotting amplitude against frequency. A steady signal by itself takes up no room in the spectrum.

they differ by 10 cycles, and a receiver or other device that could separate signals 10 cycles apart could separate these two.

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

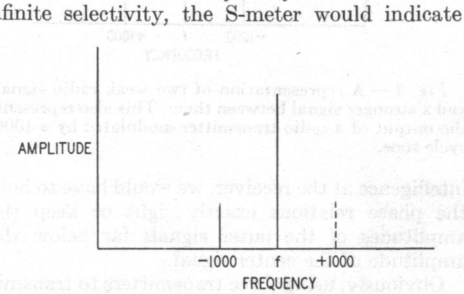


Fig. 2 — Two radio signals, separated by 1000 cycles, will give a 1000-cycle audio signal when they are mixed in a detector or other nonlinear circuit.

the amplitude at *one* setting of the tuning knob as you tuned across the frequency range shown, and *nothing* at any other setting.

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

From *QST*, July, 1950.

"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 minimized when the center signal is made large in proportion to the other signals. Thus if we didn't wish to introduce some extraneous or false

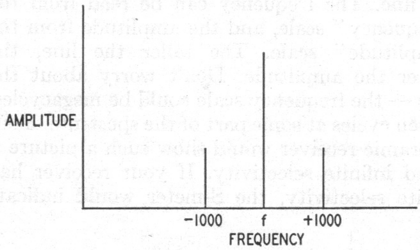


Fig. 3 **— A representation of two weak radio signals and a stronger signal between them. This also represents the output of a radio transmitter modulated by a 1000-cycle tone.**

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 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 very 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 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).

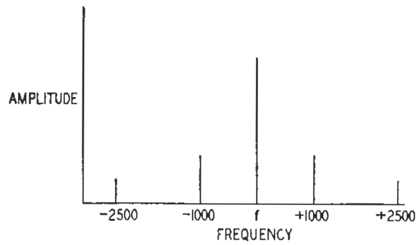


Fig. 4 — **A representation of five radio signals or a single transmitter modulated by 1000- and 2500-cycle tones. The 1000-cycle tone has almost twice the amplitude of the 2500-cycle tone.**

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 unworkable with present, and possibly future, 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.

If insufficient carrier is supplied at the transmitter, extra signals will be generated and radiated, as in the familiar case of overmodulation splatter. If insufficient carrier is supplied at the receiver, extra signals will be generated and heard, as in the case of trying to receive a single-sideband signal with insufficient b.f.o. injection.

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 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. For mental exercise, visualize what happens when you pull out the carrier during transmission and reinsert it at the receiver, or lop off one sideband at the transmitter or the receiver — it will all add up easily. Then try it again, thinking only in terms of "modulation envelopes" and "percentage of modulation"!!

» Another picture of the 'phone signal from the sideband standpoint, with a brief description of the two current methods of generating s.s.b. signals.

Getting Acquainted with Single Sideband

HOWARD WRIGHT, W1PNB

SEVERAL years ago, having built and operated several successful amateur radiotelephone transmitters, I was reasonably satisfied with my knowledge of 'phone principles. After all, they didn't seem too complicated, if one could manage to ignore the inconsistencies that showed up now and then. To the best of my memory, I used to consider modulation from about the following point of view:

"The r.f. section of a transmitter consists of a carrier-generating exciter and a final amplifier that amplifies the carrier and passes it along to the antenna. To use this typical c.w. transmitter for 'phone operation, we merely couple voice power to the final amplifier through a modulation transformer. The voice power is then in series with the power supply to the final. Therefore, the level of the carrier is varied above and below its original value at an audio rate. This is called 'modulating' the carrier and is done to allow the voice signal to be recovered by a receiver."

As I said before, I was quite happy with the above understanding of a 'phone transmitter. I suspect that there are many amateurs who are getting along nicely today on similar ideas.

And then came single sideband! Formerly, I had considered sidebands as a condition somewhat resembling a case of measles, occurring only on unhappily adjusted transmitters. With the introduction of amateur s.s.b. techniques, we were made aware that sidebands are completely normal and honorable. Concerning modulation, as the saying goes, "I didn't know from nothing."

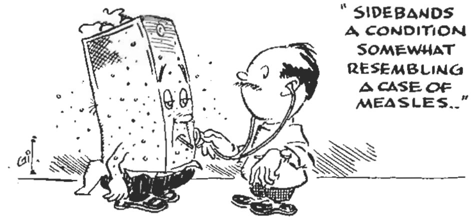
In my estimation, a main reason for the lack of a better general understanding of 'phone principles is that, in the hands of a person without much theoretical knowledge, even the best of receivers tends to create a false impression of the true nature of incoming signals.

The Receiver

Before you rise to the defense of your particular high-priced beauty, let me hurry to state that receivers only tell lies to people who don't realize that a receiver, designed for broadcast-band type reception, inherently disguises the true nature of incoming signals. Here is a typical example:

A neophyte tunes his receiver across an unmodulated carrier. The receiver tells him that the carrier is a certain number of kilocycles wide. The neophyte immediately starts a frantic and futile investigation to discover why one carrier is broader than another.

Now let's have a man who has studied receivers tune the same receiver across the same carrier.



He also sees that the carrier occupies space on the dial but, knowing that a carrier has no width, he realizes that the carrier is telling him the selectivity, or "bandwidth," of the receiver.

This case of the unmodulated carrier is bad enough, but the receiver is designed to perform a masterpiece of deception in the case of a modulated 'phone signal. It does a perfect job of gathering in the various parts of the signal, eliminating any evidence of the presence of the sidebands theory tells us were transmitted, and combining the sidebands with the carrier in such a way that it appears that the voice is simply superimposed upon the space supposedly occupied by the carrier. So complete is this deception that it might be compared to the reproduction of a color photograph in a magazine. How would we ever know that, to be reproduced, the picture was broken down into its primary colors, if all we had to go by was the original print and the magazine?

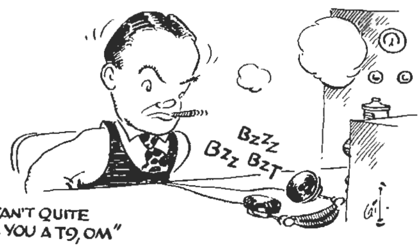
Spectrum Space

There is one very important concept to grasp. No intelligence (modulation) can be transmitted without taking up room in the spectrum. Couple this statement to the previously mentioned fact that a carrier occupies no space and there is only one conclusion to be drawn. The modulation can in no way be "on the carrier." It must consist of appropriate new signals at frequencies "alongside the carrier."

Take the case of an ordinary amateur 'phone transmitter. For the sake of discussion, let's say that the carrier frequency is adjusted to exactly 3900 kc. Now this happens to be a fine transmitter, except for one fact. There isn't enough filtering in the audio power supply. Of course, the result is a signal plagued with 120-cycle hum or, in

From "Low-Pressure Modulation Facts," *QST*, July, 1953.

the words of c.w., "I can't quite give you a T9, OM."



" I CAN'T QUITE GIVE YOU A T9, OM"

Let's take a theoretical look at our hum-modulated signal. The hum voltage of 120 cycles should mix with the 3900-kc. carrier and produce new signals of 3900.120 and 3899.880 kc. In other words, we should now have three separate signals, the strongest being the original, flanked on either side by a "hum side-frequency" 120 cycles away.

Until now, references to receivers may not have seemed too flattering. This, however, had only to do with the listener's lack of ability to interpret what he heard. Now, let's use our receiver to tie down theoretical reasoning to what we actually hear. Simply turn on the receiver's b.f.o. and tune carefully across the hum-modulated signal. Presto! We hear three distinct points of "zero beat." We have three signals. We have exact confirmation of the heterodyne theory of modulation.

If you're somewhat confused by my use of hum voltage in the above example, don't be. It was simply used in place of "a single audio tone," which is often used in explanations of sideband generation. Of course, hum is a far cry from the actual voice signals we use to modulate our transmitters. The voice contains a great number of individual frequencies which modulate the carrier. Each resulting new r.f. signal generated still maintains its original audio-frequency relationship with each of its neighbors, even though the whole business has been shifted up into the r.f. part of the spectrum.

Due to the heterodyne action, our complete band of audio frequencies is reproduced, not only once, but in exact duplicate on either side of the carrier. Thus we have the sidebands that have been discussed so much in recent years. Considering the original audio frequencies, we might think of the sidebands as being "back-to-back." The lowest-pitched sounds are close together alongside the carrier and the higher ones, progressively removed from each other, cause the complete signal to be twice as wide as the highest tone transmitted.

S.S.B. Techniques

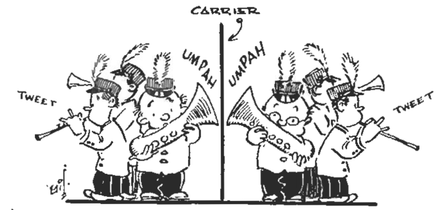
Now, let's take a look at single-sideband techniques. This definitely is not a drop-off point into the mysteries of complicated electronics. If you once manage to grasp a firm understanding of the regular double-sideband signals we have been discussing, single sideband is only a small step away.

Rather than start directly with s.s.b. transmitters, let's return to the thought of converting in a regular receiver. From one point of view, every superhet is an s.s.b. receiver in two respects. The first, one which seldom needs to be considered, is this: In converting incoming signals down to the intermediate frequency, the new frequency (or sideband) caused by the *difference* between the incoming signal and oscillator is the one that is used. The theory of heterodyning tells us that the *sum* of these frequencies is also present at the output of the converter. This sum frequency is so far removed from the i.f. that it is eliminated by the filtering action of following stages.

The more important reason for considering a receiver as having s.s.b. action concerns "image" reduction. Due to the heterodyne process, if no selectivity precedes the converter, the receiver is sensitive to two frequencies. One is above and the other below the oscillator by an amount equal to the i.f. I believe that most of us are familiar with the drawbacks of having bad r.f. "images" or, in other words, having each signal appear at two points on the tuning dial.

Here is the connection between receiver images and s.s.b. transmitters: The act of adding r.f. selectivity to the front end of a receiver to reduce the image is exactly the same process, in reverse, as adding a selective filter to a double-sideband transmitter to reduce the "image sideband." The only difference is that the receiver is purposely designed so that the image can be reduced by the use of a few simple tuned circuits preceding the converter. In a transmitter, the sidebands produced by modulation ("conversion" in a receiver) are separated only by a relatively few cycles and are therefore more difficult to divide by filtering methods.

You may have noticed that I have not stressed the "suppressed carrier" part of s.s.b. There is enough material contained in this subject to fill a book, but it is distinctly a separate subject from "single sideband." A very large part of both the superiority of s.s.b. systems and the furor caused by the appearance of s.s.b. signals on receivers tuned for regular operation can be attributed directly to carrier suppression and not to the



" THE LOWEST-PITCHED SOUNDS ARE CLOSE TOGETHER ALONGSIDE THE CARRIER..."

elimination of one sideband. This, however, is an article on modulation, so let's stick to the sidebands.

Now, if we had to give a definition of single sideband, we could call it the suppression of an "image sideband" for the purpose of reducing to a minimum the frequency band necessary to transmit a given amount of intelligence. Because the filter method is used for reducing the unwanted sideband or "image" in a receiver, we will first consider this method as applying to transmitters.

Filters

A carrier is modulated in the ordinary way, producing identical sidebands on either side of the carrier. These sidebands and the carrier are fed into a very sharp filter which passes one sideband and suppresses the other even though they are very close together. There are *LC* filters, crystal filters, and mechanical filters. They can all be built to do a good job of separating sidebands, but all have the common property of having better selectivity as their design frequency is made lower. This is the reason why practically all filter-type single-sideband transmitters use receiver-type heterodyne methods to convert to the desired band from the lower frequency at which the filter works well.

Phasing

The "phasing" method of s.s.b. generation employs theories which certainly seem to belong to people with engineering degrees. However, the theory of filtering is also basically very complicated, but we have been using different types of filters so long that we tend to leave their mystic properties to the experts. Let's describe the phasing system in terms similar to those used for filtering.

Each sideband is broken up into two parts by the use of a few craftily chosen resistors and condensers, a couple of tuned circuits, and a certain amount of adjustment. These parts of each sideband differ from each other only in that the times when any given thing happens are different ("phase shift" to an expert). The four signals thus produced are combined in another tuned circuit so that the two parts of one sideband "beat each other's brains out." The two parts of the other sideband take an immediate liking to each other and combine to form the signal intelligence to be transmitted.

The phasing method is not limited to low frequencies. It works as well at 50 Mc. as at 50 kc. However, for reasons of operating convenience, the signal is often generated at some point outside the band and heterodyned in.

In conclusion, let me say that the previously mentioned s.s.b. properties of receivers should, in no way, be confused with the general meaning of the term "selectable-sideband receiver." Such a receiver is actually able to remove the "audio image" from any incoming signal. In plain words, it listens to either sideband and rejects the other. Either phasing or filter methods are used in selectable-sideband reception. In fact, the very parts used in a transmitter can almost always be used in a receiver.

S.S.B. POWER LIMIT

Some of the single-sideband gang who use big tubes have wondered what the FCC interpretation on the s.s.b. power limit is. We quote from a Commission letter addressed to ARRL:

> The following . . . may be considered as a presently acceptable method for determining the d.c. plate power input to the final r.f. stage of a single-sideband amateur transmitter:
>
> The maximum d.c. plate power input to the radio frequency tube or tubes supplying power to the antenna system of a single-sideband suppressed-carrier transmitter, as indicated by the usual plate voltmeter and plate milliammeter, shall be considered as the "input power" insofar as Sections 12.131 and 12.136(d) of the Commission's rules are concerned, provided the plate meters utilized have a time constant not in excess of approximately 0.25 second, and the linearity of the transmitter has been adjusted to prevent the generation of excessive sidebands. The "input power" shall not exceed one kilowatt on peaks as indicated by the plate meter readings.

»A critical appraisal of single sideband in comparison with amplitude modulation. This is a basic article that every prospective "sideband" user should study.

What Single Sideband Has To Offer

DONALD E. NORGAARD, 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 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.

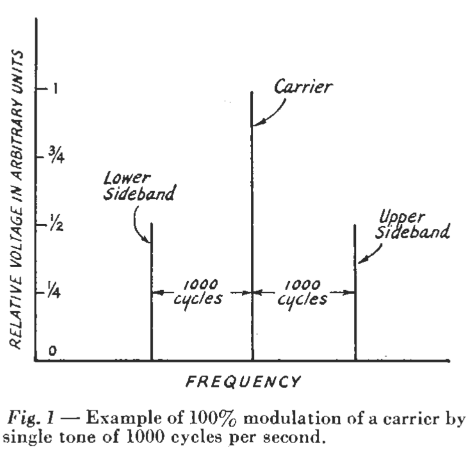


Fig. 1 — **Example of 100% modulation of a carrier by a single tone of 1000 cycles per second.**

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

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.)

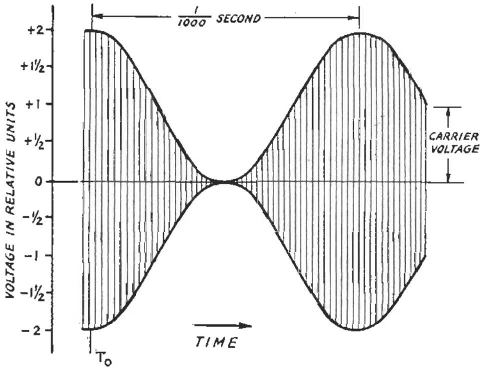


Fig. 2 — **Envelope of carrier 100% modulated by a 1000-cycle sine wave.**

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 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

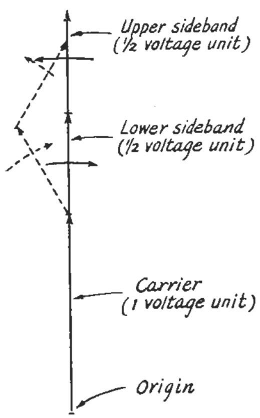


Fig. 3 — Vector diagram of 100% modulation of an a.m. carrier at the instant (corresponding to T_0 in Fig. 2) when peak conditions exist. The broken vectors show the relationships at an instant when the modulating signal is somewhat below its peak.

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 hatrack — something to hang sidebands on. It seems silly to carry a hatrack 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 hatrack home where it belongs. One hat? Certainly. It is ridiculous to go around trying to wear two dinky hats at the same time — especially in the rain!

Leaving the Carrier at Home

Sure, take a look at Figs. 1 and 3. Suppose we leave the carrier home and double the amplitude of each of our sidebands. This will still run our transmitter at its peak output capacity of 400 watts, all it can do. Well, the sideband power goes up all right. The sideband voltages are doubled, so our sideband power is four times what it used to be. That means each sideband is 100 watts, and our transmitter is not overloaded on peaks. The total sideband power is, of course, 200 watts. But this sideband power doesn't do much for us if it can't all be put to work. That is the situation with two sidebands and no carrier; nobody can take advantage of this sideband power, for it is in such a form that it doesn't lend itself to readability, no matter how you try to use it. Yet, the power is there and it can be read on a meter, but that's about all.

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 hatrack 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 hatrack, and your hat won't know the difference, either, because the hatracks 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 hatrack really means. First of all, his hatrack 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 hatrack 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 VFO 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. It can do it 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 learn how to talk with waveforms having a much lower ratio of peak to average power. Actually, the steady 100-watt carrier of an a.m. signal causes most of the dissipation in the 100-watt transmitter, but it was built to stand up under that kind of treatment.

While shrouded in theory, we were talking about *output* power, and managed to show that we could get 400 watts of sideband power output with single sideband at the same peak power that gave only 50 watts of sideband power in the case of a.m. That's fine for comparison purposes on a theoretical basis, but there is the practical matter of efficiency to consider. Let's lean over backward and say that a *good* Class C plate-modulated amplifier such as the one in our ideal 100-watt a.m. transmitter runs with an efficiency of 80%. Neglecting the fact that the total input under modulation with speech is somewhat higher than the carrier input (which is 100/0.80 = 125 watts), the dissipation in the output stage is 25 watts. Let us say, however, that the modulation still drives the transmitter to its peak output power of 400 watts, but has very low average power. Therefore, the peak sideband power output is 50 watts, with very low average power. Here is a strange way of rating things, but it means something: The peak *useful* sideband power is 50 watts obtained with a final-stage dissipation of slightly over 25 watts in the a.m. transmitter. The peak input power is, of course, 400/0.80 = 500 watts, since the efficiency of 80% is pretty nearly constant with this type of operation. You have already guessed what the next thing is. The peak useful efficiency is

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

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, then 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_g 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_g 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 × 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.

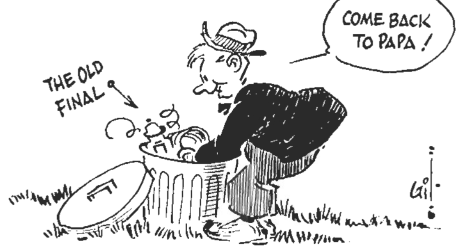
The foregoing comparison isn't absolutely accurate, since the actual waveform of speech input is unknown. But it is a fair comparison, and experience and tests support the argument. That is what really proves the point.

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 crystal 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 crystal 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 crystal notch may be used to eliminate one carrier heterodyne, but that is about all it can do. The advantage of single-sideband reception in this case is principally that, on the average, only half the number of heterodynes will be heard, where interference is the only disturbance to otherwise flawless reception. Well, that helps.

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

SIDEBAND SUPPRESSION AND RECEIVER TUNING

D. C. Bakeman, WØHKX, wrote an interesting thesis on s.s.b. for his M.S. at the University of Illinois. Since some of his experiments have a bearing on technical standards for amateur s.s.b., his findings are being passed along. For example, it was observed that attenuating all voice frequencies up to 500 cycles made little audible difference in intelligibility or volume, although some of the naturalness was destroyed.

This is useful information if you have been worrying about the characteristics of your sideband filter or phase-shift network not being too good at the low audio frequencies. The low-frequency burble you get on reception when the carrier reinsertion isn't exact was used as the criterion in some experiments on acceptable sideband rejection figures, and it was found that this burble was only barely noticeable when the sideband attenuation was 30 db. Thus 30 db. is probably an acceptable minimum figure to shoot for, so far as ease of reception is concerned. However, from the standpoint of minimizing QRM from your suppressed sideband, a figure of 40 db. is a more worthy objective.

Permissible carrier-resinsertion tuning error was also studied, and it was decided that the naturalness of the voice had definitely disappeared at 50 cycles high (audio components made higher) and 20 or 30 cycles low, although it was still easily understandable.

» 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, W1DX

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 kc." it corresponds to the notch being centered on 3903 kc. on Fig. 1, and all you could see (and hear) would be "Signal A" and a bit of "Signal B" that also shows 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 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 celluloid 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.

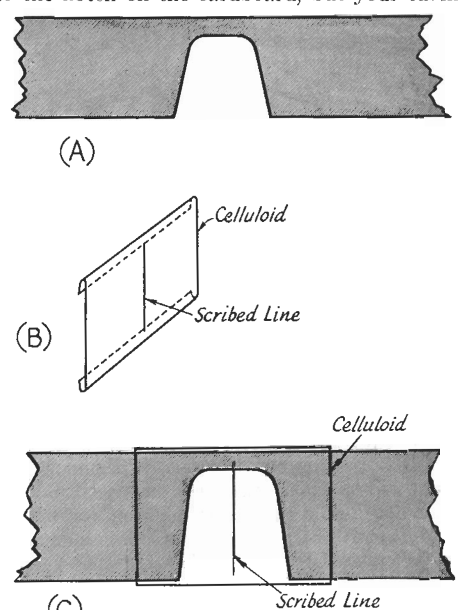


Fig. 2 — Parts required for model receiver to be used with Fig. 1. The strip at A could be a piece of cardboard, notched as shown to represent the "passband." A celluloid slider with a line scribed on it, to represent the b.f.o. frequency, would represent the b.f.o., as shown at B. With b.f.o. on, the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

Your receiver with b.f.o. off looks like Fig. 2A — 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

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

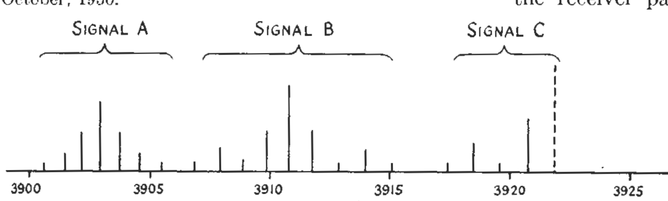


Fig. 1 — An instantaneous frequency-*vs.*-amplitude representation of a portion of an amateur band. Signals A and B are a.m. signals, and C is a single-sideband suppressed-carrier signal. Signal C is using the lower sideband, and the suppressed carrier is represented by the dashed line.

allow him to come through, and he would be centered at 3919.5 kc. You can do this with the a.v.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.v.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.v.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 in Signal C. (The dashed line represents the suppressed carrier.) Sit back and relax — you have just accomplished something that amateurs with 25 years of experience have found difficult or impossible!

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.

RATING S.S.B. TRANSMITTERS

The d.c. power input to almost every s.s.b. linear amplifier varies over a considerable range with varying modulation levels — just about the same sort of variation, in fact, that occurs in a Class B audio amplifier. Audio amplifiers, which likewise are linear amplifiers, are rated in terms of the maximum power output they can deliver with some specified value of distortion; or, if distortion is not specified, it is understood that it will be within limits commonly considered to be acceptable. This specification (or understanding) with respect to distortion is indispensable because the amount of power that can be obtained is critically related to the amount of distortion that is permissible.

An exactly similar method of rating is ideal for the s.s.b. linear amplifier. It tells you what you can expect to get in the way of "peak envelope power output," which is the thing that does the business in producing the signal at the receiving end. It is independent of the type of amplifier operation, so is equally meaningful for amplifiers falling anywhere between pure Class A and pure Class B.

However, it is not always easy to measure r.f. power output. An alternative method is to specify power in terms of "peak envelope power input." This is simply the d.c. power input at the modulation-envelope peak. It is responsible for the peak envelope power output mentioned above, and is related to it by the plate efficiency at the modulation-envelope peak. For this rating to mean much, it must be based on the assumption that average plate dissipation with typical voice modulation (not necessarily the dissipation at the modulation peak) does not exceed the rating of the tube type used in the final stage. Depending on the peak plate efficiency that happens to be obtained as a result of amplifier adjustment, the peak envelope power input may run 1.4 to 2 or more times the peak envelope power output.

In a Class AB or B linear amplifier of ordinary design operating with low distortion, the peak envelope power input is, with reasonable accuracy, 1.57 times the d.c. input as measured with a two-tone test signal. The factor 1.57 will change somewhat with no-signal plate current, but in amplifiers where the design objective is to utilize the available tube plate dissipation capacity to best advantage the difference will not be great enough to be of practical importance. This suggests that still a third method of rating might be used: d.c. power input with a two-tone test signal.

A fourth, and much less useful, method of rating would be based on the d.c. power input as registered by the flickering plate meter on voice peaks. Since this depends on a number of factors that vary with the transmitter and operator, it is almost meaningless. It is principally of interest in high-power transmitters, since it is the basis for the kilowatt limitation on power input to amateur transmitters.

The same transmitter rated by the four different methods might look like this in a typical case:

Peak envelope power output	250 watts
Peak envelope power input	400 watts
Two-tone power input	250 watts
Average voice power input	200 watts

Too much emphasis cannot be placed on the fact that negligible distortion is a necessary ingredient in all such figures. Peak flattening, for example, will generally increase the power output by a substantial factor. The increase may look good on an r.f. ammeter, but its only contribution is splatter in adjacent channels.

» Using the receiver's b.f.o. is one, but not the only, method of inserting the carrier at the receiver for detection of s.s.b. signals. B.f.o. injection is compared with signal-frequency injection in this article.

Carrier Insertion in S.S.B. Reception

PAUL N. WRIGHT, W9OHM

In the reception of an a.m. signal, it is necessary to provide a device that is able to detect the frequency difference between the sidebands and the carrier wave. This device is the amplitude-modulation detector. The amplitude detector transforms the frequency difference between the sidebands and the carrier wave into pulsating d.c. corresponding to the frequency difference, enabling us to recover at the receiving end the intelligence-bearing frequencies with which we started at the transmitter.

So far as the detector in the receiver is concerned, it isn't particular as to the source of the carrier. It can just as well be furnished from an oscillator at the receiving end. Since an s.s.b. signal is the same as an a.m. signal with the carrier and one sideband removed, the principal thing that needs to be done in order to restore the original intelligence is to replace the carrier on the signal before it reaches the detector in the receiver.

In a superheterodyne receiver, the carrier may be replaced by injecting the carrier from an oscillator at the i.f. frequency into the i.f. section of the receiver, or by injecting the carrier from an oscillator at the signal frequency at the antenna terminals of the receiver.

I.F. Carrier Insertion

If carrier injection from the b.f.o. in the receiver is used, the receiver should be adjusted as follows: First, with the receiver set up in the regular a.m. position, tune for maximum deflection of the S-meter from the s.s.b. signal. Do not touch the dial after this. Next, reduce the r.f. gain to zero and increase the audio gain to maximum. Bring up the r.f. gain until the signal is heard at a comfortable level; then turn on the b.f.o. and carefully adjust the frequency of the b.f.o. until the voice sounds natural. If this procedure is followed closely, little difficulty should be experienced tuning the signal, regardless of which sideband is being transmitted.

In using the b.f.o. method of carrier insertion, it should be pointed out that practical reception of s.s.b. signals depends upon the stability of the h.f. oscillator in the front end of the receiver, as well as the stability of the beat oscillator that supplies the carrier. Any frequency change in the h.f. oscillator produces the same effect as changing the frequency of the transmitter on the other end. The h.f. oscillator in most receivers is fairly stable on the lower frequencies. However, at frequencies above 5 Mc. the stability of many h.f. oscillators leaves much to be desired, when thinking in terms of the stability required from these oscillators when using i.f. carrier insertion.

Signal-Frequency Carrier Insertion

In using carrier insertion at the signal frequency from an external oscillator, the procedure is as follows: With the receiver set up in regular a.m. position, first tune for maximum indication of the S-meter from the signal. Then adjust the frequency of the external oscillator to the approximate frequency of the incoming signal, and increase the amplitude of carrier injection to a point that approximates the amplitude of the s.s.b. signal. When this point is reached, the S-meter will no longer swing with modulation. Carefully adjust the frequency of the external oscillator until the voice sounds natural. Rock the receiver dial back and forth across the carrier. You will easily be able to tell which sideband is being transmitted. As you leave the carrier, on one side the audio will drop off; as you swing on the other side, the audio will come up. The more selective the receiver, the more pronounced this effect.

An s.s.b. signal suffers a certain amount of nonlinear distortion when demodulated by a linear rectifier. Increasing the carrier injection above the 100 per cent modulation point will reduce this distortion to a negligible amount. Increased carrier also helps swamp out adjacent-channel QRM.

The advantages of front-end carrier insertion are:

1) Stability.

2) S-meter reports may be given on s.s.b.

3) Round tables including s.s.b. and a.m. stations become practical, since the receiver remains in the a.m. position at all times.

4) Oscillators in the s.s.b. exciter may be used to furnish the stable carrier to the receiver, providing consistent "on frequency" operation of the transmitted signal.

Point 4 is very important from the standpoint of pleasurable operation and good operating practice of a s.s.b. station. Since the oscillators in the s.s.b. exciter furnish the carrier to the receiver, the transmitted signal is automatically on the same frequency as the received signal. This means that only one oscillator has to be adjusted to get both the receiver and the transmitter on the same frequency.

From "The Reception of Single-Sideband Signals," *QST*, November, 1952.

»A practical circuit for utilizing the signal-frequency method of carrier insertion described in the preceding article.

VFO Signal-Frequency Carrier Injection

PAUL N. WRIGHT, W9OHM

The essential requirements of a signal-frequency carrier generator for reception of single-sideband suppressed-carrier 'phone signals are:

1) Frequency stability.

2) Output amplitude control over a wide range.

3) Sufficient bandspread to simplify the mechanics of precise frequency spotting.

Harmonic Generator

The circuit of Fig. 1 illustrates a stable carrier frequency should be used in order to provide enough output.

Power for this unit may be obtained from the receiver. The heater and plate voltage may be obtained from Pins 2 and 7 and Pin 4, respectively, if the output tube is a 6V6, 6K6, 6F6, etc. The regulated 150 volts may be obtained from Pin 5 of the VR-150 in the receiver. Simply wrap about 3 turns of wire around each tube pin and reinsert the tube in its socket. Be sure the wire insulation is dressed right up to the tube pin to avoid shorts when the tube is plugged in.

A 4 × 5 × 6-inch utility cabinet will house the

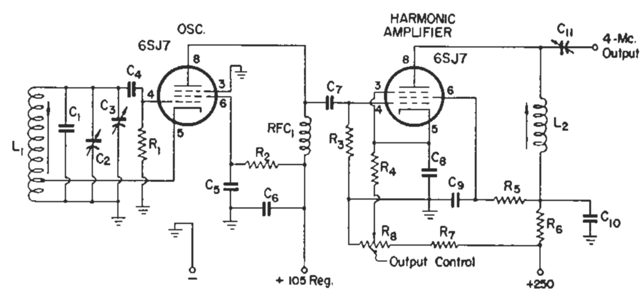


Fig. 1 — Circuit of a l.f. oscillator with good bandspread, for using harmonics for carrier reinsertion.

C_1 — 2500-μμf. zero temp. coefficient.
C_2 — 25-μμf. variable (bandspread control).
C_3 — 140-μμf. variable (bandset control).
C_4 — 47-μμf. zero temp. coefficient.
C_5, C_6, C_8, C_9, C_{10} — 0.01-μf.
C_7 — 47-μμf. mica.
C_{11} — 100-μμf. variable.
R_1, R_7 — 0.1 megohm.
R_2 — 39,000 ohms.
R_3 — 1 megohm.
R_4 — 220 ohms.
R_5 — 82,000 ohms.
R_6 — 1000 ohms.
R_8 — 25,000-ohm potentiometer.
L_1 — 10-μh. coil, on ceramic form.
L_2 — Slug-tuned inductor (CTC 5-Mc. coil).
RFC_1 — 2.5-mh. r.f. choke.

generator using a 1-Mc. VFO and a harmonic amplifier. Carrier output is controlled by R_8 in the circuit. About 40 db. of control is available with R_8. The unit is intended for use on 75 meters; however, output on other bands may be brought up to a usable level by providing a plug-in coil or bandswitching arrangement for the plate circuits of the oscillator and amplifier.

The unit serves as a band-edge marker at 4-Mc. by zero beating WWV at 5 Mc. with the 5th harmonic of the oscillator. The fourth harmonic will then provide a marker on 4 Mc. Of course, it will also provide output at 1-Mc. intervals up through the spectrum, enough to be used as a marker to 30 Mc. However, if it is to be used to provide carrier for receiving s.s.b., plate tank circuits tuned to the desired harmonic

From "Carrier Generators for S.S.B. Reception," *QST*, December, 1952.

unit nicely. Mount the tubes externally on the back of the cabinet and mount the dials on the front. The sides of the cabinet are removable, which makes wiring easy. Mount the resistors on the back of the cabinet, also. This keeps the heat away from the frequency-determining components of the oscillator. R_8 is mounted internally since it must come out the front panel.

Adjust the oscillator tank to 1 Mc. by beating against a local b.c. station or a signal generator. Adjust the slug of L_2 for maximum output on 4 Mc. with R_8 advanced far enough to provide a good signal in the receiver.

General

Many fine articles have appeared in print in the past regarding the design considerations of stable self-controlled oscillators, so the subject will be disposed of with a few reminders:

1) Keep all possible temperature rise away from frequency-determining components.

2) Use ceramic forms for VFO oscillator coils.

3) Mount all components very securely and in such a manner that vibration or jarring the oscillator will not cause any physical displacement of the components.

4) Use regulated heater voltage if possible.

5) Use regulated voltage for the plate and screen of the oscillator.

6) Use widely-spaced variable condensers to minimize the effects of vibration and humidity.

7) Cabinet and chassis should be very rigid and of sturdy construction.

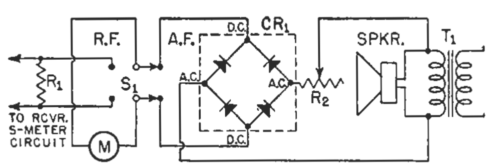
8) Use as little plate and screen voltage as possible, consistent with sufficient output.

9) Care in layout, choice of components, and construction will pay big dividends in stability.

S-METER FOR A.M. AND S.S.B. SIGNALS

Few amateur-type receivers provide for use of the S-meter when the set is tuned to an s.s.b. signal. Fortunately, this deficiency can usually be easily overcome by switching the indicator over to the audio circuit during s.s.b. reception. The circuit diagram shows how the arrangement has been applied to a National type NC-183D receiver.

In the modified circuit, the S-meter terminals are connected to the center arms of a d.p.d.t. toggle switch, S_1. When this new control is set at the "r.f." position, it ties the meter back into the original indicator circuit. When the switch is flipped to the "a.f." position, it connects the meter to the output terminals of an instrument-type full-wave copper-oxide rectifier. The input side of the rectifier is connected



Circuit diagram for the a.m.-s.s.b. S-meter.
R_1 — 1000 to 2000 ohms; see text.
R_2 — 2500-ohm potentiometer.
CR_1 — Instrument rectifier.
M — Original S-meter.
S_1 — D.p.d.t. toggle switch.
T_1 — Receiver transformer.

in series with a calibration potentiometer, R_2, and the secondary of the output transformer, T_1. Naturally, the 'speaker-transformer connections do not have to be disturbed when the modification is being made.

If the receiver on hand does not employ a shunt across the S-meter, it will be necessary to add R_1 to the original indicator circuit. This resistor prevents the a.m.-indicator circuit from opening up whenever the meter is switched over to the s.s.b. position.

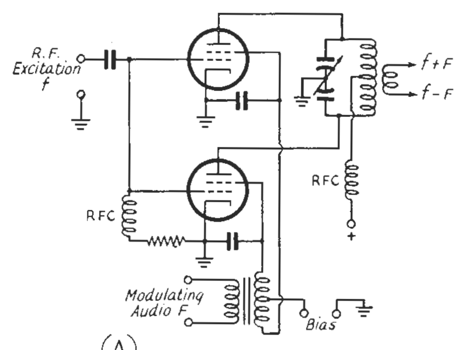
A calibration for the s.s.b. S-meter can be made most easily by comparing its readings with those obtained on a general-purpose test meter. Most of the latter have scales that are directly calibrated in terms of decibels.

— Wayne W. Cooper, YN1WC/W6EWC

» Many of the advantages of s.s.b. transmission stem from not transmitting the "carrier" frequency. The circuits that permit sidebands to be generated in the output without the carrier frequency also being present are called "balanced modulators," and these circuits are relatively new to most amateurs. Here is what balanced-modulator circuits look like and how they work.

Suppressing the Carrier

The carrier can be suppressed or nearly eliminated by an extremely sharp filter or by using a balanced modulator. The basic principle in any balanced modulator is to introduce the carrier in such a way that it does not appear in the output but so that the sidebands will. This requirement is satisfied by introducing the audio in push-pull and the r.f. drive in parallel, and connecting the output (plate circuit) of the tubes in push-pull, as shown in Fig. 1A. Balanced modulators can also be connected with the r.f. drive and audio inputs in push-pull and the output in parallel (Fig. 1B) with equal effectiveness. The choice of a balanced modulator circuit is generally determined by constructional considerations and the method of modulation preferred by the builder. Screen-grid modulation is shown in the examples in Fig. 1, but control-grid or plate modulation can be used equally as well. Balanced-modulator circuits using four rectifiers (germanium, copper oxide, or thermionic) in "bridge" or "ring" circuits are often used, particularly in commercial applications. Two-rectifier circuits are also available, and they are widely used in amateur s.s.b. equipment. Examples of rectifier-type balanced modulators are shown in Fig. 2.



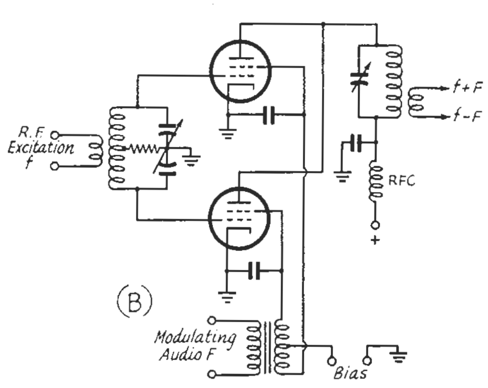


Fig. 1 — Two examples of balanced-modulator circuits using screen-grid modulation. In A the r.f. excitation is in parallel in both tubes, and the audio and output are in push-pull. In B the excitation and audio are in push-pull, the output is in parallel. In either case, the carrier frequency, f, does not appear in the output circuit — only the two sideband frequencies, $f + F$ and $f - F$, will appear. The bias fed to the screens is a practical requirement with all screen-grid tubes for proper linear operation, and is not a special requirement of balanced modulators.

From *The Radio Amateur's Handbook*, 1954 edition.

In any of the vacuum-tube circuits, there will be no output with no audio signal because the circuits are balanced. The signal from one tube is balanced or canceled in the output circuit by the signal from the other tube. The circuits are thus balanced for any value of *parallel* audio signal. When push-pull audio is applied, the modulating voltages are of opposite polarity, and one tube will conduct more than the other. Since any modulation process is the same as "mixing" in receivers, sum and difference frequencies (sidebands) will be generated. The modulator is not balanced for the sidebands, and they will appear in the output.

The amount of carrier suppression is dependent upon the matching of the two tubes and their associated circuits. Normally two tubes of the same type will balance closely enough to give at least 15 or 20 db. carrier suppression without any adjustment. If further suppression is required, trimmer condensers to balance the grid-plate capacities and separate bias adjustments for setting the operating points can be used.

In the rectifier-type balanced modulators shown in Fig. 2, the diode rectifiers are connected in such a manner that, if they have equal forward resistances, no r.f. can pass from the carrier source to the output circuit via either of the two possible paths. The net effect is that no r.f. energy appears in the output. When audio is applied, it unbalances the circuit by biasing the diode (or

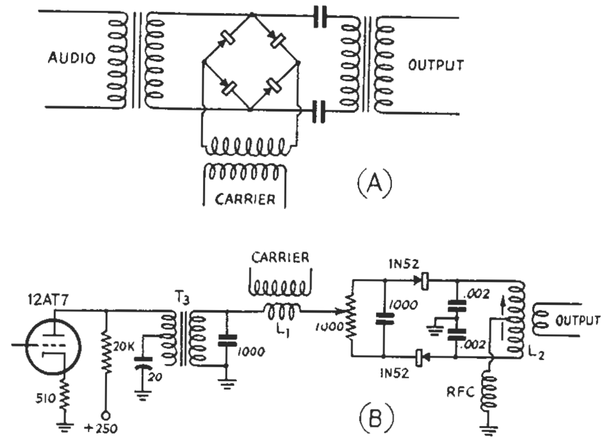


Fig. 2 — Two rectifier-type balanced modulators.

The circuit at A is called a "bridge" balanced modulator and has been widely used in commercial work.

The balanced modulator circuit at B is shown with constants suitable for operation at 3.9 Mc. T_3 is a small step-down output transformer (UTC R-38A), shunt-fed to eliminate d.c. from the windings. L_1 can be a small coupling coil wound on the "cold" end of the carrier-oscillator tank coil, with sufficient coupling to give two or three volts of r.f. across its output. L_2 is a slug-tuned coil that resonates to the carrier frequency with the effective 0.001 μf. across it. The 1000-ohm potentiometer is for carrier balance.

♦

diodes) in one path, depending upon the instantaneous polarity of the audio, and hence some r.f. will appear in the output. The r.f. in the output will appear as a double-sideband suppressed-carrier signal.

In any diode modulator, the r.f. voltage should be at least 6 or 8 times the peak audio voltage, for minimum distortion. The usual operation involves a fraction of a volt of audio and several volts of r.f. The diodes should be matched as closely as possible — ohmmeter measurements of their forward resistances is the usual test.

In any balanced modulator, the stray coupling between the output circuit and the source of carrier frequency should be held to a minimum by proper shielding and circuit isolation. If this is not done, a critical "composite balance" must be obtained in the balanced modulator.

HOW MUCH CARRIER SUPPRESSION?

We sometimes wonder if the s.s.b. gang doesn't worry a little too much about carrier suppression. As a matter of pride it's nice to get the carrier down so low that no one can possibly hear it, but with no carrier at all it leaves nothing for a receiving operator to work on. With a little carrier he can zero-beat, or his YRS-1 can lock in, but with none at all there is absolutely no reference whatsoever. Are we missing a point somewhere along the line?

— "On the Air with Single Sideband," *QST*, Oct., 1951

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In the October column it was wondered out loud if perhaps there wasn't too much emphasis put on carrier suppression, point out that with a little carrier one could readily zero-beat or get his YRS-1 to lock in. The s.s.b. gang picked us up on it, hashed it out over the air, and W3ASW was kind enough to forward the consensus.

". . . we are not quite ready for [less carrier suppression] yet, and the main reason is that not enough fellows have gotten around to the stability of their oscillators. Quite a few of the boys do not have very good carrier suppression, and when they drift their birdies are very annoying. When you have been listening to a station on a certain frequency and another fellow takes over (in the voice-controlled round table) who is 200 to 400 or more cycles away, everyone has to grab for the tuning to clean him up. It may be a coincidence, but the ones who have the best carrier suppression also have the cleanest signals and are most tolerant of tuning.

". . . quite a few owners of YRS-1 adapters say they have disabled the lock-in because near-by interfering signals take control and louse up things in general. Further, the adapter will give a false indication of zero beat when they are trying to set up on us, because it is pulled in by the signal as it approaches the frequency.

"The other evening I was talking to W4OLL about this carrier business and he brought up another instance that makes us against it. We were having our own private QSO on 3999.5 and another couple of fellows were having theirs 875 cycles lower than us. That is too far away for pulling a YRS-1 *but* it falls in the passband of any adapter, and we had a darned birdie to listen to every time this one fellow came on. The other fellow was clean, and the gibberish from his sideband wasn't nearly as annoying as his carrier.

"At any rate, until such time as all fellows have hit upon a simple means for stabilizing frequency and/or removing sufficient distortion and having the correct ratio of highs to lows so that their signals are tolerant of some mistuning, we should continue to suppress carriers as much as we can suppress them!"

And there you have the case of maximum carrier suppression. Bring the subject up again when everyone has the frequency-stability control licked!

— "On the Air with Single Sideband," *QST*, Dec., 1951

» 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 example 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

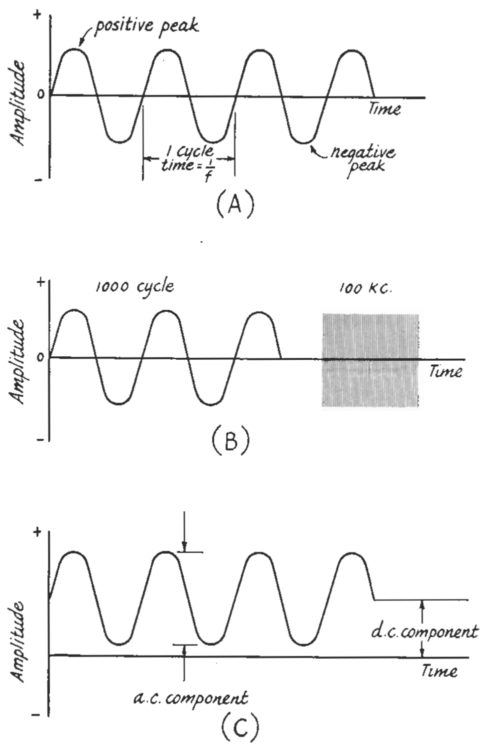


Fig. 1 — The old sine wave, familiar to one and all, is shown at (A). It is a plot of amplitude *vs.* time of a single-frequency a.c. wave.

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).

would fall on both sides of the zero-amplitude axis, but not symmetrically.)

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

From *QST*, April, 1953.

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

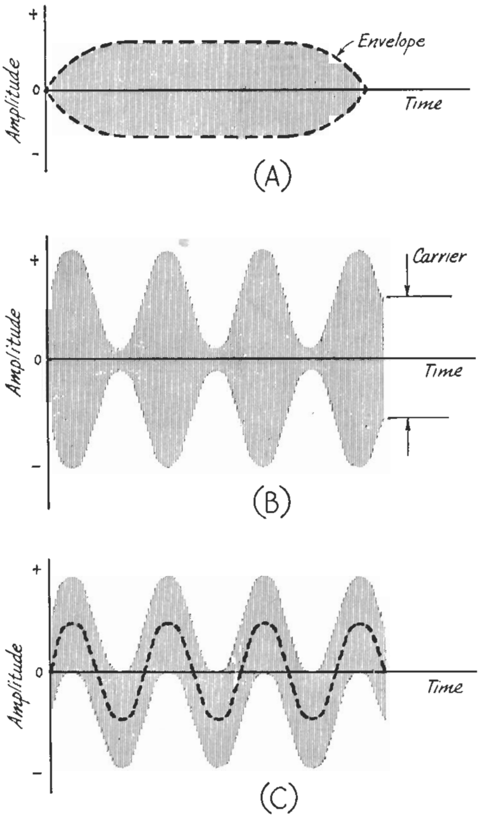


Fig. 2 — **High-frequency waves don't start and stop instantaneously, and the outline of their rise and fall is called the "envelope" (A). Each cycle swings equally above and below the axis, however.**

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, 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.

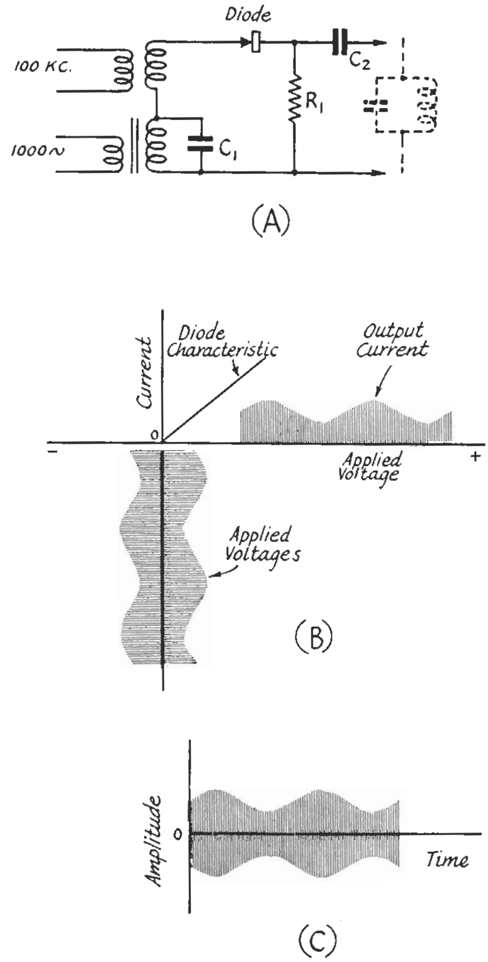


Fig. 3 — **A basic diode-modulator circuit is shown in (A). C_1 and C_2 are by-passes for the 100-kc. signal.**
The modulator action is shown at (B), where the envelope of the superimposed signals becomes a modulated envelope in the output. The a.c. coupling in the output of the modulator, and the tuned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

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 running 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.

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

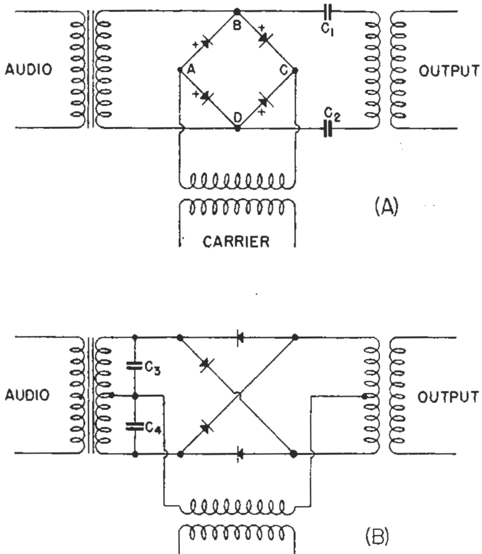
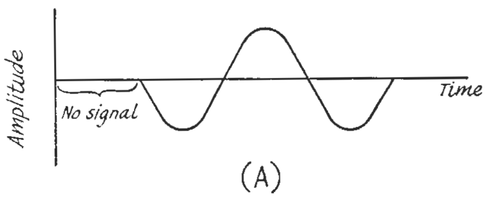


Fig. 4 — **The two common diode balanced-modulator circuits are (A) the bridge and (B) the ring. Condensers C_1, C_2, C_3 and C_4 are r.f. by-pass condensers, used to complete r.f. paths without short-circuiting the audio.**



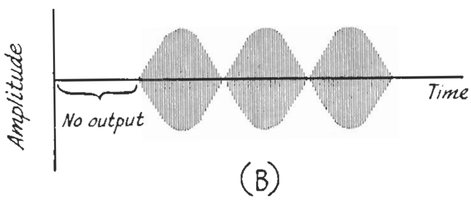


Fig. 5 — **A modulating signal as in (A) gives an r.f. output from a balanced modulator as in (B).**

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 transformer.

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 and copper-oxide rectifiers, 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. The rectifiers are in common use up to 4 Mc., and will be usable at higher frequencies with careful construction. A bugaboo at the higher frequencies is the variation in internal capacity of the rectifiers, and consequently they must be operated at lower impedance levels as the operating frequency is increased. From 600 to 1000 ohms is a practical level at 500 kc., but 50 to 100 ohms is recommended at 4 Mc.

In some of the first single-sideband articles back in 1948 *QSTs*, you will run across the abbreviation "s.s.s.c." for "single-sideband suppressed-carrier." Later on the abbreviation "s.s.b." was adopted, to conform with commercial practice and because the amateur work eventually became carrierless at all times, so "suppressed carrier" is understood when s.s.b. is talked about.

» *Here is another type of diode balanced modulator that will be found in many pieces of s.s.b. gear. Its operation is explained here — an example of its application is shown later in this book, in the crystal-lattice filter s.s.b. exciter by Weaver and Brown.*

The Series Balanced Modulator

FRED M. BERRY, WØMNN

THE "series" balanced modulator is presented because of its simplicity, good linearity, and excellent carrier suppression. In converting from audio to r.f., no iron-cored transformer is required in the audio portion, which makes it possible to keep the hum down without expensive components.

The basic circuit of the series balanced modulator is shown in Fig. 1. The action of the carrier

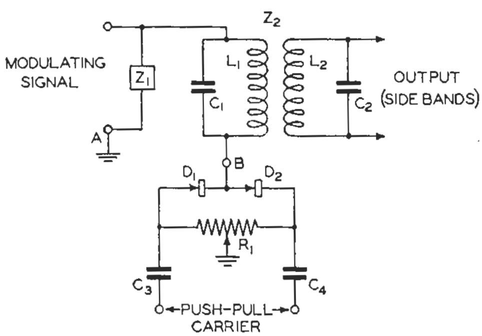


Fig. 1 — The basic series balanced-modulator circuit. L_1C_1 and L_2C_2 resonate at the output frequency, and R_1 is the carrier balance control.

is to switch point B to ground at the carrier rate. A basic requirement of the circuit is that Z_1 (the impedance across which the modulating voltage is developed) have low impedance for the carrier frequency, and that Z_2 (the impedance across which the output is developed) have low impedance for the modulating frequency. In practice, this requirement will be met automatically by using parallel-tuned circuits at Z_1 and Z_2.

Z_1 and Z_2 should be approximately the same impedance, but exact equality is not necessary unless a very minimum of loss is demanded. Impedance ranges between 500 and 50,000 ohms have been used — the optimum impedance depends upon the carrier power and the bias built up across R_1 and C_3 and C_4.

Since the diode resistances appear in series as the carrier keys the diodes, the diode resistance must be low for low-loss operation and to prevent mismatch in filter input and output impedances.

From *QST*, September, 1952.

Practical Circuits

The use of a series balanced-modulator circuit in the audio modulator ahead of the sideband filter is shown in Fig. 2. The 6C4 cathode follower eliminates the need for any step-down transformer, and C_2 offers a low-impedance return to ground for the carrier. The coupling coil, L_1, should be sufficient to develop about 6 volts across R_3. The alternative connection for the coupling coil eliminates R_3, C_3, and C_4 and uses R_4 for the balance control. Some improvement in linearity is obtained with this connection, but it requires more oscillator power to develop the proper voltage at the diodes.

An example of the series circuit used in a 0.5 to 4-Mc. converter is shown in Fig. 3. Here a 6AL5 is shown as the modulator diode, and a 6BH6 phase splitter is used to get push-pull excitation from the VFO. The VFO should de-

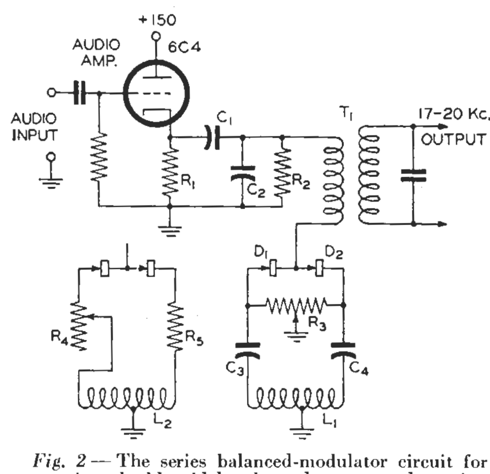


Fig. 2 — The series balanced-modulator circuit for generating double sidebands and suppressed carrier. An alternative oscillator connection is shown at the lower left.

C_1, C_3, C_4 — 0.5 µf. 200-volt paper.
C_2 — 0.05 µf.
R_1 — 1500 ohms.
R_2 — 2200 ohms.
R_3 — 3000-ohm potentiometer.
R_4 — 200-ohm variable.
R_5 — 100 ohms.
L_1, L_2 — Coupling coil on oscillator coil.
D_1, D_2 — 6AL5, 1N34 or copper-oxide rectifier.
T_1 — 500-ohm-to-sideband-filter coupling transformer (toroid core).

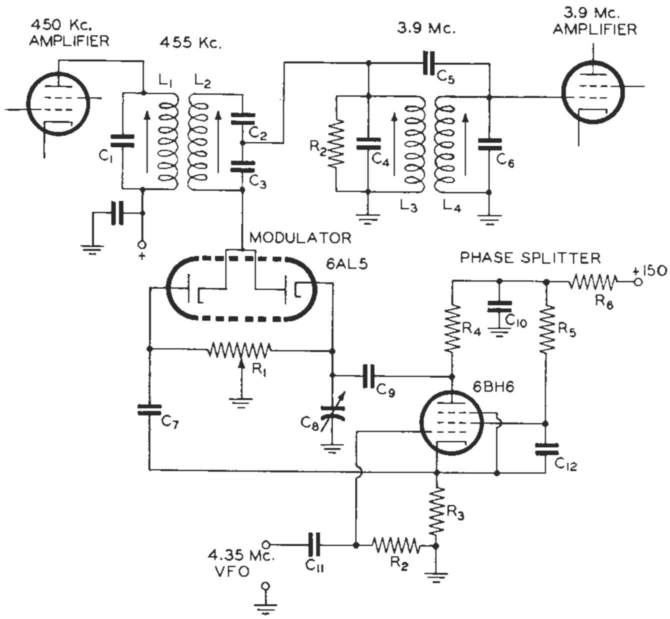


Fig. 3 — The series balanced modulator for heterodyning a 450-kc. s.s.b. signal to the 75-meter band with VFO control.

C_1, C_4, C_6 — 120 $\mu\mu$f.
C_2 — 200 $\mu\mu$f.
C_3 — 300 $\mu\mu$f.
C_5 — 15 $\mu\mu$f.
C_7, C_9 — 0.001 μf.
C_8 — 45-$\mu\mu$f. variable trimmer.
C_{10}, C_{11}, C_{12} — 0.01 μf.
R_1 — 30,000-ohm potentiometer.
R_2, R_5 — 0.1 megohm.
R_3, R_4 — 10,000 ohms.
L_1, L_2, L_3, L_4 — Slug-tuned coils to resonate as shown.

liver enough voltage to furnish approximately 6 volts across R_1. R_1 and C_8 are adjusted for best carrier balance, as indicated by the absence of VFO-frequency signal in the output of the 3.9-Mc. amplifier. The inductive coupling between L_3 and L_4 is adjusted to give best band-pass characteristics over the 200-kc. 'phone band. This circuit also illustrates the fact that the diodes can be connected between ground and point *A* or point *B* of Fig. 1.

The modulator circuit of Fig. 2 can, of course, be used at 450 kc. (ahead of a crystal filter) with germanium diodes or a 6AL5 for the modulator and with a 450-kc. tuned circuit at T_1.

A double-balanced series-modulator circuit can be built for frequency conversion in which neither the signal nor the oscillator frequency appears in the output. In such a circuit, the signal is fed to the center tap of the primary. One side of the push-pull oscillator is fed through resistors and diodes (of opposite polarity) to the ends of the primary, and the other side of the push-pull oscillator is also fed through resistors and reversed diodes to the ends of the primary. A diode anode and a diode cathode connect to each end of the primary. Two of the resistors that run to one end of the primary should be made variable, for adjusting the carrier (oscillator) balance.

General

Referring again to Fig. 1, care should be taken to keep the capacity to ground low at point *B* (or *A*, if the modulator is connected on that side). The primary-to-secondary capacity of Z_2 should also be held as low as possible.

The circuit also works well as a demodulator for receiver work where a locally-injected carrier is used (s.s.b. or c.w.). Its main advantage is that no critical balancing or special components are required. In a receiver, the push-pull carrier and its suppression permits the use of high carrier level for maximum linearity without introducing oscillator noise.

Intermodulation products are better than 60 db. down, using a carrier level of 3 volts and a maximum signal level of 0.1 volt.

CARRIER SUPPRESSION WITHOUT A BALANCED MODULATOR

A balanced modulator is not absolutely essential in a single-sideband suppressed-carrier exciter. The Edmunds crystal-filter exciter (described later) used a notch in the filter characteristic to suppress the carrier, although balanced modulators are perhaps less critical. The absence of a balanced modulator from a phasing-type exciter is perhaps best illustrated by one of the rigs used at OZ7T. It used audio and r.f. phase-shift networks, of course, but no balanced modulators. Instead, a pair of grid-modulated 6SJ7s were used to generate s.s.b. *with* carrier, and the carrier was then balanced out by introducing equal and opposite carrier through a third 6SJ7. The resultant s.s.b.-suppressed-carrier signal was amplified by an EBL21 (similar to 6L6).

A system like this has no inherent balance and is rather sensitive to variations in voltage. It is recommended that it be used only with stabilized power supplies.

» If you feel more at home with plate modulation than with a diode modulator, use one of the circuits described below. They show plate modulation applied to balanced modulators, both with and without the need for transformers. And they offer the opportunity to run things at a slightly higher power level than is customary with diodes.

Plate-Modulated Balanced Modulators

BYRON GOODMAN, W1DX

THE circuit of a plate-modulated balanced modulator is shown in Fig. 1. If you look at just one tube, you will see that it is simply the familiar plate-modulated triode, with the audio modulating power fed into the negative instead of the positive lead. The condensers from cathode to ground are r.f. by-passes. So far, no difference. However, when you look at both tubes, you will see that the excitation is fed to the grids in parallel, while the output is connected in push-pull. Since each tube acts as a neutralizing circuit for the other, none of the excitation voltage passes through the stage. However, when the tubes are modulated with push-pull audio, the double sidebands (minus carrier) appear in the output, and we have a "balanced modulator." No steady plate power is applied to the tubes, because any such power does no good and only dissipates itself on the plates of the tubes. The output sideband power is a transformation of the applied audio power, with the usual tube loss. The circuit is quite tolerant so far as excitation is concerned, requiring only that there be sufficient drive for Class C operation over the range of modulating voltages. At the small powers involved, it also seems to be reasonably tolerant of loading, although it does require some load (which can be the losses in the circuits) and it can't stand overloading beyond its linear operating conditions. It is fairly simple and something everyone can understand.

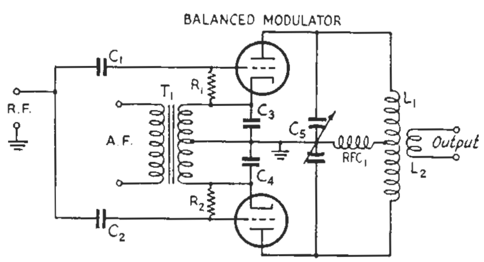


Fig. 1 — The plate-modulated triode balanced-modulator circuit. The r.f. excitation is fed to the grids in parallel and the output is taken in push-pull. The audio modulation is applied in push-pull in the negative plate lead.

For an exact balance, both triodes should have identical characteristics and interelectrode capacities. However, it will be found that two tubes of the same type will normally give adequate carrier suppression, and even dual triodes like the 6SN7GT can be used without selection. When using dual triodes, which usually do not have identical interelectrode capacities, it may be helpful in reducing the carrier to add additional grid-plate capacity to one triode. Only a slight amount is necessary — it can be nothing more than a wire soldered to the grid socket pin and bent toward the plate socket pin.

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Al Prescott, W8DLD, devised a variation of the plate-modulated balanced modulator which he calls a "series" balanced modulator. With it he avoids the use of transformers in his phasing-type exciters. The circuit is shown in Fig. 2. Although he uses the phase-shift networks that involve coupling tubes, there is no good reason why the tubeless or "passive" types couldn't be used. Potentiometers R_5 and R_6 control the carrier balance (or injection, if you prefer). The switch S_2 simply disables one modulator, for use when tuning up with the two-tone method.

The grid currents to the 6SN7s run about 2 to 3.5 ma., and if they aren't reading within about 10 per cent of each other the audio-balancing controls (not shown — they are ahead of the phase-shift networks) will seem not to work.

The r.f. phasing networks are a little different than may be found elsewhere. W8DLD uses wire-wound resistors on 4 Mc. and makes use of their inherent inductance. At 4 Mc., L_4R_{15} is simply a 1250-ohm 10-watt Ohmite wire-wound resistor, and $L_5R_{16}C_{31}$ is a 1000-ohm 10-watt Ohmite wire-wound. At 14 Mc., L_4R_{15} is 69 inches of No. 26 enameled wound on a 1000-ohm IRC BT2 resistor, and $L_5R_{16}C_{31}$ is 36 inches of No. 26 enameled on a similar resistor, with a small mica compression trimmer across the works.

The output of the unit runs about 3 watts with single-tone modulation and good carrier suppression.

From "A 75- and 20-Meter Single-Sideband Exciter," *QST*, November, 1949.

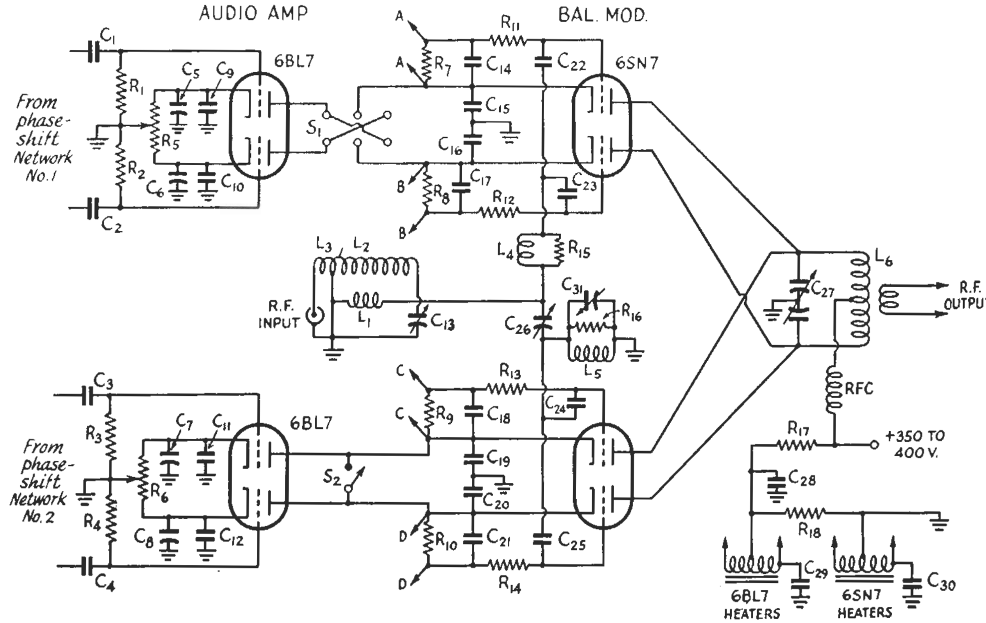


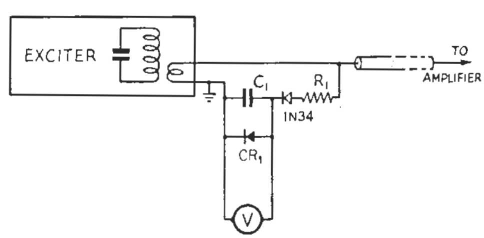
Fig. 2 — The series balanced-modulator circuit of W8DLD eliminates the need for audio coupling transformers. By-pass condensers and metering circuits make it look more complicated than it actually is.

C_1, C_2, C_3, C_4 — 0.01-μf., 600 volts.
C_5, C_6, C_7, C_8 — 10-μf. electrolytic.
C_9, C_{10}, C_{11}, C_{12}, C_{14}, C_{15}, C_{16}, C_{17}, C_{18}, C_{19}, C_{20}, C_{21}, C_{28}, C_{29}, C_{30} — 0.001-μf. ceramic or mica.
C_{13} — 100-μμf. midget variable.
C_{22}, C_{23}, C_{24}, C_{25} — 47-μμf. ceramic or mica.
C_{26} — 75-μμf. midget variable.
C_{27} — 100-μμf.-per-section variable.
C_{31} — See text.
R_1, R_2, R_3, R_4 — 0.47 megohm.
R_5, R_6 — 3000- or 4000-ohm potentiometer, 4 watts.
R_7, R_8, R_9, R_{10} — 47 ohms, ½ watt.
R_{11}, R_{12}, R_{13}, R_{14} — 4700 ohms, ½ watt.
R_{15}, R_{16} — See text.
R_{17} — 0.1 megohm, 2 watts.
R_{18} — 0.2 megohm, 1 watt.
L_1 — 13 turns interwound with L_2.
L_2 — 39 turns No. 26 enam.
L_3 — 3 turns.
L_1, L_2, and L_3 wound on 1⁵⁄₁₆-inch diam. tube base.
L_4, L_5 — See text.
S_1 — Sideband selection switch, d.p.d.t. toggle.
S_2 — Balanced-modulator disabling switch, s.p.s.t. toggle type.
Points *A*, *B*, *C*, and *D* go to a meter switch (25-ma. meter) for measuring grid current.

A CARRIER NULL INDICATOR

The usual method of balancing out the carrier of a s.s.b. exciter is to tune it in on the receiver, turn down the transmitter audio gain and then adjust the transmitter carrier control for minimum on the S-meter. This can be very bothersome when one is often reinserting and removing carrier for demonstration or educational purposes. Mark Moynahan, W2ALJ, passes along a simple gadget that he uses with his Central Electronics 10-A exciter, and it has the advantage that it can be left in the circuit at all times, without pinning the needle when the operator starts talking. It shows excellent sensitivity in the milliwatt range and yet it is not overloaded by the 10-watt peak output of a 6AG7.

As can be seen, it uses a 1N34 r.f. voltmeter, with a selenium rectifier across the indicator. At low voltages the selenium rectifier has little or no effect, and the residual carrier will indicate in a normal manner. At higher voltages the selenium conducts and prevents "pinning" of the meter. At W2ALJ, final-amplifier outputs of 3 and 225 volts (across a 52-ohm line) give voltmeter readings of 0.2 and 2.6 volts. The basic idea is a useful one that should find other applications around the ham shack.



W2ALJ uses the above circuit for the continuous monitoring of carrier suppression. A selenium rectifier, CR_1, "shorts" the meter at high values.

C_1 — 0.005 μf.
R_1 — 2000 ohms.
CR_1 — 120-volt 40-ma. selenium rectifier.
V — 0-3 1000-ohm-per-volt voltmeter.

» *Proper operation of a suitable detector circuit can improve the effective selectivity of any s.s.b. (or c.w.) receiver.*

Demodulation and Selectivity in S.S.B. Reception

OSWALD G. VILLARD, JR., W6QYT

With relatively little difficulty, the amateur can convert his existing receiver for single-sideband reception, and obtain performance rivaling that of the finest commercial installations.

It is the purpose of this article to review the mechanism of detection both in the case of double- and single-sideband reception. The way in which the shortcomings of double-sideband detection can be circumvented by single sideband will then be outlined. It is pointed out that while some of the advantages of single-sideband detection can be realized with unmodified standard communications receivers, a great improvement can be obtained through addition of a simple external single-sideband detector and low-pass audio filter.

The fact that single-sideband transmission makes possible a 50 per cent saving in transmitted bandwidth is almost academic when considered in the light of the effective selectivity of present-day receivers. It is all very well to halve the spectrum occupied by the transmitter, but if the receiver doesn't show the difference nothing has been gained.

Selectivity

To clarify the concept of selectivity in the case of conventional detection, it is helpful to consider a specific case. Suppose one is listening to a station that is very weak — perhaps one microvolt across the receiver input terminals. For a given i.f. gain, a certain voltage is delivered to the diode detector. Now what we wish to do is to prevent any station on any *other* frequency from delivering a signal of approximately equal strength to the detector. It is a characteristic of linear detection — or rectification, as it is sometimes called — that if another signal, no matter how far it may be separated in frequency from the one being listened to, *does* get through at the same strength, it will also be rectified and the modulation on it, as well as the modulation on the desired signal, will be heard. The two signals may actually be so far apart in frequency that the beat between the carriers is inaudible, yet both modulations will be heard if the two signals are of roughly equal strength at the detector.

From "Selectivity in S.S.S.C. Reception," *QST*, April, 1948.

Now it is an inherent characteristic of linear diode detectors that if either signal becomes two or three times as strong as the other, the modulation on the weaker tends to be completely suppressed. It is a desirable one, in point of fact, because it means that the *effective* selectivity of a receiver with a linear detector — for double-sideband reception — is actually greater than that of the receiver's i.f. circuits alone.

However, suppression of the weaker signal can be very annoying if that signal happens to be a weak DX station, say some five or ten kilocycles away from a strong local. When the local comes on, the DX signal disappears. All that comes out of the loudspeaker is the local's voice. Yet the DX signal is still being received; the suppression effect has simply taken him out.

In order to set up some performance specifications for an "ideal" ham receiver, we might consider what sort of selectivity problem we are actually up against. It is clear that the ideal receiver is one that will receive a barely audible signal without interference on one channel, while the loudest signal we would normally expect to encounter is blasting away on the channel immediately adjacent. We know that the weakest signal we can receive is approximately one microvolt. To find the strongest signal we are likely to have to reject (locals around the corner excepted) it is necessary to do some estimating, and estimates based on propagation factors as well as actual experience show that the ratio of strongest signal received to weakest signal received may occasionally exceed 1000 to 1.

Now the question is, how much frequency separation is necessary between a 1-microvolt desired signal and a 1000-microvolt undesired signal, in order for the latter to be attenuated down to the 1-microvolt level? The answer will be found in published receiver selectivity curves. In one commercial receiver, which has four i.f. tuned circuits, the response is down 1000 times at plus or minus 8 kilocycles from the center of the passband for the narrowest setting of the "bandwidth" control. For another receiver, which has three i.f. tuned circuits, the response is down 1000 times at plus 11 or minus 10 kilocycles. Other communications receivers will, in general, fall in the same range.

This means, then, that if we are listening to a

1-microvolt DX signal, the nearest 1000-microvolt interfering signal must be at least 8 kilocycles away in one case, and 10 to 11 kilocycles away in the other case, in order for both DX and undesired signals to be of equal strength at the second detector. When both signals have equal strength, both stations will be heard; for the undesired signal to be eliminated completely, it must be attenuated still further, until it is less than 1/3 or 1/4 as strong as the desired signal.

The tendency for a strong adjacent station completely to blot out the one to which we are listening is familiar to everyone, and is the reason for the desirability of extra i.f. selectivity such as is provided by arrangements like the Q5-er. Without such extra selectivity we cannot make full use of the frequency space now available to us.

Single-Sideband Detection

The situation is quite different in the case of single-sideband reception, because a fundamentally different process of detection is used. S.s.b. or c.w. signals are detected by frequency conversion, rather than by rectification. Here it is helpful to review some theory. A single-sideband voice signal may be thought of as a band of frequencies simply displaced in the frequency spectrum. To each of the frequencies present in the voice wave, a constant frequency is added. Thus a speech sound, which might consist at some instant of three component frequencies — e.g., 500, 1000 and 2500 c.p.s. — can be translated into the radio-frequency spectrum by adding a constant 1,000,000 cycles per second to each component. We then have three new frequencies, namely 1,000,500, 1,001,000 and 1,002,500 cycles per second, forming a voice single sideband that can be transmitted by radio. A single-sideband transmitter, then, is fundamentally a frequency-translating device that shifts the incoming speech frequencies to the desired position in the radio-frequency spectrum. To receive these signals, it is only necessary to reverse this process: by subtracting the constant frequency of 1,000,000 cycles we can recover the original speech frequencies of 500, 1000 and 2500 cycles.

Note that the reception of code signals is carried out by a very similar process. Assume an incoming keyed c.w. signal of 1,000,000 cycles per second. If we subtract exactly 1,000,000 c.p.s., what we have left is, of course, zero frequency, or keyed d.c. In actual practice, something like 999,000 cycles per second is subtracted. The c.w. signal is thereby translated to a frequency of 1000 c.p.s., which, when amplified and fed to a loudspeaker, is heard as an audible tone.

Now, frequency conversion, in reception, is a process with which everyone is familiar. For code reception in the ordinary receiver, the frequency conversion is actually done in two steps: the local oscillator converts the incoming signal to the i.f.; and the beat oscillator, in conjunction with the second detector, converts the i.f. down to an audio frequency. For either c.w. or single-sideband voice reception, the usual diode second detector could equally well be replaced by a converter tube. For both types of transmission we are interested in frequency conversion — nothing else. We do not want any normal rectification to take place, particularly in the conversion at the second detector, because this rectification would permit the modulation on undesired amplitude-modulated signals to be heard along with the desired signal resulting from the frequency-translation process.

Overcoming Rectification Effects

One way to suppress the spurious signal resulting from rectification is to make the voltage injected by the beat-frequency oscillator very strong in relation to the incoming r.f. signals. In the ordinary communications receiver the amplitude of the b.f.o. voltage is fixed, and the only way to make it strong in relation to the incoming signals is to weaken the latter. This is why it is always recommended that a.f. gain be on full and r.f. gain be kept at a minimum for single-sideband reception. Under these conditions the audio output resulting from rectification of the incoming signals is small compared to the audio output resulting from the beat between the desired single-sideband signals and the b.f.o.

However, even then the rectified audio is unfortunately not negligible. The usual diode second detector is designed as a rectifier rather than as a pure frequency converter, and its use in the latter service is a compromise. It is generally considered that the audio output attributable to rectification is negligible in comparison with the desired output when the incoming signal voltage is roughly one-tenth as strong as the local-oscillator voltage. When this ratio is achieved by reducing i.f. gain in the ordinary communications receiver, the audio output from the detector is down quite a bit and the reserve audio gain may not be able to compensate for the loss.

Furthermore, it often happens that although the c.w. or single-sideband signal one is listening to is only one-tenth as strong as the local oscillator, an interfering signal on an adjacent channel may be one-half or one-third as strong. It will therefore be rectified, giving rise to interference that will be equally bad no matter how far the carrier frequency of the interfering signal is separated from that of the desired signal.

An objection to running a receiver at reduced r.f. gain is the resulting reduction in signal-to-noise ratio in the first r.f. stage, unless the set happens to be one in which this stage operates at full gain all the time.

Of course, it is possible to use the opposite approach and increase the amplitude of the injected b.f.o. voltage while keeping the r.f. gain normal. This procedure is to be recommended, but it can only be carried to the point at which the second detector overloads.

For these reasons, it is hard to get the full benefits of single-sideband reception in QRM reduction when using a conventional receiver with b.f.o. But it is not difficult to build an external frequency-converter unit (or detector unit), especially designed for single-sideband reception, which can be added to any communications receiver. The method of operation of such a unit is somewhat different from that of ordinary converters, so here we must digress a moment to review some more theory.

Frequency Conversion vs. Rectification

There are two basic ways in which frequency conversion may be carried out. In the first, the incoming signal is linearly added to the oscillator voltage, and the combination is rectified in a diode detector. It will be found that the envelope of the combination pulsates at a frequency that is the difference between the incoming signal (or signals) and the local oscillator. Since a diode detector follows the envelope variations, the audio output is the desired difference frequency. This is how the b.f.o.-second-detector frequency conversion works. In the second method, the local oscillator modulates the incoming radio-frequency signal, thereby setting up two sidebands whose frequencies are the sum and difference of the incoming and oscillator frequencies. The lower sideband, or difference frequency, is the desired audio signal. The 6L7 converter tube, for example, works on this principle: the local oscillator simply suppressor-grid-modulates the signals being carried from the control grid to plate by the electron stream. The 6L7 is essentially a modulated amplifier.

In a modulated-amplifier converter where we are changing the incoming frequencies down into the audio range — instead of into the i.f. range as in most receivers — we must avoid distortion that would cause partial rectification of the incoming signals and thus produce undesired a.f. output. As an amplifier, therefore, the 6L7 must be very linear. The best way to keep nonlinear distortion low in any amplifier is to restrict the amplitude of the incoming signal. But where such a restriction is impractical, as is the case if the frequency converter must operate at a fairly-high signal level, it is possible to use the expedient illustrated in Fig. 1 — the push-pull or balanced frequency converter. In this circuit, each tube produces the same amount of audio output from rectification caused by nonlinearity of the grid-voltage/plate-current curve. But since the tubes are connected in push-pull so far as their outputs are concerned, these audio signals cancel out. The local-oscillator voltage, on the other hand, is fed to the two tubes in push-pull, and consequently the audio outputs resulting from the beat between this oscillator and the incoming signal add up in phase at the output transformer. Thus the desired beats are heard, while undesired signals because of rectification are balanced out.

Actually, the rejection of the unwanted signals cannot be absolutely complete because of the presence of third-order distortion which is not affected by the push-pull connection. However, remote-cut-off tubes such as the 6K7 and 6L7 are especially designed to have low third-order distortion, and their use makes possible a rejection that is quite adequate in practice.

Adjusting for Balance

The correct "balance" of the balanced detector circuit may easily be found. An ordinary modulated signal is applied to the detector, and the beating oscillator is either turned off or, preferably, is detuned so far away from the signal that any beats between it and the signal are above the limit of audibility. Then the amplitude of the signal is increased until the modulation on it just begins to be heard — in other words, the signal just begins to ride through. Disregard any distortion. To balance the detector, the cathode balancing resistor in Fig. 1 is adjusted until the audio output is minimum.

A balanced detector of the type shown in Fig. 1 may easily be added to any standard receiver without disturbing normal operation in any way.

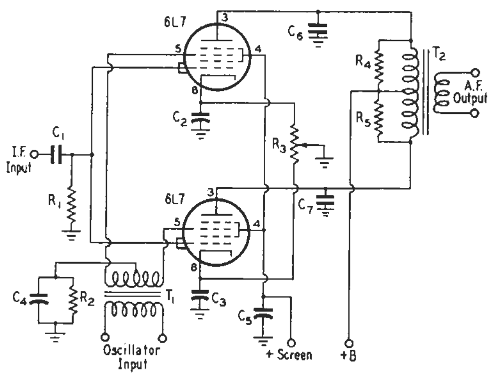


Fig. 1 — **Balanced frequency converter for single-sideband reception. This circuit substitutes for the conventional rectifier (usually a diode) serving as a second detector in a superhet receiver.**

C_1, C_6, C_7 — 0.0022 μf.
C_2, C_3 — 50-μf. electrolytic, 25 volts.
C_4 — 0.01 μf.
C_5 — 0.1 μf.
R_1 — 33,000 ohms.
R_2 — 15,000 ohms.
R_3 — 500-ohm variable.
R_4, R_5 — 10,000 ohms.
T_1 — i.f. transformer with center-tapped secondary (coupled to beat oscillator).
T_2 — Audio transformer; push-pull primary capable of carrying about 10 ma.

A cathode follower will serve to couple energy from the last i.f. stage of the receiver to the external adapter. An external beating oscillator must be provided, however, and likewise a separate audio amplifier, unless the output of the single-sideband detector is fed back into the set's audio system in some way.

In actual operation, a detector of this sort makes possible an enormous improvement over single-sideband reception by means of the set's own b.f.o. It is no longer necessary to keep the i.f. gain control down low in order to keep the b.f.o. voltage large compared with the signal voltage. Strong adjacent modulated signals no longer ride through and cause distraction by adding another voice to the one being copied; they are heard, if at all, only as "monkey chatter," or scrambled speech.

A Wien bridge or "Hetrofil" may be used with good effect after a detector of this sort to eliminate steady tones caused by the beats between the carriers of interfering double-sideband signals and the conversion oscillator. A sharp-cut-off low-pass filter connected to the output provides an inexpensive way of achieving the 'phone man's dream of a virtually square-topped straight-sided passband. This is possible because the over-all selectivity is largely determined by the audio filter, whose performance can be made very good. To achieve the same square-topped bandpass characteristic through i.f. selectivity alone would probably require conversion to a low i.f. frequency plus use of many cascaded over-coupled tuned circuits, which are awkward to align. Simple selective circuits such as crystal filters, the Q5-er, etc., provide a peaked, rather than a square, response characteristic, unless special alignment procedures are used.

The combination of single-sideband transmission, balanced detector, and low-pass filter has only one drawback, and that is the audio image. Signals on either side of the beating-oscillator frequency can produce an audio output within the passband of the filter. Therefore the combination has twice the effective bandwidth required for single-sideband reception.[1] Nevertheless, even by itself the balanced-detector low-pass filter combination represents a considerable improvement over present-day receiving techniques.

[1] This image, of course, can be reduced or eliminated if the detector system is preceded by an i.f. system having a bandpass characteristic just wide enough to accommodate a single-sideband signal. The Q5-er in conjunction with a communications receiver can give excellent performance in this respect. However, such a system is more difficult to tune; the received sideband must be properly placed on the rounded nose of the selectivity curve. — Ed.

Conclusion

The conclusions reached in this article may be summarized in the following way. With conventional modulation the linear-detection process used, plus finite receiver i.f. selectivity, results in two undesirable features: first, the possibility that the modulation on strong interfering signals quite far from the frequency of the desired signal will ride through; and second, the "suppression" effect which results in complete disappearance of the desired signal when the interfering signal is very strong. These two disadvantages of conventional reception, familiar to all 'phone men, tend to prevent maximum utilization of existing frequency assignments in the sense that very weak stations cannot be copied immediately adjacent to very strong ones. Single-sideband reception by means of the b.f.o., on standard receivers, reduces the suppression effect but still is vulnerable to the modulation on undesired signals because of the possibility of rectification occurring along with frequency conversion. Finally, single-sideband reception by means of a standard receiver equipped with a balanced detector eliminates both the suppression effect *and* the possibility of interference due to rectification. In a combination of this sort, audio selectivity becomes the complete equivalent of i.f. selectivity (except for the audio image), provided that the i.f. selectivity of the receiver is sufficient to prevent strong adjacent interfering signals from actually overloading the balanced detector. Should such overloading become a problem because of insufficient i.f. selectivity, it is possible to reduce the i.f. input to the balanced detector and to make up for the loss by increased audio gain.

SINGLE-SIDEBAND DEFINITIONS AND JARGON

Anti-Trip — System of voice-controlled break-in operation that prevents signal from loudspeaker actuating the transmitter.

Barefoot — Operation of s.s.b. exciter without high-powered linear amplifier.

Modulation Envelope — Envelope of modulated signal. When recovered by rectification, it is the modulating signal of an a.m. signal. In s.s.b. the rectified envelope does not represent the modulating signal — the carrier must first be reinserted. Envelope of s.s.b. signal is of prime importance in determining limits of linearity and power of amplifier.

Phasing Exciter — S.s.b. exciter using the phasing principle. See page 89.

***Q* Multiplier** — Electronic means for increasing the selectivity of a circuit. Often used at receiver intermediate frequencies in feed-back circuit to reject narrow band of frequencies.

QT-1 — A commercial anti-trip circuit.

Select-O-Ject — Audio-frequency selective/rejection circuit using *RC* circuits.

Slicer, Sideband or Signal — A selectable-sideband adapter for receiving, using audio phasing system as described on page 103.

YRS-1 — Commercial selectable-sideband adapter using phasing principle similar to the Slicer. Also includes a.f.c. (automatic frequency control) circuit for locking on carrier.

Zeroing In — Tuning suppressed carrier frequency of transmitter to zero beat with suppressed carrier frequency of other s.s.b. transmission.

» Here is a useful detector for s.s.b. reception, devised by Murray Crosby, W2CSY. It gets its name from its operation — the output signal is proportional to the "product" of signals in two channels.

The Product Detector

BYRON GOODMAN, W1DX

THE "product detector" circuit of Fig. 1 is useful in s.s.b. and c.w. reception because it minimizes intermodulation at the detector and doesn't require a large b.f.o. injection voltage. Two triodes are used as cathode followers, for the signal and for the b.f.o., working into a common cathode resistor (470 ohms). The third triode also shares this cathode resistor, but has an audio load in its plate circuit. The grid of this third triode is grounded for signal but has an adjustable negative bias obtained from a 5000-ohm potentiometer. The signals and the b.f.o. mix in this third triode, but its adjustable bias permits setting the bias on the signal cathode follower (through the common cathode resistor) to the point where minimum intermodulation takes place in the cathode follower. Thus if the b.f.o. is turned off, a modulated signal passing through the signal cathode follower will yield no output from the detector at one setting of the 5000-ohm potentiometer. Turning on the b.f.o. brings in the audio, because now the detector output is the product of the signals in the two channels.

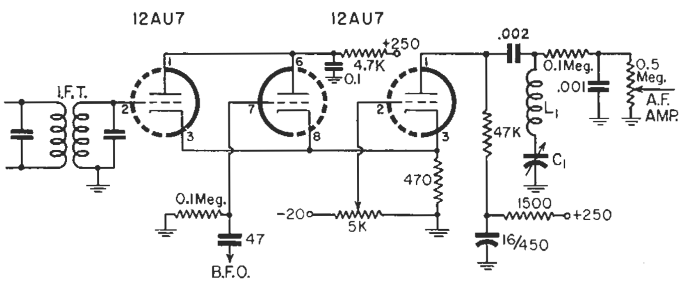


Fig. 1 — The "product detector" circuit is useful for minimizing intermodulation that occurs in a detector circuit.
L_1C_1 is a low-C circuit resonant at the i.f. It attenuates the b.f.o. and signal voltages that might otherwise reach the audio system and reduce its signal-handling capabilities.

The negative bias supply should be a well-filtered affair, because any a.c. introduced on the grid of the third triode will appear in the output.

Another advantage of the product detector is that a signal-level indicator circuit connected to the grid of the signal-circuit cathode follower (left-hand triode in Fig. 1) doesn't indicate b.f.o. voltage, so the signal-level meter reads the same with b.f.o. on or off.

From "An All-Purpose Super-Selective I.F. Amplifier," *QST*, March, 1953.

PRODUCT VS. ENVELOPE DETECTORS

The interesting and significant thing about a product detector is that there is no output with the b.f.o. turned off. Unlike an envelope detector (diode, grid, plate, etc. detectors), where two or more signals coming in will give a beat or beats, the product detector requires that the b.f.o. (local oscillator) voltage be present. Thus it is very similar to a mixer or converter stage, which also gives no output unless oscillator voltage is applied.

The advantage of the product detector is consequently that the output consists solely of beats with the b.f.o. and not cross-modulating beats between signals.

» Basic principles of low-frequency LC filter design, with practical information on adjustment and the winding of toroidal coils.

Low-Frequency Filter Design for the S.S.B. Transmitter

FRED M. BERRY, WØMNN

In the filter method of generating single sideband, a double sideband is first generated in a balanced modulator (where the carrier is eliminated), and the filter removes the undesired sideband by "brute force." Since the filter is a passive network, sideband suppression is not affected by other circuit variations, tube gains, etc.

Filters using only inductors and capacitors are practical only at frequencies in the order of 10 to 50 kc., and the sideband must be obtained at some low frequency and heterodyned to the desired output frequency. This is not a serious handicap, and enables the output frequency to be varied without disturbing the sideband-generating portion of the circuit. The selection of either upper or lower sideband is simple, requiring only a frequency change of one of the oscillators.

The block diagram of a practical single-sideband transmitter is shown in Fig. 1. The selection of upper or lower sideband is accomplished by switching the frequency of the second oscillator. In the notation of Fig. 1, "USB" and "LSB" indicate the position of sideband at the points noted with respect to the input speech frequencies. This is not to be confused with the particular sideband of the second oscillator that is selected by the second i.f. Careful study of Fig. 1 will make this clear.

Although not directly indicated in Fig. 1, the requirements of the various filters might be briefly reviewed. The first i.f. filter must select a band of frequencies from about 19.8 to 17 kc. and have high attenuation to frequencies of 20.2 to 23 kc. (the other sideband). The second i.f. selectivity must be such that it will greatly attenuate the frequencies of the unwanted sideband generated in the second modulation process. This unwanted sideband will be removed by twice the frequency of the first i.f. (34 to 40 kc. for 17–20 kc. first i.f.). The second-oscillator frequency must also be prevented from being transmitted. A balanced modulator for the second modulator will remove most of this undesired signal but is not to be relied upon for complete elimination; therefore, the second i.f. should also have considerable attenuation at ±20 kc. Coupled tuned circuits of conventional i.f. transformer design are satisfactory at the frequencies chosen.

The third i.f. requirements are similar to those of the second i.f. except that, with the frequencies chosen, the selectivity requirements are more lenient. It would be quite practical to employ a mixer (modulator) not of the balanced type, for the third modulator.

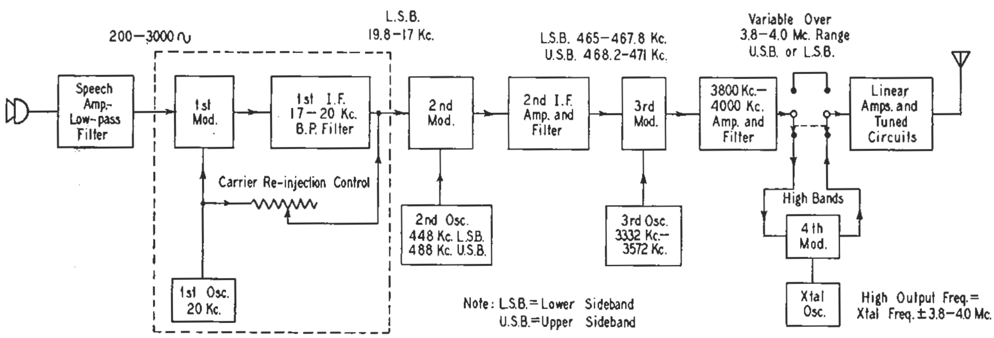


Fig. 1 — Block diagram of a typical single-sideband suppressed-carrier transmitter or exciter. The equipment inside the dashed area is described in this article.

The primary purpose of this article is to describe in detail the construction of a highly-selective sideband filter that can be built at a reasonable cost and with a minimum of special test equipment.

Filter Design

In considering the design of this filter, quality of components and desirable characteristics were of first consideration; low cost and ease of con-

From *QST*, June, 1949.

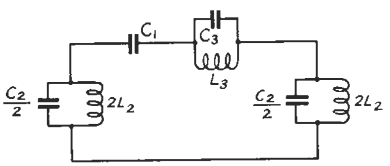
struction were achieved by selection of the type of filter sections and impedance transformations. Sharp cut-off is restricted to the high-frequency edge of the passband, concentrating the attenuation where most needed, and resulting in a minimum number of inductors. This filter is designed for selection of the lower sideband, but since the position of the sideband may be altered at the output of the transmitter in a succeeding modulator stage, this is no handicap.

A figure of approximately 40 db. reduction of the undesired sideband was selected as a practical value. It is believed that values much lower may tend to limit operation on the adjacent channel (when sufficiently selective receivers are in use). Values much over 40 db. would probably not be worth while even if a greater ratio were obtained at the output of the filter, because intermodulation in succeeding stages of the transmitter is likely to introduce spurious beat products of low intensity.

A bandwidth of 2800 to 3000 cycles has proved satisfactory for commercial voice communication and is thought to be a practical one for amateur use.

A frequency band of 17 to 20 kc. was chosen in preference to one of lower frequency to reduce the selectivity requirements of the second i.f. filter, as previously noted. This rather high frequency (for a single-sideband filter) also has the feature of lower component values, lowering cost and making hand winding of the inductors feasible.

The filter consists of two m-derived pi sections of the type shown in Fig 2. This type of section has one frequency of infinite rejection on the high-frequency side of the passband. By using two sections, one with the rejection frequency at 20.5 kc. and the other at 21.5 kc., the resultant attenuation on the high-frequency side is quite



$$C_{1K}=\frac{f_2-f_1}{4\pi f_1 f_2 R}\ ;\ L_{2K}=\frac{(f_2-f_1)R}{4\pi f_1 f_2} \qquad h=\sqrt{\left(1-\frac{f_1^2}{f_{2\infty}^2}\right)\left(1-\frac{f_2^2}{f_{2\infty}^2}\right)}$$

$$C_1=a\,C_{2k}\ ;\ C_2=m_1\,C_{2k}$$

$$C_3=\frac{1-m_1^2}{4h}\cdot C_{2k} \qquad a=\frac{(1-m_1^2)f_{2\infty}^2}{4f_1 f_2}$$

$$L_2=L_{2k}$$

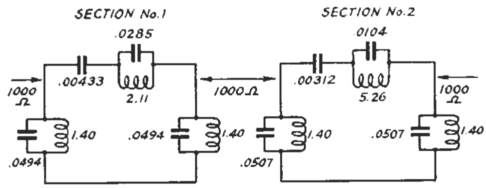
$$L_3=\frac{h}{a}\cdot L_{2k} \qquad m_1=\frac{f_1 f_2}{f_{2\infty}^2}+h$$

Fig. 2 — The basic m-derived pi section used in the filter.

high. When the two sections are combined, the inductors and capacitors at the junction may be combined, to reduce to five the total number of inductors in the complete filter.

The input and output impedance characteristics are the same as that of the midshunt constant-k type of filter of the same cut-off frequency. This sort of termination impedance is most economical and works very well either directly from a ring modulator or resistance terminations.

The design impedance R of 1000 ohms was selected to give desirable component values and



$R = 1000\,\Omega$ $\quad$ $f_1 = 17$ KC. ; $f_2 = 20$ KC. $\quad$ $f_{2\infty} = 20.5$ KC.

$R = 1000\,\Omega$ $\quad$ $f_1 = 17$ KC. ; $f_2 = 20$ KC. $\quad$ $f_{2\infty} = 21.5$ KC.

Fig. 3 — Component values of the individual pi sections of the filter. Values are in μf. and mh.

desirable input and output impedances obtained by transformer action in the end inductors.

The resultant values calculated from the design formulas of Fig. 2 for each filter section are given in Fig. 3, and in Fig. 4 for the two sections combined.

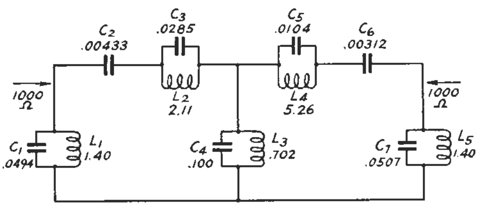


Fig. 4 — Component values of the filter after combination of the parallel components. Values are in μf. and mh.

To those who may wish to calculate similar filters, note that if sections are to be joined, the design impedance and the cut-off frequencies must be the same for both sections, although the frequencies of infinite attenuation may be different.

The filter of Fig. 4 might now be constructed, and if proper components are available, the insertion loss between 1000-ohm resistive impedances would be approximately that of Fig. 5. A low dissipation factor (high Q) is necessary in most of the components to obtain the required characteristics. Resistive losses internal to the filter not only will cause a greater loss at all frequencies but will "round off" the edges and prevent the rapid rise of attenuation needed just outside the passband.

Except in the case of C_1 and C_7 (Fig. 4), mica or other low-loss types of capacitors are necessary for proper filter action. C_1, C_4 and C_7 are large values for mica capacitors and would be expensive. Ordinarily they would have to be made from a large number of paralleled units. C_1 and C_7 appear directly in parallel with the terminating resistances and it has been found that good-quality paper capacitors are satisfactory here.

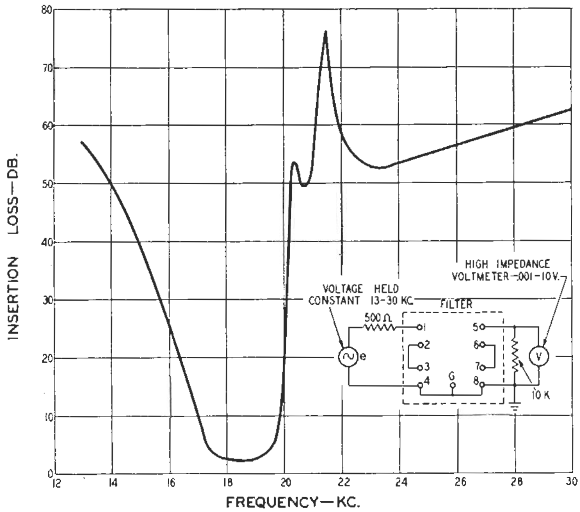


Fig. 5 — Attenuation characteristic of the filter shown in Fig. 4. The test set-up diagram refers to Fig. 6.

C_4 is located internally and must be of low-loss type for best results. It is possible, however, to use impedance transformation at L_3C_4 and permit a smaller mica capacitor to be used for C_4 at the expense of a larger value for L_3. The method of impedance transformation employed also permits a relaxation of the capacitor tolerance. Any reasonable value may be used for C_4 provided the inductor is adjusted to the correct antiresonant frequency. The correct impedance may then be regained by tapping L_3. With inductors of high coupling between turns, the proper point of tapping is such that the inductance between common and tap is approximately that of the value of L_3 before transformation. The modification with the more desirable values is shown in Fig. 6.

Component Tolerances

In the filter of Fig. 4, the tolerance of some of the elements is quite critical, particularly that of the series arms. It has been found in the design of filters of this type that the tolerance of LC ratios is not particularly critical provided the correct resonant and antiresonant frequencies are maintained. Practically, this leads to the selection of capacitors to a tolerance of ±5 per cent or better, and resonating each LC circuit to the correct frequency by turn adjustment of the inductor. The maximum possible error of 10 per cent in the impedance match between junctions of the filter arms is not serious. Greater tolerances will cause a "ripple" in the passband and other deviations from the desired characteristics. In following this procedure, note that the series arms of the filter have both a resonant and an antiresonant frequency, with the inductor as a common element for both. Obviously, the inductor could not be adjusted independently for both frequencies. To permit this desired independent adjustment, a tapped-inductor arrangement is used.

Considering the series arm $C_2L_2C_3$, C_3 is selected with a tolerance such that it will always be larger than the calculated nominal value. L_2

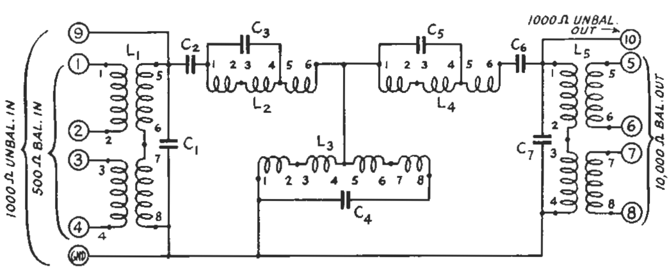


Fig. 6 — Revised filter of Fig. 4, with provision for balanced or unbalanced input and output.

C_1 — 0.049 μf. ±5%, paper.
C_2 — 0.0043 μf. ±5%, mica.
C_3 — 0.03 μf. + tol., mica.
C_4 — 0.03 μf. ±20%, mica.
C_5 — 0.011 μf. + tol., mica.
C_6 — 0.0031 μf. ±5%, mica.
C_7 — 0.051 μf. ±5%, paper.
L_1 (1–4) — 0.7 mh., 33 turns No. 26, bifilar.
L_1 (5–8) — 1.4 mh., approx. 47 turns No. 26, bifilar.
L_2 (1–4) — 2.0 mh., approx. 80 turns No. 26, bifilar.
L_2 (5–6) — 0 to 20 turns No. 26, single.
L_3 (1–4) — 0.7 mh., approx. 47 turns No. 26, bifilar.
L_3 (1–8) — 2.3 mh., approx. 86 turns No. 26, bifilar.
L_4 (1–4) — 5.0 mh., approx. 125 turns No. 26, bifilar.
L_4 (5–6) — 0 to 20 turns No. 26, single.
L_5 (5–8) — 14.0 mh., 160 turns No. 28, bifilar.
L_5 (1–4) — 1.4 mh., approx. 47 turns No. 26, bifilar.
All wire Formvar or d.s.c. — see text.

L_1 and L_5 wound on Western Elec. P476930 core ring.
L_2, L_3 and L_4 on Western Elec. P284395 core ring.

$$\text{Approx. turns P476930} = 1000/\sqrt{\frac{164}{L}}$$

$$\text{Approx. turns P284395} = 1000/\sqrt{\frac{79}{L}}$$

Resonant Frequencies

Inductor	*Capacity*	*Freq., kc.*
L_1 (5–8)	C_1	19.1
L_2 (1–4)	C_3	20.5
L_2 (1–6) with C_3 connected	C_2	19.1
L_3 (1–4)	0.1 μf. ±5%	19.0
L_3 (1–8)	C_4	19.0
L_4 (1–4)	C_5	21.5
L_4 (1–6) with C_5 connected	C_6	18.9
L_5 (1–4)	C_7	18.9

may then be adjusted with this new value of C_3 to the correct antiresonant frequency and will have fewer turns than the original calculated value of L_2. Leaving C_3 connected across the exact number of turns necessary for antiresonance, turns may be added to L_2 until the entire combination of L_2, C_3 in series with C_2 and the added winding of L_2 will series-resonate at the correct frequency. The exact value of C_2 will set the impedance of the entire arm, and $\pm$ 5 per cent is permissible.

The series arm $C_5C_6L_4$ is considered and modified in the same manner.

This now leaves only four capacitors, C_1, C_2, C_5 and C_7, to be selected to plus or minus 5 per cent. Each capacitor may of course be made up of two or more units in parallel if necessary to obtain the correct value.

The filter may be further modified by the addition of separate windings to L_1 and L_5 to permit operation directly from a ring modulator and into the grids of a balanced tube modulator. This adds little additional cost, and accurate balance can be easily obtained by using bifilar windings.

In the design given, an impedance of 500 ohms was selected for the input winding of L_1, since a copper-oxide ring modulator operates quite satisfactorily into this impedance. The impedance of the output winding of L_5 is a compromise between desired voltage step-up and keeping the number of turns to a value that permits hand winding. The completed filter design after all modifications is shown in Fig. 6. In the event other input or output impedances are desired, the number of turns and method of connection of these added windings may be altered to meet the requirements. Since the impedance varies directly with the inductance of the windings (with 1.4 mh. the inductance for 1000 ohms impedance), the required inductance in millihenrys for any new impedances may be found by dividing the new impedance in ohms by 1000 and multiplying the result by 1.4. The number of turns required can be found from the formulas for the inductors given in Fig. 6.

Filter Alignment

As has previously been mentioned, the *LC* combinations must be resonant at the desired frequency. In an *m*-type filter with closely-spaced rejection frequencies, it is very important to hold to very close frequency tolerances; while a constant error is not serious the spacing of one frequency to the next is critical.

Heretofore, it has been considered necessary to use expensive laboratory equipment, which is out of the reach of many. Signal generators for the range of 15 to 30 kc. are not common, and those available are usually not of sufficient accuracy. However, with the aid of a BC-221 frequency meter the main obstacle has been removed. The fundamental frequency range of the low band of the BC-221 is 125–250 kc., and it has sufficient output voltage to give a reasonable indication on most oscilloscopes. The BC-221 is used only as a calibration means for the test signal generator. The test generator may easily be made from the junk box, and the usual calibration problem solved by the BC-221. In fact, good procedure is to use only a rough calibration and use the BC-221 continuously for frequency set. The method of connection of the frequency-check system is shown in Fig. 7. The oscilloscope vertical and horizontal inputs are used to give the familiar Lissajous figures as a means of comparing frequencies. Since most of the frequencies needed from the test generator are one-tenth that of the BC-221 it is convenient to use the chart calibration points for 125–250 kc. By moving the decimal point one place to the left and obtaining a 10:1 Lissajous pattern on the oscilloscope the frequency may be read directly. Other multiples must, of course, be used for some frequencies.

An *LC*-type signal generator is recommended for best stability, and particularly if one has to be constructed.

A method of resonating that gives accurate results is shown in Fig. 7. This method measures all *LC* combinations as a series-resonant circuit. With the *LC* combination connected as shown, a sharp dip in amplitude occurs when the frequency is at the exact series-resonant point, since the impedance is then a minimum. Although not critical, *R* of Fig. 7 should be the smallest value that will still give a readable indication. When an entire series arm is resonated the dip will not be as great but will be very sharp.

Coil Construction

In selecting inductors for the filter, the *Q* is of primary importance. *Q* values of at least 150 are necessary for all inductors except possibly L_1 and L_5. L_1 and L_5, as in the case of C_1 and C_7, are in parallel with the terminations, and losses here

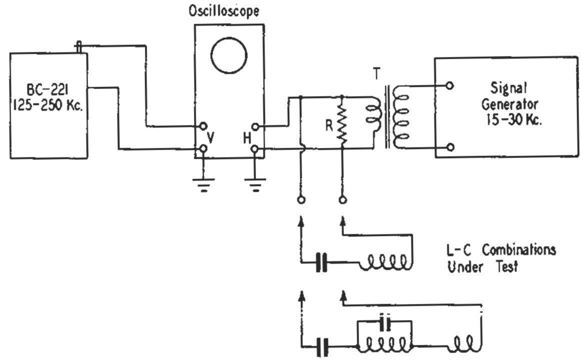


Fig. 7 — The method used for checking coil-and-condenser combinations. An accurate frequency check is obtained by using a BC-221 to check the 10th harmonic of the test signal generator. The *LC* combination under test is adjusted for minimum horizontal amplitude at the desired frequency.

R — 1 to 10 ohms, ½ watt. See text.
T — Step-down transformer. A 500-to-6-ohm audio transformer is suitable with generator outputs of 500 ohms or less.

have much less effect. While many types of inductors might be used, the toroidal type has many advantages and core rings of molybdenum Permalloy are now available to the amateur. Toroidal coils of this material are small in size and have a very low external field, and the inductance remains quite constant with power level and temperature. The coupling between turns is high, so that leakage reactance may be neglected in the design of the built-in transformers and tapped coils. The one disadvantage of using toroidal coils is the difficulty of winding, since the wire must be threaded through the core. However, in this filter special attention was given to keeping the inductances low, and winding is not too difficult. Two grades of core material were used in the inductors for the filter of Fig. 6 (attenuation characteristics shown in Fig. 5).

Inductors L_2, L_3 and L_4 use cores having an effective permeability of 60, producing Qs of 200 to 250 at 20 kc. L_1 and L_5 cores were of 125 permeability, reducing the required number of turns and still permitting Qs of over 100. The construction data in Fig. 6 give the approximate number of turns of the inductors when using Western Electric core rings P476930 for L_1 and L_5, and P284395 for L_2, L_3 and L_4. P476930 and P284395 have nominal inductances of 164 and 79 millihenrys respectively per 1000 turns. The approximate number of turns for a specified inductance, as given by the manufacturer, is found by the formulas given in Fig. 6.

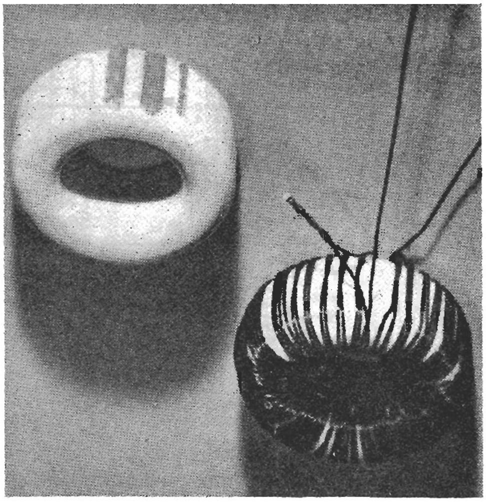
Since a tolerance is allowed on the capacitors, and the permeability of the cores varies slightly, the exact number of turns will vary and must be determined by measurement. For this reason sufficient length of wire should be allowed for the windings so that the additional number of turns necessary may be found by test. The extra length of leads will not affect the test, and these leads may later be cut to proper length after the correct number of turns has been determined.

Wire size is not critical and deviation from that given in Fig. 6 may be made if winding area does not limit. "Formvar" insulation, or the equivalent, is recommended and is easy to wind, but single-silk or nylon is satisfactory. Plain enamel not of the Formvar type is to be avoided, because of the possibility of shorted turns.

In order to reduce the number of times the wire must be threaded through the core ring, all windings are bifilar except the adjustment windings *5–6* of L_2 and L_4. In the bifilar type of winding, two wires are held together and wound as one. After winding, the start of one wire (*3*) may be connected to the finish of the other (*2*), thus connecting the two in the series-aiding manner. As in telephone practice, the numbering of windings connected for series aiding is such that the direction of current at one instant is from *1* to *2*, *3* to *4*, *5* to *6*, etc. Thus if *2* and *3* and *4* and *5* are connected together and external connections made to *1* and *6*, the windings are series-aiding.

In winding, the length of wire to be pulled through for each turn may be halved by starting at the center of the bifilar (doubled) length of wire in the core ring, and then winding in opposite directions through the core ring respectively with each end of the bifilar wire. The wire should be evenly distributed around the core ring, but this is not particularly critical.

The following procedure for proper identification and labeling of bifilar windings may be used:



A finished toroid coil of the type used in the sideband filter. An unwound core is shown at the left.

Select one of the ends of the completed winding and arbitrarily label them 1 and 3. Now by use of an ohmmeter, locate the wire at the opposite end of the winding which checks continuity to "1." This of course will be "2" and may be spliced to "3." With the exception of the input and output winding of L_1 and L_5, the free wires, 1 and 4, may be left long and any additional turns necessary may be obtained by winding on singly, with care that the wire continues through the core in the same direction.

The bifilar windings *5–6* and *7–8* follow the same procedure. However, when two windings are to be series-connected such as *1–2*, *3–4* and *5–6*, *7–8*, care must be taken in selection of the end of winding to label *5*, *7*. The proper labeling is such that the wire ends *5*, *7* pass through the core center in the same direction as the wire ends *1*, *3*.

The input windings of L_1 (*1–4*) and output windings of L_5 (*5–8*) are not critical in inductance and may be wound first to the specified number of turns. If desired, a layer of tape may be applied over these windings before application of the second windings.

L_3 (*1–4*) is wound and resonated with a 0.1-μf. test capacitor to 19.0 kc. Adjust to the nearest turn that produces resonance closest to the exact frequency. C_1 and C_7 may be paralleled and used temporarily for the test capacitor. The second winding of L_3 (*5–8*) is now applied and series-connected with the inner winding, *1–4*. Turns are adjusted to secure resonance with C_4 at 19 kc. No connection is made to the tap during adjustment.

Note that wide tolerances on C_4 are allowed and the exact number of turns of L_3 will depend on this tolerance.

L_1 (*5–8*) and L_5 (*1–4*) are wound and resonated with their associated capacitors, C_1 and C_7.

L_2 (*1–4*) is now wound and resonated with C_3. As previously mentioned, the value of C_3 may vary over a wide range (plus tolerance), and will determine the number of turns of windings *1–4*. Note that the total number of turns for L_2, including adjustment winding *5–6* depends only on the exact value of C_2. Thus if C_3 is large, winding *1–4* will have fewer turns, and *5–6* will have more. After resonating C_3 connect it in parallel with windings *1–4* and the combination in series with C_2. Check the resonant frequency of the entire series arm, less winding *5–6*. It should be higher than the frequency as given in Fig. 6. Now, wind turns on for the trimming winding *5–6*, and, with it connected, recheck the resonant frequency. Adjust turns of *5–6* until correct frequency is obtained. In the event that the resonant frequency was lower than the correct value before the addition of *5–6* it is an indication that C_3 was too low, and the entire adjust-and-check procedure must be repeated with a larger value for C_3.

L_4 is now wound and resonated with C_5 and C_6 following the same procedure as for L_2, C_3, C_2.

The filter may now be wired temporarily for test before mounting. The method of connection for test is shown in Fig. 5. While a sensitive voltmeter or decibel meter of high impedance is necessary for accurate measurement, an oscilloscope may be used instead for an approximation. If the filter is flat through the passband and attenuates rapidly on the high side it is likely that no errors have been made. If the oscilloscope gain is adjusted for full deflection at the center of the passband, the deflection at points above about 20.4 kc. should be barely visible if at all.

The mounting of the components will be left to the builder, but it is to be noted that the inductors may be mounted very close together and near metal surfaces without harmful effect, with the possible exception that L_1 and L_5 should be given some separation from one another. A metal screw may be used through the center of an inductor without harm provided it does not constitute a shorted turn, as it would if metal washers were used on both sides and the washers connected together.

A suggested schematic using the filter is shown in Fig. 8. The speech amplifier should feature low- and high-frequency cut-off as with any 'phone transmitter. Some high-frequency attenuation may be obtained by the action of the secondary of T_1 with C_1, C_2. It is well first to run a frequency-response check on the speech amplifier including T_1, C_1, C_2 with the modulator disconnected and a 500-ohm resistor substituted.

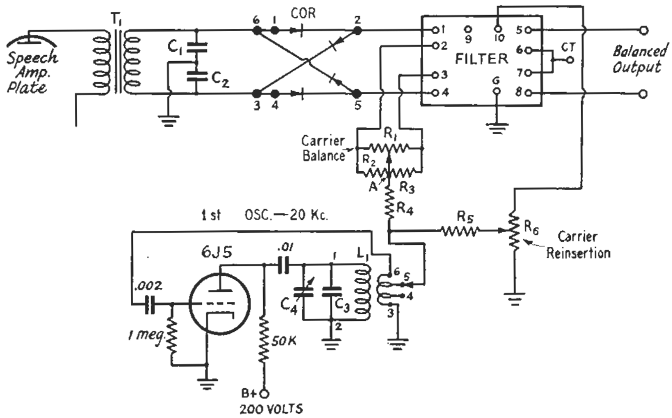


Fig. 8 — **A suggested circuit for the 20-kc. oscillator and balanced modulator to be used with the filter.**

C_1, C_2 — 0.05-µf. ±20%, matched to within 1% by trial.
C_3 — 0.01-µf. silver mica.
C_4 — 200-µµf. variable or adjustable.
R_1 — 5000-ohm potentiometer.
R_2, R_3 — 150 ohms, ½ watt.
R_4 — See text.
R_5 — 3000 to 5000 ohms.
R_6 — 1000-ohm potentiometer.
L_1 — 284 turns or 142 bifilar, No. 26 Formvar or s.s.c., for coil 1–2. Coil 3–6 is 40 turns of same tapped every 10 turns. Both coils are wound on the same W.E. P284395 core ring.
COR — Copper-oxide modulator (Varistor). See text.
T_1 — Output transformer, plate to 500 ohms.

The 20-kc. oscillator shown uses a toroidal inductor. Other types of oscillators will perform satisfactorily if the output impedance is held low. The number of turns of inductor L_1 and value of C_3 may be adjusted for proper frequency using the BC-221 and the proper feed-back adjusted by the secondary winding *3–4–5–6*. The taps on this winding are desirable to adjust the voltage at the junction of R_2, R_3 from 2 to 5 volts. Selection of 100 to 500 ohms for R_4 also permits some adjustment. R_5 should be as high as possible for least loading on the oscillator and filter, still permitting enough 20-kc. output for any desired amount of carrier reinjection.

One point not obvious is that R_2 and R_3 with R_1 in parallel are actually in series with the input to the filter. The values chosen normally give a good impedance match between the modulator and filter. If 1N34s or vacuum tubes are used instead of copper oxide for the modulator, a resistor may have to be placed across filter terminals *1* and *4*, and R_2 and R_3 lowered in value. Proper match may be noted when audio is fed into the speech amplifier and varied from 200 to 3000 cycles. If the speech amplifier has previously checked flat, the output from the filter at terminals *5* and *8* as measured with a voltmeter or noted by the oscilloscope should vary as the response through the filter alone with frequencies of 19.8 to 17.0 kc. A ripple in output amplitude indicates incorrect modulator match.

Copper-oxide "Varistors" that have proven satisfactory with the values given are Western Electric D162258, D163139 and D98914. The terminal numbering given for "COR" of Fig. 8 is for these types.

Modulator balance for maximum carrier reduction is normally quite simple. A sensitive voltmeter or oscilloscope is connected to terminals *5* and *6*. With no input to the speech amplifier and R_6 tuned for minimum carrier, adjust the carrier balance control R_1 for minimum output.

Balance should be obtained near the center of the adjustment range. If not, a trimming resistor may be paralleled with R_2 or R_3. Some capacity unbalance in the Varistor or input winding of the filter may prevent sufficient carrier balance and small values of capacity may be added from filter terminal *1* or *4* to ground. Capacity may also be tried across C_1 or C_2.

Note that any hum in the speech amplifier will appear as an output carrier, but of course will be 60 or 120 cycles from the true carrier. Hum may be identified by temporarily shorting the primary of T_1.

Audio is now connected to the input of the speech amplifier and the level adjusted to a maximum of 0.25 volt at the output of T_1.

If the output of the speech amplifier is a pure tone the output of the filter should be a single frequency of 20 kc. minus the audio frequency. Using a sweep rate that is a submultiple of the audio input frequency, a check may be made for the presence of a modulation envelope. Such a trace represents more than one frequency in the output and may be caused by distortion in the speech amplifier or overloading of the modulator. A slight modulation pattern is permissible as this represents only a slight distortion of speech and not spurious signals out of the passband.

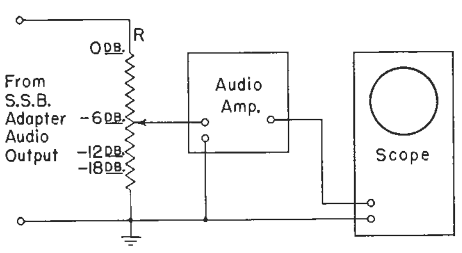
The modulator is now ready to be connected to the succeeding stages of the exciter.

MEASURING SIDEBAND SUPPRESSION

Howard Wright, W1PNB, suggests this simple stunt for measuring sideband suppression of your own or the other fellow's signal. It requires that you have a selectable-sideband receiver of some kind (filter, YRS-1, etc.) and an oscilloscope, and the only other requirement is a calibrated volume control.

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

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



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

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

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

» *How television receiver components can be used in a low-frequency filter.*

An Inexpensive Sideband Filter

DAVID O. MANN, W6HLY

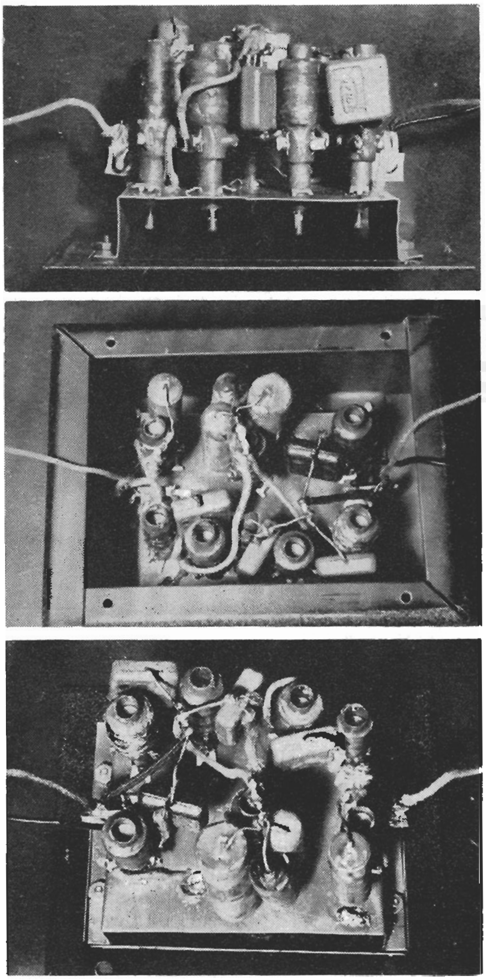
Once constructed, an exciter of the filter type has some distinct advantages over other systems. As an example, once the filter is constructed, intelligent use of a standard communications receiver (something most amateurs have) is all the test equipment required to tune up the exciter. With the addition of a potentiometer, any desired amount of carrier can be inserted, and the resultant a.m. (A3 minus one sideband) signal used to establish contacts before switching to single sideband.

This article is primarily intended to describe a bandpass filter suitable for use in the circuit and within the reach of even the relatively inexperienced amateur.

The Filter

The filter details to be given are the outgrowth of difficulties in constructing the first exciter. The first filter design was successful, but a good filter wasn't built until after about three tries at getting suitable iron for the inductances. The toroid is generally considered the best physical configuration for high-*Q* inductances, but if a filter is to be built by amateurs with a minimum of investment and test equipment, a compromise has to be made between performance and ease of construction. To make this clear, the toroid cores for the inductances were cut from the core of a television horizontal-sweep transformer by boring and slicing a cylindrical section of it. This material is called "sponge iron" and was tried because it was designed for use at 15 kc. It made some pretty good coils, but such construction isn't at all inviting and of course would be extremely difficult to describe to anyone else. One of the inductance values required in this first filter was 21 mh., much larger than the others for which the toroids were used. To make up this value using the toroid core would have required an enormous number of turns, and this started a search of all available coils having this larger inductance and a reasonable *Q*. An RCA standard television variable inductance, used in the horizontal-sweep circuit, was found to have a range of inductance between 5 and 21 mh., and a *Q* at 10 kc. of from 10 to 35. This served the purpose at the time, but it appeared to be such a handy component that various filter designs were computed, in an effort to find one in which it could be used throughout. Two revisions resulted in the one described here. These little coils are called Horizontal Linearity Coils, RCA part No. 201R3, and they can be obtained for less than a dollar. They contain about 1300 turns of No. 33 enamel wire, and to use them for the lower values of inductance in the filter it is necessary that they be pruned, to permit keeping the slug well in the winding and thus maintain a higher *Q*.

Another variable in the filter design is the image impedance, and the value of 200 ohms was



Three views of the homemade single-sideband filter. The top view shows how the components are mounted on a metal channel that is fastened to one side of the shield box. The other two views show the unit in and out of the box.

From *QST*, March, 1949.

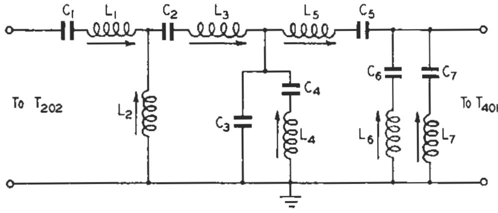


Fig. 1 — Circuit diagram of the single-sideband filter.
C_1–C_7 — Small mica, molded-paper or paper condensers — not electrolytic — combined as below.
L_1–L_7 — RCA Horizontal Linearity Coils (RCA No. 201R3) modified as described below.

Element	Make-up	Tune to
C_1 L_1 7.5 mh.	C_1–.02-µfd. paper L_1–Remove 400 turns	13 Kc.
L_2 2.9 mh.	L_2–Remove 650 turns	Tune with 0.1-µfd. condenser in series to 9.4 Kc.
C_2 L_3 18.8 mh.	C_2–.01-µfd. paper or mica L_3–No turns removed	11.6 Kc.
C_4 C_3 L_4 13 mh.	C_3–.11 µfd. (.1 and .01 in parallel) C_4–.02 µfd. L_4–Remove 200 turns (Save wire for L_6)	9.8 Kc.
L_5 C_5 17 mh.	C_5–.012 µfd. (.01 and .002 in parallel) L_5–No turns removed	10.9 Kc.
C_6 L_6 27 mh.	C_6–.01 µfd. L_6–Add 175 turns (wire from L_4) in same direction	9.6 Kc.
C_7 L_7 19 mh.	C_7–.007 µfd. (.005 and .002 in parallel) L_7–No turns removed	13.6 Kc.

selected because audio line transformers for this impedance are very reasonable and, because of unity transformation ratio, they will pass 10 to 13 kc. This is important, because audio components for this frequency range are usually tagged "Hi-Fi" and priced accordingly. Fortunately they are not required in this case. The transformers listed in the diagram are quite reasonable and have been very satisfactory.

Construction of the Filter

The photographs show three views of the filter in a standard 3 × 4 × 5-inch box, as one method of assembly. It is almost obvious that no attempt was made to give the job a commercial look, but it does indicate the relative size to be expected when completed. Fig. 1 is a complete diagram of the filter and, though it may look complicated, it really isn't any worse than some of the clipper filters in common use today. The frequency range is not an old stamping ground, but the same techniques are still good and the measuring methods used shouldn't scare anyone away from tackling the job. In addition to the filter schematic, Fig. 1 contains a table dividing the filter into seven elements and illustrating the recommended method of making up the odd values of capacitance from standard condenser units. The center column of this table gives detailed information for altering the standard coils so that the required inductance can be set with the slug. Reasonably uniform results can be expected of a standard procedure, since several of these coils were measured and the individual variation was quite small.

The condensers are small enough so that the lugs of the coils can be used for tie points. It is suggested that each element be made up as shown and all mounted, leaving pigtails long enough to permit tune-up and interconnection. Before interconnection, each element is then tuned to series resonance at the frequency shown in the right-hand column of Fig. 1, using a test set-up as shown in Fig. 2. Since the values of standard condensers vary, setting the inductances compensates for the error by making the elements have the same resonant frequencies they would have if all components had exact design values. This also helps to compensate for any fixed error in the calibration of the oscillator used to tune the elements. To guard against errors, it is suggested that the best available type of condensers be used; i.e., a mica condenser can usually be expected to be more closely watched than a tubular paper during manufacture, but the paper condensers should be satisfactory for the larger sizes. Referring to the photographs, the elements were mounted on the "U"-shaped bracket by forcing the heads of the coil forms through the proper-sized hole (approx. ¼ inch). The elements were then tuned and, after bolting to the side of the box, the slug screw adjustments are protected from accidental change during wiring and final insertion in the box. The open face of the box can then be secured down on the chassis to enclose the filter and protect the components.

As a word of encouragement, there is no need for extreme accuracy in either the number of turns pruned from the coils or the calibration of the audio oscillator used to line up the elements. The specified turns to be removed includes a fair margin of safety, and if the combination doesn't tune to the given frequency the condenser value is probably too far off and another should be tried. As mentioned above, should the marked frequencies on the oscillator be off, the

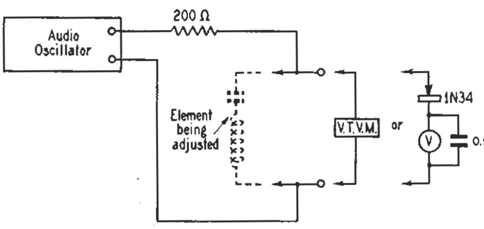


Fig. 2 — The test circuit required to adjust the elements of the sideband filter. Two types of indicators are shown. The inductance is adjusted for minimum indication by the meter at the test frequency (Fig. 1).

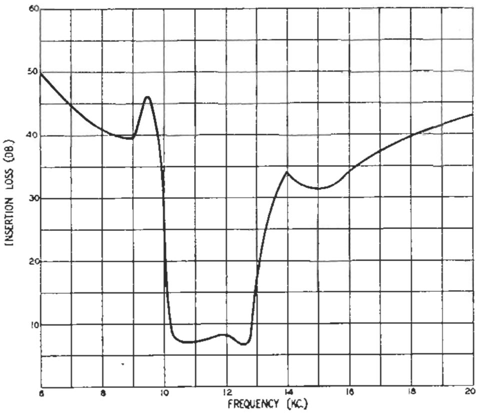


Fig. 3 — The frequency characteristic of the home-made sideband filter.

eventual operation of the filter will not be impaired, provided the *same* instrument is used to tune all elements.

The matter of obtaining use of the necessary test equipment may seem troublesome, but since it is at worst a ten-minute job to complete the tuning (assuming the elements are made up and mounted), it does not appear the least brazen to request this favor of your parts supplier should other sources fail. Audio oscillators and v.t.v.m.s are rather common test equipment in the present-day laboratory, and a reasonable will to get the job done should be all that is necessary to get over this barrier.

The finished filter can be expected to have an insertion loss roughly as shown in Fig. 3 which, though not ideal, will produce easily-copied good-quality single-sideband signals.

It should be emphasized that the filter is not symmetrical, and therefore care should be taken to see that the connections in the exciter are as indicated in Fig. 1. This means that the right end of the filter will be terminated in approximately 200 ohms as required (see Fig. 4).

Alignment Procedure

By varying R_C (Fig. 4), any desired amount of carrier can be by-passed around the balanced modulator to T_{401}, and being able to do this offers two very important advantages. The first is that the carrier can be used to tune up the exciter and any following amplifiers. The second is that having the carrier-reinsertion control during operation permits adding enough carrier to permit the signal to be copied just like any conventional a.m. station, or it can be operated with a 20-db. suppressed carrier, or no carrier. Operating experience has revealed a great deal of controversy concerning the value of a "pilot" carrier suppressed only 20 db., in view of the usual blanket of other carriers, but this arrangement is versatile enough for most requirements.

Barring unusual difficulties, any receiver with an S-meter is the only essential to tuning up the whole circuit, but it should be appreciated that a receiver is a very sensitive instrument and that the r.f. gain should be kept as low as possible, to reduce the chance of false readings. A test probe will be found convenient and can be made by connecting the receiver antenna input lead to the test points through a 50-$\mu\mu$f. or smaller condenser to an alligator clip.

If you are fortunate enough to have good ears, you can probably hear the 10-kc. oscillator in the vicinity of T_{203}, but if not you can test the oscillator later in another way. Connect the probe to either plate of the 550-kc. balanced modulator, and check for and maximize the sig-

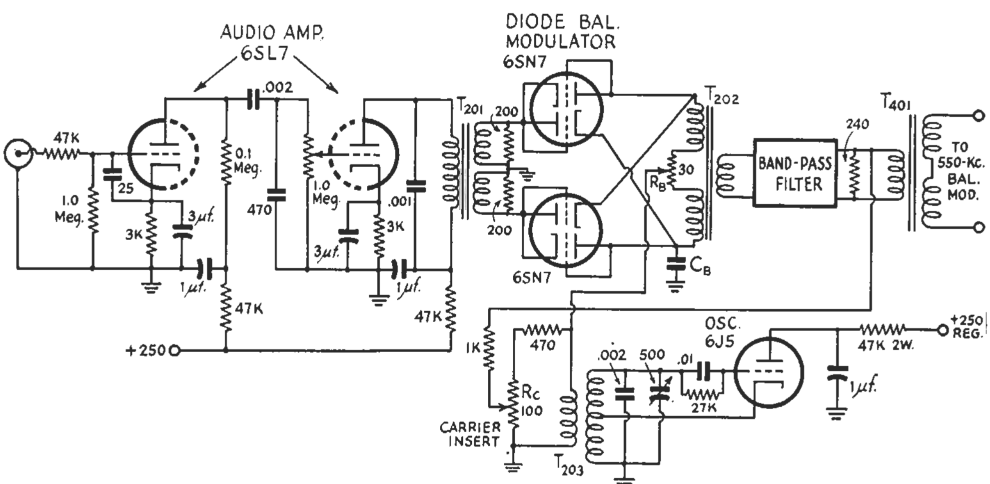


Fig. 4 — Wiring diagram of the speech amplifier, oscillator and first balanced modulator.

C_B — Balancing condenser, approx. 0.004 μf. See text.
R_B — 30-ohm wire-wound potentiometer, for balancing carrier.
R_C — 100-ohm w.w. potentiometer, for carrier insertion.
T_{201} — Single-plate-to-200-ohm-balanced-line (SNC 1P152 or Thordarson T55A15).
T_{202}, T_{401} — Balanced-line-to-line, 200 ohms (SNC 1P161).
T_{203} — Push-pull output transformer (Thordarson T22S86) with iron core removed and replaced by wood for mounting.

nal from the 540-kc. crystal-oscillator stage. A little exploring 10 kc. above or below this frequency may reveal two signals, which should be present if the 10-kc. oscillator is working. If present, tune to one of the side signals and see if its intensity can be varied by the carrier-insertion control, R_C. If so, all is fine so far, since the receiver is tuned to one of the sidebands produced by modulating the 540-kc. carrier with the 10-kc. oscillator signal. Let's assume we want to tune to the upper sideband coming from this modulator, so set the receiver to about 550 kc. (the signal just above the 540-kc. oscillator that can be controlled by R_C), shift the probe to the secondary of the 550-kc. balanced-modulator output circuit, and tune the trimmers for maximum signal. Vary R_C again just to be sure you have the *sideband* and not the 540-kc. oscillator signal.

The proper frequency for the 10-kc. oscillator, with respect to the filter characteristic, can be set with this arrangement. Watching the S-meter, set R_C for minimum indication. This now means that the only path for the 10-kc. oscillator signal is through the unbalance in the ring modulator, T_{202}, and the filter. Run up the sensitivity of the receiver until a reading of about S5 is obtained; R_B in the ring modulator can be varied to increase this reading if necessary. Start with maximum capacity in the 10-kc. oscillator tank circuit and increase the frequency (decrease capacity) until the receiver indicator shows a fairly rapid rise to a maximum. If no pronounced rise is noted (3–4 S units) by the time the 500-$\mu\mu$f. compression condenser is all out, it may be necessary to reduce the fixed capacity in the tank from 0.002 to 0.0015 μf. and try again. When the response levels off to a maximum, the oscillator is up in the passband of the filter and thus has been "located." To get the oscillator on the proper part of the filter curve it is merely necessary to decrease the frequency (increase capacity) until the response drops two S units.

When R_B is set for a minimum we don't want any of the 10-kc. carrier sneaking through, and to prevent this the ring modulator has to be balanced. This isn't difficult, and even rough balance will produce a carrier attenuation of around 30 db. To do the balancing, set the receiver for a good indication, S5 or so, with the probe still connected to the secondary of the 550-kc. balanced modulator. Check again to see that R_B controls the signal and that it is set for a minimum (center point ground). Take C_B, 0.004 μf., and connect to the side of the primary of T_{202} that produces the smaller S-meter reading. If this reduces the signal so low that the receiver sensitivity will not bring it back on scale, leave further balancing until more of the transmitter is tuned up and a higher level is obtained. With all this gain on the signal, when R_C is tuned to a minimum the carrier leakage at the ring modulator can easily be detected, so vary R_B and see if a sharp null can be obtained. If the null is broad, it will be necessary to try a little different value of C_B until the minimum attained by varying R_B is sharp. When good balance is obtained, a hum will be audible in the receiver; i.e., the carrier is so weak that the heater-cathode leakages cause appreciable modulation. Ten kc. above or below the carrier frequency (which side depends whether a 14.7- or 13.7-Mc. carrier is used in the last balanced modulator) another fairly strong signal will be found which is *not* controlled by R_C. This is an undesired output that comes from the 540-kc. oscillator and must be eliminated by balancing the 550-kc. modulator.

The two-stage speech amplifier is simple enough so that little trouble should be encountered. An r.f. filter has been added to the input circuit and some condensers inserted to make the response fall off above 4000 cycles. The 10–13 kc. bandpass filter will trim the radiated sideband down to an effective 300 to 3000 cycles, but the response of the speech amplifier to frequencies above 17,000 cycles must be well down to prevent their modulating any second harmonic of the 10-kc. oscillator and producing spurious sideband frequencies within the passband of the filter.

Balanced I.F. Transformers

To simplify the job of the balanced modulator at 550 kc., a balanced-coil assembly can be built as in Fig. 5. The "doubled" winding should be connected to the push-pull portion of the balanced-modulator circuit and, if the assembly has been made carefully, this resultant center-tapped coil will show equal capacities from its ends (A and B) to ground. The inside turn of the other coil should be connected to the "cold" side of its circuit.

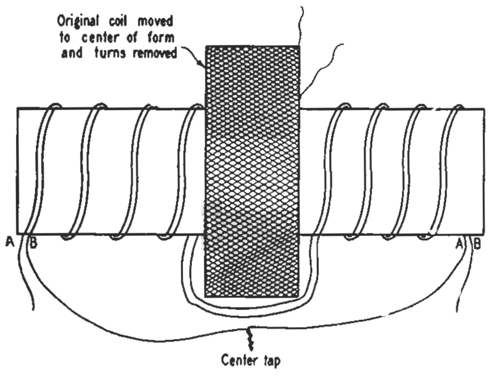


Fig. 5 — Modification of 450-kc. slug-tuned i.f. transformer for good inherent balance. One coil was unwound, and the other moved to the center of the form. About 40 per cent of the turns was then removed from this coil. The wire from the first coil was doubled and used for the new winding, starting at the center so that all of the wire would be used. After doping the windings and connecting them as shown, the two iron slugs were stuck in the center of the form with wax. The assembly was then replaced in the i.f. can.

» Simplified crystal-filter design is the basis for the well-known and widely-used "Edmunds" exciter.

A Crystal-Filter S.S.B. Exciter

F. E. EDMUNDS, W1JEO

This exciter uses a quartz crystal filter operating at 450 kc. (or vicinity). The filter allows a passband of 300 to 3000 cycles; the sideband rejection, as measured with a YRS-1, indicates 35–40 db. over 300 to 3000 cycles. At no time within the reject range is rejection less than 30 db.; at some places it approaches 60 db. Suppression of the carrier is obtained *without* the use of balanced modulators. Crystals suitable for use in the filter are available for less than one dollar each. The most useful of these crystals are in the series that runs from 375 to 525 kc. in 1.388-kc. steps; this series is marked at 72 times the crystal frequency in a series of channels from 28.0 to 38.0 Mc. The crystals were manufactured by Western Electric for the Signal Corps, and are of the plated variety, mounted in an FT-241A holder. The holder pins have ½-inch spacing. The crystals may be socket-mounted or soldered directly into the filter at the builder's discretion.

The filter is of bridge design with complex entry and terminating sections. The complex sections are used to suppress the carrier and modify the response characteristics of the bridge. Fig. 1 shows the filter proper, set for rejection of the upper sideband. The transformer, T_1, is a replacement-type 455-kc. interstage i.f. transformer, mica-tuned, and air-cored. T_2 is also a replacement type, designed to feed into a diode detector.

The original filter was designed to operate at a carrier frequency of 450 kc., although the filter has since been tested at frequencies between 425 and 490 kc. without alteration of the circuit or transformers. Under the condition of design for 450-kc. carrier, crystal "B" is 2.78 kc. *higher* than 450 kc., or 2 channels higher in the crystal series previously described. Crystal "C" is 1.39 kc. *lower* than 450 kc., or 1 channel lower. Crystal "D" is 450 kc. Crystal "A," also at 450 kc., is used in a crystal oscillator to generate the initial carrier. Channel markings on these crystals are as follows:

"A" — 32.4 Mc., Channel 324
"B" — 32.6 Mc., Channel 326
"C" — 32.3 Mc., Channel 323
"D" — 32.4 Mc., Channel 324

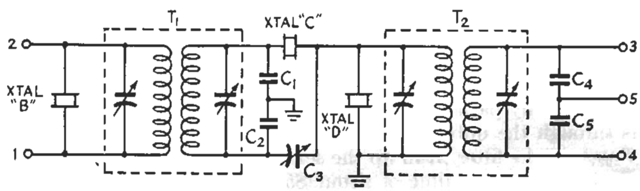


Fig. 1 — The 450-kc. crystal filter for sideband and carrier rejection.

C_1, C_2, C_4, C_5 — 100-μμf. mica or ceramic.
C_3 — 3- to 30-μμf. ceramic trimmer.
T_1 — 455-kc. interstage i.f. transformer (Meissner 16-6659).
T_2 — 455-kc. diode i.f. transformer (Meissner 16-6660).
For a carrier frequency of 450 kc., the crystals are:

Crystal	*B*	*C*	*D*
High-freq. reject	452.8 kc.	448.6 kc.	450.0 kc.
Low-freq. reject	447.2 kc.	451.4 kc.	450.0 kc.

Any other group within the range of the i.f. transformers may be utilized; only the channel relationship need be retained.

Response characteristics of the filter are as indicated in Fig. 2. The graph shows the high sideband being rejected; the lower sideband may be rejected in a similar manner with appropriate crystals, as suggested in Fig. 1.

Fig. 3 shows the entire exciter in block diagram. The 6K8 hexode-triode serves as 450-kc. oscillator and audio mixer. Approximately 3 volts of audio is required at the signal grid of the 6K8 for opti-

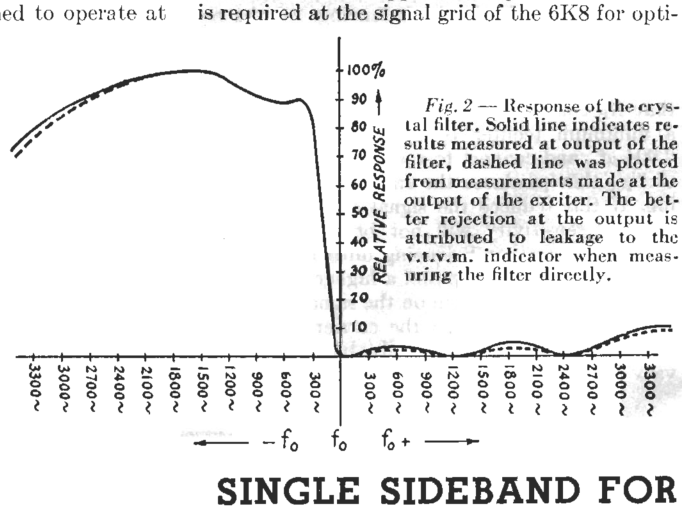


Fig. 2 — Response of the crystal filter. Solid line indicates results measured at output of the filter, dashed line was plotted from measurements made at the output of the exciter. The better rejection at the output is attributed to leakage to the v.t.v.m. indicator when measuring the filter directly.

From *QST*, November, 1950.

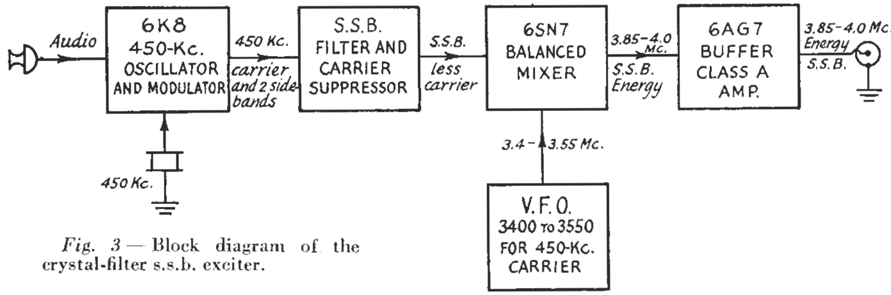


Fig. 3 — Block diagram of the crystal-filter s.s.b. exciter.

mum results. The 6K8 delivers a carrier (450 kc.) and sidebands to the input of the filter. The filter rejects one sideband (depending upon the selection of crystals) and delivers single-sideband energy to the 6SN7 mixer. The filter also suppresses the carrier some 60 db. below the peak sideband energy. The 6SN7 mixer combines the single-sideband energy (in the vicinity of 450 kc.) with the output of the VFO (3400 to 3550 kc.) and the sum products are recovered in the output (3850 to 4000 kc.). The balanced mixer is used to remove the VFO component from the output tank. Balance is not critical and no adjustments are required or provided. A VFO signal of about 6 to 8 volts is required. The output of the mixer is fed to the grid of a 6AG7 which runs as a Class A tuned r.f. amplifier. The output of the 6AG7 is sufficient to drive a pair of 807s Class AB_2. Most VFOs in use cover or may be easily made to cover 3400 to 3550 kc. A single untuned 6SJ7 or 6AC7 Class A amplifier following a BC-221 can be used at the home station with completely satisfactory results as a driver for this exciter.

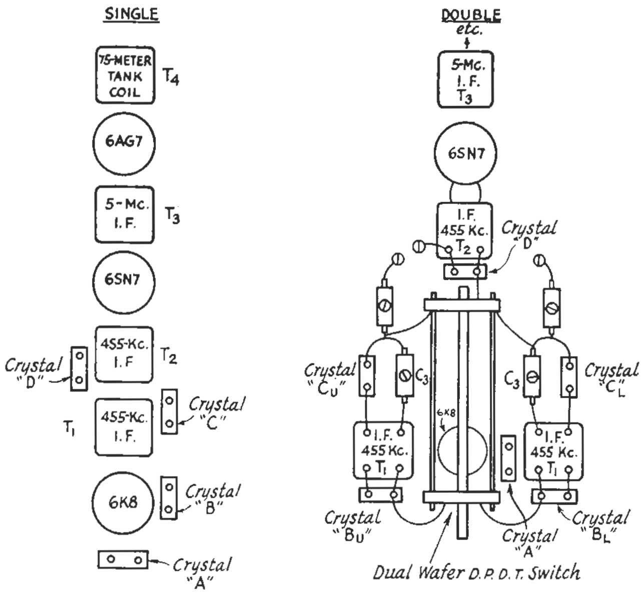
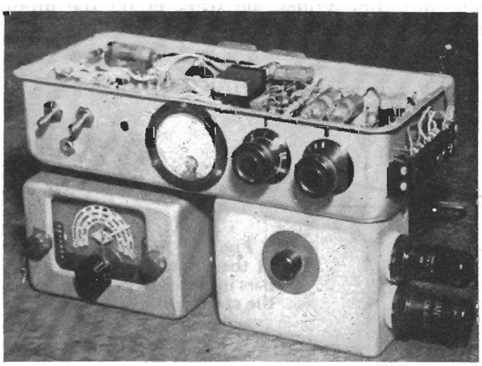


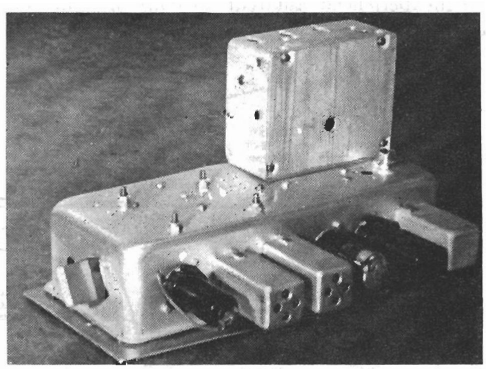
Fig. 4 — Recommended physical layout of the single- and double-channel filters. The double-channel filter is required where instantaneous selection of upper or lower sideband is desired.

Construction

The original transmitter was built for mobile operation and much hole drilling and experimentation has occurred on the chassis. Fig. 4 shows a suggested physical layout for home use.



The dashboard-mounting s.s.b. exciter, complete with receiver converter and VFO. The top dish is the exciter (with cover removed), showing some of the "innards." The meter reads cathode current to a pair of 807s driven by the unit, and the two knobs handle carrier reinsertion and 6AG7 plate tuning.



Another view of the exciter, with the converter removed. The oscillator crystal is mounted on the left-hand end, the tubes and transformers along the side are 6K8, T_1, T_2, 6SN7, 6AG7 and output tank. The Jones output jack can be seen above the output tank shield can, in front of the VFO unit.

This plan will keep stray capacity coupling at a minimum. No shielding other than that provided by the i.f. cans and the output tank can is required. It is important that capacity coupling *around* the crystal filter be minimized — in other words, no modulated signal must reach the 6SN7 mixer by any route except through the filter. Before construction is started, a decision must be made as to whether or not choice of sidebands is desired. If choice of sidebands *is* desired, a dual

Alignment

Alignment of the filter is straightforward, and once aligned it will need little attention.

1) Crystal "A" is first removed from the circuit. This crystal is best provided with a socket mount so it can be removed during alignment.

2) A calibrated signal generator covering the crystal range is connected to the grid of the triode section of the 6K8.

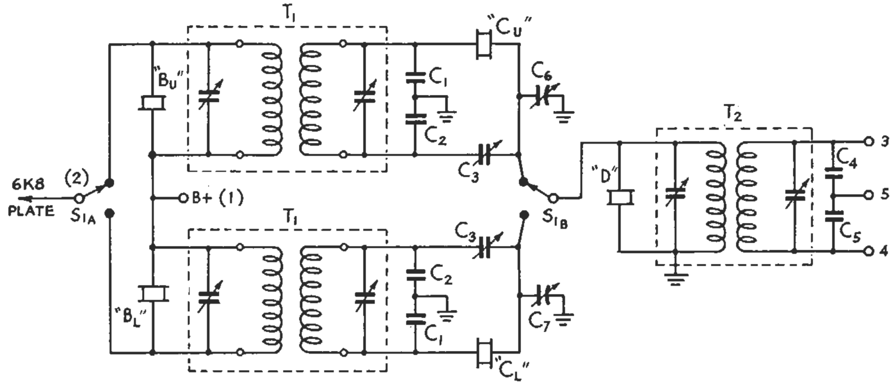


Fig. 5 — The double-channel crystal filter. All components are the same as in Fig. 1, except for the addition of the d.p.d.t. wafer switch, S_1, and the compensating condensers, C_6 and C_7 (3- to 30-$\mu\mu$f. ceramic). The trimmer on the input side of T_2 is set at minimum and the alignment procedure is followed with C_6 or C_7 wherever the instructions call for adjusting the input condenser.

filter using 5 crystals will be required. This filter is shown schematically in Fig. 5. A double-section wafer switch selects the upper or lower sideband. These wafer sections must be separated by approximately 3 inches to minimize stray coupling. In general, if the recommended physical layouts are followed, no undesirable coupling effects will be encountered. It is recommended that the crystals be wrapped with several layers of adhesive tape and then strapped to the chassis with metal brackets; connections may then be made by soldering to the holder pins. The physical layout and schematic diagrams (Fig. 7) with specifications provide sufficient information to allow even an inexperienced amateur to construct an efficient exciter.

3) A vacuum tube voltmeter is connected from one of the 6SN7 grids to ground.

4) Swing the signal generator through the crystal range until a maximum response is noted at the voltmeter. This will indicate the series-resonant frequency of crystal "C" and with the crystals described, based on a 450-kc. carrier, will be approximately 448.6 kc.

5) Align all transformer trimmers for maximum response on this frequency.

6) Next, adjust the signal generator *slowly* in the higher-frequency direction for a *null.* This will be the series-resonant frequency of crystal "D," 450 kc. with the crystals indicated.

7) Move the signal generator ½ kc. *lower* than this null and adjust the trimmer on the input side of T_2 for maximum response.

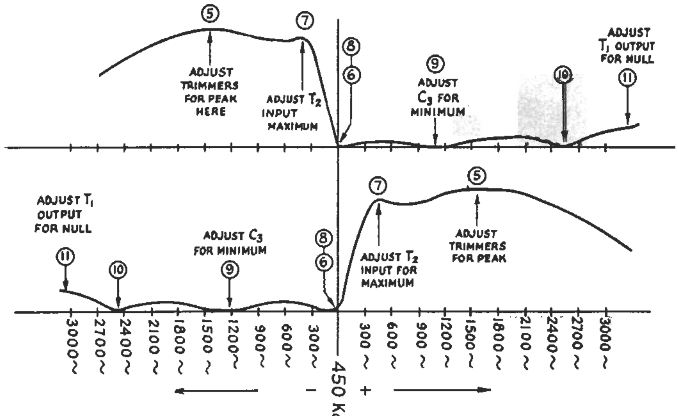


Fig. 6 — An alignment chart of the crystal filter. The numbers in the circles correspond to the steps outlined in the text.

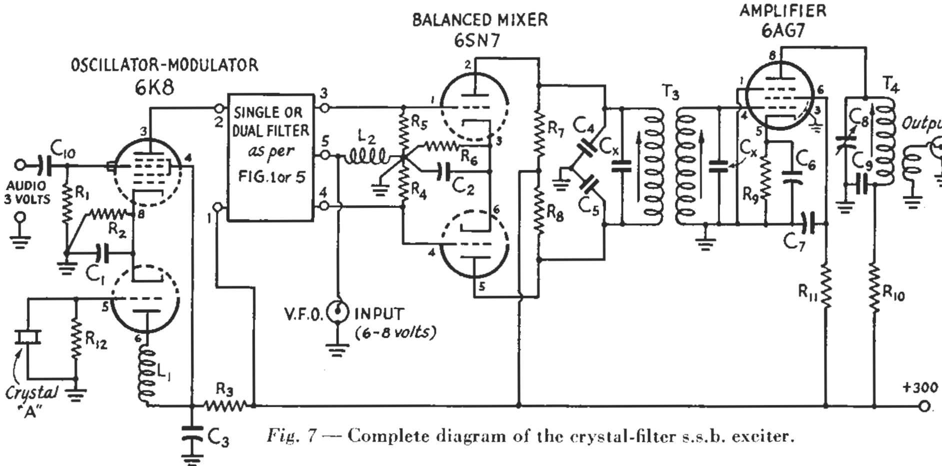


Fig. 7 — Complete diagram of the crystal-filter s.s.b. exciter.

C_1, C_2, C_3, C_6, C_7 — 0.1-μf. 400-volt paper.
C_4, C_5 — 39-μμf. ceramic.
C_8 — 100-μμf. variable air condenser.
C_9 — 0.02-μf. 600-volt mica.
C_{10} — 0.01-μf. 400-volt paper.
C_x — Trimmers in T_3.
R_1 — 0.47 megohm.
R_2 — 220 ohms.
R_3, R_{11} — 20,000 ohms, 1 watt.
R_4, R_5 — 0.1 megohm.
R_6, R_7, R_8 — 10,000 ohms.
R_9 — 150 ohms, 1 watt.
R_{10} — 1000 ohms.
R_{12} — 47,000 ohms.
All resistors ½ watt unless specified otherwise.
L_1 — 2.5-mh. r.f. choke.
L_2 — 0.5-mh. r.f. choke.
T_3 — 5-Mc. slug-tuned i.f. transformer.
T_4 — 5-Mc. slug-tuned i.f. transformer. Secondary removed and 8-turn link wound over cold end of primary. All fixed capacitors removed.

8) Return signal generator to null.

9) Move the signal generator approximately 1 to 1.2 kc. *higher* than the null and adjust C_3 for *minimum* response.

10) Move the signal generator *higher* until another null is found; this will be the series-resonant frequency of crystal "B," approximately 452.8 kc. with the crystals shown.

11) Continue approximately ½ kc. higher than this null and adjust the output trimmer on T_1 slightly for a moderate null.

12) Repeat Steps 7 through 11 to compensate for interaction, and alignment is complete.

For alignment of the dual filter the procedure is identical but must be done once for each sideband. However, when adjusting the filter for rejecting the *lower* sideband and where Steps 1–12 mention "higher" you must insert "lower" and vice versa. The alignment chart, Fig. 6, will simplify the alignment procedure of either filter.

The slug-tuned i.f. transformer is peaked at 3930 kc. and then staggered slightly to provide coverage of the entire 'phone band. The 6AG7 plate tank capacitor is adjustable from the front panel and is touched up when shifting frequency, as in the case of any transmitter amplifier stage.

Carrier Insertion

It is convenient to be able to reinsert carrier at the transmitter, for testing or in order to raise stations not used to tuning in s.s.b. signals. Carrier reinsertion in this exciter requires only one more tube, as shown in Fig. 8. A triode cathode-follower stage couples some of the r.f. from the oscillator to one grid of the 6SN7 through a potentiometer that permits setting the carrier at any desired level.

Miniature-Tube Equivalents

If it is desirable to use miniature tubes in the exciter, a 6BE6 can be substituted for the 6K8, but it is then suggested that a 12AU7 twin triode be used for the oscillator and the carrier-inserting cathode follower. A 12AU7 can be used in place of the 6SN7 balanced mixer, and a 6CL6 can be used instead of the 6AG7.

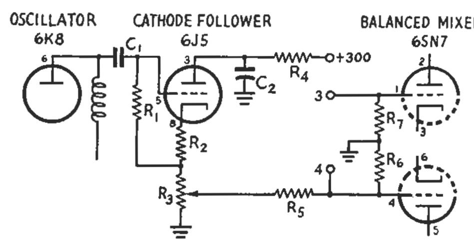


Fig. 8 — A cathode-follower circuit for carrier reinsertion.

C_1 — 47 μμf.
C_2 — 0.01-μf. 400-volt paper.
R_1, R_6, R_7 — 0.1 megohm.
R_2, R_4 — 1000 ohms.
R_3 — 20,000-ohm carbon potentiometer.
R_5 — 0.2 megohm.

Carl Eckhardt, W7BBK, has the following suggestions to make for improving the Edmunds exciter:

1) For T_3, use a Miller 1466 TV 4.5-Mc. sound i.f. transformer.

2) Use a shielded coil and a shielded variable condenser at T_4.

3) Try small resistors in the screen and grid leads of the Class A 6AG7, for eliminating instability.

» The popularity of the Edmunds exciter was increased when additional alignment-procedure information was made available, as detailed here. The method can be applied to any filter-type exciter, however.

Aligning the Edmunds Exciter

BOYCE S. WEBB, W4PIX

For anyone who has built a crystal-filter s.s.b. exciter of the type described by Edmunds and has had difficulty in aligning it properly, we would like to pass along a different tune-up procedure. The method, devised by W3MBY (aided and abetted by W4PIX), uses an audio oscillator and oscilloscope (or receiver with an S-meter and crystal filter). If a conventional audio oscillator is not available, the BC-221 or equivalent can be used. When using the BC-221 as an audio oscillator, the variations in output amplitude must be taken into account.

Although this method has only been used so far in aligning the Edmunds exciter, it should be equally applicable to other types of s.s.b. exciters, with a few alterations.

Alignment Procedure

The first step is to connect the audio oscillator and 'scope to the s.s.b. exciter, as shown in Fig. 1. A shielded lead should be used between the audio oscillator and the exciter. Fig. 1 indicates capacity

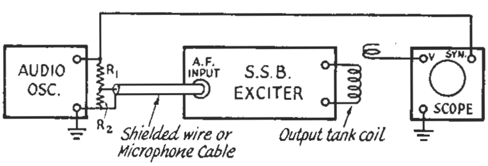


Fig. 1 — Test set-up for aligning a crystal-filter s.s.b. exciter, using an audio oscillator and an oscilloscope. Capacity coupling to the 'scope is shown, but link coupling to a tuned circuit at the vertical plates can be used.

R_1 — 0.2 megohm, approx.
R_2 — 100 ohms.

Value of R_2 determined by output amplitude of audio oscillator. R_1 and R_2 probably not required if BC-221 used as audio source.

coupling between the output tank coil and one vertical plate of the 'scope, but you can, of course, use link coupling between the s.s.b. exciter tank and a tuned circuit connected between the vertical plates. The exciter should be in complete working condition, with all voltages on and all r.f. circuits aligned. To avoid distortion in the exciter, the audio input should be kept low. The coupling to the output tank should be enough to display an adequate 'scope pattern.

If no 'scope is available, a receiver with an S-meter and crystal filter can be substituted. If the receiver has no S-meter, a v.t.v.m. can be connected across the diode load resistor of the second detector of the receiver. The crystal filter should be switched to the sharpest position, with the b.f.o. off. To insure that the receiver is picking up its signal only from the exciter output tank, a length of shielded wire or coaxial line should run from a 1-turn pick-up loop at the exciter over to the receiver antenna terminals. The antenna terminals should be short-circuited with a short

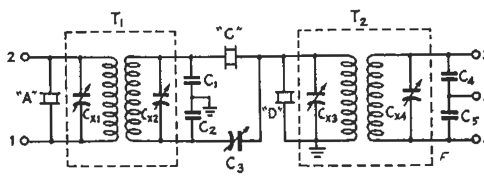


Fig. 2 — The crystal filter under discussion is the same as the original Edmunds filter. The trimmers are identified for use with the alignment table.

piece of small (No. 22 or so) wire. If the signal is insufficient for a reasonable S-meter reading, the shorting wire can be lengthened slightly.

Using the audio oscillator and the 'scope or receiver (or both), the procedure shown in Table I can be followed to align the Edmunds exciter for lower-sideband transmission. Dual filters (as described in the original article) can be aligned by appropriate changes in the procedure.

The circuit of the Edmunds crystal filter is shown in Fig. 2, for convenience in following the tune-up procedure. The transformer trimmers have been assigned symbols for ready reference in aligning the unit.

After the alignment procedure has been carried out, a check should be run on the response characteristics of the filter. This can be done very easily by varying the audio oscillator from 0 to 4000 cycles and observing the changes in output amplitude of the signal on the 'scope. If a 'scope is not available, a rough check can be obtained by watching the grid meter of the stage being driven by the exciter.

Determining Sideband Suppression

The approximate sideband rejection can be determined by either of the tune-up methods described in Table I, but reasonable care must be taken to insure that the correct frequencies or patterns are obtained. With a receiver connected as described, however, the method is straightforward and the carrier and sideband suppression can be measured in db., provided the S-meter calibration is known. Any S-meter can be checked

From "Aligning the Crystal-Filter S.S.B. Exciter," *QST*, August, 1952.

(and most of them should be) by the method shown in Fig. 3. Using a VFO or other signal source, switching from the top to bottom will give an approximate 6-db. decrease in signal at the receiver input. Thus it is a simple matter to

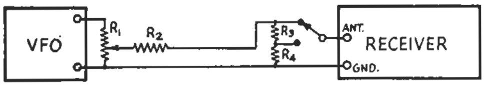


Fig. 3 — Any receiver can be calibrated in 6-db. steps by using a signal source (such as VFO) and suitable resistors.
R_1 — 1000-ohm potentiometer.
R_2 — 1000 ohms.
R_3, R_4 — 50 ohms.

start out at, say, 20 db. over S9 and work back down the scale, checking first the 6-db. interval and then resetting the level with R_1. This method is, of course, only satisfactory at high signal levels, and all stray pick-up should be minimized. The calibration should be made with a fixed setting of manual gain control and remains valid if the same setting of manual gain is always used, even when the receiver input is shorted as mentioned earlier. R_3 and R_4 should be mounted as close to the receiver input as possible.

When making measurements on the s.s.b. exciter, the r.f. input to the receiver should be controllable so that the receiver can always be run at the same setting of gain control that was used during calibration. The input can most easily be adjusted by changing the length of shorting wire at the antenna terminals, as mentioned earlier.

[1] Provided the receiver selectivity is sufficient to reject any carrier or other sideband signal that might otherwise affect the reading.

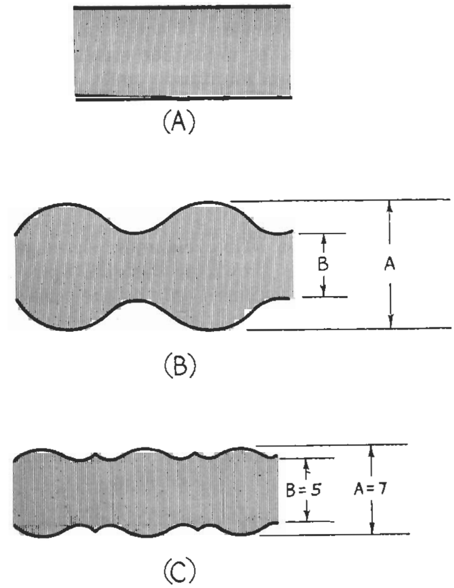


Fig. 4 — An oscilloscope can be used to determine the degree of carrier or sideband suppression. A perfect s.s.b. signal (single audio frequency) is shown at (A). With incomplete suppression, some modulation of the envelope will appear, as in (B) and (C). See text for calculation of the suppression.

Before each measurement, all modulation should be removed from the exciter and the carrier located, with the receiver in "sharp crystal." The test modulation frequency can then be fed to the exciter and the amplitude of the sidebands above and below the carrier noted on the S-meter. The difference between the two readings on the meter is the amount of sideband suppression.[1] To reduce the possibility of limiting in the exciter, the audio signal should be kept to the lowest level for which an indication of the unwanted sideband can be obtained.

The unwanted sideband suppression can be approximated by observing the output envelope on an oscilloscope, provided the carrier is suppressed sufficiently and certain other precautions are taken. A perfect s.s.b. exciter would put out a single radio frequency for a single audio input frequency, and the output envelope of such a signal would be the same as for a normal A3 unmodulated carrier, as shown in Fig. 4A. With a single audio frequency applied to an imperfect s.s.b. exciter, however, more than a single radio frequency appears in the output, and the envelope shows "ripples" that are simple or complex, depending upon how many separate frequencies are present and their relative amplitudes. The "two-tone" pattern is a simple and familiar

TABLE I

Alignment Procedure for the Edmunds S.S.B. Exciter

Step	*Audio Oscillator C.P.S.*	*Adjust*	*Using 'Scope, Adjust for . . .*	*Using Receiver with S-Meter and Sharp Xtal, Adjust for . . .*
1	1400	C_{x1}, C_{x2}, C_{x3}, C_{x4}, and resonate succeeding r.f. amplifiers	Maximum height	First locate carrier (audio off), tune 1400 cycles lower (audio on), adjust for maximum reading
2	500	C_{x3}	Maximum height	Relocate carrier, tune 500 cycles lower, adjust for maximum reading
3	1100	C_3	Average flatness of troughs	Relocate carrier, tune 1100 cycles higher, adjust for minimum
4	3300	C_{x2}	Average flatness of troughs, maintaining height	Relocate carrier, tune 3300 cycles higher, adjust for minimum
5	1400	C_{x4}, C_{x1}	Average flatness of troughs, maintaining maximum height	Relocate carrier, tune 1400 cycles lower, adjust for maximum
6	Repeat all of the above steps except No. 1, to correct for interaction.			

example of this that, as the name implies, contains only two frequencies, of equal amplitudes. In using the 'scope to determine sideband suppression, the carrier amplitude must be small enough to be negligible or its presence will complicate the patterns. The carrier amplitude can be checked by removing any audio signal and then turning the exciter off. A decrease in the thickness of the 'scope trace will indicate the presence of carrier.

If, for example, a 1000-cycle audio signal produces a pattern as shown in Fig. 4B, and the envelope variation is caused only by the presence of the unwanted sideband, then the ratio of desired sideband to undesired sideband is given in decibels by

$$S = 20 \log \frac{A + B}{A - B}.$$

If either the audio signal or the audio amplifier has distortion, the contour of the envelope will not be sinusoidal because more than one audio input signal is being applied. The effect will be practically the same if carrier is present. A good approximation can only be made if the signal input is kept low, to avoid distortion and flattening.

Fig. 4C shows a complex envelope, obtained with a single audio frequency, and is an example of how the peak instantaneous sum of all unwanted frequencies can be approximated using this method. In the case shown, where the ratio of maximum to minimum is

$$\frac{7 + 5}{7 - 5},\ S = 20 \log \frac{12}{2} = 20 \log 6 = 15.6 \text{ db.}$$

An attempt should be made to keep these undesired emissions as small as possible, and 30 to 40 db. down for the unwanted sideband and 50 db. down for the carrier represent reasonable values for amateur s.s.b. equipment.

Hints on the Edmunds Exciter

Ken Stiles, W2MTJ, passes along some useful tips on the Edmunds exciter. To quote his letter, "A number of fellows have trouble getting adequate carrier suppression in the exciter. The trouble appears to be that the crystal oscillator operates on one mode of crystal resonance and the suppression crystal operates on the other, with a consequent difference in frequencies of 500 cycles or more. A change in the oscillator circuit of the 6K8 triode section will take care of the situation. The arrangement in Fig. 1A, using a Pierce oscillator, works fine at W2MTJ and has been successfully adapted by several others. It results in the oscillator frequency being very close to the frequency of the suppression crystal. By tuning C_1, the oscillator frequency can be set exactly in the slot of the suppression crystal. No changes are required in the remainder of the 6K8 circuit. The oscillator injection voltage obtained is adequate for several volts of audio input. In

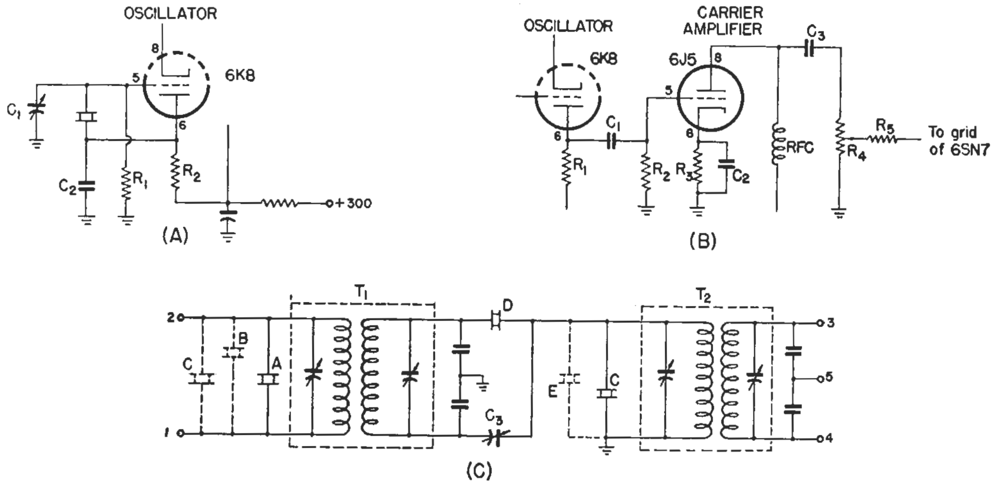


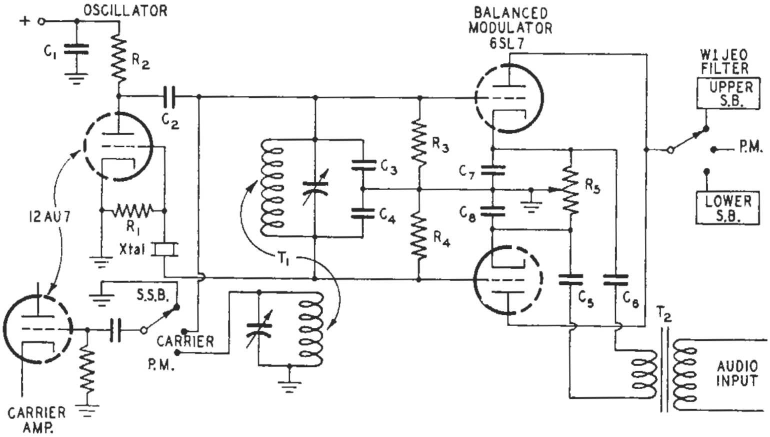
Fig. 1 — Improvements for the Edmunds crystal-filter exciter suggested by W2MTJ.

The revised oscillator circuit at A permits adjustment of the oscillator frequency and consequently better carrier rejection. C_1, 100-$\mu\mu$f. adjustable; C_2, 470 $\mu\mu$f.; R_1, 0.1 megohm; R_2, 20,000 ohms.

The revised oscillator circuit does not offer as much voltage available for carrier reinsertion, and the circuit of B gives more amplification. C_1, 47 $\mu\mu$f.; C_2, 0.01 μf.; C_3, 0.001 μf.; R_1 and R_4, 20,000 ohms; R_2, 0.12 megohm; R_3, 1000 ohms; R_5, 0.2 megohm.

Better sideband rejection is obtained by paralleling additional crystals, as shown by the dotted lines in C. Examples of crystal-channel numbers (although these don't have to be used — any adjacent channels will be suitable) are: *A*, 323; *B*, 322; *C*, 321; *D*, 320; *E*, 321, edge-ground to raise frequency several hundred cycles.

Fig. 2 — The VE6CN revision of the Edmunds exciter.
C_1 — 8-µf. 450-volt electrolytic.
C_2, C_7, C_8 — 0.005-µf. mica.
C_3, C_4 — 100-µµf. mica.
C_5, C_6 — 0.25-µf. paper.
R_1 — 47,000 ohms.
R_2 — 33,000 ohms.
R_3, R_4 — 0.15 megohms.
R_5 — 2500-ohm potentiometer.
T_1 — 450-kc. i.f. transformer.
T_2 — Small Class B input transformer.



my exciter I limit the audio input to 1 volt, however, to get the proper level for the two-tone test pattern.

"When the Pierce oscillator was first used like this, the amount of carrier at the output of the carrier-reinsertion amplifier was inadequate, and a slight change was required in the circuit, as shown in Fig. 1B. With this arrangement, sufficient carrier reinsertion is obtained to be equal (at the 6SN7 grids) to the component obtained with 1 volt of audio input to the 6K8.

"The use of additional crystals in the filter has proven beneficial. These crystals, placed in parallel with the normal suppression crystals, will provide more uniform suppression of the unwanted sideband and can even be used to control the 0–200-cycle portion of the wanted sideband to obtain a more desirable speech-equalization characteristic. The additional crystals can be added either in parallel with the sideband-suppression crystal, *A* of Fig. 1C, or in parallel with the carrier-suppression crystal, *C*. Example crystal-channel numbers are given in the sketch, but any crystals in the range can be used, provided they are for adjacent channels and the same relationship is retained. The use of multiple crystals results in a slight increase in capacity across the transformer winding, and the transformer tuning condenser must be retuned when the crystals are added. It should be noted that crystals of the two-digit series that fall in the desired frequency range may also be used to fill in between the three-digit-series crystals normally used. If these are not available, the crystals at hand can be edge ground to fit them in where a hump is observed in the suppression characteristic of the unwanted sideband. With a little time spent in adding crystals, and retuning C_3 to obtain the best suppression characteristic, it is possible to obtain an average suppression of the unwanted sideband near 30 db. C_3 should be checked for optimum tuning each time a crystal is added or changed."

—•••—

Balanced Modulator

G. O. Kincaid, VE6CN, uses an interesting balanced-modulator circuit in his version of the Edmunds exciter. Shown in Fig. 2, it has the additional feature of providing for p.m. The oscillator circuit starts faster than the original, and it can be "pulled" by tuning, for better alignment with the filter. The filter is first aligned by removing the oscillator crystal, introducing a BC-221 signal across C_4, and using the original alignment procedure. Then the oscillator crystal is replaced and the oscillator frequency trimmed by adjusting the trimmer in the primary of T_1.

The p.m. is obtained by tuning the secondary of T_1 until the voltage fed to the carrier-amplifier tube is 90 degrees different than that applied to the balanced modulator. For p.m., the crystal filter must be out of the circuit, of course, since the p.m. signal is obtained by combining the two sidebands (less carrier) with a carrier that has been shifted 90 degrees. Checking with a 'scope, one simply

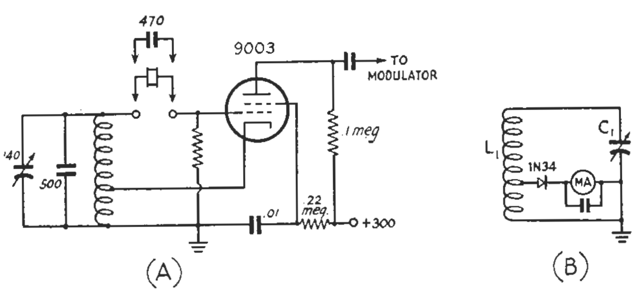


Fig. 3 — W7CJB uses the circuit at A for the oscillator in his Edmunds exciter. Plugging in the 470-µµf. condenser in the grid circuit gives a VFO for aligning the crystal filter, and plugging in the crystal gives a crystal oscillator that can be "pulled" about 400 cycles for accurately dropping the frequency into the filter notch.
The absorption wavemeter at B is used to indicate output from the 6AG7 stage of the exciter, by placing it near the 6AG7 tank coil and tuning the wavemeter to resonance.
L_1C_1 — Tunable to 75-meter band. L_1 is tapped about ¼ total turns.
MA — 0–150 microammeter.

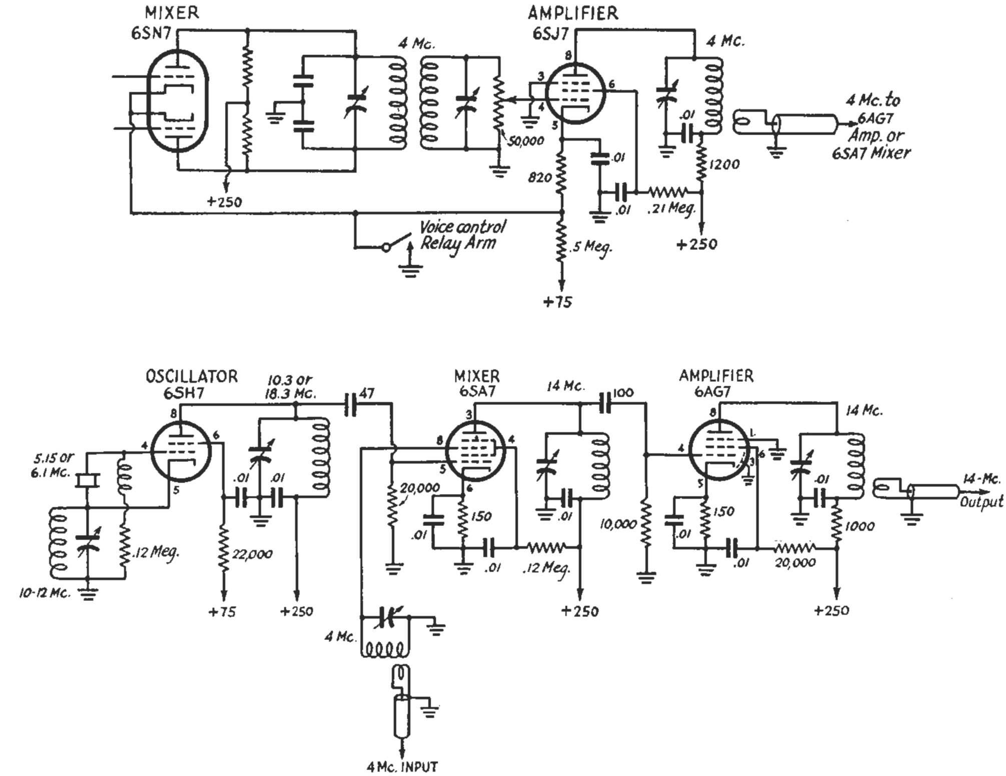


Fig. 4 — Here is the way W2FGV converts the 4-Mc. output from his Edmunds exciter to 14 Mc. The top circuit shows the only modification in the original circuit, the addition of the 6SJ7 amplifier, and the bottom circuit shows the 6SA7 mixer and a 6AG7 amplifier working on 20. The exciter is built on a 13 × 17 × 3-inch chassis and is used to drive a pair of 807s running 125 watts on either 75 or 20.

sets the trimmer in the secondary until there is minimum amplitude change on the carrier with modulation. The two circuits of T_1 "pull" a little, so the complete adjustment may require several trial runs.

— • • • —

An Oscillator for the Edmunds Exciter

Two of the problems of construction of an Edmunds crystal-filter exciter are finding a suitable test oscillator and accurately setting the oscillator frequency in the filter notch. Woody Davey, W7CJB, solves the problems by using the oscillator circuit shown in Fig. 3A (page 57). When the 470-$\mu\mu$f. grid condenser is plugged in (it's mounted in a crystal holder), a VFO is available for aligning the filter. Replacing the condenser with the proper crystal gives a crystal oscillator that can be pulled about 400 cycles by tuning the 140-$\mu\mu$f. variable.

Woody uses the simple absorption wavemeter of Fig. 3B to indicate output at the 6AG7 plate circuit — he likes it a lot better than a v.t.v.m. for the job. When used in conjunction with the VFO, he can align the crystal filter in about 90 seconds.

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Converting to 20 Meters

John Grubb, W2FGV, thinks many of the fellows with Edmunds exciters may stay off 20 because there have been no descriptions of suitable frequency converters, so he sends along the circuit of the unit he has been using successfully for some time. Shown in Fig. 4, it should require no further details or explanations, since the techniques involved are standard receiver and low-level transmitter practices.

And while we're talking about the Edmunds exciter, Harold Klaiss, W4QN, thinks it would be well to point out that you don't need the double-channel filter originally described to get a choice of upper or lower sideband. If, for example, your suppressed-carrier frequency starts out at 450 kc., using an oscillator at either 3450 or 4350 kc. will put your (suppressed) carrier at 3900 kc., with upper sideband in one case and lower sideband in the other.

» 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 condenser-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 line 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 already considered. 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.

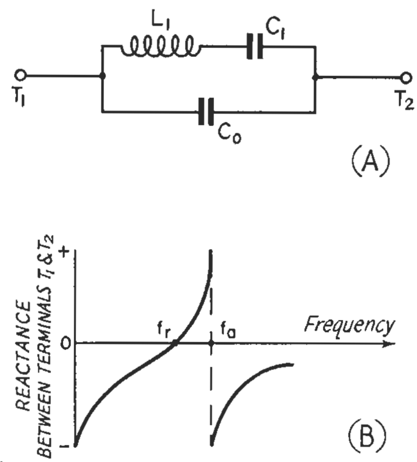


Fig. 1 — The equivalent electrical circuit of a piezoelectric crystal (A). The reactance varies with frequency as in (B).

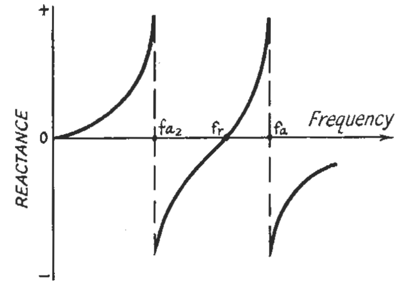


Fig. 2 — Reactance plotted against frequency for a crystal shunted by an inductance.

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

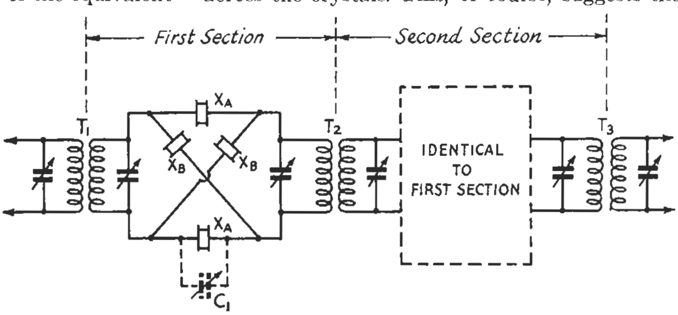


Fig. 3 — Schematic diagram of a two-section crystal lattice filter.
T_1, T_2, T_3 — Replacement-type i.f. transformers.
X_A, X_B — Matched pairs of FT-241 crystals (see text).
C_1 — 1- to 5-$\mu\mu$f. ceramic condenser (see text).

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

use of radio-frequency transformers (ordinary i.f. transformers) as input and output devices as well as spreading coils for f_r and f_a. It might be well to mention at this point that when f_r and f_a are spread, f_r remains fixed in frequency and only f_a is moved higher in frequency.

Let us briefly consider what happens inside the lattice filter. Assume that the pair of crystals connected in shunt (X 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 condensers, 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,

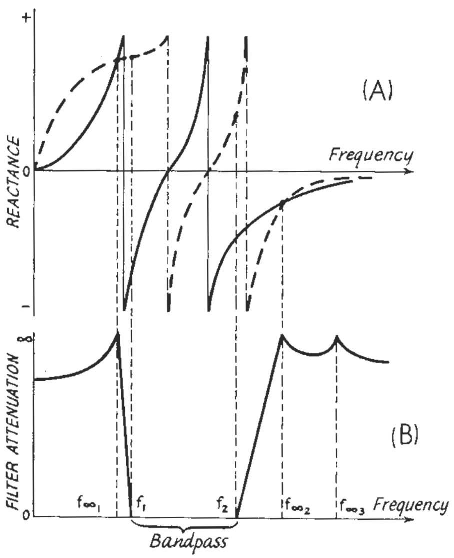


Fig. 4 — **The reactance-vs.-frequency characteristic for the two pairs of crystals in a lattice filter section is shown at (A). The resultant attenuation characteristic is shown in (B).**

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 $f_{\infty 3}$) where 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.

If the reader is familiar with the operation of any of the various forms of bridge circuits, it will be reasonably obvious how the lattice (another name for a bridge) yields the characteristic shown in B. When the reactances of the bridge arms are equal and of the same sign the signals through the two possible paths of the bridge will cancel out. When the reactances are of opposite sign there will be partial transmission through the network with the maximum signal transmitted in the passband at the points where the reactances are equal in amplitude, but still opposite in sign.

Practical Filters

A workable filter can be constructed at the cost of only a very few dollars! The FT-241-A low-frequency surplus crystals were tried with very good success. Very inexpensive coupling devices were used, ordinary replacement i.f. transformers (Meissner No. 16–5712). There is one sacrifice made that was probably due either to an improper choice of transformers or an impedance mismatch between crystals and transformers. This was an insertion loss of approximately 12 to 15 db. in the middle of the passband. However, the authors felt that this did not handicap the system too greatly, as this was less than could be gained in a single stage of ordinary i.f. amplification. A more careful study would undoubtedly be helpful in this matter of insertion loss, but available time and practical considerations forced us to the solution presented.

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?

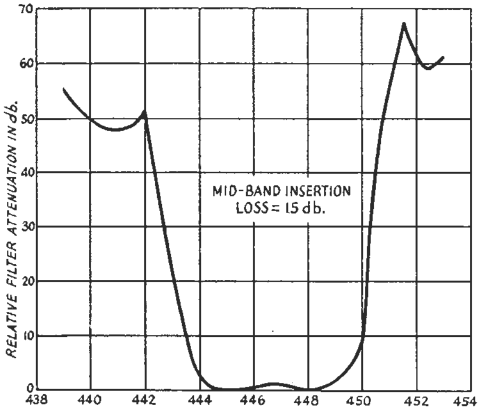


Fig. 5 — **Attenuation characteristic of an experimental crystal lattice filter (two sections) suitable for receiver use. The crystals were Channels 40 and 42 of the FT-241 series.**

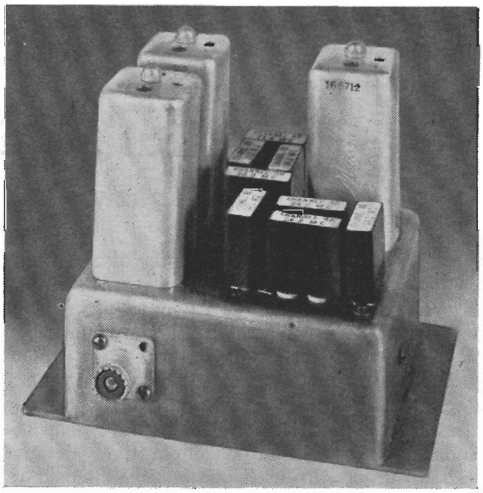
There will be very narrow attenuation slots in the edges of the passband of the filter. The commercial companies get around this problem by putting two sets of silver plating on a crystal and attaching four terminals, making the one crystal serve as two identical crystals. It's a very nice trick but 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 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 where serious trouble is encountered.

When the narrow-band lattice filter is used for transmitting, as described in the article on page 62, the primary of the first i.f. transformer is connected for series tuning instead of the parallel tuning shown in Fig. 3.

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., 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.



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

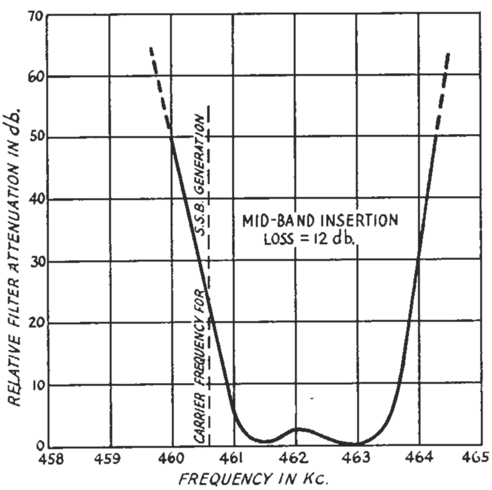


Fig. 6 — **Attenuation characteristic of an experimental crystal lattice filter (two sections) suitable for a s.s.b. receiver or transmitter. The crystals were Channels 49 and 50 of the FT-241 series.**

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" condenser[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 condenser. Usually only one or two μμf. will be sufficient to align a typical off-frequency crystal. You will notice that the trimmer adjustments on the i.f. transformers may be used to equalize the passband characteristic and make it as flat as possible.

[1] A gimmick condenser 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.

Applications

The two obvious places where these filters could be put to good use are in receivers and in single-sideband transmitters. The best place to add such a filter in a standard communications receiver would be between the first detector and the first i.f. amplifier. This might be a bit cumbersome to add to an existing receiver without a major rebuilding job. The other alternative is to build a separate i.f. strip incorporating the filter in the line-up and use it much as the popular "lazy man's Q5-er" is used.

» *The lattice filter in an exciter, with some circuits for balanced modulators and an all-electronic voice-controlled break-in system.*

A Crystal Lattice S.S.B. Exciter

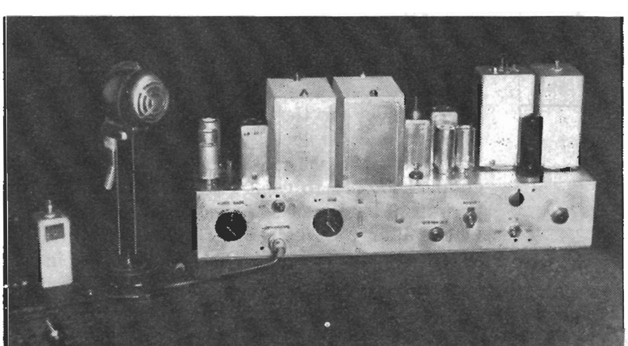
C. E. WEAVER, W2AZW, AND J. N. BROWN, W3SHY, EX-W4OLL

This exciter contains a majority of the essential features which most s.s.b. operators consider desirable. No attempt was made to design an exciter with a bare minimum number of tubes or parts. It was felt that the exciter should not require the addition of external audio stages, voice-control or other appendages to make it complete. These features have been included in a unit that is comparatively simple and easily lined up with test equipment usually available to most ham stations. When once adjusted properly, the exciter should stay in alignment for months without further attention.

The VFO was not included as part of the exciter because this item is usually available as part of the existing station equipment. Generally, the VFO can be modified to cover the frequency range required by the exciter. This is done by adding enough capacity to the existing tuned circuits of the VFO so that it will tune approximately 475 kc. lower than the 'phone band to be used. For 3.8- to 4.0-Mc. operation, this will be in the vicinity of 3.325 to 3.525 Mc. The exact frequency range will depend on the choice of crystals for the lattice filter in the exciter.

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

The exciter shown in the photograph is the original experimental model. The chassis is one that happened to be on hand and is adequate for the purpose, but rather difficult to wire. The builder can use any size chassis or arrangement desired. However, a few precautions should be observed. The low-frequency oscillator (V_3, Fig. 1) and associated components should be isolated as far as practicable from the other equipment, and particularly from the output of the crystal filter, to prevent leakage of the carrier around the filter that would result in incomplete carrier suppression. The crystal lattice sections and



A complete s.s.b. exciter using the crystal lattice filter. The small chassis in the left foreground contains the low-frequency oscillator circuit and the sideband selector switch.

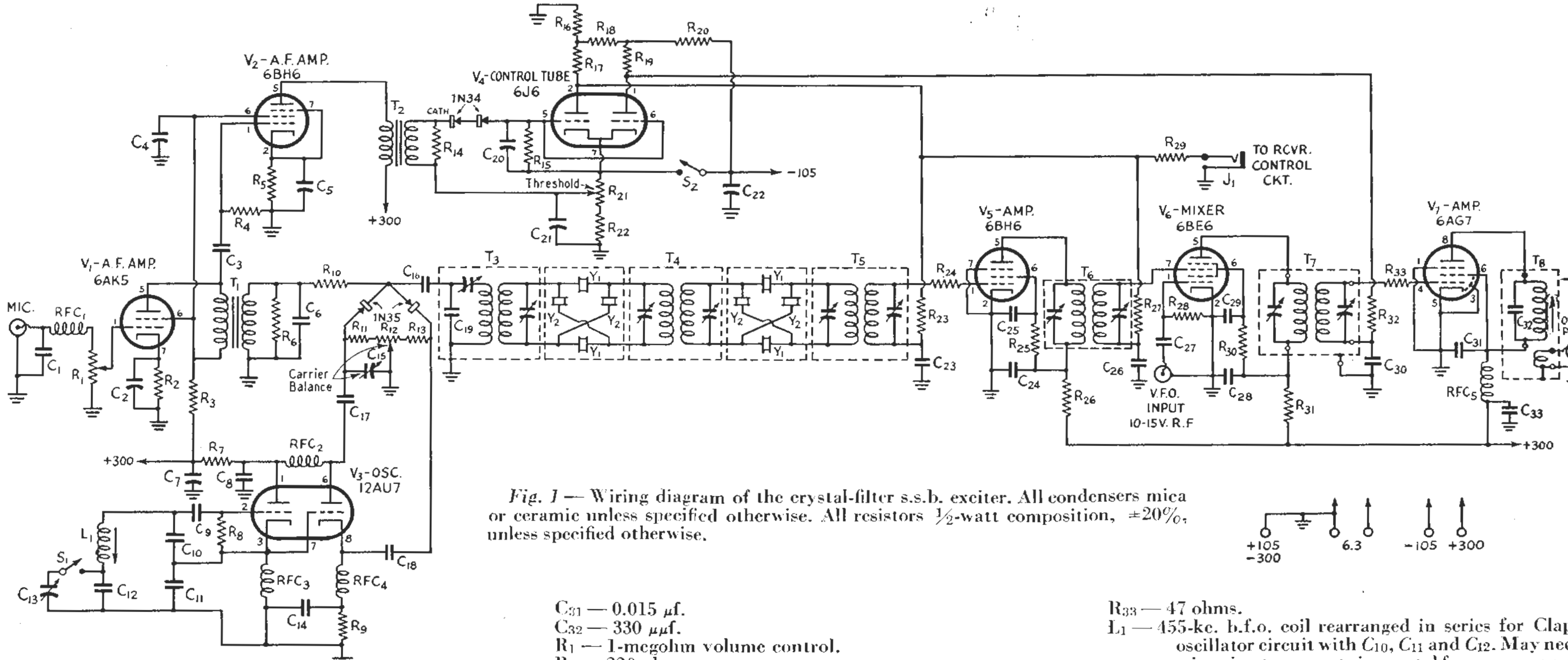


Fig. 1 — Wiring diagram of the crystal-filter s.s.b. exciter. All condensers mica or ceramic unless specified otherwise. All resistors ½-watt composition, ±20%, unless specified otherwise.

C_1, C_9 — 220 μμf.
C_2, C_5 — 10-μf. 25-volt electrolytic.
C_3, C_{16}, C_{19} — 0.001 μf.
C_4 — 10-μf. 450-volt electrolytic.
C_6, C_{17}, C_{18} — 330 μμf.
C_7 — 20-μf. 450-volt electrolytic.
C_8, C_{14} — 0.01 μf.
C_{10}, C_{11} — 0.0039 μf., silvered mica.
C_{12} — 550 μμf. (or sufficient to resonate L_1 to crystal-filter range).
C_{13} — 50-μμf. air trimmer.
C_{15} — 50-μμf. air trimmer. Required only if complete carrier balance cannot be obtained with R_{12}. If balance not obtained with C_{15} across R_{11}, try across R_{13}.
C_{20} — 0.1-μf. oil-filled or high-grade paper.
C_{21}, C_{22} — 0.05-μf., 300-volt paper.
C_{23}, C_{24}, C_{25}, C_{26}, C_{28}, C_{29}, C_{30}, C_{33} — 0.0082 μf.
C_{27} — 100 μμf.
C_{31} — 0.015 μf.
C_{32} — 330 μμf.
R_1 — 1-megohm volume control.
R_2 — 220 ohms.
R_3 — 15,000 ohms, 5 watts.
R_4 — 0.22 megohm.
R_5 — 150 ohms.
R_6, R_{10} — 1000 ohms.
R_7 — 18,000 ohms, 5 watts.
R_8, R_{14}, R_{17}, R_{19} — 47,000 ohms.
R_9 — 2700 ohms.
R_{11}, R_{13} — 820 ohms ±5%.
R_{12} — 250-ohm linear carbon potentiometer.
R_{15} — 2 megohms.
R_{16} — 510 ohms.
R_{18} — 4700 ohms ±5%.
R_{20} — 47,000 ohms ±5%.
R_{21} — 20,000-ohm wire-wound potentiometer.
R_{22} — 82,000 ohms.
R_{23}, R_{27} — 0.1 megohm.
R_{24} — 2200 ohms.
R_{25}, R_{30} — 47,000 ohms, 2 watts.
R_{26}, R_{29} — 4700 ohms.
R_{28}, R_{32} — 22,000 ohms.
R_{31} — 1500 ohms, 1 watt.
R_{33} — 47 ohms.
L_1 — 455-kc. b.f.o. coil rearranged in series for Clapp oscillator circuit with C_{10}, C_{11} and C_{12}. May need trimming to resonate in crystal frequency range.
T_1 — 20,000-ohm plate to 600-ohm line audio transformer (Stancor A-3250).
T_2 — Midget 2:1 interstage audio transformer, reversed.
T_3, T_4, T_5, T_6 — 455-kc. interstage i.f. transformers. Must tune through crystal-filter range. Note revision of T_3 primary connection.
T_7 — 4-Mc. i.f. transformer (capacity-loaded 5-Mc. unit), remounted in National PB-10 plug-in shield can.
T_8 — Plate coil is 31 turns No. 22 enam. close-wound on Millen 69016 coil form. Link winding, 4 turns over B+ end of winding. Assembly mounted in National PB-10 can with C_{32}.
RFC_1, RFC_3, RFC_5 — 2.5-mh. r.f. choke.
RFC_2, RFC_4 — 60-mh. r.f. choke, iron core, shielded.
S_1 — S.p.s.t. ceramic switch for sideband selection.
S_2 — Toggle switch for disabling voice-control circuit.
Y_1, Y_2 — Filter crystals (surplus FT-241-A, adjacent channels of two-digit series with fundamental frequency preferably between 430 and 490 kc. Y_1 is higher, Y_2 is lower adjacent channel.

TABLE I — ALIGNMENT PROCEDURE

Test No.	Test	Signal Generator Frequency[1]	Signal Generator Connection	Measuring Set Connection[3]	Measuring Set Adj.[2]	Test Condition	Adjust . . .	Adjust for . . .
1	Crystal-filter alignment	Center of crystal-filter passband	Apex junction of 1N35, R_{10} and C_{16} through 20-$\mu\mu$f. condenser	Plate of V_6 through 20-$\mu\mu$f. condenser	Tune to sig. gen. freq. AVC-On BFO-Off	S_2 open, V_3 removed, b.f.o. disconnected, R_1 off (CCW)	T_3, T_4, T_5, T_6	Maximum S-meter reading
2	Same	Vary sig. gen. plus and minus 3 kc. as required under "Adjust for"	Same	Same	Same	Same	None	Note S-meter reading at several points above and below center freq., devia. and loss *vs.* freq. should be approx. same as Fig. 6, page 61 [4, 5]
3	Converter V_6 and output amplifier V_7 alignment	Center of crystal-filter passband	Same	Output of T_8 (link) loaded with 50-ohm 5-watt carbon resistance. Antenna lead in close proximity to 50-ohm res. but not directly connected	Tune to desired operating freq. (3.8–3.999 Mc.) AVC-On BFO-Off	Same except VFO connected and adj. to desired operating freq. (3.8–3.999 Mc.) minus center freq. of crystal filter [8]	T_7, T_8	Maximum S-meter reading [3]
4	Carrier adjustment	None	None	Same	Same	Same except V_3 in place, S_1 open, R_{12} full CCW or CW [7]	L_1	Maximum S-meter reading [3]
5[9]	Same	None	None	Same	Same	Same	L_1 toward high freq. side of filter. (Decrease inductance)	Decrease of five S points from max. obtained test No. 4 (25–30 db.)
6[9]	Same	None	None	Same	Same	Same except S_1 closed	C_{13} from min. toward max.	Same as test No. 5
7	Carrier Suppression	None	None	Same	Same	Same	R_{12}, C_{15}[6]	Minimum S-meter reading [6]
8	Voice quality check	None	None	Same	AVC-Off, BFO-On, a.f. gain-full on, r.f. gain for normal receiver level	Same except mic. connected	None	Tune receiver slowly for normal intelligibility while talking into mic. — observe speech quality [10]
9	Voice control operation and adjustment	None	None	Same	Same	Same except S_2 closed	R_{21}	Adjust slightly below the point where clipping disappears

[1] Preferably a BC-221 frequency meter. A good signal generator should also be satisfactory. A shielded cable should be used between the signal generator and point of connection.

[2] A communications receiver used as a measuring instrument. If 200-kc. to 500-kc. band is not available, use crystal-controlled converter suggested on page 61. A shielded cable should be used between the point of connection and receiver input.

[3] For best accuracy, maximum S-meter readings should be taken around S9 or slightly above. The output of the signal generator should be adjusted as required. A carbon potentiometer output control may be necessary if a BC-221 is used.

[4] It is assumed that the S-meter calibration is approximately 5 db. per unit.

[5] If S-meter reading is too high, adjust R_{12} to lower.

[6] Try adjustment first with C_{15} disconnected. If suppression is insufficient, connect C_{15} as shown (Fig. 1) and alternately adjust R_{12} and C_{15} for minimum carrier. If no improvement is noted, move outside connection of C_{15} to the opposite side of the bridge and repeat the adjustments.

[7] The r.f. voltage from either side of the bridge will ordinarily be between 1.5 and 2.5 volts with the components shown. It should be measured with a v.t.v.m. if available.

[8] The r.f. voltage at the No. 1 grid of V_6 (Pin No. 1) should be checked initially, by placing a 0–1 ma. meter in series with the ground end of R_{28}, by-passed with a 0.01-μf. condenser at the point of connection. The d.c. through R_{28} should be between 0.5 and 0.8 ma.

[9] Repeat several times alternately.

[10] Repeat test with S_1 set for other sideband.

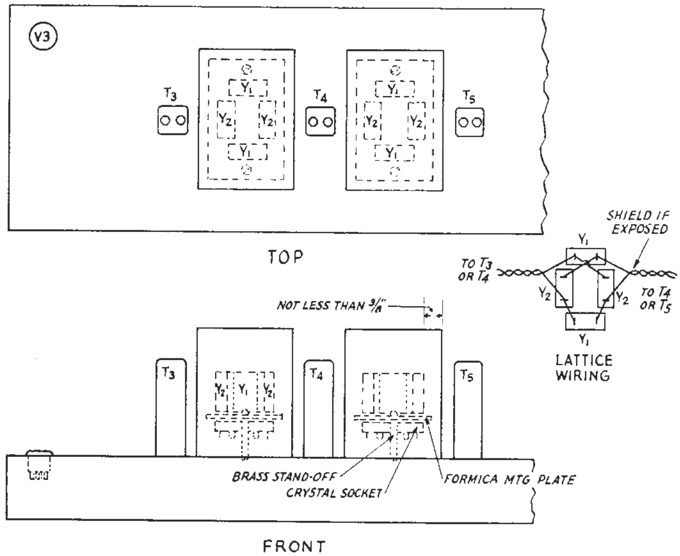


Fig. 2 — **A suggested chassis arrangement for the crystal filter.**

associated transformers should be arranged approximately as shown in Fig. 2. The crystals and their sockets should be mounted above the chassis in a shield cover, large enough so that it does not add appreciable distributed capacity to the crystals. It is also advisable to shield the wires between the transformers and the crystal lattice sections to prevent stray pick-up. This wiring should be kept as short as possible.

The narrow-band crystal filter was chosen for the exciter because it is easier to adjust than the wide-band filter and normally should not require additional shunt capacity across the crystals. Tests on several sets of crystals indicated that the capacity due to the plating alone is sufficiently uniform for all practical applications. The crystals used were Channels 49 and 50 of the surplus FT-241-A type, although any adjacent channels of the two-digit serial number with fundamental frequencies between approximately 430 kc. and 490 kc. should be satisfactory.

The Exciter Circuit

Referring to Fig. 1, the circuit diagram of the exciter, briefly the operation is as follows: V_1 amplifies the voice frequencies from the microphone and steps the output down to approximately 600 ohms impedance, which feeds into the balanced modulator (twin crystal diode 1N35) to combine with the carrier from oscillator V_3. This produces an amplitude-modulated signal with the carrier suppressed. The resulting upper and lower sidebands feed into T_3 through an impedance-transforming network consisting of C_{19} and the trimmer condenser in series with the primary winding of T_3. As the signal passes through the filter, one sideband is removed. V_5 amplifies the remaining sideband signal from the filter output and feeds it into the r.f. input grid of V_6 where it is mixed with an external VFO to produce the final frequency. Mixer tube V_6 is coupled to amplifier tube V_7 through a 4-Mc. i.f. transformer. V_7 is a 6AG7 operating Class A and will deliver approximately 3 watts peak power into a 50-ohm load. T_7 and T_8 are enclosed in National plug-in shields so that transformers can be constructed to cover other bands if desired. The coil values given here are for the 75-meter 'phone band only.

Voice Control

V_2 is the additional audio amplifier for the voice-control circuit. V_4 is a 6J6 that serves as the voice-control tube. The secondary of T_2 connects to the 1N34 rectifiers (in series for voltage and back-resistance considerations) and R_{15} and C_{20}. Potentiometer R_{21} is a threshold sensitivity control, which is advantageous where room noise is present and it is desired to lower the sensitivity of the voice control to prevent false operation. When there is no audio input to the exciter, the grid of V_4 has zero bias, thus permitting it to conduct. This in turn causes a voltage drop across R_{17} and R_{19} that biases V_5, V_6 and V_7 to cut-off, disabling the exciter and permitting the receiver to pick up incoming signals without being masked by thermal and room noise from the transmitter output stage. When the microphone is energized, V_1 and V_2 amplify the audio, which in turn is rectified by the 1N34s. The resultant voltage is used to bias the grids of V_4 negative and to cut off the plate current. This turns on the exciter by placing normal operating bias on V_5, V_6 and V_7 from voltage divider R_{16}, R_{18} and R_{20}.

If the receiver control unit, Fig. 3, is connected to the transmitter voice control circuit, J_1, Fig. 1, the receiver will be automatically disabled during transmission and restored to receiving condition when the operator stops talking. Referring again to Fig. 1, no clipping is discernible even on the first syllable because the voltage produced across the secondary of T_2 is comparatively high and the bias required to cut off

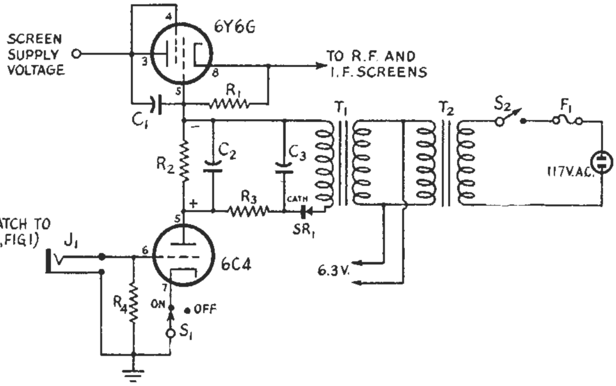


Fig. 3 — The receiver voice-control circuit.
C_1 — 0.01 μf., 600 volts.
C_2, C_3 — 16-μf. 150-volt electrolytic.
R_1 — 0.47 megohm, ½ watt.
R_2, R_4 — 0.22 megohm, ½ watt.
R_3 — 22,000 ohms, 1 watt.
T_1 — 6.3-volt 1-amp. filament transformer.
T_2 — 6.3-volt 1.5-amp. filament transformer.
F_1 — ½-amp. fuse.
J_1 — Open-circuit jack.
S_1, S_2 — S.p.s.t. toggle.
SR_1 — 50-ma. selenium rectifier.
If battery bias is preferred, replace R_2 with a 50- or 90-volt hearing-aid battery and omit T_1, SR_1, C_2, C_3 and R_3.

the plate current of V_4 is quite low; therefore, the transmitter is energized in the matter of a few milliseconds. When the operator stops talking, it takes a longer time for the bias to drop to zero through the R_{15} and C_{20} network. The network values shown are about optimum, but the "hangover" time can be changed to suit the individual by the proper choice of values for R_{15} and C_{20}. The transmitter and receiver control circuits are timed so that the receiver is completely disabled a few milliseconds before the transmitter is energized. When the operator stops talking, the receiver is again energized approximately 0.1 second after the transmitter is de-energized. This split-second timing permits the operator to engage in a normal conversation with one or more single-sideband stations similarly equipped and operating on the same frequency, without clicks or howls from the receiver as it is automatically switched on and off.

Alignment

The alignment is covered step by step in Table I. Where measurements are made with the signal generator and receiver operating on the same frequency, previous tests should be made to insure that the receiver does not pick up an appreciable amount of signal from the generator through stray coupling, even when they are disconnected.

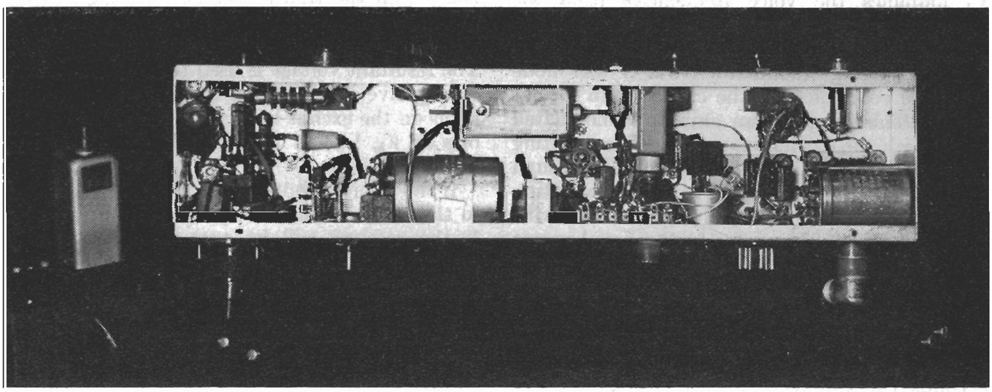
When selecting the crystals for the filter, they should be "paired" for each section; that is, the two like crystals in each section should be as nearly identical in frequency as possible. Tests indicate that the frequency of a pair can deviate as much as 50 cycles from each other without serious effect. Generally, most of the 241-A crystals with the same channel designation will be found to be within these tolerances. Occasionally, one may be found that deviates 100 cycles or more. If this occurs, it should be replaced with another crystal, or the lower-frequency one can be edge-ground slightly until it matches the higher-frequency crystal of the pair.

Operation

The exciter can be used directly as a low-power transmitter, but it is not recommended, particularly on 75 meters. A very creditable job can be done with a pair of Class B 807s following the exciter. If everything is properly lined up, the unwanted sideband at the output of the exciter will be suppressed 50 db. or more and distortion products should be more than 35 db. down at 3 watts output. The linear amplifier following the exciter will more than likely determine the amount of distortion products radiated by the antenna.

The authors express their appreciation to W. S. Felch, W2EAS, for his many constructive suggestions during the preliminary work on the crystal filter and subsequent design of the exciter.

A view underneath the chassis of the s.s.b. exciter, showing i.f. transformer T_4 in Fig. 1.



» The half-lattice crystal filter is quite similar to the full-lattice filter, but it uses only half as many crystals and thus is quite attractive for amateur use.

Half-Lattice Crystal Filters

WILLIAM E. GOOD, W2CVI

A LATTICE-TYPE filter with quartz crystals can have a bandpass of twice the spacing between the series- and parallel-resonant frequencies of one of the crystals. In an X-cut crystal this separation may be 1.5 to 2 kc. at 465 kc., so that a bandpass of the order of 3 to 4 kc. should be possible. A circuit equivalent to the four-crystal lattice is the typical balanced crystal filter circuit with a second crystal substituted for the neutralizing or balancing condenser. This is illustrated in Fig. 1.

The crystals used in the following experiments were low-frequency crystals from the FT-241-A series, used in several filters described earlier. These crystals are in the range from 370 to 500 kc. The labels on the crystals in one group run from 20.0 to 27.9 Mc. in 0.1-Mc. steps — the fundamental frequency is found by dividing the label frequency by 54, which makes the frequency steps 1.85 kc. at the fundamental. These crystals are being sold at reasonable prices, and they are usually listed at the low-frequency value. Another group is labeled between 28 and 38 Mc., and the fundamental frequency is found by dividing the label frequency by 72.

The frequency separation of the crystals in the first group is just about what the theory says it should be for an optimum bandpass filter. The group of curves shown in Fig. 2 was obtained from an RME-69 with its crystal filter circuit modified as in Fig. 1B.

Bandpass Characteristics

The curve of Fig. 2A is the typical response of a single-crystal filter with *LC* resonated to the crystal frequency and C_1 just balancing the crystal capacitance. As C_1 is made smaller, a rejection notch (parallel resonant frequency) moves in from the high-frequency side. Conversely, as C_1 is made larger than the balance capacity, the notch moves in from the low-frequency side.

By replacing C_1 with X_2 (a crystal 1.8 kc. higher than X_1), the curve of Fig. 2B is obtained. The shunt capacitances of the two crystals balance each other and no rejection notches are noticed. The top of the curve should be flat if the filter is properly terminated and *LC* is tuned properly. However, in most cases a 3- to 6-db. dip was noticed between the two peaks. The peaks are separated by approximately 2 kc., and one may be slightly higher than the other. The skirts are about 10 kc. wide at 60 db. down. The dotted curve is the i.f. passband without the filter.

If a trimmer, C_2, is placed across the higher-frequency crystal, as shown in Fig. 2C, two rejection notches will appear, and they will move in toward the center frequency more or less symmetrically as C_2 is increased.

Fig. 2D is the same as Fig. 2C except that the value of C_2 is larger and it shows how the side responses tend to rise higher as the notches come closer together. The general result is that the sides of the response curve become steeper as C_2 is increased, without appreciably affecting the separation of the two peaks. Practical values for C_2 are around 1 or 2 $\mu\mu$f., obtained readily by twisting together two short pieces of insulated wire.

It should be emphasized that *in every case LC is tuned to resonance at the center of the passband.* If this is not done, there will be a very pronounced dip between the two peaks, and the value of the filter will be lost. Experimentally, *C* is varied until the minimum dip occurs at the center of the response curve.

If C_3 is placed across the lower-frequency crystal, as shown in Fig. 2E, the skirts widen out

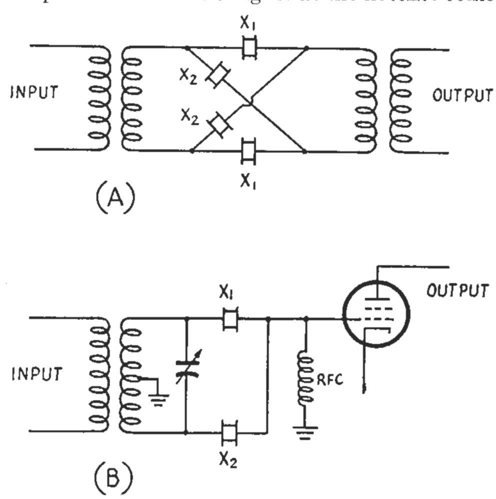


Fig. 1 — **A conventional crystal lattice filter is shown at A. The bandwidth is determined by the frequency separation between the crystals X_1 and X_2.**
The circuit in B behaves the same as that in A but requires only two crystals. The circuit is similar to a single-crystal filter, such as is used in a communications receiver, with the phasing condenser replaced by the second crystal.

From "A Crystal Filter for 'Phone Reception," *QST*, October, 1951.

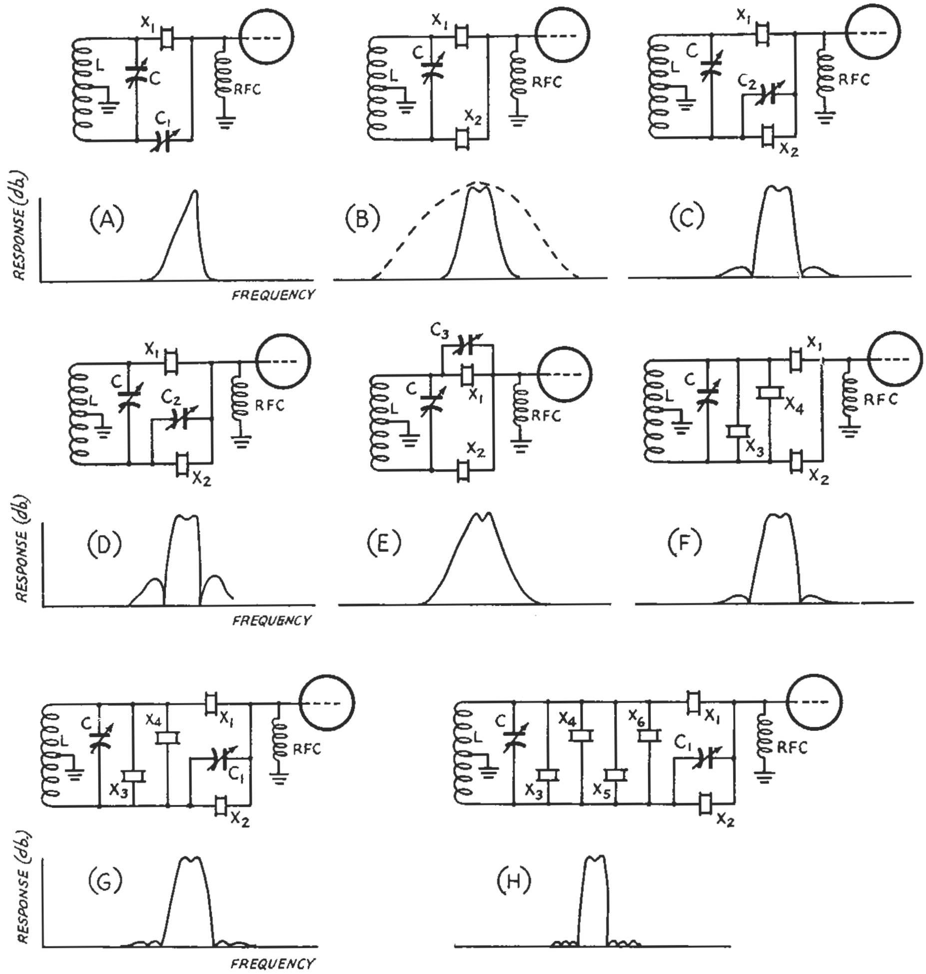


Fig. 2 — Typical response curves for the various crystal-filter circuits. See text for discussion and values of L, C, C_1, C_2, and C_3

X_1 — 464.81 kc. (marked 25.1 Mc.).
X_2 — 466.67 kc. (25.2 Mc.).
X_3 — 462.96 kc. (25.0 Mc.).
X_4 — 468.52 kc. (25.3 Mc.).
X_5 — 461.11 kc. (24.9 Mc.).
X_6 — 470.37 kc. (25.4 Mc.).

and the dip becomes more pronounced. This is undesirable from a selectivity point of view but important to know for the experimenter who intends to work with this device.

When two crystals, higher and lower in frequency than X_2 and X_1, are placed in shunt across the circuit of Fig. 2B, the result is as shown at Fig. 2F. Their capacitance is compensated for by reducing C slightly. The two notches appear at the series resonant frequencies of the two new crystals. Essentially, they are series-resonant traps shorting out LC at their resonant frequencies. The effect, that of steepening the sides of the response curve, is similar to that obtained in Fig. 2C. However, these notches will stay put and are not subject to variations like those obtained by tuning adjustments.

If C_2 is added to X_2 in the combination shown in Fig. 2F, the original pair of notches may be introduced and their frequencies set to reduce the size of the extra humps of F, as shown in Fig. 2G. C_2 may be increased to bring the new notches inside those caused by X_3 and X_4. This will give steeper sides, at the expense of somewhat higher side lobes, and the result is a curve about 4 kc. wide at 60 db. down and 2 kc. wide at the top. By adding still more crystals in shunt, the side lobes can be reduced still further, as shown in Fig. 2H.

Figs. 3 and 4 show typical curves made on an RME-69 receiver that had its crystal circuit modified. The readings were made with the S-meter, using a harmonic from a 100-kc. oscillator as the signal source, and the calibrated band-spread dial for frequency indications. The band-spread on the receiver was calibrated by using the series-resonant frequency of the crystals themselves in the circuit of Fig. 2A.

Some Practical Considerations

The i.f. transformer in the RME-69 has a center-tapped secondary (L in Fig. 2). However, there is no reason to believe that a straight secondary with two condensers (capacity divider) wouldn't be just as effective, and it would certainly be more convenient, particularly where the receiver does not have a crystal filter to start with. As for proper termination of the filter, rough calculation shows that it should be around 80,000 ohms. In practice, it doesn't seem to be too critical. The 16-mh. r.f. choke that was already in the RME-69 seemed to be satisfactory, as did 50,000- and 100,000-ohm resistors that were substituted. A tuned circuit in place of the choke did not hold down the side lobes unless it was loaded down or detuned. Presumably a fine adjustment on the center tap of L would have permitted a better balance between the two peaks. Perhaps more careful adjustment of the driving and terminating impedances would have reduced the dips between the two peaks.

Rather than use half a dozen (or more) crystals as in Fig. 2H, it is more desirable to cascade two crystal filters using either two or four crystals each. In other words, make another more or less identical crystal filter in the next i.f. stage and perhaps stagger the rejection notches for the best side-lobe reduction. This will steepen the sides and reduce the skirts, without appreciably affecting the nose of the response curve.

To obtain still steeper sides, it seems feasible to move X_3 and X_4 of Fig. 2F close to the center frequency. This could be done (with the present crystals) by further grinding. Apparently these wire-mounted crystals were originally trimmed to final frequency by grinding the top edge with a piece of fine emery paper in the same way that one might file his fingernails. The edge of the crystal is the proper place to grind, because the vibration is a surface-shear mode, and removal of quartz from the edges will cause the frequency to increase. Some way of holding the crystal should be devised for this grinding operation, as the wire mounting is quite delicate.[1] The author ground the edge of one crystal for several minutes with some No. 1 emery paper and the frequency moved about 1 kc. This offers a possible method for moving these crystals around or making use of the ones that don't fall within the bandpass of the i.f. amplifier. Mr. Roy Lewis of the General Electric crystal laboratory recommends that the crystal be washed in alcohol after grinding and that, if the frequency is going to be moved more than a few per cent, the crystal should be ground equally on opposite edges. He also suggests that the frequency might be lowered by plating with silver or possibly copper.

The bandwidths shown seem to be adequate for reasonably good 'phone reception. When the signal is tuned in, the carrier is placed on one or the other of the two peaks, and then the reception is essentially single sideband. If there is interference on that sideband, the receiver is tuned to place the carrier on the other peak. The effect of tuning through a 'phone signal is perhaps more pronounced than with the usual Q5-er. As soon as the carrier is far enough up one side of the curve to permit demodulation (detection) without too much distortion, the passband has its maximum width for one sideband. As you tune through the signal, the modulation sounds more bassy as the carrier approaches the center of the passband, but it quickly returns to a more natural-sounding signal as the carrier reaches the other edge. With the filter of Fig. 2H, it is surprising to see how quickly an interfering heterodyne drops out as the receiver is tuned. It is also possible to tune the carrier quite far down on the side and then inject a local carrier (b.f.o.) tuned to zero beat with the carrier.

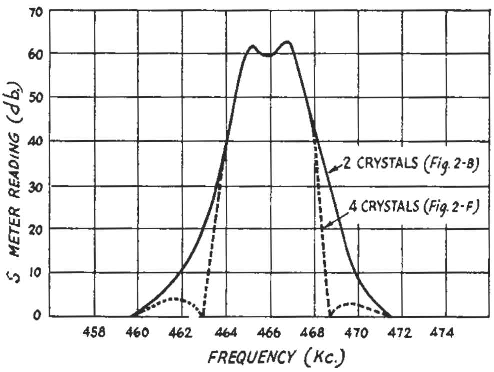


Fig. 3 — S-meter readings plotted against frequency, for the filters of Figs. 2B and 2F, made with an RME-69 receiver. Both of these filters give good voice reception, particularly if tuned for single-sideband reception.

This arrangement also works quite well for c.w. reception, giving good selectivity without the critical tuning of a sharp single-crystal filter. The hardened c.w. man would probably prefer two crystals much closer together in frequency. It should not be too difficult to switch between three crystals to give a 'phone or c.w. filter that would surpass the usual crystal filter. For c.w. the extra (off-frequency) crystal could be switched to the shunt position, to give better rejection.

Some crude tests were made with the two crystals two channels (3.7 kc.) apart and three channels (5.55 kc.) apart. The passbands were relatively wider; however, the center dip began to be very pronounced in the latter case.

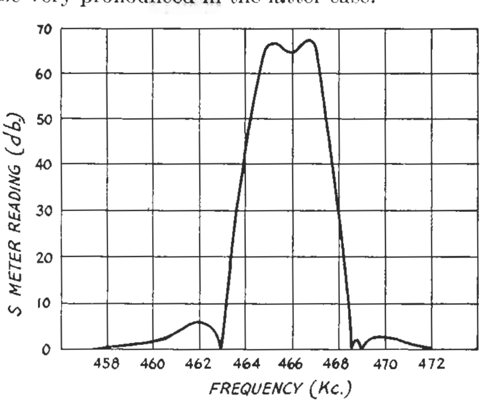


Fig. 4 — S-meter readings plotted against frequency for the filter of Fig. 2C, made with an RME-69 receiver. The value of C_2 was about 1 or 2 $\mu\mu$f.

[1] W1SCO uses a fly-tying vise to hold the crystal. — Ed.

» Practical examples of the half-lattice filter applied to s.s.b. exciters. Both units are also interesting for their balanced-modulator circuitry.

Exciters Using Cascaded Half-Lattice Crystal Filters

The s.s.b. rig used by Ken Stone, W7BMF, has several novel features in the exciter. For example, the balanced-modulator circuit (swiped from Motorola) doesn't require push-pull inputs of any kind, an advantage or convenience in many cases. As can be seen from Fig. 1, the carrier voltage is applied to the cathodes in parallel, and the modulating voltage is fed to one grid. A similar circuit is also used to convert the 450-kc. output of this unit to the operating frequency — VFO output is fed to the two cathodes and the 450-kc. s.s.b. signal goes to one grid.

The crystal filter uses adjacent-channel crystals (Y_1 and Y_2). Tests on a single section of the filter show about 35-db. rejection, and the two cascaded sections measure up around 55 db.

The final at W7BMF uses p.p. 807s, triode-connected with the grids and screens connected together and operated at zero bias. Plate voltage is 750.

W2JJC has an exciter design that is passed along by Fred Huff, W2AMB. It borrows ideas from several designs.

The circuit is given in Fig. 2. The transformer T_2 is opened in the primary for a series-tuned connection, since the varistor (four germanium diodes) modulator wants to look into a low impedance. The "bifilar" windings on L_1 are simply two interwoven windings, with the left-hand end of one winding connected to the right-hand end of the other. You can find a picture of such a winding in Fig. 5 of Mann's, "An Inexpensive Side-

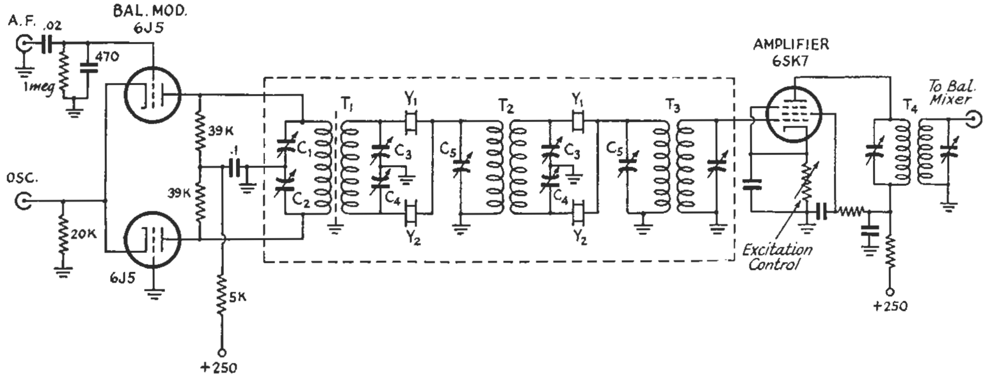


Fig. 1 — The balanced-modulator and crystal-filter circuit used by W7BMF. All transformers (T_1, T_2, T_3 and T_4) are standard types, with the phase-splitting condensers (C_1–C_5) added. These condensers are made of good quality mica in parallel with 50-μμf. air trimmers. The Faraday shield between the windings of T_1 is made by winding several turns of No. 18 d.c.c. between primary and secondary and grounding one end.

The filter was aligned with a signal generator, introduced at the audio-input terminal (oscillator turned off). The first step was to align all circuits to the mid-frequency between Y_1 and Y_2. The signal generator is then set to a frequency about 3 kc. higher (or lower), and C_3 and C_4 are adjusted simultaneously until a very sharp null is obtained. C_5 should be adjusted to the center of the passband, since improper adjustment will cause a large dip in response between the two crystal frequencies. It has been found that a 6-db. sag causes no impairment of voice quality, however, and it does improve the skirt selectivity. The adjustments may require several go-rounds, because they interlock slightly.

From *QST*, April, 1953, and June, 1953.

band Filter." The FT-241-A crystals do not need to be the exact channels shown — any combination of channels that gives the necessary 1.8-kc. separation should be satisfactory. Trimmers C_1 and C_2 across the high-frequency filter crystals are used to trim the shape of the filter characteristic and sharpen the rejection.

Provision for carrier reinsertion is included, by the use of the 6J5 cathode follower. The 20K potentiometer sets the level of carrier as desired.

The VFO signal can be tuned either 433 kc. (approximately) higher or lower than the desired output frequency. Using these crystals, or any giving a similar oscillator/filter relationship (lower sideband out of the filter), the VFO should be tuned 433 kc. lower for lower-sideband output.

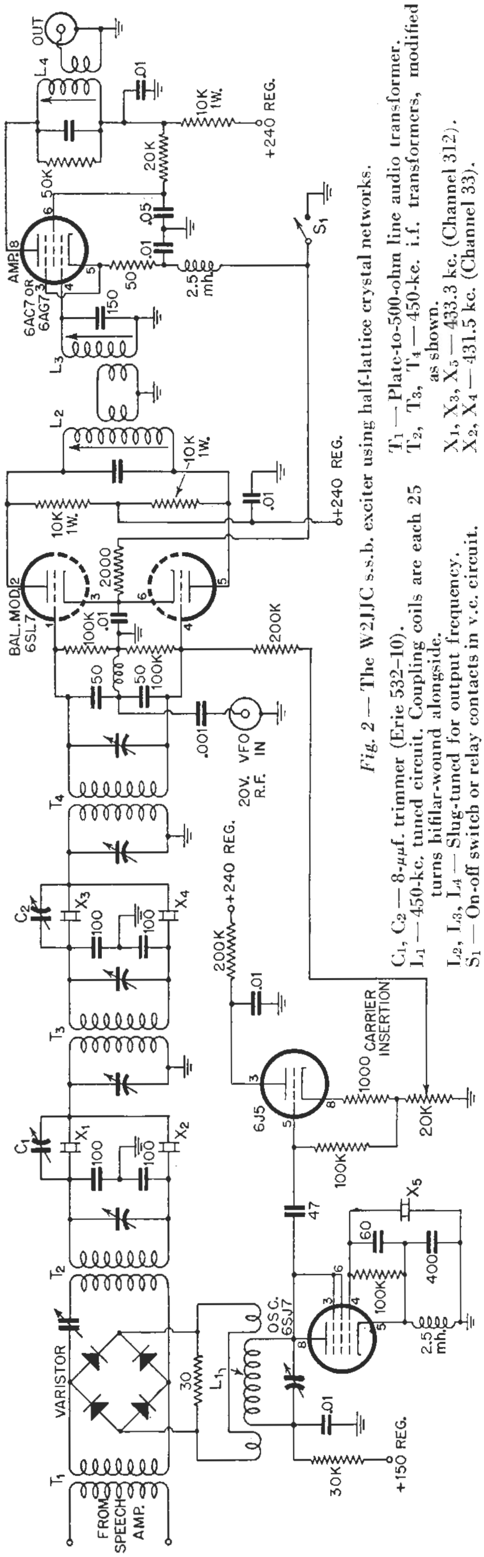


Fig. 2 — The W2JJC s.s.b. exciter using half-lattice crystal networks.

C_1, C_2 — 8-μμf. trimmer (Erie 532-10).
L_1 — 450-kc. tuned circuit. Coupling coils are each 25 turns bifilar-wound alongside.
L_2, L_3, L_4 — Slug-tuned for output frequency.
S_1 — On-off switch or relay contacts in v.c. circuit.
T_1 — Plate-to-500-ohm line audio transformer.
T_2, T_3, T_4 — 450-kc. i.f. transformers, modified as shown.
X_1, X_3, X_5 — 433.3 kc. (Channel 312).
X_2, X_4 — 431.5 kc. (Channel 33).

Audio Quality and Filter-Type S.S.B. Exciters

Assuming no significant distortion in the microphone or audio amplifier, the positioning of the carrier with respect to the sideband filter frequency, and the bandwidth of the filter, are the determining factors in the "quality" or sound of the signal. Moving the (suppressed) carrier frequency further away from the filter frequency will accentuate the "highs" in the signal and suppress the "lows," while moving the carrier closer to the filter frequency will reduce the highs and bring in the lows. The limit to this latter process, of course, depends upon the "steepness" of the side of the filter characteristic, since moving the carrier closer to the filter frequency will reduce the attenuation of the "lows" in the unwanted sideband. A reduction of low-frequency response through the speech amplifier will alleviate this latter condition somewhat and result in better sideband rejection at low audio frequencies.

Fortunately the higher audio frequencies contribute more to the intelligibility of a signal than do the low frequencies, so there is no real need to strive for excellent low-frequency response in an amateur 'phone signal. Low-frequency response in the speech amplifier can be decreased by reducing the capacity of interstage coupling capacitors.

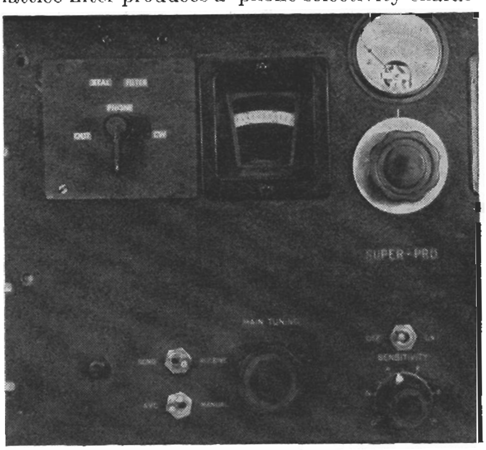
While checks with an audio oscillator are useful in determining just what is happening through the system, a practical test should always be made by the operator to determine the best-sounding conditions for any given filter. This will involve his speaking into the microphone and trying several different settings of the (suppressed) carrier frequency, while another operator is monitoring the signal on a low-selectivity receiver. The carrier frequency should be adjusted for maximum intelligibility. This will not involve too much compromise with "naturalness" unless the sideband-filter bandwidth is rather narrow, on the order of 2000 cycles or so. With narrow-bandwidth filters, small changes in the carrier positioning will have a marked effect on the "naturalness" of the output signal, and it will be worth while to spend some time in careful positioning of the carrier.

» The addition of a cascaded half-lattice filter is an inexpensive way to improve the selectivity of any receiver. Here are some suggestions and the alignment procedure.

Cascaded Half-Lattice Crystal Filters for 'Phone Reception

HOWARD L. MORRISON, W7ESM

The improvement in receiver selectivity obtained by a single half-lattice crystal filter, together with the published selectivity characteristics of some of the newer commercial receivers, were incentives for trying a cascade arrangement in the main receiver at W7ESM — a prewar Super-Pro. Despite its age this model of the Pro is an effective ham receiver without any modifications, as long-time owners or those who have acquired them as war-surplus items know. Its variable coupling in the i.f. amplifier provides good skirt selectivity in the minimum-bandwidth position. Consequently, even though the installation of a two-stage half-lattice filter produces a 'phone selectivity characteristic which compares favorably with those obtained by mechanical filters or 50-kc. i.f. systems, the improvement in this particular receiver is not as startling as might at first be expected. However, there is a noticeable improvement in readability when the QRM is really rough. In the absence of QRM, better readability is often obtained on a.m. signals with the filter switched out, however. On c.w. both skirts of the selectivity curve are steep, and the operation is more "single-signal" than that of conventional crystal filters. The filter circuit herein described can be applied to other receivers, possibly with more beneficial effect.



Front panel of the receiver with the new filter arrangement replacing the original crystal filter. The switch mounting plate is slightly larger than the former one, but the four screws are in their original positions. The original bandwidth control in the lower left-hand corner has been "blacked out" since it has been fixed in position during the alignment process.

Mounting

The photos show how the filters squeeze into the space occupied by the original filter. There are two cascaded filters, one for 'phone with a nose bandwidth of 3 kc., and one for c.w., considerably sharper. Two crystals are common to both 'phone and c.w. filters so that six crystals do the work of eight. A three-position switch selects either filter, or straight-through operation. Even though the spacing of the two gangs of the switch was cut down, the mounting plate extended about an eighth inch beyond the front panel, so a second plate, with a hole in it large enough to recess the front of the switch, was used to fill the gap. A thin piece of aluminum shields the two sections of the switch from each other and is also sandwiched in between the two small boxes which contain and shield the three crystals associated with each of the two i.f. stages. A small angle is bent along the bottom edge of this piece so as to shield the leads from the first i.f. transformer to the first switch section from the wiring of the second switch section immediately above. This amount of shielding might seem scarcely enough, but no amplifier instability or adverse effects on the selectivity characteristic were noted.

Circuit

The basic filter circuit for each stage is that of Bill Good, W2CVI. The inductance in the Pro i.f. transformers is enough to allow the low-*C* tuned circuits which are necessary for half-lattice filters. The complete circuit is shown in Fig. 1. Balanced-to-ground i.f. transformer secondaries are obtained by sawing off all but three rotor plates of the original tuning condensers and adding additional condensers as shown.

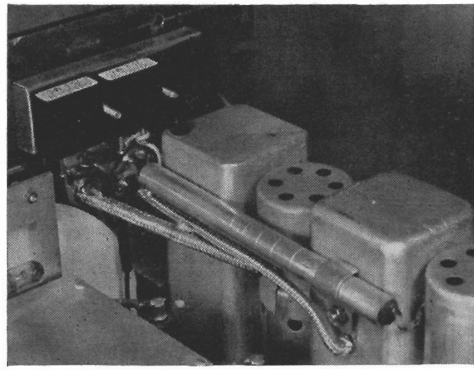
Early tests revealed that the i.f. tank circuits would drift out of alignment as the set warmed up — an effect evidenced by the flat top of the selectivity curve becoming tilted. The combination of negative and zero temperature coefficient

From *QST*, May, 1954.

tank condensers specified in Fig. 1 sufficiently minimizes this effect for practical purposes. The purist might want to add more negative coefficient condensers to all of the i.f. tank circuits, since the drift in alignment is due to the all-brass variable condensers used in this model of the Pro. However, getting inside the i.f. transformers and reassembling them, especially those with variable coupling, is no fun.

The original first i.f. transformer can be used by removing the inner shield can which contains the primary circuit, and relocating both primary and secondary coils by fastening their mounting boards to the outer edge of their respective ceramic supports. The latter procedure increases the mutual inductance and also permits 180-degree rotation of the trimmers. The link-coupling coils associated with the original crystal filter circuit can be left in place, but must be completely disconnected. There is ample room for the additional mica and ceramic condensers on the coil boards. The second i.f. transformer has only the secondary (upper) coil remounted as just described. (A tip on reassembling this transformer: Wrap several turns of string around the upper end of the guide rod so as to hold back the long compression spring until the rod is again in place; the string then can be pulled out with a little persuasion.) While both i.f. transformers are apart it will make future alignment easier by marking the trimmer adjustment nuts in some way so as to indicate the position of the rotor with respect to the stator.

The shielded pair leading to the filter from the second i.f. transformer, along with the low-capacity grid leads for the first and second i.f. tubes, can be seen in the back-view photo. Up to 10 $\mu\mu$f. capacity in the grid leads doesn't do much to the filter except lower the gain; larger capacity will lower the terminating impedance of the filter, and may affect the selectivity characteristic. Auto-antenna lead-in cable is very practical for this use, but since none was at hand "home-brew" low-capacity leads were used. Ordinary shielded wire or the RG types of coax cable are not suitable in this case.



Rear view with the aluminum box which contains the three crystals in the second stage removed. The second switch section is directly beneath the middle crystal, which is the second *B* crystal. The loop of wire around one prong of this crystal is the trimming capacity for the second stage. The lowest shielded lead comes through the plate which shields the two switch sections, and goes to the grid of the first i.f. tube; it consists of large-diameter varnished tubing with a small-sized hook-up wire inside to keep the capacity low. The center shielded lead is for the connections to the secondary of the second i.f. transformer. The upper "pipe" is a low-capacity lead to the grid of the second i.f. transformer, and has its center conductor supported by a ceramic bead at each end. Auto-antenna lead-in cable would be satisfactory for the two low-capacity leads. The second-i.f. grid resistor, partially covered with varnished tubing, is visible near the dial.

Two-pole switches must be used in each filter stage in order to disconnect completely the unused crystal and prevent irregularities in the selectivity characteristic.

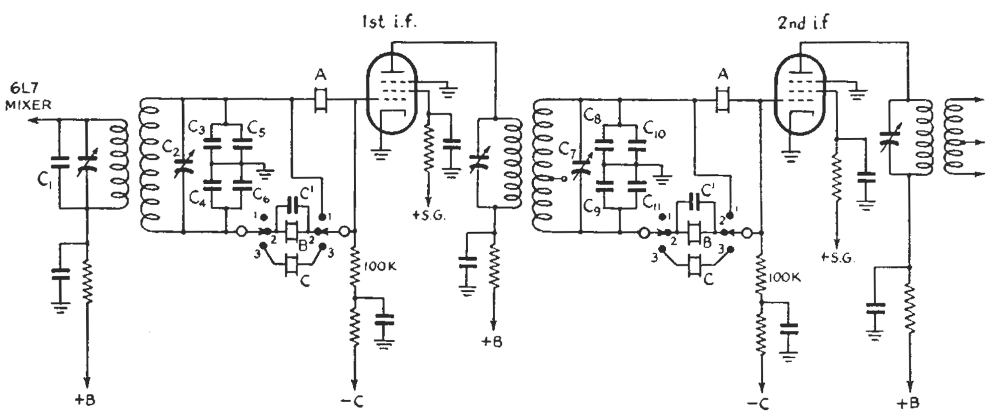


Fig. 1 **— Circuit of cascaded half-lattice crystal filters as applied to the i.f. amplifier of the SP-110X. Unmarked components as in original receiver.**

A, B, C — See Table I.
C_1 — 25-$\mu\mu$f. ceramic, negative temp. coeff. (N750 type).
C_2, C_7 — Original condensers with all but 3 rotor plates removed.
C_3, C_4, C_8, C_9 — 47-$\mu\mu$f. ceramic, negative temp. coeff. (N750), 5 per cent tolerance.
C_5, C_6, C_{10}, C_{11} — 100-$\mu\mu$f. zero temp. coeff. ceramic (NPO), or silver mica, 5 per cent tolerance.
C' — Crystal trimming capacity. See text.
Switch — 2-gang with 2 poles and 3 positions per gang.
Switch position 1: filter out
" " 2: 'phone
" " 3: c.w.

Crystal Trimming Capacity

With filters in cascade, crystal trimming capacity is not as critical as with a single stage. This can be seen by taking the example of a half-lattice filter with enough excess capacity across the higher frequency crystal to produce side lobes only 30 db. down from reference level — a condition easily brought about by an extra $\mu\mu$f. or two. If two such filters are cascaded the side lobes would be 60 db. down, because the second stage responds 30 db. less to the side lobes produced by the first stage than it does to the first stage's peak. Hence in cascaded filters we can use relatively large trimming capacity to produce steeper skirts without introducing large side lobes as in the case for a single-stage filter. Trimming capacity is used only in the 'phone filters, as indicated in the diagram, and consists of one loop of hook-up wire in each stage as shown in the photo.

Crystals

FT-241-A-series crystals are used, each whose labeled frequency is the 72nd harmonic of the crystal frequency and can be directly used in the 'phone filters because alternate channels are 2.8 kc. apart; the selectivity curve will then be about 0.3 kc. wider than that shown in Fig. 2. Crystals in the 54th-harmonic series can be used, but must be shifted by plating or grinding so that they are at least 2.5 kc. apart; the nominal 1.9-kc. spacing of adjacent channels in this series is too narrow for a satisfactory 'phone bandwidth when cascaded filters are used, and the 3.8-kc. spacing of alternate channels is too wide. However, crystals in this series can be used in conjunction with those of the former, as shown in the first arrangement of Table I, to provide reasonable bandwidths for both 'phone and c.w. without requiring any crystal modification. This particular set of crystals also brackets the 465-kc. receiver i.f. very nicely; if a set of crystals is used which has the desired frequency separations, but does not bracket the original i.f., it will be necessary to touch up the r.f. alignment if the calibration of the main tuning dial is to be held. Realignment of the head end is not a bad idea anyway, especially in the case of a receiver which has been in service a good many years. The arrangements listed in Table I are representative only.

Alignment

The crystals are measured and modified as may be necessary, according to methods detailed in the next article. Since there is little difference in aligning either a one- or two-stage half-lattice filter, the procedure and test equipment also described in that article can be used. In this case make the following preliminary settings: filter switch to 'phone position, bandwidth control (variable i.f. coupling) to *3.7*, a.v.c. switch to MANUAL, b.f.o. switch to MOD, SENSITIVITY (r.f. gain) between *3* and *4*, bandswitch to one of the higher-frequency bands, audio gain to zero, h.f. oscillator tube out, high-impedance d.c. voltmeter (v.t.v.m. on 10-volt scale, or 20,000 ohms/volt meter on 2.5-volt scale) across second-detector load resistor (PHONO terminals), and signal generator midway between the series-resonant frequencies of crystals *A* and *B*. Clip the signal generator through a blocking condenser to the grid of the third i.f. tube and adjust the trimmers of the fourth and fifth i.f. transformers for maximum output. Next clip

TABLE I

Some Possible Crystal Arrangements

Crystals A Channel	Kc.	*Crystals B* Channel	Kc.	*Crystals C* Channel	Kc.	*B − A* Kc.	*'Phone bandwidth at 6 db. down*	*C − A* Kc.	*C. W. bandwidth (estimated)*
1. 334	463.9	336	466.7	51	464.8	2.8	3.3 kc.	0.9	800 cycles
2. 334	463.9	336	466.7	51 shifted 0.4 kc. higher than crystal *A*		2.8	3.3 kc.	0.4	150 cycles
3. 333	462.5	335	465.3	50	462.9	2.8	3.3 kc.	0.4	150 cycles
4. 335 shifted 1.4 kc. lower	465.3	337 shifted 2.5 kc. higher than crystal *A*	468.0	51	464.8	2.5	3.0 kc.	0.9	800 cycles
5. 51 shifted 0.6 kc. lower	464.8	52	466.7	51	464.8	2.5	3.0 kc.	0.6	400 cycles

Crystal frequencies are nominal, and are for parallel resonance (labeled frequency divided by 72 for the 300-series channels, and by 54 for the 50-series channels). The series-resonant frequencies of the actual crystals used must be measured, and the difference frequencies, *B* − *A* and *C* − *A*, must be as listed above in order to obtain the listed bandwidths.

If the 110-cycle c.w. bandwidth shown in Fig. 1 is desired, *C* − *A* should be 0.3 kc.

the generator to the grid cap of the second i.f. tube, adjusting the third i.f. transformer, and so on until the generator is connected to the grid of the 6L7 mixer, and the first i.f. transformer aligned. Then tune the generator between 460 and 470 kc. and note the two peaks of maximum output; the frequency exactly midway between them is the final alignment frequency, and all the trimmers in the amplifier should be touched up with the generator at this frequency and connected to the grid of the 6L7. Go over the entire series two or three times and again check the two peaks; they should match within a few per cent, and the dip between should not be down more than 20 per cent of peak. In Fig. 2 the dip is 15 per cent down, or 1.5 db. If the dip is greater than 20 per cent, reduce the i.f. transformer mutual inductance by turning the bandwidth control counterclockwise a little, and repeak the amplifier. If there is little dip, with the peaks scarcely noticeable, so as to make a sharp-nosed selectivity curve, increase the mutual inductance. (If a receiver with fixed i.f. coupling is being modified, mutual inductance can be reduced by sliding the coils apart, or by a shorted turn of heavy copper wire around the form midway between them. If the transformer has tank capacity greater than 100 $\mu\mu$f., mutual coupling can be reduced along with the *C/L* ratio by means of an additional coil in series with each secondary.) If one of the peaks is greater than the other, the response can be leveled by slightly readjusting some of the trimmers so as to favor the lower peak. For example, if the low-frequency peak is down, some of the trimmers can be set a little higher in capacity than the setting which provides maximum output at the alignment frequency. The final alignment adjustments should be made only after the set has had plenty of time to reach normal operating temperature.

The side lobes and nulls should be checked next. If the lobes are greater than those illustrated in Fig. 2, the crystal trimming capacity is too large; if the lobes are smaller, or even nonexistent, but the skirts of the main curve wider, the value of the trimming capacity should be increased.

When the desired selectivity characteristic is obtained it is a good idea to fasten the bandwidth control permanently in position. A short length of No. 12 copper wire soldered to the brass cam nearest the front panel and bolted to the chassis accomplishes this readily.

Insertion Loss and Audio Considerations

With the filter switched out, the i.f. gain is somewhat greater than in the original receiver because the grids of the first two i.f. tubes are effectively tapped across a greater portion of the resonant circuits. In the 'phone position the loss is about 25 db. with respect to the *out* position, and in the c.w. position about 10 db. Gain should come after, and not before, the selectivity circuits in order to prevent overloading and cross-modulation, which undo the good work of these circuits. Some increase in i.f. gain is had by shorting out the 50,000-ohm resistor which is in series with the screen-voltage supply lead to the first three i.f. tubes, but most of the loss is more easily made up in the audio section. There are several possibilities: A 6J5 can be substituted for the triode-connected 6F6 audio driver by merely reconnecting the grid resistor of this stage to the same bias-supply tap which feeds the first audio tube. An alternate scheme is to use a 6SN7 dual triode for the first two audio stages, and take advantage of the leftover tube socket to install a noise limiter. Still more gain can be had by connecting the 6F6 audio output tubes as pentodes.

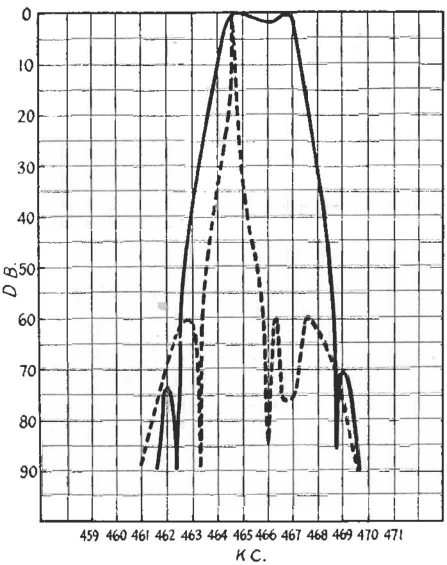


Fig. 2 — The solid curve shows the i.f. selectivity with the filter switched to the 'phone position. The bandwidth is 3.0 kc. at 6 db. down, and 6.1 kc. at 70 db. down. The dotted curve is for the c.w. position, and is 110 cycles wide at 6 db. down, 370 cycles at 20 db., and 1320 cycles at 40 db.

Since the 'phone filter limits the high-frequency audio response to 3000 cycles, some attenuation of low audio frequencies is desirable in order to restore a reasonable proportion of highs and lows, and to prevent signals from sounding mushy. One of the original 0.05-μf. audio coupling condensers should be replaced with 0.001 μf. Since the original variable i.f. bandwidth feature is now gone, the set no longer has enough bandwidth, even with the filter out, to serve as a hi-fi b.c. receiver, anyway. It is FB for communication work, though.

Several European amateurs have used high-frequency crystal filters — around 5 Mc. — in their s.s.b. generators, but the system has apparently never been tried in the U. S.

» For the reception of 'phone signals, even a single half-lattice filter section is superior to the simple single-crystal filter. And here is how to "move" crystal frequencies slightly.

Alignment of Half-Lattice Filters

HOWARD L. MORRISON, W7ESM

A BC-312 was modified by adding a half-lattice filter of the type described by W2CVI, since crystals in the neighborhood of 470 kc. were available. These are the FT-241-A series, labeled according to channel number and transmitter (not crystal) frequency. Of crystals in the group whose labeled frequency is the 54th harmonic of the crystal frequency, Channels 53

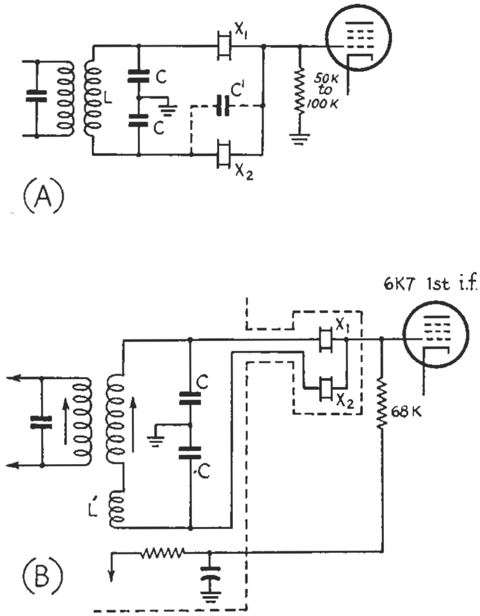


Fig. 1 — A — Basic dual-crystal filter circuit. The trimmer, C', is discussed in the text.

B — Basic circuit applied to the BC-312. Unmarked components are the same as in the original receiver. L' is discussed in the text. Condensers marked C should be zero-temperature coefficient ceramic or silver mica. Values between 150 and 200 $\mu\mu$f. are satisfactory, but both should have the same capacitance.

(468.5 kc.) and 54 (470.4 kc.) are closest to the 470-kc. i.f. of the BC-312. Channels 54 and 55 (472.2 kc.) are also suitable. Of crystals in the group whose labeled frequency is the 72nd harmonic, Channels 338 (469.4 kc.) and 340 (472.2 kc.) are the most suitable, but will provide a wider bandwidth, since their frequencies are 2.8 kc. apart as compared to the 1.9-kc. separation in the first group. Means of changing the crystal frequency by plating and grinding methods will be discussed in detail further on, and make it possible to utilize crystals other than those in the desired channels as well as to obtain the desired bandwidth.

Modification Details

Since all BC-312 type war-surplus receivers have a first-i.f. transformer with either center-tapped secondary or split tank capacity, it seemed at first that the dual-crystal filter circuit, Fig. 1, could be applied with very little trouble, and a pair of crystals was installed alongside the i.f. shield can.

The whole i.f. amplifier was aligned exactly on a frequency midway between the two crystal frequencies, but the double-humped selectivity curve shown dashed in Fig. 3 was the best that could be obtained. Such a pronounced dip between the peaks will destroy the value of the filter. Separation of stations was greatly improved, but the resulting harsh voice quality was very undesirable. After a considerable amount of experimenting it was found that both the L/C ratio of the tuned circuit which feeds the crystals and the degree of coupling between it and the primary affect the selectivity curve considerably. The i.f. tank circuits in these receivers have 400 $\mu\mu$f. capacity. When the capacity across the first secondary was reduced to 100 $\mu\mu$f. by making each of the condensers (C in Fig. 1) 200 $\mu\mu$f., with an additional coil placed in series with L so that the resonant frequency was adjustable to 469.8 kc., the solid curve in Fig. 2 resulted.

The coil can be either dielectric core and adjusted by peeling off turns, or slug tuned. A fixed b.f.o. coil from the junk box was used here, although a slug-tuned one would have obviated the tedious process of peeling turns a few at a time. The coil was mounted adjacent to the main secondary inside the i.f. can, and equidistant from the wider sides. Since it has some mutual coupling with the main secondary, connecting it one way will give a greater total inductance than the other; the connection which gives the greater inductance (as indicated by a lower resonant frequency) should be used in order to realize the greatest Q. For the same reason, a coil so large that it comes closer than a quarter inch to any side of the can should not be used

From "'Phone Selectivity for the BC-312," *QST*, February, 1954.

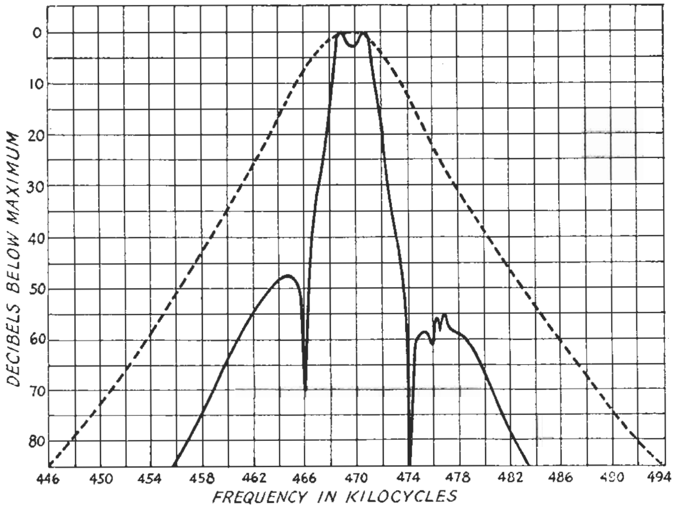


Fig. 2 — The dashed curve is the straight i.f. selectivity of the unmodified BC-312. The solid curve is the selectivity with the dual-crystal filter installed as described. The latter is 2.7 kc. wide at 3 db. down and 6.5 kc. wide at 40 db. down.

Shunt Crystals

The method of increasing skirt selectivity by crystals shunted across the tuned circuit ahead of series crystals did not seem very effective when tried on two different receivers. The skirts became a little steeper, but new side lobes appeared at the antiresonant frequencies of the shunt crystals, and were sometimes only 40 db. down. Perhaps other shunting crystals with resonant frequencies equal to the antiresonant frequencies of the crystals closest to the passband could be used, but the whole affair becomes rather cumbersome for the improvement obtained. More selectivity than that provided by a single-stage filter is best obtained by adding another stage.

without cutting a one-inch hole in the latter. This procedure is necessary for most any coil if a BC-312N is modified, because this model has a smaller i.f. can. If a slug-tuned coil is used, the slug should be insulated from the can so as to minimize additional unbalanced capacity to ground.

Trimmer Condenser Considerations

If the physical arrangement and the construction details mentioned above are followed the trimmer C', Fig. 1A, will not be required to obtain the best selectivity characteristic. Fig. 3 shows the effect of trimmer capacity across the crystals. If a different physical layout is used, it may be necessary to add a slight amount of trimming capacity across one or the other of the crystals. W2CVI, for example, used between one and two $\mu\mu$f. across the high-frequency crystal to obtain the best characteristic. Just how slight the trimmer capacity need be to cause large changes in skirt shape is shown by the solid curve of Fig. 3. The "trimmer" consisted of the inner conductor and polyethylene insulation of a short length of RG-59/U coaxial cable running at *right angles* to the pins of the low-frequency crystal, one end of the wire being soldered to one pin, with the insulated end lying across the other pin and extending beyond it for only 1/16 inch. This probably represents a capacity of less than ½ $\mu\mu$f.

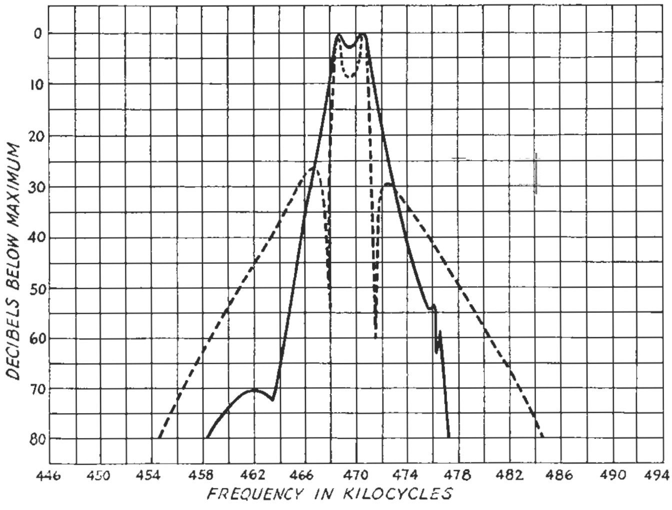


Fig. 3 — Effect of trimmer capacitance on shape of selectivity curve. Dashed curve: i.f. selectivity obtained with dual-crystal filter and original L/C ratio in i.f. transformer secondary (C in Fig. 1 equal to 800 $\mu\mu$f.), with 2-$\mu\mu$f. trimmer at C', Fig. 1A, across high-frequency crystal. Solid curve: same conditions as solid curve in Fig. 2, but with very small C' (less than ½ $\mu\mu$f.) across low-frequency crystal.

Alignment

The alignment procedure is easy if you have a test oscillator or signal generator with plenty of bandspread in the neighborhood of 470 kc. A BC-221 or LM-7 frequency meter is ideal, but the average serviceman's oscillator is out. It is not very difficult, however, to make a test oscillator for the occasion: A standard circuit taken from *The Radio Amateur's Handbook* and built around a tuned circuit from an old b.c. set is readily made. Such an oscillator should be padded to 470 kc. by an air, silver-mica, or ceramic condenser, together with a 25-$\mu\mu$f. bandspread tuning condenser that has semicircular (straight-line capacity) rotor plates. A BC-221 was used here; connections are shown in Fig. 4. For the home-built oscillator, the outer terminals of the potentiometer can be connected to a one-

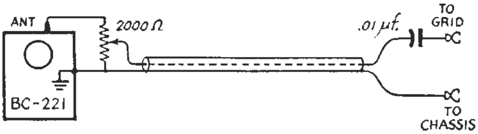


Fig. 4 — Using the BC-221 frequency meter as a signal generator for filter and i.f. alignment. If a 221 is not available, any oscillator covering the necessary frequency range with adequate bandspread may be used, as described in the text.

or two-turn pick-up coil loosely coupled to the oscillator tank. Since the bandspread tuning condenser represents only a small fraction of the total tank capacity, two or three calibration points (obtained by beating a harmonic of the oscillator with local b.c. stations) are sufficient to determine the straight-line calibration curve for the oscillator.

A sensitive (20,000 ohms/volt or a v.t.v.m.) d.c. meter is necessary for measuring i.f. amplifier output. It is connected across the diode load resistance of the receiver, and should be *kept* on the lowest scale in the case of a 20,000 ohms/volt meter, or on the 10-volt scale of a v.t.v.m., to prevent overloading the i.f. amplifier. The pointer can be reset to zero with the receiver turned on, in order to buck out the small voltage developed by the diode emission current. The h.f. oscillator tube in the receiver should be removed, the b.f.o. turned off, the manual gain control used, and the bandswitch set to one of the higher frequencies during the course of the alignment. The receiver manual volume control and test oscillator output always are adjusted so as to keep all readings on the same scale of the output meter.

Before installing the crystals in the set, their series-resonant frequencies should be measured by connecting them one at a time in series with the lead from the test oscillator to the 6K7 grid, and isolating this junction from the tuned circuit by a 50,000-ohm resistor. As the oscillator is tuned from lower to higher frequency, a sharp rise in the output meter reading will occur at the series-resonant frequency of the crystal, followed by a dip which indicates the antiresonant point. This test can also be used to check roughly the Q of the crystals; the amplitudes of peaks and dips should be the same for both crystals, and the ratio of peak to dip about 50 to 1. A defective crystal is indicated by a small ratio of peak to dip, and such should not be unexpected among quantities of crystals at bargain prices. Crystals which are stamped "Limited Test" are not necessarily low Q, however.

With the test oscillator still connected to the 6K7 grid, set it midway between the crystal series-resonant frequencies and align the 2nd and 3rd i.f. transformers for maximum output. Modify the 1st i.f. transformer as shown in Fig. 1B with 200-μμf. condensers (as low as 150 μμf. can be used) and a trial series coil, and reconnect the original grid lead to the 1st i.f. 6K7. Then clip the test oscillator to the grid lead of the 6L7 mixer and adjust or prune the series coil until peak response is obtained at the alignment frequency with the core in the main secondary coil about half in. Caution! Do not use this core as a means of determining whether the secondary circuit is higher or lower than the alignment frequency; it will give misleading information because an increase in output which may result from screwing in the core can be due to increased coupling to the primary and not to lowering the resonant frequency to bring it nearer the alignment frequency, as might be thought. The best indication of which direction to head with the series coil is had by using a small (around 10 μμf.) test condenser. If the output decreases when this condenser is touched between the 6K7 grid and ground, the series coil has too much inductance, and vice versa.

Next, connect up the two crystals as in Fig. 1B, and with the test oscillator on the alignment frequency, peak both primary and secondary of the 1st i.f. transformer. Then tune the test oscillator between 460 and 480 kc. and note the two maximum-response peaks. The frequency exactly midway between them is the final alignment frequency, and the entire i.f. amplifier should be carefully tuned for maximum output at this frequency. Again check the two peaks; they should now be equal within a few per cent, and the depression between should be around 70 per cent of the peak. If one peak is noticeably less than the other, the cores can be *very slightly* readjusted so as to favor it.

Measurement

The following method is recommended for finding the side-lobe response and the final selectivity characteristic:

With the test oscillator set on one of the peaks, and the manual volume control at about 2.5, adjust the test oscillator output until the output meter reads at some division near full scale which can be taken as "10." (For example, on a 20,000 ohms/volt meter, the 2.5-volt meter range is used, but the 5-volt scale multiplied by 2 is *read*.) Since decibels are obtained from voltage *ratios*, the actual voltage measured is unimportant so long as all voltages are measured proportionately. Tune the oscillator from the peak until the meter reads "1" on the scale of "10." This will be the 20-db. point. Leaving the test oscillator output fixed, increase the receiver's manual gain until the meter again reads "10"; then tune the oscillator until the meter reads "1." The output is now 100 times down, which is the 40-db. point. Again increase the manual gain until the meter reads "10." This is the range in which the side lobes can be measured. Full scale ("10") corresponds to 40 db.; "1" corresponds to 60 db., and "0.1" to 80 db.

Intermediate values are calculated according to the formula

$$\text{db.} = 20 \log_{10} \frac{e_1}{e_2}$$

in which e_1 is taken as 1000 if full-scale e_2 is 10, because we are two decades (100 times) down from our original starting point.

If any side lobe is greater than "5" on this scale (46 db.), a slight amount of trimmer capacity should be added across the low-frequency crystal; if the main nulls are more than 8 kc. apart, trimmer capacity should be added to the high-frequency crystal.

Moving the Crystals Around

It may happen that the two crystals to be used have resonant frequencies less than the nominal 1.9 kc. apart due to manufacturing tolerances. It can also happen that the only crystals available are not in the vicinity of 470 kc. Two procedures can be used to remedy these situations: plating and edge grinding.

Crystals can be lowered several kc. by plating them. However, it was found here that whenever a simple copper sulfate solution was used, a "black deposit" was formed regardless of the concentration of the solution. The addition of a small amount of sulfuric acid and alcohol cleared up the trouble. It is recommended that before any crystal plating be attempted, the process be tried out on less expensive objects such as alligator clips! The sulfuric acid and alcohol are added until a copper-colored plating is formed. A standard formula is 15 gm. of copper sulfate, 5 cc. of sulfuric acid, and 5 cc. of alcohol in 100 cc. of distilled water.

In all cases the electroplating scheme was found superior to merely dipping the object in the solution. Put the solution in a glass tumbler, bend a piece of clean No. 12 copper wire so that it clamps the edge of the tumbler and extends the depth of the solution, and connect it in series with a 330-ohm resistor (not critical) to the *positive* terminal of a 1.5 volt dry cell or flashlight cell. Better control of the frequency change is had by thus limiting the plating current, and the plating is more uniform. Without the resistor, repeated "dunkings" will cause a noticeably thicker plating on that part of the crystal which enters the solution first and leaves it last. The object to be plated is connected to the negative terminal. When plating crystals connect both pins in parallel. The crystal frequency is lowered according to the amount of plating, and changes up to 2 kc. can be obtained without seriously lowering the Q. Because of differences in solutions no time *vs.* frequency-change data are given, but a preliminary short immersion and frequency check will provide a basis for estimating the total time needed. The nice thing about plating is that if you go too far you merely reverse the battery polarity and take off some of the plating. With a few trials, a crystal can be jockeyed around to just where you want it. After removing the crystal from the plating solution it is important to rinse it in clean water and dry it completely before making any measurements.

With tough fingers and a big supply of patience a crystal can be raised 15 kc. or more by edge grinding. For small frequency changes, grinding the upper edge alone is sufficient; but for changes greater than a kilocycle it is necessary to grind all four edges equally and squarely. If grinding is attempted, buy three or four extra off-frequency crystals to practice on and acquire the knack! Before any grinding can be done it is necessary to remove the crystal from the supporting wires: Fasten an octal socket to the bench; bring a lamp near, and have several different-sized blocks available for hand rests. A sheet of metal or asbestos is used so that the barrel of the soldering iron can be solidly rested. Having carefully pried off the bakelite holder cover, plug the crystal in the bench socket and apply the tip of a well-tinned iron to one junction of crystal wire and holder wire, and with tweezers hold the latter away from the former until both have cooled. *Every effort must be made to avoid straining the crystal wires;* once they come loose from the crystal you are, for all practical purposes, through with that crystal. Next, select a solid hand rest, and with the tweezers hold the crystal carefully but firmly; apply the soldering iron to the opposite junction and lift away the crystal with its two wires.

For grinding, the crystal is held by its edges between the thumb and index fingers and stroked back and forth with fairly strong pressure on a Carborundum No. 106 razor hone, or something similar. A combination of figure-8 and back-and-forth strokes can be used. The edges should be checked frequently for squareness under a high-power magnifying glass, and the squareness of the sides checked by silhouetting the crystal against the sky. As the crystal approaches the desired frequency it should be checked for Q by the method already described. Low Q is due to lack of squareness and/or grinding the edges unequally. The necessity for resoldering the crystal to the holder wires each time frequency and Q are checked is an unavoidable evil. The crystal is remounted by a reverse of the taking-out process. After mounting, it should be rinsed in alcohol and dried thoroughly.

It is sometimes difficult to get the exact parts used by the author of an article describing the construction of a piece of equipment. If so, don't lose sleep over it. Substitution of another component having the electrical values specified will rarely lead to anything but satisfactory results. If the part the author used is the *only* one that will work, he'll tell you so.

» A receiver doesn't have to be in the Rolls-Royce class to be capable of good s.s.b. performance.

Modifying the S-40 for S.S.B. Reception

EDWARD H. SOMMERFIELD, W3SGF

Although any communications receiver worthy of the name is capable of receiving s.s.b. signals quite satisfactorily, it is generally agreed that there are a few things that the receiver should have if the advantages of s.s.b. communication are to be realized to the utmost. In the writer's opinion, the main requirements for good reception of s.s.b. signals are a steep-sided 4-kc.-wide i.f. bandpass characteristic, a stable b.f.o., and a means for varying the b.f.o. injection.[1] Several changes were made in the author's S-40 receiver in an effort to meet these requirements, and they resulted in a marked improvement in performance.

The Bandpass Filter

To obtain the desirable sharp i.f. characteristic, a half-lattice crystal filter was added between the first and second i.f. tubes. Fig. 1 shows the circuit — the crystals are the FT-241-A type.

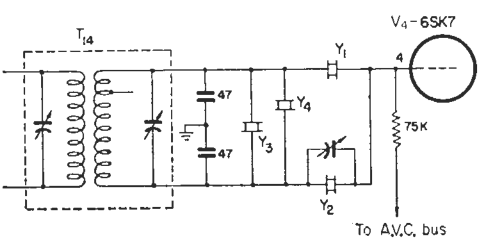


Fig. 1 — The bandpass crystal filter is inserted between the first and second i.f. amplifier stages.
Y_1 — Channel 346, 480.55 kc.
Y_2 — Channel 348, 483.31 kc.
Y_3 — Channel 345, 479.16 kc.
Y_4 — Channel 349, 484.69 kc.
Crystals of these exact frequencies are not required, but the same relative channel (frequency) intervals should be used.

The grid-tap lead was removed from the secondary of T_{14} and the full secondary used, as shown in Fig. 1. A metal bracket for holding the four crystal sockets of the filter was built and mounted under the chassis. In order to place the bracket directly under T_{14} and thereby take advantage of the transformer's mounting bolts, it was necessary to remove the S-meter socket, SO_2, from the rear of the chassis. Wires that had used the pins of socket SO_2 as tie-points were replaced by wires running straight through.

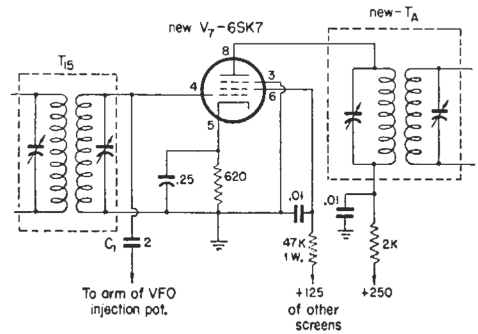


Fig. 2 — Wiring diagram of the new third i.f. amplifier stage.
T_A — 456-kc. output transformer (Meissner 16-6660).

Third I.F. Stage

Adding the crystal filter results in a loss of gain in the receiver, and a third i.f. amplifier stage was added. To do this, the old V_7 socket was rewired for the new 6SK7 amplifier (the V_7 functions were taken over the additional changes to be described later), and the old V_5 socket was

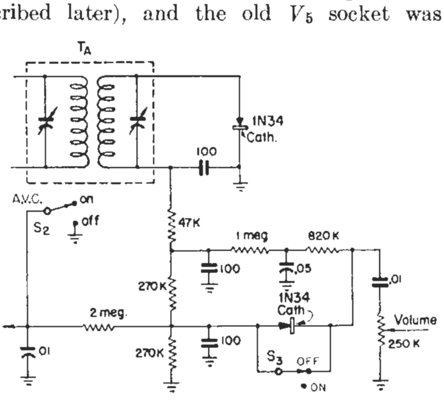


Fig. 3 — The modified 2nd detector and noise-limiter circuits use 1N34 crystal diodes.
S_2 S_3 — As originally in receiver.
T_A — See Fig. 2.

From *QST*, April, 1954.

[1] Two of these are controversial points. Many operators prefer a bandwidth of only 3 kc. or less, arguing that this is sufficient to pass all of the essential frequencies. No one argues the stable b.f.o. point, because frequency stability is vitally important throughout the receiver, from high-frequency oscillator(s) to b.f.o. The advantages of variable b.f.o. injection should be negligible, since the only requirement is that the b.f.o. voltage be appreciably greater than any signal voltage reaching the detector. It is considered to be more important to prevent any b.f.o. voltage getting into the "front end" of the i.f. amplifier, because this will then reduce the signal-handling capabilities of later stages. Another desirable feature of an s.s.b. receiver is a slow tuning rate. — Ed.

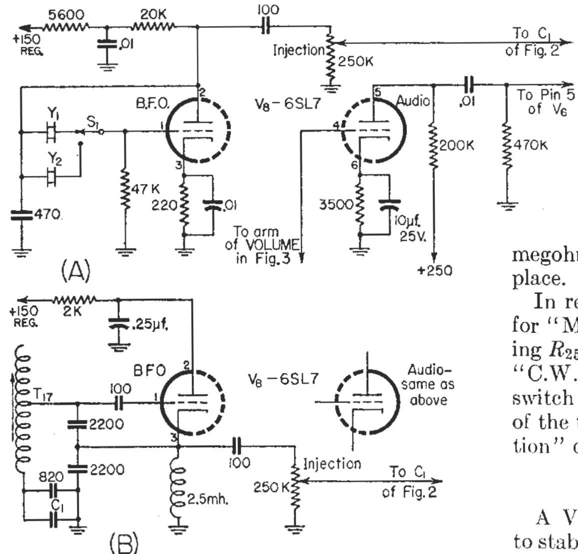


Fig. 4 — **Circuit diagrams of (A) the crystal-controlled b.f.o. and (B) the self-controlled b.f.o.**

C_1 — **Sufficient to bring b.f.o. within range. About 100 or 200 $\mu\mu$f.**

S_1 — **S.p.d.t. rotary switch mounted in "Pitch Control" hole.**

Y_1 — **Channel 348, 483.31 kc.**

Y_2 — **Channel 346, 480.55 kc.**

removed to make room for the additional i.f. transformer. The old V_7 socket and the new transformer were wired as shown in Fig. 2. The b.f.o. injection is at the grid of this third i.f. stage because it was found to give more adequate injection than at the second detector. With variable injection amplitude, getting the proper b.f.o. voltage is no problem.

2nd Detector, Noise Limiter and A.V.C.

The audio stage that was originally combined in V_5 was replaced by an audio stage combined with the b.f.o. and will be described later. The second-detector, a.v.c. and noise-limiter functions were taken over by the circuit shown in Fig. 3. To conserve space, 1N34 crystal diodes were used.

The small components of the circuit of Fig. 3 were mounted on the terminal strip that runs alongside the V_5 and V_7 socket holes.

B.F.O. and 1st Audio

Later models of the S-40 use a dual triode for the b.f.o. and 1st audio stages, and the same dodge was used in this revision. A prime requisite for single-knob s.s.b. reception is a stable b.f.o. Although both self-controlled (Clapp-circuit) and crystal-controlled b.f.o. circuits have been tried in this receiver with good results, our personal preference lies with the crystal oscillator, for stability and ease of alignment (described later). However, since all operators might not prefer the crystal oscillator, both circuits are given in Fig. 4.

Admittedly the crystals used in the b.f.o. circuit do not fall exactly on the outer edges of the crystal-filter passband. However, it was found that they lie close enough to work satisfactorily. The two crystals were mounted in crystal sockets that were in turn mounted on a small metal bracket fastened to the crystal-selector switch S_1 (Fig. 4-A). The switch was then installed in the hole from which T_{17} was removed. The tone control (S_4 in the original wiring diagram) was removed and the 0.25-megohm "Injection" control installed in its place.

In removing the tone control, the tone circuit for "Medium" was left in the circuit by grounding R_{25} (original wiring diagram), and the former "C.W.-A.M." switch (S_5) was used as the a.c. switch (S_{4A} in original). The b.f.o. is left on all of the time — for a.m. reception the new "Injection" control is turned to zero.

Voltage Stabilization

A VR-150 voltage regulator tube was added to stabilize both the b.f.o. and the high-frequency oscillator. This required cutting a new socket hole in the chassis, near the front panel about halfway between the loudspeaker and the tuning flywheel. The circuit was changed as shown in Fig. 5.

Tuning Indicator

A tuning eye was added for ease of alignment. It is a standard 6E5 circuit with the grid tied to the a.v.c. bus. The mounting bracket for the tube was mounted in the upper left-hand corner of the panel.

Alignment

The i.f. amplifier was aligned by plugging a Channel 347 crystal into the b.f.o. and peaking the i.f. transformers at this frequency. If other crystals were used in the filter, requiring that the i.f. amplifier be shifted in frequency, the b.f.o. could still be used as a signal generator for aligning the i.f. by feeding some of its output to Pin 8 of V_2. Just bringing the lead near the stator of the tuning condenser C_{7B} will suffice in most cases. (The high-frequency oscillator can be turned off during this procedure by shorting C_{7C}.) Then peak all of the i.f. transformer trimmers for maximum closing of the tuning eye.

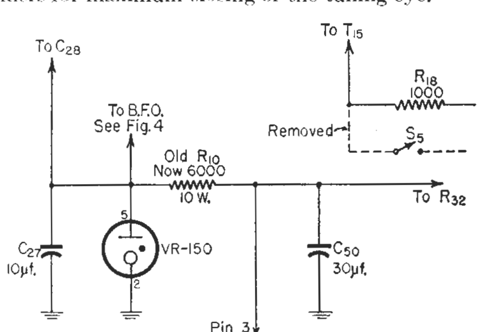


Fig. 5 — **Voltage stabilization is added to the S-40 by changing one resistor (R_{10}) and a few leads.**

» Here is a wealth of practical ideas for a receiver. The trick for selectable-sideband reception without double conversion is particularly ingenious.

Notes on a Specialized 'Phone Receiver

ROBERT W. EHRLICH, W4CUU, EX-W2NJR

While for years it has been customary for amateurs to buy their receivers rather than make them, the modern commercial receiver, expensive as it is, usually falls short of furnishing the best in reception of amateur signals.

Most of the complexity and expense of the standard communications receivers is attributable to the features of broad coverage: bandswitching and gang tuning. These features require engineering compromises all along the line, and precision craftsmanship is needed to get even fair performance. Fortunately, the amateur who builds his own receiver is in a unique position to by-pass all these problems by designing his receiver just to cover his favorite ham band, relying on crystal-controlled converters to pick up any other bands that may be wanted. In this way, the construction job can revert to the standard chassis-and-panel technique, leaving the builder free to concentrate on the circuit design features he wants to incorporate.

The amateur who builds his own receiver has several other advantages over the commercial designer. He can incorporate the exact combination of features to handle his particular needs. He can also avail himself of high-grade surplus components that would be out of the question commercially. And he can take advantage of the latest available techniques that usually take years to find their way into commercial products. As an example of this last item, the now familiar Q5-er was first described in *QST* in 1947, yet it was about five years later that receivers incorporating this degree of selectivity first appeared on the market. Meanwhile, still better selective systems were devised.

A receiver is described here that illustrates the principles just mentioned. No elaborate machine work was involved in making it, and its cost was moderate, yet for its specific job it will completely outperform anything on the market. A detailed discussion of how to make this particular receiver would not be appropriate, because very few amateurs would want to copy it exactly, but it is hoped that a description of the significant features might offer some helpful ideas to the amateur who is seriously interested in getting the most out of his favorite band.

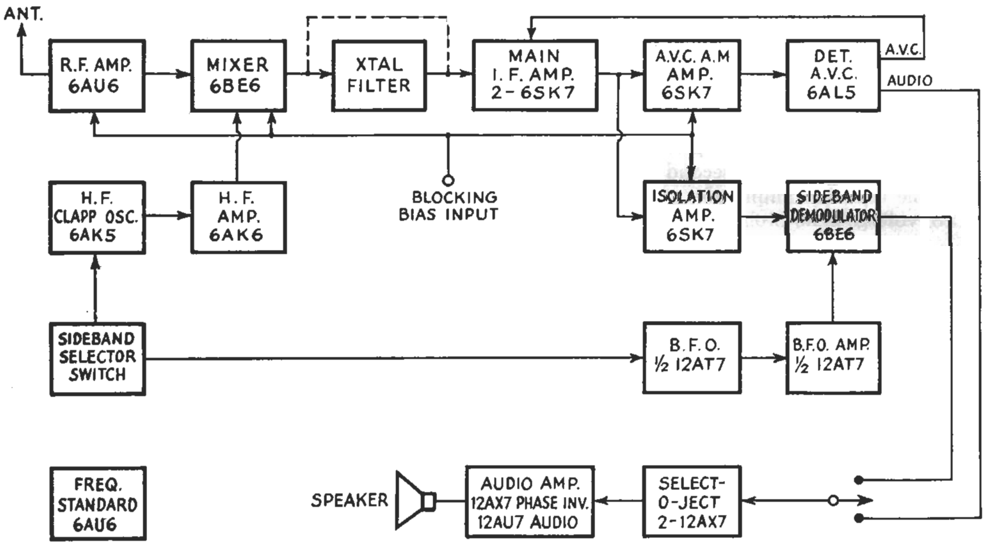


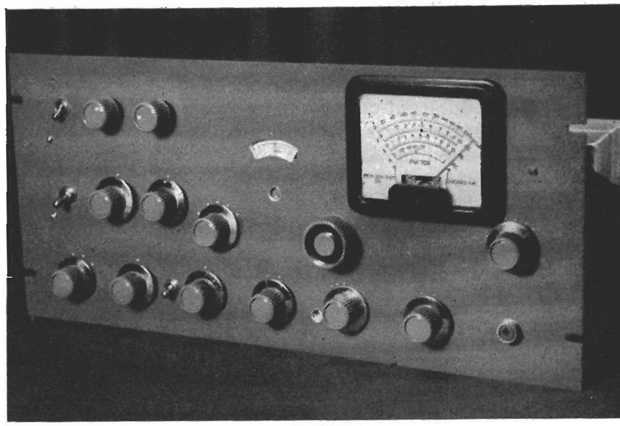
Fig. 1 — Block diagram of the homemade receiver.

The Circuit

A block diagram of the receiver is shown in Fig. 1. The circuit was designed for 75–80 meters, with primary emphasis on the reception of single-

From *QST*, April, 1953.

Front view of the home-made receiver. Panel layout problems are minimized by the absence of bandswitching or gang tuning.



sideband signals. Here, stability is the first requirement — the receiver should be exceptionally stable and capable of being tuned just a few cycles at a time. To take full advantage of s.s.b. communication, the selectivity should be high — high enough to accommodate just one sideband and reject strong adjacent-channel signals without intermodulation effects. Such other features as image ratio and noise limiting, which would be important in a 10-meter receiver, for example, seem to require only secondary consideration.

To start with the front end, its circuit is perfectly ordinary, but the mechanical arrangements for tuning are a departure from the usual complex assembly of gears and shafts. The two r.f. circuits are gang-tuned with an ordinary two-section 50-$\mu\mu$f. variable, with 100-$\mu\mu$f. condensers added across each section to establish the right ratio of minimum to maximum capacitance for tuning the band. The tuning circuits have slug-tuned coils, making it easy to set them to tune together. This condenser is brought out to a panel knob that works about like the old antenna trimmer; it is only necessary to peak up the front end occasionally.

The high-frequency oscillator was designed by pretending it was a transmitter VFO. It uses the Clapp VFO circuit, and a separate box houses just the coil and condenser forming the tuned circuit. The tuning box was made by cutting down a surplus BC-458 transmitter chassis, which provided an excellent main tuning condenser and a geared dial to go with it. In addition to the usual bandsetting fixed condensers, a 5-$\mu\mu$f. variable was also added to provide a $\pm$2 kc. vernier adjustment on the front panel. This has proved to be very helpful in actual receiver operation.

Following the oscillator, an amplifier was found necessary to get enough drive for the 6BE6 mixer. This amplifier is fixed-tuned and peaked near the high-frequency end of the band, to compensate for the tendency of the Clapp oscillator to lose output at that end.

The crystal filter is, of course, the heart of the receiver's selectivity. The filter circuit is just as described by Weaver and Brown, using eight crystals of the 2-digit series. A switching circuit, shown in Fig. 2, enables the filter to be cut out when desired but still leaves two i.f. transformers in the circuit to retain moderate selectivity. The coupling resistor, R_1, is selected to make the overall level of transmission through the i.f. system about the same whether the filter is in or out. The switching circuit and its shielding had to

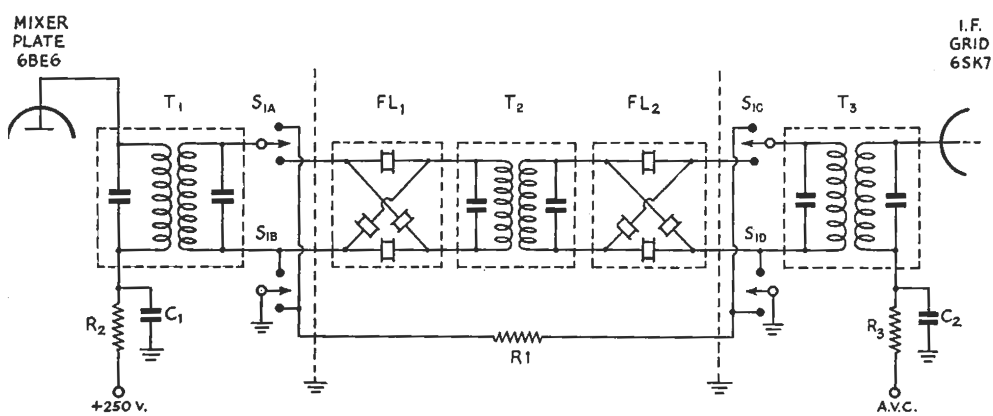


Fig. 2 — Crystal filter and switching circuit.

C_1 — 0.005-μf. ceramic.
C_2 — 0.01-μf. ceramic.
R_1 — 0.47 megohm.
R_2 — 1000 ohms.
R_3 — 0.1 megohm.
FL_1, FL_2 — Crystal-lattice filter section. See text.
T_1, T_2, T_3 — 456-kc. iron-core interstage transformer.

be designed to contribute no stray capacitance paths around the filter when it is being used.

It is significant that the filter is the first thing in the i.f. system. This follows the principle used in producing high adjacent-channel selectivity in commercial mobile receivers, the idea being to eliminate all unwanted signals at the lowest possible power level before they are amplified. There was some apprehension that the attenuation of the filter might degrade the over-all signal-to-noise ratio, but it was found that the front end had enough gain so that the first r.f. grid circuit still is the controlling noise source.

After two stages of amplification, the i.f. system splits into two branches. One branch feeds a carrier-type demodulator, using a 6BE6 tube, for detection of c.w. and s.s.b. signals. With this kind of detector, shown in the upper portion of Fig. 3, no intermodulation products are developed as long as the signal input is held below about one volt. The i.f. stage feeding this detector does not provide any additional gain; its main function is to isolate the main i.f. and a.v.c. systems from the effects of the strong b.f.o. signal (about 20 volts) injected into the 6BE6.

The second branch feeds a combination a.m. detector and a.v.c. system, using a 6AL5 double diode. This circuit, together with the switching arrangement for the two detectors, is shown in the lower part of Fig. 3. Of particular interest is the diode section that is cut in series with the a.v.c. line on switch position No. 3, for a.v.c. reception of c.w. or s.s.b. signals. This diode causes the a.v.c. to charge up quickly but discharge slowly, so that in effect the a.v.c. bias "hangs up" and rides with the peaks of the received c.w. or s.s.b. signal.

The i.f. stage feeding the a.v.c. detector operates at a fixed gain of about 40. With this arrangement, there can be 40 volts of a.v.c. bias for every 1 volt of signal at the 6BE6 demodulator grid. Since 40 volts is enough to cut off the main i.f. amplifiers, it follows that, with the a.v.c. operating, no signal can overload the 6BE6.

Coming to the matter of sideband selection, nothing further would need to be done if only a.m. signals, with carrier, were to be received. It is only necessary to tune the receiver a little to one side or the other, keeping the carrier just within the edges of the crystal filter response. Since the filter has a flat-topped characteristic, the signal remains perfectly intelligible over a

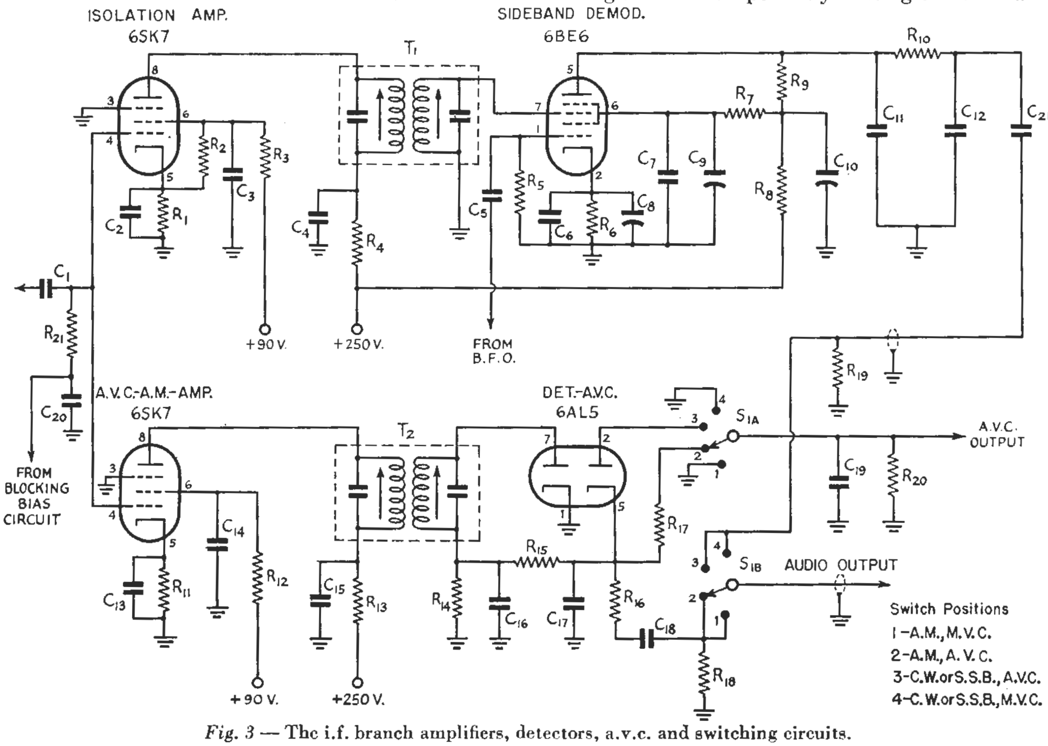
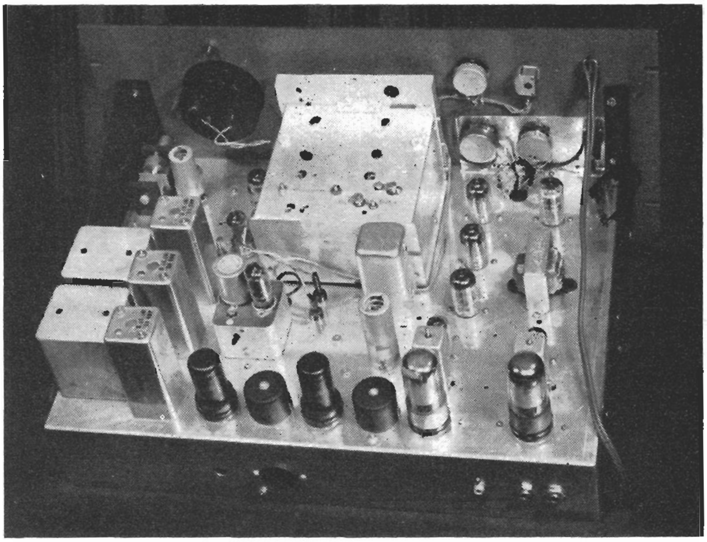


Fig. 3 — **The i.f. branch amplifiers, detectors, a.v.c. and switching circuits.**

C_1 — 100 μμf.
C_2, C_3, C_4, C_6, C_7, C_{13}, C_{14}, C_{15}, C_{18}, C_{20}, C_{21} — 0.01-μf. ceramic.
C_5, C_{11} — 470 μμf.
C_8 — 10-μf. 50-volt electrolytic.
C_9, C_{10} — 8-μf. 450-volt electrolytic.
C_{12}, C_{16}, C_{17} — 270 μμf.
C_{19} — 1 μf., oil-filled.
R_1, R_5 — 20,000 ohms.
R_2, R_{10}, R_{15} — 47,000 ohms.
R_3, R_4, R_{12}, R_{13} — 1000 ohms.
R_6, R_{11} — 220 ohms.
R_7 — 10,000 ohms, 1 watt.
R_8 — 2700 ohms.
R_9 — 33,000 ohms.
R_{14} — 0.1 megohm.
R_{16}, R_{20} — 10 megohms.
R_{17} — 2.2 megohms.
R_{18}, R_{19} — 1 megohm.
R_{21} — 0.47 megohm.
S_1 — 2-circuit 4-position wafer switch.
T_1 — 456-kc. iron-core interstage transformer.
T_2 — 456-kc. iron-core output transformer.



Ordinary chassis and panel construction can be used. The large box at the center houses the coil and tuning condenser for the high-frequency oscillator. The crystal-filter network elements are mounted in the group of shield cans at the left.

range of about 2½ kc. of tuning. Heterodyne QRM falling on one sideband can be completely eliminated by judicious tuning.

For s.s.b. signals, or for exalted-carrier reception of a.m., tuning from one sideband to another requires that both the high-frequency oscillator and the b.f.o. injection oscillator be moved simultaneously in order to maintain zero beat. The switching circuit of Fig. 4 is used for this purpose. With just the two switching condensers, C_1 and C_2, the circuit would not perform properly because the shift in the high-frequency oscillator would be different at different parts of its tuning range. Compensating capacitor, C_3, takes care of this problem. Its shaft is brought out to the front panel, and a calibration mark is made for each

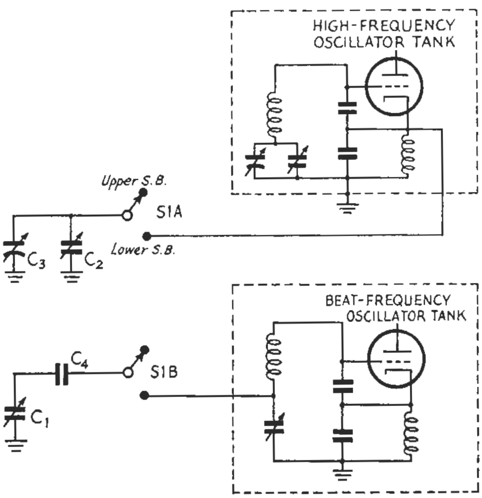


Fig. 4 — The sideband-switching circuit. The connections are shown for use with a receiver with the h.f. oscillator on the high side of the signal frequency.

C_1, C_2 — 7-45 μμf. trimmer.
C_3 — 15-μμf. midget variable with shaft.
C_4 — 10 μμf.
S_1 — 2-circuit 2-position ceramic wafer switch.

100 kc. across the band. The setting need only be within the proper 100-kc. segment to keep the switching error within a few cycles. If one desired to go to the trouble, this condenser might easily be ganged with the main tuning.

The audio system includes a standard Select-o-ject to help with the heterodynes, etc., that are not eliminated by the sideband filter. Following this, it is important that there be plenty of gain, so that neither the Select-o-ject nor the 6BE6 stage need operate above their distortion limits to produce enough audio output.

A somewhat unusual method is used for disabling the receiver during transmissions. Applying negative bias to various amplifiers is a convenient scheme, but such an arrangement is usually encumbered by the time constants in the a.v.c. circuit. In this receiver, a.v.c. is applied only to the two i.f. amplifying stages, so the blocking bias is applied to all *other* stages: r.f., mixer, and the i.f. branch stages. The latter circuit has only a few 0.01-μf. r.f. by-pass condensers involved, so its time constant is quite short. Normally, an external bias is applied when it is desired to cut the receiver off. If this is not available, however, the first r.f. grid will rectify the incoming signal and produce enough bias to cut off all the other stages and silence the receiver. Meanwhile, the a.v.c. detector experiences no signal because its branch amplifier is cut off, so the receiver comes back with full life the instant the transmitter is turned off.

Conclusion

It is, of course, a pleasure to have in the shack a receiver that within itself is capable of doing the full job for reception of all kinds of 'phone signals in a crowded band. At the same time it has been a refreshing experience to reaffirm the notion that with time and thought the amateur can still do as well or better than his commercial brethren. Try building your own "dream" receiver; you'll enjoy it.

» *This is a recent approach to providing several degrees of receiver selectivity, to meet varying QRM conditions.*

Variable I.F. Selectivity for the Communications Receiver

ONE feature of the SX-88 receiver is the variable-selectivity i.f. amplifier that provides six bandwidths, from 250 cycles to 10 kilocycles. The heart of this 50-kc. i.f. amplifier is the tuned circuits that are used. These are special coils tuned by a ferrite slug and surrounded by a ferrite sleeve. The special design

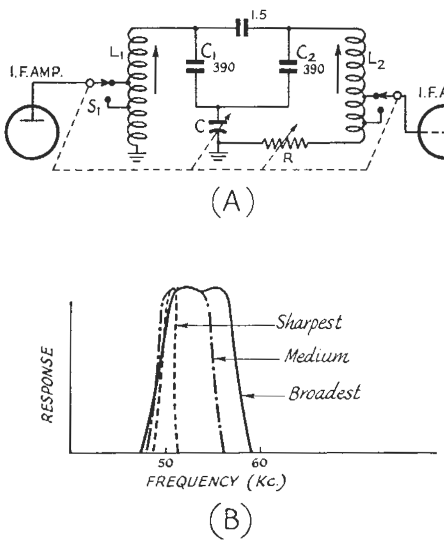


Fig. 1 — (A) Basic circuit of the variable-bandwidth i.f. used in the SX-88. The coupling is increased as C is made smaller, and the Q of L_2C_2 is reduced as R is increased. The stage gain is held constant with changes in bandwidth by tapping the grid and plate up or down on the coils.
(B) The effect of varying C and R in (A) is that the passband "grows" out to a higher frequency, as illustrated here.

gives a coil with a Q of 175 to 185 at 50 kc. An interesting sidelight is that it was found impossible to obtain a Q of higher than 130 until a metal screw was removed from the ferrite core and a means was found for threading the glass-hard and glass-brittle ferrite.

Almost anyone can build a *sharp* i.f. amplifier if he is handed a batch of $Q = 180$ coils, but the SX-88 i.f. has the wide range of bandwidths mentioned earlier. This poses quite a problem, because the frequency must not be changed radically by the bandwidth-variation method, and the gain must be held substantially constant. This was accomplished by the Hallicrafters engineers in the general way shown in Fig. 1A. This simplified diagram shows a variable condenser ganged with a variable resistor — in the actual receiver these are step-switched. It can be seen that the smaller the capacity of C, the tighter will be the coupling between the two tuned circuits, L_1C_1 and L_2C_2. Furthermore, the larger the value of R is made, the lower becomes the Q of the grid tuned circuit, L_2C_2. By proper proportioning of the various values of C and R (at different switch positions), the wide range in bandwidth variation is obtained. One of the three 50-kc. i.f. stages has taps on the coils, as represented in Fig. 1A by the leads to S_1, and this enables the gain of the i.f. amplifier to be held relatively constant over the entire range.

The *midband* frequency of this i.f. system does not remain constant — the low-frequency edge remains substantially constant. This is illustrated in Fig. 1B, and it is something the operator must remember if he is to understand fully the performance of the receiver as the bandwidth is changed. Here three conditions ("sharpest," "medium" and "broadest") are shown — the effect is as though the bandwith "grows" to the higher frequency. It is pointed out here to explain what will undoubtedly puzzle some operators when they switch bandwidths and find that sometimes the carrier drops out and sometimes it doesn't. Obviously, it will depend on whether one has the carrier centered at around 50 kc. or on the high-frequency side of the i.f. passband.

From "New Equipment — the SX88 Receiver," *QST*, June, 1954.

EASY-TO-TUNE S.S.B. SIGNALS

Insufficient sideband suppression will make a signal difficult (if not impossible!) to tune in on a receiver with no selectivity, but good selectivity in the receiver can increase the sideband suppression *in the receiver* and thus make a decent s.s.b. signal out of a mediocre one.

What really makes a s.s.b. signal — any s.s.b. signal — "hard to tune" is a fast tuning rate in the receiver. The receiver has to be set with an accuracy measured in *cycles* to receive the signal faithfully, and this doesn't leave much room for error. A receiver that covers 100 or 200 kilocycles per revolution of the tuning knob makes tuning a single-sideband signal much more difficult than one that covers 20 or 25 kc. per knob rotation.

» *"Selectable-sideband" receiving permits the operator to switch instantly to one or the other sideband of an incoming a.m. signal and thus receive the signal with a minimum of QRM.*

A Mixer-Oscillator for Selectable-Sideband Reception

The McLaughlin selectable-sideband system of receiving uses a selective low-frequency i.f. amplifier following a normal receiver's i.f. amplifier. If, for example, the selective amplifier is centered at 50 kc. and the normal receiver i.f. is centered at 450 kc., the signals can be heterodyned from 450 to 50 kc. by beating them against a 400- or a 500-kc. signal. If, however, the sharp i.f. amplifier is not centered exactly at 50 kc., but instead has a 3-kc. bandwidth extending from 50 to 53 kc., a 'phone signal at 450 kc. in the normal i.f. amplifier will have its carrier heterodyned to 50 kc. by the 400- or 500-kc. signal mentioned above, but only one sideband will fall within the selective i.f. amplifier's range.

To insure good stability of the system, crystal-controlled oscillators are used to heterodyne the signal to the low frequency. A circuit for use with a 455-kc. i.f. and a 50-kc. i.f. is shown in Fig. 1. The 455-kc. signal from the receiver is picked up by coupling through a 1- or 2-$\mu\mu$f. capacity from the grid of the second i.f. amplifier to a length of shielded cable that connects to the 455-kc. input jack of Fig. 1. The position of switch S_1 determines which crystal oscillator is in operation. The third position on the switch permits both oscillators to run simultaneously, and this position will be found useful in the original alignment of the entire receiving system. The 750-μh. r.f. chokes in the screen leads of the mixers provide tuning for the oscillator portions, and it may be found necessary to shunt them with capacities other than that shown in Fig. 1, depending upon the crystals and r.f. chokes that are used.

Other combinations of frequencies can be used, of course. For example, an 85-kc. selective i.f. amplifier working from a 455-kc. amplifier would require oscillator crystals of 370 and 540 kc.

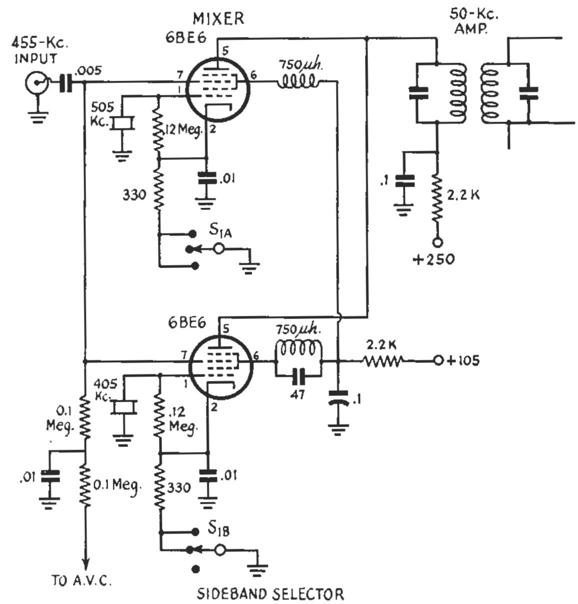


Fig. 1 — Circuit diagram of the mixer/oscillator portion of a selectable-sideband i.f. system. By switching the oscillator to the "high" or the "low" side of a selective i.f., either sideband of a 'phone signal can be selected without retuning the receiver.

From "An All-Purpose Super-Selective I.F. Amplifier," *QST*, March, 1953.

Just to prove that one really doesn't need too complicated a receiver for s.s.b., W2NJR (now W4CUU) cites a QSL card he received from North Carolina, where the SWL was using a 6F8G regenerative receiver!

» Here is a trick for using a single-sideband filter and switching sidebands without shifting the carrier frequency. A modulator circuit for use with a Collins mechanical filter is also shown.

Selectable Sideband with VFO and a Filter-Type Generator

This idea, passed along by Jim Freund, W5QMI, may be helpful. One of the problems with a filter-type s.s.b. generator is in devising a method for selecting either sideband in the output. If the filter is one that has a characteristic steeper on one side than on the other, as many of them do, it is desirable to keep the (suppressed) carrier always on the same side of the filter. W5QMI's system, shown in Fig. 1, allows you to do just that. The output from the VFO (the actual operating frequency) is mixed with a signal from the filter-frequency oscillator. The sum or difference beat is selected, depending upon the sideband to be used in the amateur band, and this is used to heterodyne the low-frequency s.s.b. signal to the operating frequency. The output of the second mixer can, of course, be ganged to the VFO. There are several advantages to the system: only one crystal oscillator is required, an operating-frequency VFO can be used and it sets the stability of the system, only one sideband filter is required but choice of sidebands is available, VFO output is available for carrier reinsertion at the receiver, and the "best" side of filter is always used.

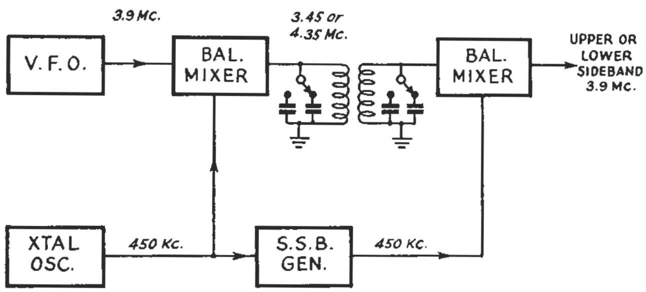


Fig. 1 — Block diagram of a heterodyne method for selecting upper or lower sidebands with a filter-type s.s.b. generator. This method has the advantage over changing the modulated-carrier frequency in that the modulated carrier is set up once for best filter action and left there.

Fig. 2 shows a suggested low-frequency s.s.b. generator, using the Collins mechanical filter or a crystal-lattice filter, that would be applicable in such a system. The balanced modulator, using two 1N35s, is adjusted for best carrier balance by the setting of the 250-ohm potentiometer. Care should be exercised in the construction of the generator to insure a minimum of coupling between the 6AK6 oscillator plate coil and the 6AK6 amplifier plate coil, for best carrier suppression.

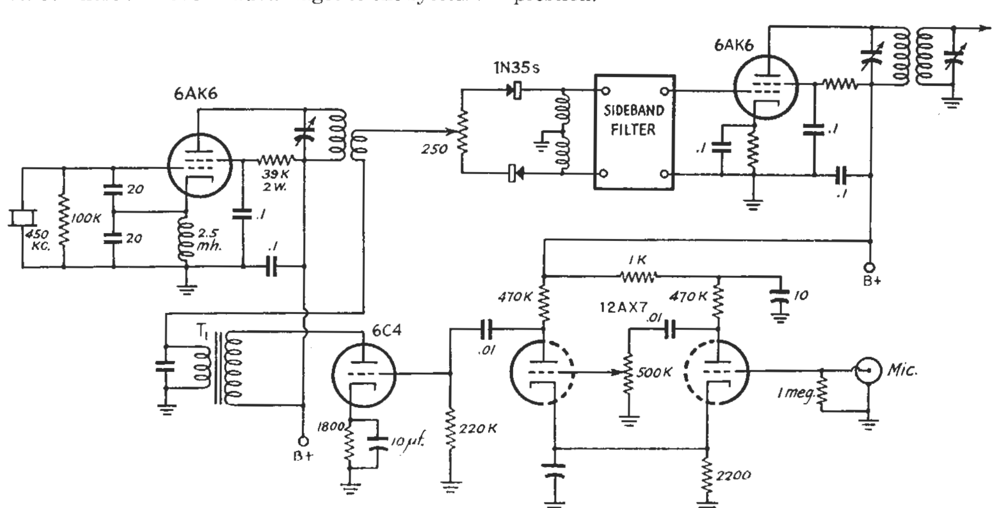


Fig. 2 — A modulator circuit for use with the Collins mechanical filter. T_1 is a small universal output transformer.

» 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.

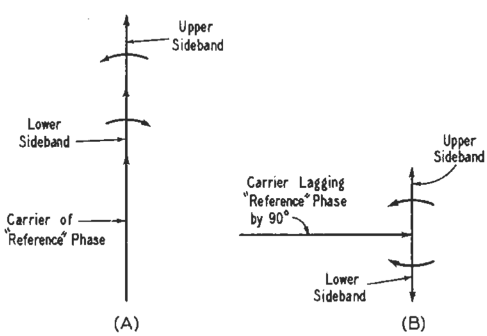


Fig. 1 — The carrier and sideband relationship required to generate a single-sideband signal by the "phasing" or "balancing" method. The modulating signal in B leads the modulating signal in A by 90°. When the two signals represented by A and B are combined, the upper sidebands add and the lower side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

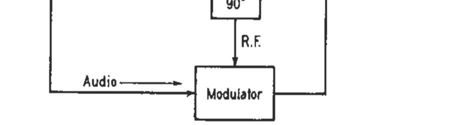


Fig. 2 — A block diagram showing the circuits required to generate a single sideband by the method of Fig. 1. This is an impractical method because there is no known means for obtaining the 90° audio shift over a wide range of frequencies.

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.

It so happens that two phase-shift networks having a *differential* phase shift of 90° can be inserted between the source of modulating signals

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

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

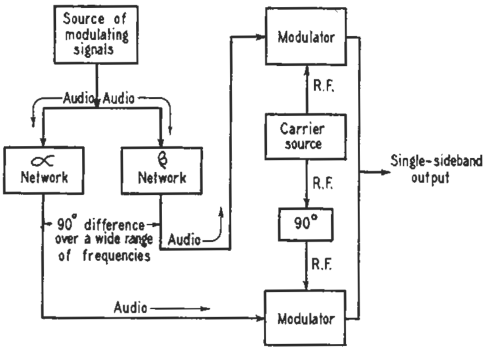


Fig. 3 — The system outlined in Fig. 2 becomes practical by using two audio channels (α and β networks) with a constant phase *difference* of 90°.

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:

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

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

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

4) Modulation may be accomplished at any chosen power level. In the interest of efficiency, it is generally wise to carry out this portion of the process at receiver-tube level, using linear amplifiers to build up the power.

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

The characteristics of typical wide-band phase-shift networks are shown in Fig. 4. It can be seen that the differential phase shift averages 90° over a frequency range of at least 7 octaves. Of course, the ideal differential phase shift is exactly 90°, and the excursions of the actual phase-shift curve are $\pm$ 2° 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 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, 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-

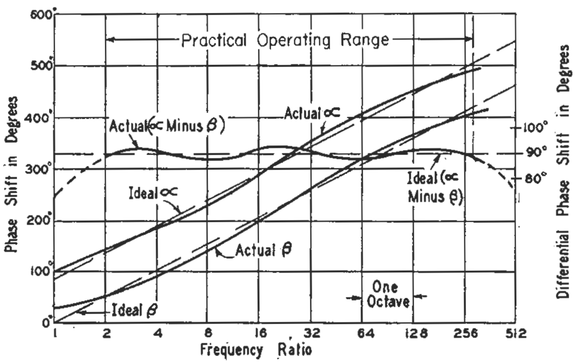


Fig. 4 — This plot shows how the 90° difference between the α and β networks is maintained over a wide frequency range. The scale for the phase difference is given on the right-hand side of the graph.

sideband exciter; rather, the purpose is to point 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 watts 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.

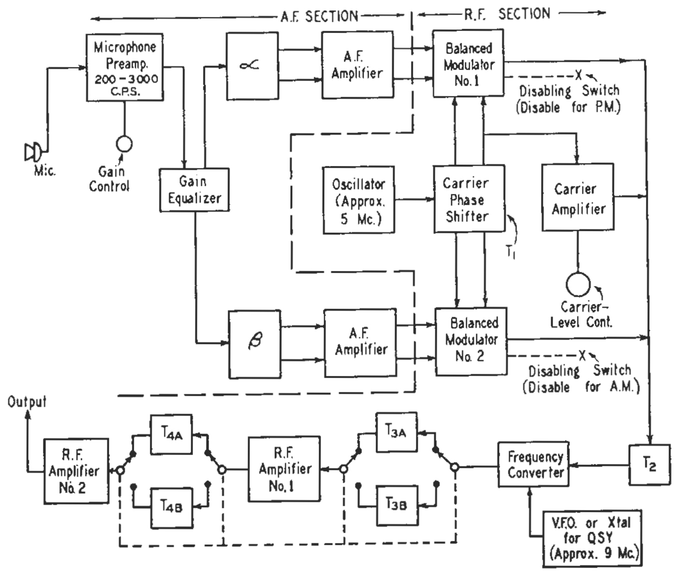


Fig. 5 — Block diagram of an exciter capable of generating s.s.s.c., a.m. and p.m. signals on either lbe 75- or 20-meter bands. Table I gives a description of the various components.

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 objectives 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. (See Fig. 5.)

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 3000 c.p.s. for maximum intelligibility. Why do anything about it at all, if the system as such does not require it? The answer has two important aspects — important to *you* as an occupant of the bands we share:

1) Intelligible speech does not require transmission of frequencies higher than 3000 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 bandwidth so much that he receives only what is necessary, if even that much. Not only is "high fidelity" wasted, but also its use is downright selfish.

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. It's worth it!

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. feed-back.

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 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,

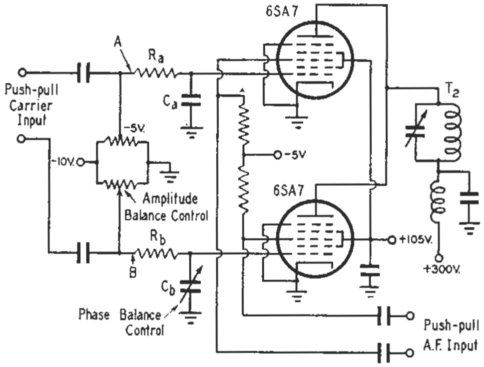


Fig. 6 — A typical balanced modulator, using 6SA7 tubes. Provision is included for obtaining amplitude and phase balance of the r.f. (carrier) input.

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 ±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 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.

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.s.c.–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 beam-power output tube. Can be operated as Class A or B amplifier.

AUDIO PHASING NETWORKS

The 90-degree audio phasing network described in the preceding article is an early "active" type (containing amplifiers) that has since been superseded by simpler "passive" networks using only resistors and condensers. An example of the latter is the network devised by R. B. Dome, W2WAM, and used by W2UNJ in the phasing exciter described on page 96. A passive network using even fewer parts was incorporated in W2KUJ's "SSB Jr." (*G. E. Ham News*, Vol. 5, No. 6) and has since been made available commercially as a precision-adjusted inexpensive unit. These last two networks have substantially identical performance, the principal difference in their characteristics being that the Dome network can be terminated in resistance while the Norgaard network must work into an open circuit.

» The necessity for a high degree of care in preventing distortion subsequent to modulation has been emphasized in the preceding article. 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 such 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.

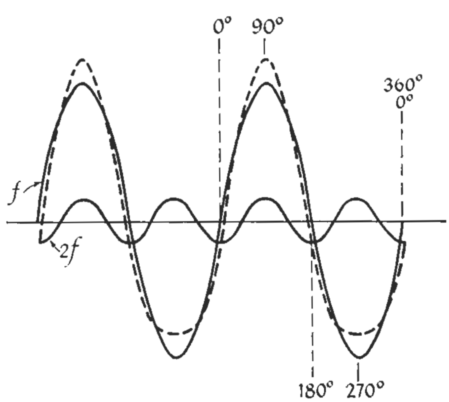


Fig. 1 — The common type of distortion in vacuum tubes can be resolved into a fundamental and second harmonic having the time relationships shown. The fundamental, *f*, is the amplified grid signal, and 2*f* is the generated harmonic.

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 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 modulators with 90-degree r.f. phasing, the composite diagram is as shown in the

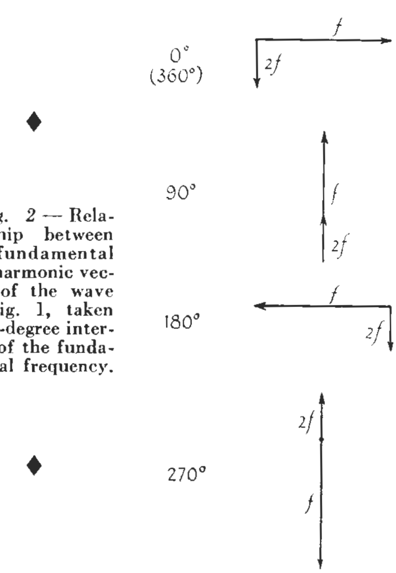


Fig. 2 — Relationship between the fundamental and harmonic vectors of the wave in Fig. 1, taken at 90-degree intervals of the fundamental frequency.

From *QST*, February, 1954.

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.

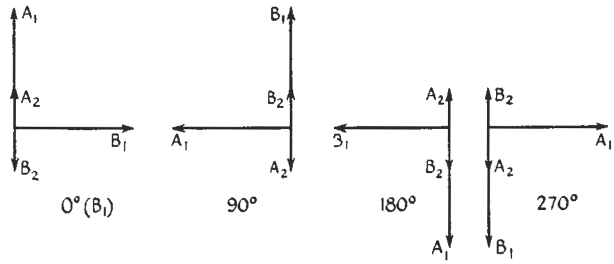


Fig. 3 — **Phase relationships in two channels when the fundamental in one (A_1) leads the fundamental in the other (B_1) by 90 degrees. A_2 is the second harmonic associated with A_1, and B_2 the second harmonic associated with B_1.**

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.

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.

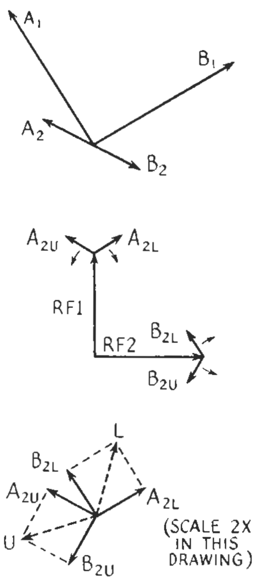


Fig. 4 — **Sideband output when the harmonics (A_2 and B_2) in the upper drawing are applied to balanced modulators having 90-degree r.f. phasing. The r.f. carriers are shown for reference in the middle drawing. When carriers are eliminated the upper and lower side components, subscripts U and L respectively, combine as shown in the lower figure.**

» The "W2UNJ Exciter" has given many hams their first taste of single-sideband operation. It's easy to build, and any good amateur-band VFO can be used as the r.f. source. Also, it has enough power output to give a surprising amount of communication when worked directly into an antenna.

A Phasing-Type S.S.B. Exciter of Simple Design

WILLIAM M. RUST, W2UNJ

THIS article will describe an exciter, built mostly of junk-box parts, that is suitable for the average ham who might like to try single sideband with a minimum of cost and effort. It is small and compact; five inches wide, ten inches deep behind the panel, and seven inches high overall. No special or close tolerance components are used, with the exception of six condensers and six resistors that are carefully chosen from a stock of standard units.

Assuming that an a.m. transmitter is already at hand, the exciter takes its r.f. driving power from the present exciter and audio power from the present speech amplifier. It will deliver single-sideband output either to an antenna for local work or to an r.f. amplifier that is adjusted for linear operation. The operating frequency can be varied over a wide range without seriously impairing the adjustment. Provision is made for transmitting either the upper or the lower sideband.

The Circuit

The circuit uses the phasing method of single-sideband generation and is shown in block diagram in Fig. 1.

An r.f. source is fed into two phase-shifting networks. One network has an output voltage that leads the input voltage by 45 degrees, and the other has an output voltage that lags the input voltage by 45 degrees. Therefore, there is a 90-degree phase difference between the output voltages of the two networks. Each of these r.f. voltages is used to drive a pair of balanced modulators. Each pair of balanced modulators balances out the excitation frequency, so that no carrier appears in the output of either pair of balanced modulators.

The audio source is fed into a Dome-type phase-shifting network. This network requires push-pull input voltage, and delivers two output voltages differing in phase by 90 degrees. These two voltages are used to drive separate Class A amplifiers that serve the dual purpose of amplification and isolation of the network from the audio load (the balanced modulators). Each pair of balanced modulators, when supplied with audio from its amplifier, produces two sidebands in its output, but when the outputs of both pairs of balanced modulators are combined, one sideband is canceled out and the resultant output is single sideband.

The complete schematic of the exciter is shown in Fig. 2. Four 6V6 tubes are used as balanced modulators. The plate circuit of the balanced modulators uses a push-pull-parallel arrangement. The grids of one pair of balanced modulators are fed through a phase-shift network consisting of a 300-ohm resistor and an inductance that is adjustable to 300 ohms reactance at the operating frequency. The grids of the second pair of balanced modulators are fed through a phase-shift network consisting of a 300-ohm resistor and a condenser which is adjustable to 300 ohms reactance at the operating frequency. The input impedance of the two phase-shift networks in parallel is 300 ohms.

Screen modulation is used, and the screen of each modulator tube is by-passed to ground for

From "Single Sideband for the Average Ham," *QST*, August, 1949, and incorporating subsequent improvements communicated by the author.

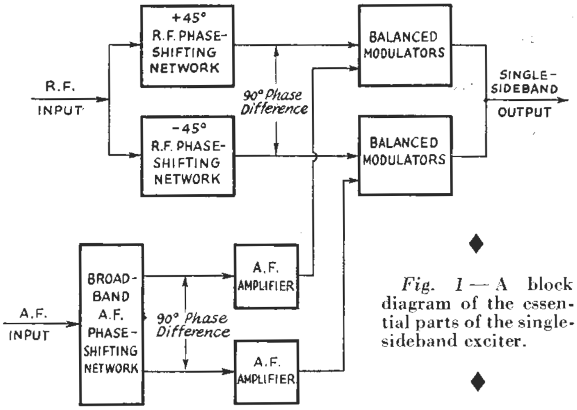


Fig. 1 — A block diagram of the essential parts of the single-sideband exciter.

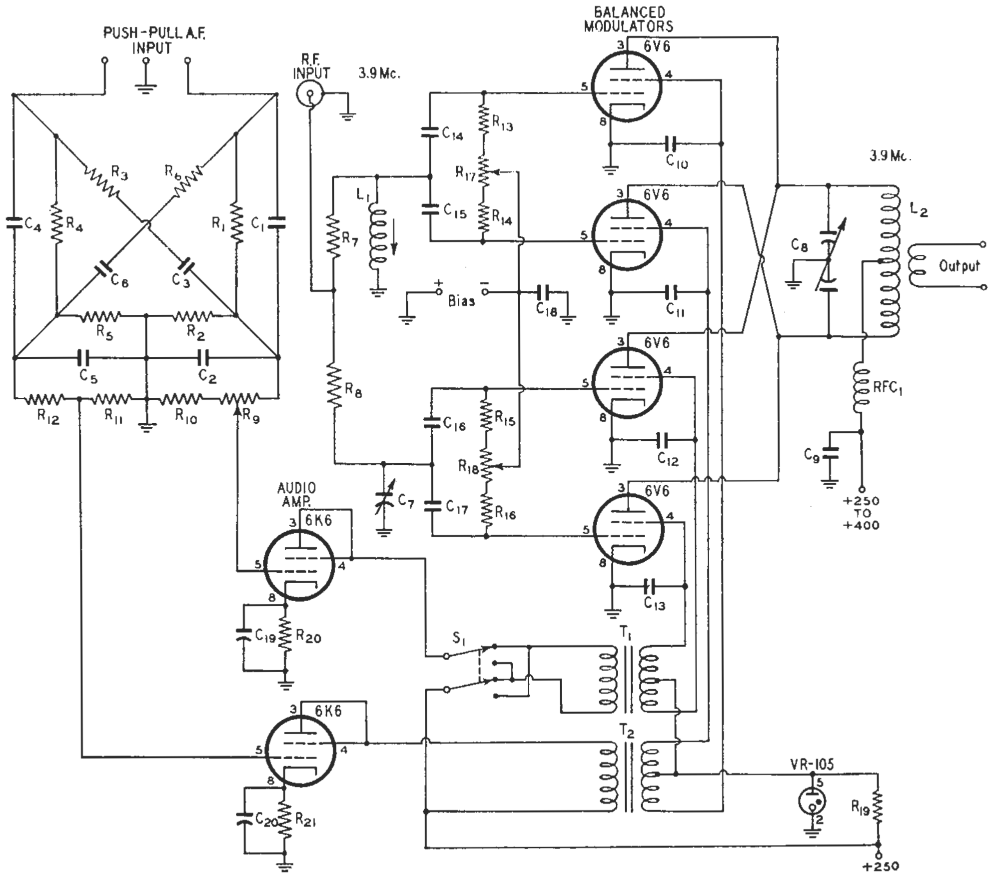


Fig. 2 — Circuit diagram of the single-sideband exciter.

C_1–C_6, inc. — See Table I.
C_7 — Air padder condenser. 3.9 Mc.: 150 $\mu\mu$f.; 7 Mc.: 100 $\mu\mu$f.; 14 Mc.: 35 $\mu\mu$f.
C_8 — Approx. 400-$\mu\mu$f. per section, b.c. receiver tuning condenser.
C_9 — 0.001-μf. 1000-volt mica.
C_{10}–C_{18}, inc. — 0.001-μf. 500-volt mica.
C_{19}, C_{20} — 4-μf. 150-volt electrolytic.
R_1–R_6 — See Table 12-II.
R_7, R_8 — 300 ohms, 5 watts (5 1500-ohm 1-watt in parallel).
R_9 — 0.5-megohm linear volume control.
R_{10} — 0.47 megohm.
R_{11} — 0.75 megohm.
R_{12} — 0.24 megohm.
R_{13}–R_{16}, inc. — 10,000 ohms.
R_{17}, R_{18} — 15,000-ohm potentiometer, wirewound.
R_{19} — 7500 ohms, 10 watts.
R_{20}, R_{21} — 680 ohms, 2 watts.
All resistors 1-watt unless specified otherwise.
L_1 — 3.9 Mc.: 25 turns No. 28 enam.
7 Mc.: 18 turns No. 22 enam.
14 Mc.: 12 turns No. 20 enam.
All coils close-wound at mounting end of slot of National XR-50 slug-tuned form.
L_2 — 3.9 Mc.: 40-meter 75-watt tank coil with swinging link (Bud OLS-40).
7 Mc.: 20-meter 75-watt tank coil with swinging link (Bud OLS-20).
14 Mc.: 15-meter 75-watt tank coil with swinging link (Bud OLS-15).
RFC_1 — 2.5-mh. r.f. choke.
S_1 — D.p.d.t. toggle, preferably with center off position. See text.
T_1, T_2 — 5-watt modulation transformer, 10,000 ohms c.t. to 4000 ohms (Stancor A-3812).

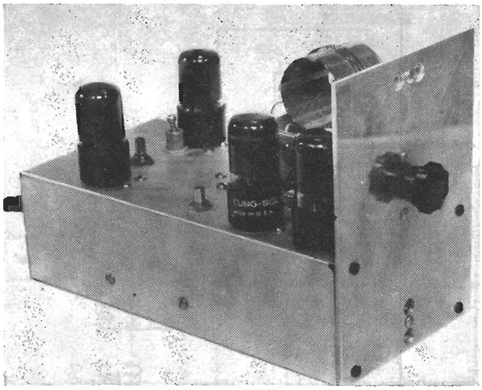
r.f. A transformer with a center-tapped secondary winding is employed in the output of each audio amplifier to provide push-pull modulating voltages. The d.c. screen voltage is regulated by a VR-105.

A voltage divider is inserted between each output of the audio phase-shift network and the corresponding amplifier grid. One of these voltage dividers is made variable to provide for balancing the two audio channels. The network constants are compensated for the load of these voltage dividers.

A reversing switch, S_1, allows switching to either the upper or lower sideband.

The Audio Phase-Shift Network

Rumor has it that audio phase-shift networks are difficult to construct because they require odd values of resistance and capacitance, made to very close tolerances. However, advantage can be taken of the fact that most resistors and condensers available in junk boxes and in stock at local dealers do vary considerably from their nominal values.



The original W2UNJ exciter used a circuit differing in several details from that given in Fig. 2, but the identical constructional layout can be used. This model represents a very simple approach to single sideband for the newcomer, since it contains only those elements essential to generating s.s.b. The r.f. required can be taken from any existing source, and the audio can come from the speech amplifier of an existing a.m. transmitter. The VR-105 shown in Fig. 2 is not included in the unit itself, but there is ample room on the chassis to accommodate it.

Table I is used in selecting the network components. The procedure is to collect as many resistors and condensers as possible with nominal values as indicated in the second column of the chart. Measure all of the condensers first, and select the six condensers whose measured values are closest to the "target values" in the third column. Enter the measured values of these condensers in the fourth column of the chart. Then calculate the "target values" for the resistors and select the six resistors whose measured values are closest to these target values.

A capacity bridge, of the type used by servicemen, and a good ohmmeter should give sufficient accuracy in selecting the network components. Absolute accuracy is not important, if the components are all in correct proportion to each other. A difference in percentage error between the resistance measurements and the capacitance measurements will merely shift the operating range of the network.

The network components are mounted on a small sheet of insulating material to facilitate wiring. If the network components have been carefully chosen and assembled, no test of the network should be necessary.

Perhaps some local amateur supply houses can be persuaded to furnish the stock of resistors and condensers, and the measuring equipment, as a service to amateur customers.

The R.F. Phasing Inductance

The only other "tricky" component of the exciter is the r.f. phasing inductance, L_1. This inductance is wound on a slug-tuned form. The coil should resonate, respectively, at the center of the band with a variable condenser set at about 135 $\mu\mu$f. for 3.9 Mc., 75 $\mu\mu$f. for 7.25 Mc., and 27 $\mu\mu$f. for 14.3 Mc. Resonance can be checked by using the coil and condenser as a wavetrap connected in series with the antenna on the station receiver.

Construction

The exciter is assembled on a 5 × 10 × 3-inch chassis. The plate tank tuning condenser is mounted on top of the chassis, front and center, with two of the 6V6 modulator tubes on each side. The plate tank coil is mounted on top of the condenser. Plate leads from the four 6V6s are brought directly to the tuning condenser through four ⅜-inch holes drilled through the chassis near each tube-socket plate connection. The 6V6 screen grids are by-passed to ground directly at the sockets. R_{17}, L_1, C_7 and R_{18} (all adjustable components) are mounted in a row directly behind the 6V6s. The two 6K6 amplifiers are mounted at the rear of the chassis, one on each side, with R_9 and S_1 between them. The audio phase-shift network is mounted inside the chassis at the rear. Crystal sockets are used for r.f. input and output connections. A cable is brought out at the rear of the chassis for audio and power connections. Layout, construction and wiring are all conventional. The 5 × 7-inch front panel is optional.

Associated Equipment

The r.f. input impedance of the exciter is 300 ohms, but a link line of lower characteristic impedance will operate satisfactorily for the short distance usually required. A means for adjusting the r.f. driving power is desirable. A surplus Command set transmitter (BC-696 or T-19/ARC-5), operating at low plate voltages, makes an ideal r.f. source for 75 meter work, but any VFO or crystal oscillator with a few watts output will do.

In most stations, the handiest source of push-pull audio for the exciter will be the secondary of the modulator driver transformer. A single triode-connected 6F6 output tube in the speech ampli-

TABLE I
Phase-Shift Network Design Data

Part	*Nominal Value*	*Target Value*	*Measured Value*
C_1	.001	.00105	(C_{m1})
C_2	.002	.00210	(C_{m2})
C_3	.006	.00630	(C_{m3})
C_4	.005	.00475	(C_{m4})
C_5	.01	.00950	(C_{m5})
C_6	.03	.0285	(C_{m6})
R_1	100,000	$\frac{100}{C_{m1}} =$	
R_2	50,000	$\frac{105}{C_{m2}} =$	
R_3	15,000	$\frac{100}{C_{m3}} =$	
R_4	100,000	$\frac{453}{C_{m4}} =$	
R_5	50,000	$\frac{476}{C_{m5}} =$	
R_6	15,000	$\frac{453}{C_{m6}} =$	

All condensers mica, and all resistors 1 watt.

fier will provide sufficient audio. The modulator tubes should be removed from their sockets, and the center tap of the driver-transformer secondary should be grounded, after removing the bias connection. An alternative method is to use blocking condensers in the audio leads to the single-sideband exciter to isolate the modulator bias from the audio phase-shift network in the exciter. If some other source of push-pull audio is used, it should have low internal impedance (Class A triodes, or beam tubes with negative *voltage* feed-back).

The exciter may be coupled directly to an antenna for use as a low-power transmitter, but most amateurs will wish to use it to drive a buffer or final amplifier. All stages following the exciter must be operated under Class A, AB, or B conditions. In general, the correct operating conditions for stages following the exciter may be found by referring to the audio operating conditions for the tube under consideration. Grid-bias and screen voltages should have very good regulation. For amateur voice operation, tubes may be operated considerably beyond the ratings given in the tube manuals, but such operation is beyond the scope of this article. When the r.f. amplifier is operated Class AB_2 or Class B, the grid tank circuit should be shunted by a resistor in order to provide better regulation of the exciting voltage. The value of this resistor is not critical and may be determined by experiment.

Operating Conditions

The operating conditions for the exciter are determined by the required output. If the required output is low, it is better to run the exciter with low plate voltages. This will reduce the amount of residual carrier present in the output in relation to the sideband output. Also, the exciter will be more stable and maintain adjustment longer with lower plate voltages.

A pair of 807s operating Class AB_2 can be driven by the exciter with only 60 ma. (at 120 volts) input to the balanced modulators, and with the exciter amplifiers also operating at 120 volts. Part of the output of the exciter is, of course, dissipated in the load resistor across the grid tank circuit of the 807s. The balanced modulators require sufficient r.f. drive to develop 12 volts of grid bias under these operating conditions.

With 400 volts applied to the balanced-modulator plates and 250 volts to other plate supply inputs, the operating currents will be approximately as follows:

Total balanced-modulator plate current	85 ma.
VR tube supply current	20 ma.
Total 6K6 amplifier current	62 ma.

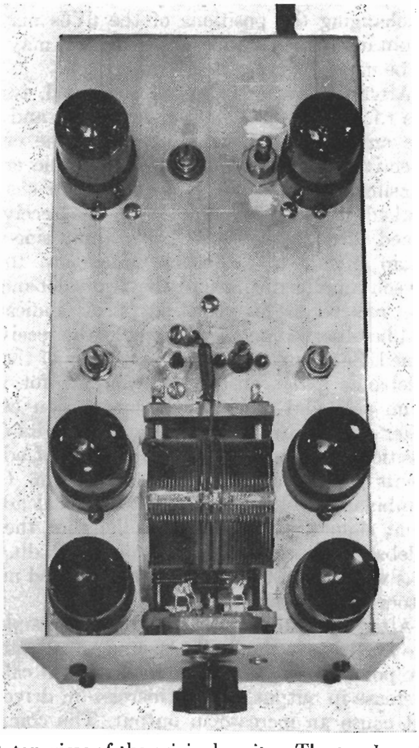
The total balanced-modulator grid current, measured at the bias terminals, will vary with excitation, but it should be in the range 3 to 5 ma.

These currents will not change appreciably with varying audio input and, with the exception of the grid current, will not change appreciably when the excitation is removed, provided that 4½ volts of fixed bias is used on the balanced-modulator grids.

The power input to the modulator plates should not exceed 30 watts with no audio input. If the d.c. operating voltages are removed when excitation is removed for stand-by, then no fixed bias is required on the balanced modulators and a jumper can be placed across the bias terminals. When excitation is removed with d.c. voltages applied, as in voice-controlled operation, then 4½ volts of fixed bias should be used to limit the plate and screen currents on the balanced modulators.

Adjustment

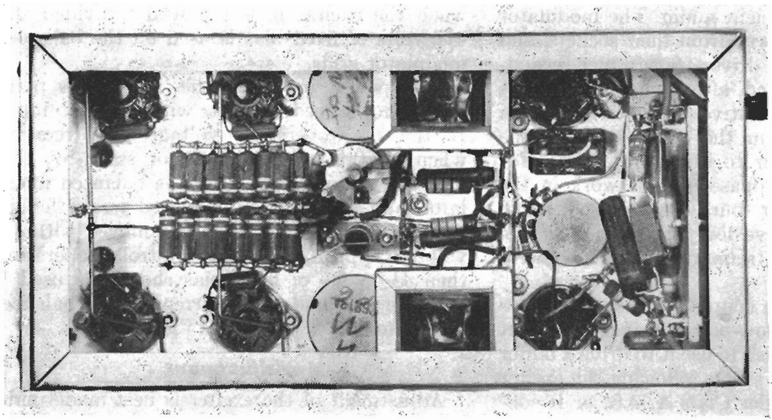
Adjustment of the exciter is best made under actual operating conditions. Connect the exciter to the amplifier, load the amplifier into a dummy



A top view of the original exciter. The toggle switch at the rear selects the sideband in use.

load, apply r.f. excitation to the exciter, feed a source of sine-wave audio into the speech amplifier, and tune the r.f. circuits in the conventional way for maximum output.

Reduce the audio input to zero, and adjust potentiometers R_{17} and R_{18} for minimum carrier output. Minimum carrier output may be determined by any sensitive r.f. indicator coupled to the final-amplifier plate circuit. A 0–1 milliammeter, in series with a crystal detector and a two-turn coupling loop, will make a satisfactory indicator. The meter should be by-passed with a 0.005-μf. condenser. If a null indication cannot



Bottom view of the original W2UNJ exciter. The audio phase-shift network is mounted on a panel at the right-hand side. The double string of resistors at the left is a load for the r.f. excitation; this does not appear in the circuit diagram since it is used principally to absorb excess driver power and will not be required in every case. Other components can easily be identified by reference to Fig. 2 and the text.

be obtained within the range of the potentiometers, the 6V6 tubes are not evenly matched. Exchanging the positions of the 6V6s may aid in obtaining the balance, or other tubes may have to be used.

After the carrier balance is obtained, tune in the r.f. source on the station receiver, and with the antenna terminals shorted and the crystal selectivity in sharp position, adjust the crystal phasing to the point where only one sharply-peaked response is obtained as the receiver is tuned through the signal. Now apply sine-wave audio of about 1500-cycle frequency to the speech amplifier, and find the two sidebands on the receiver. Three distinct peak indications will be observed on the S-meter as the receiver is tuned. Set the receiver on the weaker of the two sidebands and adjust L_1, C_7 and R_9 for minimum sideband strength. If suppression of the other sideband is desired, throw S_1 to its other position. A dip obtained with one set of adjustments is not necessarily the minimum. Other combinations should be tried. The final adjustment should give S-meter readings for the two sidebands which differ by at least 30 db. The bias voltages on the two pairs of balanced modulators will be approximately equal.

After the adjustments have been completed, the r.f. drive to the exciter should be adjusted to the point where a decrease in drive will cause a decrease in output, but an increase in drive will not cause an increase in output. The complete adjustment procedure should then be rechecked. The rig is then ready for a microphone, an antenna, and an on-the-air test.

If an oscilloscope is available, a simpler and more reliable adjustment procedure may be used. Either linear or sine-wave horizontal sweep may be used on the oscilloscope. The vertical input should be coupled to the output of the transmitter in the same manner as is used for observing amplitude modulation. The sine-wave audio-frequency input to the speech amplifier should be any convenient multiple of the oscilloscope sweep frequency. A 60-cycle sweep frequency and a 600-cycle audio frequency are commonly used.

When the exciter is modulated with a single sine-wave audio frequency, the output should be a single radio frequency. Therefore, the oscilloscope should show a straight-edged band across the screen, the same indication as is given by an unmodulated carrier. If carrier output, or unwanted sideband output, is present, it will be indicated by "ripple" on the top and bottom edges of the oscilloscope picture. A small amount of ripple can be tolerated, but if the exciter is badly out of adjustment the output will appear to be heavily modulated. Adjustment with the 'scope is accomplished by adjusting all controls to obtain the smallest possible amount of ripple. The oscilloscope may also be used for continuous monitoring during transmissions to avoid overloading of any stage of the transmitter. Overloading is indicated by flattening of the modulation-peak patterns at the top and bottom.

Speech Amplifier and Voice Control for the W2UNJ Exciter

The basic s.s.b. circuit described above can readily be combined with a speech amplifier and voice control circuit to make it a complete exciter in every respect except for an r.f. source. The latter, which is usually a VFO, is a separate unit in most station set-ups. The additional circuits required are shown in the schematic of Fig. 3, and the accompanying photographs show the revised model which incorporates Figs. 2 and 3 in one unit.

The speech amplifier is designed to attenuate both low and high frequencies, amplifying only the audio range required for good intelligibility. Its output is coupled to the input of the audio phase-shift network through a transformer with a center-tapped secondary, to provide push-pull audio for the phase-shift network.

The plate voltage for the speech amplifier must not be taken from the same point in the power supply that furnishes voltage for the 6K6 amplifiers, since interaction may occur that will upset the phase relationship at the output of the two 6K6s. If separate plate voltage sources are not

»

This version of the W2UNJ exciter uses the modulator-r.f. circuit of Fig. 2 in combination with the audio and voice control circuits of Fig. 3. The microphone input jack and audio gain control are at the left-hand end of the front wall of the chassis. The sideband reversing switch is toward the right.

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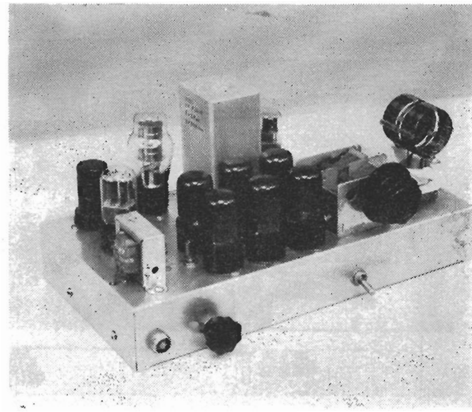


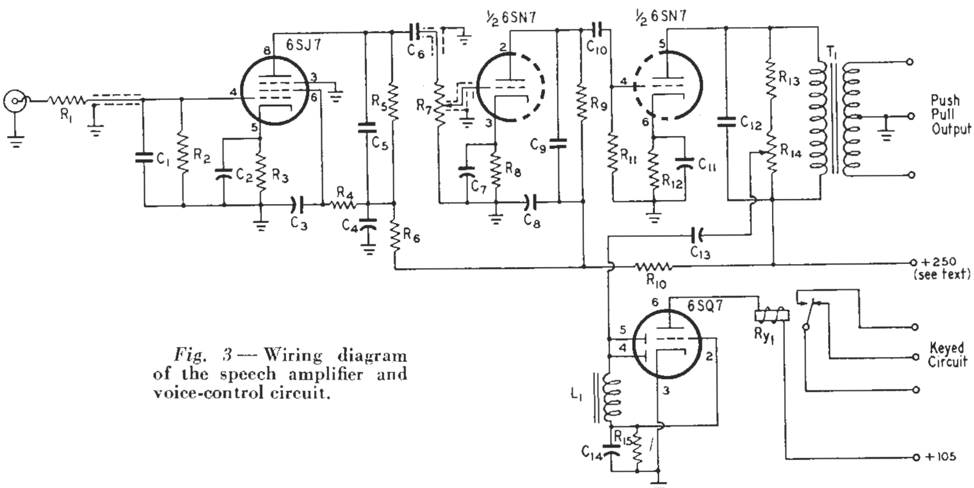


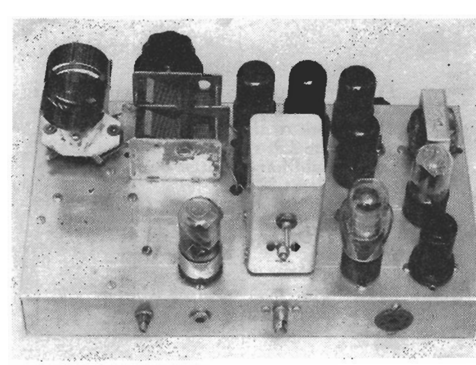
Fig. 3 — Wiring diagram of the speech amplifier and voice-control circuit.

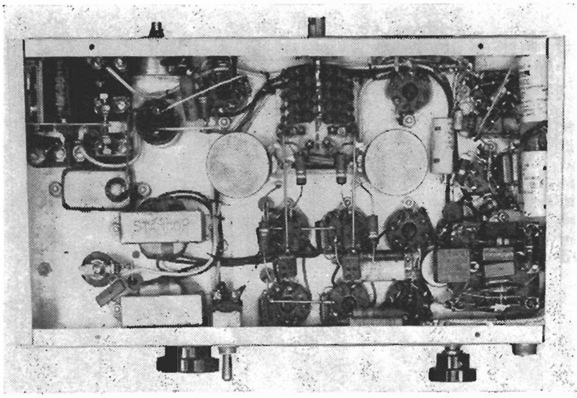
C_1 — 100-μμf. mica or ceramic.
C_2, C_7, C_{11} — 4-μf. 150-volt electrolytic.
C_3 — 0.02-μf. 400-volt paper.
C_4, C_8 — 8-μf. 450-volt electrolytic.
C_5 — 270-μμf. mica or ceramic.
C_6 — 0.001-μf. mica or ceramic.
C_9 — 0.0033-μf. mica or ceramic.
C_{10} — 0.002-μf. mica or ceramic.
C_{12} — 0.005-μf. ceramic or mica.
C_{13} — 0.01-μf. 400-volt paper or ceramic.
C_{14} — 0.5-μf. 200-volt paper.
R_1, R_9 — 0.1 megohm.
R_2 — 2.2 megohm.
R_3, R_{12} — 910 ohms.
R_4 — 1.0 megohm.
R_5 — 0.27 megohm.
R_6 — 27,000 ohms.
R_7 — 0.5-megohm volume control.
R_8 — 2700 ohms.
R_{10}, R_{13} — 10,000 ohms, 1 watt.
R_{11}, R_{15} — 0.47 megohm.
R_{14} — 15,000-ohm volume control.
All resistors ½-watt unless specified otherwise.
T_1 — 5-watt modulation transformer, 10,000 ohms c.t. to 4000 ohms (Stancor A-3812).
L_1 — Small filter or audio choke (Stancor C-1707).
Ry_1 — Sensitive 10,000-ohm relay.

»

A rear view of the complete exciter. The two r.f. phasing adjustments project from the shield can. The potentiometer shaft at the left sets the voice-control threshold level. The jack is for the keyed circuit, the r.f. connector takes the excitation cable, and the octal socket is for the power cable.

»





Underneath the exciter chassis. The two potentiometers are the bias balancing controls, R_{17} and R_{18}, Fig. 2.

available, an added filter section may be used to isolate the voltage to the speech amplifier.

Voice Control

Part of the output of the speech amplifier is taken off through an adjustable voltage-divider circuit, R_{13}, R_{14}, and blocking condenser, C_{13}, to the voice-control circuit. There it is rectified by the diodes of the 6SQ7, and the resulting d.c. voltage is used to charge C_{14} negative. An audio choke prevents audio components from appearing across C_{14}. The triode section of the 6SQ7 is normally conducting and holding the relay closed, but when the negative voltage appears across C_{14} the 6SQ7 plate current is cut off and the relay opens. When the audio signal is removed, C_{14} discharges through R_{15} and the triode again conducts, closing the relay.

The built-in voice-controlled relay can be used in a number of ways to provide the rapid voice break-in commonly used on 3.9-Mc. s.s.b. 'phone. If a good c.w. break-in system is already in use at the station, the voice-control relay contacts may be substituted for the key, and no other changes are necessary.

If the local oscillator in the receiver will key in the plate voltage lead satisfactorily, then a simple voice break-in system may be obtained by using the relay contacts to shift the plate voltage from the receiver local oscillator to the VFO. A drifting receiver oscillator must be avoided in this system, however.

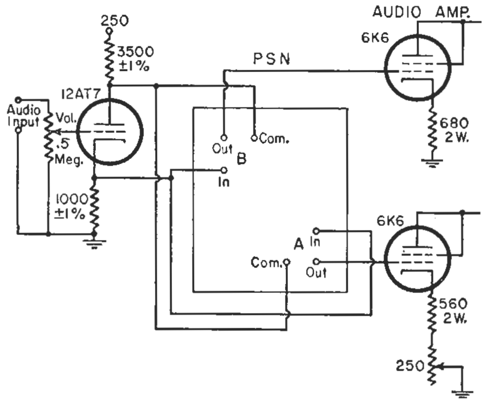
Construction

The exciter and its associated audio equipment are assembled on a 13 × 7 × 2-inch aluminum chassis. The four 6V6 balanced-modulator tubes are arranged in a square pattern toward the front center of the chassis, with the plate tuning condenser and coil off to one side and the 6K6 audio amplifier tubes on the other. The two modulation tranformers are under the chassis directly below the plate tuning condenser. The speech amplifier is arranged along the left-hand side of the chassis, with the 6SJ7 at the rear and the output transformer on the top of the chassis at the front. The audio phase-shift network is below the output transformer.

The reactive components of the r.f. phasing network, L_1 and C_7, are mounted in a plug-in shield can that mounts directly behind the balanced-modulator tubes. The shield can is grounded to the chassis through the spare pins of its plug. The voltage regulator tube is mounted to the left of the shield can, and the 6SQ7 voice-control tube is to the right. The components in the voice-control circuit are mounted under the chassis at the rear.

COMMERCIAL PHASE-SHIFT NETWORK WITH THE W2UNJ EXCITER

Fig. 4 — Schematic diagram giving circuit revisions for including a commercial phase-shift network in the s.s.b. exciter of Fig. 2, page 97. The 250-volt supply for the 12AT7 section should have a large output-capacity filter of at least 80 μf.
PSN — Phase-shift network (Millen 75012 Phasing Unit).



» 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, 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 demodulate 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.s.c. 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.s.c. '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

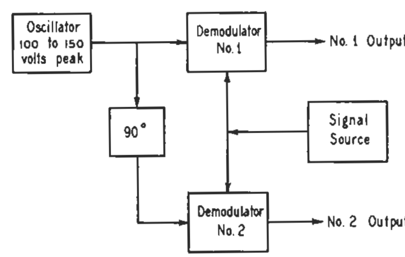


Fig. 1 — The elements of a dual exalted-carrier demodulator.

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 circuit 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

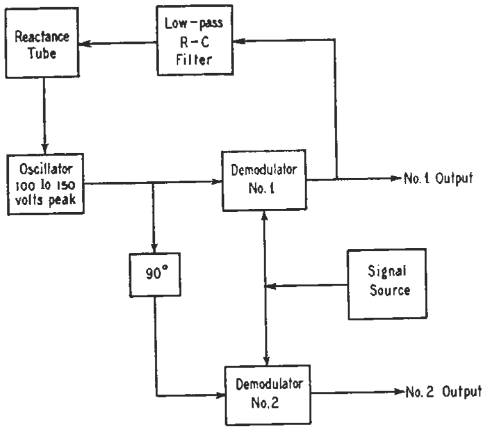


Fig. 2 — A block diagram of a dual exalted-carrier demodulator with automatic carrier synchronization. Such a system can be used to "lock in" on a small amount of carrier sent along with an s.s.s.c. signal, or it can be used for single-sideband (or double-sideband) reception of conventional a.m. or p.m. signals.

the oscillator, a plus-90° phase relationship is produced; if the 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 set-up.

Use can be made of the d.c. output of one of the demodulators to control the frequency of the oscillator so that it will synchronize automatically with a selected component of the input signal. If this selected component is the carrier of a 'phone signal, the sidebands will be demodulated against the synthetic carrier acting as a substitute for the real carrier. The block diagram of Fig. 2 illustrates an exalted-carrier demodulator that has an automatic synchronizing arrangement in addition to the elements shown in Fig. 1. Fig. 3 is a schematic diagram of such a circuit.

Single-Sideband Receiving System

The block diagram of Fig. 4 illustrates a single-sideband receiving system that employs the exalted-carrier demodulator of Figs. 2 and 3 driving phase-shift networks that have the property of 90° differencial 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. 5 will be helpful in understanding the action.

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 fre-

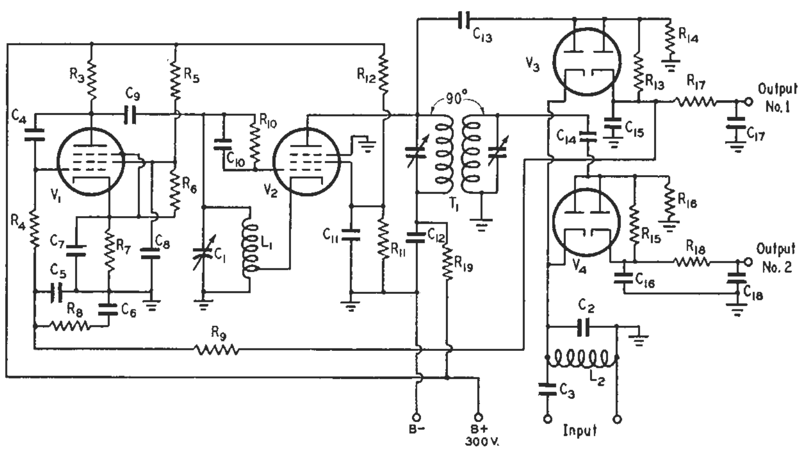


Fig. 3 — The schematic diagram of a dual exalted-carrier demodulator. Values of resistors and capacitors are ±20% unless otherwise specified.

C_1 — 5- to 25-$\mu\mu$f. air trimmer.
C_2 — 0.0022-μf. mica or ceramic.
C_3, C_7, C_8, C_{11}, C_{12} — 0.01-μf. paper.
C_4 — Approx. 5-$\mu\mu$f. mica or ceramic.
C_5 — 0.1-μf. paper.
C_6 — 1.0-μf. paper.
C_9 — 0.001-μf. paper, mica or ceramic.
C_{10} — 100-$\mu\mu$f. mica or ceramic.
C_{13}, C_{14}, C_{17}, C_{18} — 47-$\mu\mu$f. mica or ceramic, ±5%
C_{15}, C_{16} — 470-$\mu\mu$f. mica or ceramic, ±5%.
R_3, R_{10} — 0.1 megohm.
R_4 — 470 ohms.
R_5, R_{12} — 62,000 ohms, 2 watts.
R_6, R_{11} — 33,000 ohms, 2 watts.
R_7 — 820 ohms.
R_8 — 10,000 ohms.
R_9 — 2.2 megohms.
R_{13}, R_{14}, R_{15}, R_{16} — 0.24 megohm, ±5%.
R_{17}, R_{18} — 56,000 ohms, ±5%.
R_{19} — 3300 ohms.
All resistors ½-watt unless otherwise mentioned.
L_1 — 2.3-millihenry choke. (National R-100 tapped between first and second sections.)
L_2 — 50-microhenry coil, approx. (to resonate with C_2 at intermediate frequency).
T_1 — I.f. transformer with coupling adjusted to provide equal primary and secondary voltages.
V_1 — 6SJ7 reactance tube.
V_2 — 6SJ7 oscillator.
V_3 — 6H6 demodulator No. 1.
V_4 — 6H6 demodulator No. 2.

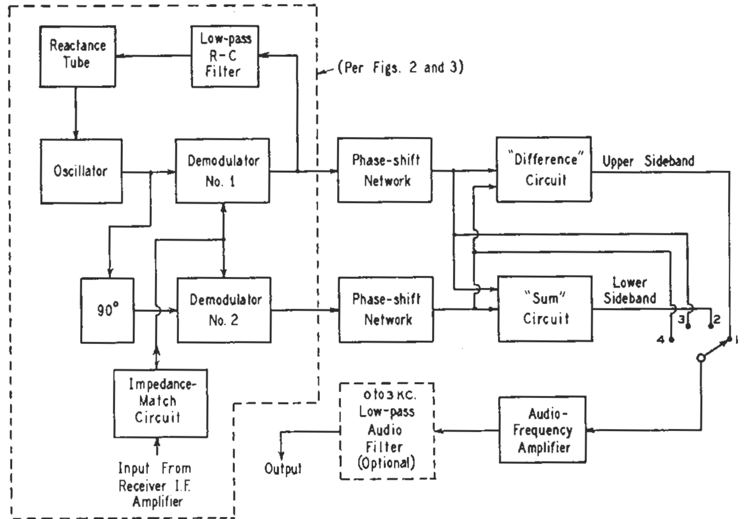


Fig. 4 — A block diagram of a single-sideband receiving system incorporating exalted-carrier demodulation.

quency 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

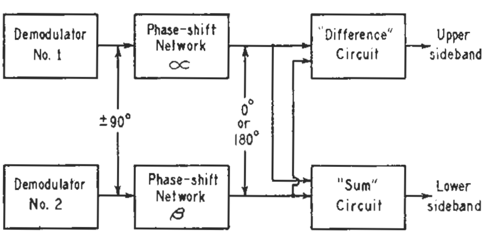


Fig. 5 — The single-sideband selection is obtained by coöperative action of two demodulators and phase-shift networks.

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 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. 6. 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. 4 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 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. The use of the filter (indicated by dashed lines in Fig. 4) is optional. Its purpose will be explained later.

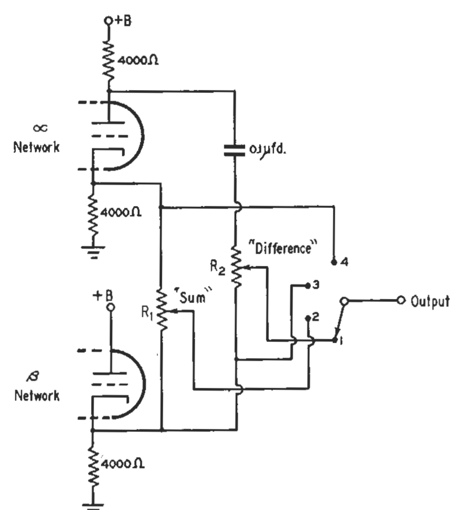


Fig. 6 — The method of connecting the "sum" and "difference" circuits to the phase-shift networks. In positions *1* and *2*, the switch gives output from one sideband or the other, while positions *3* and *4* give demodulated double-sideband outputs.

It is important to drive the demodulators of Fig. 4 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. 4 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. 7A and 7B, 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. 4. 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 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

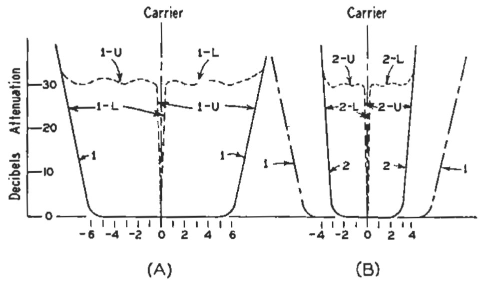


Fig. 7 — **Over the range for which the audio phase-shift networks hold close to 90° difference, the apparent i.f. response of the system is determined (in combination) by the i.f. and audio bandwidths. The characteristic with no audio filter is shown at (A), and (B) demonstrates how greater effective selectivity is obtained with an audio filter.**

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. 7B. 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.

It can be appreciated that single-sideband reception of double-sideband signals offers substantial opportunity to avoid interference by choice of a sideband region least contaminated by QRM. The signal-to-noise ratio suffers with this mode of operation because only one of the transmitted sidebands is heard. This is not the fault of the receiving system, however. The real fault is that double-sideband transmission uses twice as much bandwidth as is necessary and has only half of its initially-small "communication" power in each sideband. In addition, carrier heterodynes are the greatest source of QRM. Single-sideband transmissions overcome most of the problems of operating in congested bands, even though there is no alternate choice of sideband possible at the receiver. Such transmissions reduce the carrier-heterodyne problem to the vanishing point.

Getting maximum performance from an audio phasing network requires that the signal source have low impedance. This same requirement applies in the case of the plate power supply for the tube or tubes feeding the network. The output filter condenser of the supply should be large — 40 μf. or more.

» Aside from the heterodyne beat note caused by the carrier, amplitude-modulated 'phone signals do not, on the average, cause serious interference with single-sideband reception. The circuit described here is simple to build, requires no changes in the receiver to which it is connected, and will balance out a heterodyne with little effect on the desired signal.

A Heterodyne Eliminator

OSWALD G. VILLARD, JR., W6QYT, AND JOSÉ MIGUEL DÍAZ, XE1RZ

This article describes an improved version of the "Select-o-ject" in which only one connecting lead to the associated receiver or audio amplifier is necessary to achieve rejection of a particular audio frequency. Connection to the amplifier is made by wrapping a wire around the plate pin of a low-level stage, and by providing a common ground. *No modification whatever of the wiring of the amplifier is required.* Use is made of "constant percentage bandwidth" operation, which is desirable for rejecting heterodynes or beat notes in 'phone reception because this

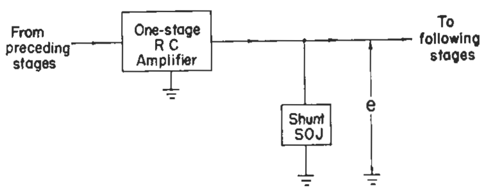


Fig. 1 — Block diagram showing how shunt SOJ is connected to existing low-level audio amplifier.

method provides the narrowest obtainable "notch" and thereby removes as little of the desired intelligence as possible.

Principles of Operation

The basic idea behind the operation of the shunt SOJ will be briefly set forth. Fig. 1 shows an ordinary one-stage resistance-coupled audio amplifier, which could be the first audio stage of a communications receiver, across whose output terminals the shunt SOJ is connected.

Now it is well known that such an *R-C* amplifier can be represented, insofar as a.c. quantities are concerned, by the equivalent circuit shown in the left-hand half of Fig. 2. The input, in reality connected to a high-impedance grid and therefore essentially floating above ground, is shown in the diagram as a wire terminated in midair. The action of the amplifier may be explained in terms of an equivalent generator acting in series with an equivalent plate circuit resistance. (The voltage of this generator is, of course, directly related to the voltage applied to the grid of the tube.)

The shunt SOJ essentially consists of another one-stage audio amplifier connected in parallel with the first, as shown in the right-hand half of Fig. 2. This SOJ amplifier may be represented by the same equivalent circuit. For the moment, the question of where the SOJ amplifier's grid signal comes from will be ignored. Let it be assumed that the SOJ's equivalent generator has a voltage of the opposite phase to that of the amplifier's equivalent generator. The a.c. voltages in the circuit can then be replaced for convenience by d.c. voltages, and the generators assigned polarities just as if they were d.c. generators.

Suppose further that the equivalent voltages e_1 and e_2 are equal, and that the equivalent resistances R_1 and R_2 are equal. (This can always be arranged, if desired.) If the generator polarities are opposite, as indicated in the figure, a little thought will show that the output voltage e will then be zero. The generators are acting in series, and a current will be flowing through the resistors, but the two resistors taken together form a voltage divider, and their midpoint is at ground potential. Since there is no voltage across the amplifier's output terminals, they could be short-circuited without changing currents or voltages in any way. So far as the amplifier can tell, connection of the SOJ disposed as shown is

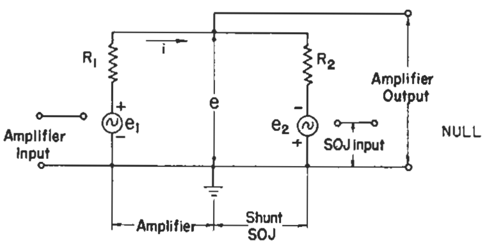


Fig. 2 — Equivalent circuit for null condition.

the equivalent of placing a short circuit across the amplifier's output terminals.

If R_1 and R_2 are not equal, it is still possible to cause the net output voltage to be zero, by adjusting the relative magnitude of the two generator voltages. Thus if R_2 were twice R_1, the output null could be restored by making e_2 twice e_1.

Now comes the question of providing the SOJ tube's grid with a suitable input voltage, so that the right-hand equivalent generator of Fig. 2 will have a voltage of the correct magnitude and

From "The Shunt Select-o-ject," *QST*, October, 1952.

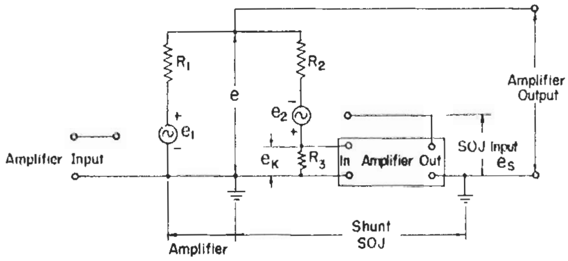


Fig. 3 — Equivalent circuit for null condition, showing derivation of SOJ amplifier grid voltage.

phase to cause a null. It might at first seem hard to find a usable voltage, since the object is to cause the amplifier's output voltage e to disappear. Fortunately, even when e disappears there is still a current flowing around the loop, and a voltage proportional to this current (e_k) can be developed across R_3 in Fig. 3. Note that e_k is available *inside the SOJ.* This voltage, properly amplified, given the correct phase, and fed back to the SOJ tube's grid, will serve to make the equivalent generator voltage e_2 have just the right value to cause e to disappear and produce a null.

Practical Circuits

Fig. 4 shows how the SOJ's tube (V_2) is connected to the amplifier in practice. To produce a null, the cathode voltage e_k is amplified without phase change and applied between the grid of V_2 and ground. This voltage will be called e_s.

Since it is desired that the null occur at only one frequency, the amplifier connecting e_k and e_s must be frequency-sensitive. It is convenient to make the frequency-sensitive portion of this circuit a variable all-pass *R-C* phase-shifter. A complete schematic of the SOJ so connected as to produce a null is shown in Fig. 5. The phase-shift circuit uses ganged 5-megohm linear potentiometers as the frequency-controlling elements. The 120K resistor limits the highest "resonant" frequency to 6600 cycles. The low-frequency limit of response is 160 cycles. V_5 in this circuit provides isolation and phase reversal.

When the SOJ is used to produce a null, an important point is to keep its insertion loss low by making the plate resistance of V_2 in Fig. 4 high, and likewise R_L. The SOJ then affects the gain of V_1 only to a minor extent, at frequencies far from that at which the null occurs.

The shunt SOJ draws approximately 7 ma. at 150 to 200 volts, so can be powered from an existing receiver or amplifier, if desired. Whatever power supply is used *must* have a large (order of 20 μf.) filter condenser directly across its output terminals. This is important.

The performance of the shunt SOJ will vary slightly depending on the internal impedance

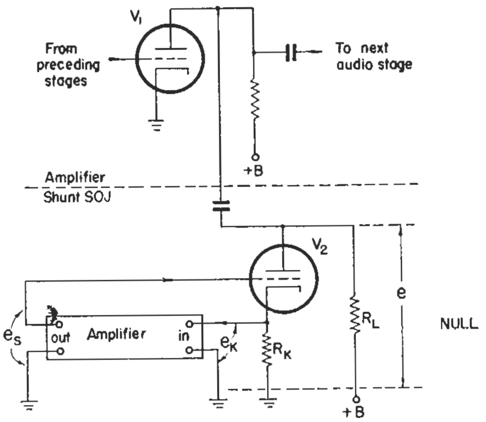


Fig. 4 — Detail of SOJ amplifier connection, null condition.

of the circuit to which it is connected. For most receivers this will vary from 50K to about 200K. The circuit of Fig. 5 will give satisfactory performance over this range of impedances.

In constructing the unit no special precautions are necessary other than to minimize hum, both from the power supply and by direct pick-up from the filament leads, as much as possible. It is suggested that the tube types and component values be adhered to as closely as possible. It is desirable to provide good filtering in the B supply and to twist the filament leads and place them close to the chassis.

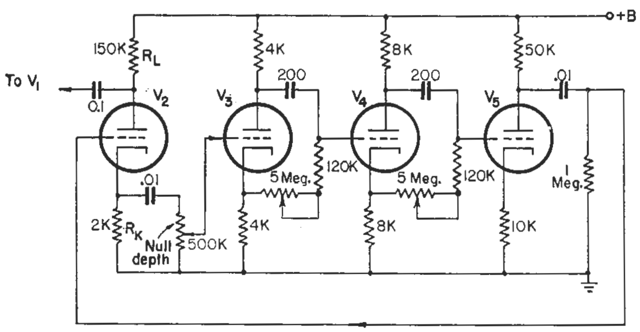


Fig. 5 — Actual circuit of shunt SOJ for null condition. Tubes are 12AU7. Capacitance values below 0.001 μf. are in $\mu\mu$f.; higher values in μf. Resistors are ½ watt. The 5-megohm potentiometers should be ganged.

» Good linearity in an r.f. amplifier is based on a nice balance among several factors, including plate voltage, bias voltage, r.f. driving voltage, and plate loading. This is an outline of the design procedure.

Linear Amplifier Design

STYRK G. REQUE, W2FZW

A LINEAR AMPLIFIER is one in which the output voltage is proportional to the input voltage. All of our audio amplifiers are of this type, or we get very objectionable distortions. Similarly the r.f. and i.f. amplifiers of our receivers are linear r.f. amplifiers, for if there were any serious distortion of the modulation envelope the detector would give us a distorted output signal. In fact, any amplification of a signal with a modulation envelope must be linear if we are to be able to recover the modulation in a detector system without severe distortion.

The simplest form of linear amplifier (r.f. or audio) is the Class A amplifier, which is used almost without exception throughout our receivers and our low-level speech equipment. While its linearity can be made phenomenally good, it is unfortunately quite inefficient. The theoretical limit of efficiency in this case is 50 per cent, while most practical amplifiers run 25–35 per cent efficient at full output. At low levels this is not worth worrying about, but when we exceed the 2- to 10-watt level something else must be done to improve this efficiency and reduce tube, power-supply and operating costs.

The use of Class B amplifiers for high-level audio amplifiers is now well known and common amateur practice. Class B amplifiers are theoretically capable of 78.5 per cent efficiency at full output, and practical amplifiers run at 60–70 per cent efficiency at full output. The same amplifier tubes, with suitable tank circuits substituted for the driver and output transformers, will make good linear r.f. power amplifiers of the same power rating and efficiency. In fact, we can even generalize this and make the following statement: Any reasonably distortion-free audio amplifier may be converted to a linear r.f. amplifier by replacing the input and output transformers with properly designed and loaded r.f. tank circuits, provided, of course, that the tubes are suitable for use at the desired frequency. In r.f. circuits running Class B, only one tube need be used if only half the power is wanted, because the flywheel action of the tank circuits will smooth out the missing half cycle.

One side issue is well worth considering at this moment. If you look up the Class B *r.f.* amplifier ratings of a given tube, you will undoubtedly be shocked to see that the efficiency given is in the order of 33 per cent and not the 60–70 per cent quoted above. The discrepancy arises because the figures given are for a conventional a.m. system. The efficiency of a Class B amplifier is proportional to the signal voltage; i.e., at full output it is 60–70 per cent and at half voltage it is 30–35 per cent. In a conventional a.m. system the carrier is always at half voltage, and so when no modulation is applied the efficiency of a properly-adjusted Class B r.f. amplifier will be in the order of 33 per cent. This need not concern the amateur running a single-sideband system with suppressed carrier, since his resting or no-modulation condition corresponds to zero signal input to the amplifier and he observes only the small resting d.c. input to the amplifier.

Amplifier Design

In most cases the design of a Class B linear amplifier will be rather simple, since most of the common power-amplifier tubes are rated for Class B audio work. In a case of this sort the proper plate voltage, bias voltage, load resistance and power output are given, and the sole job is to provide proper tank circuits and drive for the tubes. As an example, let us choose a tube of good reputation as a Class B audio amplifier, such as the GL-805. Typical operating conditions are given in Table I.

TABLE I
Class B Audio-Amplifier Data GL-805 Tubes
Values given for two tubes

D.c. plate voltage	1250 volts
D.c. grid voltage	0 volts
Peak grid-to-grid voltage	235 volts
Zero-signal plate current	148 ma.
Max.-signal plate current	400 ma.
Max.-signal driving power	6 watts
Max.-signal plate input	500 watts
Effective load plate-to-plate	6700 ohms
Max.-signal power output	300 watts

Fig. 1 is a schematic diagram of the usual Class B audio amplifier. Fig. 2 is a diagram of the amplifier changed over for use as a linear r.f. amplifier. Our first concern will be the design of the proper tank circuits for the grid and plate circuits. The subject of proper loading will be discussed under the section on practical adjustment.

Let us design the proper plate-tank circuits first. As in all r.f. amplifiers, this tank circuit

From "Linear R. F. Amplifiers," *QST*, May, 1949.

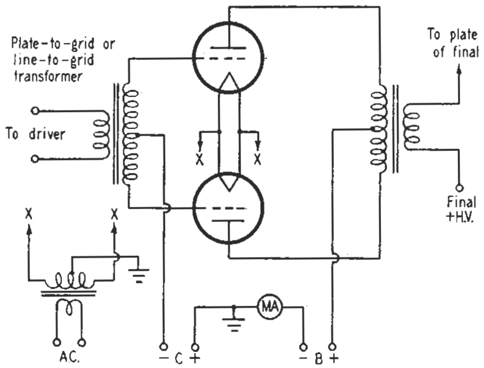


Fig. 1 — A circuit diagram of the familiar Class B modulator.

should have a loaded Q of at least 12, if we want to have reasonable efficiency and low harmonic output. The loaded Q is defined in terms of the tank-capacitor reactance (equal to the tank-inductance reactance at resonance) and the load resistance by the equation

$$Q = R_L/X_C \quad (1)$$

Rearranging, and substituting in the figures,

$$X_C = R_L/Q \quad (2)$$
$$= 6700/12 = 560 \text{ ohms}$$

But we also know that

$$X_C = 1/2\pi fC \quad (3)$$

If we choose the 75-meter 'phone band as our example of design, and hence substitute 4 Mc. for f and 560 ohms for X_C in (3), we will find the value for C to be approximately 70 $\mu\mu$f. This is the value of the capacity across the tank, and we must double it to find the value for each section of our split-stator condenser, or 140 $\mu\mu$f. per section. Note that this is the value of the capacity actually in use, and that for proper adjustment a capacitor with a rating of at least 150 (and preferably 200) $\mu\mu$f. per section would be clearly indicated. The coils should be chosen or pruned until the proper amount of capacity is required to tune them to resonance, with the error if any on the low-inductance (high-capacity) side where it can do little harm. Many troubles in amateur transmitters can be traced to the use of too little capacity in the r.f. tank circuits. This is not a peculiarity of the Class B linear amplifier, but is equally true of the Class C, perhaps to an even greater degree.

The calculation of the grid tank circuit is performed in just the same way as we calculated the plate tank. However, the loading of the grids, which must be substituted for R_L, is not given. Our present example, GL-805s, involves a pair of zero-bias tubes. Tubes in this class draw grid current even when very small signals are applied, and the equivalent loading of the grid tank is very nearly constant regardless of signal level. This will mean that a nearly constant load will be reflected to the driving stage and only a small amount of loading or "swamping" will be necessary to insure that the driving signal is not distorted.

If, on the other hand, we choose tubes that operate at a normal bias of 50–60 volts (such as GL-810s) it is apparent that the grids will not draw any current at all until the driving signal exceeds this bias. In a case of this sort the grids load the grid tank circuit, and hence the driving stage, in a variable manner. Unless some further step is taken, this will result in distortion of the driving signal, and our amplifier system is not linear. This can be avoided if sufficient fixed loading is supplied for the driver stage, and if suitable impedance matching is done so that the variable grid loading is negligible. In any case, this will require that the driver be capable of supplying several times the listed value of grid driving power. A full discussion of the possible ways of impedance matching and controlling this variable grid loading is unfortunately far beyond the scope of this article. However, as a guide to those who care to delve into the subject, we can state that the necessary conditions which must be satisfied are two in number:

a) The load presented to the driving stage must be constant.

b) The voltage at the grids must have good regulation.

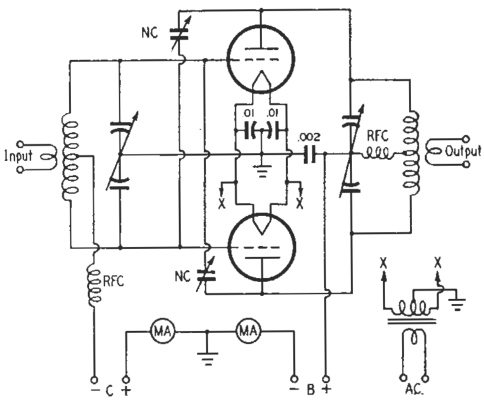


Fig. 2 — A Class B linear r.f. amplifier circuit resembles Fig. 1 but uses r.f. instead of audio components.

Returning now to our original example, the GL-805s, we can calculate the loading effect of the grids from the known grid-driving power and known peak grid-to-grid voltage by means of the simple formula,

$$R_{EQ} = E^2_{G-G}/2P_G \quad (4)$$

Substituting the proper values of grid-to-grid voltage and grid driving power from the date in Table I gives an equivalent grid loading of 4600 ohms. To be conservative, we might well put a 5000-ohm damping resistor across the tank, so that the net effective resistance across the tank will be approximately 2400 ohms. Substituting this value in Equation (2), and the resultant value of reactance in (3), we find the necessary value of C to be 200 $\mu\mu$f. across the tank or a split-stator capacitor of 400 $\mu\mu$f. per section. A broadcast-receiver condenser of 420 $\mu\mu$f. per section is readily available and will easily stand the low peak voltage on the grids.

Here again it may not be amiss to mention that the large value of capacity indicated is not a result of Class B operation, but in this case is purely a function of the tube chosen. For linear amplifiers it is necessary that the tanks be properly designed. If the Class C stage seems to be tolerant of errors in tank design, it is because few of us have given full consideration to the proper handling of our amplifiers and have been content to operate with the efficiency and the harmonic output accident has provided.

Design from Tube Curves

There are actually very few power tubes which we might choose to use as linear amplifiers that do not carry a Class B audio rating. However, there are a few tubes, designed for v.h f. use, which are not so rated. If we care to use a tube of this sort we must determine the quantities equivalent to those given in Table I from the characteristic curves and a few formulae which have been worked out by the engineers. Fig. 3 shows the grid and plate characteristics of such a tube, the GL-829-B.

As a first assumption, let us suppose that the plate-supply voltage is 500 volts. The proper bias is our next consideration, and one good rule of thumb in determining this is to choose the bias such that the resting plate current will produce approximately ⅓ rated plate dissipation. Bias determined in this way will usually allow better linearity (less distortion) than a bias chosen closer to complete cut-off. Since our GL-829-B has a rated dissipation of 20 watts per section, the proper bias will allow 6.7 watts resting dissipation per section. At 500 volts plate supply this means a resting plate current of 13.3 ma. per section. From the characteristic curves of Fig. 3 it will be seen that approximately 18 volts of bias will be required. The resting point is marked "A" in Fig. 3.

Since the maximum signal efficiency is going to be in the order of 65 per cent, we can now determine the maximum input power. The 35 per cent power loss must equal the maximum plate dissipation, which is 40 watts (both sections) for our GL-829-B. Therefore, the total power input (100 per cent) must be 115 watts maximum, or 57.5 watts per section.

Dividing the maximum power input by the plate voltage will give the maximum signal plate current. In this case the total current will be 115/500 = 0.230 amp. = 230 ma. This is 115 ma. per section d.c. plate current at maximum signal.

The plate-current pulses of each tube of our Class B linear amplifier are half sections of a sine wave, such as might have been produced by a half-wave rectifier. In such a waveform, the peak current is 3.14 times the value read by a d.c. meter, and this permits us to find the peak current flowing through the tube. Since the d.c. input per section is 115 ma., we know then that the peak current through each section should be 115 × 3.14 = 360 ma.

Returning to Fig. 3, we see that 360 ma. will flow on the crest of the cycle if the grid is driven up to +10 volts on the peak and the plate is not allowed to swing lower than 75 volts. Since the grid starts from −18 volts (the bias), this will be a peak r.f. grid swing of 28 volts, or a peak grid-to-grid voltage of 56 volts.

The grid driving power may be calculated from the peak grid-to-grid voltage and the grid current that will flow at the operating point "B." This is marked as "B" on the grid current curves shown in Fig. 3. The grid driving power is one-quarter [1] of the product of this peak grid current and the peak grid-to-grid voltage, or 0.39 watt in this case.

The power output of this amplifier may now be calculated by the aid of the formula

$$P = 0.78\,(E_B - E_{P\,min})\,I_{dc\,max} \qquad (5)$$

Substituting the value of minimum plate voltage, the plate-supply voltage and the maximum-signal d.c. plate current we find the output power to be 0.78 (500–75) 0.23 = 76 watts.

As a double check we subtract this from the power input of 115 watts and find 39 watts plate dissipation for both sections. The actual efficiency is 66 per cent, a bit higher than assumed at first.

The plate-to-plate load resistance is readily ob-

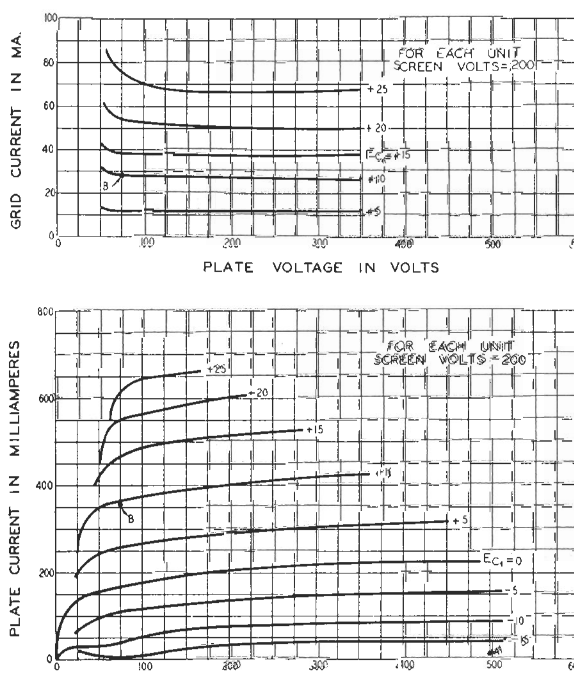


Fig. 3 — **Average grid and plate characteristics of the GL-829-B.**

[1] Approximate value commonly used for design purposes.

tained from the formula:

$$R = 2.6\ (E_B - E_{Pmin})\ /I_{dc.\ max} \qquad (6)$$

Substituting the same values used with Equation (5), we find the plate-to-plate load resistance to be 2.6 (500–75)/0.23 = 4800 ohms.

Collecting all the values calculated, we can now make up a table similar to the one given for the GL-805s which will apply to the GL-829-B. This is shown in Table II.

The calculation of the specific amplifier will now be the same as the case of the GL-805s, since we have determined all the significant values.

TABLE II

Class B Audio or Linear R.F. Amplifier Data—GL-829-B

Values given for both sections

D.c. plate voltage	500 volts
D.c. grid voltage	—18 volts
Peak grid-to-grid voltage	56 volts
Zero-signal plate current	27 ma.
Max.-signal plate current	230 ma.
Max.-signal driving power	0.39 watts
Max.-signal plate input	115 watts
Effective load plate-to-plate	4800 ohms
Max.-signal power output (audio or peak r.f.)	76 watts

General Considerations

If proper operation is expected, it is essential that the amplifier be so constructed, wired and neutralized that no trace of regeneration or parasitic instability remains. Needless to say, this also applies to the stages driving it.

The bias supply to the Class B linear amplifier should be quite stiff. The Class C stage thrives on grid-leak bias, but for really good operation the Class B linear should be supplied from a very stiff source, such as batteries or some form of voltage regulator. If nonlinearity is noticed when testing the unit, the bias supply may be checked by means of a large electrolytic capacitor. Simply shunt the supply with 100 µf. or so of capacity and see if the linearity improves. If so, rebuild the bias supply for better regulation. *Do not rely on a large condenser alone.*

POWER-SUPPLY REGULATION

The necessity for good regulation of the output voltage of the plate power supply is stressed by every author who writes on linear amplifiers. Two kinds of regulation must be considered — "static" and "dynamic." Static regulation is determined by measuring the change in output voltage under different values of steady loads, and can be found from measurements made with ordinary d.c. instruments. A choke-input filter of good design will lead to quite satisfactory values of static voltage regulation.

"Dynamic" regulation is a term used to describe the transient variation in d.c. output voltage with a rapidly-varying load. When a load is applied suddenly, there is an interval during which practically the only energy available for it is what is stored in the output filter condenser, since the current cannot change rapidly through the filter choke. There may, therefore, be a momentary drop in output voltage of proportions far exceeding the drop that would be expected from the static regulation curve. It is not unusual for the output voltage to drop to less than half its no-load value when full load is suddenly applied, as shown by W2KUJ in *G. E. Ham News,* Vol. 9, No. 1. The recovery time may easily exceed the duration of the voice peak that caused the transient. Thus the power output is limited and the modulation envelope is distorted. This dynamic behavior will not show up on the d.c. meter readings and can only be checked with an oscilloscope.

Good dynamic regulation requires a large capacity in the filter output condenser. The value needed depends on the extremes of load variation with the particular linear amplifier used. Class A operation imposes no special requirements since the load current is the same regardless of r.f. output level. With other classes of operation it may be said that the required capacity (over that needed for adequate hum filtering according to usual design practice) increases with the percentage change in plate current from no-signal to maximum-signal conditions. With Class B, which swings from a very low no-signal current to a large maximum-signal current, experience has shown that the minimum output capacitance in the filter should be of the order of 40 µf., and values up to 100 µf. usually will show continuing improvement.

It is advantageous to operate with the largest possible value of no-signal plate current — up to the limit of the plate dissipation capabilities of the tubes — and thus minimize as much as possible the current variation with voice operation.

LINEAR-AMPLIFIER TUBE-OPERATION DATA FOR SINGLE SIDEBAND

Except where otherwise noted, ratings are manufacturers' for audio operation. Values given are for one tube. Driving powers represent tube losses only—circuit losses will increase the figures.

Tube	Class	Plate Voltage	Screen Voltage	D.C. Grid Voltage	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. Avg. Grid Current	Max.-Sig. Avg. Driving Power	Max.-Rated Screen Dissipation	Max.-Rated Grid Dissipation	Avg. Plate Dissipation	Max.-Sig. Useful Power Output
2E26	A_1	250	200	−14	35	42	7	10	14	—	0	2.5		10	5
	AB_2	400	125	−15	10	75		16	30		.2	2.5		10	21
		500	125	−15	11	75		16	30		.2	2.5		12.5	27
6146	AB_1	600	200	−50	26	120	.6	13	50	—	0	3		25	47
		750	200	−50	29	114	.5	14	50	—	0	3		25	60
	AB_2	600	185	−50	21	135	.5	15	57	.4	.02	3		25	58
		750	165	−45	18	120	.3	11	51	.4	.02	3		25	65
807 1625	AB_2	600	300	−30	30	100	.4	6	39		.1	3.5		25	40
		750	300	−32	26	120	.3	8	46		.1	3.5		30	60
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
4-65A	AB_1	1500	480	−105[1]	30	90 (70)[4]	—	13 (4.2)[4]	105	0	0	10			75
		2000	450	−100[1]	22	80 (60)[4]	—	11 (3.0)[4]	100	0	0	10			100
		2500	405	−90[1]	17	70 (50)[4]	—	8.5 (2.5)[4]	90	0	0	10			115
	AB_2	1000	250	−30[1]	30	150	0	23	105		2.5	10	5	65	85
		1500	250	−35[1]	30	125	0	15	100		1.6	10	5	63	125
		1800	250	−35[1]	25	110	0	13	90		1.1	10	5	63	135
	B[2]	1500	300	−50[1]	33	200	0	35[3]	190	13	2.4	10	5	60	150
		2000	400	−75[1]	25	270	0	50[3]	270	17	4.6	10	5	65	300
		2500	500	−100[1]	20	230	0	35[3]	300	6	1.8	10	5	65	325
813	AB_2	2000	750	−90	20	158	.8	29	115		.1	22		100	228
		2250	750	−90	23	158	.8	29	115		.1	22		100	258
		2500	750	−95	18	180	.6	28	118		.2	22		125	325
4-125A	AB_1	2000	615	−105[1]	40	135 (100)[4]	—	14 (4.0)[4]	105	0	0	20			150
		2500	555	−100[1]	35	120 (85)[4]	—	10 (3.0)[4]	100	0	0	20			180
		3000	510	−95[1]	30	105 (75)[4]	—	6.0 (1.5)[4]	95	0	0	20			200
	AB_2	1500	350	−41[1]	44	200	0	17	141	9	1.25	20	5	125	175
		2000	350	−45[1]	36	150	0	3	105	7	.7	20	5	125	175
		2500	350	−43[1]	47	130	0	3	89	6	.5	20	5	122	200
4-250A	AB_1	2500	660	−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[1]	50	243	0	17	96	11	1.1	35	10	150	214
		2000	300	−48[1]	60	255	0	13	99	12	1.2	35	10	185	325
		2500	300	−51[1]	60	250	0	12	100	11	1.1	35	10	205	420
		3000	300	−53[1]	63	237	0	17	99	10	1	35	10	190	520
304TL	AB_1	1500	—	−105	135	286	—	—	105	—	0	—		300	128
		2000	—	−160	100	273	—	—	160	—	0	—		300	245
		3000	—	−260	65	222	—	—	260	—	0	—		300	365

[1] Adjust to give stated zero-signal plate current.
[2] Single-sideband suppressed-carrier linear amplifier ratings, voice signal.
[3] Due to intermittent nature of voice, average dissipation is considerably less than max.-signal dissipation.
[4] Values in parentheses are with two-tone test signal.

» 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

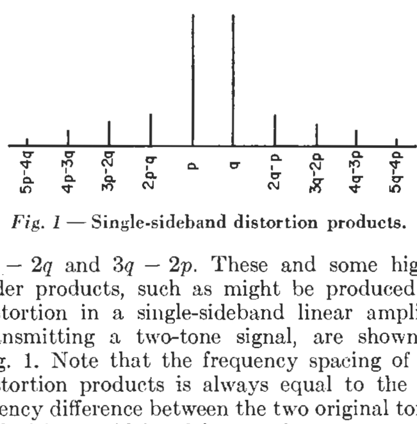


Fig. 1 — Single-sideband distortion products.

$3p - 2q$ 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, 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 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

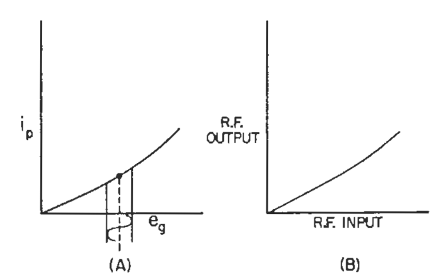


Fig. 2 — Effect of nonlinear plate characteristics.

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 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

From *QST*, November, 1954.

of a series of cascaded amplifiers have similar curvatures, the distortion products generated by each add together in phase.

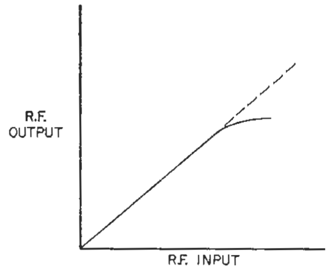


Fig. 3 — **Effect of grid loading on linearity.**

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

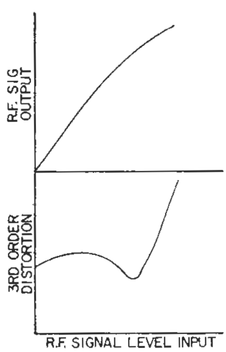


Fig. 4 — **Distortion cancellation.**

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.

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

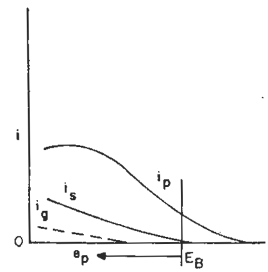


Fig. 6 — **Instantaneous plate current characteristic.**

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!)

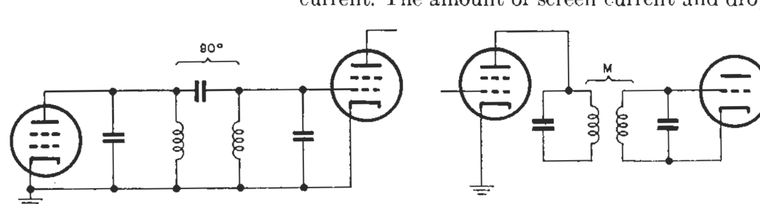
It is apparent that it is best to choose tubes and operating conditions for low grid driving power. Tubes are available that will operate Class AB_1 at power levels up to 500 watts, and their use greatly simplifies the driver and bias regulation requirements.

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 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-

[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.**



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

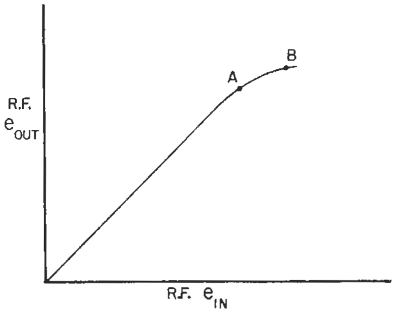


Fig. 7 — Linearity curve of typical tetrode amplifier.

curve of a 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,

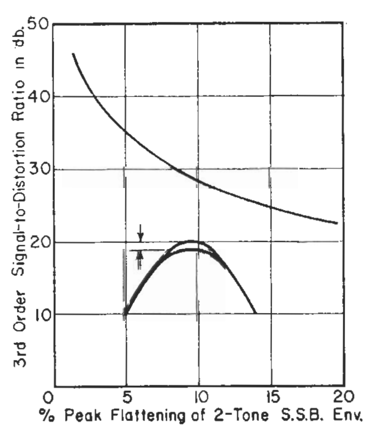


Fig. 8 — **Relationship between third-order distortion and envelope peak flattening.**

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 elements. It can be substantially eliminated by proper adjustment of bias, screen and plate

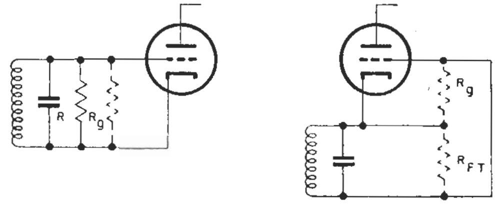


Fig. 9 — **Grid loading. A — Grounded-cathode input circuit; B — Grounded grid.**

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 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.

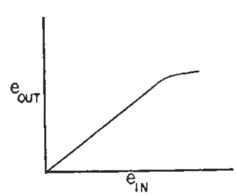


Fig. 10 — **Linearity curve with plate saturation.**

Usually the third-order distortion component is at least 6 db. greater than the fifth- or 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 characteristic of the tube. Fig. 11 shows such a curve. As stated earlier, this type of curve is obtained

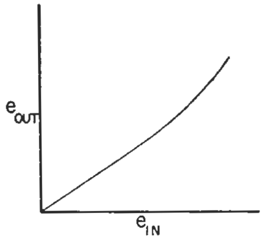


Fig. 11 — Nonlinearity due to exponential plate current characteristic.

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 relatively stronger fifth- and higher-

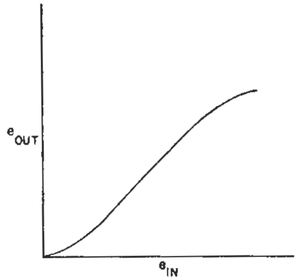


Fig. 12 — Linearity curve with compound curvature.

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.

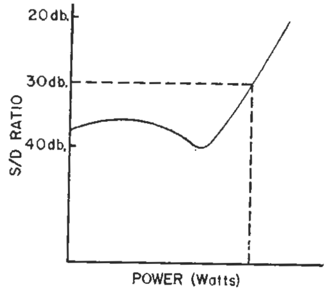


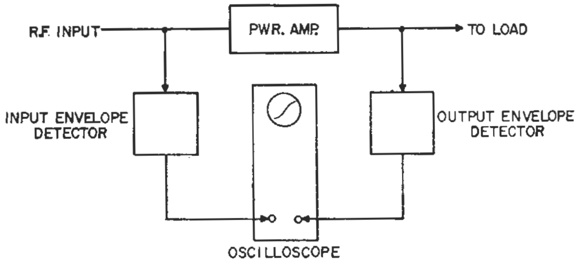
Fig. 13 — Signal-to-distortion ratio *vs.* power output.

envelope detectors with the output of one connected to the horizontal input of an 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 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

Fig. 14 — Block diagram of linearity tracer.



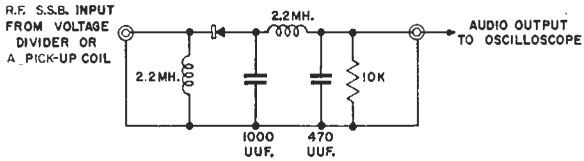


Fig. 15 — Envelope detector.

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.

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.

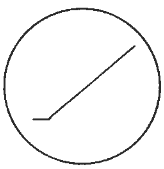


Fig. 16 — Effect of inadequate response of vertical amplifier.

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.

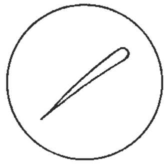


Fig. 17 — Double trace caused by audio phase shift.

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.

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

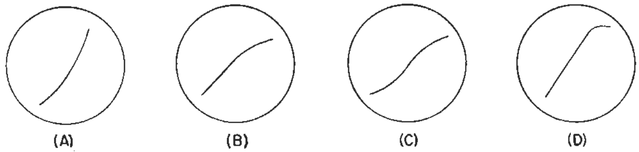


Fig. 18 — Typical linearity traces.

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

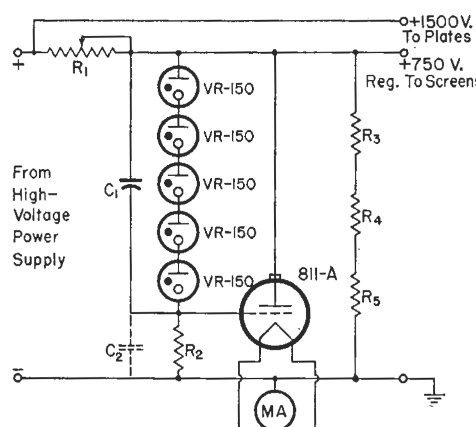
As everyone knows, or soon finds out, tetrode linear amplifiers require "stiff" screen-voltage supplies for lowest distortion. Earl Weaver, W2AZW, uses a pair of 813s in his output amplifier, and devised the circuit shown here to stabilize the screen voltage. It is a shunt-type regulator that derives 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 varies from about 5 to 60 ma., and the shunt regulator holds the screen voltage constant to within 10 or 15 volts.



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

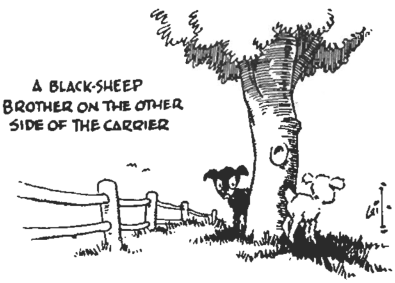
C_1 — 0.01 μf., 2000 volts.
C_2 — 0.01 μf., 400 v., if needed to prevent oscillation.
R_1 — Adjustable wire-wound, resistance and wattage as required.
R_2 — 22,000 ohms, 2 watts.
R_3, R_4, R_5 — 0.1 megohm, 2 watts.
MA — Milliammeter required for original adjustment.

» *The workings of linear amplifiers explained in terms any amateur can understand. Goes into the things that cause nonlinearity, and what to do about them.* Must *reading.*

Why Linears Go Wrong

RICHARD E. LONG, W3ASW

Some single-sideband signals are beautiful to listen to, but some leave much to be desired. Analysis of these latter signals reveals that, while the portion of the signal that carries the intelligence is substantially a good clean s.s.b. affair, it has a "black sheep" brother on the other side of the carrier frequency that does no



earthly good insofar as delivering a signal is concerned, and it only wastes precious frequencies.

The main difficulty seems to be that most fellows will spend a lot of time and money in building a good exciter and then forget all about what follows in the transmitter line-up. Many hams have adopted the attitude of making the old final handle the job. Then when it doesn't deliver the goods the way the book says it should, they start overloading all stages ahead of it until the plate current kicks up to where they think it is really going to town. This is where they usually undo all the work put in on the exciter and, incidentally, where they ruin a good s.s.b. signal.

Since all the aforementioned observations were made through experience on the air it may be well to recount how all this hullabaloo got started. Some of the brethren may see a similarity to their own troubles if they are just getting on the air with a linear amplifier. My exciter was built and put on the air in the days when W2KUJ was the only ham who had enough know-how and equipment to analyze a s.s.b. signal. He operated 20 meters and I operated 75. That made things just dandy so far as signal reports were concerned. If I got a report that did not suit me, I credited it to the fact that the other fellow didn't know what he was talking about. After all, didn't I have a filter that only passed 2.5 kc. of signal? And didn't the final plate current always fall back to nearly zero when I stopped talking? That settled it. No parasitics in the final and a 2.5-kc. filter — it must be the other fellow's receiver, etc., etc. One day I got a letter from a nonham who was doing some listening work for the National Bureau of Standards and some incidental ham monitoring. The gist of the letter was that *he thought he could determine some splatter on either side of the main signal!* Sounds nasty, doesn't it? It stuck in the back of my head, and I asked a fellow ham with a selective receiver to give me a good going-over the next time he heard me on the air and see what he could find. He reported splatter, too, and suggested I try to do something about it. Frankly, I didn't know *what* to do about it!

Along about this time, W2KUJ came up to 75 from 20 and began looking over the existing s.s.b. signals. The reports he passed out were anything but gratifying. Yes, I got poor ones, too. I didn't want to believe him, but then he was the engineer who had developed a system of s.s.b. for GE, and who was I to argue with him? Don wrote me an eleven-page letter describing types of distortion in linear amplifiers, as a follow-up to a discussion we had over the air. I'm saving it for a day when I can understand mathematics. However, he also made some statements and drew some pictures that I could understand, and that was the beginning of my seeing daylight. If I can pass along some of this daylight so that others can understand a few things about linear amplifiers without resorting to slide rules, vector analysis, and other math, maybe they can apply the principles to their equipment, as I did at W3ASW, and some of these "double-sideband" single-sideband signals will be eliminated.

Sources of Distortion

As a starter, let me quote from Don's letter some of the things which clicked with me. Here's the first page:

DISTORTION IN LINEARS
OR
GRAY HAIRS ON S.S.B. OP'S HEADS

In a linear amplifier, departures from a truly linear response fall into two main categories:

1) Amplification which increases with increased signal level (caused in many cases by overbiasing); and

2) Amplification which decreases with increased signal level. This is "limiting action," "peak squashing," or

From "Sugar-Coated Linear-Amplifier Theory," *QST*, October, 1951.

whatever term one wishes to apply to the action. Combinations of these effects can and do exist in such amplifiers.

Nothing tough about that so far. Even I understand it. A good idea of the sound of the first type of distortion could be heard on those old receivers that used a 6C6 Class C stage as a squelch to silence between stations. When the thing was about to open, the speech would be all chopped up as the amplifier swung about the cut-off point with variation in signal level. On Type 2 distortion, I've always called it "saturation," or "flattening off." You have heard it splattering all over the band long before hams began using s.s.b. Getting on with Don's letter, he draws curves and proves, by means of two pages of math, that these two types of distortion do occur. I'll believe him. Here's page 3 of his letter:

It should be pointed out that the transmission of a single pure signal through an amplifier having either of the two types of distortion will cause a series of harmonic signals to be generated. In general, these harmonic signals are not transferred to the radiating antenna (in case the amplifier is operating at radio frequencies) but the harmonic currents must flow in the output circuit. If the amplifier is "broadband," [1] such as an audio amplifier, the output will contain the whole series of harmonics generated, within the limits of the bandpass circuits.

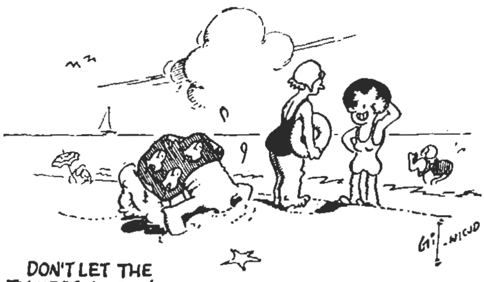
No spurious signals are created by an ideal amplifier, nor are spurious signals produced by a nonlinear amplifier if adequate harmonic attenuation is provided in the output circuit *when, in this last case, only one tone is presented to the input terminals of the amplifier.*

When more than one signal is impressed upon the input terminals of a nonlinear amplifier, spurious signals are generated. Many of these will not appear in the antenna circuit (if harmonic suppression is adequate), but many of them will have frequencies that are essentially in the same band as the desired signals, and therefore will appear in the output along with the desired signals.

Phew! Seems to be getting a little deep, but the main thing to remember seems to be the fact about only *one* tone applied to a distorting amplifier may not show up in the output circuit as a distorted signal in all cases. This leads some fellows astray in using only one tone, or carrier, and a 'scope in adjusting a linear. Notice his statement about "more than one signal" being impressed. That's where a two-tone test comes into the picture for amplifier adjustment, and you will find that proper interpretation of what you see with a two-tone test is an easy and simple way to adjust your amplifier. More on this later.

The next two and one-half pages of the letter contain the most gosh-awful looking mathematics and are the real reason the letter was written because they prove statements Don made to me in the contact and which I found hard to believe. I loaned the papers to several "bushy brains" whom I know and they said it is all true, so I believe it and will just pass along to you the example given and you can take it from me, it does happen "like he sez."

It seems a distorting amplifier can *put signals back on* where you spent so much time and money to *take them off!* They won't be readable things, to be sure, but they are still signals. They take up room in the spectrum, they would be classed as splatter, and they can smear an adjacent channel. I've heard this happen on numerous occasions when the operator was not aware that his signal was distorting. Here we go again with more of the letter. Don't let the figures scare you — they are really simple:



Suppose $F_1 = F_o - 200$ cycles/sec. and
$F_2 = F_o - 100$ cycles/sec.,
where F_o is the carrier frequency of a single-sideband transmitter carrying a two-tone test signal on the lower sideband, using modulation frequencies of 200 and 1000 cycles per second, respectively.

The output frequencies are:

a. $F_1 = F_o - 200$ c.p.s.

b. $F_2 = F_o - 1000$ c.p.s.

c. $2F_1 - F_2 = 2F_o - 400 - F_o + 1000 = F_o$ *plus* 600 c.p.s.

d. $2F_2 - F_1 = 2F_o - 2000 - F_o + 200 = F_o - 1800$ c.p.s.

It can be seen that the signal *c* is on the "high" side of the carrier! Signal *d* is on the low side as are *a* and *b*.

So this is how amplifier distortion produces "hash" on the "other" side of the carrier.

What do all those figures mean? Just that in the output of an amplifier that is distorting due to improper bias, you will find the original two tones applied plus the intermodulation products which are shown as the second harmonic of the first beating with the fundamental of the second and vice versa. Since these are from two tones only, imagine what speech frequencies will do! Don later proves that similar products are generated in the limiting type of distortion and comes up with the following combinations:

$2F_1 - F_2$, $2F_2 - F_1$ (same as Case 1)
$3F_1 - 2F_2$, $3F_2 - 2F_1$ (new one)
$4F_1 - 3F_2$, $4F_2 - 3F_1$ (also new)
etc., etc.

These figures mean that the third harmonic of the first will beat with the second harmonic of the second, the fourth of the second with the third of the first, etc. They indicate the existence of "intermodulation" or "cross modulation" products. If you want further proof that they exist, listen to some of the gang with limiting amplifier systems and by means of a selectable-sideband receiver adapter, check the "unwanted sideband." You will no doubt find a lot of hash. If you must have mathematical proof, send a self-addressed stamped envelope along with your right arm for payment.

How does all this apply to the actual equipment? Let's see how we can put some of these things to work for us.

[1] "Broadband" here refers to the *ratio* of the frequencies to which the output circuit will respond.

Many s.s.b. transmitters consist of the generator itself (filter or phasing type) followed by one or more linear amplifiers. Low power levels (up to a few watts) are most easily handled by receiving tubes run as Class A amplifiers, so if you have followed the figures in published tables for receiving-type tubes in Class A service, you should have no trouble with these low-power stages. Remember though, *no instability caused by regeneration can be tolerated.* This means adequate, or better than adequate shielding, good separation of grid and plate leads, etc., and a fairly good Q in the tuned circuits. Some resistance across a grid or plate circuit may be needed. Here I had to put a 100,000-ohm resistor from a 6SK7 plate to ground to tame a 456-kc. amplifier. This all follows receiver design practices and applies wherever receiver-type tubes are used.

Driver Stages

Whatever type of exciter is used, you eventually come to the first power stage, usually an 807, acting either as a final, or as a driver for the final. It is here where trouble can and usually does occur. Many words have been written on the troubles encountered with 807 stages and Class B driver stages and, in all probability, no two cases will ever respond to the same treatment. I'll tell you what I had to do here, and you might get some ideas which may help in your case. I used an 807 driving a pair of 811s. While the voltages available would not make even an old-type 807 blush, I couldn't find the proper ratings for an 807 in Class A service. Since the 807 is just a 6L6 with a top cap, inspection of the ratings of a 6L6 gave a set of voltages that are doing very nicely here with my old 807. Values of 350 plate volts, 250 *regulated* screen volts, and a 250-ohm cathode resistor will set the tube up in Class A operation with a load resistance of around 5000 ohms and an output of from 6 to 10 watts.

Now, with all these voltages applied, my 807 promptly took off on its own, and I didn't get it to calm down until I applied the v.h.f. chokes to cure the parasitics. Then, with neutralization, I began to get a "cold" 807 but with not quite enough drive for the 811s. Somebody said proper impedance matching is hard to obtain if you don't have enough C in the plate tank. Checking the *Handbook* showed something in the vicinity of 300 $\mu\mu$f. for a single-ended amplifier at 4 Mc. with my voltages and currents. I had been using a 100-$\mu\mu$f. condenser here, and substitution of a 365-$\mu\mu$f. broadcast type for it and pruning the coil to make use of about 350 $\mu\mu$f. made a world of difference. I had plenty of drive for the final now, but since the 807 stage was still skittish at times, I suspected that some regeneration was left in it. Looking again at the load-resistance figure, and trying to remember what a tuned circuit looked like at resonance, brought out the idea that maybe a 5000-ohm resistor across the tank circuit would give a better match to the tube than a tank circuit alone. While all these are probably not the right answers to taming 807 drivers, the final result is that my 807 with the voltage specified, a high-C tank circuit loaded with 5000 ohms and *no neutralization* delivers enough output to drive the 811s to RCA's specification of 38 grid milliamperes in Class B service. As for that high-C tank, it improved the matching of impedance between driver and final, and it meets Norgaard's statement about adequate harmonic suppression in the output circuit. The 5000-ohm resistor helps a lot with another of the bugaboos: regulation of the driver-stage voltage.

The Final Amplifier

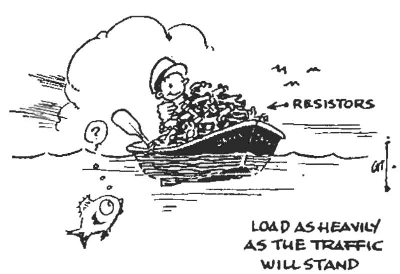
This brings us up to the final amplifier, which also is capable of either making or breaking a good s.s.b. signal. Once more, while what I did here may not be the criterion of treatment for linear amplifiers, it may give you some ideas along the right path to the best operation of yours. Although triodes are used here, and they do present a varying load when they draw grid current, many of the ideas applied will hold for a tetrode or pentode final that operates in the region of no grid current.

Let's look at the grid tank, which is the first item encountered. The 811s will draw grid current, and that means a variable load on the driver. Don said that it is important to keep harmonic content of all circuits down by the use of adequate Q. Furthermore, the experience of matching the final to the driver plate indicated that high Q or high C was a good idea. Consulting the chart for tank circuits in the *Handbook* shows that for Q's of 12 at 4 Mc., effective capacities may range from 50 to 1000 $\mu\mu$f., depending upon the ratio of plate voltage to plate current. Despite the fact that these actual figures won't apply for a grid tank, they will convey the general idea of what may be encountered with various ratios, or with varying loads, which is what we have in a Class B grid circuit. Under the heaviest loading, or the least equivalent resistance, you will need the most capacity to maintain good Q and, although I can't tell you exactly what your load limits will be, I can tell you to use the most capacity available in order to stay on the safe side. When in doubt, always use more capacity than is needed.

I took an average from the chart, and wound up with a 365-$\mu\mu$f.-per-section dual broadcast-type condenser. The coil was pruned in order to use about 340 $\mu\mu$f. of each section. This gave an

effective capacity of 170 $\mu\mu$f. in the circuit. Compare this value with the usual grid tank circuit and you will get the idea I am trying to get across. The usual dual 75- or 100-$\mu\mu$f. grid tuning condenser just won't do for a 4-Mc. triode amplifier in Class B where the tubes draw grid current.

What about that varying load? Let's try swamping to steady it a bit. What do I mean by swamping? Just this — connect a *noninductive* resistance across the tank circuit, to dissipate some of the excitation and offer a more nearly-constant steady load. Then, when the tubes draw more or less grid current, the over-all load on the driver will not vary as greatly as without it. How much swamping? That question has always started a lot of arguments on the air, and each fellow has his own idea about what is correct. In my case, I started with a rather high value of resistance, to keep the peaks under control, and found that I could reduce it until I started losing excitation to the final. This value proved to be somewhat below that calculated from the formulae given by Reque and was near to that used to load the exciter plate tank. The natural thing to try was to make them both the same. You're right — I now have a 5000-ohm swamping resistor across each tank. This cut-and-try may not be the elite method of obtaining optimum swamping but by using it you will arrive at a good value, and it will work right



along with the available excitation. Load as heavily as the "traffic will stand." I can see eyebrows raise and arguments start on that one, but I'll stick by my guns.

Grid bias? That's easy; follow the specifications for the tube. They will be OK for a starter and may not need changing in the last adjustment, but more on that later.

In figuring the final plate tank, stay with the *Handbook.* Use the expected or wanted value of plate current at full signal with the plate voltage available and try then to go a bit more on the high-capacity side. Uusually your tuning condenser won't be anywhere near large enough in capacity. Mine wasn't, and I had to make up the difference with padders. Those 50-$\mu\mu$f. vacuum units from the Command set antenna relay boxes will do nicely, but *don't* put a single unit across the tank from plate to plate. Use two or four in a "split stator" arrangement, because unwanted harmonic currents will find a better path to ground with this arrangement. The net result here was a dual 180-$\mu\mu$f.-per-section variable with each section padded with a 50-$\mu\mu$f. vacuum padder. I pruned the coil to make use of the maximum capacity available.

One more item before we try to adjust the amplifier. Let's look at the plate supplies. Since the exciter uses all receiving-type tubes except the 807 driver, the best arrangement found was one good husky low-voltage supply with the 807 plate taken off ahead of the regulator. The 807 screen and all other plates are regulated with a VR-105 and VR-150 in series, and the dropping resistor adjusted for an average current through them of 25 ma. The plate supply for the final should be as "stiff" as you can make it. The line regulation here at W3ASW is very poor; turning on a 100-watt load will make quite a flicker in the lights. This had me worried and I knew I would do a lot of head scratching to get around it. Line regulating transformers of the size needed are quite expensive, and electronic regulation would probably not compensate for the poor line voltage supplies. Well, here again Don Norgaard came to the rescue, with the suggestion that I use as large an output capacity on my final filter as it was possible to use. The plate voltage here averages 1400 volts and the final idles at 30 ma., so I made use of two 10-μf. 1500-volt units along with a 10-henry 500-ma. choke (nonswinging, by the way) as a filter for the final. Use the choke-input system with all the capacity on the output side. I manage to stay fairly linear under voice operation, but a steady tone will pull the output voltage way down. Incidentally, 60 to 80 μf. on the exciter plate supply output won't do a bit of harm.

Before getting into the loading, it is assumed that you have good neutralization and parasitic suppression in your amplifier. Here again, *instability can not be tolerated!* With plate voltage applied, the bias should be adjusted to allow the tubes to draw their maximum rated plate dissipation; then rotate both tuning condensers without the antenna or exciter coupled and see if the final spills over at any point. If it does, you have more work to do. This may be a tough baby to

shave, but it is a *must.* When you are sure the final behaves, return the bias to normal. My 811s required 25 ohms with 7 turns of hook-up wire wound around them in each grid lead to tame them.

Testing with a 'Scope

If you don't have a 'scope, borrow one and make notes of optimum conditions on your final coupling and final plate current, and then try to maintain these.

Figs. 1, 2 and 3 represent what you should see on a 'scope when a two-tone test is applied to your rig. Fig. 1 shows an ideal picture with an improper bias distortion curve plotted on it. Notice the nice clean "X" crossover on the ideal, and then look at the curves or loops on the distorted curve. Fig. 2 shows another ideal picture, with the distortion curve resulting from saturation in the plate circuit plotted on it. Here the crossovers are clean but notice the rounding and flattening on the peaks. Fig. 3 shows an ideal curve compared with one having both types of distortion. Quoting Don, "Have you ever seen Figs. 1, 2 or 3 on your 2-tone test? If you can see it, it's pretty bad — believe me!"

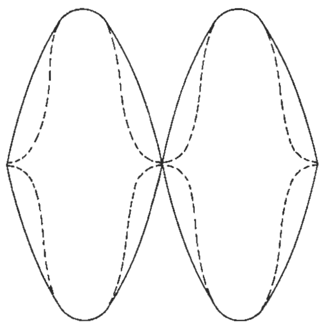


Fig. 1 — **Ideal two-tone pattern (solid line) and the result of too much bias (dotted line).**

What is a two-tone test? Just two a.f. signals separated by about 1000 cycles applied to the amplifier under test. How do you get it? On a filter rig with a "carrier insertion" or an "unbalance" control, open up a bit of carrier and at the same time apply an audio signal of approximately 1000 cycles into the front end. Vary the levels, while watching the 'scope screen, until the picture looks like or nearly like the ideal curves shown. With the phasing rigs, the introduction of a tone of about 1000 cycles into the front end of the set and the disabling of *one* balanced modulator will give a similar picture.

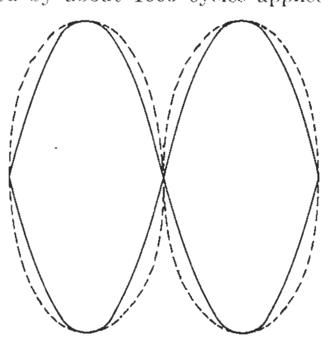


Fig. 2 — **With too much drive or incorrect loading, the two-tone pattern turns into the fat, squashed signal shown by the dotted line. Compare this with the ideal (solid line).**

Now carefully advance the amount of both signals applied until you either reach the limit of the power-handling capability of your final, or until the picture shows limiting distortion.

If, by adjusting coupling and drive, you can eliminate the limiting distortion, fine business. If the limiting appears at a point beyond where you want to operate your final, forget it. But then *don't overdrive that final.* This goes for all stages, by the way. A good way to stay within these limits, if you own a 'scope, is to monitor continuously with it. Make some kind of calibration mark on the screen where the flattening begins to show, and then stay below that point in talking. If you have borrowed the 'scope, make note of what the final plate current is when this flattening occurs and stay below that point. A still better system is to employ a limiter or clipper in the speech circuits so that you don't drive out of linearity, no matter how loud you shout. This latter system is employed here and has proven very successful. With a phasing rig, the use of a limiter or clipper means that it *must* be followed by a low-pass filter, to keep the harmonics generated in the clipper from being radiated. The adjustment of the limiter is simple: make note of the point of distortion on either the 'scope or the plate meter, and then set the output control of the limiter so as not to exceed this point.

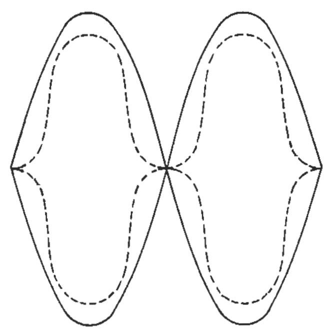


Fig. 3 — **You are really in the soup with too much bias *and* too much drive or incorrect loading, because the two-tone pattern will look like the dotted line shown here.**

As for the biasing type of distortion (Fig. 1), I never have seen it here with zero-bias triodes. If it should occur, check your bias supply carefully and make sure that no grid-leak action is taking place, for this would distort severely. This goes for all types of amplifiers in linear service. For my 811s, I use three flashlight cells in series, which is about right for my plate voltage.

Before I get along too far, let me say that the above-mentioned two-tone tests should be applied to your driver stage as well as the final. I find here that I limit in the plate circuit of the driver at a point that gives me about 300 watts input to the 811s. That is where I operate my rig and keep it below that point with the limiter. Three hundred watts of s.s.b. is no mean signal and, in order to get a higher input without distortion, I would have to rebuild the entire plate supply for the exciter, to raise the 807 plate voltage. It isn't worth it, according to my way of looking at it.

That about winds up the story for using zero-bias triodes in the final. A few words were promised about multielement tubes and here they are: Practically all of these should be operated as Class A, Class AB_1, Class AB_2 or as "Modulator Service." Tubes in Class AB_1 or less should not draw grid current at any time, therefore the loading of the grid circuit should not be necessary. The loading should be applied to the output of the driver in all cases, however. With tetrodes or pentodes, where the screen voltage is the big controlling factor, the screen supply should be regulated. In such types as 807s, 829s and 4-125s, the ordinary VR tubes should suffice. "Bleed" the VRs up to nearly their maximum current so they won't go out when signal is applied. Regulation of 813 screens is something different. They draw too much screen current when going from no signal to full signal, and the VR tubes just won't handle the job. The ideal answer is electronic regulation with tubes such as 2A3s, or triode-connected 6L6s, handling the load. This same care is needed in the control-grid bias. It should be very stiff. All other items, such as plate tank, loading to antenna, etc., are the same. Don't let the tetrodes or pentodes scare you — good signals can be gotten from them.

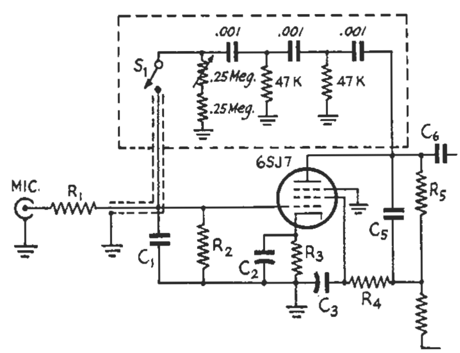
Some swamping may be helpful on the grid tank to control transients, and it will be best determined by cut-and-try. Start high and work downward. It won't do any harm.

A final word about antenna loading. Try to use some kind of indicator in the feed line or in the output circuit somewhere. Many fellows have gone overboard in coupling the antenna to the final. Their plate current will show that they are driving way up near the limit the law allows, with the exciter loafing along. Inserting a radio-frequency ammeter in the feed line (a 52-ohm coax line, in the case I'm referring to) showed that the actual output doubled when the coupling was reduced to where the final drew considerably less input.

If your work's slipshod, you just don't get the right results with single sideband.

A SIMPLE AUDIO OSCILLATOR FOR TUNE-UP

The trend certainly seems to be to build an audio oscillator into your speech amplifier, for quick testing and tune-up of the s.s.b. rig. Curt Smith, W6VCM, sends along the dope on the audio-oscillator circuit he built into the speech amplifier and voice-control circuit of his s.s.b. exciter. As shown here, the only components needed to make this addition are a few resistors and condensers. The switch S_1 is



This audio test oscillator can be added to the speech amplifier and voice-control circuit of the s.s.b. exciter described on pages 96 through 101. It is turned on by S_1, mounted on the 0.25-megohm variable (pitch) control. Values not given are the same as in the original circuit, shown in Fig. 3, page 101.

mounted on the 0.25-megohm variable pitch control. With these constants, the frequency range is approximately 450 to 600 cycles, and different frequencies can be obtained with a different set of constants. The output is not constant over the range of adjustment of the 0.25-megohm variable, dropping off as the resistance is made lower.

» To make your linear amplifier put out a signal that is above criticism you need three things: this article, the preceding one by W3ASW, and an oscilloscope. Wait a minute — we almost forgot the fourth ingredient: You! Unless you really *put into practice what the other three tell you, you'll be right in the ranks of the splatterers.*

How To Test and Align a Linear Amplifier

ROBERT W. EHRLICH, W4CUU, EX-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 much less than one thousandth of the total power (30–40 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 a poor indicator 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.

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.

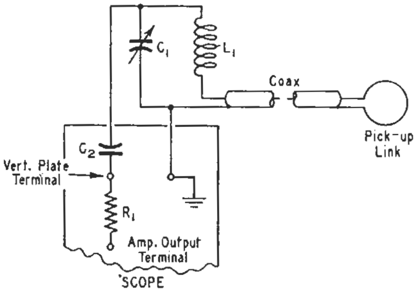


Fig. 1 — The recommended method for sampling r.f. and applying it to the vertical plates of a 'scope. The pattern height can be varied by changing the location of the pick-up loop or by varying C_1.

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.

Normally, the pick-up loop 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 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.

From *QST*, May, 1952.

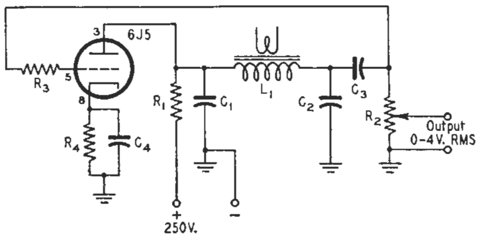


Fig. 2 — Fixed-frequency audio oscillator having good output waveform. The frequency can be varied by changing the values of C_1 and C_2.

C_1, C_2 — 0.02 μf., 600 volts.
C_3 — 0.01 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.

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 modul tion; 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. In the W2UNJ exciter, for example, pull out one 6K6GT; or in the SSB Jr., place a short from plate to B+ on one section of the 12AT7 audio tube.

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.

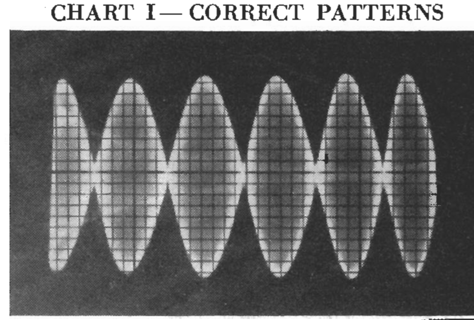
Double-Trapezoid Test

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. transmitters will recognize this set-up as being the same, except that instead of one trapezoid, this test produces two triangles pointing toward each other.

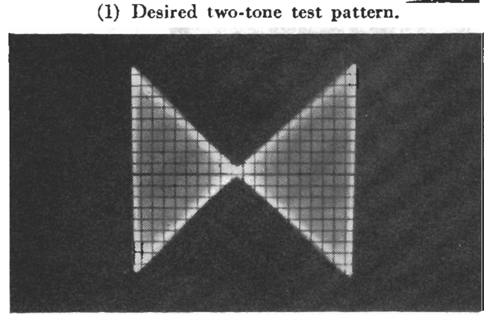
Each individual triangle is subject to the same analysis as the regular trapezoid pattern; i.e., the sloping sides of the pattern should be straight lines for proper operation. Since it is much easier to tell whether a line is straight or not than to judge the correctness of a sine curve, the double trapezoid has the advantage of being somewhat more positive and sensitive to slight 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 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 set-up, 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 fre-

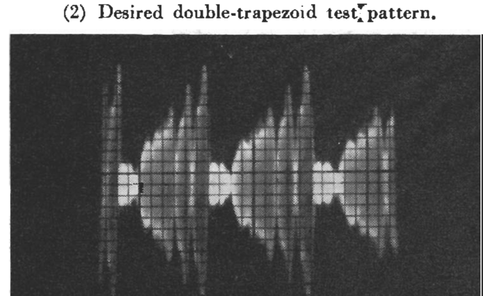
CHART I — CORRECT PATTERNS



(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.

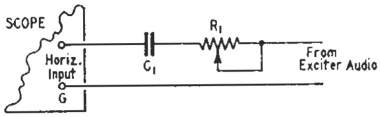


Fig. 3 — "Phaser" circuit for the oscilloscope.
C_1 — 200 $\mu\mu$f. or as required.
R_1 — 0.5-megohm potentiometer.

quency or by putting a phaser in series with the horizontal input to the 'scope, as shown in Fig. 3.

Ratings

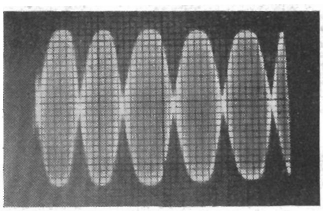
Before proceeding with linearity tests, it is well to have in mind the current and power levels to expect. A suppressed-carrier signal is exactly like an audio signal, except for its frequency, so the audio ratings for any tube are perfectly applicable for linear r.f. service where no carrier is involved. 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.

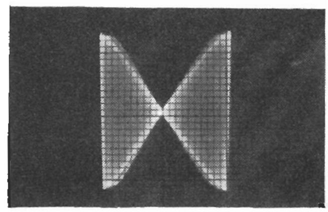
The maximum-signal conditions determined from tube data correspond in s.s.b. work to the very peak of the r.f. envelope. In a correctly-

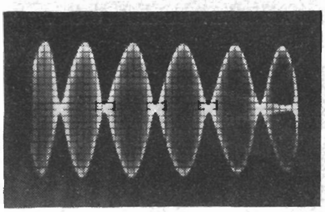
CHART II — IMPROPER AMPLIFIER OPERATION



« (4) Overdrive, indicated by flattening of peaks.

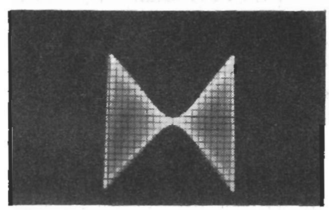
» (5) Same as (4), double-trapezoid test.

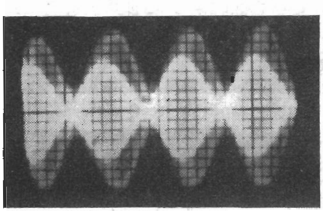




« (6) Too much bias, causing crossover to become pinched together rather than cutting straight across center line.

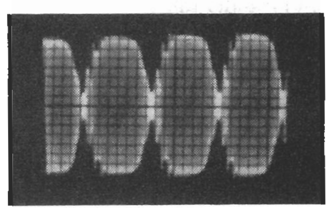
» (7) Same as (6), double-trapezoid test.

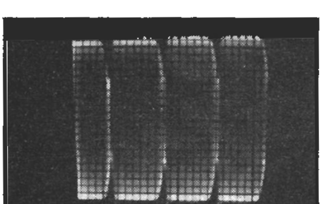




« (8) Two-tone test with v.h.f. parasitics. Note fuzzy halo or fringe. In milder cases the fuzziness will appear just at the peaks.

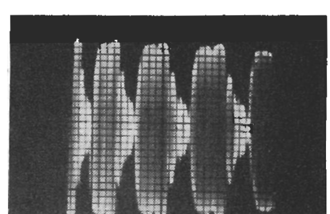
» (9) Two-tone test with fundamental frequency parasitics, accompanied by overdrive.





« (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.



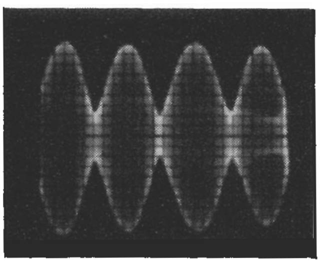
adjusted amplifier, the rated peak input would register on the meters only if one were to whistle into the microphone, otherwise the meters will always read less. In particular, the average input under two-tone linearity-test conditions is close to 65 per cent of the actual peak input for a Class B amplifier, 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

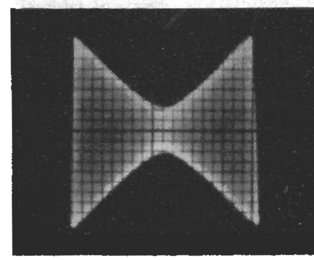
The photos accompanying this article have been taken to show many of the typical patterns that may be encountered with either of the test

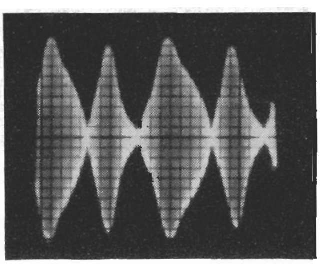
CHART III — IMPROPER TEST SET-UP



« (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.

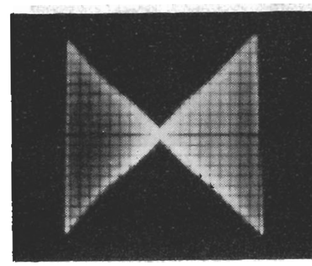
» (13) Same as (12), double-trapezoid test (Method B).

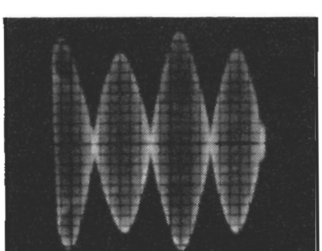




« (14) Distorted audio. A clue to this defect is that successive waves are not identical.

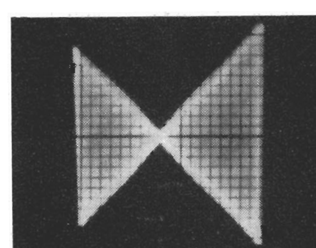
» (15) Same distortion as (14), but switched to double-trapezoid test pattern. Note that correct pattern prevails regardless of poor audio signal.

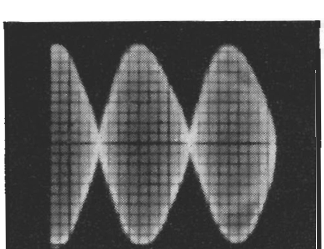




« (16) Carrier leakage through working modulator (Method B only).

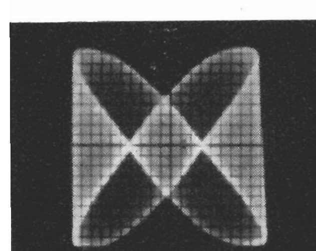
» (17) Same as (16), double-trapezoid.





« (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 set-up.



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 set-up 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 can then be 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.

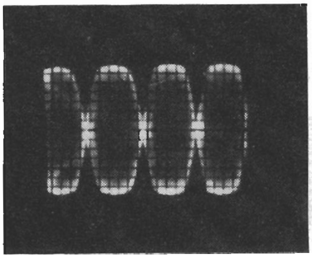
2) If the final is not being driven to its limit, it should be loaded less heavily until such is the case.

3) If the power level obtained above is less than should be expected, use more driving power.

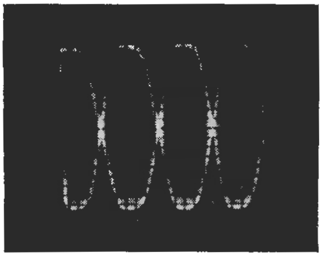
There are several ways to tell whether or not the final is being driven to its limit. One way is

CHART IV — AMPLIFIER LOADING CHARACTERISTICS

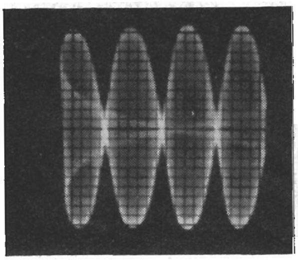
Two-tone patterns taken at the output of a Class B linear amplifier with constant drive and successively heavier loading. Measured input power is indicated.



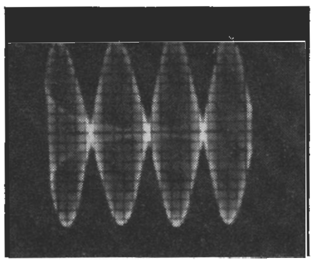
(20) 90 watts.



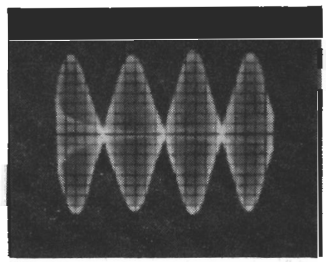
(21) 135 watts.



(22) 250 watts.



(23) 330 watts.



(24) 400 watts.

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.

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 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 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 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-condenser 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.

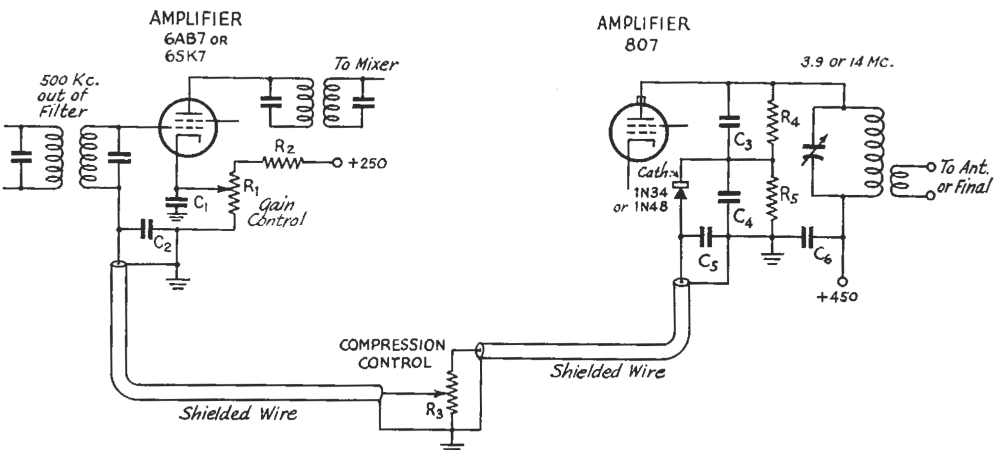
Conclusion

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.

PEAK-LEVEL CONTROL

Dave Mann, W6HLY, has a worth-while method for insuring that his s.s.b. rig isn't hit hard enough to be driven beyond the linearity limits. He finds this very useful with visitors, where the level may vary considerably, depending upon the visitor. It is no more than an output (or high-level) rectifier that is used to control a low-level r.f. amplifier. Some of the r.f. developed at the 807 is rectified and fed back (through level-setter R_3) to the grid of a variable-μ r.f. amplifier following his sideband filter. The gain of the amplifier is set by R_1, but the fed-back control voltage is set by R_3.



This peak-level control circuit will readily handle variations in audio input over a 20-db. range.

C_1 — 0.01 μf.
C_2, C_6 — 0.002 μf.
C_3 — 15-$\mu\mu$f. mica.
C_4 — 100-$\mu\mu$f. mica.
C_5 — 0.1 μf.
R_1 — 10,000-ohm potentiometer.
R_2 — 0.22 megohm, 1 watt.
R_3 — 1-megohm volume control.
R_4 — 1 megohm.
R_5 — 12,000 ohms.

The crystal rectifier is biased (through divider R_4R_5), so it doesn't rectify immediately but only when this bias is exceeded. Thus the effect is more like limiter action than straight compression, but without introducing appreciable distortion. The capacity divider, C_3C_4, makes the r.f. voltage applied to the rectifier substantially independent of frequency.

The control circuit cannot be installed in every existing exciter without modification, because it requires that the existing exciter have 20 or 30 db. of gain *to spare*. However, installing the control tube in an existing design will insure that this requirement is met. In the interests of good linearity, the controlled tube should be used in the circuit at a point where the signal level is low (less than 1 volt). To avoid feedback troubles, the controlled stage should be on a frequency different than that from which the control voltage is derived.

» The importance of proper loading for optimum operation of a linear amplifier has been stressed in the preceding four articles. The bewilderment with which many amateurs contemplate an amplifier that "won't load" is entirely unnecessary, provided the simple design procedures outlined below are followed.

Coupling to Matched Coaxial Lines

GEORGE GRAMMER, W1DF

THE problem of designing a tank circuit to load an amplifier properly is solved quite simply if a sensible attitude is adopted toward the question of coupling. Many amateurs look for coupling systems that will, through some black magic, work into any antenna-feeder system that happens to strike their fancy. This is about as reasonable as going into a radio store and asking the salesman for "a transformer," without specifying voltage, power, or type of service, and then demanding that he produce one that will do any job you may happen to have in mind.

Thanks to the necessities of TVI prevention, transmitters now are almost universally designed to work into a coaxial line. This is good, because if the line is properly matched at its output end, a definite value of resistive impedance is established as the load for the transmitter. In round figures, this load will be either 50 or 75 ohms, depending on which type of line is chosen (either kind is equally good). The only other thing that needs to be known is the load resistance that the final amplifier tubes must "see." With these two resistances known, the design of a coupling circuit requires only reading a few values from a graph of substitution in some simple formulas.

Reactance

In coupling-circuit calculations it is often more convenient to work with inductive and capacitive reactance than to attempt to use inductance and capacitance directly. Reactance values apply at any frequency, and once they are found it is easy to convert them to L and C for a particular frequency by the following formulas:

$$L_{\mu h.} = \frac{0.159\, X_L}{f_{Mc.}}$$

$$C_{\mu\mu f.} = \frac{159{,}000}{X_C f_{Mc.}}$$

where $L_{\mu h.}$ = Inductance in microhenrys
$C_{\mu\mu f.}$ = Capacitance in $\mu\mu$f.
X_C = Capacitive reactance in ohms
X_L = Inductive reactance in ohms
$f_{Mc.}$ = Frequency in megacycles.

Load Resistance

In using the audio ratings of tubes as the operating conditions of a linear amplifier the tube load resistance, R_L, is usually specified in the tube data. Generally, it will be a "plate-to-plate" value for a push-pull amplifier. If only one tube is used, the proper load resistance is one-half the plate-to-plate figure. When tubes are used in parallel, divide the proper load resistance for one tube by the number of tubes in parallel. In case other than the published operating conditions are to be used, an approximate formula for the optimum tube load resistance is:

$$R_L = 500\, \frac{E_B}{I_B}$$

where E_B = D.c. plate voltage
I_B = Peak plate current in milliamperes.

Plate Tank Q

In a parallel-tuned tank circuit (the conventional type) the operating Q of the tank is equal to the load resistance divided by the tank reactance in use. The inductive and capacitive reactances are equal at the resonant frequency, so either may be used in this relationship:

$$Q = \frac{X}{R_L}$$

In a push-pull amplifier, where R_L is the plate-to-plate load resistance and the tank condenser is of the balanced or split-stator type, the capacitive reactance is the *total* reactance of the tank condenser. Each section of the condenser contributes half of the total reactance, so each section must have twice the capacitance required for the whole circuit. To take a specific case, suppose that the plate-to-plate load resistance is specified to be 8000 ohms. Then if $Q = 10$, the required X_C is 8000/10 or 800 ohms. If the frequency is 4 Mc., the formula given earlier shows that this X_C is equivalent to a capacitance of 50 $\mu\mu$f. Each section of the split-stator condenser therefore should be 100 $\mu\mu$f. Note that if only one tube is used R_L would be 8000/2 or 4000 ohms, so the required X_C for the same Q would be 4000/10 or 400 ohms. Thus the capacitance required for the single-tube single-ended circuit would be 100 $\mu\mu$f., and 200 $\mu\mu$f. if two tubes were used in parallel instead of push-pull.

Matching the Line

The design methods outlined below are based on a matched line — one whose input impedance is equal to its rated characteristic impedance. They will accommodate a moderate amount of mismatch, but may not work if the standing-wave ratio is over 1.5 to 1 or so. To make the s.w.r. less than 1.5 to 1 it may be necessary to use a matching circuit between the output of the transmitter and the input terminals of the transmission line that goes to the antenna. An s.w.r. bridge is a practical necessity for determining whether such a matching circuit is needed; and if it is, for adjusting the matching circuit for minimum s.w.r. on the coax line. Such a bridge can be made at such low cost — a dollar or two — that no amateur can afford to be without one. The construction of s.w.r. bridges and matching circuits is beyond the scope of this book, but full details can be found in the ARRL *Handbook*.

Inductive Coupling

Fig. 1 is a circuit often used for coupling the output of an amplifier to a coaxial line, C_1L_1 being the usual plate tank circuit and C_2L_2 a series-resonant circuit having R, the characteristic resistance of the line, as its load.

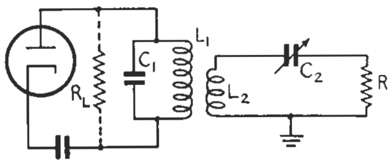


Fig. 1 — Series-tuned coupling circuit for coaxial lines operating at a low standing-wave ratio.

The coupling reflects an equivalent resistance, R_L, across the plate tank circuit; thus R_L is the load "seen" by the tube.

The Q of the tank or primary circuit is

$$Q_1 = \frac{R_L}{X_L} = \frac{R_L}{2\pi f L_1}$$

when C_1L_1 is resonant at the operating frequency. If C_2L_2 also is tuned to resonance, its Q is

$$Q_2 = \frac{2\pi f L_2}{R}.$$

With these conditions, the coefficient of coupling that will just cause the proper value of R_L to be reflected across the primary is

$$k = \frac{1}{\sqrt{Q_1 Q_2}}$$

With any smaller value of k the reflected resistance will be too high; that is, the amplifier cannot be loaded heavily enough to obtain the desired power input.

For reasons other than coupling, it is desirable that the primary should have a Q of the order of 10 or more, 10 being a value that it is usually convenient to obtain with available coils and condensers. Substituting 10 for Q_1 and rearranging gives

$$Q_2 = \frac{1}{10k^2}$$

as the minimum value of Q_2 that will give sufficient coupling for a given coefficient of coupling between the two coils.

Coefficient of Coupling

The coefficient of coupling between two coils is principally a function of their relative spacing and dimensions and not particularly of the number of turns. Fig. 2 shows the shapes of several typical cylindrical coil combinations of the type used for transmitting circuits, together with the measured coefficient of coupling in each case. The coefficient is least when a small coil, such as a link of a few turns, is coupled at one end of a large coil, and increases if the same small coil is moved to the center of the large coil. The largest values of coefficient will be obtained when the smaller of the two coils is fairly sizable in comparison with the larger, and when it is placed on the outside of the larger coil at its middle.

When the series-tuned method of coupling is used for coaxial lines the coupling coil usually is larger than the conventional "link," so the coupling coefficient can be expected to run between 0.5 and 0.6, depending on whether the coupling coil is at the end or center of the tank coil. Using these values in the formula above (for a tank Q of 10) shows that Q_2 should lie between about 0.4 and 0.28, the larger value

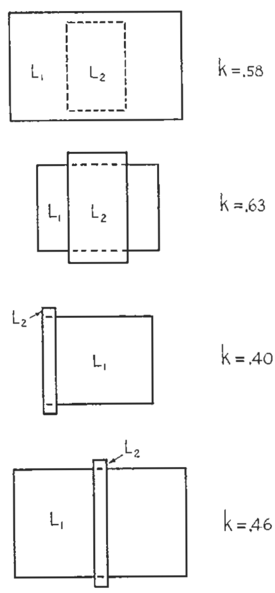


Fig. 2 — Measured coupling coefficients for typical transmitting-coil configurations (side views of cylindrical coils). So long as the same *relative* dimensions are maintained for a given pair of coils the actual physical dimensions do not affect the coupling coefficient. This also holds approximately for varying numbers of turns.

Based on "Coupling to Coaxial Lines," *QST*, May, 1954.

being required for the smaller coefficient of coupling.

Applying the figures to a practical case, suppose that the frequency is 3.9 Mc., that R is 75 ohms — that is, a 75-ohm line properly terminated so that the s.w.r. is 1 to 1 — and that the coupling coefficient is estimated to be 0.5. Then the required Q_2 is 0.4 and the value of inductance needed at L_2 is

$$L_2 = \frac{Q_2 R}{2\pi f_{MC}} = \frac{0.4 \times 75}{2 \times 3.14 \times 3.9} = 1.23 \ \mu h.$$

The capacitance required at C_2 for tuning the secondary circuit to resonance is

$$C_2 = \frac{10^6}{4\pi^2 f^2 L_2} = \frac{10^6}{4 \times 9.9 \times 15.2 \times 1.23} = 1350 \ \mu\mu f.$$

Although this is a rather large value for C_2, it can be obtained by paralleling ganged broadcast-receiver type condensers since the voltage across the condenser will be relatively low even with fairly high power. Alternatively, a 250- to 500-$\mu\mu$f. variable can be paralleled with fixed capacitors to make up the required total. Still another method is to use some available value of variable — such as 250 $\mu\mu$f. — and increase the inductance accordingly to maintain resonance. The ratio of inductance with 250 $\mu\mu$f. in the example above would be 1350/250 = 5.4, so the required inductance would be 6.7 μh. This would increase Q_2 by the same factor.

The principal disadvantage of using a higher Q_2 than the minimum required is that the secondary cannot be operated at exact resonance, which would cause R_L to be lower than the desired value and thus overload the amplifier, but must be detuned somewhat to cause the proper value of resistance to be reflected. Also, the tuning of the secondary circuit becomes more critical. Since the circuits are overcoupled when Q_2 is larger than necessary, the reactance reflected into the tank circuit is fairly large when C_2L_2 is detuned. There is thus some effect on the setting of the plate tank condenser, C_1, for resonance. Although undesirable from a tuning standpoint, a moderate amount of such detuning is unimportant so far as performance of the amplifier is concerned. C_2 makes a smooth and convenient coupling control.

External Loading

The reaction on tank tuning just mentioned can be minimized by using the arrangement shown in Fig. 3. Here the secondary circuit is split into two parts, L_2 and L_3, it being assumed that there is negligible coupling from L_3 to either L_1 or L_2.

There is a basic difference between Fig. 1 and Fig. 3 that has an important bearing on the practical application of these circuits. In Fig. 1, increasing the inductance of L_2 also increases the voltage induced in the secondary (even if the coefficient of coupling does not change) because as L_2 is made larger the mutual inductance between L_1 and L_2 also increases. In Fig. 3, adding L_3 has no such effect, since the mutual inductance between L_1 and L_2 remains the same.

For this reason the inductance of L_3 does not enter into the value of Q_2 when the formula given earlier relating k, Q_1 and Q_2 is used. For purposes of determining coupling, Q_2 must always be based on the value of inductance in L_2 alone. Hence L_2 may have the proper value for sufficient coupling and L_3 used simply to supply enough additional inductance to permit resonating the secondary circuit with a condenser of reasonable size. Continuing the example above, if C_2 is 200 $\mu\mu$f. (to keep well inside the range of a 250-$\mu\mu$f. condenser), then the total secondary inductance required for resonance at 3.9 Mc. is

$$L_{total} = \frac{10^6}{4\pi^2 f^2 C_2} = \frac{10^6}{4 \times 9.9 \times 15.2 \times 200}$$
$$= 8.32 \ \mu h.$$

Subtracting the inductance of L_2 from this gives 7.09 μh. as the inductance required at L_3.

In the circuit of Fig. 3, then, L_3 and C_2 simply provide a variable reactance by means of which the secondary circuit can be tuned to resonance, and L_3 may be looked upon as being merely a convenient way of extending the effective range of C_2. (For example, if the reactance of L_3 is just a bit smaller than that of C_2, the resultant of the two will be a very low value of capacitive reactance; i.e., the combination is equivalent to a condenser of very large capacitance.)

The other side of this picture is, of course, that if L_2 by itself is not large enough to load the amplifier properly when the secondary circuit is tuned to resonance, the only means available for increasing the coupling is to increase Q_1. This can be done by making the primary circuit higher-C. If the C/L ratio is increased by removing turns from L_1, the turns should be taken off the end or ends of the coil farthest from L_2 — probably the most convenient way in nearly all cases — since this will tend to increase k and give a further improvement in coupling.

The links provided on manufactured coils are almost invariably too small, on the lower frequency bands, to provide sufficient coupling to anything but very small load resistances. Measurement shows that the link inductance on some of the medium-power coils for the 3.5-Mc. band is not even as much as 1 μh. As the calculation above shows, this is not large enough for coupling into a 75-ohm line operating at a reasonably low s.w.r., unless a plate tank Q considerably higher than the usual 10 or 12 is used. If L_2 is 0.75 μh., Q_2 becomes

$$Q_2 = \frac{2\pi f L_2}{R} = \frac{2 \times 3.14 \times 3.9 \times 0.75}{75} = 0.245.$$

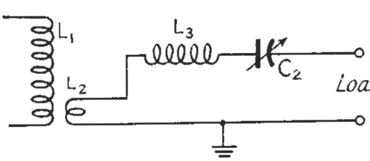


Fig. 3 — **External loading coil, for cases where the inductance of L_2 cannot be altered.**

SINGLE SIDEBAND FOR

Assuming the same coefficient of coupling, then the required Q_1 is

$$Q_1 = \frac{1}{k^2 Q_2} = \frac{1}{(0.5)^2 \times 0.245} = 16.3.$$

This assumes, of course, that L_3 and C_2 are used to resonate the secondary circuit. Without such tuning a still higher tank Q would be necessary for sufficient loading.

Checking Circuit Values

Coil dimensions can be calculated from the formula in the *Handbook* or by means of the *Lightning Calculator*, in most cases, although when the inductance is small the results tend to become inaccurate. In this event, a better method is to adjust the coils to the proper inductance by measurement. A grid-dip meter and an inexpensive "standard" condenser are all that is needed using the method described in the chapter on measurements in the *Handbook*. Measurement can also be used for L_1, the plate tank coil, although either the formula or *Calculator* is amply accurate for such coils.

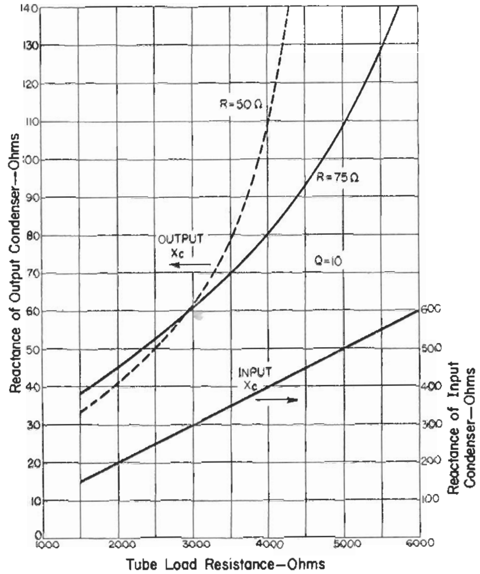


Fig. 1 — Pi-network design curves for working into 50- or 75-ohm loads with a Q of 10.

Pi-Network Tanks

For the special case where a pi network is to couple a tube to a flat coax line, much of the necessary design information can be presented in a few curves.

Fig. 1 gives the values of input and output reactance needed when the Q is held at 10. The curves are carried only as far as 140 ohms for the output reactance, since higher values can easily be obtained with a 300-$\mu\mu$f. variable condenser even at 3.5 Mc. When the Q is 10 (or more) the reactance of the coil averages 25 or 30 ohms higher than the reactance of the input condenser. It should be noted that the output reactance becomes infinite — that is, the output capacitance required is zero — when the ratio of actual load resistance to tube load resistance is equal to the square of the Q. Since the square in this case is 100, the output capacitance is zero for a 50-ohm line when the tube load is 5000 ohms. A higher tube load resistance cannot be "matched" unless the Q is increased above 10. Corresponding conditions obtain with the 75-ohm line when the tube load is 7500 ohms.

Fig. 2 shows the values required when the reactance of the output condenser is fixed at 60 ohms, a median value for both 50- and 75-ohm lines, and one that is useful over at least the range 35 to 100 ohms when the inductance is continuously adjustable. The approximate Q variation for various tube load resistances also is shown. If a larger Q is wanted for low tube load resistances (low ratios of plate voltage to plate current) it can be obtained by reducing the reactance of the output condenser. This in turn requires reducing both the input reactance and the coil reactance, as comparison of Figs. 1 and 2 for values of tube load resistance below 3000 ohms will show.

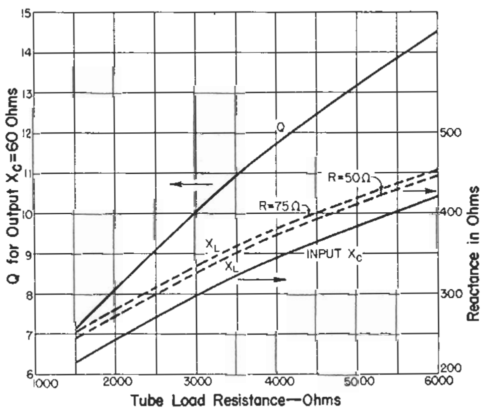


Fig. 2 — Pi-network design curves for working into 50- or 75-ohm loads with a fixed output capacitive reactance of 60 ohms.

From "Pi-Network Design Curves," *QST*, April, 1952.

A 'scope is the best monitor of your transmission you can have, provided you check the height of the pattern at the flattening point as shown by the two-tone test *and then keep below that height with voice.* Don't worry if the plate meter doesn't kick the way you'd like it to. The 'scope tells you you're getting all the peak power your amplifier can give, and it's telling the truth.

» An inexpensive linear amplifier of simple construction. The 807 will deliver a peak r.f. output of 50 watts in Class AB_2. A watt or two of s.s.b. signal is enough to drive it.

The "Single Side-Saddle" Linear

CARL W. ECKHARDT, W7BBK

THE Side-Saddle Linear is designed for the low-power station and will operate from existing 50-watt 600-volt power supplies. The 807, in the Side-Saddle Linear, performs efficiently as a Class B linear and will deliver approximately 50 watts of peak s.s.b. output at 75 watts peak input. Furthermore, the 807 is literally loafing at 600 volts plate voltage.

Every effort was made to "de-bug" the amplifier before construction by referring to the excellent articles on linear design by Reque and Long, and using straightforward mechanical and electrical layout. The resulting amplifier is stable and behaves properly, just as the experts say it should.

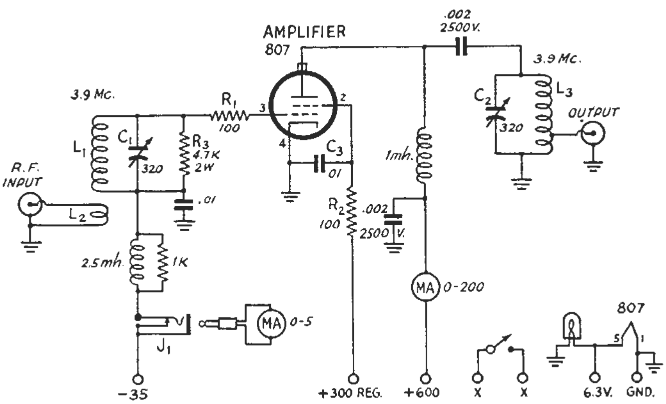


Fig. 1 — Circuit diagram of the "Side-Saddle" linear amplifier.

C_1, C_2 — Hammarlund MC-325-M or equivalent.

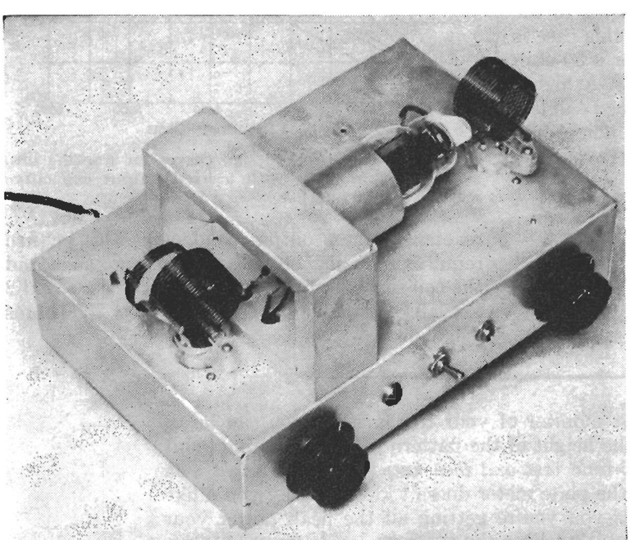
L_1, L_2 — 8-µh. Bud OES-40 with 3 turns removed from grid end of L_1

L_3 — 8-µh. Bud OES-40 with 3 turns and end link removed.

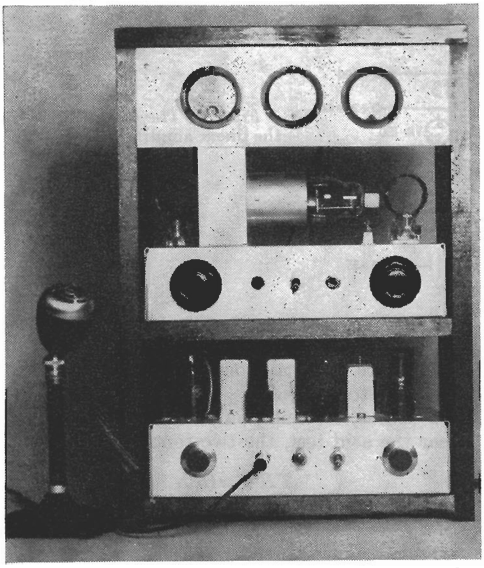
Resistors are 1-watt composition unless otherwise specified.

The receiver has been connected directly to the antenna coax transmission line through an electronic "TR" switch, and although the 807 is not biased to cut-off during receive periods (the exciter is), the linear is perfectly quiet, with no trace of thermal noise.

From *QST*, November, 1953.



The "Side-Saddle" linear amplifier uses a single 807 mounted horizontally.



In use, the linear amplifier is mounted in a wooden rack above the Edmonds crystal-filter exciter. The meters measure antenna current, grid current and plate current.

The drive requirement is approximately 2 watts. As Fig. 1 shows, the grid is series fed, and 35 volts of bias is supplied by batteries. Since approximately 1 ma. of grid current will flow on peaks, batteries provide the cheapest and most convenient method of supplying the required well-regulated bias voltage.

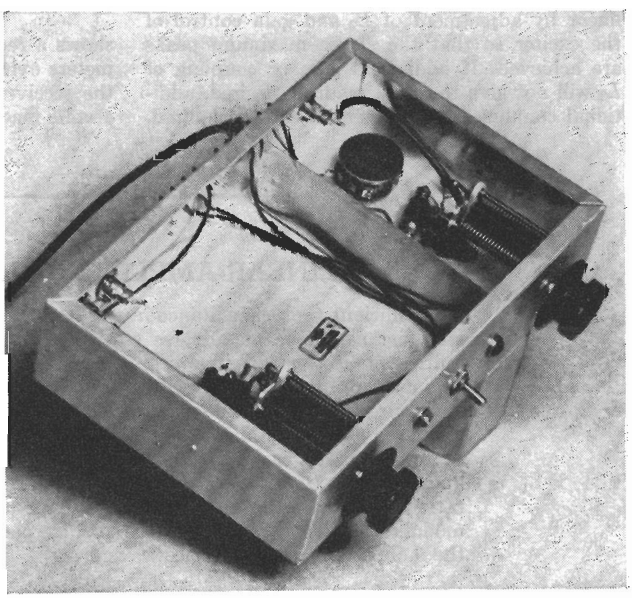
The plate is shunt fed, and 300 volts regulated is provided for the screen. Instability cannot be tolerated in a linear amplifier. The grid and plate circuits are effectively isolated above the chassis as illustrated, and by a 3-inch aluminum shield running the depth of the subchassis, near its center, thereby isolating C_1 from C_2. Further to guard against instability, small 1-watt noninductive resistors R_1 and R_2 are placed in the grid and screen leads directly at the socket. The cathode is grounded to the tube-support chassis with a short lead at the socket. The screen by-pass condenser C_3, a disk ceramic, is wired directly across the tube socket, keeping leads as short as possible. Noninductive resistor R_3 provides the proper amount of swamping.

The values given for the grid and plate tank circuits should be followed if proper circuit Q is to be maintained. The Bud coils specified must be pruned to get the desired L/C ratio. Three turns are removed from both L_1 and L_3. The end link of L_3 did not provide sufficient coupling and was cut from its supporting leads at the coil. One of the remaining link support leads is used as the antenna tap to L_3. Scrape a small section of the enamel covering from the 3rd, 4th, 5th and 6th turns from the ground end of L_3 and carefully solder the antenna tap to the proper turn during the tune-up procedure.

The amplifier is constructed on an 8 × 12 × 3-inch aluminum chassis. A 4 × 6 × 2-inch aluminum chassis provides the tube support as shown. The 807 is provided with a base shield. Grid and plate tank circuits are near the left- and right-hand ends of the chassis as illustrated. You will note that J_1 and the rotor of C_1 are at −35 volts bias potential and must be insulated from the chassis. This is done by slightly enlarging the mounting holes for C_1 and J_1, and insulating each of the bushings with a couple of fiber washers. The terminal strip and input and output coaxial sockets are on the rear of the chassis. An aluminum subchassis shield 7½ inches long by 2¾ inches high, with a ½-inch lip, should be placed as mentioned earlier.

A pilot light, a power switch, and J_3 are mounted on the front center of the chassis as shown. (The potentiometer located below the chassis is not being used and should be ignored.)

A bottom view of the linear amplifier. The potentiometer is not used.



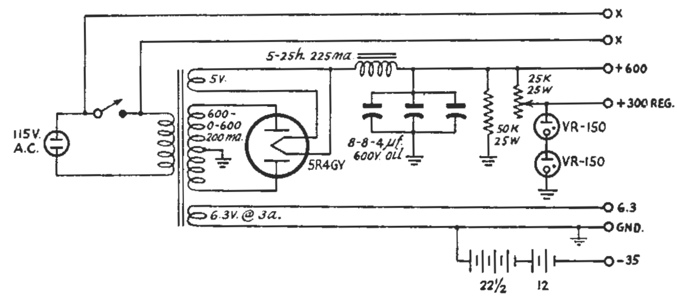


Fig. 2 — The power supply for the linear amplifier.

The power supply shown in the schematic of Fig. 2 delivers approximately 600 volts at 150 ma. to the plate of the 807, and 300 volts regulated to the screen.

Tune-up Procedure

Before applying power to the 807, check the bias and screen voltages. Without excitation to the linear, and applying power, the grid current should be 0 and plate current 10–15 ma. The amplifier should be perfectly stable without a trace of self-oscillation or parasitics.

In applying grid drive and antenna loading, adjust for the following goals:

	Grid Current	*Plate Current*	*Relative Ant. R.F. Current*
Idling (no voice)	0	10–15	0
Maximum peak, tone or whistle	1 ma. max.	140	max.
Normal speech peaks		100	high
Normal voice		10–15 to 60–70	0 to high

Relative readings are suggested for output r.f. current, as this will depend on the impedance of the line and the s.w.r.

Tap up from the ground end of L_3 with the antenna tap until the above maximum peaks are reached. Provide just enough excitation to the linear by adjustment of L_2 and gain control of the exciter so that the above maximum peaks are achieved. If additional gain or coupling of L_2 will not give the peak readings desired, additional coupling to the antenna will be required.

If you are overdriving, reduce the drive to the point where the peak conditions of plate current just begin to drop off.

Of course, in time you will wish to give your linear the acid test, the two-tone adjustment.

TR Switch

The TR switch in use is shown in Fig. 3, and is a modification of Cronin's system. It

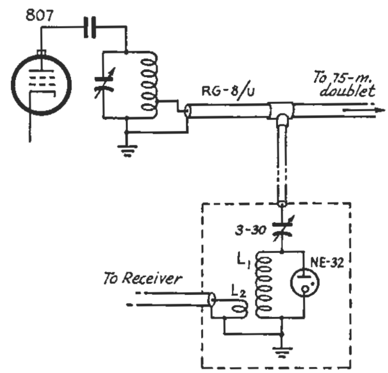


Fig. 3 — The TR switch allows the same antenna to be used for transmit and receive.

L_1 — 76 turns No. 30 d.c.c., 1¼-inch diam.

L_2 — 4 turns No. 20 flexible hook-up wire.

shows a receiving loss of about 1 S-point on 75 meters over a direct connection, and it protects the receiver to the point where a maximum of 4 volts reaches the receiver input terminals.

LINEAR-AMPLIFIER TANK Q

Insufficient tank capacitance in the tuned plate circuit of a linear amplifier can have an adverse effect on the linearity, but the requirements in this department are no more severe than in r.f. amplifiers used for other purposes. Theoretically, the tank Q could be zero with pure Class A operation, with more and more C needed as the operation passes through AB_1 and AB_2 to pure Class B. However, even in the latter case a tank Q of 10 or 12 is high enough to preserve linearity. The Q should be based on the plate-voltage/plate-current ratio under *peak* conditions — that is, at the maximum point of the modulation envelope — not on the average plate current as read by the d.c. plate meter with voice waveforms.

One hundred and twenty watts peak output from a pair of 6146s in parallel. No grid power is needed since the tubes are operated in Class AB_1.

The "Little Firecracker" Linear

BEN RUSS, W2QZ

Amplification of the output of any low-powered s.s.b. exciter can be done only by using a *linear* amplifier. A linear is an amplifier so adjusted that its output voltage is proportional to its input voltage. Use can be made of Class A, AB_1, AB_2 or B. Class A is generally used for very low power levels, as in the exciter output stage.

Not only do we want to amplify our s.s.b. signal, we want to amplify it without adding new and possibly undesired signals. If the unwanted sideband is 40 db. down in the signal coming from the exciter, we expect the same (or very close to the same) ratio in the antenna after amplification. We want no intermodulation products added that will either degrade the desired sideband or appear as "crud" outside the sideband. In short, we want a really "high-fidelity" amplifier for r.f. As in audio work, you can't expect a good linear to clean up a degraded signal fed to its input. That's expecting too much.

Desirable characteristics of any linear amplifier include:

1) Good linearity up to the power-handling limit of the tubes.

2) High power sensitivity.

3) Multiband operation without opening the cabinet.

4) High-*Q* *LC* circuits.

5) Constant-voltage plate, screen and grid supplies.

6) Stability.

From *QST*, September, 1953.

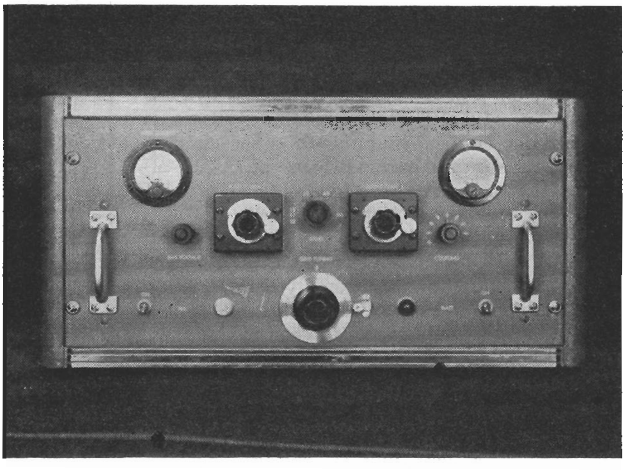
Linearity and Output

It should be recalled that the figures given for audio service in Class A, AB_1, AB_2 or B can be used for r.f. linear amplifiers used with s.s.b. suppressed carrier. Two tubes can be used in parallel or push-pull — we elected to use two tubes in parallel, for circuit simplicity. In push-pull or parallel, however, it is desirable to use tubes with similar characteristics so that the tubes will share the load equally.

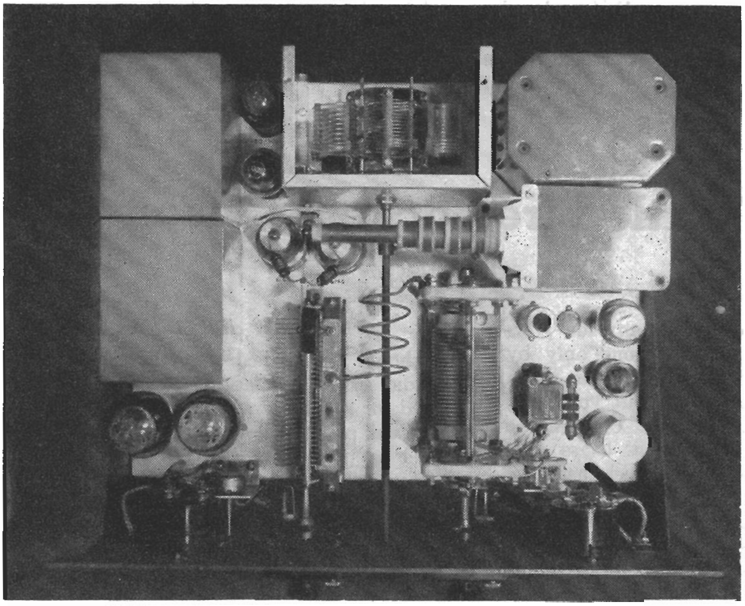
Table I gives the Class AB_1 and AB_2 ratings of the 6146, an excellent tube with a plate-dissipation rating of 25 watts. Slightly more output can be obtained in Class AB_2 operation, although running the tubes in Class AB_1 (no grid current) simplifies the driver problem and greatly reduces the chances for distortion on signal peaks in the grid-current region. In s.s.b. suppressed-carrier operation the maximum screen voltage (250) can be used, resulting in higher power sensitivity and slightly more peak output. The "maximum-signal d.c. plate current" is not what your meter reads on *speech.*

High Power Sensitivity

The 6146 is tailor-made for this desirable feature. Like all beam tubes, the 6146 requires very little drive, and you can figuratively "blow" on the grid and get high power output. Preliminary checking of this unit on all bands was done by driving the amplifier with the output from a Millen grid-dipper. A pair of 6146s will deliver full rated output when driven by any of



The "Little Firecracker" s.s.b. linear uses two 6146s in parallel. It operates on 80 through 10 meters without lifting the lid, and all power supplies are housed in the same cabinet as the r.f. components. The two dials at the center were taken from some war-surplus equipment.



This top view of the linear amplifier shows the grid-tank shield removed. The power-supply components are on each side of the chassis — the two screen VR tubes can be seen to the left of the grid circuit.

the commonly-used s.s.b. exciters. The exciter will not be required to work heavily, and swamping can be used to absorb the extra power from the exciter and add to its linearity. Swamping is used in both the exciter tank circuit and the linear-amplifier grid tank at W2QZ. The amplifier operates Class AB_1, with 120 watts peak output.

Multiband Operation

Multiband operation is not essential to s.s.b. but, as in any other transmitter, it is a nice feature to have. If a shelf full of plug-in coils can be eliminated, so much the better.

In this amplifier, a revised B & W turret is used in the grid circuit, and a B & W variable inductor is used in the output. The unit covers all amateur bands from 80 through 10 without opening the cabinet. The 160-meter band could have been included by switching in an additional inductor in the plate circuit.

High C

The subject of high Q (or high C) in tuned circuits has been stressed in many articles relating to s.s.b., as well as in the *Handbook*. High-Q circuits are used in linear amplifiers for two main reasons: for ease of coupling to other circuits, and to minimize the harmonic content. A rule-of-thumb that the s.s.b. gang follows is "Use a 10-meter coil on 20, a 20-meter coil on 40, etc."

Table 2 shows how this principle was applied to the B & W BTEL 35-watt turret used in the grid circuit (an end-link assembly with separate link windings for each of its five coils). A 250-$\mu\mu$f. variable is used to tune the coil in use.

The Q in the plate circuit can be set with reasonable accuracy, since a variable inductor and a high-capacity variable condenser are used in a pi-network circuit. The condenser is a 190-$\mu\mu$f.-per-section dual, and the stators are connected together by a ½-inch-wide copper strap. Connected this way, the condenser measures 40 to 400 $\mu\mu$f. (including stray capacity to the chassis). It is a simple matter to plot the capacity *vs.* dial setting, since the condenser plates are semicircular and give straight-line-capacity tuning, and hence the tuning condenser can be set to any given capacity on any band. The circuit is then tuned to resonance by adjustments of the variable inductor, and load changes are made with the output condenser stack of S_2. The Q chart in the *Handbook* was used for working out these settings — the maximum value of d.c. plate current (227 ma.) is used in the denominator.

Regulation of Power Supplies

In no other type of transmitter is the regulation of power supplies as important as in s.s.b.

TABLE I

Amplifier Ratings for the 6146 (Values Are for Two Tubes)

A.f. power amp. & mod.	AB_1
D.c. plate voltage	750
D.c. grid No. 2 screen voltage	200
D.c. grid No. 1 control grid	−50 volts
Zero-signal d.c. plate current	57 ma.
Maximum-signal d.c. plate current	227 ma.
Zero-signal d.c. grid No. 2 current	1 ma.
Maximum-signal d.c. grid No. 2 current	27.5 ma.
Maximum-signal driving power (approx.)	0 watt
Maximum-signal power output	120 watts
A.f. power amp. & mod.	AB_2
D.c. plate voltage	750
D.c. grid No. 2 screen voltage	165
D.c. grid No. 1 (control grid) voltage	−45
Zero-signal d.c. plate current	35 ma.
Maximum-signal d.c. plate current	240 ma.
Zero-signal d.c. grid No. 2 current	0.6 ma.
Maximum-signal d.c. grid No. 2 current	21 ma.
Maximum-signal d.c. grid No. 1 current	0. 7 ma.
Maximum-signal driving power	0.07 watt
Maximum-signal power output	130 watts

TABLE II

L_1 L_2 — B & W BTEL 35-Watt Coil Turret

All primaries left as is. Secondary turns shorted from hot end (opposite primary link).

Original Use	*Revision*	*Band Use*
10–11 meters	4 turns shorted	10–11–15 meters
15 "	3 " "	20 "
20 "	3 " "	40 "
40 "	4 " "	80 "
80 "	None	Not used

transmission. For a clean s.s.b. signal, the only factor governing the output signal should be the r.f. driving signal at the input. With the exception of Class A operation (used only at low power levels because of its relative inefficiency), the average plate and screen currents vary over wide ranges at a syllabic rate. If grid current is drawn during part of the operating cycle, as in AB_2 and B operation, the grid current also varies. If these variable current demands affect the voltages of the power and bias supplies, the effect is to add some plate or screen (or grid) modulation, or a combination of them, to the output. These will generate new and undesired components, or distortion products, and they will degrade the overall signal. It is therefore important to use well-regulated power supplies with a s.s.b. linear amplifier.

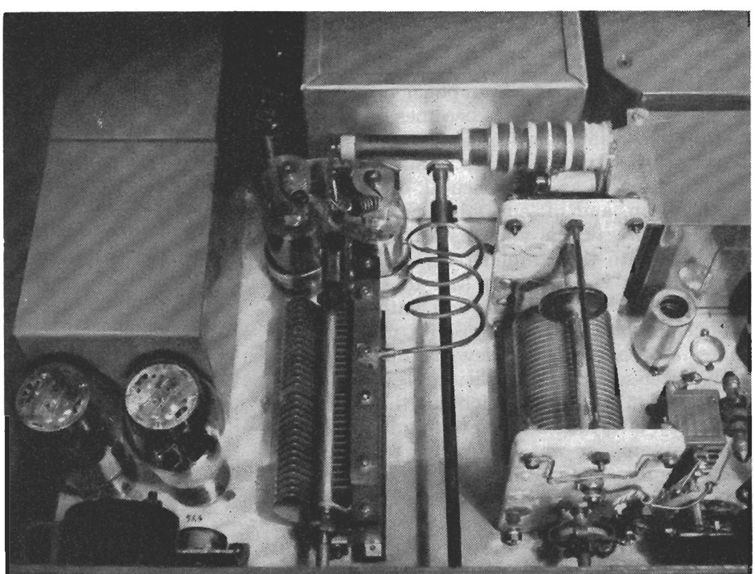
The requirement of a grid-bias supply well regulated over a wide current range is avoided by using 6146s, which draw a very low value of grid current on peaks, and by using a "stiff" bias supply. The bias voltage is adjustable and, in practice, there is no detectable voltage change, even on modulation peaks. The bias supply is "stiffened" by using a low-resistance voltage divider and a large electrolytic condenser across the output.

The screen supply is easy to regulate, since the maximum screen-current demand is less than the maximum current a VR tube will handle, and a VR-105 and VR-150 in series give the necessary 255 volts. The dropping resistor to the VR tubes is adjusted so that, with no modulation, the VR tubes pass their maximum current of 40 ma.

Getting the best possible regulation from the average plate power supply is no easy task. There are many components that contribute to the *IR* drops; e.g., the transformer secondary, the rectifier tube (if high-vacuum), the input and smoothing chokes, and the a.c. line. To smooth out the syllabic variations, a large amount of output capacity is recommended, and too many microfarads at this point are not within the realm of possibility. One full-wave rectifier could have been used in the rig to be described, but two were used, to reduce the *IR* drop. Mercury-vapor tubes would be even better, perhaps, in this respect. Originally, one surplus plate transformer was used. There were no current ratings on the case, but the transformer looked big enough and felt heavy enough. Later, a duplicate transformer was acquired and connected in parallel. Immediately the regulation was improved, and the *IR* drop through transformer resistance was cut in half. Don't go right out and buy two similar transformers — the point is mentioned just to emphasize the need for conservatively-rated power-supply components when good regulation is your objective.

It was always considered desirable to have two matched filter chokes, one swinging and the other smoothing. Yes, you read it right — it says "was." In s.s.b., only one is sufficient, and using only one cuts down another source of series resistance and poor regulation. With 24 μf. of filter condenser and the swinging choke, the regulation is improved and the ripple is only 2½ per cent.

A close-up view of the r.f. section shows the neutralizing condenser mounted close to the two 6146 plate connectors.



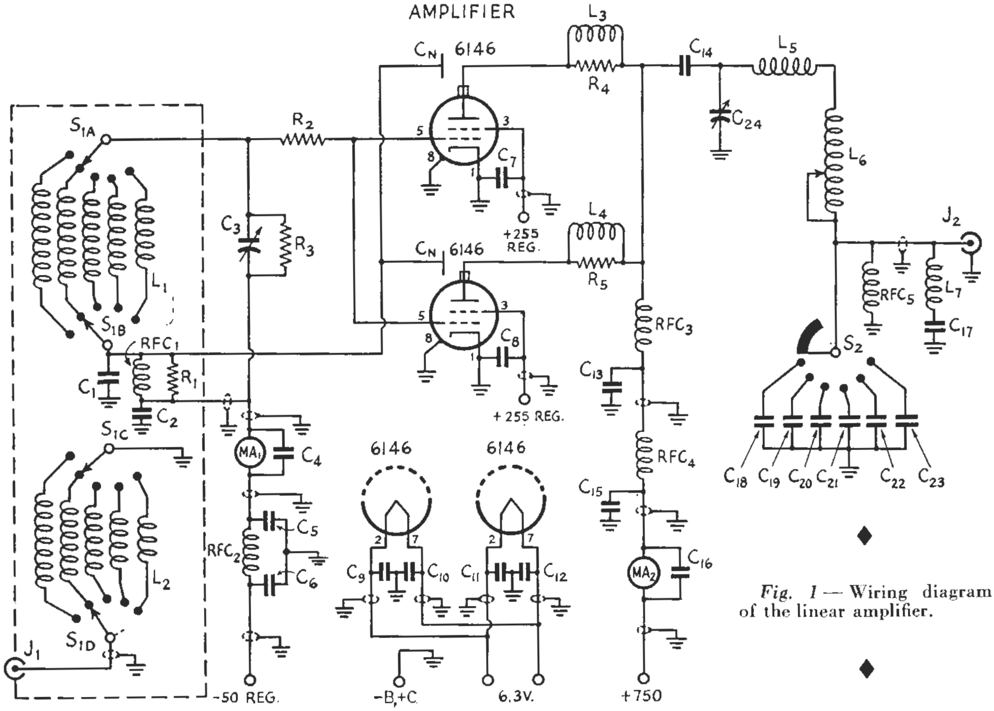


Fig. 1 — Wiring diagram of the linear amplifier.

C_1 — 250-$\mu\mu$f. mica.
C_2, C_4 – C_{12}, C_{16} — 0.001-μf. disk ceramic.
C_3 — 250-$\mu\mu$f. variable (Hammarlund MC-250M).
C_{13} — 500-$\mu\mu$f. 10,000-volt ceramic (CRL TV3-5301).
C_{14} — 0.002-μf. 2500-volt mica.
C_{15} — 0.001-μf. 2500-volt mica.
C_{17}, C_{18}, C_{19} — 100-$\mu\mu$f. 2500-volt mica.
C_{20}, C_{21} — 200-$\mu\mu$f. 2500-volt mica.
C_{22}, C_{23} — 500-$\mu\mu$f. 2500-volt mica.
C_{24} — 380-$\mu\mu$f. variable (Cardwell MO-180-BD, stators in parallel).
C_n — Neutralizing condenser (see text).
R_1 — 1000-ohm 2-watt carbon.
R_2, R_4, R_5 — 100-ohm 2-watt carbon.
R_3 — 10,000-ohm 10-watt noninductive (Sprague Koolohm NIT).
L_1, L_2 — See coil table.
L_3, L_4 — 6 turns No. 20 tightly wound around R_4 and R_5
L_5 — 3½ turns No. 10, 2-inch diam., 3½ inches long.
L_6 — Variable inductor (B & W 3852).
L_7 — Series-resonated with C_{17} to TV channel most likely to be interfered with.
MA_1 — 0–15 milliammeter.
MA_2 — 0–500 milliammeter.
RFC_1, RFC_5 — 2.5-mh. r.f. choke.
RFC_2 — 7-μh. r.f. choke (Ohmite Z50).
RFC_3 — 225-μh. r.f. choke (National R-175).
RFC_4 — 4-μh. r.f. choke (National R60).
S_1 — 4-pole 5-position switch of L_1L_2 assembly.
S_2 — Progressive shorting switch (Centralab PIS wafer and P-121 index).

Stability

Stabilization of any r.f. amplifier is important, of course, but it sometimes becomes more important in linear-amplifier work because of the high power sensitivity of the linear. Instabilities manifest themselves as fundamental oscillations or regeneration, and l.f. and v.h.f. parasitics. As a matter of precaution, v.h.f. parasitic suppression was built into this amplifier before any testing was done. It is reasonably safe to assume that there is everything to gain and nothing to lose by exercising this precaution with multigrid r.f. tubes, especially where they are operated in a high-power-sensitivity condition.

Oscillation or regeneration at the operating frequency is evidence of plate-to-grid feed-back and the amplifier was neutralized to offset any such coupling. As originally built, no neutralization was included, although every precaution was included to minimize feed-back, such as grounding all cathode pins with heavy wire and isolating the grid and plate circuits. The amplifier was stable, but after several days of band-hopping between 75 and 20 it was noticed that maximum output did not occur at the resonance dip indicated by the plate-current meter. Adding capacitance bridge neutralization brought maximum output at the plate-current dip.

The introduction of this neutralization introduced another factor in the form of "birdies" across the band, similar to ITV. This was an unexpected l.f. oscillation that was eliminated by the addition of a 1000-ohm resistor across the grid r.f. choke. Another value of r.f. choke would probably have cured the trouble, but none was available at the time.

The amplifier circuit is shown in Fig. 1, and the power supply diagram is shown in Fig. 2. The thermal time delay, K_1, was included to insure proper warm-up time for the 6X4 and 6146s.

Construction

The panel and chassis are aluminum, with an 8¾ × 19-inch panel and a 13 × 17 × 3-inch chassis. The large chassis allows for mounting of many of the smaller components on the underside. The chassis is set back from the panel, to allow room for the mounting of parts between the panel and the front apron of the chassis. This spacing also serves to bring the rear apron, with its coax connectors and terminals, flush to the inside back of the cabinet.

All wiring was done with shielded wire where possible, with no attention being paid to its length. It is bonded and grounded to the chassis at all convenient points. Disk ceramic condensers are used liberally throughout, in keeping with present practice. The grid-tuning condenser is mounted directly under the grid-turret shield box, permitting short lead lengths.

A series of ¼-inch holes is drilled directly above R_6, the series bleeder, for better heat dissipation. A current of 40 ma. flows through this resistor when the plate switch is "on."

Condenser C_{17} is mounted as close to the output coax connector, J_2, as possible, and flush against the chassis. It is "grid-dipped" with L_7 to the frequency of the TV channel most likely to be interfered with. J_2 is shorted during the grid-dipping process.

The "C" bias dual filter condenser, $C_{11}C_{12}$ (Fig. 2), is mounted on an insulating wafer because the can is negative and does not ground to chassis. A disk ceramic condenser, C_{10}, is used to by-pass the can to the chassis.

Type 4 mica condensers are used in the pi-network output stack because they lend themselves to firm mechanical mounting. Their voltage rating is higher than required for the power used, but they were on hand. The "safety choke," RFC_5 in Fig. 1, is mounted across the condenser stack.

Both sides of the plate-tuning stators are connected together with ½-inch-wide copper strap to reduce lead inductance and make use of the condenser's maximum capacitance. The self-resonance of this condenser, C_{24}, connected this way, measures 120 Mc. — well outside any TV band.

The plate choke, RFC_3, is mounted horizontally, to bring the "hot end" directly centered over the plate caps of the 6146s. Here, too, ½-inch copper strap is used to connect the end of the choke to the plate blocking condenser, C_{14}, resulting in short leads and low inductance. This choke works as well on the 15-meter band as it does on 10, when we used the grid-dipper as the source of excitation for preliminary tests. This may be due to the proximity of the adjacent components adding sufficient capacitance to

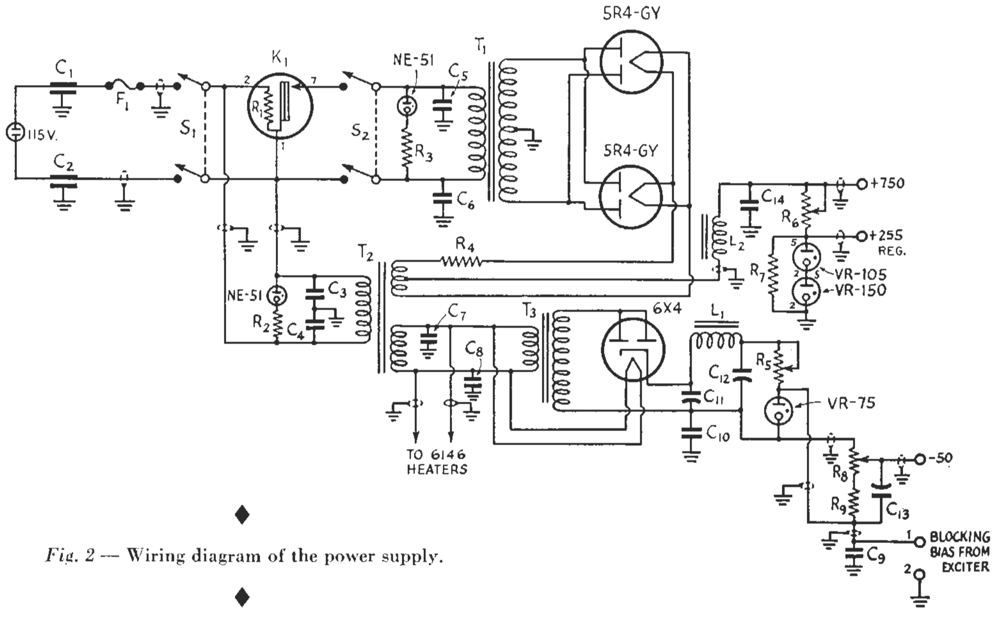
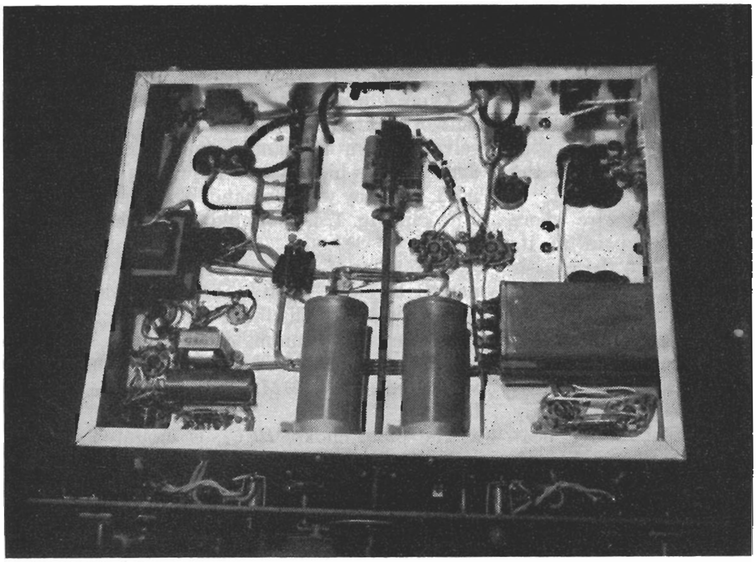


Fig. 2 — Wiring diagram of the power supply.

C₁, C₂ — 0.1-µf. feed-through capacitor (Sprague Hypass).
C₃ – C₈ — 0.005-µf. disk ceramic.
C₉, C₁₀ — 0.001-µf. disk ceramic.
C₁₁ — 10-µf. 450-volt electrolytic (in same can with C₁₂).
C₁₂ — 80-µf. 450-volt electrolytic (Mallory FP-245).
C₁₃ — 80-µf. 150-volt electrolytic.
C₁₄ — 24-µf. total, 1000 volts.
R₁ — Heater built in *K₁*.
R₂, R₃ — 0.2 megohm, 1 watt.
R₄ — Homemade 0.35-ohm resistor. Can be omitted if *T₂* has 5-volt winding.
R₅ — 1000-ohm 10-watt adjustable.
R₆ — 20,000-ohm 50-watt adjustable.
R₇ — 0.1-megohm, 2 watts.
R₈ — 2500-ohm 2-watt potentiometer (Ohmite AB CU2521).
R₉ — 2400 ohms, 2 watts.
L₁ — Small filter choke, not over 250 ohms resistance.
L₂ — 5–20 hy. 300-ma. swinging choke.
F₁ — 5-amp. fuse.
K₁ — Thermal time delay (Amperite 115NO30 wired for 6 amp.).
S₁, S₂ — D.p.s.t. toggle.
T₁ — Plate transformer, to deliver 750 v. d.c. at 300 ma., approx.
T₂ — Filament transformer, each secondary 6.3 volts at 4 amp.
T₃ — 6.3-volt 3-amp. filament transformer.



The 24 μf. of filter condenser in the plate power supply is mounted under the chassis — it consists of the two center cylinders plus the two condensers on the right-hand side. Shielded wires and disk ceramics are liberally used throughout the unit.

move an expected "hole" out of the 15-meter band.

Parts R_1, RFC_1, and C_1 are mounted just inside the input turret shield can, and directly in back of the 6146s. A lead is brought through close to the bottom of the vertical side of the shield can, by means of a feed-through insulator. This lead connects to the vertical part of the neutralizing condenser, C_N. The neutralizing condenser is made of a thin aluminum sheet cut in the shape of the letter "T," with the horizontal part of the T 1 inch high and 4 inches wide. It is mounted on a ceramic post so that it can be bent away from the plates of the 6146s.

Tuning and Loading

Tuning is straightforward as with any amplifier of similar design. However, careful attention should be paid to loading.

Light loading cannot be tolerated because (a) the screen dissipation may be exceeded. (An indication of this can be seen by the screen VR tubes being extinguished. When the amplifier is properly loaded, these VR tubes should not "go out" during any part of the modulation cycle). (b) "Flattening" of the peaks will occur before rated output is reached, assuming proper drive requirements are met. If reduced power output is wanted, just turning down the speech level control will do the trick. There is a point of loading where maximum rated output occurs without flattening, and loading beyond this point serves only to reduce the efficiency and output.

A word about tuning. As with any r.f. device that covers a wide band of frequencies, caution must be exercised to be sure that only the desired output frequency is amplified. This is especially advisable when using beam or tetrode tubes, because of their high power sensitivity.

With the grid circuit tuned to resonance at the fundamental, the plate circuit should be tuned starting with the *maximum* number of turns in the variable inductor. The first resonant point is the desired frequency.

Results

This linear has been on the air regularly on 20 and 75 meters, and has been dummy-load tested on the other bands. Stable as the Rock of Gibraltar, all reports indicate that it transmits a high-quality signal with excellent sideband suppression, and a minimum of distortion products. This result was the hoped-for goal when the unit was in the design stage. That it is so is due to the application of good design practices for s.s.b. transmission.

It is affectionately called "Little Firecracker"; for its small size, it sure gives a loud report.

A word about TVI — this is the first transmitter in use at this station to prove itself completely TVI-proof. With this unit operating, the family, watching any one of the seven local TV stations, was completely unaware of its being in use. Furthermore, at the time, the linear was out of its cabinet and no low-pass filter was used.

For his patient efforts photographing the "Little Firecracker," special thanks to Henry Marcus, W2AJX.

There is no magic in s.s.b. that makes it cause less TVI than a.m. It is, however, the most effective type of 'phone signal to use with a linear amplifier, and a properly-adjusted linear amplifier will not generate a very high percentage of harmonic energy. "Fundamental overload" can still be a problem with s.s.b. and a good linear amplifier, of course, but it must be cured at the receiver, in any case.

» The comparatively low duty cycle of s.s.b. offers opportunities for power-supply economy, and Class AB_1 offers ease of adjustment and convenience in operating. These features are the keynote of this amplifier design. Four 1625s (or 807s) in push-pull parallel give 200 watts peak output.

The "Four-in-Line" Linear

GEORGE GRAMMER, W1DF

Class AB_1 operation has only one important disadvantage: Nearly all tubes (the 6146 is about the only exception) will deliver considerably more power — of the order of twice as much, in most cases — when driven into the grid-current region than they will in AB_1. However, for a given power *input* practically the same output can be obtained from either AB_1 or AB_2. This means that a given plate power supply can be utilized with equal efficiency with either type of operation. In order to obtain equal power inputs it will be necessary, as a rough approximation, to use twice as many r.f. tubes in AB_1 as in AB_2. This is not always an uneconomical approach, because the savings in other equipment that become possible with operation without grid current often can offset the extra r.f. tube cost. Also, the amplifier design automatically becomes more conservative, since doubling the number of r.f. tubes doubles the available plate-dissipation capacity.

There are distinct advantages in Class AB_1 with ample plate dissipation capabilities as against grid-current operation with just-enough plate dissipation:

1) The extra plate dissipation provides a worth-while safety factor for making tests and adjustments.

2) With more plate-dissipation capacity available, the resting d.c. input can be greater. This reduces the percentage change in d.c. input with voice excitation, helping to relieve the problem of plate-voltage regulation. The output condenser in the supply filter has less work to do.

3) Similarly, there is less strain on the screen supply.

4) It may even be possible to get an a.m. carrier of useful size out of the amplifier.

Just to round out the picture, the well-known advantages of Class AB_1 operation should be tabulated:

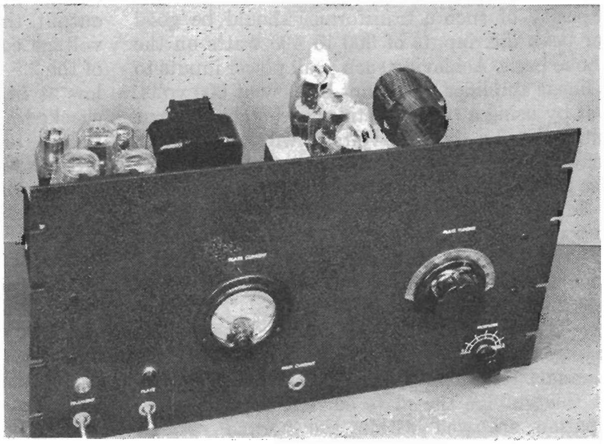
5) The driver can be a very small tube and needs no swamping for maintaining linearity.

6) The bias supply can be a quite inexpensive, noncritical arrangement instead of requiring special means for insuring that the bias does not change with grid current.

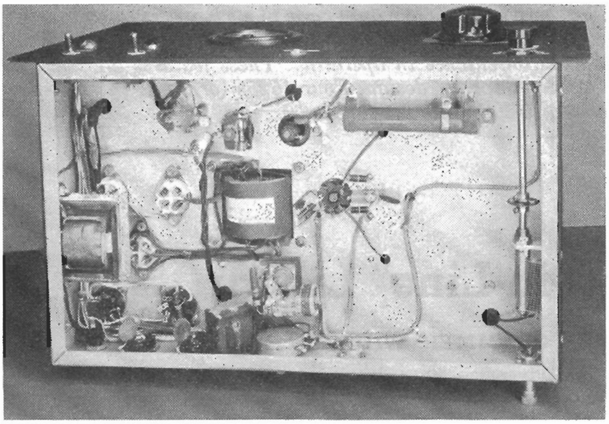
7) A meter in the d.c. grid circuit makes a simple and positive "overmodulation" indicator, practically independent of the modulation waveform.

The list is extensive enough to warrant serious consideration, we believe. Perhaps the greatest advantage of all is that this approach is practically painless, both designwise and operationwise, compared with the "squeeze 'er to the limit" method. No battling with grid-circuit regulation and coupling problems, and the rather comforting knowledge that if you do happen to forget and whistle a fraction of a

From "The Case for the AB_1 Linear," *QST*, April, 1954.



This linear amplifier uses four 1625s in push-pull parallel in Class AB_1. It is complete with power and bias supplies on a 17 × 10 × 3-inch chassis. R.f. input and output circuits are designed for coax lines, with the grid circuit bandpassed so that only the plate tank and output coupling need be adjusted. The peak output from the tubes is approximately 200 watts. The grid-current jack is for a meter for monitoring peaks. The panel is 10½ by 19 inches.



The only r.f. components underneath the chassis are the socket for the grid tank, grid loading resistors, and the variable condenser for output coupling adjustment. The bias supply is the group of components in the lower center in this view. The 12.6-volt filament transformer is mounted on the left chassis wall and the filament transformer for the 83 rectifiers projects through the chassis near the center. The latter transformer is a homewound job, but transformers of similar ratings are available ready-made.

second too long in the mike you won't burn a hole in the plate.

A Practical Example

The fact that the ratio of r.f. output to d.c. plate input is substantially independent of the type of operation suggests the thought that there may be some merit in selecting the number and type of r.f. tubes to fit the power supply, rather than the more usual method of selecting the power supply to fit the tubes.

Speech waveforms are such that in s.s.b. linear operation the maximum demand — that is, d.c. as indicated by the plate milliammeter — on the plate supply is about half the peak d.c. input. Also, this demand occurs only on peaks that in turn occur only at a syllabic rate — that is, intermittently — so the *average* demand is considerably less. Since it is the average demand that determines the heating in the plate-supply components, it is safe to assume that the power capacity of the supply, based on continuous ratings, need not be more than one-third to one-fourth the peak d.c. power required.

In the search for compact, low-cost power the replacement transformer offers a fertile field for moderate power outputs. The largest of these transformers, in terms of voltage, is the 600-volt 200-ma. type. The 120-watt output capacity of such a transformer should be good for peak d.c. inputs of 300 to 400 watts on the above basis. Achieving such peak power inputs to a linear amplifier would not be easy at 600 volts, but by using a bridge rectifier and a choke-input filter the d.c. output voltage will be approximately 1000 volts, a more reasonable figure for the purpose.

Several types of tubes would work well at this voltage, but 1625s were used in the amplifier described here because they can still be purchased cheaply in surplus; also, 807s can be substituted and are not at all expensive.

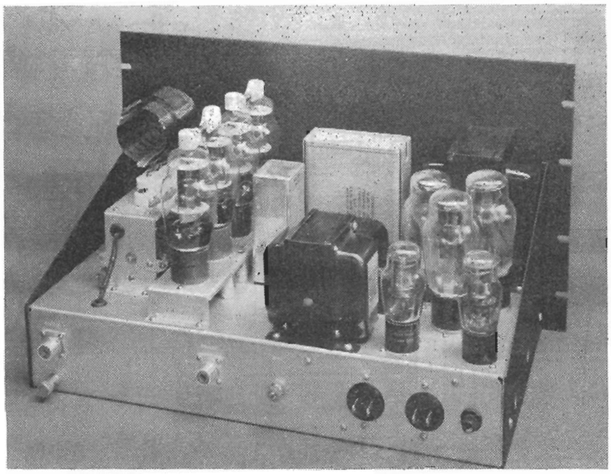
A plate voltage of 1000 is in excess of any ratings explicitly given for these tubes in the tube manuals. However, with Class AB_1 operation at this voltage the plate current and plate dissipation are well within the normal ratings. The plate voltage, while higher than the c.w. rating, is not as high as the maximum value reached on peaks at the plate-modulation rating.

Optimum Class AB_1 operation of 1625s or 807s calls for a peak instantaneous plate current per tube of 220 ma. at full drive. The corresponding d.c. plate current depends somewhat on the bias voltage, but is approximately 70 ma. per tube. To utilize the power-supply capacity fairly well four tubes are needed, taking a total peak power input of 280 watts. As stated above, this peak is about twice the maximum demand on the supply, with speech, so the maximum d.c. power is no more than 150 watts. The peak tube output, as taken from the characteristic curves, is approximately 200 watts from the four tubes.

R.F. Circuit

The logical circuit arrangement for four tubes is to use them in push-pull parallel. As shown in Fig. 1, parallel plate feed is used in this amplifier, principally to take the d.c. off the plug-in tank coil for safety reasons. The chokes originally were the familiar 2.5-mh. type, which worked satisfactorily in normal operation on 75 meters ("normal operation" means operation as a linear amplifier on voice; although the tubes have enough capacity to operate continuously at peak output, these chokes will not stand the peak r.f. voltage continuously). However, the impedance of the 2.5.-mh. type turned out to be undesirably low at 14 Mc. so a number of other types were checked. The ones finally used were Millen type 34107, 1 mh.; these showed good characteristics on all three bands.

Tubes in AB_1 require no driving power except that necessary to overcome circuit losses, but most exciters do have a small amount of power output available. This "waste" power can be used to eliminate a tuning control. As shown in Fig. 1, a fixed-tune grid circuit is used. A circuit having a Q of 8 will have substantially uniform response over a 200-kc. band centered at 3900 kc., so the L/C ratio of $L_1C_1C_2$ is chosen to give approximately this Q in conjunction with the loading resistors, R_1 and R_2. The values of R_1



The power supply occupies the right-hand half of the chassis and the r.f. section the left-hand half in this view. The power transformer and filter condenser are near the panel and the filter choke is at the edge of the chassis next to the voltage-regulator tubes.

The four r.f. tubes are mounted on an elevated subchassis so that the cathodes can be directly grounded to the top of the main chassis. The plug-in grid circuit is in the can to the right of the tubes. The small ceramic stand-offs visible beneath the subchassis support the metal tabs which form one of the neutralizing condensers. A similar pair, hidden by the shielded grid circuit, supports the other neutralizing condenser.

and R_2 were chosen so that the total power dissipated in them would be about one watt at peak excitation. These resistors constitute the only load on the tuned grid circuit — they are in no sense swamping resistors — and since the load is constant it is possible to adjust L_2, the coupling coil, to offer a definite input impedance to the connecting line from the exciter. This can be done quite easily with a standing-wave bridge (the amplifier tubes do not even have to be lit) and in the case of the amplifier shown, the inductances of the coils were adjusted to give

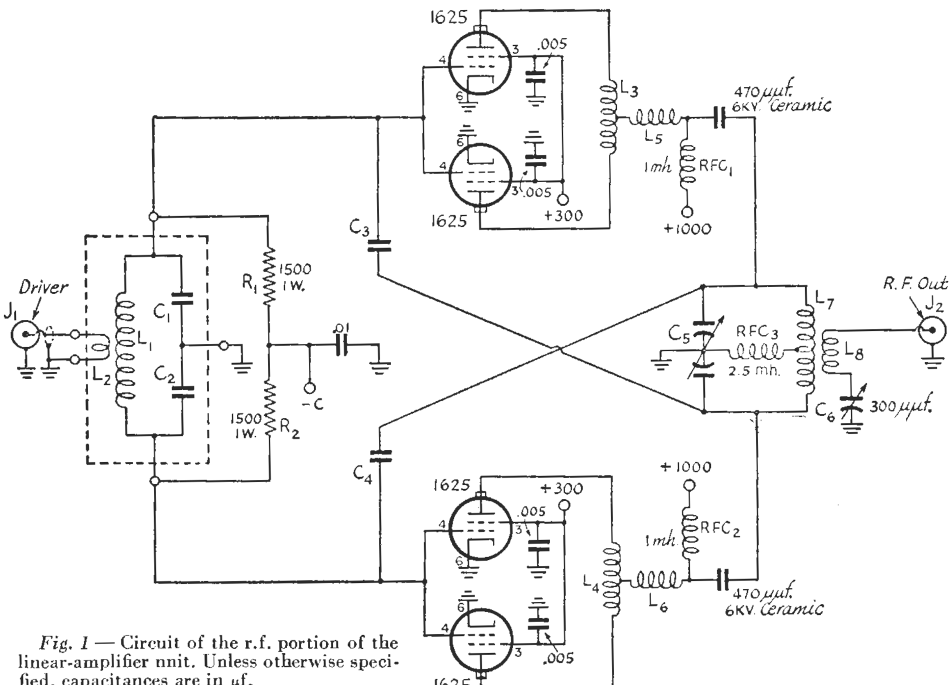


Fig. 1 — Circuit of the r.f. portion of the linear-amplifier unit. Unless otherwise specified, capacitances are in μf.

C_3, C_4 — Copper tabs 3/8″ wide, app. 1/4″ separation, 1/2″ overlap.
C_5 — 180-μμf.-per-section, 0.07-inch spacing.
C_6 — 300 μμf., receiving spacing.
L_3, L_4 — 18 turns No. 22 enam. on 1-watt resistor (any high value) as form, tapped at center.
L_5, L_6 — 12 turns No. 22 enam. on same type form.
L_2 wound over L_1 at center on 3.5 and 7 Mc.; interwound with L_1 on 14-Mc. coil. Coil forms 1-inch diam.
L_7 and L_8 made from B & W coil stock, L_7 2-inch diam. (3907 and 3900), L_8 2 1/2-inch diam. (3906), assembly mounted on Millen 40305 plug base.
The grid tuned circuit, enclosed by dashed line, is mounted in Millen 74400 plug-in base and shield.

Tuned Circuits

	3.8–4.0 Mc.	*7.2–7.3 Mc.*	*14 Mc.*
L_1	31 turns No. 22 enam. close-wound	17 turns No. 22 enam. close-wound	12 turns No. 22 enam. length 5/8-in.
L_2	4 1/2 turns No. 22	2 2/3 turns No. 22	2 3/4 turns No. 22
C_1 C_2	200 μμf. silver mica	100 μμf. silver mica	50 μμf. silver mica
L_7	28 turns No. 16 10 turns/in.	18 turns No. 14 8 turns/in.	8 turns No. 14 8 turns/in.
L_8	10 turns No. 14 8 turns/in.	6 turns No. 14 8 turns/in.	2 turns No. 14 8 turns/in.

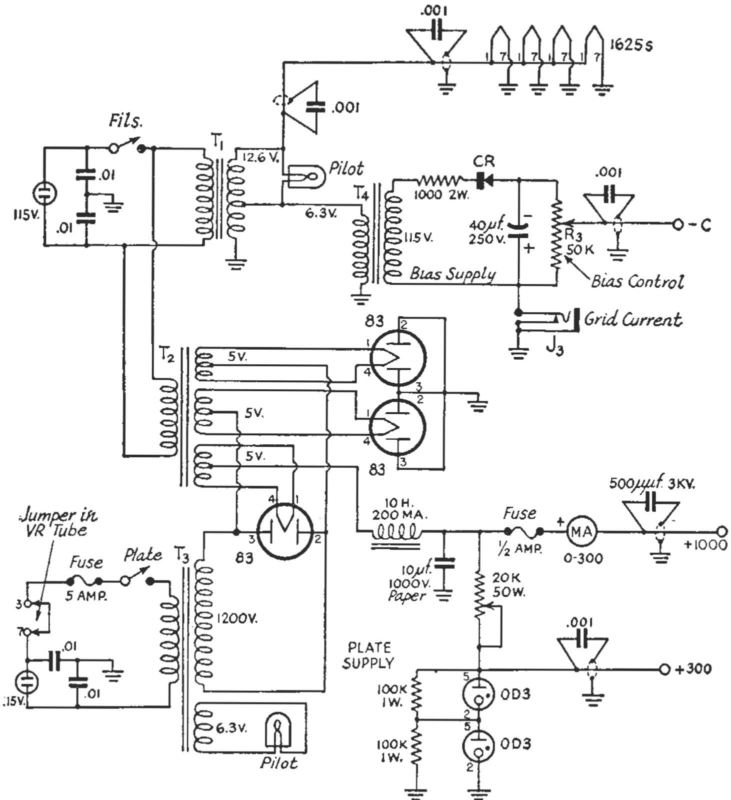


Fig. 2 — Power and bias supplies. Capacitance values are in μf. unless otherwise specified.
CR — Selenium rectifier.
T_1 — Filament transformer, 12.6 volts, 2 amp.
T_2 — Rectifier filament transformer, three 5-volt 3-amp. secondaries.
T_3 — 600-volt 200-ma. replacement-type transformer. Filament windings not used except for pilot light.
T_4 — Filament transformer, 6.3 volts, 1 amp.

close to a 1-to-1 s.w.r. in 75-ohm line at the band center. On the 75-meter band the maximum s.w.r., which occurs at the band edges, 3.8 and 4 Mc., is under 1.5 to 1. The deviation over the 7- and 14-Mc. 'phone bands is less since the percentage bandwidth is less.

This method of adjusting coupling is a great convenience, since the exciter and amplifier can be connected by any length of 75-ohm line with no change in the coupling conditions. The method is more difficult to use with an amplifier that takes grid current because the load varies with the driving voltage. Also, the small resistance-type s.w.r. bridges cannot be used in such a case because the s.w.r. has to be measured at the full driving power level.

Parasitic oscillations were anticipated, and two v.h.f. modes turned out to be present. One, a push-pull type oscillation at around 180 Mc. between the two tubes of each pair, was cured by installing the detuning inductances L_3 and L_4. The other, at about 150 Mc., was the usual type with these tubes and was similarly cured by L_5 and L_6. Except in cases where resonance at a frequency in a particular TV channel has to be avoided to prevent TVI, there is nothing very critical about these coils beyond the fact that they must be large enough to put the plate-circuit lead resonance below the self-neutralizing frequency. With the constants given, this resonance is at about 50 Mc. This is close to Channel 2, so anyone in a Channel 2 region who is interested in a similar circuit arrangement would be well advised to increase the inductances of L_5 and L_6 to move the resonant frequency lower.

The circuit is cross-neutralized by means of C_3 and C_4. Although the amplifier was wholly stable, insofar as self-oscillation at the operating frequency was concerned, without neutralizing, tuning the plate circuit with the plate voltage applied and the grids driven just above the grid-current point would swing the grid current through a range of a few hundred microamperes, so the neutralizing condensers were installed to reduce this reaction in the thought that it would improve linearity.

Power Supply

Tests on a "600-volt 200-ma." transformer had shown that it could deliver an r.m.s. current of 200 ma. from the entire high-voltage secondary without undue heating, operating continuously over a period of a few hours. This is an a.c. power of 240 watts, good for a d.c. output of about 200 watts with a choke-input filter having a choke of adequate inductance. Allowing for voltage drops in the rectifier tubes and choke, the output voltage is close to 1000 at a load current of the order of 200 ma.

The entire output current could be available for the amplifier plates if the tubes were used as a "free bleeder," but in the circuit of Fig. 2 part of the current is used for the screen grids to avoid the necessity for a separate screen supply. The peak d.c. screen current does not exceed 40 ma. for the four tubes, and is thus within the current rating of VR tubes. Since the screen current is practically zero with no r.f. on the control grids, the 20,000-ohm dropping resistor is adjusted to make the current in the voltage-regulator circuit 40 ma. under this condition. With a 10-henry filter choke the bleed current required to prevent the output voltage from building up appreciably above its load value is about 100 ma., so the r.f. tubes should be biased to take a minimum of 60 ma. with no signal.

The filter, a single-section affair using a 10-μf. condenser with the 10-henry choke, provides ample filtering and good voltage regulation under AB_1 conditions of operation. A 1000-volt rating on the condenser might seem to be a bit skimpy, considering the fact that loss of bleed and load might send the output voltage up to the a.c. peak, or 1700 volts. However, there is a considerable safety factor in the condenser ratings (commonly, the test voltage is at least 1000 above the rating in a condenser of this size) and the transformer primary circuit is interlocked with the VR tubes, which provide enough bleed to keep the voltage from getting near the peak.

If the amplifier plate voltage is left on during receiving periods the bias preferably should be left at the operating value and not switched to a value beyond cut-off, since this would leave only the VR circuit as a load on the supply. Although a TR system using the same antenna has not been tried, there is no trace of noise from the amplifier, at normal bias, when using a separate receiving antenna so long as the last tube in the exciter is cut off. In other words, the amplifier itself does not generate enough noise to be heard, with separate antennas, but it will amplify any noise fed to its grid circuit to the point where it can be annoying. The solution is of course simple.

The bridge rectifier uses three 83s, which are quite capable of handling the current but are slightly above rating on inverse peak voltage. Since this was a popular combination at the same applied voltage some years back, the writer had no hesitancy in using them. Four 816s could be substituted at a small increase in cost.

The bias supply uses a low-current 6.3-volt filament transformer with its 6.3-volt winding connected to half of the secondary of T_1. With the constants given in the diagram this provides a maximum of about 100 volts bias. The value of the bias control resistor can depart widely from that shown, although it should not be so low as to load the circuit unduly since this would require a larger filter condenser. Neither should the resistance be too high (in the hundreds of thousands of ohms) because high resistance has a limiting effect when peaks run into the grid current region and thus tends to destroy the usefulness of the grid meter as a peak indicator.

Operating Conditions and Adjustment

Operating tubes in push-pull in a linear amplifier requires very good balance in the driving voltages applied to each side of the circuit. If the driving voltage is higher on one side than the other, the tube or tubes on that side will be driven to peak output before those on the other side, and will start saturating or "flattening" before the full output of the amplifier is realized. The condensers in the grid tank circuit, C_1 and C_2, should be matched in capacitance within a per cent or two, and the usual precautions as to maintaining circuit balance should be observed. The r.f. voltage balance can be checked with an r.f. probe and v.t. voltmeter. Another checking method would be to provide individual by-passes at the cold ends of R_1 and R_2, running out separate d.c. return leads to separate grid-current jacks to note whether or not the tubes on both sides of the circuit start taking grid current at the same time.

The oscilloscope patterns of Fig. 3 show the effect of grid bias on the linearity. The peak output was the same in both cases, but in the upper pattern the grid bias was set so that the total plate current of the four tubes was 110 ma. without signal (almost full plate dissipation) while the lower one was the best that could be obtained when the bias was near plate-current cut-off. The improvement in linearity resulting from operating at the lowest possible bias is striking, to say the least. The pronounced curvature in the bottom region of the characteristic is typical of over-biased tetrodes. In selecting the resting plate current by adjustment of the bias voltage it is advisable to make sure that no one tube is overloaded. This can occur even though the total input is less than 120 watts, since there is some variation in the plate currents taken by various tubes at the same bias voltage. Check the tubes individually with the other three out of their

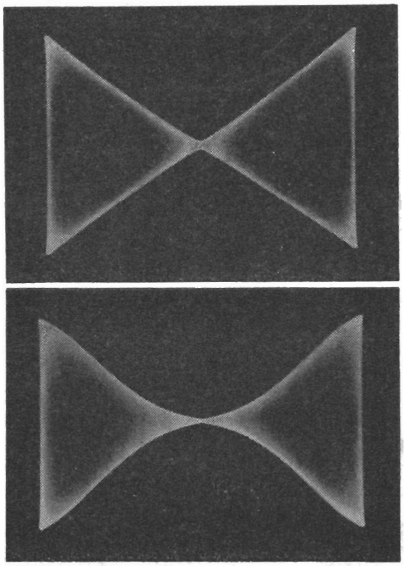


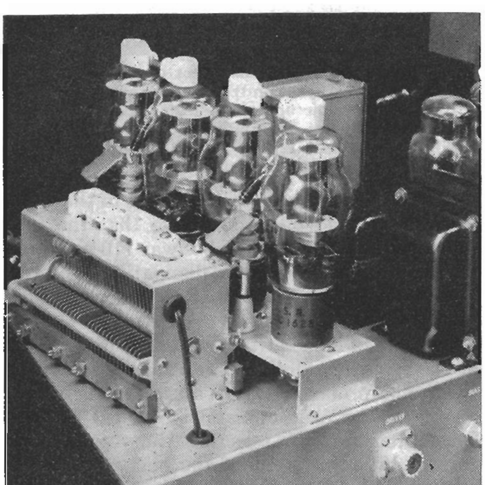
Fig. 3 — Showing the effect of grid bias on linearity. The peak output is the same in both cases. Upper picture, bias set for approximately full rated plate dissipation with no excitation; lower picture, bias near cut-off (approximately 10 ma. total plate current).

sockets and, if a selection is possible, choose four that take substantially the same plate current.

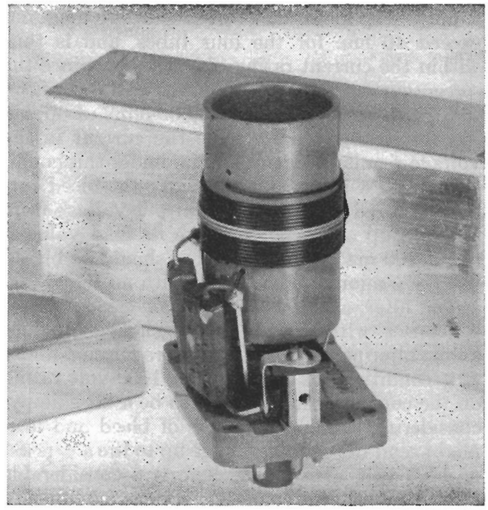
The preferable method of adjusting the amplifier tuning for optimum output and linearity is of course to use an oscilloscope with the two-tone test. If the audio oscillator generates a good sine wave and the distortion in the exciter itself is low, the optimum conditions should be secured with a plate current of 180 to 190 ma. when the driving voltage is just at the point where a trace (a few microamperes) of grid current shows. A fairly good job of adjustment can be done without the 'scope, provided the two-tone test can be used and there is independent assurance that the distortion in the exciter is low. Maintain the driving voltage just at the grid-current point and adjust the antenna coupling, keeping the plate circuit at resonance, for about 180 ma. plate current. The off-resonance plate current should be only 10 ma. or so larger than the "in-tune" current. Some sort of r.f. output indicator such as an antenna ammeter is helpful; the output should start to drop immediately on even a slight reduction in driving voltage. If the output tends to stay up when the driving voltage is cut slightly, the amplifier is saturating on the peaks and is not loaded heavily enough. The trick is to get the loading just right so that the maximum output is obtained (too-heavy loading will reduce both the output and plate efficiency) at exactly the point where a bit more drive will cause flattening.

Since there is ample plate-dissipation capacity for a.m. operation at the same peak output, the linearity also can be checked by the customary a.m. method if the exciter can furnish an a.m. signal. The trapezoidal pattern should be used, and a very simple 'scope such as is shown in the *Handbook* will suffice.

In voice operation using a resting plate current between 100 and 120 ma., the maximum plate current as registered by the plate meter on voice peaks is about 150 ma.



Close-up view of the plate circuit with the tank coil removed to show the blocking condensers, parallel-feed plate chokes and parasitic-suppressor coils. The double lead through the grommets runs from the output-circuit coil to the coupling condenser and coax connector underneath the chassis.



Construction of the plug-in grid tanks. The inductances of the two coils are adjusted for an input impedance of 75 ohms at the center of the band. Final pruning of the grid coil can be by adjusting the spacing of an end turn as in this 7-Mc. assembly. The coil form is mounted on a strip of thin insulating material which is mounted on the studs at the sides of the plug-in base.

Additional Notes

Fig. 1 shows that each 1625 screen has its own by-pass condenser which, so far as the operating frequency is concerned, simply means that the entire screen circuit is by-passed with a capacitance four times as large since the impedance of the connecting leads is very small. The separate condensers were used in anticipation of a possible need for installing decoupling circuits in case of parasitic oscillations involving the screen circuits. Since the small coils in the plate leads settled the parasitic question very nicely, no such decoupling was needed. The condensers may be useful, however, for good suppression of v.h.f. harmonics.

Although the constructional practice of shielded wiring with disk by-passes was followed as a matter of course, the amplifier was not shielded for TVI. Shielding is not necessary for 75 meters, but is likely to be required for 14-Mc. — and perhaps 7-Mc. — operation in localities where a harmonic falls directly in a channel having a weak TV signal. Class AB_1 operation does help — it is only necessary to look at the TV screen while the driving voltage is nudged into the grid-current region to see that — but it is not a complete panacea for the tough cases.

The amplifier should be neutralized by the usual method of adjusting for minimum r.f. in the plate circuit with r.f. voltage on the grids but with plate and screen voltages off. A sensitive indicator such as a crystal detector and low-range milliammeter should be used; they may be connected to the r.f. output terminals for convenience. C_3 and C_4 are adjusted by bending the metal tabs from which they are constructed, to vary the spacing.

» The 811-A is probably the most popular triode for linear-amplifier use; its zero-bias characteristics simplify grid-circuit regulation problems by maintaining a fairly constant load on the driver. Here are two in push-pull, capable of peak power outputs up to 400 watts. Since a Class A driver is included, practically no power is taken from the s.s.b. exciter.

A Two-Stage Linear

BYRON GOODMAN, W1DX

PROBABLY any tubes can be made to operate satisfactorily as Class B r.f. amplifiers, but the driver problem is greatly reduced if zero-bias tubes can be used. When negative-bias tubes are used, such as any of the tetrodes and most of the triodes, the load on the driver changes as the signal swings in and out of the grid-current region, and the driver is hard put to deliver an undistorted signal. With zero-bias tubes, the driven tubes draw grid current under any and all signal conditions, and the load on the driver is more nearly constant. When using negative-bias tubes, "swamping resistors" are used across the grid tank to offer a more nearly constant load to the driver. These resistors waste driver power, however, and it is nice to be able to avoid using them.

The 811-A is our choice for a Class B stage, since it can be used with from 1000 to 1200 volts on the plate with zero grid bias, and a pair will deliver from 250 to 300 watts peak in this plate-voltage range.

From *QST*, March, 1951.

The driver stage is logically something running Class A or AB_1, to minimize its drive requirements and offer maximum power sensitivity. It is hard to find a tube better suited for the purpose than the 807, since it is inexpensive and will deliver the required power.

The Circuit

The circuit of an amplifier like this doesn't look much different than the usual Class C amplifier arrangement, since the only real differences are in the operating conditions (bias and loading). It can be seen in the wiring diagram, Fig. 1, that the 807 stage looks like any other, except that it uses

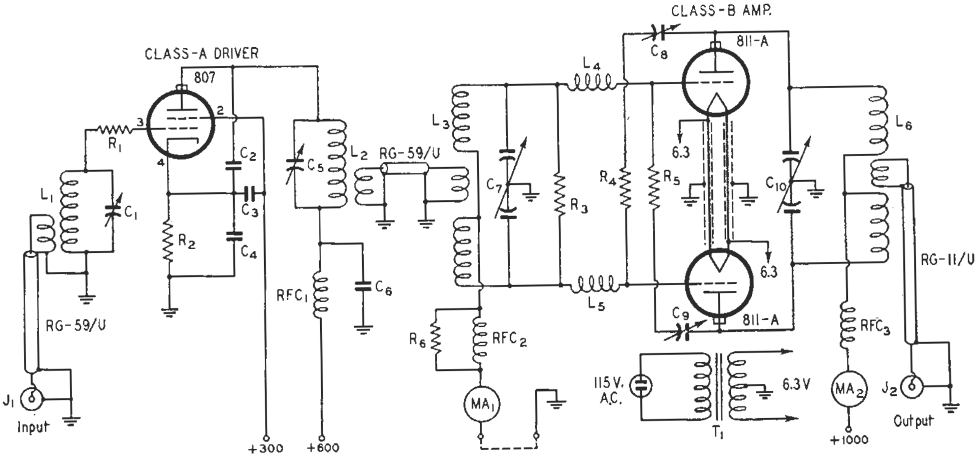


Fig. 1 — Wiring diagram of the linear amplifier.

C_1 — 140-μμf. variable (Millen 19140).
C_2 — 13-μμf. tubular, made of RG-58/U. Active length, 6 inches.
C_3, C_4 — 0.005-μf. disk ceramic.
C_5 — 140-μμf. variable (Millen 22140).
C_6 — 0.001-μf. 1200-volt mica.
C_7 — Dual variable, 100 μμf. per section (Millen 24100).
C_8, C_9 — Disk-type neutralizing condensers with feed-through base (Bud NC-853).
C_{10} — Dual variable, 200 μμf. per section, 0.077-inch spacing (National MC-200D).
All resistors are composition, not wire-wound.
R_1 — 100 ohms, ½ watt.
R_2 — 680 ohms, 2 watts.
R_3 — 2700 ohms, 4 watts (four 2700-ohm in series-parallel).
R_4, R_5 — 20 ohms, 2 watts.
R_6 — 1000 ohms, 1 watt.
J_1 — Input connector (Jones S-101-D).
J_2 — Coaxial-line connector (Amphenol 83-1R)
L_4, L_5 — 9 turns No. 12 enam., 1½-inch diam., 1¼ inches long.
MA_1 — 0–50 milliammeter.
MA_2 — 0–500 milliammeter.
RFC_1 — 2.5-mh. 300-ma. r.f. choke.
RFC_2 — 200-μh. 75-ma. r.f. choke.
RFC_3 — 5-mh. 300-ma. r.f. choke.
T_1 — 6.3-volt 10-amp. transformer (Stancor P-6308).

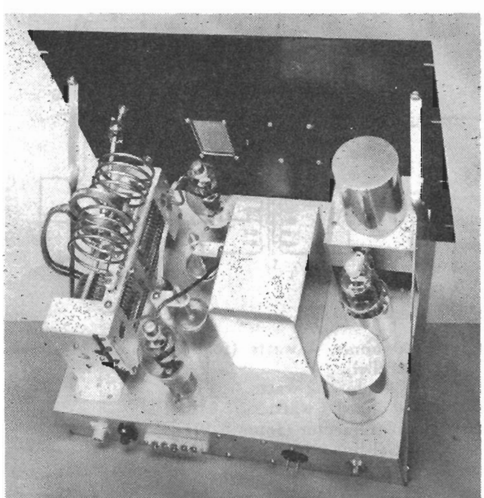
cathode bias. A small resistor, R_1, in series with the control grid, and a coaxial plate-to-cathode condenser, C_2, are included to eliminate the tendency to oscillate at various and assorted frequencies, as can be expected of a tetrode running Class A. The 807 is link-coupled to the grid circuit of the 811-As, to confine the ground returns to their respective stages and to provide a simple means for adjusting the coupling. Since the over-all gain of a two-stage amplifier like this is quite high, it is advisable to return the ground leads within a stage to a single point, to reduce the chances for over-all feed-back.

In the 811-A stage, the chokes L_4 and L_5 were required to detune a v.h.f. parasitic that showed up, and R_6 knocked out a low-frequency one. The two resistors, R_4 and R_5, in the neutralizing circuit may seem a little unusual, but they were found necessary to kill a v.h.f. parasitic. Loading the grid circuit slightly with R_3 killed the last traces of instability. If all of these suppression devices make it sound like the amplifier is a hot-bed of oscillations, just remember that this stage has considerably more gain than the usual Class C amplifier. Several of the parasitic oscillations could be killed easily by using grid-leak bias, but that isn't the way you operate a linear amplifier.

No filament by-pass condensers were used because they weren't found to be necessary. The filament wiring was done with shielded wire, however, and this adds a little filament-to-ground capacity. All of the non-r.f. leads were made with shielded wire, as has become the custom of many of us. Although it is probably not necessary in many cases, it seemed like a good idea here because of the possibilities for over-all feed-back.

Construction

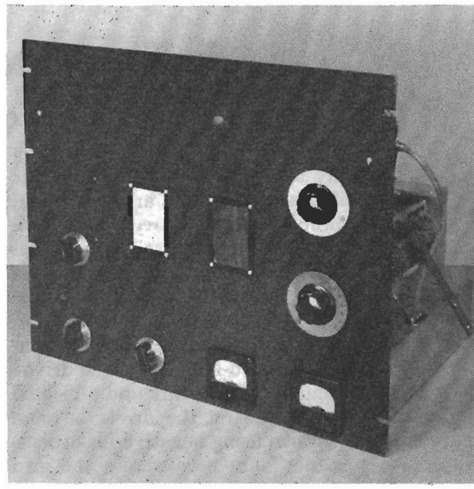
The amplifier is built on a 13 × 17 × 3-inch aluminum chassis. A 15¾-inch-high aluminum



A rear view of the linear amplifier, showing the push-pull 811-A output amplifier and the 807 driver. The cover of the rectangular shield can slides off for access to the final grid coil. The round shield cans are for the 807 grid and plate coils.

relay-rack panel is fastened to the chassis by the meters and two shaft bushings, and it is further reinforced by a pair of brass straps.

The grid coil for the 807 plugs into a four-pin socket mounted at the rear of the chassis and is shielded by an ICA No. 1549 3-inch diameter shield can. The plate coil plugs into a five-pin



A two-stage linear amplifier for boosting the power level of a s.s.b. signal. The large knobs control the antenna coupling and output plate tuning. Meters indicate grid and plate currents of the output stage.

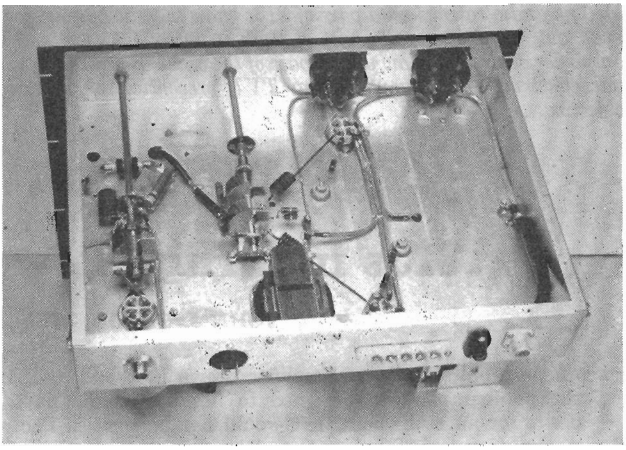
socket mounted 4 inches above the chassis. The platform for the socket also shields the plate condenser, C_5. Another 3-inch diameter shield can protects the 807 plate coil. The plate by-pass condenser, C_6, is mounted under the chassis near the 807 socket, and the "cold" lead from C_5 and L_2 is brought down to it in shielded wire. The coaxial condenser, C_2, is made of a length of RG-58/U and drops down through the same chassis hole as does the shielded "cold" lead.

The grid coil for the 811-As is shielded by an ICA No. 29842 4 × 5 × 6-inch aluminum utility cabinet. To simplify coil changing, the cabinet is fastened to the chassis and a friction-fit cover is made from a piece of sheet aluminum. The inside lips on the top of the cabinet should be bent down to allow more room for the hand that changes coils.

The output tank condenser, C_{10}, is mounted on the chassis with aluminum brackets that also support the jack bar for the output coil, L_6. The variable link mounts on the jack bar and is connected to the panel control through two flexible couplings and an extension shaft. A B & W 3-turn shielded link was used for the output link, but an ordinary link might serve just as well in cases where TVI precautions are unnecessary.

Adjustment

During the initial testing it is advisable to connect a milliammeter in the plate circuit of the 807, to check the static current of the tube. With 600 volts on the plate and 300 on the screen, the plate current will run around 40 ma. If it differs



Underneath the chassis, showing all but r.f. leads in shield braid. The coils in the leads from the split-stator grid condenser are parasitic chokes.

from this value, it should be brought back by changing the cathode resistor, R_2, or the screen voltage. If the available plate voltage is something other than 600, adjust the plate current for a static plate input of about 25 watts. With no input signal the screen current will be insignificant, and with maximum signal it will be only a few ma. The screen supply therefore has no great drain on it, but it should be "stiff" and not vary in output voltage. A low-level exciter stage is probably the best source for this voltage. Since the 600-volt plate supply will have practically a constant drain on it, the regulation of this supply is not important.

Couple a signal from the exciter in at J_1 and, with plate and screen voltage on the 807 but no plate voltage on the 811-As, resonate the circuits for maximum grid current in the 811-As. It should be easy to run this over 50 ma. without any change in the 807 plate current. Cut the excitation back to where the 811-A grid current is about 25 ma. and neutralize the output stage. You can use the "flick" in the grid current as C_{10} is tuned through resonance, but a more sensitive indication can be obtained by using a crystal diode and 0–1 milliammeter connected to J_2. You may find an irreducible minimum feed-through current with a sensitive indicating device, but this is the neutralization point.

Couple a dummy load to J_2 and apply plate voltage to the 811-As. Couple an oscilloscope to the dummy load and apply a "two-tone" test signal at the input. The 811-A no-signal plate current should run around 40 or 50 ma., depending upon the plate voltage. Adjust the two-tone signal amplitude for 10 or 15 ma. grid current and resonate all circuits. Then increase the excitation until the two-tone pattern just begins to flatten on the peaks. When using 1000 volts on the plates of the 811-As, this flattening should not occur until MA_2 indicates 160 ma. or so — with 1200 volts the current should run up to 190 ma. without noticeable flattening. If distortion occurs sooner, it indicates that the 811-A stage should be coupled more tightly to the dummy load, or that the 807 stage is not delivering enough drive. It will probably be found that the 811-A output coupling is at fault — if the link at L_3 is coupled closely the 807 should behave at all power levels. The 811-A grid current will be around 25 or 30 ma. when the pattern flattens.

When the linearity has been checked, turn off the power long enough to couple the antenna to the amplifier instead of the dummy load, and then couple the antenna to give the same plate current with the same excitation condition as before. You are then in business.

If an exciter having a peak output of about 10 watts is available, it may be used to drive the 811-As directly. Simply disregard the 807 driver stage and connect the output coil of the exciter to the RG-59/U link, replacing the link coil coupled to L_2.

COIL TABLE FOR TWO-STAGE LINEAR AMPLIFIER

Band	*Turns*	*Wire No.*	*Diam.*	*Length*	*µh.*	*Link*	*Spacing*
L_1*							
3.9	22½	20 enam.	1	¾	10	4	1/16
14	10½	20 enam.	1	¾	2.5	3	1/16
L_2**							
3.9	25	20 enam.	1	⅞	11.2	4	1/16
14	11	20 enam.	1	¾	2.5	3	⅛
L_3***							
3.9	22	22 enam.	1¼	1¼	9.4	6	Adj.
14	12	18 enam.	1¼	1⅛	3.3	4	Adj.
L_4****							
3.9	22	16 enam.	2½	2¼	20	3	Adj.
14	8	0.15 tubing	2½	3¾	2.3	3	Adj.

*Wound on Millen 45004 plug-in form.
**Wound on Millen 45005 plug-in form.
***National AR-16-40S and AR-16-20S. 75-meter coil shunted by 150-$\mu\mu$f. mica condenser.
**** B & W 8TVL0 with 18 turns removed, and B & W 15TVL.

» The audio ratings on a pair of 813s show them to be good for a peak output of 650 watts with a peak input of 900 watts. The amplifier described below uses a pair of the tubes in a rather unusual constructional layout having TVI prevention as a major objective.

813s in a High-Power Linear

JOHN J. SIMON, W5SCE

The push-pull 813 linear amplifier shown in the photographs was designed to fulfill a desire for an s.s.b. final that would handle the maximum permissible input, and be driven by a Central Electronics 10A exciter or its equivalent.

The circuit, shown in Fig. 1, is quite conventional, except for the use of the National MB-40SL multiband tuner in the grid circuit. The chokes L_3 and L_4, wound directly in the plate leads, in conjunction with L_1 and L_2 in the grid leads, eliminated all traces of v.h.f. parasitic oscillation. The combination of RFC_1 and the 1000-ohm resistor in the bias lead similarly takes care of any low-frequency parasitic. The amplifier is neutralized by C_1 and C_2 whose construction is described later. It is believed that the individual filament transformers help in maintaining circuit balance.

From *QST*, July, 1954.

Construction

The amplifier is built on a 17 × 13 × 4-inch aluminum chassis. The tank condenser is fastened directly to the chassis, along the rear. The jack bar for the tank coil, elevated on ceramic pillars, is placed immediately in front of the tank condenser. The tubes are mounted symmetrically in

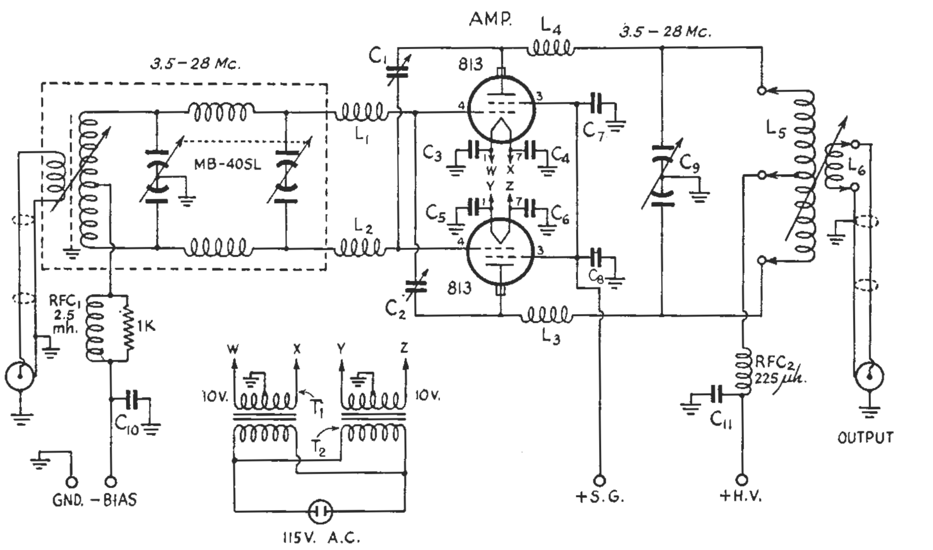
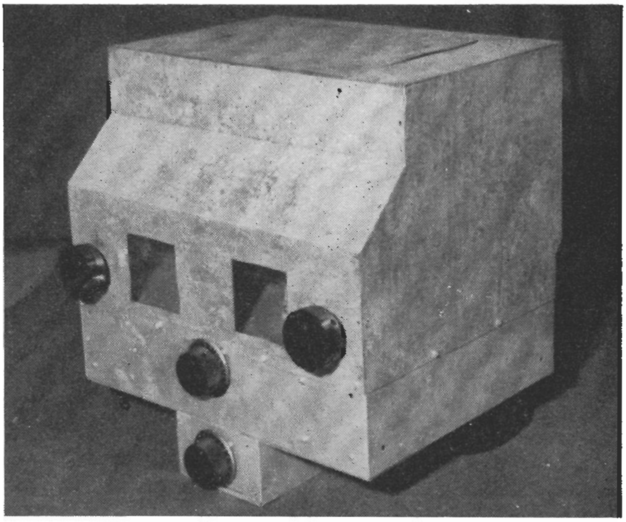


Fig. 1 — Circuit of the high-power push-pull linear amplifier.

C_1, C_2 — Neutralizing condenser — see text.
C_3, C_4, C_5, C_6 — 0.004 μf. disk ceramic.
C_7, C_8 — 0.001-μf., 1.5 kv., disk ceramic.
C_9 — Dual-section variable, 200-μμf.-per-section, 0.125-inch spacing (Johnson 200DD45).
C_{10} — 0.001 μf., disk ceramic.
C_{11} — 500-μμf., 3 kv., disk ceramic.
L_1, L_2 — 11 turns No. 20 d.c.c., ⅜-inch i.d., close-wound.
L_3, L_4 — 11 turns No. 14 bare, ⅜-inch i.d., turns close but not shorting (see text).
L_5 — 3.5 Mc. — 28 μh. — each half 13 turns No. 10 (Johnson 1000LC80).
— 7 Mc. — 15 μh. — each half 8 turns No. 8 (Johnson 1000LCS40).
— 14 Mc. — 5 μh. — each half 5 turns ¼-inch tubing (Johnson 1000LCS20).
— 21 Mc. — 3 μh. — each half 4 turns ¼-inch tubing (Johnson 1000H/LCS14).
— 28 Mc. — 1.8 μh. — each half 3 turns ¼-inch tubing (Johnson 1000H/LCS10).
All above 3½-inch diam., 2½ inches long, each half, 1-inch space at center.
L_6 — 5-turn variable link.
MB-40SL — National multiband tuner.
RFC_2 — National R-175A.
T_1, T_2 — Merit P3145 (5-amp.) or P3146 (10-amp.). (See text.)



The high-power linear amplifier with all shielding in place. The tubes may be seen through the screened cut-out in front. The unusual shape leaves room to mount meters near the top of the 17½-inch rack panel to which the unit will be attached.

front of the tank coil, and their sockets are sub-mounted to bring the internal shields level with the chassis. Both the tank-condenser and the variable-link shafts are driven from the panel controls by means of right-angle gear drives.

The multiband tuner, fitted with National type AM vernier dials and used in the grid circuit, is mounted centrally between the two tube sockets underneath the chassis. Individual filament transformers for each tube are mounted in the two forward corners. The plate r.f. choke and by-pass are toward the bottom and to the left in the bottom-view photograph. It was found necessary to enclose these two components with a partition shield, to remove the last trace of instability at the operating frequency. It was also necessary to shield the high-voltage lead with copper braid, from where it leaves the partition shield to the point where it goes up through the chassis to the tank-coil center tap.

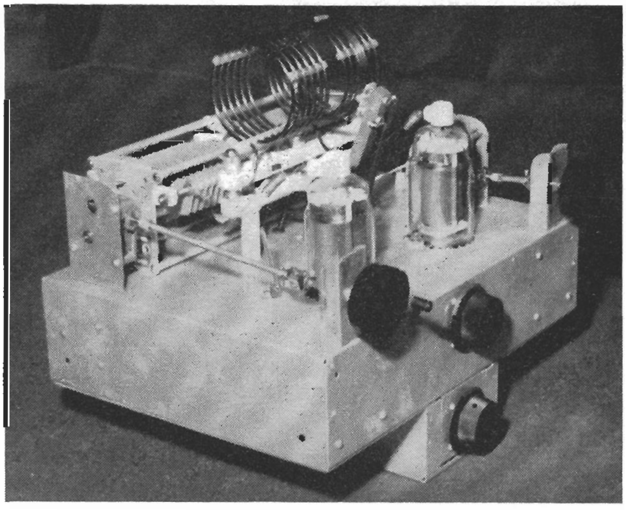
Power terminals and a coax connector for r.f. input are along the rear edge of the chassis.

Since the multiband tuner protrudes below the chassis limits, a cut-out was made in the bottom plate, and a 5 × 10 × 3-inch aluminum chassis serves as a box cover over the opening. A ventilating blower, mounted alongside the cover, with its motor inside the chassis, exhausts into the box, and thence into the chassis proper.

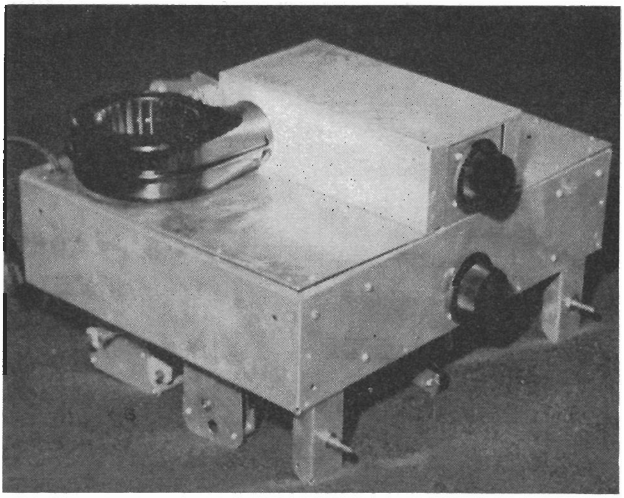
A shielding enclosure is not necessary for amplifier stability but, of course, is desirable in the consideration of TVI. The unusual shape, evident from one of the photographs, is a result of the discovery that cabling to meters mounted along the upper portion of the panel, to which the amplifier was eventually attached, was causing unbalance in the r.f. circuit. The indentation at the top of the enclosure still allows the original meter mounting, but isolates the meters and cabling from the r.f. circuit.

The enclosure shown in the photograph is made of galvanized sheet iron. It has a removable top cover to permit changing plate coils, and screened openings in front to allow a peek at the tubes while they are operating.

Originally, capacitive neutralization was used,



A high-power push-pull linear amplifier using 813s. The plate tank condenser and the variable link are operated through right-angle drives. The boxed-in section below contains a multiband tuner for the grid circuit. The feed-through insulator close to the tube to the right serves as a neutralizing condenser. A similar one is installed alongside the second tube.



A ventilating blower mounted underneath the chassis exhausts into the chassis through the box shield covering the multiband tuner.

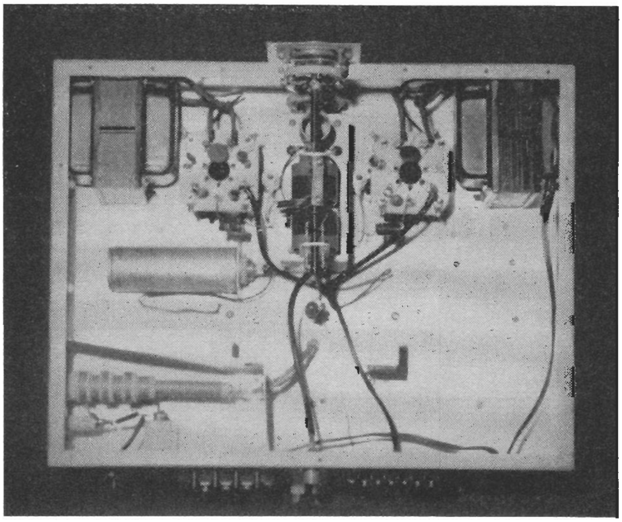
as shown in the diagram. The neutralizing condensers consisted of a pair of ceramic feed-through insulators, mounted as close to the tubes as possible. Adjustment consisted of running the second nut up or down on the threaded rod until neutralization was obtained. When neutralization was complete, the nuts were cemented in place.

Recently, however, link neutralization was tried with very satisfactory results. A 2-turn link was wound symmetrically on the center of the low-frequency coil of the multiband tuner, and 75-ohm ribbon run up through a hole in the chassis, central and behind the plate tuning condenser. This terminates in a 2-turn self-supporting link, 1 inch in diameter, wound with No. 12 wire, mounted on a ceramic stand-off. The stand-off, in turn, is mounted on a bracket fastened to the center brace of the tuning condenser. In general, this system is easier to adjust. Drive requirements are increased slightly, but linearity is just as good.

Attention to some details was necessary to obtain balanced operation of the amplifier. The 75-ohm 1-kw. Twin-Lead from the swinging link to the terminals at the rear should be brought under the tank-coil mounting, and out between the two sections of the tank condenser. Other routings tried caused unbalance. The capacitance between the plate of the right-hand tube and the mechanism driving the variable link was sufficient to unbalance the amplifier. This was compensated by mounting a metal tab near the left-hand tube, and bending it closer to or farther away from the tube until the circuit balanced.

Power Supplies

Since little or no grid current is drawn, almost any good bias supply will suffice. VR tubes or batteries will be satisfactory. The only requirements are that the source be good d.c., and that it be constant. The author uses an electronically-regulated supply for the purpose of facilitating experiments with various operating points and classes of operation.



Bottom view of the high-power linear amplifier, showing the tube sockets, filament transformers, and the grid-circuit multiband tuner. The plate r.f. choke and by-pass are enclosed in the partition shield in the lower left-hand corner. The electrolytic near the center was used in this instance to augment the filter capacitance in the author's screen supply.

When properly loaded, the maximum screen current drawn will not exceed 20 ma. Therefore, VR tubes are ideal for stabilizing screen voltage. A good steady screen voltage improves linearity and makes the tubes easier to drive. Screen voltage can be taken from the plate supply, if the latter has good regulation, by using VR tubes to stabilize the voltage dropped. The plate supply should also have as good regulation as possible. Use choke input, the lowest value of bleeder resistance the supply will tolerate, and as much output capacitance as possible. The proper idling current for this amplifier is 50 to 100 ma. This idling current should be considered part of the bleeder current. In this connection, it is strongly recommended that the amplifier be allowed to idle, and that cut-off bias not be used between transmissions.

Under normal s.s.b. operation with 2500 volts on the plates, 750 volts on the screens, and a biasing voltage of 80, the grid current should just barely flicker with modulation, the screen current should rise from about 1 ma. to 20 ma., and the plate current increase from an idling value of 50–100 ma. to 400 ma. on peaks. The amplifier has been operated with excellent linearity with the following range of applied voltages:

Plate	*Screen*	*Bias(–)*	*Pwr. Input*
3000	750	90	1000
2500	750	80	1000
2000	600	65	600
1500	600	60	450

At 2000 and 1500 volts, it is necessary to operate in the region of grid current, and some swamping of the grid circuit is desirable. In this connection, it should be recognized that the 10A exciter has some swamping built into it. Since the amplifier as operated by the author just barely invades the grid-current region, no swamping is included in the amplifier.

This amplifier is completely stable, free from parasitics, and has excellent linearity. Operating on 75, 40 and 20 there is absolutely no TVI on Channels 6 and 10 in a fringe area, and only very weak BCI has been noticed. It should be mentioned, however, that presumably because of the high Q of the multiband tuner on 20, only sufficient power is available using the 10A to drive the amplifier to 600 watts. No operation has been attempted on 10 or 15, due to the lack of suitable injection frequency for the exciter.

F.C.C. REGULATIONS GOVERNING CALLING AND SIGNING PROCEDURES

§ 12.82. *Transmissions of call signs.* (a)

(1) The operator of an amateur station shall transmit the call sign of the station or stations (or may transmit the generally accepted identification of the network) being called or communicated with, or shall identify appropriately any other purpose of a transmission, followed by the authorized call sign of the station transmitting:

(i) at the beginning and end of each single transmission or;

(ii) at the beginning and end of a series of transmissions between stations having established communications, each transmission of which is of less than three minutes' duration (the identification at the end of such a series may be omitted when the duration of the entire series is less than three minutes), and;

(iii) at least once every ten minutes or as soon thereafter as possible during a series of transmissions between stations having established communications, and;

(iv) at least once every ten minutes during any single transmission of more than ten minutes' duration.

(2) The required identification shall be transmitted on the frequency or frequencies being employed at the time. . . .

(b) . . . When telephony is used, the call sign of the station shall be preceded by the words "this is" or the word "from."

(d) When using telephony, phonetic aids to identify the call sign of the station may be employed.

GROUNDED-GRID LINEARS

With the current interest in grounded-grid linear amplifiers, you will probably be interested to see how Bill Felch, W2EAS, has his rigged up. Bill uses a pair of 805s, as shown in Fig. 1A, and the heart of the set-up is the low-capacity filament transformer, T_1, that makes it easier to keep the filaments above r.f. ground. The transformer is one that was kicking around in surplus, about the most foolproof amplifier he ever tried. The 240 volts bias is obtained from a VR-150 and a VR-90 connected in series, and this holds the idling current to around 100 ma. On peaks, the indicated plate current is approximately 300 ma. Parasitic suppressors were originally included in the plate leads but they were found to be unnecessary.

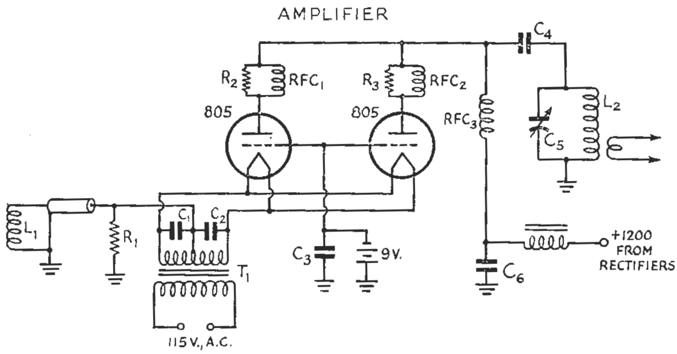
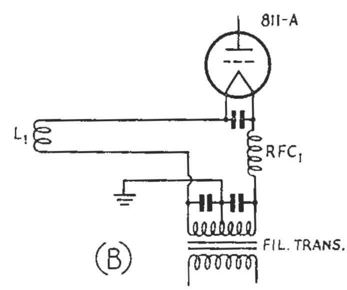


Fig. 1 — (A) The grounded-grid linear amplifier at W2EAS uses a pair of 805s in parallel. A special (surplus) low-capacity filament transformer solves the problem of keeping the filaments above ground for r.f.



C_1, C_2, C_3 — 0.01-μf. 600-volt mica.
C_4 — 0.004-μf. 2500-volt mica.
C_5 — 360-$\mu\mu$f. variable, 0.05-inch spacing (Cardwell MO-180-BD, both halves in parallel).
C_6 — 15-μf. 1500-volt filter condenser.
R_1 — 800 ohms.
R_2, R_3 — 100 ohms, 1 watt.
L_1 — 4½-turn link on ARC-5 driver.
RFC_1, RFC_2 — 30 turns No. 30 enam., wound on R_2, R_3.
T_1 — Low-capacity filament transformer, surplus.

(B) — W4PIX uses the excitation link, L_1, and an r.f. choke, RFC_1, to keep the filament above ground in his grounded-grid amplifier.

with about ½-inch spacing between windings and core. Originally with a 5-volt secondary, it was rewound for 10 volts to keep the 805s happy.

The resistor R_1 has no particular significance except to limit the d.c. voltage appearing at the end of the cable if the link is disconnected. The 4½-turn link, L_1, is part of a converted BC-696 driver. Peak plate input to the driver runs around 75 watts. Idling current to the 805s is around 85 ma., kicking to 350 to 400 ma. on peaks. The linear worked right off, after a minor oscillation was killed by the parasitic suppressors in the plate leads, and no neutralizing is required, of course.

Bill also passes along the sketch of Fig. 1B, which is the way W4PIX runs his grounded-grid 811-A linear final. The method takes a higher-voltage filament transformer, to make up for the drop in the link and the r.f. choke.

Push-Pull 304-TLs

Ed Brown, W9ROQ, has a pair of 304-TLs in his 14-Mc. output amplifier driven by push-pull 811-As. The 304s are in the grounded-grid circuit shown in Fig. 2, and Brownie says it is

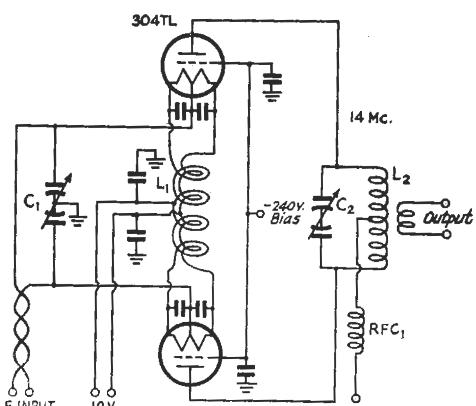


Fig. 2 — The 14-Mc. grounded-grid linear at W9ROQ uses a pair of 304-TLs.

C_1 — 100-$\mu\mu$f.-per-section variable.
C_2 — 50-$\mu\mu$f.-per-section variable.
All other condensers are 0.004-μf. 600-volt mica.
L_1 — 20 double turns No. 12 enam., wound on 1-inch diam. form.
L_2 — B & W HDVL-20
RFC_1 — R.f. choke (National R-175).

SINGLE SIDEBAND FOR

THE VIKING II AS A LINEAR AMPLIFIER

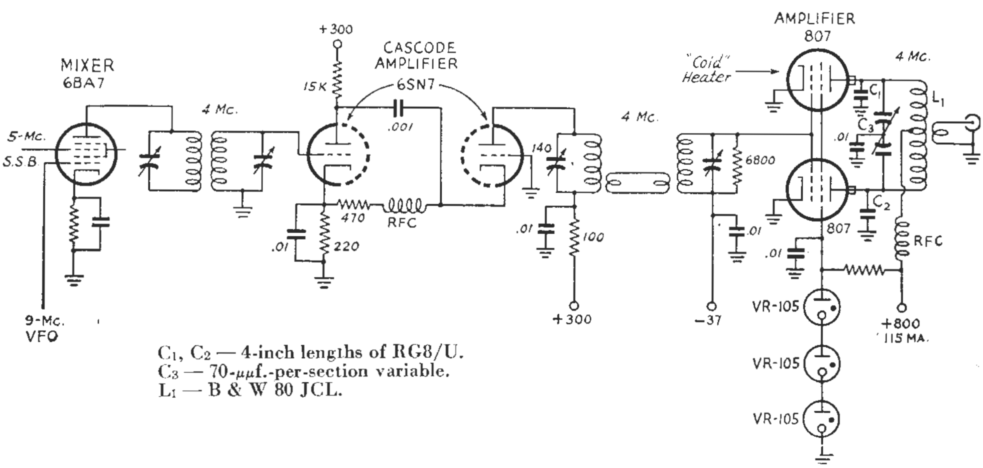
The Viking II can be used as a linear amplifier on both 75 and 40 meters without modification. W4JMU was the first to try it — he has a home-made phasing rig with 6AG7 output. The s.s.b. signal is introduced at the "VFO" socket, and the Viking is switched to "C.W." and "VFO." The stages can be tuned by injecting carrier from the s.s.b. exciter. W4JMU then talks into the mike and backs off the "Drive" control of the Viking until no grid current is indicated, even on voice peaks. The output loading is adjusted so that the voice peaks kick the indicated plate current up to about 240 ma. Naturally the best checks on loading can only be made and confirmed with the help of a 'scope, if one is looking for maximum undistorted output.

Both W4JMU and WØGPT (who drives his Viking II with a Central Electronics 10A exciter) report that the oscillator current in the Viking runs high if nothing else is done, but it is a simple matter to plug in a 500-ohm cathode resistor at the key jack and bring the current down to around 8 or 9 ma.

All this doesn't mean that linear amplifiers are no longer needed for s.s.b.! What it does mean is that W4JMU found that the biases and frequency sequence in the Viking II permit *its* use as a linear. You can't use a rig where frequency multiplication is involved, and you can't use a rig (without modification) in which some of the amplifier tubes are biased to cut-off or beyond or, on the opposite end of the scale, would run into grid current. The tubes in the Viking are running Class A or AB_1 under the above conditions.

CASCODE DRIVER STAGE

For a number of years the "cascode" circuit has been the exclusive property of the v.h.f. men and the TV set manufacturers, but such is no longer the case.



C_1, C_2 — 4-inch lengths of RG8/U.
C_3 — 70-μμf.-per-section variable.
L_1 — B & W 80 JCL.

The cascode circuit is used as a driver by W4AWS for his neutralized 807 output stage. A "cold" 807 is used for neutralizing and for circuit balance.

its signal from a modified SSB Jr. exciter and a 9-Mc. VFO. The output from the cascode is link-coupled to an 807 amplifier that is a little unusual in that it uses an extra 807 for neutralizing and balance. With 6800 ohms swamping in the 807 grid circuit, a peak grid current of about 2 ma. shows plate-meter peaks of about 115 ma. Battery control-grid bias is used, and the screen is held at 315 volts by three VR-105s in series.

S.s.b. exponent Art Hale, W4AWS, finds that it makes an excellent replacement for a 6AG7 amplifier, if you have had trouble taming one of those fiery pentodes. As used at W4AWS, the 6SN7 cascode follows a 6BA7 converter that gets

» To get from one band to another frequency conversion or mixing is the common method. "High-level" mixing is often a convenience, making it possible to get on a new band without digging into the exciter and offering a way to use the several watts output available from most existing exciters.

High-Level Converters

There frequently is need for a good high-level converter to aid in bandchanging at powers above what the small receiving-type mixers will handle, and Norman Roller, W6EDD, has a neat solution. Shown in Fig. 1, it can be used as a Class A grounded-grid amplifier or as a mixer, depending upon how the plate circuit is tuned. W6EDD has used a 6Y6, 6V6, 6L6 and 6AQ5 in the circuit, all with equally good results. As an amplifier, the plate current runs around 15 ma. (at 300 volts), with no variation with signal. As a mixer, the plate current idles at around 15 ma. and kicks to 30 ma. on peaks. It requires less than 1 watt of drive, and its output is sufficient to overload an 807.

The heterodyning signal has no effect on straight-through operation; W6EDD uses the

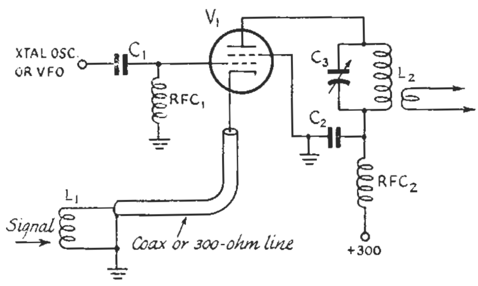


Fig. 1 — This simple high-level mixer/amplifier has been used by W6EDD and others. The heterodyning signal is present at the grid in either amplifier or mixer service — the tuning of the output circuit, L_2C_3, determines the function.

C_1 — 100 $\mu\mu f$.
C_2 — 0.001 μf.
RFC_1 — 2.5-mh. r.f. choke.

See text for suitable tube types.

device straight-through on 75, and a 10.4-Mc. oscillator and a new coil at L_2 put the output on 20 meters.

Suitable tubes other than those already mentioned include the 807, 2E26 and 6AR6.

75- and 40-Meter S.S.B. Operation

To simplify two-band operation with his 450-kc. crystal-filter exciter, Ralph Porazzo, WØLLW, uses the double-conversion system shown in Fig. 2. A block diagram is shown at A, illustrating the dual injection of the VFO. The output of the high-level mixer is tuned to the desired band.

The circuit of the high-level mixer is shown at B, and is self-explanatory. It is essentially the same as the W6EDD circuit described earlier. With a system like this, if the lower sideband is obtained on 75 mete s it will also be available on 40 meters.

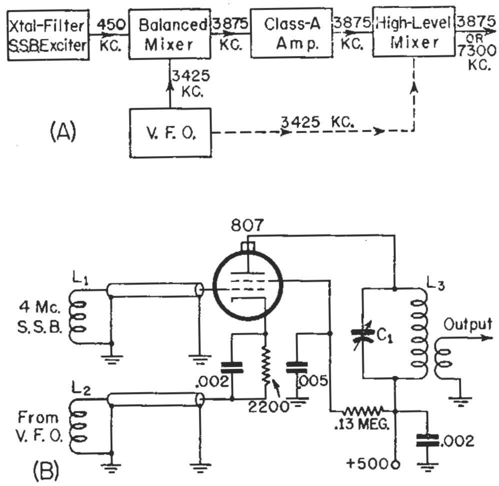


Fig. 2 — WØLLW simplified bandchanging between 75 and 40 with his crystal-filter s.s.b. rig by using the same VFO in two different parts of the circuit, as shown at A. The high-level mixer circuit is shown at B. C_1 is large and tunes to either 75 or 40 with the same L_3. L_1 and L_2 are 3- or 4-turn links.

BC-457 as a Mixer

The following description of a BC-457 conversion to a mixer is due to Dr. Milton Schwalbe, W4VP. Any low-level 4-Mc. single-sideband signal can be fed in, and the resultant 7-Mc. output can be fed to the antenna or used to drive a linear amplifier. The same principle is applicable to other bands by working out appropriate crystal frequencies and coil-condenser combinations. When crystal and exciter frequencies are mixed to obtain output at their sum frequency, the original sideband appears in the mixer output. However, when the mixer output is the difference of the input frequencies, the sideband is inverted. This may be especially useful when the exciter has no provision for switching sidebands, as in some of the crystal-filter jobs.

As shown in Fig. 3, the output of a 12A6 crystal oscillator on 3.3 Mc. is fed to the control

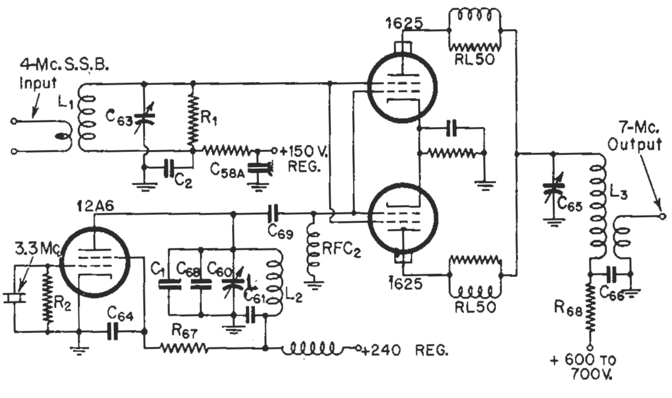


Fig. 3 — The BC-457 as W4VP's 40-meter mixer.

New Parts:

C_1 — 220 $\mu\mu$f.
C_2 — 0.01 μf.
R_1 — 4700 ohms.
R_2 — 0.1 megohm.
R_3 — 300 ohms, 10 watts (1625s cathode).
RFC_1, RFC_2 — 2.5 mh.
L_1 — 20 turns No. 22 enam. on 1⅛-inch diam. plug-in form. Link is 5 turns at cold end.

BC-457 Components:

C_{58A-B} — 0.05–0.05 μf.
C_{60} — Oscillator padding.
C_{61} — 0.006 μf.
C_{63} — Oscillator tuning.
C_{64} — 0.002 μf.
C_{65} — Amplifier tuning.
C_{66} — 0.01 μf.
C_{68} — 3 $\mu\mu$f.
C_{69} — 50 $\mu\mu$f.
R_{67} — 51,000 ohms.
R_{68} — 20 ohms.
R_{78} — 51 ohms (1625s screen decoupling).
RL_{50} — Parasitic suppressor.
L_2 — Oscillator coil T_{53}. Use "A_6" winding as is, after removing surplus leads.
L_3 — Amplifier coil T_{54}. Trim down to 10 turns evenly spaced. Add 3-turn link at cold end.

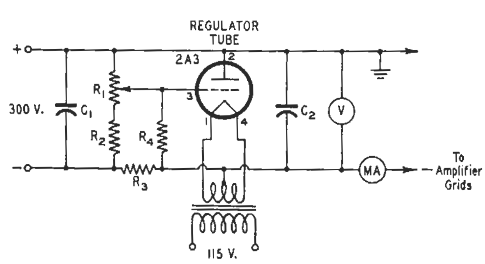
grids of paralleled 1625s. The output of the 4-Mc. s.s.b. exciter is injected at the screen grids of the mixer. The plate tank circuit of the mixer is tuned to the sum frequency, 7.3 Mc.

The three sockets at the rear of the BC-457 chassis will accommodate the crystal, a 12A6 crystal oscillator, and the 4-Mc. plug-in coil, L_1. Remove the flexible shaft under the chassis that couples the front and rear variable condensers. The front condenser is used to tune the plate circuit of the 1625s. The rear condenser (C_{63}) will tune the plug-in coil to resonate the screen-grid circuit to the 4-Mc. excitation frequency. The padding condenser across the plate tuning condenser of the 1625s is disconnected and not used. The former grid-circuit air-spaced padder (C_{60} in the shield above the chassis) and the former grid coil will serve as the output circuit of the 3.3-Mc. crystal oscillator. Shaft couplings to adjust the various condensers can be brought out without encountering any serious mechanical difficulties.

With power supplied as indicated in the schematic, tune C_{60} and check with receiver or absorption wavemeter for crystal-oscillator output on 3.3 Mc. Detune C_{60} on the high-frequency side so that the 1625 resting plate current with no s.s.b. excitation runs about 100 ma. Connect the s.s.b. exciter and, using steady tone input, tune C_{63} for maximum 1625 plate current. Using a dummy load, tune C_{65} for maximum mixer output on 7.3 Mc. The 4-Mc. injection level is controlled by varying the audio gain control of the s.s.b. exciter. This can be adjusted by monitoring the output signal on a receiver (if you have a delicate ear) or, preferably, by 'scope and the two-tone test. In W4VP's set-up an r.f. output of about 15 watts can be obtained before flattening of the peaks is apparent.

REGULATED BIAS SUPPLY

One of the requisites of any linear Class AB_2 or Class B amplifier (except those using zero-bias tubes like the 811-A) is a "stiff" bias source. Although batteries are used in many instances, they are really at their best only when the peak grid current is relatively low. Dave Mann, W6HLY, worked out a variation of an earlier regulator circuit that he uses to give a constant 40 volts of bias, over a grid current range of 0 to 80 ma. As shown, the circuit has another advantage in that a 1000-ohms-per-volt grid voltmeter can be hung across the regulator without affecting its operation.



The regulator circuit for stabilizing the bias voltage on a linear amplifier. The bias voltage can be adjusted by varying R_1

C_1 — Power-supply output condenser.
C_2 — 50-μf. electrolytic, 150 volts.
R_1 — 10,000-ohm potentiometer.
R_2 — 56,000 ohms.
R_3 — 35,000 ohms, 5 watts.
R_4 — 0.22 megohm.

» *Aside from frequency ranges, which differ with different types of s.s.b. exciters, the general design principles of variable-frequency oscillators are the same whether they are used for c.w., a.m. 'phone, or single sideband. Therefore VFOs as such are not given detailed treatment in this book. However, frequency drift is an especially acute problem in s.s.b. communication, and this article by W3ASW describes a neat and simple way of overcoming it.*

Cutting Down VFO Drift

RICHARD E. LONG, W3ASW

Some years ago the writer went through the throes of trying to sell s.s.b. to a bunch who would come up with almost any reason for not getting into it. One night a W3 said to me, "That stuff may be all right, but if you're tied to crystal, you're licked. You've got to have VFO so you can move around."

I looked at my exciter and it seemed like it ought to work with VFO. So I started out to build one for it. Knowing that a drift of more than 50 cycles would throw the stuff into inverted speech in the other fellow's receiver, I realized that I would have to build something stable.

Mounting Components

The first attempt was the construction of a series-tuned oscillator in a 3 × 4 × 5-inch box that would fit into the space allowed for the crystal oscillator it was to replace. Making this one work brought home the first lesson. Solid construction alone is not enough; the components must also be mounted so that there is a minimum of strain on the frequency-determining parts. The coil, tuning condenser and padder were all mounted solidly, but I had fastened the padder in such a manner that it was supported by two *opposite* walls of the shield box, as shown in Fig. 1A. For two weeks I hunted the cause of a burble in the signal. One night I tried the rig with the rear cover of the box removed. The burble disappeared! Put the cover back on, and the frequency would jump continually. The cure was to mount the padder from two *adjacent* walls of the box, as shown at B, thus reducing the compression and stretch on the condenser.

Temperature Effects

Although this solved the business of frequency jumping, I was still bothered by drift with temperature. The operating time has always been from 6 to 8 in the evening, and this VFO would drift for about two hours, after which it settled down. That was just dandy! My signal drifted all evening until I was ready to shut down! A thermometer placed on the VFO box showed that it actually required those two hours for the heat of the chassis to permeate the parts enclosed in the can and level off.

Why not eliminate the metal-to-metal contact between the chassis and the VFO box? I remounted the box on four ceramic pillars and hoped that it would help in the heating problem. It did! It was now about two hours before the heat would seep through the ceramic to the oscillator parts.

Isolating the Tuned Circuit

That was OK for the evenings. But on week ends, after two hours, I would take off and slide on down the band the same as before. I had only postponed the starting of the drift. Very annoying! This was in the summer of 1949. September *QST* came out with By Goodman's exciter arrangement that used switches and short pieces of coax cable connected between the tube and the frequency-determining circuit. We'll go Goodman one better. We'll leave the tube on the

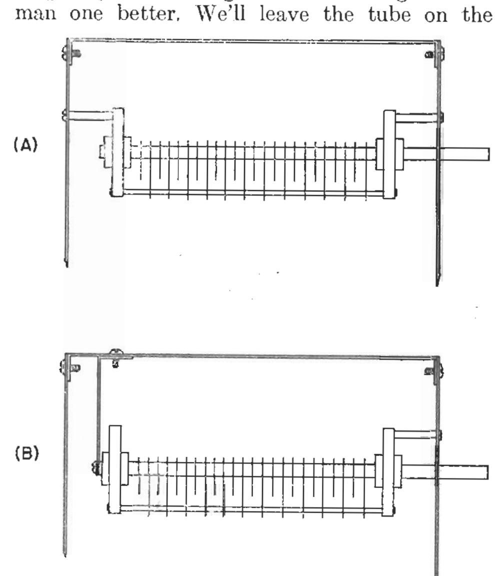


Fig. 1 — Sketches showing (A) original mounting of the variable padder and (B) the alteration in mounting that eliminated frequency jumping in a Clapp oscillator.

From *QST*, August, 1952.

chassis and put the tuned circuit in a box on the operating table and connect the two with a couple of long pieces of coax.

Yes, it worked! I put the coil, tuning condenser padder and the two 0.001-μf. fixed bridging condensers in a 3 × 4 × 5-inch box and ran 4-foot lengths of RG-59/U over to the grid and cathode of the 6AG7 in the rig, as shown in Fig. 2A. The total drift of this VFO was something around 100 cycles from scratch and, furthermore, it settled down in about ten minutes and stayed put. That was more like it. Incidentally, I used a 4-$\mu\mu$f. negative-temp. compensator (C_3) in parallel with the tuning condenser.

Simplification?

Ah me! Things went quite smoothly for a year or so. Then I moved across the hall and could not get up on the roof for several weeks to change the antenna feed lines and drop them to the new apartment. This gave time for new ideas to brew. I had passed along to several of the gang this idea of the remote tuning circuit and they responded with some elaborations of their own. One was an arrangement needing only *one* piece of coax. One cable looks simple and neater than two. I made up a new VFO using the scheme of Fig. 2B with a 6-foot length of coax.

The first thing that happened was that the darned thing wouldn't oscillate with the same circuit values. I had to remove one third of the turns on the coil and increase the series tuning capacitance before the circuit would oscillate reliably. In this arrangement the cable is in the tuned tank circuit and apparently has a very detrimental effect on the Q. Worse yet, when I put the rig back on the air, the gang began again to ask me to get back on frequency. Brother, I'm going backward here! It doesn't seem probable that losses in 6 feet of coax at 3500 kc. could be sufficient to generate enough heat to account for the drift. My guess is that the increased tank current through the coil, as a result of the lower L/C ratio after the coil was pruned to make the circuit oscillate, was responsible. At any rate, that settled the hash for the single coax line so far as W3ASW was concerned.

Improving the Q

This experience led me to consider ways in which I might improve the Q so that the L/C ratio could be made as great as possible, still maintaining oscillation. I had noticed at one time during my experiments that the oscillator behaved a lot differently when one side of the box was removed. The same components gave a much lower frequency and the circuit would oscillate with a lower value of C. Well, let's throw out the little box and try a larger one. Charley Atwater, W2JN, sent me a 4 × 5 × 6-inch box and I mounted the coil as near to the center of it as I could and then fitted the other components around the coil, as far away from it as possible. Then, mounting the two 0.001-μf. bridging condensers back in the box with the tuned circuit, as it was in the first model, I replaced the two coax lines between the box and the tube in the rig. Now the coil is larger than the one I started out with and the series tuning capacitance is almost down to half what it was then. This ought to cut the mustard. Fire up the

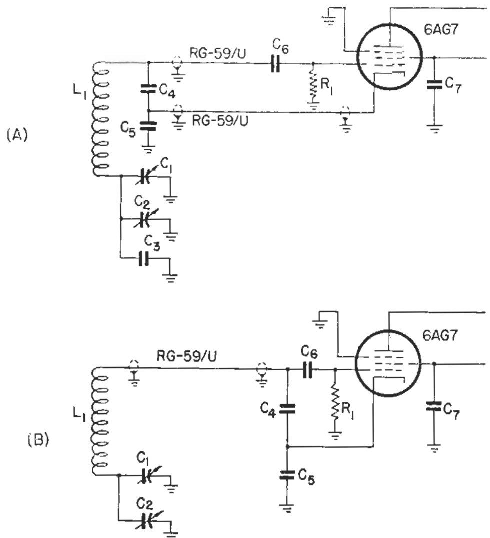


Fig. 2 — **Two methods of remote tuning for a Clapp VFO. In (A) the bridging condensers, C_4 and C_5, are mounted with the tuned circuit and coax cables connected to the grid and cathode. In (B) the bridging condensers are mounted with the tube and a single coax line connects the grid to the tuned circuit. The arrangement of (A) is preferred as discussed in the text.**

C_1 — Tuning condenser.
C_2 — Band-set condenser.
C_3 — Negative-temp. compensator (see text).
C_4, C_5 — Bridging condenser.
C_6 — Grid condenser.
C_7 — Screen by-pass condenser.
R_1 — Grid leak.
L_1 — Oscillator coil.

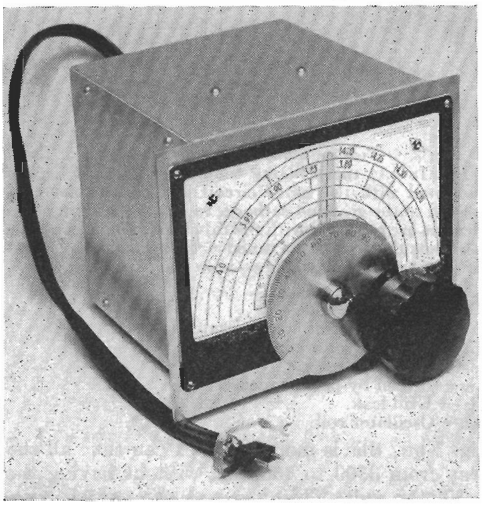
rig. Now, this is more like it. I can tune all the way from 3500 to 4500 kc. without having the oscillator quit. This means I can switch sidebands merely by tuning the VFO. Now let's set up the old modified BC-221 that has brought some fair results in the ARRL Frequency Measuring Tests. Let it run all day to make sure that it has warmed up thoroughly. Snap on the transmitter filament switches and apply plate power after 30 seconds or so. As soon as you can find the VFO beat, measure the frequency. Let the rig run for the total two-hour evening session, or longer, and check the frequency at short intervals. You come up with something like two-tenths of a dial division drift on the BC-221 which represents about 35 cycles. With heat in the house all day, there is a more even room temperature and the drift characteristics may be different as the seasons change. So far I have never seen the drift reach 50 cycles which is better than twice as good as the first model and this one uses no temperature compensator. Here again, it looks like the drift is mostly a matter of how much tank current you put through the coil to heat it up. This VFO "sits still." How about yours?

» *This is a practical application of the remotely-tuned VFO system described in the article immediately preceding. Designed to cover a few hundred kilocycles in the 5-Mc. region, it provides the VFO tank circuit for any exciter that employs the popular 5–9 Mc. combination to give outputs on 4 and 14 Mc.*

A 5-Mc. Remote-Tuned VFO

GEORGE GRAMMER, W1DF

The remote-tuned VFO circuit shown in Fig. 1 uses the popular series-tuned oscillator arrangement to cover a bandspread range in the vicinity of 5 Mc., and the VFO output can be mixed with that from a fixed-frequency single-sideband generator at 9 Mc. to give output in either the 4- or 14-Mc. 'phone bands. The constants given are for a range of 5.0 to 5.35 Mc. so that both bands are covered when the s.s.b. generator frequency is very close to 9.0 Mc. The 350-kc. spread can be reduced to 200 kc. (from 5.1 to 5.3 Mc.) for somewhat greater ease of tuning if the s.s.b. generator is on 9.1 Mc., although not everyone would consider this necessary since one of the features of the unit as shown is that it includes a mechanical dial arrangement that provides a very slow tuning rate. Setting the oscillator to zero beat with another s.s.b. station is a noncritical adjustment.



This remote-tuned circuit for a variable-frequency oscillator can be used with conversion-type s.s.b. exciters in which the single-sideband signal is generated at a fixed frequency in the vicinity of 9 Mc. Its 5.0–5.35 Mc. range permits converting to either the 75 or 20-meter 'phone bands. Ample bandspread and a slow tuning rate insure ease of setting to zero beat with other stations. The 6 by 6 by 6 aluminum enclosure is provided with a separate panel of the same height as the box and slightly wider than the dial frame.

Based on "A Tubeless VFO for the 10A," *QST*, June, 1954, and "Re the Tubeless VFO," *QST*, October, 1954.

The circuit values shown provide the low-C high-Q tuned circuit required by the series-tuned arrangement, and include sufficiently large values of capacitance at C_1 and C_2 to give good isolation between the oscillator tube and the tank circuit. The total capacitance includes the capacitance of the length of RG-59/U cable used to connect the circuit and the tube. This length should be of the order of three feet, adding about 60 $\mu\mu$f. across C_1 and C_2 individually, so that the effective capacitance in each case is about 750 $\mu\mu$f. A temperature-compensating condenser could be paralleled with C_3, C_4 and C_5, if desired, although in the unit shown the frequency drift over a period of hours was too small to justify much effort along these lines. This assumes, of course, that the location for the circuit under operating conditions will be chosen with some care so that the frequency will be affected only by room-temperature variations, which are usually small. Remote tuning gives no benefit if the circuit is installed on or close to other equipment that gets 40 or 50 degrees hotter than its surroundings after operating for a while!

The fixed condensers in the unit are silver micas, since these are far more temperature-stable than the ordinary mica condenser. Zero-temperature-coefficient ceramic condensers designed for r.f. use should be equally good. C_4 is for setting the minimum circuit capacitance to cover the proper frequency range and C_5 is the

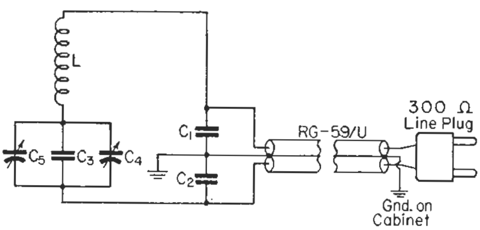


Fig. 1 — Oscillator circuit diagram. The 300-ohm line plug fits a standard crystal socket when the circuit is used as a "crystal substitute." A shielded connector is recommended when used with the unit shown in Fig. 2.

C_1, C_2 — 680-$\mu\mu$f. silver mica.

C_3 — 50-$\mu\mu$f. silver mica.

C_4 — (Trimmer) 25-$\mu\mu$f. midget variable (Bud LC-1642).

C_5 — (Tuning) 20-$\mu\mu$f. midget variable (Millen 20920 with one stator plate removed).

L — 40 turns No. 22, 1-inch diameter, 16 turns per inch (B & W 3015).

tuning condenser. The actual band covered is determined by the change in capacitance of this condenser from maximum to minimum, and the desired limits can be obtained by removing one or more plates.

Mechanical Construction

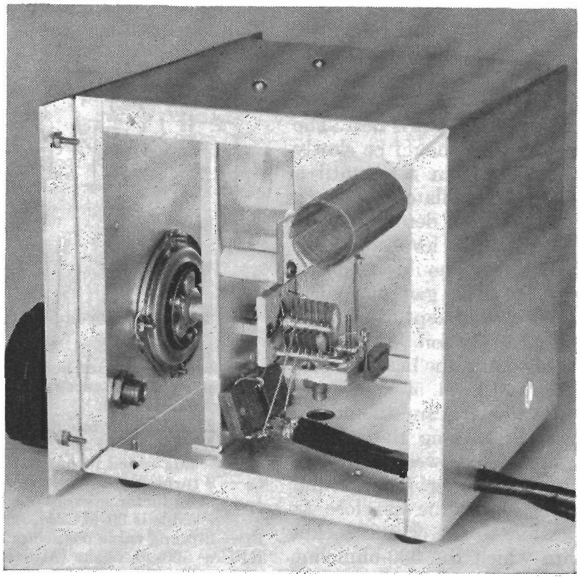
The construction shown in the photographs was adopted as a means of stiffening the assembly and avoiding mechanical strains; it also makes building easier since the upright member has all the parts mounted on it and all the work can be done outside the box. The aluminum bracket on which the oscillator circuit is mounted is three inches wide and extends from the top to the bottom of the box. The bracket has its edges bent to stiffen it, and also has a bent-over "foot" for mounting at the bottom. It is fastened to the top of the box by a piece of aluminum angle having elongated mounting holes to permit exact fitting. C_1 and C_2 are mounted to a two-terminal tie point strip that also provides connection points for the inner conductors of the coax cables. The mounting point for this strip is the only ground point for the circuit; the cables are insulated from the box by their vinyl coverings where they leave it at the rear. C_4 is mounted with its shaft downward and is screwdriver adjusted through a hole in the bottom of the box.

The tuning condenser, C_5, should be of the type that mounts on studs in the end plate, rather than the single-hole mounting type. The stud mounting avoids the necessity for grounding the rotor plates through the mounting and thus eliminates multiple paths for the ground currents. Careful alignment of the condenser shaft with the dial is essential for backlash-free operation, since misalignment will tend to distort the condenser shaft in a direction dependent on the direction of rotation of the dial. In this unit, proper alignment is provided for by elongating the mounting holes for the supporting member sidewise, and elongating the condenser mounting holes vertically. The test for good alignment is that with the condenser and dial in final position, but with the dial setscrew not tightened, it should be possible to turn the dial throughout its range without any movement of the condenser shaft. This indicates that the shaft and dial coupling are both "on center." (It is assumed, of course, that the condenser shaft is sufficiently free in the coupling to permit independent rotation. If it is not, use steel wool on the shaft to give it a free, but not loose, fit.)

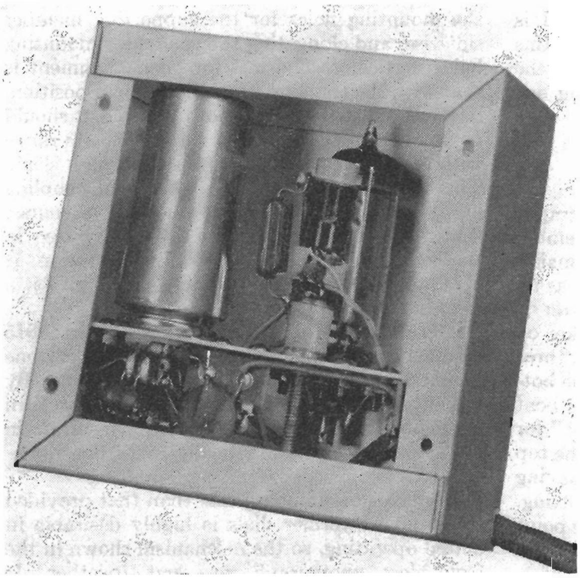
The tank coil, a length of B & W No. 3015 Miniductor, is cemented to a strip of polystyrene mounted on two 1-inch stand-off insulators. By cementing the whole length of the coil each turn is anchored to the supporting strip, resulting in an assembly that is practically free from vibration effects.

A much slower tuning rate than that provided by standard vernier dials is highly desirable in actual operating, so the mechanism shown in the front-view photograph was put together. It combines a National type ACN dial with a Type K and a large knob. Since the ACN has a 3/16-inch shaft while the K has a 1/4-inch hole, it is necessary to put a 1/4-inch sleeve on the shaft. The sleeve can be made by drilling out a short piece of quarter-inch brass tubing, slipping it over the ACN shaft, and running some solder in the end. This combination has a mechanical step-down ratio of about 35 to 1, giving a "light" feel and ample adjustment range for frequency changes of a few cycles.

With the construction shown, provided the precautions described above are taken, the unit should be immune to vibration and free from all backlash except that inherent in the original dial mechanism. This is detectable on critical examination, but is of too small magnitude to constitute an annoyance in fine tuning.



Side view of the VFO tank with the box cover plate removed. Rubber feet on the bottom help provide insulation from mechanical vibration. The end plate of the trimmer condenser, C_4, is visible behind the tuning condenser.



Interior construction of the oscillator unit. The small shelf is the full width and depth of the box, thus making full use of the available space. The coil-form slug is adjusted through a hole in the bottom side of one L-shaped piece.

plug fits. The best place to ground is at the screw holding the crystal socket, with the very minimum of lead length.

Existing Exciters

The remote tuned circuit as described above can be directly substituted for the crystal in a crystal-controlled oscillator that is designed to operate in the vicinity of 5 Mc. It has been used successfully in this way with the Central Electronics 10A exciter, and although this circuit arrangement provides no buffering to isolate the oscillator, tests have shown that frequency-modulation effects are negligible at levels up to where the exciter output begins to "flatten." Overdriving will cause the VFO frequency to shift, but since this represents an improper operating condition it should be avoided in any event.

Whatever the exciter arrangement used, the stability should be carefully checked before putting the unit on the air. This can be done by listening to the oscillator with the receiver b.f.o. on, the receiver being tuned directly to the oscillator frequency and the signal pick-up adjusted so that there is no overloading in the receiver. Single-tone input is best. The audio input should be slowly increased from zero until a change in oscillator frequency can just be detected. The power-output level at which this occurs should be noted and the audio gain and voice input always kept below it in actual operation. An oscilloscope makes the best monitor of output level, of course.

Particular care should be used in making the ground connection between the cables and the exciter. The ground point should be as close as possible to the crystal socket into which the 300-ohm line

Companion Oscillator Unit

Where there is any doubt about sufficient isolation to eliminate f.m. effects, or an exciter is under construction, provision should be made for a separate oscillator tube with an adequate buffer arrangement. The circuit of Fig. 2 incorporates these features. It is built as a small unit that can be used externally with an exciter such as the 10A or 10B, but could logically be made an integral part of a homebuilt exciter. As shown installed on a 10B, the filament and plate voltages are obtained from the socket provided for that purpose on the exciter. The oscillator tank circuit remains as shown in Fig. 1 with the exception that a shielded 3-conductor plug is substituted for the 300-ohm line plug.

Except for the fact that in recent years it has become customary to use an e.c.o. version of the series-tuned oscillator while Fig. 2 shows a

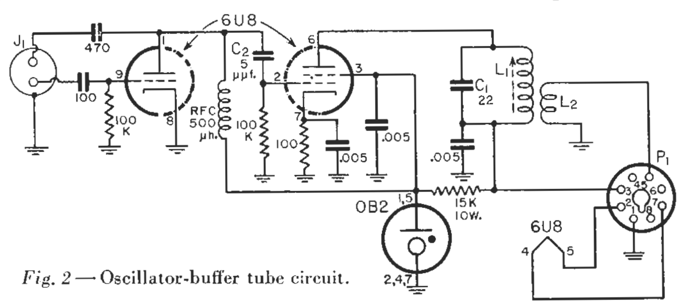


Fig. 2 — Oscillator-buffer tube circuit.

Capacitance values below 0.001 μf. are in μμf., 0.001 and above in μf. All condensers ceramic except as iudicated below. Resistors are ½-watt composition except as indicated.

C_1 — 22-μμf. silver mica (20 or 25 μμf. also satisfactory).
L_1 — Adjustable to app. 30 μh.; 1-inch winding of No. 30 enam. on ⅜-inch slug-tuned form (National XR-93).
L_2 — 9 turns No. 24 d.c.c. wound over ground end of L_1.
J_1 — 3-conductor microphone connector (Amphenol 80-PC2F; mating plug for cable is 80-MC2M).
P_1 — 8-prong cable connector, male (Amphenol 86-PM8).
RFC — 500-μh. choke (Millen 34300–500).

grounded-cathode triode, the circuit is quite straightforward. The separate amplifier was used in the thought that it would give better isolation than an e.c.o. plate circuit, particularly when a tuned output circuit is used. The only unusual component value is the 5-$\mu\mu$f. condenser, C_2, used to couple the oscillator plate to the buffer grid. This was made just large enough to drive the buffer to the required output. A slug-tuned tank, L_1C_1, is used in the buffer plate circuit, and is coupled by a small coil, L_2, to the No. 1 grid of the 6BA7 in the exciter through the socket connection on the rear chassis wall. The number of turns on L_2 was adjusted to deliver maximum voltage to the 6BA7 No. 1 grid through a short (7-inch) length of ordinary 5-conductor cable. It is possible that a different cable length would require a different number of turns; in this case just enough length was used to permit mounting the oscillator unit on the back of the exciter cabinet.

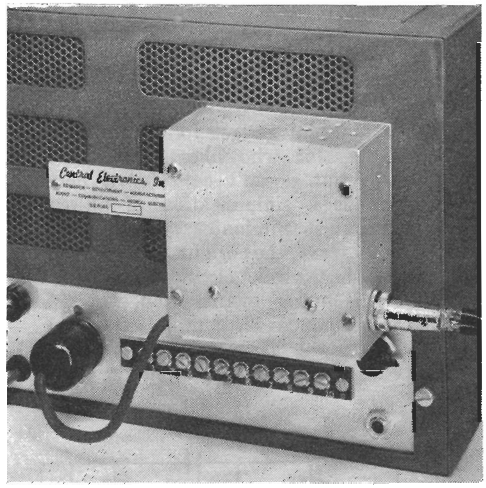
The circuit of Fig. 2 is built on a small shelf inside a 2 × 4 × 4 aluminum box. The shelf is mounted on one of the removable sides, and all the wiring except the connections to J_1 is done on just these two pieces before the rest of the box is attached. The box portion is sawed through at diagonal corners as shown, and J_1 is mounted on one of the L-shaped pieces. To insure a good ground connection between J_1 and the oscillator circuit an oversize soldering lug is made from a piece of sheet copper, with a hole large enough to fit under the mounting nut and a shank long enough to reach the oscillator ground point on the small "deck."

The second L-shaped piece has a small cut-out in one corner to allow the cable to pass through, and is attached after the wiring is done and the unit has been tested. The second cover plate, when attached, makes the final assembly practically as rigid as the original box. This plate is used as a template for drilling the back of the exciter cabinet, where the whole unit is held by self-tapping screws put in from inside the cabinet.

Note that the 6U8 is fitted with a shield. This was found to be of considerable benefit in reducing the effect on the oscillator frequency of adjusting the slug in L_1.

Because the tube gets rather hot and the space in the box is limited, a good deal of heat is transmitted to the other components unless some means of ventilation is provided. Some holes drilled in the top and bottom edges of the box, over and under the 6U8, help keep things cooler and thereby minimize frequency drift.

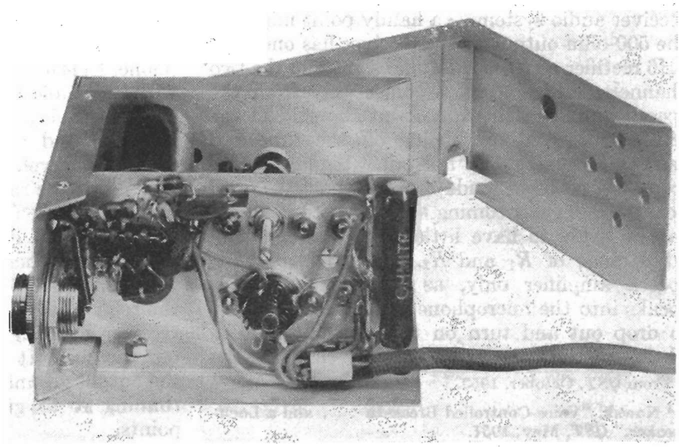
To cover the necessary frequency range — 350 kc. when the s.s.b. generator frequency is 9.0 Mc. — without adjustment a broadly tuning low-C circuit would be desirable at L_1C_1. However, a 1-inch winding is about all that can be



Showing the oscillator-buffer unit mounted on the back of the exciter cabinet.

used on the coil form if the slug is to have much tuning effect, so it is necessary to use some C in the buffer tank. With L_1C_1 set at the center of the range the output, as measured by the rectified voltage at the No. 1 grid of the 6BA7 (using a v.t. voltmeter) drops off to about one-half at the ends of the range. The constants were chosen so that the rectified voltage at the center of the range was about 20 volts. The voltage is not critical, since the normal exciter output power can be obtained with identical waveform with as little as 5 or 6 volts.

Practically all the components are grouped around the oscillator tube socket. With one L-shaped side piece removed as shown, all wiring is readily accessible.



» During the early days of amateur s.s.b. work, it was hoped that the suppression of the carrier would permit "duplex" operation, but this was found to be unworkable. It was just a step, however, to voice-controlled break-in, which resulted in a very close approximation of true duplex operation. The crowning achievement was "break-in with a loudspeaker," in which the operator's voice will operate the transmitter but a sound from the loudspeaker will not.

Simplified Voice Control with a Loudspeaker

WALTER N. HUNTER, W6IBR

This is a useful gadget for the single-sideband operator who likes voice-controlled break-in but doesn't care for a headset. This circuit is a simplification of an earlier circuit, in that it eliminates two transformers and a tube from the original design. Hank Turkel uses the Nowak[1] circuit at W6LZE and has tried this newer modification. He reports that the performances are comparable.

Referring to the circuit diagram in Fig. 1, the 6SN7 is a two-channel amplifier. The top channel is connected to the transmitter speech amplifier, ahead of the audio gain control. The lower channel is connected to some point in the receiver audio system — a handy point might be the 500-ohm output, if the receiver has one. The 6H6 rectifier rectifies the audio signals in the two channels. When a signal comes only from the speech amplifier, the voltage at the grid of the 6SL7 goes negative, cutting off the tube and causing the relay to "fall out." Signal coming from the receiver tends to bias the 6SL7 grid positive — signals coming through both channels can be made to have little or no net effect by adjustment of R_1 and R_2. A signal from the speech amplifier only, as when the operator speaks into the microphone, will cause the relay to drop out and turn on the transmitter. The "hold-in" time can be modified by changing the value of C_6 — a larger value of capacitance will give a longer hold-in.

It will be noted that the components specified are different for each channel. This is because the writer's speech amplifier has shaped frequency response and capacitors C_3 and C_5 tend to shape the lower-channel response in a like manner.

W6LZE has suggested that the only control needed on the front panel is R_8, which is used to balance for different operators. His experience has been that, once adjusted, R_1 and R_2 will need no further attention, and I have also found this to be the case.

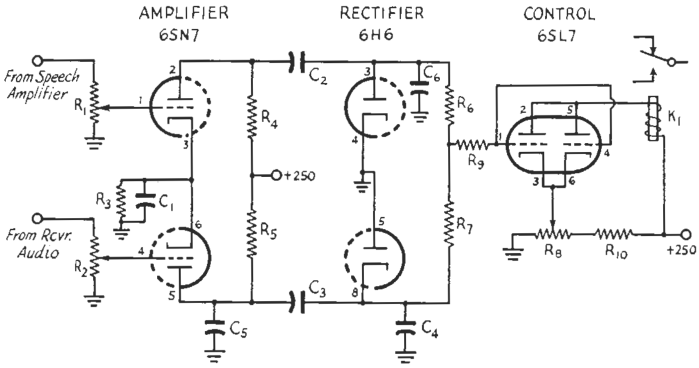


Fig. 1 — Circuit of the voice-operated control circuit for use with a loudspeaker.

C_1 — 25-μf. 25-volt electrolytic.
C_2, C_6 — 0.1-μf. paper.
C_3, C_4 — 0.05-μf. paper.
C_5 — 0.01-μf. paper.
R_1 — 0.5-meg. volume control.
R_2 — 50,000-ohm volume control.
R_3 — 470 ohms.
R_4, R_5 — 0.1 megohm.
R_6, R_7 — 0.47 megohm.
R_8 — 500-ohm 4-watt wire-wound potentiometer.
R_9 — 1 megohm.
R_{10} — 10,000 ohms, 10 watts.
K_1 — 10,000-ohm sensitive relay.

To put the unit in operation, first adjust R_8 to the point where the relay is held closed with a positive action. While speaking into the microphone, R_1 is now adjusted until the relay operates with negligible lag. Now turn on the station receiver (placing the microphone at its normal location) and adjust the receiver audio for a fairly high level. The relay will now probably be tripped by the receiver output. Adjust R_2 to just eliminate this effect.

One word of caution. If your receiver gives a loud "pop" when the relay operates, it will feed through the 6SN7 receiver amplifier and may pull the relay in again. The answer to this is to use a receiver silencing system that will operate without the click. It was done at W6IBR by lifting the output ground connection on the receiver and running it to ground through a set of relay points.

From *QST*, October, 1953.

[1] Nowak, "Voice-Controlled Break-In . . . and a Loudspeaker," *QST*, May, 1951.

» Here is still another voice-controlled break-in circuit, together with such useful operating accessories as an anti-trip ("break-in with a loudspeaker") circuit and an audio oscillator for furnishing a test signal.

A Speech Amplifier and Operating Accessories for S.S.B.

DANA A. GRIFFIN, W2AOE, AND DONALD H. FRYKLUND, W2HLP

A FIXED-FREQUENCY oscillator is sufficient for tune-up and checking of an s.s.b. transmitter, and provision is made for such an oscillator in the speech amplifier-voice control circuits shown in Fig. 1. Using an *RC* network that can be switched in and out as desired, the first speech-amplifier stage can be converted into a 1500-cycle audio oscillator.

From "Delay-Line Phase Shift," *QST*, March, 1954.

In addition to provision for converting the first speech stage to an audio oscillator, there are other features in Fig. 1 that are useful and of interest to s.s.b. operators. By means of S_2, the voice-control rectifier and amplifier can be used as a v.t. voltmeter for checking both a.f. and r.f. levels at various points in the transmitter. This is a distinct advantage both in initial adjustment of the transmitter and in monitoring its performance. With S_2 in the extreme left-hand

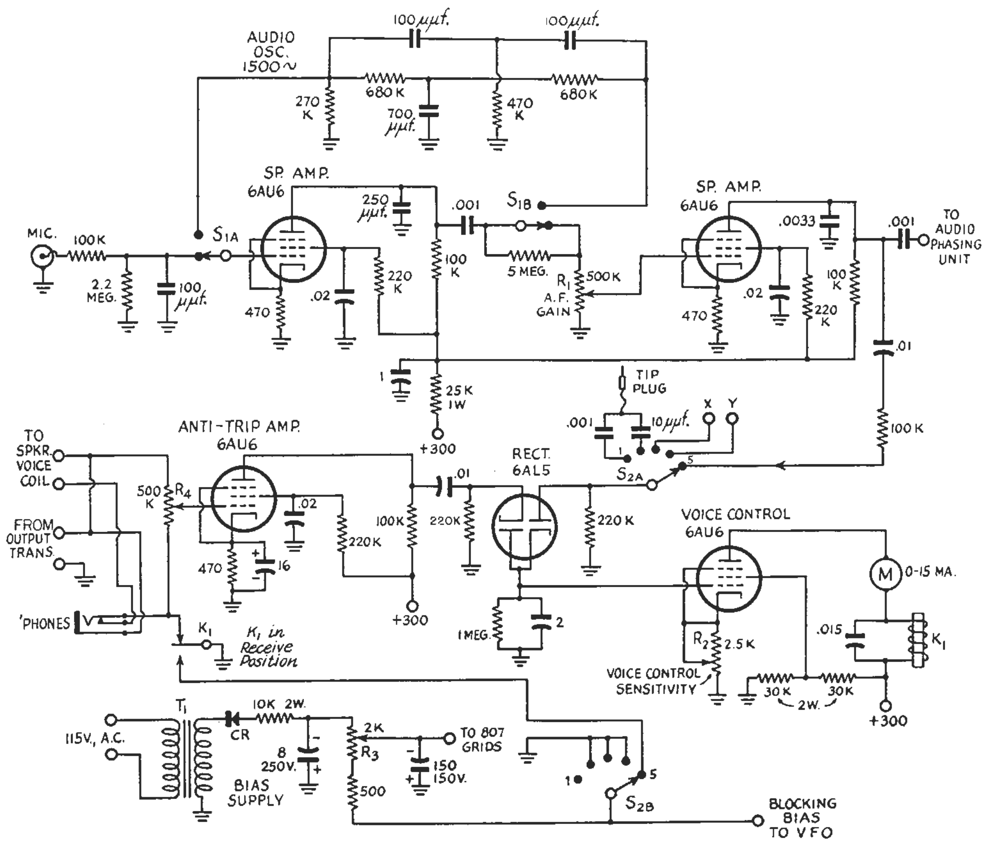


Fig. 1 — Speech amplifier and operating accessories of the s.s.b. transmitter shown in the photograph. Capacitance values are in μf. and resistors are ½ watt, unless otherwise indicated.

CR — Selenium rectifier.
K_1 — D.p.d.t. relay (one set of contacts used), 10,000-ohm coil.
T_1 — 1-to-1 transformer, 115 v.
See text for connection to *X*. *Y* goes to the final r.f. output link through a 2-μμf. coupling condenser.

position, a test prod is connected to the voice-control rectifier through a 0.001-μf. condenser for audio measurements. By the use of pin jacks installed at appropriate points, the a.f. voltages applied to the modulators can be balanced up. In the second position of S_2 a 10-μμf. condenser is connected between the probe and rectifier for balancing the r.f. voltages applied to the modulators at test points in the exciter. In the third position, X, the rectifier can be connected to the output of the exciter, where the circuit is used for checking carrier suppression. The fourth position can connect the rectifier to the output link of a linear amplifier through a condenser of a few μμf., where it serves as a level indicator. The fifth and final position connects the speech amplifier to the voice-control rectifier and is the normal operating position for S_2.

The audio gain and operating sensitivity for proper functioning of the voice-control circuit are controlled by R_1 and R_2, respectively. The anti-trip sensitivity control, R_4, is adjusted for the minimum gain necessary to hold the transmitter off with normal output from the speaker.

The bias supply shown in Fig. 1 provides both operating bias for a pair of 807s in the linear amplifier and, through the voice-control relay, blocking bias for the 807 and the VFO.

Voice-Controlled Break-In with a Loudspeaker and No Relays

SINGLE-SIDEBAND enthusiast Ray Brandt, W9LIJ, uses the circuit shown in Fig. 1 to give him smooth break-in with a loudspeaker, and it has the additional attraction that no relays are required. The unit delivers −90 volts blocking bias for the transmitter and −65 volts blocking bias for the receiver. These biases exist at either full value or not at all, so the action is positive.

In the "receive" condition, audio from the receiver (speaker) is rectified by the right-hand diode and holds the right-hand 6AG5 conducting. The left-hand diode is also rectifying audio picked up from the speaker and passed through the speech amplifier, but the setting of the "Threshold" control holds the left-hand 6AG5 just below cut-off. Talking into the microphone unbalances this condition but causes no output in the receiver until the circuit is tripped, because the transmitter is off. Once the circuit is tripped, the receiver no longer delivers audio, and the bias developed by the right-hand diode decays. The transmitter is held on until the rectified output of the audio from the speech amplifier decays below the threshold value.

W9LIJ adds that if the receiver is to be disabled by applying the −65 volts to the a.v.c. bus, it is recommended that the bias be applied through a diode, the plate to the a.v.c. bus and the cathode to the "disable receiver" lead. This also requires that there be some resistance left between ground and the a.v.c. bus, of course, when the a.v.c. is switched off for s.s.b. reception.

The time constants of the diode circuits can be made variable, but the values shown have proven quite satisfactory at W9LIJ. Mica condensers are to be preferred, to stabilize the time constants

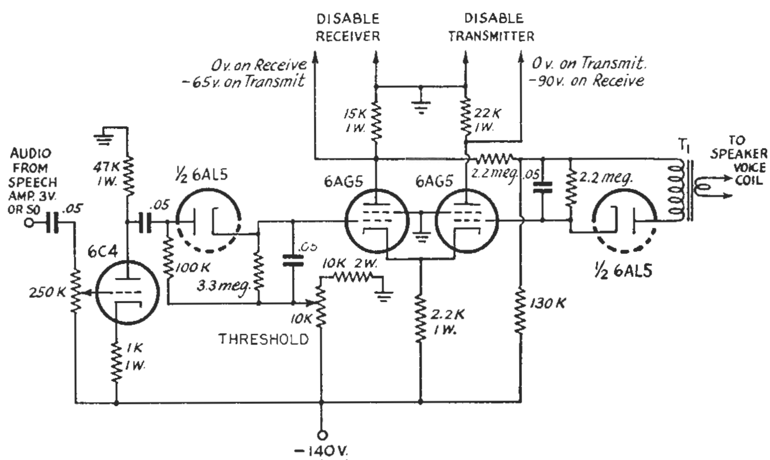


Fig. 1 — This voice-control circuit requires no relays and permits loudspeaker reception. T_1 is a midget a.c.-d.c. output transformer with the low-impedance winding connected to the speaker voice coil.

From *QST*, November, 1953, and February, 1954.

under various conditions of temperature and humidity.

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Laurence Smith, W7FOM, has been using the W9LIJ voice-control circuit and is pleased with its operation. He made some modifications and had some experiences that he passes along for others who might be interested in the circuit. In his words, "Any sharp-cut-off pentode r.f. tube can be substituted for the 6AG5s used in the circuit. I ended up with one 6AK5 and a 9001 because that's what the junk box yielded. However, 6AK5s were tried in both sockets with good operation. Improved 'no-trip' action may be obtained by altering the value of the common cathode resistor and/or the plate resistor of the right-hand 6AG5. I used a 3000-ohm cathode resistor and 15,000 ohms in each plate circuit. This reduced the cut-off bias for the transmitter to about 30 volts (with a 120-volt negative supply) but was sufficient to bias off the 6SN7 mixer and 6AG7 output tube of the crystal-filter exciter.

"Receiver bias was about 45 volts, which worked on my BC-312 when applied to the a.v.c. bus. These three resistors should be changed to the values that give the required voltages for each individual set-up. If a germanium diode is used in series with the bias to the receiver a.v.c. bus, pick one with a very high back resistance. I had some trouble with a 1N34 diode in that the back resistance decreased after the receiver operated for a period of time. This upset the S-meter readings and reduced the a.v.c. action. One diode section of a 6AL5 dual diode was installed in place of the 1N34 and eliminated the trouble.

"I originally tried using half a 6J6 in place of the 6C4. However, the circuit tripped too easily from the loudspeaker. A triode-connected 6AQ5 (screen tied to plate) with a 10,000-ohm plate load resistor was substituted, and operation was much improved.

"If low-pitched sounds from the speaker have a tendency to trip the circuit, the coupling condenser to the input might be changed to a smaller value to reduce this effect. If it is sensitive to high-pitched sounds, a small by-pass condenser from plate to ground should help to minimize this trouble.

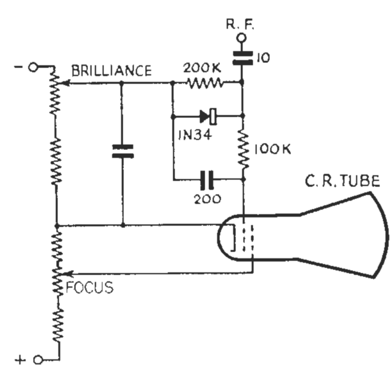
"I use one of the cheaper dynamic mikes with this circuit, about two feet from a loudspeaker that faces the operating position. Excellent no-trip action is obtained with a fair amount of speaker volume.

"The power supply for this circuit consists of a pair of small 6.3-volt heater transformers connected back-to-back (6.3-volt windings tied together). This furnishes heater power for the tubes, and −120 volts d.c. from the 120-volt winding of the second transformer through a selenium rectifier, small filter choke and two 40-μf. 150-volt condensers. The + side of the supply is grounded, of course, so the condensers and selenium rectifier are reversed from a normal power-supply circuit."

'SCOPE INTENSIFIER

If you don't read the *RSGB Bulletin* and thus get a chance to see G3CU's excellent column, "CQ Single Sideband," you will have missed the neat trick shown here. It's a dodge for reducing the intensity of the 'scope trace when there is no signal, and thus reduce the chances for "burning" a line on the screen. It's easy with an a.m. rig, of course, where you're clunking on and off with all sorts of power supplies, and turning off the 'scope can be made part of that procedure, but with s.s.b. and break-in, the screen can go fast if you turn up the brightness high enough to do some good on peaks. However, by adding a germanium rectifier and a few resistors, the 'scope will brighten up when r.f. is delivered to the antenna and dim out when none is forthcoming. The point marked "R.F." can be connected to the feeder or some convenient point where r.f. is being developed.

It is, of course, a simple thing to tie in your 'scope to your voice-controlled break-in circuit, but G2IG's system becomes self-powered and automatically adjusts itself to the proper level. G2IG reports that, in practice, the center quarter inch of vertical deflection is invisible, while the remaining deflection is easily visible but not bright enough to cause a burn.



By using rectified r.f. as an intensifier, G2IG monitors his s.s.b. transmitter without danger of burning a line on the 'scope face.

» *The TR box system used in radar can be applied to an amateur station.*

Break-In with One Antenna

M. E. HIEHLE, EX-W2SO

When using the transmitting antenna for receiving, the usual procedure is to employ an antenna change-over relay, but this precludes practical break-in operation.

In radar, the "TR box," or "duplexer," was the solution to this problem. Essentially, it gave the effect of the circuit in Fig. 1. When the transmitter was "on," the quarter-wave-

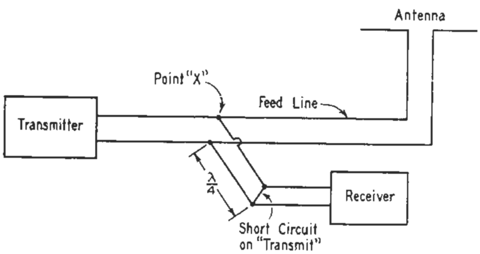


Fig. 1 — This arrangement will permit use of the same antenna for transmitting and receiving if the quarter-wavelength stub is shorted during "transmit" periods. It will be recognized as the "TR" circuit used in radar.

length line was short-circuited, and the receiver was protected. The quarter-wave line had no appreciable effect on the feed line from transmitter to antenna because a short-circuited quarter-wave line shows high impedance at the open end, and hanging a high impedance across the line at point X has no effect on the line. When the transmitter was "off" the short was removed, and if point X was the correct distance from the transmitter (the "off" transmitter looked like a high impedance), all of the energy coming down the feed line would go into the receiver. In radar work, the short circuit was obtained by either an open spark gap or one in a "TR tube."

From *QST*, Nov., 1949.

The TR-tube system won't work on the amateur bands because the keyed spark would bring the FCC on the double (spark has been outlawed for some time now) and, anyway, you would probably have quite a time with a feed mechanism to replace the gap points. However, since amateur rigs aren't keyed as fast as radar rigs, it is possible to use a relay to short circuit the line. "Yeah, but the contacts will burn up or arc over or something," you say. Well, let's see.

If your transmission line is matched to the antenna, the voltage on the line is

$$E = \sqrt{PZ_0}$$

where P is the power output of the transmitter and Z_0 is the line impedance. For any standing-wave ratio on the line,

$$E_{\text{max.}} = \sqrt{PZ_0e}$$

where e is the voltage s.w.r. To take an extreme case, consider 1 kw. into a 600-ohm line with a 20-to-1 s.w.r. $E_{\text{max.}}$ works out to be 3500 volts. Hence the voltage across the line at point X might run this high. The current through the short will be

$$I = \frac{E_{\text{max.}}}{Z_1}$$

where Z_1 is the impedance of the stub line. Assum-

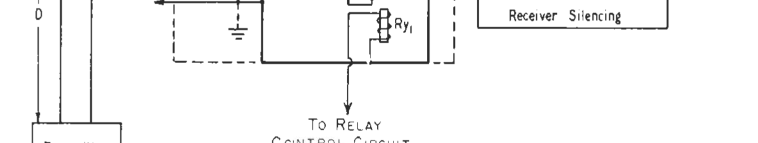


Fig. 2 — A practical amateur application of the "TR" system. The short circuit is obtained with relay Ry_1, and the job is made easier by the step-up in impedance. L_1C_1 tunes to the frequency in use, C_2 is a 0.001-μf. mica receiving-type condenser. The transmitter oscillator is turned on and off through the two leads marked "To Osc." R_1, a 5000-ohm potentiometer, establishes the receiver gain when the transmitter is on.

ing a 300-ohm line for the stub, the current through the short would be

$$I = 3500/300 = 11.7 \text{ amperes.}$$

One of the peculiar characteristics of shorted quarter-wave stubs is the fact that the short current is independent of the short resistance. This means that any relay used for shorting the stub must have low contact resistance. But the foregoing calculations were made for the worst conditions, and most practical cases will not require a relay with such high-current requirements. Even then, the 12-ampere figure is not a large one and can be met by a number of different relays.

A Practical System

Several features can be added to make the system foolproof. First of all, it is necessary to insure that the relay does not open with the transmitter power on. This means that the following sequence of events be obtained: (1) relay closes, (2) transmitter goes on, (3) transmitter goes off, (4) relay opens.

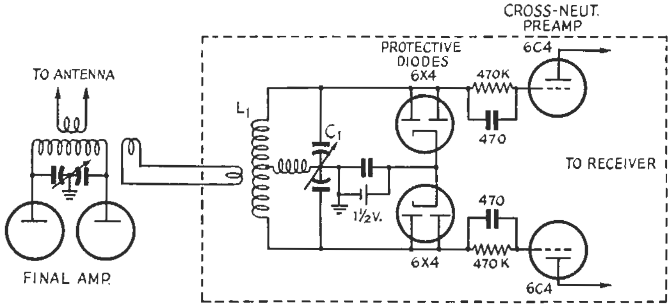
Fig. 2 shows a circuit that combines the "T-R" system with a means for reducing the gain of the receiver during "transmit" periods. The high-current requirement for the relay is decreased by transforming the impedance at the short-circuit point to a higher value. For a 3-to-1 turns ratio, the current is about 4 amperes instead of 12. The separate coil and Faraday shield represent good engineering practice and reduce capacity coupling. Using the shield also permits balanced-to-balanced or balanced-to-unbalanced line without undue capacity coupling. The length of stub shown as $\lambda/4$ requires an electrical length of a quarter wavelength. For 300-ohm Twin-Lead this is equal to $178/f$ and for coaxial line is $146/f$, for the length in feet and the frequency in Mc. These formulas include a 10 per cent shortening factor for the coil reactance. The bold lines indicate where short heavy leads should be used. Relay Ry_1 should be a fast one.

Some precautions should be observed with the quarter-wavelength stub. It should be kept several inches away from the wall and metallic objects, but it can be folded back on itself if good separation is maintained. As for the relays, a check on whether or not Ry_1 opens too soon can be made by connecting a neon bulb across the contacts. If the bulb lights, the contacts open too soon. The neon bulb can be left permanently in the circuit as added protection, although it is unnecessary.

Send-Receive with One Antenna

During the course of some scatter-sounding experiments [1] at Stanford University, the men (and also radio amateurs) in charge of the project devised a means for using the same antenna for transmitting and receiving that is also applicable to s.s.b. As shown in Fig. 1, the tank circuit becomes small and negligible power transfer occurs. The high series resistors to the triode grids further protect the input tubes of the receiver. As soon as the preamplifier grids go positive, a high bias is developed across these resistances that limits the flow of grid current.

Fig. 1 — A receiver-protection circuit that uses 6X4 diodes for decoupling the receiver from the antenna when the transmitter is operating. The diodes short the tuned circuit, L_1C_1 (resonant to the signal frequency) and thus lower its Q markedly. The series resistors to the grids of the preamplifier tubes further limit the power reaching the receiver.



final amplifier of the transmitter is link-coupled to the receiver through a preamplifier using diodes and neutralized triodes. The diodes protect the receiver and prevent loss of transmitter power. The Q of the preamplifier input circuit is so reduced, when the diodes are conducting, that the coefficient of coupling with the transmitter

Since the shunting capacitors are small, the bias built up discharges in a very short time interval after the transmitter goes off.

The triode amplifier ahead of the receiver was used in the Stanford work primarily to improve the noise figure of the receiver. Don Kinney, W8FSA, has tried the system in his receiver without adding a neutralized-triode preamplifier, and he has found that it gives good results with a

[1] Villard and Peterson, "Instantaneous Prediction of Radio Transmission Paths," *QST*, March, 1952.

final running up to a kilowatt on peaks. The receiver used a 6K7 input stage, and his only modification was to add a single shunt 6X4 and the series resistor and capacitor of one side of the circuit in Fig. 1.

Vacuum Tube Coupling

Another solution to the problem is that devised by Tom Puckett, W2JXM.[2] The basic principle is to use a low-noise coupling tube between antenna line and receiver, and to bias off the tube during transmit periods. The method has a few limitations as described, and the most serious one is that it is applicable only to a matched 50-ohm transmission line if the power level is as high as 500 watts. As shown in Fig. 2, a 6AJ4 grounded-grid r.f. amplifier couples the receiver to the low-impedance line to the antenna or antenna coupler. When the transmitter is off, the 6AJ4 acts as a normal grounded-grid amplifier for the receiver, but during transmit periods it is biased high enough to prevent any appreciable signal reaching the receiver. Additional circuitry can be added that will simultaneously reduce the gain of the receiver.

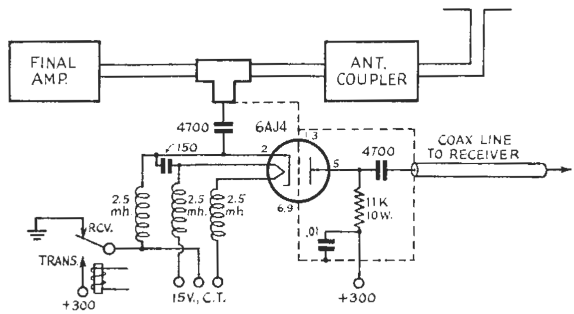


Fig. 2 — In this receiver-protection circuit, a 6AJ4 coupling tube is biased back when the transmitter is on and thus minimizes the power reaching the receiver.

The physical construction of the 6AJ4 stage should be such that the plate circuit, which feeds the receiver input, is completely shielded, so that the only coupling is through the 6AJ4 itself. A metal shield should enclose the 6AJ4 plate terminal and the receiver connection, which should preferably be coaxial. The shield can cross the 6AJ4 socket directly across the grounded grid terminals. The lead from the 6AJ4 plate resistor should be by-passed where it leaves the shielded enclosure.

The coaxial transmission line between the antenna coupler and the transmitter should pass directly through the control unit, as shown in Fig. 2, rather than reach the unit through a "T" connection. The "T" connection may cause resonance effects, giving a key-down voltage at the control unit of more than the 200 volts r.m.s. that is the maximum allowable for proper operation. This 200 volts, by the way, corresponds to 800 watts in a matched 50-ohm line, and 570 watts in a 70-ohm line.

It is necessary to feed the 6AJ4 filament in the manner shown in Fig. 2 to keep within the heater-cathode rating. With the National R-300S chokes used, about 15 volts of heater supply voltage was needed because of the drops in the resistance of the chokes.

[2] Puckett, "Break-In with One Antenna," *QST*, March, 1954.

Automatic Antenna Switching

Although relays can be used for quick switching of antenna from receiver to transmitter when working voice-controlled break-in, it is much nicer to do it electronically. Two circuits used for this purpose by Bill Rust, W2UNJ, are shown in Fig. 3. The circuit at A is along the lines of that described by W2OUA (Cronin, *QST*, June, 1952). The system at B is presently in use at W2UNJ. The circuit C_2L_3 should be low-C and tuned for maximum received signal. It is broad enough to hold over a 'phone band without retuning. The neon bulb must have the resistor removed, of course, and a ¼-watt neon will suffice for 75 watts or so. The pilot lamp is a safety fuse to protect the receiver in case of failure of the "TR" switch. In some cases it may be necessary to shield the TR circuit to prevent the radiation of harmonics and subsequent TVI.

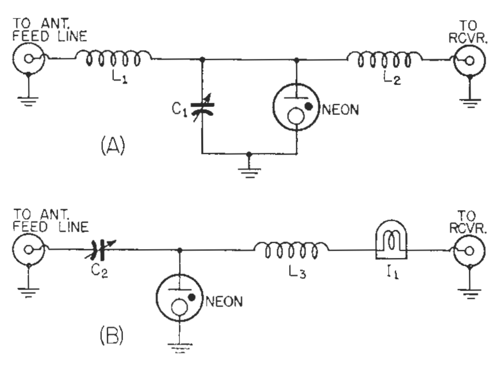


Fig. 3 — Two TR ("transmit-receive") switch circuits that have been used by W2UNJ. The circuit at A uses two large inductances, L_1 and L_2, a small condenser, C_1, and a neon bulb. The circuit at B uses only one coil and adds a small flashlight bulb for added protection of the receiver.

C_1, C_2 — 50-μμf. variable.
L_1, L_2 — 70 turns No. 28 on 1-inch diam. form (for 3.9 Mc.).
L_3 — 90 turns No. 28 enam. on 3¼-inch diam. form (for 3.9 Mc.).
I_1 — 6–8 volts 150 ma.
The neon bulb can be ¼ watt with a low-powered rig and 2 or 3 watts with a high-powered transmitter.

INDEX

Hot Ham News from hallicrafters

SINGLE SIDEBAND

- **TRANSMITTER/EXCITER**
- **RECEIVER**

We're excited about these two new single sideband units developed here at the "world's largest ham shack." The compactness of each unit, the completeness, the new styling and, above all, the features that will make both the Hallicrafters transmitter/exciter and receiver the most talked about ham news in years.

Check the features below, see your Hallicrafters jobber or write for complete specifications—we're sure you'll agree with us.

Bill Halligan, Jr.

Model SX-96 SELECTABLE SIDEBAND RECEIVER

- Covers Broadcast 538-1580 kc plus three S/W 1720 kc—34 Mc.
- Full precision gear drive dial system.
- Double conversion with second oscillator crystal controlled.
- Selectable side band reception of both suppressed carrier and full carrier transmissions.
- Mixer type second detector.
- CW operation with AVC on.
- Delayed AVC.
- Calibrated bandspread—"S" meter—double superhet.
- 11 tubes, 1 rectifier and voltage regulator.

- Highly stable VFO with full 100:1 ratio gear drive system built-in.
- Stability comparable to most crystals .01%.
- Ample gain for 55 db microphone with hum and noise 40 db down.
- Full 40 watt lineal peak power output.
- Unwanted sideband at least 40 db down.
- Undesired beat frequency down 60 db or more.
- T. V. I. suppressed.
- Provisions for coaxial output fitting.
- Built-in voice control circuit with bias switching for final amplifier.
- AM—CW—SSB—21 tubes plus voltage regulator and 2 rectifiers.

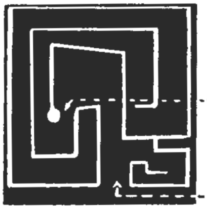
Model HT-30 SINGLE SIDEBAND TRANSMITTER/EXCITER

hallicrafters

CHICAGO 24, ILLINOIS

178

NO MAZES? AMAZING!



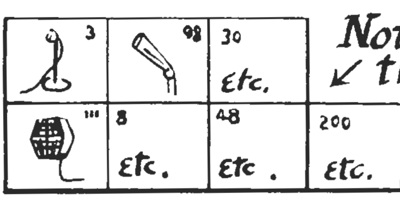
A VERY HAND-Y GADGET!

PANELS ONE ON EVERY PAGE!

TUBES *unreversed*

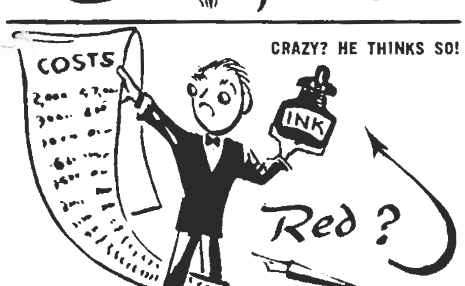
TUBES *reversed*

SPLIT-LISTINGS VERBOTEN!



EVERYTHING'S IN STOCK!

CRAZY? HE THINKS SO!



You need this free 224-page book!

AT RADIO SHACK we like to think our catalog is as indispensable to the Radio Amateur as any other reference manual he uses.

Our just-published 1955 edition is probably the easiest catalog to find your way around in you ever saw. The front cover, for example, has *two indexes* instead of the traditional science-fiction assortment of mikes, CR tubes, towers and surrealistic real estate.

The manufacturers' index on the cover lists some 230 of our suppliers and tells you on which pages their products are shown. Just to the right there's a tricky thumb index of 18 main categories; by bending Radio Shack's '55 catalog so the right hand page margins are exposed, these 18 categories line up with 18 little black blocks that enable you to turn *directly* to the category you want. All this without a magnifying glass, compass, or Baedeker.

On every page inside there's a 3" x 5/16" reverse panel telling you what's on the page; and beside each panel, in type ¼" high, is the page number. Hence by merely riffling the top corners of the catalog you can tell what's where — just like a deck of cards. Add to this a three page product index *1500 entries long*, and you know why Rand & McNally think we're on the proverbial!

Just one more thing: we don't have microphones listed in seven different hiding places, or anti-static powder spilled all over insulated shaft extensions. Components and equipment are 98.648% separated. And nowhere, amigos, have we claimed that our deliveries are faster than *immediate* because hyperbole isn't listed in Radio Shack's index.

As you glance over the ±10% 30,000 entries you may look down your nose and grunt "They can't *possibly* have all this stuff in stock."

D'ja ever look down your nose and grunt? It's not the easiest thing in the world. Nor is it a cinch to "stock all this stuff" — but we *do*, doggone it, and if you care enough to come to Beantown we'll show it to you, and the bills as well.

Every year our accountants tell us we're crazy to give away for free a 224-page book that costs us so much per copy to produce and mail that we have special ulcers reserved for catalog-thoughts only. "You oughta charge 'em a quarter," say our CPA's, "and get back at least a fraction of the cost."

Maybe we *are* a trifle balmy to hand out these expensive compendiums gratis?

Tell you what: write today for your free copy of Radio Shack's new 1955 catalog. It'll be sent to you by return mail.

And if you're *terribly* worried about getting such a valuable book "for no", send along a mail order for a ¼-watt resistor, a kilowatt swinging link or similar token of your regard for our extravagance. We are now 32 years of age and well able to survive the shock!

THE RADIO AMATEURS AT RADIO SHACK

RADIO SHACK CORPORATION

167 WASHINGTON ST., BOSTON 8, MASS.

Rush me free Catalog-55A.

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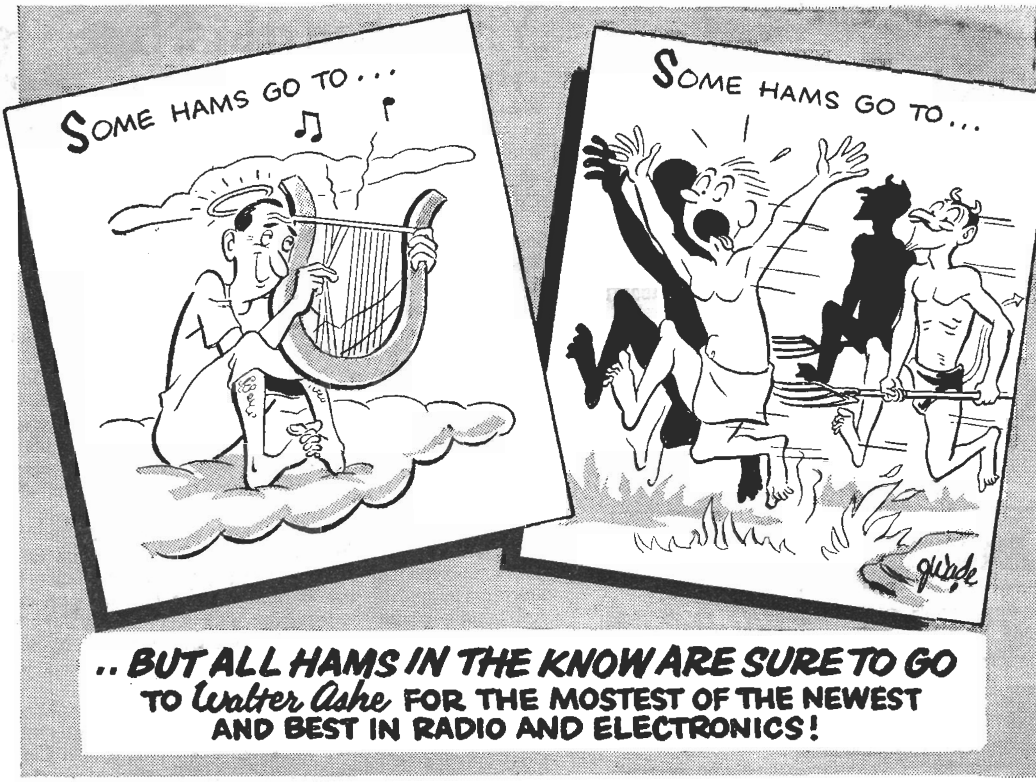

Whether you're a beginner, just getting a start in ham radio or an old timer who knows all the answers, you'll be way ahead if you depend on Walter Ashe . . . the centrally located, dependable source of supply, known from coast to coast as the house of "Surprise" trade-ins. Our big four story building is stocked with everything in radio, electronics, parts and supplies for every conceivable need. Novices and hams buy from us because they know we have what they want. Thousands of satisfied customers can't be wrong. So buy where the hams buy . . . at Walter Ashe.

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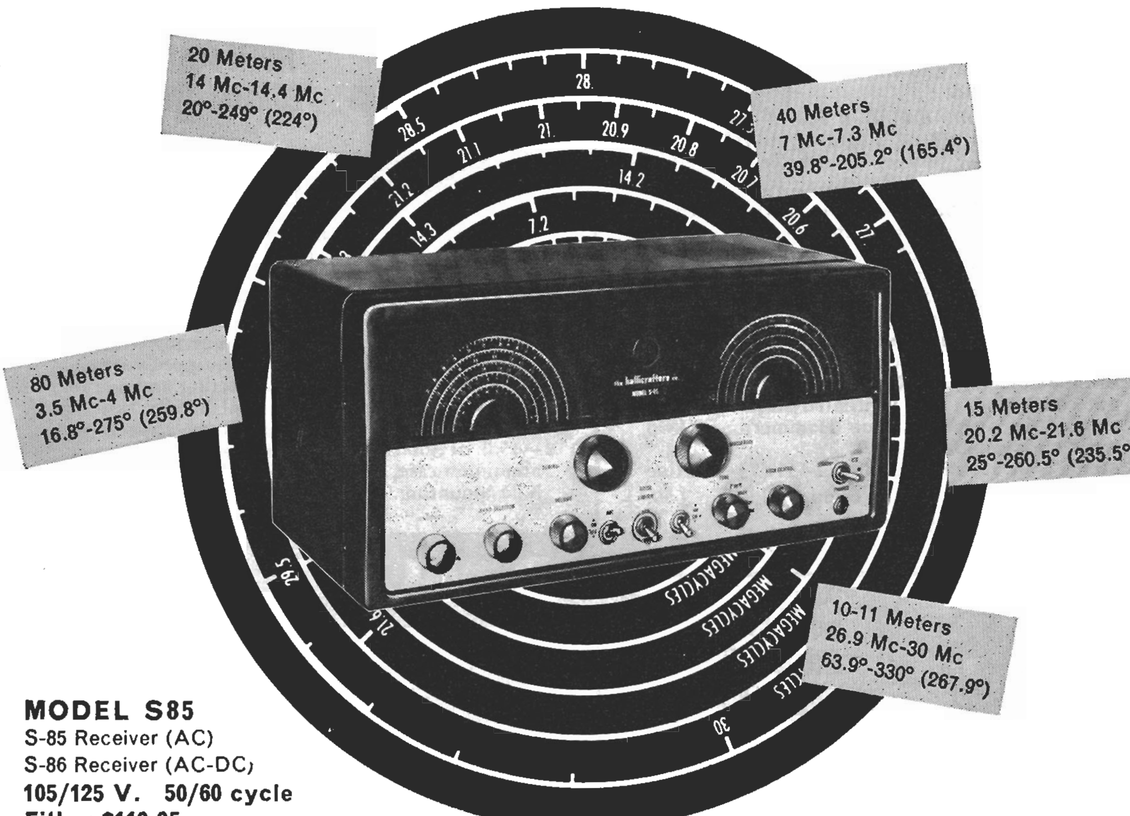

MODEL S85
S-85 Receiver (AC)
S-86 Receiver (AC-DC)
105/125 V. 50/60 cycle
Either $119.95

We know you expect the finest from Hallicrafters and you get it with this exciting new receiver. Here at the "World's Largest Hamshack," we are proud of our new models including the S-85 with the 10, 11, 15, 20, 40 and 80 meter amateur bands calibrated over 1000° for superior selectivity. Broadcast band 540-1680 kc and three short-wave bands 1680 kc—34 Mc are displayed on a large easy-to-read dial. Separate bandspread tuning condenser, seven tubes plus rectifier—there's so much more about this new receiver, why not see your Hallicrafters jobber or mail coupon below for complete specification sheet.

Bill Halligan, Jr.

Mail coupon today:
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Specification Sheets ☐ S-85 (S-86) ☐ S-95 (S-94) ☐ S-38D

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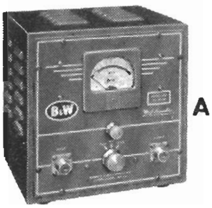


MATCHMASTER

Models 650 and 651

A Dummy Load, R-F Watt Meter, SWR Bridge, All in One

Here's the instrument you asked for. And once you've tried it, you'll wonder how you ever got along without it. It provides, in one completely self-contained cabinet, 6" x 8" x 8",—

A Dummy Load—for all kinds of tests on your transmitter without putting a signal on the air. Maximum SWR 1 to 1.2 over a frequency range of 300kc to 30mc.

A Direct-Reading R-F Watt Meter—for precise adjustments of all r-f stages up to 125 watts, and even higher powers by sampling. Excellent repeat accuracy over full 125 watt scale.

Integral SWR Bridge—for matching antennas and other loads to your transmitter, giving you precise adjustment of beam antennas, antenna tuning networks, and mobile whip antennas.

Controls — including a 3-position function switch, and a meter adjusting knob—are conveniently grouped on the attractive, silk-screen-gray front panel, which also contains a 3-inch calibrated meter, and Type SO239 input and output connections. The ventilated steel cabinet is finished in attractive blue Hammertone. Two types are available;

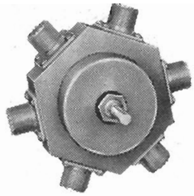
Model 650: 52-ohm line–Model 651: 73-ohm line

For details, write for descriptive Bulletin 650.

AUDIO PHASE SHIFT NETWORK

Type 2Q4—Model 350

This octal based, audio phase shift network provides a constant 90° phase shift, ± 1.5°, over the audio range of 300 to 3000 cycles, yet requires no more space than a 6J5 tube. Designed especially for single sideband receiving and transmitting applications.



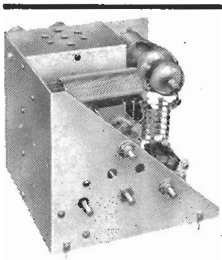
MULTI-POSITION COAXIAL SWITCH

Model 550

Takes The Mess Out of Switching Circuits

At last you can have an inexpensive, multi-position coaxial type switch—for selecting antennas . . . transmitters . . . exciters . . . receivers . . . and other r-f generating devices using 52-75 ohm coaxial line—without fumbling or breaking your back trying to screw and unscrew connections. This B&W Model 550 is equipped with six SO239 type connections for selecting any one of five 52 or 75 ohm lines. It will handle 1kw of modulated power with a maximum crosstalk of —45db at 30mc. Housed in a heavy, 4" diameter aluminum case, the Model 550 is made for single hole mounting.

ON THE

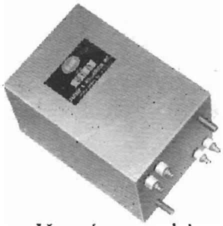


MULTI-BAND FREQUENCY MULTIPLIER

Model 504C

Gives You Any Band At The Flip of a Switch

Here is a newly conceived and designed exciter unit that makes transmission on any band available at the flip of a switch. Compact in its 8" x 7" x 9½" size, the Model 504C covers the 80 through 10 meter bands with a nominal power output of 25 watts from the 807 amplifier stage through a flexible pi-network output circuit. Its broad band type amplifiers require no tuning, and the unit comes equipped with four 6AQ5's that make up its multiplier string. An external VFO or crystal oscillator (80 meter fundamental) is required, as well as a suitable power supply. Sturdily constructed of heavy gauge frosted aluminum, the Model 504C also makes an ideal basic mobile foundation unit for multi-band operation.



PRECISION TOROIDAL TYPE SSB BANDPASS FILTER

Model 360 and 361

Here is a precision bandpass filter valuable for use in heterodyne type sideband generation. Containing eight toroidal type coils in an LC type filter, it is designed to pass the frequencies 16.9 to 20kc. Extreme skirt attenuation. Two types are available: a receiving type (Model 360) for 20,000 ohm input and output; and a universal transmitting or receiving type (Model 361), for 20,000 ohms input and an output of 20,000 ohms unbalanced, plus two 500 ohm balanced outputs. Both types are precision adjusted and housed in hermetically sealed, tinned steel cases measuring 2⅝" x 2¼" x 3¾", exclusive of mounting studs and terminals. Write for Bulletin 360.

BARKER & WILLIAMSON,

AIR WITH B&W

SINGLE SIDEBAND GENERATOR — Model 51SB

For Use With B&W Model 5100 Transmitter

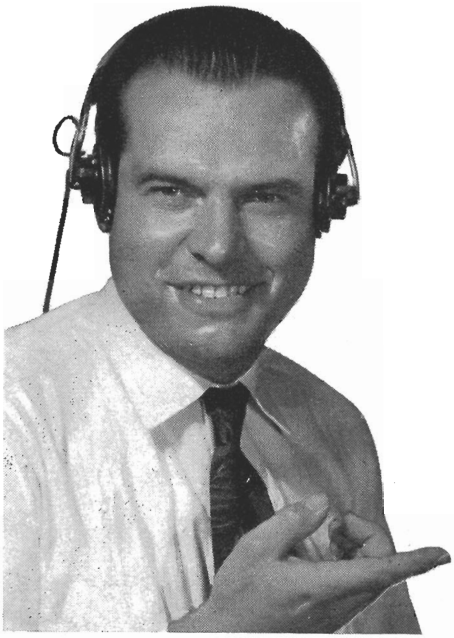
Now, for the first time, you can get really sparkling performance on either SSB, AM phone, or CW. This B&W Single Sideband Generator teamed up with the famous Model 5100 Transmitter gives you outstanding SSB operation on all frequencies provided in the 5100. Tuning and operation are a breeze. No test equipment is required. Single sideband signal is generated by a simple and efficient method perfected after two years of extensive research and testing by B&W engineers. No stone has been left unturned to give you such extras as voice operated and push-to-talk controls, a speaker deactivating circuit, TVI suppression, and unitized construction for quick and easy removal of any major section. Completely self-contained, the 51SB requires no more external accessories than a microphone.

Combine this Single Sideband Generator with the features of your Model 5100—150 watts peak envelope power input (100 watts peak envelope power output) on SSB, 150 watts on CW, 135 watts on AM phone; VFO or crystal operation; pi-network final—and you've got a combination that will flutter the heart of the most critical operator. The 51SB cabinet is made to bolt right onto the 5100 cabinet, extending the 22-inch length to 32 inches. Distinctive panel styling and appointments are the same for both. Easy to install, the 51SB comes factory wired and tested, complete with tubes and all necessary components to convert your Model 5100 Transmitter to SSB. This combination provides a superlative driver for *any* hi-powered linear amplifier.

Write for Bulletin

Inc. 237 Fairfield Avenue
Upper Darby, Pa.

These are just a few of the hundreds of products especially designed and built by B&W to meet the needs of the radio amateur. Others are described in Catalog 2PC available upon request. Write for your copy.



Dale Answers the Questions on Single Sideband

by Bill Cummings

W1RMG

Q: Why go to Single Sideband?

A: Here are 8 good reasons—the big advantages that have made SSB one of the most talked-about developments in years . . .

1. With Single Sideband, your final amplifier can provide up to 9 db gain, or a power increase of 8 times, over conventional double sideband AM.
2. No high power modulator and modulator power supply required.
3. SSB eliminates the heterodyning carriers that plague the overcrowded phone bands.
4. Less spectrum space occupied by phone signals using SSB.
5. Take full advantage of SSB technique by switching sidebands to avoid QRM.
6. Round-table operation of two independent QSOs on the same suppressed carrier frequency using opposite sidebands.
7. Distortion due to selective fading eliminated.
8. Harmonic TVI virtually eliminated through the use of linear amplifiers.

Q: Have they worked the bugs out of SSB?

A: Single Sideband is no laboratory curiosity. It's been on the air long enough to prove it delivers more consistent, stable performance than AM. The selection of rigs now available from such mainstays as Central Electronics, Hallicrafters, and Collins establishes SSB as standard equipment for any ham who wants to cut through the QRM even over long distance. My own 10-watter has worked all call areas from W1 thru WØ.

Q: How high is the price tag on SSB?

A: This is one of the big surprises in Single Sideband. SSB rigs that also work AM actually cost a fraction of the price you'd pay for an AM-only outfit.

Q: What is the best way to start?

A: If you are past the pro and con stage of SSB vs. AM, the way to make the switch is to drop me a note at Dale and I'll send you all the details about available equipment, prices, trade-in allowances and terms. Here at Dale, we've been riding SSB ever since the first models came off the breadboards. So we can help you choose the right job and get it set for the best results. What's more, we have all the models on hand for immediate delivery. When you write, list your present equipment and I'll quote you a trade-in allowance at the same time. Now, if you still have any questions about SSB, send them in. I'll be glad to give you the answers based on our long experience.

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MULTIPHASE EQUIPMENT

and for good reason. It's versatile, permits all-band operation 10 thru 160, it's extremely stable and it's a well engineered, dependable piece of communications equipment.

MODEL 20A

MULTIPHASE EQUIPMENT is the overwhelming choice of SSB OPS everywhere. Ask any ham who uses it! Listen to it perform on SSB, AM, PM or CW!

MODEL 20A

- **20 Watts Peak Output SSB, AM, PM and CW**
- **Completely Bandswitched 160 thru 10 Meters**
- **Magic Eye Carrier Null and Peak Modulation Indicator**

Choice of grey table model, grey or black wrinkle finish rack model.

Wired and tested............**$249.50**
Complete kit..............**$199.50**

SIDEBAND SLICER
MODEL A IMPROVES ANY RECEIVER

Upper or lower sideband reception of SSB, AM, PM and CW at the flip of a switch. Cuts QRM in half. Exalted carrier method eliminates distortion caused by selective fading. Easily connected into any receiver having 450-500 KC IF. Built-in power supply. Reduces or eliminates interference from 15 KC TV receiver sweep harmonics.

Wired and tested............**$74.50**
Complete kit................**$49.50**

Check These Features
NOW IN BOTH MODELS

- **Perfected Voice-Controlled Break-in on SSB, AM, PM.**
- **Upper or Lower Sideband** at the flip of a switch.
- **New Carrier Level Control.** Insert any amount of carrier without disturbing carrier suppression adjustments.
- **New Calibrate Circuit.** Simply talk yourself exactly on frequency as you set your VFO. Calibrate signal level adjustable from zero to full output.
- **New AF Input Jack.** For oscillator or phone patch.
- **CW Break-in Operation.**
- **New Gold Contact Voice Control Relay.** Extra contacts for muting receiver, operating relays, etc.
- **Accessory Power Socket.** Furnishes blocking bias for linear amplifier and voltage for optional VFO (Modified BC458 makes an excellent multiband VFO.)
- **40 DB or More Suppression** of unwanted sideband.

MODEL 10B
SUCCESSOR TO THE POPULAR MODEL 10A

- **10 Watts Peak Output SSB, AM, PM and CW**
- **Multiband Operation** using plug-in coils.

Choice of grey table model, grey or black wrinkle finish rack model. With coils for one band.

Wired and tested............**$179.50**
Complete kit..............**$129.50**

QT-1 ANTI-TRIP UNIT

Perfected Voice Operated Break-in with loudspeaker. Prevents loud signals, heterodynes and static from tripping the voice break-in circuit. All electronic — no relays. Plugs into socket inside 20A or 10B Exciter.

Wired and tested, with tube....**$12.50**

AP-1 ADAPTER

Plug-in IF stage — used with Slicer, allows receiver to be switched back to normal.

Wired and tested, with tube.....**$8.50**

PS-1 NETWORK

Plug-in prealigned 90° phase shift network and socket available separately for use with GE Signal Slicer and SSB Jr.........................**$8.95**

WRITE FOR LITERATURE

MULTIPHASE EQUIPMENT

Central Electronics, Inc.

1247 W. Belmont Ave. Chicago 13, Illinois

Watch For Early Announcement Of The New DeLuxe MULTIPHASE VFO.

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HAM HEADQUARTERS SINCE 1925 - HARRISON HAS IT!

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W2AVA

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Nine floors crammed full with the latest and finest in amateur equipment and parts. We have EVERYTHING for the novice, old-timer or short wave listener.

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Large scale purchasing and efficent business operation enable us to offer everything at lowest prices.

NO ONE CAN UNDERSELL HARRISON!

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"Harrison Has It!" is more than a slogan. It is a promise, and your assurance, of prompt delivery from stocks adequately maintained at all times.

4 *Trade-ins* ...

The best deal anywhere! Harrison is always eager to match or top any offer! And you will get better service too! Come on in with your old gear - you'll go home happy!

5 *Guaranteed Satisfaction* ...

Every transaction is backed by my 30 year reputation for fair dealing and customer satisfaction. You can always buy with full confidence - at Harrison!

6 *Free Parking* ...

When in N. Y., park free at ANY METER, ANY GARAGE, or ANY LOT. Harrison will pay for your parking while making any purchase over $10.00

Bil is Glad to Help You! ...

Whether you're a newcomer just starting, or an old timer in the game, I'm always glad to lend a helping hand.

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Better yet, visit either of our stores (New York City or Jamaica) for personal attention. Lots of active hams to help you with your problems.

Always Remember ...

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MAKE HARRISON'S YOUR NYC HQ!!!

Relax in Air Conditioned Comfort — Have a Coke — See What's New — Meet the Gang! You're Always Welcome.

73, Bil, W2AVA

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Each edition is chock-full of the newest gear, money-saving bargains, ham hints, articles, etc. A postcard puts your name on our mailing list ... FREE! Send it today!

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TYPE 2XP

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1600 to 12000 Kc.
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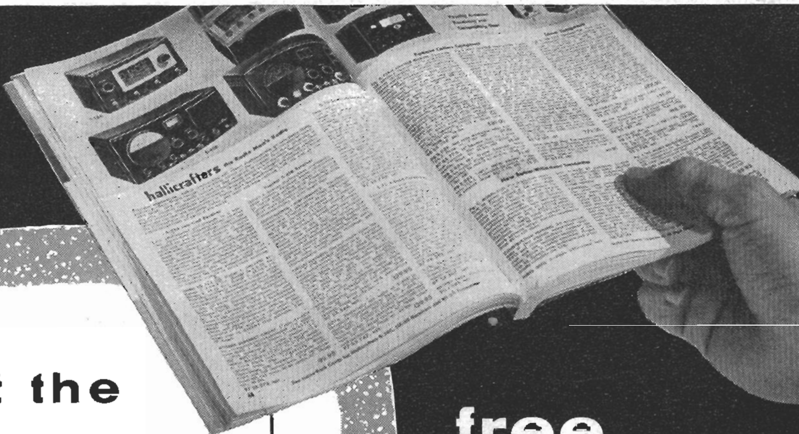
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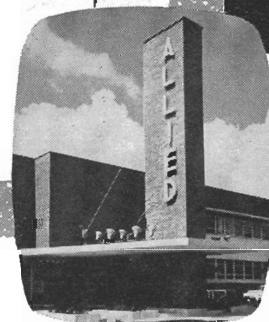
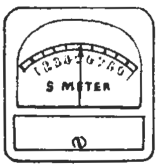

ALLIED

A good microphone can improve your results as much as a high gain antenna



Ever notice that two signals of the same "S meter" intensity sound differently? One is muddy, dull, a little hard to read—the sibilant letters like S and F almost alike. The other signal is sharp, clean and readable even in QRM and QRN—because there's usable intelligence. No mistake about the call or comments.

The greatest variation is in the microphone. A sharp peak adds no intelligibility but limits the modulation to that value. A peak of, say 6 db, which is usual in many ordinary microphones, will reduce voice power by HALF. Don't be fooled by a microphone that sounds "louder"—loudness by itself is not a criterion of performance; quite the contrary since it may indicate undesirable peaks.

An E-V microphone with smooth, peak-free response, replacing an inferior instrument, often will do more for a phone signal than a new antenna or increased power. As a further *plus*, of course, you get well-known E-V durability, style and performance. An E-V microphone, to raise stations, to carry through a QSO, is your best station investment.

Shown above are a few of the E-V microphones designed for effective communications. Amateur discount applies.

(upper left) **Model 611** high output dynamic and **Model 911** crystal. On-Off switch. **List from $25.50 to $37.50**

(upper right) **Model 950** Cardax high-level crystal cardioid, with dual frequency response. On-Off switch. **List, $42.50**

(lower left) **Model 630** wide range, high output dynamic, with exclusive Acoustalloy diaphragm. On-Off switch. **List, $47.00**

(center) **Model 636** "Slimair" wide range dynamic. Pop-proof head. Acoustalloy diaphragm. On-Off switch optional. **List, $70.00**

(lower right) **Model 623** slim-type high output dynamic, with E-V Acoustalloy diaphragm. On-Off switch. **List, $49.50.** Also **Model 926** crystal, less switch and connector. **List, $24.50**

(Other E-V microphones for mobile and aircraft communications, telecasting, broadcasting, recording, and public address.)

For further information,
see your E-V Distributor
or write for
Condensed Catalog No. **119**

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192

94

A Complete Amateur Station

IN *only* ONE CUBIC FOOT *with*

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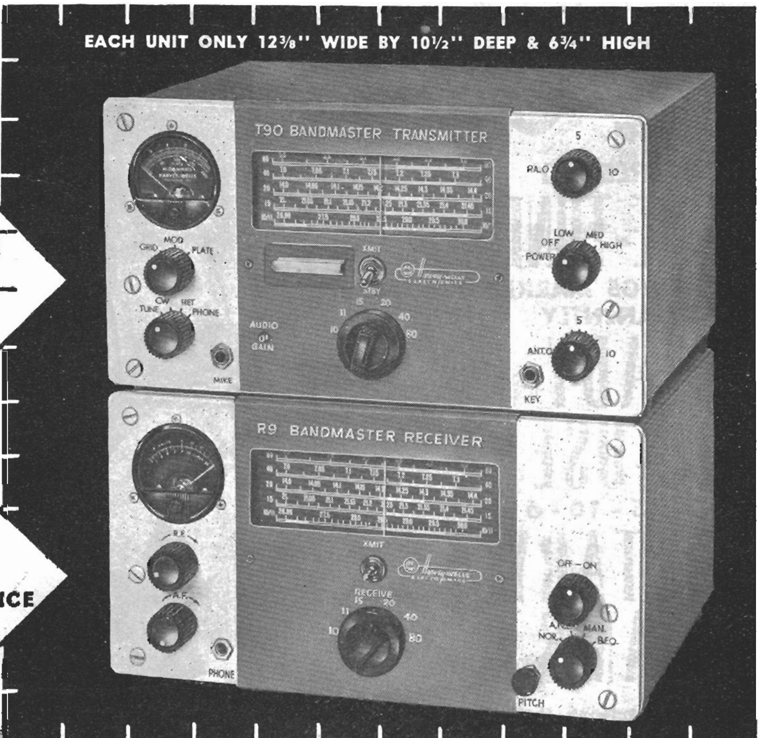
6 AMATEUR BANDS
BAND-SWITCHING
Factory built & tested
(NOT A KIT)



T-90
90 WATT TRANSMITTER
The midget with a MIGHTY PUNCH
$179.50*
Complete with tubes less power supply (NOT A KIT)

R-9
9 TUBE RECEIVER
DOUBLE CONVERSION
PACKED WITH PERFORMANCE
$149.50*
For 115 V. AC operation

TRANSMITTER FEATURES

1. TVI Suppressed
2. Complete band-switching; no plug-in coils
3. Complete Break-in Key or keying of multiplier stages only
4. VFO Tuning without carrier on
5. Cathode biased Exciter tubes and clamp tube control of Final Amplifier Screen Voltage
6. Initial tuning at reduced power
7. Three position excitation control
8. Antenna loading flexibility
9. Selector switch allows metering of PA Grid, PA Cathode and Modulator currents
10. Remote Break-in and Receiver muting provided by relay control
11. VFO voltage regulated and temperature compensated
12. Illuminated dial and meter
13. Crystal door on front panel
14. Filament Operation 6 or 12 volts AC/DC
15. Low average Modulator current
16. Built-in provision for either Carbon, Crystal or Dynamic microphone and push-to-talk

RECEIVER FEATURES

1. Double conversion on all bands
2. Three tuned circuits on each band, in R.F. section
3. All coils slug tuned, giving high "Q" circuits
4. Separate oscillator coils for each band (no spurious response)
5. Bandwidth: Four kilocycles wide at the 6 db point
6. Complete with tubes and built-in AC power supply. 6 or 12 volt DC power supply available.
7. Crystal filtering optional.
8. Approximately 6" of dial spread on all bands. Accurately calibrated
9. Rigid Steel construction. (Vibration-Proof)
10. 6¾" hgt. enables easy under dash mounting for mobile installation

Send for technical Bulletins **HW 456 *and* HW 457**

*Prices subject to change without notice.

Harvey-WELLS ELECTRONICS, INC., SOUTHBRIDGE, MASS.

196

TAKE A TIP *from* WRL

THE WORLD'S LARGEST DISTRIBUTOR OF AMATEUR RADIO EQUIPMENT

Leo I. Meyerson
WØGFQ

All of us at World Radio Laboratories want to make it as easy and as pleasant as possible for you to do business with us! So we do our own financing, allowing you the use of new equipment for only 10% down payment. We also offer our special "10 Day Free Trial" Plan, so that you may return, prepaid, within 10 days, any of our rigs for full refund or credit! Won't you give us a chance to offer you our "personalized service"?

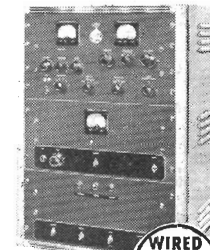
FOR THE FINEST IN AMATEUR RADIO AND ELECTRONIC EQUIPMENT!

Here's More Watts per Dollar!

The 500 WATT, BANDSWITCHING GLOBE KING!

COMPARE THESE FEATURES WITH ANY OTHER RIG!

- ★ 500 WATTS INPUT ON FONE AND CW.
- ★ COMPLETE BANDSWITCHING, 10-160M.
- ★ PROVISIONS FOR VFO AND SINGLE SIDE-BAND INPUT AND OPERATION.
- ★ PI NETWORK MATCHES ANY ANTENNA FROM 52-600 OHMS.
- ★ THOROUGHLY SCREENED AND BY-PASSED FOR TVI.



WIRED $675.00 CASH

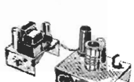
ONLY **\$36.78** PER MO.

$67.50 CASH DOWN

BARGAIN BUYS FOR BEGINNERS!

COMPLETE AMATEUR RADIO KITS!

RECEIVER
$14.70 COMPLETE
Stock No. 75A008



TRANSMITTER
$29.40 COMPLETE
Stock No. 81A052

Build your own transmitter and receiver and "go on the air" with this complete kit and simple instructions. Complete kit only $43.22. Designed for single wire antenna. Power supply included with transmitter. Units may be purchased separately. Write for full details.

BEST BUY IN BEAMS!

WRL's new line of sturdy "Plumber's Delight" beams. 1¼", hot dipped galvanized steel boom; all elements aluminum; 1⅛" center pieces; 1" end inserts. Wooden dowelled at all stress points. 20M beam has double 1¼" hot dipped galvanized steel booms for extra sturdiness. Specify choice of T or Gamma Match. All beams three element.

10M $18.95 15M $27.95
20M $49.95

Combination & short element beams to be announced!

Write For Detailed Specification Sheets TODAY!

65 WATT BANDSWITCHING GLOBE SCOUT

KIT $89.95 CASH — WIRED $99.95 CASH

KIT: **ONLY $7.15 per Mo.**
$9.00 CASH DOWN

WIRED: **ONLY $7.95/Mo.**
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Ideal Xmtr. for novice. Fully bandswitching, 10-160-M. 65 watts CW; 50 watts on fone. Metering provided. Pi Network antenna tuner. 100% modulation of Final. TVI screened cabinet, 8" x 16" x 8". Send for complete information today!

RECONDITIONED EQUIPMENT? *You Bet!*

SAVE UP TO 50% ON OVER 600 ITEMS IN OUR INVENTORY — EACH CARRYING A 90 DAY, FACTORY-NEW GUARANTEE ON ALL PARTS AND WORKMANSHIP. OR YOU MAY RECEIVE 100% TRADE-IN VALUE ON NEW EQUIPMENT WITHIN 90 DAYS OF ORIGINAL PURCHASE.

EVERYTHING IN ELECTRONICS!
SEND FOR THE **FREE** 1955 WRL CATALOG!
OVER 15,000 TOP-VALUE ITEMS

ARRL-4

Please send me your New Free Catalog and full information on the items checked below! **Quote your top Trade-In offer for my:** ______________________

(Name and Make of Equipment)

ALL PRICES SUBJECT TO CHANGE WITHOUT NOTICE.

WORLD'S MOST PERSONALIZED RADIO SUPPLY HOUSE

3415 W. BROADWAY, CO. BLUFFS, IA., Phone 2-0277

for your ______________________
(New Equipment Desired)

☐ GLOBE KING ☐ NEW WRL BEAMS ☐ GLOBE SCOUT ☐ 3'x4' RADIO MAP (25c) ☐ RECONDITIONED EQPT. LIST ☐ AMATEUR RADIO KITS.

NAME ______________________
ADDRESS ______________________
CITY & STATE ______________________

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From

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W7OVW W7DDQ W7RGD
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The only thing to come out of *this* cloud will be a *deluge of service.*

AROUND the edges of this ad appears just a few of the many lines we stock. We have the products for your particular use whether it be amateur, industrial, broadcast, communications, school, hi-fi or sound. A post card, letter or call from you will bring an instant reply or SAME DAY shipment. We welcome all requests for information or catalogs and, of course, your order for parts or equipment.

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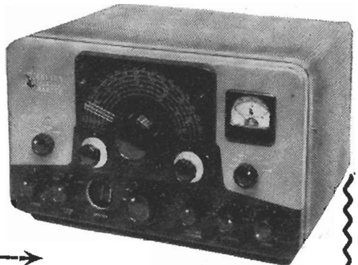
THE Johnson Ranger is just one of the many pieces of equipment upon which we can make a generour trade-in allowance. We have a plan which will realize you more than a regular trade-in. It will be worth your time to look into this.



HALLICRAFTER SX-88

THIS Hallicrafter SX-88 is another good example of the products we can offer to you on BUDGET TERMS. We will gladly answer any inquiries on our time payment plan.

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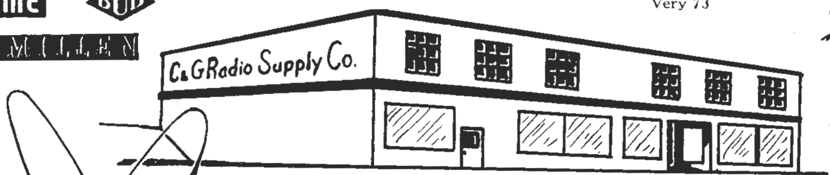
A LAST WORD to you out-of-town hams and customers. Stop in and see us during your vacation to the Pacific Northwest. A personal meeting would be greatly appreciated by the gang here at the store. May we hear from you?

Very 73



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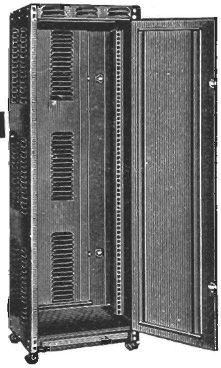
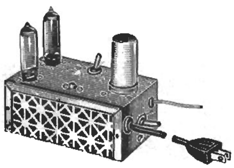
Only BUD has so many PRODUCTS TO FILL YOUR NEEDS

HERE ARE THE DIFFERENCES THAT MAKE BUD PRODUCTS THE BEST CHOICE FOR YOUR RIG



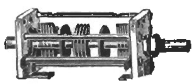
TWO-DOOR RACKS

THE ONLY REASONABLY PRICED 2-DOOR RACK WITH ADJUSTABLE PANEL MOUNTING SUPPORTS. Rear door provides access to equipment behind panel. Front door allows dials, knobs and other projections in front of panel. Choice of 3 finishes and 4 sizes.



FREQUENCY CALIBRATOR FCC 90-A

THE ONLY SELF-POWERED MODEL. Has 100 kc crystal, and will give check points on all bands to 30 mc. Uses two tubes, plugs into 110V receptacle.



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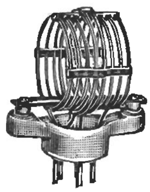
ONLY STANDARD CONDENSER THIS SMALL. DESIGNED FOR MOBILE USE. Designed particularly for high frequency use in converters, pre-selectors and receivers in amateur, T.V. and F.M. bands. Well constructed with soldered brass plates and ceramic brackets.

CODE PRACTICE OSCILLATOR AND MONITOR CPO 128-A

THE ONLY OSCILLATOR WITH BUILT-IN MONITOR WHERE NO MODIFICATION IS NEEDED TO CHANGE FROM OSCILLATOR TO MONITOR AND BACK AGAIN. It has 2 tubes and built-in 4" dynamic speaker. Operates on 110V AC or DC. Available in earphone model CPO 130-A.



75 WATT COIL

THE ONLY TRANSMITTING COIL WITH POLYSTYRENE BASE. Resists breaking or cracking; has high Q. Pins molded in place. Always in perfect alignment. Available with fixed and adjustable center links or end links

SLOPING PANEL CABINETS

THE ONLY ONE OF ITS TYPE WITH HINGED TOP FOR EASY ACCESSIBILITY TO COMPONENTS MOUNTED WITHIN. Front is also removable. Sturdily built and handsomely finished. Available in five popular sizes.

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Perfectly Balanced

to supply the finest equipment in ham radio

Two names known and trusted the world over for quality equipment and dependable service.

HT-20 Transmitter

Bandswitching transmitter that practically eliminates "TVI" interference. Covers 160-10 meters. AM power output 100 watts; CW—115 watts. Uses bandswitching exciter and continuous-tuning final amplifier. Control leads filtered thoroughly. Pi-network tank circuit is followed by 4-section low-pass filter; dial and meter scales are covered with metal screening to eliminate spurious radiation. 10-position crystal selector switch has additional positions for external VFO. Single meter with selector switch checks all stages and provides output tuning indication. Has built-in blower for cooling. Size: 20½x11¾x16¾". For 105-120 volts. 50-60 cycles AC. Wt., 130 lbs.

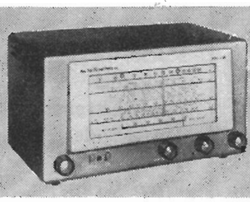
98F022. NET.. **449.95**

SX-88 Dual Conversion Receiver

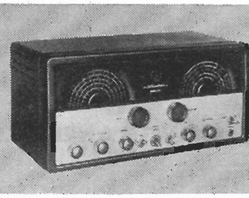
Finest amateur communications receiver ever built by Hallicrafters. Covers 535 kc to 33.3 mc in 6 bands. Calibrated electrical band-spread. First IF is 1550 kc on band 2, and 2075 kc on all other bands. Second converter is crystal controlled. Second IF 50 kc. 50 kc BFO has buffer amplifier for greater stability. Has 50 kc output jack via cathode follower for teletype converter, oscilloscope, etc. IF selectivity variable from 10 kc to 0.25 kc at —6 db. Antenna impedance—52-600 ohms. Audio output rated at 10 watts. Output impedances for speaker or line, 3.2/8/500-600 ohms. Audio response: 20 to 10,000 cycles. For 105-125v, 50/60 cyc. AC. or batteries. Less speaker. Size: 20x10⅜x18¼". Standard 8¾"x19" panel for rack mounting. Shpg. wt. 85 lbs.

98F025. NET.. **595.00**

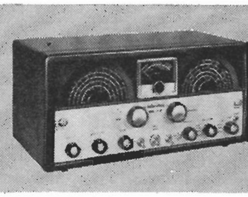
Model R-46A. Matching speaker for above.
98F002. NET.. **19.95**



S-38D All-Wave Receiver. Famous low-cost "Radio Man's Radio". Covers 540 kc to 32 mc in 4 bands. Electrical band-spread dial with scale for separating crowded bands. 5" PM speaker and headphone tip jacks. Wt. 14 lbs.
98F033. NET.................. **49.95**



S-85 4-Band Receiver. Covers 540 kc to 34 mc in 4 bands. Electrical band-spread. High signal-to-noise ratio. Includes ANL, AVC, BFO, 1-RF and 2-IF stages. 3-position tone control. 5" PM speaker. Wt. 29 lbs
98F031. NET............... **119.95**



SX-99 4-Band Receiver. Compact lightweight receiver. Covers 540 kc to 34 mc in four bands.. "S" meter, crystal filter, calibrated electrical bandspread. Large easy-to-read dial. ANL, AVC Less speaker. (Uses R46A above). Wt., 36 lbs.
98F026. NET............... **149.95**

F.O.B. Chicago, Include Shipping Charges and Insurance

Send for **FREE CATALOG**

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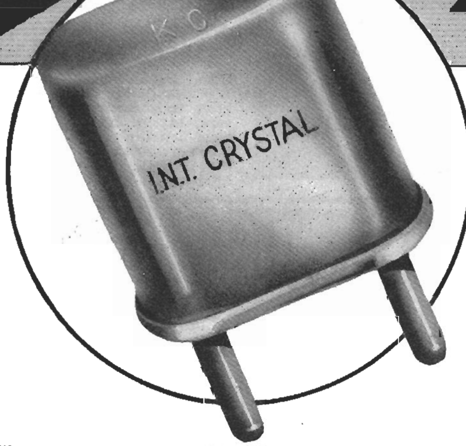
NEWARK ELECTRIC COMPANY

**223 West Madison St.
Chicago 6, Illinois**

200

Amateurs and Experimenters!

ONE DAY SERVICE!
2000 KC to 54 MC



ONE-DAY Processing

Orders for less than five crystals will be processed and shipped in one day. Orders received on Monday thru Thursday will be shipped the day following receipt of the order. Orders received on Friday will be shipped the following Monday.

International TYPE FA-9
(fits same socket as FT-243)

Pin Spacing .486 Pin Diameter .093

RANGE (kc)	TOLERANCE	PRICE
Fundamental Crystals		
2000-9999	.01%	$2.80
10000-15000	.01%	$3.90
Overtone Crystals (For 3rd overtone operation)		
15 MC- 29.99 MC	.01%	$2.80
30 MC- 54 MC	.01%	$3.90

SPOT FREQUENCY

.01% TOLERANCE—Crystals are all of the plated, hermetically sealed type and calibrated to .01% or better of the specified frequency when operated into a 32 mmf load capacitance.

HOW TO ORDER

In order to give the fastest possible service, crystals are sold direct and are not handled by any jobber. Where cash accompanies the order, International will prepay the Air Mail postage; otherwise, shipment will be made C.O.D. Specify your exact frequency and the crystal will be calibrated to .01% or better of this frequency with the unit operating into a 32 mmf load capacitance.

International Crystal Mfg. Co., Inc. DEPT. Y
18 North Lee
Oklahoma City, Okla.

Price

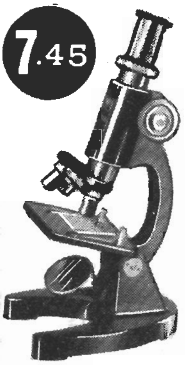
Please Send: ______Crystals Freq.________ ________
______Crystals Freq.________ ________
______Crystals Freq.________ ________
TOTAL $________

TO: Name:________________________
Address:________________________
City__________Zone:______State:__________

Enclosed: ☐ Check, ☐ Cash, ☐ M.O. for $______, or Ship C.O.D. ☐

International CRYSTAL Mfg. Co., Inc. 18 N. Lee Phone FO 5-1165 OKLAHOMA CITY, OKLA.

it's *Lafayette* for
YEAR ROUND BUYS at GREAT SAVINGS!



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3 LENS TURRET MICROSCOPE

100X-200X-300X

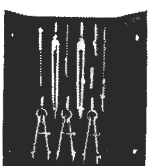
Precision built for accuracy and long lasting service. Triple position turret holds 3 achromatic color-corrected objectives for magnification of 100X, 200X and 300X power! Heavy base and stand—inclines through 90°. Dual knob focusing by rack and pinion. Adjustable plano sub-stage mirror. High-grade microoptics used throughout! Complete with fitted wooden carrying case and slides. Shpg. Wt. 4½ lbs.

F-10 ____________ NET 7.45

TIMER-SWITCH SALE

This "almost human" timer will switch your radio, television, or any other appliance on or off, fully automatically. Will turn them on at any desired pre-set time. Can be pre-set up to 12 hours. Full-vue crystal dial, 3 knob front, 3½" diameter. With mtg. ring and bracket.
MS-62, for 110V/60 Cy AC

Each 3.95



$2.75

PRECISION DRAFTING SET

11 Pieces—Fitted Felt-Lined Case

Made in Germany, of heavy brass, nickel plated and polished. Instruments include 5⅛" Compass with pencil and pen points and lengthening bar. 5⅛" Divider, three 3¾" sidewheel bow dividers with needle point, pen point and pencil point. 5" Ruling pen, extra handle for pen or pencil, capsule with extra leads. Interchangeability of these many parts makes this an exceptionally versatile set. Your money back—if this set is not worth twice our price!

F-13 ....Net 2.75

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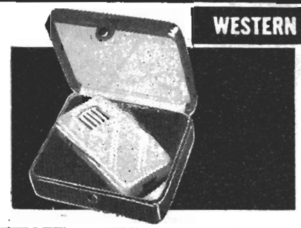
Greatest Tape Buy Ever!

1200 FT. REEL
Genuine Plastic Base
RECORDING TAPE

$1.89 per roll plus postage
C.O.D.'s accepted

LAFAYETTE made a terrific deal with one of the leading manufacturers of recording tape to supply us with their regular tape which sells for almost twice our price. **WE GUARANTEE ABSOLUTE SATISFACTION OR YOUR MONEY BACK.** The finest, professional-quality recording tape obtainable. Highest performance for thousands of playings. Red Oxide Base in a smooth, uniform coating; greater signal strength; with maximum fidelity; uniform frequency response from 40-15,000 cps.



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Reg. price ~~$185.00~~ *Our price* **$14.95**

Brand new, in original Western Electric's jeweler's case. Supplied with receiver, receiver cord, battery cord and plug (less batteries). Money back guarantee. Act now while they last! Uses Burgess XX30E and 8R batteries at $1.55 per set.

NEW POCKET AC-DC VOM MULTITESTER

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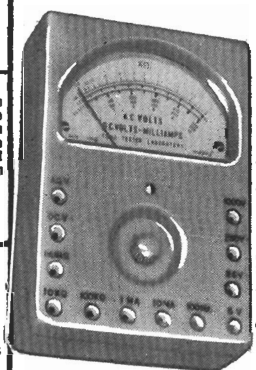

This instrument is one of the best buys that Lafayette has ever offered in a Wide Range **AC-DC MULTITESTER.** An ideal portable unit that meets the need for a compact, yet rugged test instrument. Has ease of operation usually **FOUND ONLY IN MORE EXPENSIVE INSTRUMENTS.** Has 1000 ohms/volt sensitivity on both AC or DC. Uses full 3" rectangular meter with large, easy to read scale. Uses 1% precision resistors, jeweled D'Arsonval microamp meter movement. Ranges: AC-DC and output volts 0-5, 0-25, 0-250, 0-1000V; DC current 0-1, 0-10, 0-100. Resistance 0-10K and 0-100K ohms. In handsome sturdy bakelite case. Size: 4¾" x 3¼" x 1⅜". Supplied Complete with test leads and batteries. A Must for every serviceman, shop, Laboratory or experimenter—and at Lafayette's Price you can afford to own one. Shpg. Wt. 2½ lbs.

MODEL RW-27C—Complete
In Lots of 3 **9.45**
Single, ea. **9.95**

Top Quality CRYSTAL MIKE

$5.75

- —52 db output level
- Range 30 - 10,000 cps
- 10 ft. cable and connector
- Handle and interlocking base

A really fine microphone by a manufacturer known the world over. Use for amateur broadcasting, public address and home recording. Can be used with floor stand. Shpg. wt. 2⅛ lbs.

PA-21 ....5.75

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BARGAIN CATALOG

Microscopes, Binoculars, Barometers, Compasses, Timers, Hearing Aids, Tools, Car Radios, Record Players, Tape Recorders, TV Boosters, Antennas, Intercoms, Hi Fidelity Equipment, P.A. Systems, All Radio, TV & HAM SUPPLIES. Check full of buys at great savings. Write today for your free copy.

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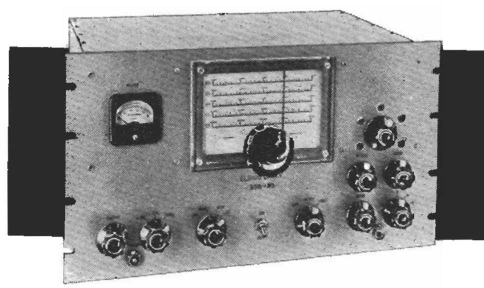
Include postage with order.

Eldico *announces* *Another* "FIRST"

the SSB 50 Transmitter/Exciter for SSB—AM—CW

with all-band VFO

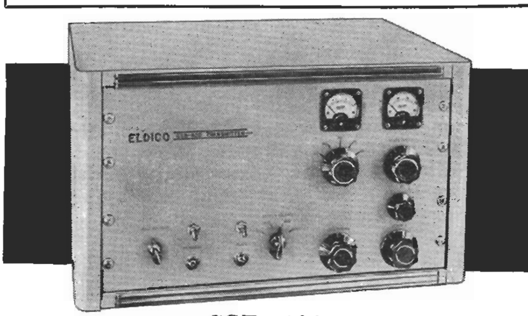
in production now!



SSB 50

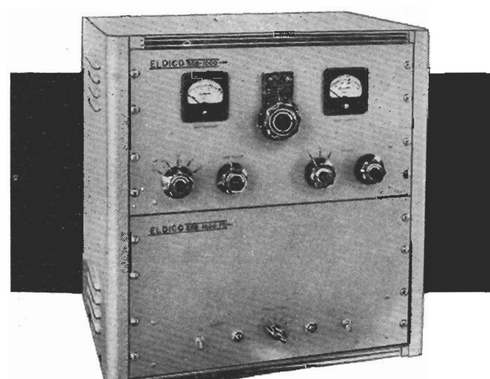
the ultimate in design and versatility — no extras needed — virtually eliminates TVI

- Instantaneous switching AM-SB1-SB2-CW.
- Smooth, simplified tuning — readjustment necessary only when changing bands.
- Single control bandswitching, 80 through 10 meters.
- Extremely stable, permeability tuned VFO, covering 200 KC per band.
- Frequency accuracy ± ½ KC on easily readable slide rule dial.
- 6146 output amplifier with pi-network output.
- "Anti-trip" voice break-in with speaker.
- Complete with power supply . . . shielded construction . . . 10½" front panel . . . cabinet 12" x 20" x 15".



SSB 400

400 W linear amplifier using 4E27 A



SSB 1000

1000 W linear amplifier using 4-250 A

featuring:

Bandswitching 80 through 10 Pi network output; well designed power supplies—adjustable for all classes

Competitively Priced

(See Your Distributor for other Eldico Products)

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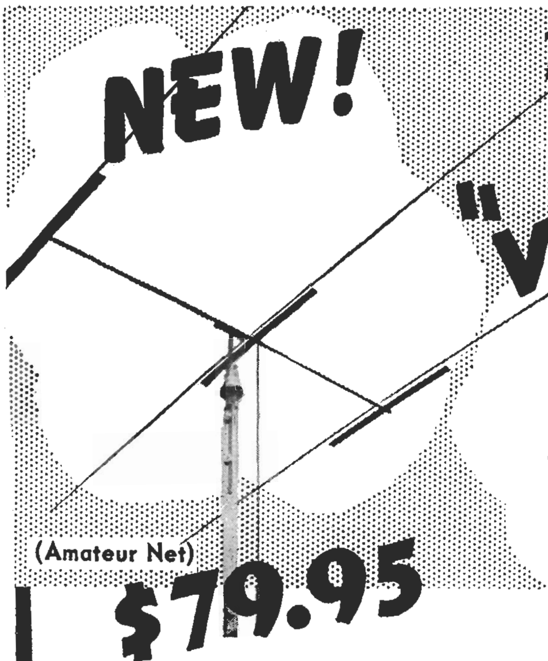

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(Amateur Net)

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Model VPA 20-3

(OTHER MODELS LISTED BELOW)

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- **Ham Proved Performance!**
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Here is the *Original Design* "Vest Pocket Beam" — the amazingly efficient miniaturized beam designed by WØVZC and WØQFG and described in May '54 QST.

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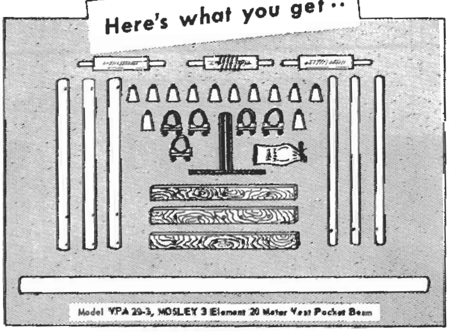

Model VPA 20-3, MOSLEY 3 Element 20 Meter Vest Pocket Beam

The MOSLEY V-P Beam is complete ... ready to assemble in 30 minutes, or less, and mount on your rotor.

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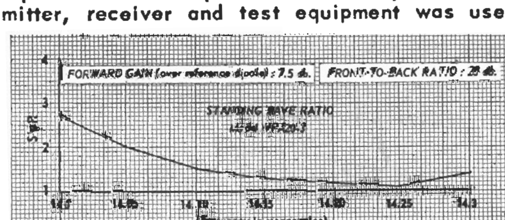
Typical Performance Data

Performance figures achieved with production model VPA 20-3, 3 element beam, in typical housetop installation. Popular commercially built transmitter, receiver and test equipment was used.



VPA 20-2, *2 Element 20 Meter Beam.*
Amateur Net $55.95

VPA 40-2, *2 Element 40 Meter Beam.*
(Price to be announced)
Write for full details

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8622 ST. CHARLES ROCK ROAD
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Dept. H-1

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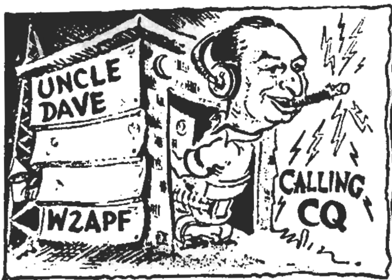

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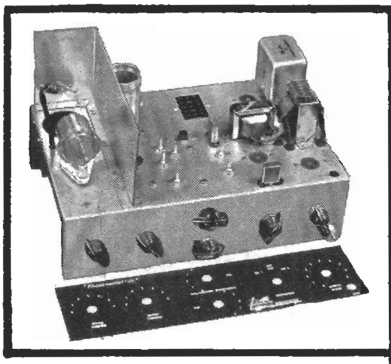

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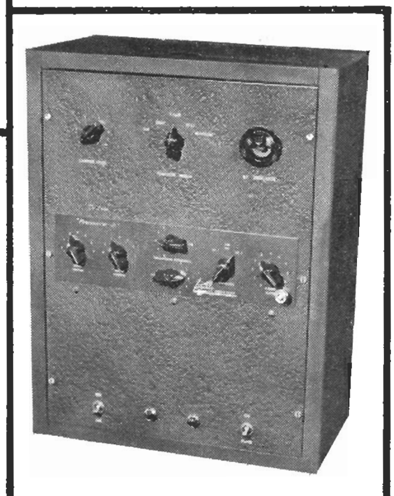

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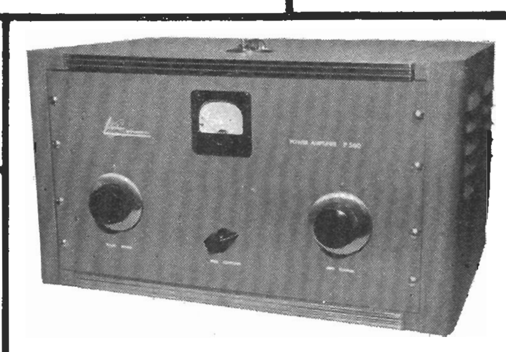

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