SINGLE SIDEBAND
for the Radio Amateur

A DIGEST OF
AUTHORITATIVE
ARTICLES ON
AMATEUR RADIO
SINGLE SIDEBAND

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE
SINGLE SIDEBAND for the RADIO AMATEUR
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 coordinated 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 suffice. 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.

West Hartford, Conn.

A. L. Budlong
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 W6NWF 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, WIDX

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

We will start with the initial statement that to understand ’phone you must first understand c.w. Practically everyone knows that an unmodulated carrier and a c.w. signal with the key held down are the same thing. Any way you tune them in on a receiver they act the same. On a panoramic receiver they look the same. Any test you can make of them at the receiving location will give the same result. They are the same. Furthermore, if they are stable they take up no room in the spectrum! Oh, sure, you tune in one or the other on your receiver, with the b.f.o. on, and you can hear it over several dial divisions. Turn your b.f.o. off and the S-meter on, and the signal gives a reading over a range of several kilocycles. But neither of these effects proves that the signal is broad — it only indicates that your receiver doesn’t have infinite selectivity. By definition, 14,200,000 and 14,200,010 kc. aren’t the same frequency, so they must be different. Actually, signals. Any single r.f. signal can be represented by an infinitely-thin vertical line on a plot of amplitude vs. frequency. Fig. 1 is such a representation, except that the draftsman couldn’t draw an infinitely-thin line that would show on the paper, and we had to settle for a finite-thickness line. The frequency can be read from the “Frequency” scale, and the amplitude from the “Amplitude” scale. The taller the line, the greater the amplitude. Don’t worry about the units — the frequency scale could be megacycles, or even cycles at some part of the spectrum. Your panoramic receiver would show such a picture if it had infinite selectivity. If your receiver had infinite selectivity, the S-meter would indicate the amplitude at one setting of the tuning knob as you tuned across the frequency range shown, and nothing at any other setting.

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

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.

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.

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 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 pass-band, 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 run
two carriers to 1000-cycle tone. The modulating signal might very well be a complex signal, made up of different frequencies, without modifying the basic concept one iota. For example, our purpose was to transmit simultaneously a 2000-
cycle tone and a 1000-cycle tone of greater am-
plitude, we could set up two transmitters as shown in Fig. 4, with careful control of the relative phases so as not to have some 1500, 2000-

![Fig. 4 - A representation of four audio signals as a single transmitter modulated by two different frequencies. The 1000-cycle tone is twice the amplitude of the 5000-cycle tone.](image)

5000- and 1000-cycle signals in the receiver out-
put. 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 con-
sideration. In an a.m. transmitter the audio frequencies modulated (you could say "are beat against") or "are mixed with") the carrier and generate corresponding side frequencies through what you call "modulation" and "modulated amplifiers." You would be just as correct, if not more so, to call your modulator a "preamp am-
plifier" and your modulated amplifier a "mixer" (as you would in a receiver).

Sidebands with Sidebands

Since the carrier carries no intelligence, it
should be possible to dispense with it and intro-
duce it at the receiver. This will save transmitter power and reduce heterodyne QRM. Unfor-
tunately, if both sidebands are received at the detector where the carrier is introduced, the receiver must have exactly the correct phase rela-
tionship with the sidebands if distortion is to be avoided. Since exact phase relationship pre-
cludes even the slightest frequency error, such a system is unworkable with present, and possibly future, techniques. However, only one sideband is present at the detector, there is no need for an exact phase relationship and there can be some degree of 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 be-
terod QRM in one band 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, the transmitter
and receiver at the receiver, its frequency must be set 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 radiotelephony, but in voice work normal 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 trans-
mitter, extra signals will be generated and radi-
atured, or in the familiar case of overmodulation, are added. If insufficient carrier is supplied at the receiver, extra signals will be generated and heard, or in the case of trying to receive a single-
sideband signal with insufficient i.f. injection. The minimum possible bandwidth, of a modu-
lated signal is the bandwidth of one sideband. Ordinary a.m. signals uses at least twice this bandwidth because both sidebands are trans-
mitted. Claims that some methods of amplitude modulation result in narrower signals thus others are obviously spurious - any normal system resulting in double sidebands will give the bandwidth or any provided transmission, and that both systems are in proper adjustment. Out of adjust-
ment, they may result in less general band-
width.

That about the whole ugsa-orted story. Think of modulation, beats, heterodyning and mixing as exactly the same thing, and forget about carriers transporting audio and all of the other misnomers, and you will be able to understand your own techniques thrown at you. Visualize the audio signal modulating the car-
rier 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 hap-
ens when you pull out the carrier during trans-
mision and reinset it at the receiver, or stop off one sideband at the transmitter or the receiver - it will all add up neatly. Then try it in trans-
mitting only in terms of "modulation envelopes" and "percentage of modulation"!!

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SEVERAL years ago, having built and operated several successful amateur multicarrier 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 incoherences that showed up and then. To the best of my memory, I used to consider modulation from almost the following point of view:

'The i.f. section of a transmitter consists of a carrier-generating center and a final amplifier that amplifies the carrier and passes it along to the antenna. To use this typical i.f. transmitter for 'phone operation, we merely couple voice power to the final amplifier through a modulator 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 .5 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 that came single sideband. Formerly, I had considered single sidebands as a condition some-what resembling a case of senility, recurring only on unprofitably adjusted transmitters. With the introduction of amateur a.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 with- out much theoretical knowledge, the very 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 set, let me hurry to add that receivers only tell lies to people who don't realize that a receiver, designed for broadcast-band type reception, inherently distorts the true nature of incoming signals. Here is a typical example:

A multiplier tunes his receiver across an unmodulated carrier. The receiver tells him that the signal is a certain number of kilocycles wide. The multiplier immediately starts a frantic and often 'Lost Providence Mediation Pants' [209: July 1943].

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 "hardness," 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 was present, 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. No complete is the 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 modulation (modulation can be traced out 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 multiplier may think 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 100 kc. Now this happens to be a fine transmitter, except for one fact. There isn't enough filtering to throw this power away. Of course, the result is a signal plugged with 120-cycle hum in, in
the words of c.w., "I can't quite give you a Th. OK?"

Let's take a theoretical look at our hum-modulated signal. The hum voltage of 120 cycles should mix with the 3000 kc. carrier and produce new signals of 300120 and 389980 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 off.)

Until now, references to receivers may have seemed too farfetched. This, however, had only to do with the listener's lack of ability to interpret what he read. Now, let's see our receiver to tie down theoretical matters to what we actually hear. Simply turn on the receiver's b.f.o. and tune carefully across the hum-modu-
lated signal. Presto! We hear three distinct points of "zero beat." We have those 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 expositions of sideband generation. Of course, this is far from the actual voice signals we use to modulate our trans-
mitters. The voice contains a great number of individual frequencies which modulate the carrier. Each resulting new c.f. signal generated still main-
tains its original audio-frequency relationship with each of its neighbors, even though the whole business has been shifted up into the i.f. part of the apparatus.

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. Consider-
ing the original audio frequencies, we might think of the sidebands as being "heterodyned." The lowest-frequency voice 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 techni-
ques. SSB is definitely not a deep subject, but into the mysteries of complicated electronics. If you once again glance at the first definition 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. transmis-
tors, we'll return to the thought of converting in a regular receiver. From one point of view, every aspect 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 (off signals), tuned 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 three 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 reasons for considering a receiver as having s.s.b. action concerns "image" reduction. Due to the heterodyne process, if no sideband precedes the converter, the receiver is sensitive to two frequencies, one 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 had r.f. "images" so, in other words, having such sig-
als 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. sideband to the beat 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 pro-
duced by modulation ("conversion" in a re-
ceiver) 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 no practical advantage in attempting to cut down a book, but it is distinctly a separate subject from sidebands. A very high sideband is necessary to keep the superimposed r.f. systems and the form 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

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the filter method is used for reducing the unwanted sideband or "image" in a receiver, we will first consider this method as applied 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 LF 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 e.h.f. generation employs theories which certainly seem to belong to people with engineering degree. 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 readily 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 beats out." The two parts of the other sideband take an immediate liking to each other and combine to form the signal intelligible 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 e.h.f. 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 "radio 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 e.h.f. who use big tubes have wondered what the FCC regulations are on the e.h.f. power limit. We quote from a Commission letter addressed to ARL:

The following ... may be considered as a presently acceptible method for determining the d.c. plate power input to the final e.h.f. stage of a single-sideband superheterodyne 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 superheterodyne transmitter, as indicated by the usual plate voltmeter and plate ammeter, shall be considered as the "input power" instead of sections 12.131 and 12.183(b) of the Commission's rules are concerned, provided the plate meters utilized have a scale constant not in excess of approximately 0.25% varied, 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.

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What Single Sideband Has To Offer
DONALD E. NORGUARD, WKEU

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." Those 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 as 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 by it at all.

A plot of the frequency spectrum (voltage versus frequency) of the simple case of steady 100% modulation of the carrier by a single tone (sinusoidal wave of 100 cycles per second, 100 cycles per second, 100 cycles per second) 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 simple 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.

The voltage of each of the sidebands is half that of the carrier. Therefore, the power in each sideband is 1/4 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 output 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 overmoderate. This roils down the desired sideband power and generates spurious sidebands called splatters.)

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 the 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 these 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.

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Leaving the Carrier at Home

Sure, take a look at Figs. 1 and 2. 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 capability 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 was needed to be. This means we can double the bandwidth to 400 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 be put to work. That is the situation with the 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 modulability, to say nothing of how you try to use it. Yet, the power is there and it can be read on a meter, but that's about it.

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 is so happens that sideband energy in this form is usable. Yes, in all, it can be used, for it is just like one. It is indeed, and we receive it in just the same way. All that is necessary is to put the 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 carrier back, and we don't have to go out with a little piston-like hat on either. Your host will let you hang your hat on his backrack, and your hat won't know the difference, either, because the backracks 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.

Looking at the spectrum, the bandwidth of 400 watts of 30 watts (8:1) is 9 db. But this isn't the complete story. We are not considering only half the spectrum of the a.m. transmission and isn't below 30 watts. This means that the backrack of 8-db gain doesn't bother the other fellow as much as it were obtained with antenna gain on a.m. transmission.

Before climbing down from the ivory tower of abstract thinking let us take a look at what hanging our hat on the backrack really means. First of all, his has no effect on the power of 400 watts. The backrack is useless — it wasn't provided for us. We are not saying what our hat is put there. The point is this. The sideband must be based on a clear clean carrier of maximum frequency stability, and our host's carrier must be stable, too. A good crystal-controlled oscillator or a really stable crystal oscillator, made up of a stable backrack trans- mitter, anyway, so there is no worry on this point. Receiving a sideband has become exceedingly important through the years and it quite likely that our host is in a position to be a fairly good receiver.

At last, to hear him tell about it

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steady it a little bit by hand or to do some tinker-
ing with it in the fine time between rig-ups and schedule or (CQs) so that he doesn't have to coo 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 call y wanted it to do. It can do it for a short percentage of the time, but it probably would burn up if we hope that one sideband generated by the 1000-cycle tone jumping through it steadily. Fortunately, speech waveforms have a high ratio of peak to average power. It is average disipated power that burns up tubes, so there is nothing to worry about on this score still we learn how to hack 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 treat-

ment.

While abound in theory, we were talking about output power, well managed to show that we could get 400 watts of sideband power output with single sideband at the same peak power that gave only 30 watts of sideband power in the case of a.m. That's far for comparison purposes on a theoretical basis, but there is the practical matter of efficiency to remember. Let's lean over back-

ward and say that a good Class C plate-modu-
lated transmitter such as the one we used in our 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 (CQ) peak (which is 1000 watts or 1250 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 out-

put is 50 watts, with very low average input.
tion would be something close to 372 × 0.06 = 22.3 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 it out) that noise power is proportional to the square of 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 impulsive 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 reduce the power of the receiver to respond to all of the sidestream power that we can get 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 i.f. 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 know half the noise power when the bandwidth is cut in half. Apparently, this would then give the single-sideband system a 3.25-b (10:1 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 the same is not necessarily true when we consider a receiver’s response 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 noise 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 1/2 voltage unit at the detector. The demodulated signal in this case runs 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 also 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 of receiver output on any audio frequency, we should get back about the same results.
we had cursed by reducing the bandwidth by two to one. Therefore, on an idealized theoretical basis we must conclude that single-sideband operation can give both signal-to-noise ratio improvement and signal-to-distortion improvement operating at the same equal power output.

Back up to the ivory tower we begin to wonder what significance this whole system gain has, since we arrive at this figure on an idealized basis. This idealized condition is looked upon as a type of the only 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 as complete mask not thermal noise in a good receiver operated on our low median-frequency bands that we should try to evaluate the performance of single sideband working under the conditions we know we have.

Impulse noise—the clicks and pops we hear—produces detector output voltage noise or has proportion 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 singlesideband 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 signal 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 desired signal and the interfering signal caused by the carrier sidebands of the undesired transmission accompanying with a relatively strong desired carrier. A crystal filter may be used to put a notch in the i.f. passband in 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 carrier is much more bothersome than the monkey chatter, so we pay to notch out the interfering carrier. With singlesideband 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 heterodyne and monkey chattering.

The crystal notch may be used to eliminate one carrier heterodyned, but that is about the best one can do. The advantage of single-sideband reception in this case is primarily 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 detected against each carrier, and there is as much monkey chatter caused by the desired sidebands beating with the detected 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 carrier by far the loudest signal band, and it consists of a fundamental heterodyne note and a series of fairly strong harmonies throughout the audio band. Add a little QRM on both sides 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 course is removed. The remaining monkey's chatter of 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 this interference, but the monkey chatter remains. Here again, the exposure to QRM is cut half, but that is about the best one can do. Interference weaker than the desired signal but, of course, with a high weightings is besieged by more than one interfering signal of equivalent strength only one of the carriers can be in the crystal notch, and the others have to be tolerated along with monkey chatter. The re-making in the heterodyne, however, would be more disturbing since they are not distorted in the detector, so what is left is then partly 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 detect, but that is not so easily disposed of. The low-level speech
sidelbands 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 sidetones are demodulated to produce objectionable noises, groups, and monkey chatter of a kind that is horrible. It’s all a wild mess in spite of anything that can be done with the very best commercial equipment. With single-sidetone reception of the desired carrier, all of the undesired noise is, of course, louder than in the previous cascs, but that is the only difference. Noticing 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 a faint 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 set in half cuts down the average probability of trouble by a factor of two or three.

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 two adjacent signals are involved. The carrier signal not only does more damage than necessary to others using the band, but it is out of lock when it is the recipient of QRM from other transmissions.

When single-sideband signals are in the case of QRM, the situation is much the same. The tone carrier unless the sidetone strength is sufficient to put the interfering to the class of a signal which exceeds the carrier strength (§, an m.m. signal). Single-sideband reception cuts up this difficulty, but does not eliminate all interference. Single-sideband reception of equalized m.m. and n.m. signals with excited 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 ear-throat 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 causing QRM.

SIDEBAND SUPPRESSION AND RECEIVER TUNING

D. C. Bokemus, W9REX, wrote an interesting article on QRM for his M.S. at the University of Illinois. Some of his experiments have a bearing on technical standards for amateur a.m. his findings are being passed along. For example, it was observed that attenuating all audio 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 burr in your get on reception when the carrier modulation isn’t exact was used as the criterion in some experiments on acceptable sidetone injection figures, and it was found that this burr was only barely noticeable when the side-tone modulation was 30 db. Thus it looks like a probably an acceptable minimum figure to shoot for, as far as sound 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-distortion tuning curve was also studied, and it was decided that the naturalness of the voice had definitely disappeared at 20 cycles high (audio components made higher) and 20 or 30 cycles low, although it was still easily understandable.

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How To Tune In a Single-Sideband Signal

Byron Goodman, WDIO

Trouble? 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 s.s.b. section of the 20-meter sideband bands is Fig. 1. Your receiver can be considered a sort of "antipode" that you slide back and forth across the band. If you are going to build a working model of the 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 the notch would be. Your working model would consist of this cardboard strip laid out 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 "5000 kc." it corresponds to the notch being centered on 5000 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 3011 kc., you would see (and hear only) "Signal B" and with the receiver (cardboard scale) centered anywhere from 2918 to 3102 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 cloth glued mounted on it. This clothoid has a single vertical line strung on it, representing the beat-frequency. A waving model would look like Fig. 2B. Assembling your receiver model, it would look like Fig. 2C. Your h.f.o. adjustment on your receiver is the knob moves the cardboard strip and the clothoid together. This simply means that the relationship between h.f.o. frequency (line on clothoid) and the receiver nullband (notch in clothoid) is constant with receiver tuning.

Your receiver with h.f.o. off looks like Fig. 2A. Now you're ready to tune into that s.s.b. station represented by Signal C in Fig. 1. With h.f.o. off, tune your receiver until Signal C is centered in the passband. As mentioned before, any setting between 2918 and 3102 would

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1. The d.c. power input to almost every s.a.h. 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 explicitly considered to be acceptable. The specification (or understanding) with respect to distortion is indispensable because the amount of power that can be obtained is critically affected to the amount of distortion that is permissible.

An exactly similar method of rating is ideal for the s.a.h. 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 it is equally meaningful for amplifiers falling anywhere between pure Class A and pure Class B.

However, it is not always easy to measure d.c. 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,
Carrier Insertion in S.S.B. Reception

PAUL N. WRIGHT, W0HJM

I\n in the reception of an s.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, en\ncouraging us to recover at the receiving end the in\ntelligente-bearing frequency with which we started at the transmitter.

So far as the detector in the receiver is con\ncerned, 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.m. sig\nual 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. sec\ntion of the receiver, or by inserting 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 i.f. in the re\ceiver is used, the receiver should be adjusted as follows: First, with the receiver set up in the regular a.m. position, tune for maximum deflec\tion of the d-c meter from the a.b.b. signal. Do not touch the dial after this. Next, reduce the r.f. gain to zero and increase the audio gain to maxi\nmum. 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 normal. If this pro\ncedure 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 injection, it should be pointed out that practical reception of a.s.b. signals depends upon the stability of the b.f.o. Since the carrier is at the front end of the receiver, as well as the stability of the beat oscillator that supplies the carrier. Any frequency change in the b.f.o. oscillator produces the same effect as changing the frequency of the transmitter on the other end. The b.f.o. oscillator in most receivers is fairly

stable on the local frequencies. However, at frequencies above 5 Mc. the stability of many b.f.o. oscillators leaves much to be desired, when thinking in terms of the stability required from these oscillators when using i.f. carrier injection.

Signal-Frequency Carrier Insertion

In using carrier insertion at the signal fre\nquency 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 8-meter from the signal. Then adjust the frequency of the external oscillator to the ap\nproximate frequency of the incoming signal, and increase the amplitude of the carrier injected to a point that approximates the amplitude of the s.m. signal. When this point is reached, the 8-meter will no longer swing with modulation. Carefully adjust the frequency of the external oscillator until the voice sounds normal. Rock the receiver dial back and forth across the carrier. You will easily be able to tell which sideband is being transmitted. As you have the carrier, or one side the audio will stop off; as you swing on the other side, the audio will come up. The more selective the receiver, the more pronounced this effect.

An a.b.b. signal suffers a certain amount of nonlinear distortion when demodulated by a linear receiver. Increasing the carrier injection above the 10% per cent modulation point will re\duce this distortion to a negligible amount. In\ncerse carrier also helps swamping out adjacent\nchannel QRM.

The advantages of front-end carrier insertion are:

1. Stability.
2. Gain points may be given on s.m.
3. Round tables including s.m. and a.m. sta\ntions become practical, since the receiver remains in the a.m. position at all times.
4. Oscillators in the s.m. receiver may be used to furnish the stable carrier to the receiver, pro\viding consistent "on frequency" operation of the transmitted signal.

Point 4 is very important from the standpoint of transmissible operating and good operating prac\ntice of a s.m. station. Since the oscillators in the s.m. carrier 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.

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VFO Signal-Frequency Carrier Injection
PAUL N. WRIGHT, WOØHM

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

1) Frequency stability.
2) Output amplitude control over a wide range.
3) Sufficient bandwidth to simplify the mechanism of precise frequency spotting.

Harmonic Generator

The circuit of Fig. 1 illustrates a stable carrier frequency generator using a 1-Mc. VFO and a harmonic amplifier. Carrier output is controlled by R₂ in the circuit. About 60 db. of control is available with R₂. 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 hand-tuning arrangement for the plate circuit of the oscillator and amplifier.

The unit serves as a bandside marker at 4 Mc. by zero beating WWV at 5.5 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 +4 Mc. interests up through the spectrum, enough to be used as a marker to 30 Mc. However, if it is to be used to provide carriers for receiving s.s.b. plate tank circuits tuned to the desired harmonic.


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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 a.m. signal. Fortunately, this deficiency can usually be easily overcome by switching the indicator over to the audio circuit during a.m. reception. The circuit diagram shows how the arrangement has been applied to a National type NC-181D receiver.

In the modified circuit, the S-meter terminals are connected to the center arms of a d.p.d.t. toggle switch, $S_3$. 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 in series with a calibration potentiometer, $R_0$, and the secondary of the output transformer, $T_2$. Naturally, the speaker-transformer connections do not have to be disturbed when the modification is being made.

If the receiver as-hand does not employ a short across the S-meter, it will be necessary to add $R_0$ to the original indicator circuit. This resistor prevents the r.f.-inductive circuit from opening up whenever the meter is switched over to the a.m. position.

A calibration for the a.c.h. 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, W4WFC/WE3WFC

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Suppressing the Carrier

The carrier can be suppressed or nearly elimi-
nated by an expensively sharp filter or by using a balanced modulator. The basic principle in any balanced modulator is to introduce the car-
er in such a way that it does not appear in the output but so that the sidebands will. This re-
quirement is satisfied by introducing the audio (in push-pull) and the r.f. drive in parallel, and con-
necting the output (plate circuit) of the tubes in push-pull, as shown in Fig. 1B. Balanced modula-
tors can also be connected with the r.f., drive and audio inputs in push-pull and the output in par-
allel (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 (germa-
nium, copper oxide, or thermistors) 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 r.f. equipment. Examples of rectifier-
type balanced modulators are shown in Fig. 2.

In any of the vacuum-tube circuits, there will
be no output with no audio signal because the cir-
cuits 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 also
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 modu-
lation process is the same as "mixing" in re-
ceivers, tone and difference frequencies (side-
bands) 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 the
amplified carrier. Normally two tubes of the
same type will balance closely enough to give at
least 15 or 20 db carrier suppression without any
adjacent circuit. If greater suppression is required
trimmer condensers to balance the grid-plate
tube leads and separate bias adjustments for the
operating points can be used.

In the two-diode 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 when the drive is at either of the two possible
paths. The effect is that no r.f. energy ap-
pears in the output. When audio is applied, it
modulates the signal by biasing the diode (or

Fig. 2 — Two examples of balanced-modulator cir-
cuits using two grid modulation. In (A) the r.f. modula-
tion is introduced in both tubes, and the audio and out-
put are in push-pull. In (B) the r.f. audio and audio are
in push-pull, the output is in parallel. In either case, the
carrier frequency, f, does not appear in the output cir-
cuit — only the two sideband frequencies, f + f and
f — f, will appear. The tone fed to the carrier is a
constant r.f. signal. Note that in neither of the circuits
is any special provision for balanced modulators.


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The circuit at A is called a "hybrid" balanced modulator and has been widely used in commercial work. The balanced modulator circuit at B is shown with components suitable for operation at VHF. It is a push-pull modulator and is designed with a Class C output stage. The output stage is coupled to the output circuit by a transformer. The transformer is designed for efficient power transfer and has a turns ratio of 1:1. The transformer is also used for impedance matching.

Diodes in one path, depending upon the instantaneous polarity of the signal, 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 actual operation involves a fraction of a volt of audio and several volts of r.f. The diodes should be matched as closely as possible -climatometer measurements of their forward resistance 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.a.h. 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, for sure, or his YER-1 can lock on, but with none at all there is absolutely no reference whatever, the "null" point and the output circuit are interchangeable, the signal is not there.

"On the Air with Sideband" June 1947, p. 66

In the October column it was wondered aloud if perhaps there wasn't too much emphasis put on carrier suppression, pointed out that with a little carrier one could readily readjust the R.C. and the R.C. biasing to get back to lock in. The s.a.h. thinged it up on it, hacked it out over the air, and W6ANW was kind enough to forward the comments:

...we are not quite sure why [less carrier suppression] yet, and the main reason is that not enough followers 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 sidebands are very noisy. Then you have been listening to them on a certain frequency and another fellow takes over on (the same-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 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 YER-1 adapters say they have ohmed the lock-in because near-by interfering signals take control and loose up things in general. Further, the adapter will give a false indication of any beat when they are trying to set up on or, because it is pulled in by the signal as it approaches the frequency.

"The other evening I was talking to W40LL about this carrier business and he brought up another instance that nicely as against us. We were having our own private QSO on 3069.5 and another couple of fellows were having theirs 747 cycles lower than us. That is too far away for pulling a YER-1, but it falls in the passband of any adapter, and we had a handy birdie to listen to every time this one fellow came on. The other fellow was clean, and the glibberish from his sideband wasn't nearly as annoying as his carrier.

"At any rate, until until, time as all fellows have hit upon a simple means for stabilising frequency and/or removing sufficient distortion and having the correct ratio of highs to lows so that there is no tolerance, or some mischief, 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 Sideband" July 1947, p. 111.
Diode Modulators

BYRON GOODMAN, WIDX

Because single sideband, amateur had little say on contact with diodes used as modulators. While they had been used for years as demodulators—"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 commercial code, 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 "sinewave" or "cosinewave" wave, from the trigonometric function that define 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/T, where T is the period. If the wave in Fig. 1A is represented as a 1000-cycle wave, 1/T is 0.0001 second, but if it were a 100-ke wave, 1/T is 0.00001 second. Drawn to the same scale, the 1000-cycle and 100-ke 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 signal. 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 an a.c. wave to a steady d.c. value, as shown. The polarity never goes negative in contrast to the a.c. wave scale 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 plus of amplitude vs. a time of a. The two different frequencies drawn to the same time scale scale back exactly the same, because the higher frequency cycles are superimposed. (B). The shape is the same, however — only the scale is different. A pure single-frequency a.c. wave must swing equally above and below the axis — if it doesn't, it has a d.c. component (C).

This a.c. wave with a d.c. component is easy to spot by eye, and notice in many places throughout radio equipment. The current in an audio ampli- fier 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.

SINGLE SIDEBAND FOR
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 dies away 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 waves, 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-cycles above the zero-amplitude axis, the outline bears a strong resemblance to Fig. 2C, except that in Fig. 2B the envelope replaces the signal, and the half-carrier amplitude replaces the d.c. component. The same picture, flipped 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 dispise 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 is a 1000-cycle signal and a 100-kc. signal exist in the same circuit. It is no longer true that the zero-amplitude axis is the same as the zero-value 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 wave, the peak-to-peak amplitude of each 100-kc. cycle is the same as that of the previous cycle, even though the envelope 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 the envelope. (A) exhibits a square wave; (B) shows the same wave with an envelope; and (C) is the waveform of a 1000-cycle wave with the less-familiar pattern of "superimposed" waves at (B). 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 rectifier (for 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. Put suppose we use no 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. Whatever the 100-kc. applied voltage swings to the right (in positive), the diode conducts and a half-cycle of r.f. passes through it. Fluctuating 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.

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coupling (through $C_3$) 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 as the signal proceeds in diode modulator to is 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 have a high per- centage 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-diode 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 characteristics (har- monics) 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 circuit 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 DB and CD will be “biased back.”

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**SINGLE SIDEBAND FOR**

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Fig. 5A - A modulating signal as in (A) gives an e-f output from a balanced modulator as in (B), by the amount of the battery voltage, and they will not conduct e-f. (at the proper polarity) until the e-f. voltage exceeds this bias voltage. The other two diodes, BC and AD, will conduct readily, however, and over more than half the e-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 e-f. will appear across the output. The fact that there are approximately half cycles of e-f. flowing through the diodes shouldn't bother you - remember that this is an a-c-coupled affair and the e-f. will be normal full cycles in the output. The more voltage applied, the more the imbalance, and the more e-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 eager ones.

Since the output depends upon the voltage across points B and D, it is essential to cut out audio transformer and apply a single audio frequency, the e-f. output will appear in proportion to the audio voltage and regardless of the instantaneous polarity. Thus we will obtain an output like that of Fig. 5A when an audio voltage like that of Fig. 5A is applied. This pattern is that of the "touhou" 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 modulator and a.e.

Except that this isn't the complete story. One thing these envelope patterns can't show is the modulating 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 (unmodulated) 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 (unmodulated) carrier frequency of 100 kc. Such an envelope pattern can be generated in a normal modulator, by modulating 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 out 1000-cycle intervals out to 10 or 15 kc., either side of 100 kc. Hence, although the envelope could look the same, the spectrums would differ greatly - the difference is in the phase of the e-f. cycle and the lack or presence of a carrier. In the balanced modulator, the phase of the e-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 e-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 basic 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 15 to 20 times the peak modulating voltage, for germanium crystals and copper-oxide rectifiers, the "e-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 50 db. Although the modulator power delivered, minus the resistances from the diodes, and the losses will run from 2 to 10 db., depending upon the carrier frequency. The rectifiers must, of course, have a sufficient g.m. up to 4 Mc., and will be usable at higher frequencies with careful construction. A breakdown 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 in a practical level at 500 kc., but 30 to 100 ohms is recommended at 4 Mc.

In some of the first single-ended devices built in 1948 QSTs, you will run across the abbreviation "a.e." for "audio-frequency suppressed-carrier." Later on the abbreviation "a.e.b." was adopted, to conform with commercial practice and because the amount of work eventually became burdensome at all times, so "suppressed carrier" is understood when g.m. is talked about.

THE RADIO AMATEUR 29
The Series Balanced Modulator
FRED M. BERKEY, W7MNN

The "series" balanced modulator is presented because of its simplicity, good linearity, and excellent carrier suppression. In converting from audio to s.s.b., an improved transformer is required in the audio portion, which makes it possible to keep the box down without expensive components.

The basic circuit of the series balanced modulator is shown in Fig. 1. The action of the carrier generator is to switch point B to ground at the carrier rate. A basic requirement of the circuit is that $Z_C$ (the impedance across which the modulating voltage is developed) have low impedance for the carrier frequency, and that $Z_A$ (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 $Z_A$ and $Z_C$. $Z_A$ and $Z_C$ should be approximately the same impedance, but not equality is not necessary unless a very minimum of low frequency is desired. Impedance ranges between 500 and 50,000 ohms have been used—the optimum impedance depends upon the carrier power and the line built up across $R_A$ and $C_A$.

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

From QST, September, 1932.

Practical Circuits

The use of a series balanced-modulator circuit in the audio modulator ahead of the speechband filter is shown in Fig. 2. The 6CA4 cathode follower eliminates the need for any step-down transformer, and $C_2$ offers a low-impedance return to ground for the rectifier. The coupling coil, $L_1$, should be sufficient to develop about 6 volts across $R_3$. The alternative connection for the coupling coil eliminates $R_2$, $C_4$, and $C_3$ 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 tube, and a 6AG6 phase splitter is used to get push-pull excitation from the VFO. The VFO should de-
Sufficient voltage to furnish approximately 6 volts across $R_1$, $R_2$ and $C_1$ are adjusted for best carrier balance, as indicated by the absence of VFO-frequency signal in the output of the 3.5-kc. amplifier. The inductive coupling between $L_2$ and $L_4$ is adjusted to give best bandpass characteristics over the 200-kc. phone band. The circuit also illustrates the fact that the diodes can be connected between ground and point $A$ or point $O$ of Fig. 1.

The modulator circuit of Fig. 2, of course, be used at 10 kc. (ahead of a crystal filter) with germanium diodes or a 0.525 for the modulator and with a 450-kc. tuned circuit at $T_1$.

A double-balanced mixer-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 led to the center tap of the primary. One side of the push-pull oscillator is fed through resistor and diodes of positive polarity to the ends of the primary, and the other side of the push-pull oscillator is also led through resistors and reverses to the ends of the primary. A diode and a diode cathode connected 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 (e.g., a.c.c.w.e.). 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 Edmonde 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 exciter designs in G0T72. S.R. used radio and r.f. phase-shift networks, of course, but no balanced modulators. Instead, a pair of grid-modulated 6DK7s were used to generate r.f. with carrier, and the carrier was then balanced out by introducing equal and opposite carriers through a third 6DK7. The resultant a.c.i.e.-suppressed-carrier signal was amplified by an E801 similar to 4L6A.

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.

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Plate-Modulated Balanced Modulators
BYRON GOODMAN, WIDX

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 grid. The condenser from cathode to ground are n.f. bypasses. 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 exciting voltages passes through the stage. However, when the tubes are modulated with push-pull audio, the double admittance (maxima minus minima) appears 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 signal power is a transformation of the applied audio power, with the grid-plate loss. The circuit is quite tolerant for all excitation in nature, requiring only that there be sufficient drive from Class C operation over the range of modulating voltages. The small powers involved, if it seems to be reasonably taken out of loading, although it does result some heat (which can be the lesser in the circuit) and it isn't much ever neutralized beyond its linear operating condition.

For a constant balance, both triodes should have identical characteristics and interelectrode capacities. However, it will be noted that two tubes of the same type will normally give adequate carrier suppression, and each dual triode like the 6SN7GT can be used without sacrifice. When using dual triodes, which usually do not have identical interelectrode capacities, it may be helpful in reducing the current 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.

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

M. Prescott, WIDX, 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 reason good why the tubes be passive types couldn't be used. Potentiometers $R_1$ and $R_2$ control the carrier balance (or injection, if you prefer). The switch $S_1$ merely simplifies another modulator, for use when turning up with the two-tone method. The grid currents to the 6SN7GT run about 2 to 3.5 ma., and if they aren't reaching within about 30 per cent of each other the audio-balancing controls (not shown — they are ahead of the phase-shift networks) will not act to work.

The r.d. phase networks are a little different thus may be found elsewhere. WIDX over wire-wound resistors on 4 MA, and makes use of their inherent inductance. At 4 MA, $L_{AB}R_1$ is simply a 1200-smd 100-ohmOhm wire-wound resistor, and $L_{BA}C_1$ is a 1000-ohm 10-ohm 12-turn 46-ohm 30-turn tap. In the latter of No. 20 enamelled wire wound on a 10000-ohm 1H 5C inductor and $L_{BA}R_1$ is a 30-turn bobbin of No. 20 enamelled on a similar resistor, with a small noise-compensation trimmer across the works.

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

SINGLE SIDEBOARD FOR
A CARRIER NULL INDICATOR

The usual method of balancing out the carrier of a s.s.b. receiver 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 meter. This can be very bothersome when one is often reworking and removing carrier for demonstration or educational purposes. Mark Meinhardt, W2AJL, passed along a simple gadget that he uses with his Central Electronics 10-A receiver, and it has the advantage that it can be left in the circuit at all times, without piercing the needle when the operator starts talking. It shows excellent sensitivity in the millivolt range and yet is not overloaded by the 100-volt peak output of a 6AL7.

As can be seen, it uses an IN34 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 W2AJL’s final amplifier outputs of 2 and 225 volts (across a 20-ohm load) give voltmeter readings of 0.2 and 2.0 volts. The basic idea is a useful one that should find other applications around the ham shack.

THE RADIO AMATEUR 33
Demodulation and Selectivity in S.S.B. Reception
OSWALD G. VILLARD, JR., WSGYJ

When relatively little-difficult, the amateur
may convert his existing receiver for
single-sideband reception, and obtain per-
formance rivalling 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 corrected by single-sideband
will then be outlined. It is pointed out that while
some of the advantages of single-sideband detec-
tion can be realized with unmodified standard
communications receivers, a great improvement
can be obtained through addition of a simple ex-
ternal single-sideband detector and low-pass audio
filter.

The fact that single-sideband transmission
makes possible a 30 per cent saving in trans-
mitted bandwidth is almost academic when con-
sidered in the light of the effective selectivity of
present-day receivers. It is all very well to hope
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 sta-
tion that is very weak — perhaps one microphone
above the receiver input terminals. For a given
I.F., gain, a certain voltage is delivered to the
diode detector. Now we wish to do is press the
button on any station on any other frequency (not deliv-
ering a signal of approximately equal strength to
the detector. It is a characteristic of linear detec-
tion — 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,
and so 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 carrier is inaudible, but both modulations will
be heard if the two signals are of equally strong
strength at the detector.

From “Selectivity in K.S.B. 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 modula-
tion on the weaker tends to be completely sup-
pressed. 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 will
be very automatic if that signal happens to be a
weak DX station, say some ten 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 specifica-
tions for an "ideal" ham receiver, we might con-
side what sort of selectivity problems we are actu-
ally up against. It is clear that the ideal re-
ciever 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 blinding away on the channel im-
mediately adjacent. We know that the weakest
signal we can receive is approximately one
millivolt. To find the strangest signal we are
likely to have to hear (just around the corner
perhaps) 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 re-
cieved 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 case, for instance, at a frequency of a
i.f., tuned circuits, the response is down 1000
times at plus 10 or minus 8 kilocycles from the
center of the passband for the narrowest setting
of the "bandwidth" control. For another re-
ciever, which has three i.f. tuned circuits, the
response is down 1000 times at plus 10 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
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t-microwave L.I.S signal, the nearest 1000-microwave
voltage interfering signal must be at least 8 kilometers
away in one case, and 10 to 11 kilometers away in
the other case. In order for both D.X. and unde-
sired 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 at-
tracted still further, until it is less than 15 or 14
as strong as the desired signal.

The tendency for a strong adjacent station
cross-talk to blot out the one thing, so that we are
listening is familiar to everyone, and is the scan-
non for the possibility of extra i.f. selecting such as
is provided by arrangements like the Q-Bars.
Without such extra selectivity we cannot make
full use of the frequency space now available to us.

Simple-Sideband Detection

The situation is quite different in the case of sin-
gle-sideband reception, because, if fundamental-
tially different processes of detection are used, e.g.,
or o.c. signals are detected by frequency convolu-
tion, rather than by rectification. Here it is
to help to review some theory. A single-side-
band wave signal may be thought of as a band
of frequency simply displaced in the frequency
spectrum. To each of the frequency groups in the
wave wave, a constant frequency is added.
Thus a speech sound, which might consist at
some instant of three component frequencies —
e.g., 990, 1690 and 2590 e.p.s. — can be trans-
mitted 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 1,990,260, 1,001,004 and
1,259,004 cycles per second, forming a voice
single-sideband which can be transmitted by
radio. A single-sideband transmitter, then, is
fundamentally a frequency-translating device
that shifts the incoming speech frequency to the
desired position in the radio-frequency spec-
trum. To receive these signals, it is only necessary
to select this process; by extracting the con-
stant frequency of 1,000,000 cycles we can re-
cover the original speech frequencies of 990,
1690 and 2590 cycles.

Now, the radio detection of code signals is ex-
celled by a very similar process. Assuming an
incoming keyed c.w. signal of 1,000,000 cycles
per second, if we add a constant 1,000,000
cycles per second, we have what we let by, i.e.,
zero frequency, or keyed c.w. In actual practice, some-
hing like 999,000 cycles per second is substracted.
The c.w. signal is thereby translated to a fre-
nquency of 1000 e.p.s., which, when amplified
up to a loudspeaker, is heard as an audible
noise.

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 by two strips of local
oscillator which converts the incoming signal to the
I.F.; and the best oscillator, in conjunction with the
second detector, converts the I.F. down to an
audio frequency. For either e.c.m. or single-
sideband voice reception, the usual dide 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 rec-
tificator to take place, particularly in the con-
version at the second detector, because this
rectification would preclude the modulation on
undesired amplitude-modulated signals to be
heard along with the desired signal resulting
from the frequency-modulation process.

Overcoming Rectification Effects

One way to suppress the unwanted signal result-
ing from rectification is to make the voltage in-
jected by the low-frequency oscillator very
strong in relation to the incoming r.f. signals. In
the ordinary communications receiver the ampli-
tude of the h.f. e.p.s. 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 r.f. gain be on full and
r.f. gain be kept at a minimum for single-sideband
reception. Under these conditions the audio out-
put resulting from rectification of the incoming
signals is small compared to the audio output
resulting from the beat between the desired single-sideband signal and the b.f.o.

However, even then the rectified audio is un-
fortunately 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 reduced by reducing r.f. gain,
the audio output from the detector is down quite a
bit, but it is better to recheck the audio output
from the detector in order to correct for the
losses. Furthermore it often happens that although the
weak or single-sideband signal one is listening to is only one-tenth as strong as the local os-
cillator, an interfering signal on an adjacent chan-
nel may be one-half or more as strong.

The radio amateur will therefore, give rise to inter-
ference that will be equally bad no matter how
weakened the frequency of the interfering signal is separated from that of the desired
signal.

An objection to using a receiver at reduced
e.f. gain is the resulting reduction in signal-to-
noise ratio in the first e.f. stage, unless the set
happens to be one in which the stage stage operator
is the output stage.

Of course, it is possible to use the opposite ap-
proach and increase the amplitude of the desired
b.f.o. voltage while keeping the receiver r.f. gain unaltered. This procedure is to be recommended, but it can
only be carried to a point at which the second
detector overloads.

For these reasons, it is hard to get the full
benefit of single-sideband reception in QBM

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reduction when using a conventional receiver with b.f.s. But it is not difficult to build an external frequency-converter unit (or detector unit), especially designed for single-sided reception, which can be added to any communication receiver. The method of operation of such a unit is somewhat different from that of ordinary converters, and 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 pulses 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.s.-second-order 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 M.G. converter tube, for example, works on this principle; the local oscillator simply suppresses-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 operate it in a class 'A' mode at the incoming signal. But where such a restriction is impractical, as in the case if the frequency converter must operate at a fairly high-signal level, it is possible to obtain satisfactory results by using those tubes in class 'B'—the push-pull or balanced frequency converter. In this circuit, each tube is fed 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 as to keep their outputs the same, the audio signal is canceled out. The local-oscillator voltage, on the other hand, is fed to the two tubes in parallel, and consequently the audio outputs resulting from the beat between this oscillator and the incoming signal add up in phase in the output transformer. Thus the desired beats are heard, while undesired sidebands 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 6ET7 and 6ZF2 are especially designed to have less third-order distortion, and their use makes possible a rejection that is quite adequate in practice.

Adjusting for Balance

The current "balance" of the balanced detector circuit may easily be found. An ordinary modu- lated signal is applied to the detector, and the heating oscillator is either turned off or, preferably, turned on, with a pilot lamp. If any beats between it and the signal are above the limit of audibility, then the amplitude of the sig- nal is increased until the modulation on it just begins to be heard — in other words, the signal just begins to ride through. Disregard any dis- tension. To balance the detector, the cathode-biasing 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.

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Fig. 1 - Balanced frequency converter for single- sided receive. This circuit makes it easier for the conventional receiver to serve as a single-detector circuit in a superhet receiver. C3, C4, C6 — 0.0002, uf. C7, C8 — 25-muf. electrolytic, 55 volts. C9 — 10,000-pf. resistor. C10 — 15,000-pf. resistor. C11 — 150-pf. resistor. R1 — 50,000-ohm. R2, R3 — 10,000-ohm. Rs — 500-ohm outside. Rs, Rs, Rs — 10,000-ohm. T1 — 6D7 transformer with center-tapped secondary (molded to heat oscillator). Y1 — 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 exter- nal adaptive. An external hearing oscillator must be provided, however, and likewise a sepa- rate 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.s. It is no longer necessary to keep the i.f. gain centered down low in order to keep the b.f.s. voltage large compared with the signal voltage. Strong adjacent-modulated signals no longer ride through and cause distortion by add-

SINGLE SIDEBOARD FOR
single-sideband definitions and jargon

anti-trip — System of voice-controlled break-in, operated so that it prevents signals from loudspeaker controlling the transmitter.

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

modulation envelope — Envelope of modulated signal. When received by rectifier, it is the modulating signal of an a.m. signal. In e.s.b. the rectified envelope does not represent the modulated signal — those must first be reenveloped. Envelope of e.s.b. signal is of practical importance in determining limits of linearity and power of amplifier.

phasing exciter — e.s.b. exciter using the phasing principle. See page 50.

q multiplier — Electronic means for increasing the selectivity of a circuit. Often used at receiver intermediate frequencies in loud-hailer circuits to assure broad bandwidth of frequencies.

ql — A commercial anti-trip circuit.

select-o-ject — Audio-frequency selective. See page 50.

slicek-sideband or signal — A switchable-sideband adapter for receiving, using audio phasing system as described on page 100.

vb — Commercial switchable-sideband adapter using physical principle similar to the slicek. Also includes a c.f. (constant-frequency control) circuit for having on carrier.

zoning — In — Toning suppressed earlier frequency of transmitter to assure best with suppressed carrier frequency of other e.s.b. trans- mission.

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bization represents a considerable improvement over present-day receiving techniques.

Conclusion

The conclusions reached in this article may be summarized in the following way. With conven- tional modulation the linearity-detector 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" ef- fect which results in complete disappearance of the desired signal when the interfering signal is very strong. These two disadvantages of conven- tional reception, familiar to all "phone men," tend to prevent maximum utilization of existing fre- quency assignments in the sense that very weak stations cannot be copied immediately adjacent to very strong ones. Single-sideband reception, by means of the d.c.m., on standfaudio, reduces the suppression effect but still is vulnerable to the modulation on undesired signals because of the possibility of nullification occurring along with frequency competition. 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 combina- tion 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 overload 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.

anti-trip — System of voice-controlled break-in circuit that prevents signal from loudspeaker controlling the transmitter.

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The Product Detector

BYRON GOODMAN, WIDX

The "product detector" circuit of Fig. 1 is useful in s.a.h. and s.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 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

![Diagram of the product detector circuit](image)

the signal and for the b.f.o., working into a common cathode resistor (170 ohms). The third triode also shares the 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 3000-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 resistance of 5000-ohm potentiometer. Tuning on the b.f.o. brings in the audio, because with the detector output is the product of the signals in the two channels. The negative bias supply should be a well-filtered affair, because the 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.

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. detector), 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 give 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-modulat-

ing beats between signals.

SINGLE SIDEBAND FOR
Low-Frequency Filter Design for the S.S.B. Transmitter

FRED M. BERRY, W3MNN

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 unwanted sideband by "dropping lines." 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 practically only at frequencies in the order of 50 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 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.

![Block Diagram of a Practical Single-Sideband Transmitter](image)

In the notation of Fig. 1, "USB" and "LSB" indicate the position of sideband at the grid voltage with respect to the input speech frequency. 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 10 to 17 kc, and

move (modulator) not of the balanced type, for the third modulator.

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, equality of components and desirable characteristics were of first consideration; low cost and ease of con-

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situation 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, controlling the attenuation where most needed, and resulting in a mini-

mum number of inducers. 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 modula-
tion stage, this is no hindering.

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 lead 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 intermodu-

lation 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 communications and is thought to be a practical one for amateur use. A frequency band of 17 to 29 kc. was chosen in preference to one of lower frequency to reduce the selectivity requirements of the second L. filter, as previously noted. This rather high frequency for a single-sided filter also has the feature of lower component values, lowering cost and mak-

ing hand winding of the inductances 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 ripple 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

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

inductors in the complete filter.

The input and output impedance characteris-
tics are the same as that of the midband constant-k type of filter of the same cut-off

frequency. This sort of termination impedance is most economical and works very well either di-

rectly from a ring modulator or resonance termi-

nations.

The design impedance R of 1000 ohms was selected to give desirable component values and
desirable input and output impedances obtained by transformer action in the end inducers.

The resultant values calculated from the design formula of Fig. 4 for each filter section are given in Fig. 3, and in Fig. 4 for the two sections com-

bined.

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 dif-

ferent.

The filter of Fig. 4 might now be constructed, and if proper components are available, the inser-
tion loss between 1000-ohm resistive imped-

ances would be approximately that of Fig. 3. A low dissipation factor (high Q) is necessary in auxiliaries of this type to avoid the required characteristics. Resistor losses internal to the filter unit only will cause a greater loss at all fre-

quencies but will "round off" the edge and prevent the rapid rise of attenuation needed just outside the passband.

Except in the case of C1 and C7 (Fig. 4), mic or other low-loss types of capacitances are necessary for proper filter action. C1, C4 and C5 are large values for mini receivers and would be ex-

pensive. Ordinarily they would have to be made from a large number of parallel units. C1 and C7 appear directly in parallel with the terminat-

ing resistors and it has been found that good-

quality paper capacitors are satisfactory here.

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**SINGLE SIDEBAND FOR**
Component Tolerances

In the filter of Fig. 4, the tolerance of some of the elements is quite critical, particularly that of the series arm. It has been found in the design of filters of this type that the tolerance of the LC ratio 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 self-adjustment of the inductor. The maximum possible error of 10 per cent in the impedance match between junctions of the filter arm 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 arm 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, C10/C1, C10/C1 is selected with a tolerance such that it will always be larger than the calculated nominal value, L0.
may then be adjusted with this new value of $C_1$ to the correct antiresonant frequency and will have fewer turns than the original calculated value of $L_0$. Leaving $C_2$ connected across the exact number of turns necessary for antireso- nance, turns may be added to $L_0$ until the entire combination of $L_0$, $C_1$ in series with $C_2$ and the added winding of $L_0$ will antiresonate at the same resonant frequency. The exact value of $C_1$ will set the impedance of the entire arm, and $C_2$ is then fixed.

The series arm $L_0C_1$ is considered and modified in the same manner.

This now leaves only four capacitors, $C_0$, $C_1$, $C_2$, and $C_3$, to be selected to play or null 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 $L_0$ and $L_2$ to permit operation directly from a radio modulator and into the grids of a balanced tube modulator. This adds little additional cost, and power 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 $L_0$ since a copper-nickel ring modulator operates quite satisfactorily into this impedance. The impedance of the output winding of $L_0$ is a compromise be- tween 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 impedances varies directly with the inductance of the windings (with 1.4 ohms, the impedance for 1000 ohm impedance), the required inductance in millihenrys for any new impedance may be readily found by dividing the new impedance by 1000 and multiplying the result by 1.4. The number of turns required can be found from the formulas for the in- duc- tance given in Fig. 6.

Filter Alignment

As has previously been mentioned, the LC combinations must be resonant at the desired frequency. In an antiresonant filter with closely-spaced rejection frequencies, it is very important to hold to very close frequency tolerances; while a constant error is not across the spacen- ing of one frequency to the next is essential.

Hence, it has been considered necessary to use expensive laboratory equipment, which is out of the reach of many. Signal generators for the range of 10 to 1000 kilocycles are not common, while available are usually not of sufficient accuracy. The use of the BC-221 frequency meter makes the measurement very easy. 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 an oscilloscope. The BC-221 is used only as a calibration means for the test signal generator.

The test generator may very well be made with the same electronics, and the 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 plates are used to give the familiar Leatherface figure as a means of compar- ing frequencies. Since most of the frequencies noted from the test generator are out beyond that of the BC-221 it is convenient to use the chart calibration points for 125-250 kc. By mov- ing the demark point one place to the left and adjusting a 125 kc Leatherface pattern on the oscilloscope the frequency may be read directly. Other multiples of the test generator may 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 obtaining that gives accurate results is shown in Fig. 7. This method uses all the LC combinations in a series-parallel circuit. With the LC combination connected as shown, a straight dip is constantly used when the frequency is at the exact series-parallel point, since the impedance is then a minimum. Although test 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.

**Cal Construction**

In selecting inductors for the filter, the Q is of primary importance. Q values of at least 150 are recommended. The input and output windings $L_0$ and $L_2$, and the core of $C_0$ and $C_2$ are in parallel with the terminations, and losses here.

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**Fig. 7.** — The method used for checking exact-resonance conditions. A frequency-check oscillator is obtained by using a BC-221 to check the 100-kilocycle range of the test signal generator. The chart calibration points for 125-250 kc are shown, with the frequency of the tube-oscillator tuned to the desired frequencies.

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**Figs. 1 and 2.** — Diagrams used in the construction of the tube-oscillator, using a 100-kilocycle filter with a transistor is suitable with moderate outputs of 500 milliamps or less.
have much less effect. While many types of inductance might be used, the toroidal type has many advantages and core rings of molybdenum Permalloy are now available for 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 trans- formers 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 inductance low, and wind- ing is not too difficult. Two grades of core mate- rial were used in the inductors for the filter of Fig. 6 (see attenuation characteristics shown in Fig. 3). Inductors $L_1$, $L_2$, and $L_3$ are twee rings having an effective permeability of 10, producing Qs of 200 to 250 at 20 kHz. $L_4$ and $L_5$ cores were of 133 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, $L_4$ and $L_5$ have noninductive resistors of 64 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 core varies slightly, the exact number of turns will vary and must be determined by measurement. For this reason suffi- cient 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 be cut to size after the correct number of turns has been determined.

It is not necessary to wind the cores to the limits that given in Fig. 6 may be made if winding area does not limit. "Fermoral" insulation, or the equivalent, is recommended and is easy to wind, but single-cork or nylon is satisfactory. Plain enamel of the "Fermoral" type may be used, because of the possibility of shorted turns.

In order to reduce the number of times the wire must be threaded through the core rings, all windings are bipolar except that adjustments, windings 5-6 of $L_4$ and $L_5$. In the bipolar type of winding, two wires are held together and wound as one. After winding, the start of one wire (1) 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 5 and 3 and 2 and 5 are wound together and also connected so that connections are made for $L_4$ and $L_5$, the windings are series-aiding.

In winding the length of wire to be piled through for each turn may be halfed 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 necessarily critical.

The following procedure for proper identifica- tion and labeling of windings may be used:

A formed tweed coil of the type used is the soldered filter. An unwound core is shown at the left. Select one of the ends of the completed winding and arbitrarily label them 1 and 2. Now use by an ohmmeter, batte the wire at the opposite end of the winding which checks continuity to "1." This of course will be "2" and may be applied to "3." With the exception of the input and output winding of $L_4$ and $L_5$, the wire ends, 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, 6, 7, 8. 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, 2. The input windings of $L_4$ (1-4) and output windings of $L_5$ (5-8) are not critical in inductance and may be wound fast to the specified number of turns. If desired, a layer of tape may be applied over these windings before application of the second windings.

$L_4$ (1-4) is wound and resonated with a 0.5 µf, test capacitor to 1500 kHz. Adjust to the nearest turn that produces resonance closest to the exact frequency, $C_4$ may be paralleled and used temporarily for the test capacitor. The second winding of $L_4$ (5-8) is now applied and series- connected with the inner winding, 1-4. Turns are stopped to secure resonance with $C_4$ at 20 kHz. No connection is made to the tap during adjustment.

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Note that wide tolerances on C4 are allowed and the exact number of turns of L4 will depend on this tolerance.

L4 (G-H) and L4 (K-L) are wound and vacuumed with their associated capacitors, C1 and C2, L4 (K-L) is now wound and vacuumed with C3. As previously mentioned, the value of C3 is

given some separation from one another. A metal wire 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 cutoff as with any 'phone transmitter. Some high-frequency attenuation may be obtained by the action of the series capacitors, C2 and C3. It is well to remember the following precaution:

One point not obvious is that R2 and R3 with R4 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 a microphone or vacuum tube is used instead of copper oxide for the modulator, a resistor may have to be placed across filter terminals 4 and 6 to match the load to the source. R4 and R5 in vacuum tubes. Proper match may be obtained when audio is fed into the speech amplifier and varied from 200 to 3000 cycles. If the speech amplifier has previously checked that, the output from the filter at terminals 0 and S as measured with a voltmeter, should be 7 to 10 volts. The speech amplifier should be resonant in the frequency of 18.8 to 17.0 kc. A diple in output
Measuring Sideband Suppression

Howard Wright, W5PBR, suggests this simple stunt for measuring sideband suppression of your own or the other fellow's signal. It requires that you have a selectible-sideband receiver of some kind (filter, YFS-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.

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 note peaks appear as individual "spikes."

To measure sideband suppression, set the control at 0 db, and advance the A.G. 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 loud speaker) and switch to the desired sideband. Adjust the calibrated control until exactly the same amount of 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, error in reading the scope, and by the maximum possible amount of sideband suppression the selectible-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.
An Inexpensive Sideband Filter  

DAVID O. MANN, WE3LY

Although constructed, an earlier of the filter types 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 circuit. With the addition of a potentiometer, any desired amount of carrier can be inserted, and the resonant a.m. (AM minus one sideband) signal tried to establish contacts before switching to single sideband. This article is primarily intended to describe a handsoper filter suitable for use in the circuit and within the reach of even the relatively inexperienced amateur.

The Filter

The details on this very narrow, but a good filter wasn't built until after three tries at getting suitable iron for the inductances. The ferrite is generally considered by 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 ferrite cores for the inductances were cut from the core of a television horizontal-sweep transformer by boring and shaping a cylindrical section of it. The material is called "mangite iron" and was first because it was designed by me, at 45 k. I made some pretty good coils, but even construction isn't as difficult as anyone else. One of the inductances required in this last filter was 21 kh., much larger than the others for which the ferrites were used. To make up this value, using the ferrite core would have required an enormous number of turns, and this started a growth of all available coils having the larger inductance and a reasonable Q. An ECA standard television variable inductance, used in the horizontal-sweep circuit, was found to have a range of inductances between 3 and 21 kh., and a Q at 50 kh. of from 10 to 35. This served the purpose at the time, but it appeared to be such a highly components that various filter designs were compared, in an effort to find one in which it could be used throughout. Two equations resulted in the one described here. These little coils are called Horizontal, Linearity, Unit, RIA part No. 30183, and they can be obtained for less than a dollar. They contain about 1000 turns of No. 33 enameled wire, and to use them for the lower values of inductance in the filter it is necessary that they be panned, to permit keeping the slug well in the winding, and that maintain a higher Q.

Another variable in the filter design is the image impedance, and the value of 20 kh. was

From QST, March, 1949.

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selected because audio line transformers for this impedance are very reasonable and, because of high transformation ratio, they will pass 0 to 15 kHz. 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 1/4 X 5 1/2" box, no one method of assembly. It is almost certain 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 sawing grass, but the same techniques are still good and the measuring methods used shouldn't even annoy you from tackling the job. In addition to the

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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 mismatched), it does not appear the least burden to request this favor of your parts supplier should other sources fail. Audio oscillators and V.U. meters 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 audio section 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 R3 (Fig. 4), any desired amount of carrier can be by-passed around the balanced modulator to T200, and being able to do this offers two very important advantages. The first is that the carrier can be used to tune up the carrier and any following amplifiers. The second is that having the carrier-rejection control during operation permits adding enough carrier to permit the signal to be copied just like my 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 "pipe" carrier suppressed only 20 db, in view of the usual blanket of other carriers, but this arrangement is versatile enough for most requirements.

Having 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-ohm 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 T200, but if not you can test the oscillator later in another way. Connect the probe to either plate of the 500-kc. balanced modulator, and check for and maximize the sig.

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

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and from the 540-kc. crystal-modulator 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-signal and see if its intensity can be varied by the carrier-inser- tion control. RC, if so, is all 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 wish to tune to the upper sideband coming from this modulator. Let’s set the receiver to X and tune the signal just above the 540-kc. oscillator that can be controlled by RC. Shift the balance to the secondary of the 540-kc. balanced-modulator output circuit, and tune the transformers for maximum signal. Very RC 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 540-kc. characteristic, can be set with this arrangement. Watching the S-meter, set RC for minimum indication. This means that the only path for the 10-kc. oscillator signal is through the modulation in the ring modulator, T 24, and the filter. R 13 puts the sensitivity of the receiver until a reading of about 85 is obtained. R 12 is 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 output shows a fairly rapid rise to a maximum. If no pronounced rise is noted (0 to 3 units) by the time the 500-μf. 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 at a maximum, the oscillator is up in the pass band of the filter and thence has been “located.” To get the oscillator on the proper part of the Smite curve it is merely necessary to decrease the frequency (increase capacity) until the response drops 2 or 3 units.

While RC 12 is set for a minimum we don’t want any of the 10-kc. carrier coming through, and to prevent this the ring modulator has to be balanced. The ring modulator isn’t of much use if the balance will produce a carrier attenuation of around 30 db. To do the balancing, set the receiver for a good indication, 0 db or so, with the bridge still connected to the secondary of the 50-kc. modulation transformer. Carefully move the R 12 until a minimum reading is noted. Take C 24, 0.004 μf., and connect to the side of the bridge. This produces the maximum change in the signal passing through the microphone. If this reduces the signal so low that the receiver becomes barely audible, with the 10-kc. gain on the signal, when R 12 is tuned to a minimum the carrier leakage at the ring modulator can be measured. Set RC 24 to X 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 24 until the minimum obtained by varying R 12 is sharp. When good balance is obtained a hum will be audible in the receiver; i.e., the carrier is on the mark. The range-balanced leakages cause appreciable modula- tion. Ten kc. above or below the carrier fre- quency (which side depends whether a 14.7- or 13.7- kc. carrier is used in the last balanced modulator) another fairly strong signal will be found which is not controlled by R 12. This is an undesired output that comes from the 540-kc. oscillator and must be eliminated by balancing the 540-kc. modulator.

The twinpeak speed amplifier is simple enough so that little trouble should be encoun- tered. 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 10-12 kc. bandwidth 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 pass band of the filter.

Balanced I.F. Transformers

To simplify the job of the balanced modulator at 550 kc., a balanced-cell 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 coil will replace the transformer in the tank circuit and thereby eliminate any need for adjusting the inductance of the coil from its ends (A and B) to ground. The inductance of the other coil should be connected to the “cold” side of its circuit.
A Crystal-Filter SSB Exciter

T

emitter uses a quartz crystal filter operat

...ng at 450 kc. (See Fig. 1.) The filter al

...n a bandwidth of 300 to 3000 cycles, the

...outside the rejecting range, it passes the

...s distant from 375 to 325 in. A 1.388 kc.

...n the reject range is rejection less than

...culation of the crystals is marked at

...is used for frequencies between 28.0 and

...crystals are manufactured by Wester

...pended on the signal grid of the 458A for upp

...cult. Crystal “B” is 2.78 kc. higher

...than 450, or 2 channels higher in the crys

...tions as described. Crystal “C” is 1.39 kc.

...1 channel higher. Crystal “D” is 450 kc. C

...receptor diode 450 kc. Crystal “A” also at

...s. When measuring the filter diode.

Figs. QST, November, 1940.

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Fig. 3 - Block diagram of the oscillator-mount.e.h. circuit.

Fig. 4 - Recommended physical layout of the single- and double- tuned filters. The double-tuned filter is preferred when instantaneous selection of upper or lower sideband is desired.

The dashboard mounting e.h. output, complete with receiver mounting and VFO. The top slab is the output with motor removed, showing most of the "ironcore." The motor, made outside current a.e., is shown. The smaller motor is shown in the lower part where the metal handle may be seen. The 6AG7 plate tank is also shown.

Another view of the circuit, with the covers removed. The output transformer is mounted on the left-hand end, the relay and transformers are at the side, and the output tank coil is at the rear of the VFO unit.

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This filter will keep stray capacity coupling at a minimum. No shielding other than that provided by the G1 case and the output tank can is re-
quired. A is important that capacity coupling around the crystal filter be minimized—in other words, no capacitive signal must reach the 68NT 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 

filter using 5 crystals will be required. This filter is shown schematically in Fig. 5. A double-section power switch selects the upper or lower sideband. These wiper sections must be repositioned 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 diagram (Fig. 7) with specifications provide sufficient information to allow even an inexperienced amateur to construct an effi-
cient exciter.

3) A vacuum tube voltmeter is connected from one of the 68NT grids to ground. 4) Swing the signal generator through the crys-
tal range until a maximum response is noted at the voltmeter. This will indicate the setenter-frequen-
ty of crystal "C" and with the crys-
tals described, based on a 450-ke. carrier, will be approximately 448.6 ke. 5) Align all transformer trimmers for maxi-
 mum response on this frequency. 6) Next, adjust the signal generator slightly in the higher-frequency direction for a null. This will be the zero-resonant frequency of crystal "D," 430 kc. with the crystals indicated. 7) Move the signal generator 1/2 ke. lower than this null and adjust the trimmer on the input side of T3 for maximum response.

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

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

Carrie Insection

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

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Aligning the Edmunds Exciter
BOYCE S. WEBB, W4PIX

From anyone who has built a crystal-filter S.A.B. exciter of the type described by Edmunds and has had difficulty in aligning it properly, we would like to pass along a different tuning procedure. The method, described by WNEB (aided and acclimated by W4PIX), uses an audio oscillator and oscilloscope (or oscilloscope 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 voltage in output amplitude must be taken into account.

Although this method has only been used as far in aligning the Edmunds exciter, it should be readily applicable to other types of S.A.B. exciters, with a few alterations.

Alignment Procedure
The first step is to connect the audio oscillator and "scope to the exciter, as shown in Fig. 1. A shielded cord should be used between the audio oscillator and the exciter. Fig. 1 indicates capacity coupling 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.m. off. To ensure 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 pickup loop at the exciter over to the receiver antenna terminals. The antenna terminals should be short-circuited with a short piece of small (No. 22 or so) wire. If the signal is insufficient for a reasonable S-meter reading, the shielding wire can be lengthened slightly.

Using the audio oscillator and the "scope or receiver (or both), the procedure shown in Table 1 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 tapings 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 resonant 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 waving a metal 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 1, 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 exciter and sideband suppression can be measured in db, provided the S-meter calibration is known. Any S-meter can be checked

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Fig. 2.—Any receiver can be calibrated in 4-dB steps by using a signal source (such as VFO) and an audio oscillator. \( R_1 = 3900-	ext{ohm potentiometer.} \) \( R_2 = 3900-	ext{ohm resistor.} \) \( R_3 = 50-	ext{ohm resistor.} \) Start out at any 20 db. over 80 and work back down the scale, checking first the finals, interstage and then the meter. Note \( R_3. \) This method is, of course, only satisfactory at high signal levels, and all stray pickup 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 decreased as mentioned earlier. \( R_3 \) and \( A_1 \) should be mounted as close to the receiver input as possible.

When making measurements on the a.c. meter, the r.f. input to the receiver should be controlled 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.

Normally the meter is driven from a signal source at a level that may vary from 80 to 40,000 microvolts. This method is good only for relatively low modulations. To check a higher modulation, the output of the oscillator can be made to feed into the receiver through a coupling network.

TABLE I

<table>
<thead>
<tr>
<th>Step</th>
<th>Adj</th>
<th>Gain Source</th>
<th>Other Receiver with A.M. and Sharp Cut, Adjusted for . . .</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>100</td>
<td>C, C, C,</td>
<td>Microphone</td>
</tr>
<tr>
<td>2</td>
<td>300</td>
<td>C, C, C,</td>
<td>Microphone</td>
</tr>
<tr>
<td>3</td>
<td>150</td>
<td>C, C, C,</td>
<td>Average final</td>
</tr>
<tr>
<td>4</td>
<td>500</td>
<td>C, C, C,</td>
<td>Average final</td>
</tr>
<tr>
<td>5</td>
<td>1000</td>
<td>C, C, C,</td>
<td>Average final</td>
</tr>
</tbody>
</table>

The radio amateur 55
Hints on the Edmunds Exciter

Received, WEMTF, passes along some useful hints on the Edmunds exciter. To quote his letter: 

"A number of persons have trouble getting adequate carrier suppression in the exciter. The trouble appears to lie in the fact that the crystal oscillator operates on an 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 GSK triode section will take care of the sinusoid. 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 

\[ 7 = 5 \quad 12 = 20 \log 6 = 15.6 \text{ db}. \]

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 20 db, down for the carrier, represent reasonable values for amateur s.s.b. equipment.

Fig. 1 - Improvements for the Edmunds crystal oscillator exciter suggested by WEMTF. The revised oscillator circuit at A permits adjustment of the oscillator frequency and consequently better carrier injection. C1, 100,000 uf., adjustable; C2, 670 uf. and 5J, 0.1 microfarad; R, 20,000 ohms. The revised oscillator circuit does not offer as much voltage available for carrier injection, and the circuit of B gives better amplitude. C1, 1 uf.; C2, 0.05 uf.; C3, 0.001 uf.; R, 20,000 ohms; and R, 20,000 ohms (as before). \[ \text{C1, } 4 \text{, } 322; \text{ } R, 322; \text{ } C, 322; \text{ } 0, 329; \text{ } E, 321, \text{ edge-grounded in} \text{ radio frequency} \text{ crystal band.} \]

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my exciter I limit the audio input to 1 volt, how-
ever, I got the proper level for the two-tone test
patent.
When the Pierce oscillator was first used like
this, the amount of carrier at the output of the
carrier-emitter amplifier was inadequate, and
so a simple change was required in the circuit, as
shown in Fig. 1B. With this arrangement, suffi-
cient carrier insertion 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
been beneficial. These crystals, placed in paral-
lel with the normal suppressor crystals, will
provide more uniform suppression of the un-
wanted sideband and can even be used to control
the 0-300-cycle portion of the wanted sideband
to obtain a more desirable speech-modulation
characteristic. The additional crystals can be
added either in parallel with the sideband-sup-
pressor crystal, A, of Fig. 1C, or in parallel with
the carrier-suppression crystal, C. Examples of
such add-on crystals are given in the sketch,
but any crystals in the range can be used, pro-
vided they are for adjacent channels and the
same relationship is retained. The use of multiple

Balanced Modulator

G. O. Kincaid, VE3CN, uses an interesting balanced-modulator circuit in his version of the
Edmunds exciter. Shown in Fig. 2, it has the addi-
tional feature of providing for p.m. The oscillator
circuit starts (rather than the original) and it can be
"pulled" by tuning, for better alignment with
the filter. The filter is first locked into the oscil-
lator crystal by removing the oscillator crys-
tal, introducing a RO-251 signal across C2, 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 T2.
The p.m. is obtained by tuning the secondary of T3 and the volt-
age fed to the carrier-amplifier tube is 90 degrees different
than that applied to the balanced mod-
ulator. For p.m., the crystal filter
must be out of the circuit, of course,
and since the p.m. signal is ob-
tained by comparing the broad-
sidebands (less carrier) with a
carrier that has been shifted
90 degrees. Checking with a scope, one simply

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sets the trimmer in the secondary until there is minimum amplitude change on the meter with modulator. The two circuits of $Y_1$ "pull" a little, so the complete adjustment may require several trial runs.

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An Oscillator for the Edmuns Exciter

Two of the problems of construction of an Edmuns crystal-filter exciter are finding a suitable test oscillator and accurately setting the oscillator frequency in the filter notch. Woody Davy, W7CJS, solves the problem by using the oscillator circuit shown in Fig. 3A (page 57). When the 470-ma. 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 out 400 cycles by tuning the 180-ma. variable.

Woody uses the simple 12s-epsilon waveverter of Fig. 3B to indicate output at the 6AG7 plate circuit — he likes it a lot better than a v.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 Crabbe, W2FUV, thinks many of the fellows with Edmuns 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 elaboration, since the technique involved are standard receiver and low-level transmitter practices.

And while we're talking about the Edmuns exciter, Harold Khier, 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 490 kc., using an oscillator at either 5400 or 4550 kc., will put your (suppressed) carrier at 3900 kc., with upper sideband in one case and lower sideband in the other.

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SINGLE SIDEBAND FOR
Crystal Lattice Filters

C. E. WEAVER, W2AZW, AND J. N. BROWN, W8SHY, EX-W4OQL

The ability of a receiver 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 Q's of the crystals themselves. In some cases, the crystals yield Q's 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 resonant frequency of the equivalent circuit is plotted for all frequencies between zero and infinity. The series-resonant frequency, $f_s$, occurs first when the curve crosses the zero-reactance line, and the parallel-resonant (nonskirt) point, $f_p$, occurs where the line rises to high values of inductive reactance (−90°) and then breaks sharply through zero to a high capacitive (−90°) 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 resonances.

Fig. 2.—Reactance plotted against frequency for a crystal damped by an inductor.

Fig. 1.—Schematic diagram of a two-section crystal lattice filter.

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use of radio-frequency transformers (ordinary i.f. transformers) as input and output devices as well as spreading coils for $f_1$ and $f_2$. It might be well to mention at this point that when $f_1$ and $f_2$ are spread, $f_2$ remains fixed in frequency and only $f_1$ is moved higher in frequency. Let us briefly consider what happens inside the lattice filter. Assume that the pair of crystals connected in shunt (i.e., connected in series to the output of a cathode follower) are identical and have a 2 or 3 kc. higher in frequency than the pair of identical crystals con• nected in series (horizontally connected). Also assume that the coils used have spread the $f_1$ and $f_2$ of each crystal. Any overmoding can be cor• rected by the i.f. transformer tuning condensers, provided the crystals are exactly paired. See later section on filter alignment.1 A of Fig. 4 shows the reactance plot for both sets of crystals, through the two possible paths of the bridge will come out. When the reactance of the opposite sign there will be partial transmission through the network with the maximum signal transmitted in the middle at the points where the re• tances are equal in magnitude, but still opposite in sign.

Practical Filters

A waveguide filter can be constructed at the cost of only a very few dollars! The FT-241-A low-frequency surplus crystal set is built with very good success. Very inexpensive coupling devices were used, i.e., ordinary replacement i.f. transformers (Microwave No. 10-5712). There is one condition made that is probably due to an improper choice of transformers or an impedance mismatch between crystals and trans• formers. 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 hinder 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 un• doubtedly be helpful at this point in the 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 band• pass. For a 5- or 6-kc. bandwidth 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 bandwidth of 2.5 or 3 kc, the channel numbers should be consecu• tive; 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 close to the same fre• quency as possible. The pairs should be selected for close or nearly equal. 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. low sweep to the signal generator start and end the frequency of the crystal, and you will dis• cover that there will be a small indication for an randomly chosen frequency. As the generator frequency is swept 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, $f_2$ and the null was the parallel or antiresonant fre• quency, $f_1$. After considerable difficulty in equalizing, it will be possible to match pairs of crystals using this method. Thus when proceeding to the lower one of a pair of crystals will fix this matching problem. But be careful only one or two very light snips on the five-gma side of a new 50-ohm tinned wire. And take care, you can break it--and it sounds worse than it actually is. What happens if these crystals are not closely matched?
shielding is recommended when serious trouble is encountered.

When the narrow-band lattice filter is used for transmitting, as described in the article on page 6, 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 100-kc. frequency meter or equivalent calibrated source of rf energy covering the range of 100 to 500 kc., a low-frequency receiver such as the BC-568, BC-663, or a parasitic adapter whose signal covers the frequency range we are concerned with. In lieu of the receiver or parasitic 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 5-meter as a tumbler makes the wattmeter a valuable aid and sometimes permits an accurate reading of the signal strength in a given crystal.

Specific step-by-step adjustments for alignment of these filters will not be given in this article. They would be long and somewhat consuming and require patience on the part of the operator. Instead, a few pointers will be given, and we have faith that the old ham instinct will fill in the rest. The best step is to peak the i.f. transformers for the highest frequency of the filter. It may be necessary to align each roughly with the signal source and then adjust the trimmer to obtain a smooth slope. If you have been careful in the matching of the crystals, the passband will be fairly well defined. Misalignment of these pairs of

**Fig. 5** — Attenuation characteristic of an experimental crystal lattice filter (two sections) suitable for receiver use. The crystals were Channels 49 and 50 of the FT-241 series.

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

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A Crystal Lattice SSB Exciter


The exciter consists of a majority of the essential features which most SSB 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-controlled or other appendages to make it complete. These features have been included in a unit that is comparatively simple and easily fixed up with test equipment usually available to most hams 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 so adjusted to cover the frequency range required by the exciter. This is done by making enough capacity to the existing tuned circuits of the VFO so that it will tune approximately 175 kc. lower than the phone band to be used. For 3.5- to 4.0-Mc. operation, this will be in the vicinity of 3.253 to 3.288 Mc. The exact frequency range will depend on the choice of crystals for the lattice filter in the exciter.

A complete s.s.b. exciter made the crystal lattice filter. The filter is designed to provide a narrow bandpass compatible with the low-frequency oscillation circuits and the sideband selector switch.

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<thead>
<tr>
<th>Step</th>
<th>Condition</th>
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<tbody>
<tr>
<td>1</td>
<td>None</td>
<td>None</td>
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<td>2</td>
<td>None</td>
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<td>3</td>
<td>None</td>
<td>None</td>
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<td>4</td>
<td>None</td>
<td>None</td>
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<tr>
<td>5</td>
<td>None</td>
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<tr>
<td>6</td>
<td>None</td>
<td>None</td>
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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 pickup. 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 capacitance across the crystals. Tests on several sets of crystals indicated that the capacitance 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 400 kc. and 400 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 ohm impedance, which feeds into the balanced modulator (from crystal diode IN521) to combine with the carrier from oscillator $V_2$. This produces an amplitude-modulated signal with the carrier suppressed. The resulting upper and lower sidebands feed into $V_3$ through an impedance-transforming network consisting of $C_6$ and the trimmer condenser in series with the primary winding of $T_2$. As the signal passes through the filter, one sideband is removed. $V_4$ amplifies the remaining side-band signal from the filter output and feeds it into the r.f. input grid of $V_4$, where it is mixed with an external VFO to produce the final frequency. The resulting $V_4$ is coupled to amplifier tube $V_5$ through a 4-Mc.

Fig. 3 - A suggested chassis arrangement for the crystal filter.

If $V_1$ is a 6AG7 operating Class A and will deliver approximately 3 watts peak power into a 600-ohm load, $T_1$ and $T_2$ are enclosed in Nortronics plug-in units so that transformers can be constructed to cover other bands if desired. The coil values given here are for the 25-meter 'phone band only.

Voice Control

$V_3$ is the additional audio amplifier for the voice-control circuit. $V_3$ is a 6bf6 that serves as the voice-control tube. The secondary of $T_2$ connects to the 150-ohm resistance in series for voltage and back resistance considerations) and $R_9$ and $C_2$. Potentiometer $R_9$ 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 $V_3$ has zero bias, thus permitting it to conduct. This in turn causes a voltage drop across $R_1$ and $R_2$ that biases $V_3$, $V_4$, and $V_5$ 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 unconnected, $V_3$ and $V_4$ amplify the audio, which in turn is rectified by diode D2. The resultant voltage is used to bias the grid of $V_4$ negative and cut off the plate current. This turns on the exciter by placing normal operating bias on $V_3$, $V_4$, and $V_5$ from voltage divider $R_1$, $R_2$, and $R_9$.

If the receiver control unit, Fig. 4, is connected to the transmitter voice control circuit, $V_5$, Fig. 1, the receiver will be automatically disabled during transmissions and returned to receiving condition when the operator stops talking. Referring again to Fig. 1, the clipping diode 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

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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 \( R_5 \) and \( C_5 \) network. The network values shown are about optimum, but the "hangover" time can be changed to suit the individ- uals by the proper choice of values for \( R_5 \) and \( C_5 \). The transmitter and receiver control circuits are timed so that the receiver is com- pletely 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 simulta- neously equipped and operating on the same fre- quency, without echoes 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 fre- quency, 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. Oc- casionally, one may be found that deviates 100 cycles or more. If this occurs, it should be re- placed with another crystal, or the lower-fre- quency one can be edgewise slightly until it matches the higher-frequency crystal of the pair.

**Operation**

The exciter can be used directly as a lower- power transmitter, but it is not recommended, particularly on 75 meters. A very variable job can be done with a pair of Class B 805's following the exciter. If everything is properly lined up, the unwanted sideband at the output of the exci- ter will be suppressed 50 db. or more and dis- tortion products should be quite less than 35 db. down at 3 watts output. The linear amplifier following the exciter will then be likely de- termine the amount of distortion products radiated by the antenna.

The authors are indebted to W. B. Flock, W2EAN, for his many constructive sug- gestions during the preliminary work on the crys- tal filter and subsequent design of the exciter.
Half-Lattice Crystal Filters

WILLIAM E. GOOD, WACVI

A lattice-type filter with quartz crystals can have a bandwidth of twice the spacing between the zero- and parallel-resonant frequencies of one of the crystals. In an X-cut crystal this separation may be 1.5 to 2 kc, at 855 kc, so that a bandwidth of the order of 3 to 4 kc should be possible. A circuit equivalent to the four-crystal lattice is the crystal 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 328 to 500 kc. The labels on the crystals in one group range from 200.2 to 27.9 Mc, in 0.1-Mc steps—the fundamental frequency is found by dividing the label frequency by 34, which gives 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 52.

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 crystals 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 resonant to the crystal frequency and C3 just balancing the crystal squarer. As C3 is made smaller, a rejection notch (parallel resonant frequency) moves in from the 3d frequency. Conversely, as C3 is made larger than the balance capacity, the same moves in from the low-frequency side. By replacing C1 with X3 (a crystal 1.8 kc high), the curve of Fig. 2B is obtained.

The squat 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 4-dB dip was noticed between the two peaks. The peaks are approximately 4.5 db, and one may be slightly higher than the other. The slits

are about 10 kc, wide at 60 db down. The dotted curve is the LF passband without the filter. If it is trimmed, C3 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 C3 is increased.

Fig. 2D is the same as Fig. 2C except that the value of C2 is larger and it shows how the side requirements tend to rise higher as the notches come

closer together. The general result is that the sides of the response curve become steeper as C3 is increased, without appreciably affecting the separation of the two peaks. Practical values for C3 are around 1 or 2 sec, obtained roughly by bridging together two short pieces of insulated wire.

It should be emphasized that in every case LC is tuned to resonance of the crystal at 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 C3 is placed across the lower-frequency crystal, as shown in Fig. 2E, the skirts widen out.

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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 fre-
cuency than X₂ and X₁, are placed in shunt ac-
cross the circuit of Fig. 2B, the result is as shown
in Fig. 2F. Their capacitance is compensated for by
adding C slightly. The two notches appear at the
series resonant frequencies of the two new
crystals. Essentially, they are series-resonant
gaps shorting out LC at their resonant frequen-
cies. The effect, that of deepening 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 C₁ is added to X₁ in the combination shown
in Fig. 2F, the original pair of notches may be
introduced and their frequency set to reduce the
size of the extra hump of F, as shown in Fig.
2G. C₁ may be increased in height, the new notches
inside those caused by X₂ and X₁. 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
RMC-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 hand-
spread dial for frequency indications. The hand-
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 RMC-69 has a
center-tapped secondary (c, in Fig. 2). However,
there is no reason to believe that a straight

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Exciters Using Cascaded Half-Lattice Crystal Filters

The half-lattice, or zig-zag, crystal lattice is used by WTBMF. It has several novel features in the exciter. For example, the balanced-modulator circuit (adapted from Motorola) doesn’t require push-pull inputs of any kind, an advantage in 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 50-kc output of this unit to the operating frequency — VHF output is fed to the two cathodes and the 50-kc a.c. signal goes to one grid.

The crystal filter uses adjacent-channel crystals (£2 and £3). Tolerant of a single section of the filter show about 35-db rejection, and the two cascaded sections measure up around 50 db. The final at WTBMF uses p.p. 56G, triode-connected with the grids and screen connected together and operated at zero bias. Plate voltage is 750.

The modulator has an entirely different circuit that is pulled along by Fred Hull, WTAMU. It borrows ideas from several designs. The circuit is given in Fig. 2. The transformer £4 is operated in the primary for a series-tuned connection, where the variation (four germanium diodes) modulator wants to look into a low impedance. The “bifilar” windings on £6 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. 3 of Mon's "An inexpensive Side-

The filter was aligned with a signal generator, inserted at the audio-input terminal (modulator turned off). The first step was to align all circuits to the mid-frequency between £2 and £3. The signal generator is then set to a frequency about 3 kc higher (or lower), and £2 and £3 are adjusted simultaneously until a very sharp null is obtained. £1 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, 0.1% area of improvement of voice quality, however, and it does improve the signal-to-noise ratio. The adjustments may require several go-arounds, but of the audio block slightly.

From April 1953, the June 1954, and June 1958.
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 the interstage coupling capacitors.

While checks on 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, with another operator monitoring the signal on a low-sensitivity receiver. The carrier frequency should be adjusted for maximum intelligibility. This will not necessarily 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.

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Cascaded Half-Lattice Crystal Filters for Phone Reception

ROBERT L. MORRISON, W7SEW

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 cascaded arrangement in the main receiver at W7SEW—a proven Super-Per. Despite its age this model of the Per is an effective low-pass receiver with 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 that skews both skirts of the selectivity curve are steep, and the operation is more “single-band” than that of conventional crystal filters. The filter circuit herein described can be applied to other receivers, possibly with more beneficial effect.

Mounting

The photo shows how the filters are mounted in the space occupied by the original filter. There are two cascaded filters, one for phone with a i.f. 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 opening of the two gaps of the switch was cut down, the mounting plate extended about an eighth inch beyond the front panel, so a second pair, with a hole in it large enough to receive the switch, was used to fill the gap. A thin piece of aluminum shields the two sections of the switch from such omissions 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 feed from the first i.f. transformer to the first switch section from the wiring of the second switch section immediately above. The amount of shielding might seem scarcely enough, but no amplifier instability or adverse effects on i.f. selectivity characteristics were noted.

Circuit

The basic filter circuit for each stage is that of Bill Gerlach, W7IV. The inductions in the Per i.f. transformers is enough to allow the use of low-C tuned circuits which are necessary for half-lattice filters. The complete circuit is shown in Fig. 1. Balanced-ground i.f. transformer construction is obtained by moving off all but three outer 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—a effect explained by the rise top of the selectivity. The interstage condenser includes a provision of negative and zero temperature coefficient.

Front panel of the receiver with the new filter arrangement replacing the original crystal filter. The switch selecting the two filters and the two small boxes containing the crystals are in their original positions. The use of transformers instead of the lower i.f.-band condenser has been “blanked out” since it has been fixed in position during the alignment process.

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tank condensers specified in Fig. 1 sufficiently minimize 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. trans- formers and reassembling them, repeatably those with variable coupling, is no fun.

The original i.f. transformer can be used by removing the inner shield can which contains the primary circuit, and rewiring both primary and secondary coils by ﬁtting 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 new and ceramic condensers on the coil boards. The second i.f. transformer has only the secondary (upper) coil reassembled as just described. (A lip on reassembling this transformer: Wrap several turns of string around the upper end of the ridge rod and in 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 station.

The shaded cut leading to the ﬁlter from the second i.f. transformer, along with the low-capacity grid leads for the 967 and second i.f. tubes, can be seen in the back-view photo. Up to 20 µf. capacity in the grid leads doesn’t do much to the ﬁlter except lower the grid-

larger capacity will lower the terminal im-

Fig. 1 — Circuit of second half-lattice crystal ﬁlter as applied to the i.f. ampliﬁer of the SP-136X. (Unmarked components are identical to those in Fig. 1 of the 1.l.f. circuit.)

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Crystal Triminity Capacity

With filters in cascade, crystal triminity capacity is not as critical as with a single stage. This can be seen by taking the example of a half-wave 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 addition of 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 does to the first stage’s peak. Hence in cascaded filters we can use relatively large triminity capacity to produce steeper skirts without introducing lobes at the cost for a step-down filter. Triminity capacity is set only in the “phone filters,” as indicated in the diagram, and consists of a series of high cut-off networks at each stage as shown in the photo.

Crystal

FT-241: A series crystals used are each, when loaded frequency is the 72nd harmonic of the crystal frequency and can be directly used in the “phone filters” because alternate crystals are 2.8 kc. apart; the selectivity curve will thus be about 0.3 kc. wider than that shown in Fig. 1. Crystals in the 54th-harmonic series can be used, but must be shifted by placing or grading so that they are at 2.5 kc. apart; the nominal 1.8 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 bandwidth for both “phone” and r.f. without requiring any crystal modification. This particular set of crystals also brackets the 40 kc. receiver if virtually nicely; if a set of crystals is used which has the desired frequency separation, but has not bracketed the original r.f., it will be necessary to touch up the u.l. alignment if the excitation of the main tuning dial is to be held. Realignment of the hand 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-wave filter, the procedure and test equipment descibed in this article can be used. In this case make the following preliminary setting: filter switch to “phone” position, bandwidth control variable (r.f. coupling) to 27, a.c. switch to MANUAL, b.c.o. switch to MOD, SENSITIVITY (i.e., gain) between 2 and 6, bandwidth to one of the higher-frequency bands, audio gain to zero, b.c.o. oscillator tube out, high-noise audio d.c. voltmeter (v.v.m. on 10-volt scale, or 20,000 ohms 1-3 volt meter on 2.5-volt scale) across second detector load resistor (PHONO terminal), and signal generator midway between the series-resistance frequencies of crystals A and B. Clip the signal generator through a blocking condenser to the grid of the third r.f. tube and adjust the trimmers of the fourth and fifth i.f. transformers for maximum output. Next clip

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TABLE I

<table>
<thead>
<tr>
<th>Crystal A</th>
<th>Channel</th>
<th>C. A.</th>
<th>Crystal B</th>
<th>Channel</th>
<th>C. B.</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>334</td>
<td>611</td>
<td>350</td>
<td>611</td>
<td>25</td>
</tr>
<tr>
<td>2</td>
<td>334</td>
<td>611</td>
<td>350</td>
<td>611</td>
<td>25</td>
</tr>
<tr>
<td>3</td>
<td>333</td>
<td>612</td>
<td>351</td>
<td>612</td>
<td>25</td>
</tr>
<tr>
<td>4</td>
<td>331</td>
<td>613</td>
<td>352</td>
<td>613</td>
<td>25</td>
</tr>
</tbody>
</table>

Crystal frequencies are measured and may be parallel resonance shifted of frequency divided by 72 for the 30-hertz oscillator and by 14 for the 60-hertz oscillator to arrive at the audio frequencies for initial crystals used. The trimmer noted and the difference frequency, D = A - C, and E = B - C, must be as flat as possible in order to obtain the desired bandwidth.

If the 114-hertz c.o. bandwidth chart in Fig. 1 is followed, C = D should be 0.3 kc.

SINGLE SIDEBAND FOR
Several European amateurs have used high-frequency crystal filters — at $2$ and $3$ Mc. — in their s.h.f. generators, but the system has apparently never been tried in the U. S.
Alignment of Half-Lattice Filters

HOWARD L. MORRISON, W7ESM

AC-312 was modified by adding a half-lattice filter of the type described by W2CWI, since crystals in the neighborhood of 470 ke. were available. These are the FT-21A series, labeled according to channel numbers and transmitter (not crystal) frequency. Of crystals in the group whose labeled frequency is the 6th harmonic of the crystal frequency, Channel 33

![Diagram of filter circuit](image)

(Fig. 3) - A - Basic dual-crystal filter circuit. The trimmer, C, is discussed in the text.

B - Basic circuit applied to the AC-312. Components are marked the same as in the original receiver. C is discussed in the text. Confidence marked C should be non-resonant-carrier ceramic ceramic or silver-xilene. Values between 190 and 400 µf, are satisfactory, but both should have the same capacitance.

(405.5 ke.) and 51 (470.4 ke.), are closest to the 470 ke. if. of the AC-312. Channel 54 and 55 (472.2 ke.) are also suitable. Of crystals in the group whose labeled frequency is the 22nd harmonic, Channel 338 (469.4 ke.) and 340 (472.2 ke.) are the most suitable, but will provide a wider bandwidth, since their frequencies are 2.8 ke. apart as compared to the 1.9-kc. separation.


SINGLE SIDEBAND FOR
without cutting a one-inch hole in the latter. This procedure is necessary for most any coil if a BC-610 is modified, because the model has a smaller 0.25" core. If a slug-tuned coil is used, the slug should be insulated from the core so as to minimize any unbalanced capacity to ground.

**Transformer Condenser Considerations**

If the physical arrangement and the construction details mentioned above are followed the trimmer C1, 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, W3CV1, for example, used between one and two µfd. across the high-frequency crystal to obtain the best characteristics. Just how slight the trimmer capacity used be to cause large changes in skirt形状 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-8/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 1/8" length. This probably represents a capacity of less than 1.2 µfd.

**Coupled Crystals**

- The method of increasing skirt selectivity by crystals situated across the tuned circuit ahead of select crystals did not seem very effective when tried on two different receivers. The skirts became a little steeper, but the sidebands appeared at the antinomeant frequencies of the select crystals, and were sometimes only 40 db. down. Perhaps other shunt- ing crystals with resonant fre- quencies equal to the antinomeant frequencies of the crystals closest to the position 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.

**Alignment**

The alignment procedure is easy if you have a test oscillator or signal generator with plenty of bandwidth in the neighborhood of 470 kc. A BC-221 or LM-17 frequency meter is ideal, but the average homebrew'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 receiver is readily made. Such an oscillator should be tuned to 470 kc. by an air, sleeve-input, or ceramic condenser, together with a 25 µfd. band-

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**The Radio Amateur**

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or two-turn pick-up coil loosely coupled to the oscillator tank. Since the handspool tuning con-
ductor 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.m. stations) are sufficient to
determine the standard-line calibration curve
for the oscillator.
A sensitive (20,000 ohm/volt or a v.v.m.)
d.c. meter is necessary for measuring i.f. amplifier
output. It is connected across the plate load
resistance of the receiver, and should be kept
on the lowest scale in the case of a 20,000 ohm/
volt meter, or on the 10-volt scale of a v.v.m.,
for preventing overloading the i.f. amplifier.
The pointer can be reset to zero with the receiver
turned off, in order to back out the small voltage
developed by the diode emission current. The i.f.
oscillator tubes in the receiver should be
removed, the b.f.a. turned off, the manual gain control
used, and the handwinder 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 load from the test oscillator to the 6X7 grid,
and isolating this junction from the tuned cir-
cuit by a 50,000 ohm resistor. As the oscillator is
turned from lower to higher frequency, a sharp
rise in the output meter reading will occur at
the series-resonant frequency of the crystal, fol-
lowed by a dip which indicates the antiresonant
point. This dip can also be used to check closely
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 5 to 1. A
defective crystal is indicated by a small ratio of
peak to dip, and such should not be expected
among quantities of crystals at bargain prices.
Crystals which are stamped "Limited Test" are
not necessarily low Q however.
If the oscillator still connected to the 6X7 grid, set it midway between the crystal
antiresonant frequencies and align the 2nd
and 3rd i.f. transformers for maximum output.
Modify the 1st i.f. transformer as shown in Fig.
18 with 200-ohm condensers (as low as 150 ohm
are used) and a total series coil, and reconnect
the original grid lead to the 1st i.f. 6K7. Then
copy the oscillator to the grid lead of the
i.f. meter and adjust or prune the series coil
until peak response is obtained at the aligned
frequency with the core in the main secondary
coil about half full. 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 arrange in the core can be due to increased
coupling to the primary and not to lowering the
resonant frequency to bring it nearer the align-
ment frequency, as might be thought. The best
indication of which direction to head with the
series coil is by bad by using a small (around 10
microh) test condenser. If the output decreases
when this condenser is touched to the 6K7 grid
and ground, the series coil has too much inductance,
and vice versa.
Next connect up the two crystals as in Fig.
18b, and with the test oscillator on the alignment
frequency, peak both primary and secondary of
the test i.f. transformer. Then tune the test oscil-
lator between 490 and 490 khz, and note the two
maximum-response peaks. The frequency exactly
midway between them is the final alignment fre-
quency, and the entire i.f. amplifier should be
carefully tuned for maximum output at this fre-
quency. Again check the two peaks; they should
now be equal within a few per cent, and the de-
pression between should be around 70 per cent
of the peak. If one peak is noticeably lower than
the other, the core can be very slightly readjust-
ed so as to favor it.
Measurement
The following method is recommended for
finding the average Q of a resonant circuit.
Connect the oscillator set on one of the
peaks, and the manual volume control at about
2.5. Adjust the i.f. oscillator output until the
output meter reads at some divisions near full
scale which can be taken as "10". (For example,
on some 800 khz receivers, the zero scales
are "5" when full scale is reached with volt-
age ratio, the actual voltage measured is
simply 10 times as large at all voltages and applied
proportionately. Tune the oscillator from the peak
until the meter reads "14" 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
"10." Observations are made in the range in which
the side lobes can be measured.
The first is the "I" point, which corresponds
"0.1" to 80 db.
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Intermediate values are calculated according to the formula

\[ d, = 20 \log_{10} \frac{v}{v_0} \]

in which \( v \) is taken as 1000 if full-scale \( v_0 \) is 10, because we are two decades (100 times) down from our original starting point.

If any scale lies in greater than \( \frac{1}{3} \) on this scale (0 db.), a slight amount of hum residue capacity should be added across the low-frequency crystal, if the neat, nulls are more than 10 db., apart, hum-residue 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 19 kc., apart due to manufacturing tolerances. It can also happen that the only crystals available are not in the vicinity of 400 kc. Two procedures can be used to remedy these situations: plating and edge grounding.

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

In all cases the electrophating 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 clings the edge of the tumbler and extends the depth of the solution; then, connect it to a 350-volt resistor (or another) to the positive terminal of a 1.5 volt dry cell or flashlight battery. The control of the frequency change is had by thus limiting the plating current, and the plating is more uniform. Without the resistor, repeated "dimplings" will cause a noticeable thickest 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 1 kc. can be obtained without secondly lowering the \( Q \). Because of differences in solution of time, frequency-change data are given, but a preliminary signal transmission and frequency check will provide a basis for estimating the total time needed. The nice thing about plating is that you go too far; you merely reverse the electrolyte polarity and take off some of the plating. With a few tinfoil, a crystal can be jockeyed around to just where you want it. After removing the crystals from the plating solution, it is important to time 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 grounding. 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 the 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 wire; fasten an orbital socket to the beam; 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 base of the soldering iron can be solidly rested. Having carefully potted off the holder holder, push the crystal in the block socket and apply the tip of a well-tinned iron to one junction of crystal wire and solder wire, and with tweezers hold the latter away from the former until both have cooled. Each crystal must be used to smooth the crystal; once they concommit 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 at an angle of 20 degrees and moved back and forth with fairly pressing pressure on a 100 r.p.m. drill or with a file of similar shape. 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 edges checked by sliding the edge against the sky. As the crystal approaches the diameter change, it should be broken off by the method already described. Low \( Q \) due is due to lack of squareness and square edge grinding the edges unevenly. The necessity for reworking the crystal is indicated by the following: the frequency and \( Q \) are checked in an unknown coil. The crystal is recounted by a reverse of the taking-out process. After meaning, it seems should be tied 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 let it bother you. Drop the component having the electrical values specified will readily lead to anything but satisfaction. To the best of our knowledge, the part the author used is the only one that will work, but I'll tell you so.
Modifying the S-40 for S.S.B. Reception

EDWARD H. SOMMERFELD, W5GFT

A common way of modifying a receiver whether or not the receiver is already in the S.S.B. class is to add a "bandpass filter" to improve the selectivity of the receiver. This is done by modifying the filter section of the receiver to make it more selective. The bandpass filter is placed in the audio output circuit of the receiver to filter out unwanted frequencies and improve the selectivity of the receiver.

The Bandpass Filter

To obtain the desirable sharp filter characteristics, a half-wave crystal filter was added. The crystal filter consists of a single crystal and is mounted in a metal case. The crystal is tuned to a specific frequency and is connected to the input and output of the filter. The filter is tuned to the desired frequency by adjusting the crystal frequency.

![Image of a crystal filter with frequency tuning knob](image)

Fig. 1 - The bandpass crystal filter is inserted between the first and second IF amplifier stages.

1. Channel 1: 280kc.
2. Channel 2: 560 kc.

Crystals of these exact frequencies are not required, but the same relative channel (frequency) intervals should be used.

The grid-leak load was removed from the grid of the first IF amplifier stage. The circuit was shown in Fig. 1. A metal bracket for holding the crystal filter is shown in Fig. 2.

![Image of a circuit diagram](image)

Fig. 2 - Wiring diagram of the new third I.F. amplifier stage, T4. T6: 6SK7, output transformer (Melbourne 16-669). The voltage of the transformer's mounting bolts, it is necessary to remove the B-motor socket, SO. From the rear of the chassis. Wires that had used the pins of socket S9 as tie-points were replaced by wires running straight through.

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 F7 socket was removed for the new 6SK7 amplifier (the F7 functions were taken over the additional change to be described later), and the old F7 socket was used.

![Image of a circuit diagram](image)

Fig. 3 - The modified 2nd detector and mixer-limiting circuits use INE crystal diodes.

S1 - As originally in receiver.

S2 - 80 SINGLE SIDEBAND FOR
Fig. 4—Circuit diagram of (A) the crystal-controlled I.F. amplifier and (B) the self-contained I.F. amplifier.

A 1.6-V 150-volt 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 knob. The circuit was changed as shown in Fig. 5.

Tuning Indicator
A tuning eye was added for ease of alignment. It is a standard 435-circuit with the grid tied to the v.e.x. bus. The mounting bracket for the tube was mounted in the upper front-hatch center of the panel.

Alignment
The I.F. amplifier was aligned by plugging a Channel 347 crystal into the b.f.o. and proceding to align the I.F. transformers at this frequency. If other crystals were used in the filter, requiring that the I.F. amplifier be aligned 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 Plate 8 of V4. Just bring the load near the meter of the tuning condenser C10 until maxium occurs in most cases. The high-frequency oscillator can be turned off during this procedure by shorting C20. 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 (R4) and a few leads.

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Notes on a Specialized 'Phone Receiver

Robert W. Ehrlich, W4CUU, EX-W2NR

Writing 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
fulfilling 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 ganging tuning. These features require engineer-
ing components all along the line, and accurate
equipment is needed to get even fair per-
formance. Fortunately, the amateur who builds
his own receiver is in a unique position to bypass
all these problems by designing his receiver just
to cover his favorite ham band, relying on
components that would be out of the question com-
erically. And he can take advantage of the
least available techniques that usually take years
to find their way into commercial products. As
an example of this last item, the new familiar
Qvar was first described in QST in 1947, yet it
was about five years later that receivers incor-
porating this degree of selectivity first appeared
on the market. Meanwhile, still better selective
systems were desired.

A receiver is described here that illustrates the
principle 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 fea-
tures might offer some helpful ideas to the ama-
teur who is seriously interested in getting the
most out of his favorite band.

The Circuit

A block diagram of the receiver is shown in
Fig. 1. The circuit was designed for 75-90 meters,
with primary emphasis on the reception of single-

Fig. 1 — Block diagram of the homemade receiver.
Wideband signals. Here, stability is the first requirement — the receiver should be exceptionally stable and capable of being tuned just a few cents at a time. To take full advantage of a. c. communication, the selectivity should be high — high enough to accommodate just one station 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, the circuit is perfectly ordinary, but the mechanical arrangement for tuning are a departure from the usual complex assembly of gears and shafts. The two v.f. circuits are geared together with a highly variable condenser added across each section to establish the right ratio of minimum to maximum capacitance. The variable-condenser tuning is provided for tuning the band. The tuned circuits have damped tuned coils, making it easy to set them up together. This condenser is brought out to a panel knob that works about like the old antenna trimmers; 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 Chapp 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 J.C.-108 transmitter chassis, which provided an excellent main tuning condenser and squared dial to go with it. In addition to the main bandswitching fixed condensers, a 10-pf. variable was also added to provide a 45-kc. vernier adjustment on the front panel. This has proved to be very helpful in actual receiver operation.

Following the oscillator, an amplification circuit is found necessary to get enough drive for the 6SK7 mixer. This amplifier is fixed-tuned and peaked near the high-frequency end of the band, to compensate for the tendency of the Chapp oscillator to free 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 Waer and Bower, using eight crystal of the 26-Sigle notes. A switching circuit, shown in Fig. 2, enables the filter to be cut out when desired but still leaves the i.f. transformers in the circuit to retain modern selectivity. The coupling resistor, , 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

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**Figure 2 — Crystal filter and switching circuit.**

- $C_1 = 0.005$ pf. ceramic.
- $C_2 = 0.01$ pf. ceramic.
- $C_3 = 0.1$ pf.

- $R_1 = 3000$ ohms.
- $R_2 = 3.5$ megohms.

- $F_{1}$, $F_{2}$ — Crystal-lattice filter sections. See text.
- $T_{1}$, $T_{2}$, $T_{3}$ — 450-kc. iron-core audio-frequency transformers.

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THE RADIO AMATEUR

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be designed to contribute no stray capacitance path around the filter when it is being tuned.

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

After two stages of amplification, the L.F. system splits into two branches. One branch feeds a carrier-type demodulator, using a 606 tube, for detection of c.w. and a.m. signals. With this kind of detector, shown in the upper portion of Fig. 2, no intermediate 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 h.f. signal (about 20 volts) injected into the grid.

The second branch feeds a combination i.f. and a.v.c. detector and a.v.c. system using a 6BY2 double diode. This signal, together with the switching arrangement for the two detectors, is shown in the lower part of Fig. 2. Of particular interest is one 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 a.m. signals. This diode causes the a.v.c. to charge up quickly but discharge slowly, so that in effect the a.v.c. line "hangs up" and rides with the peaks of the received c.w. or a.m. 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 demodulation 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 L.F.S.

Coming to the matter of sideband selection, nothing further would need to be done if only a.m. signals, wide 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 limbs of the crystal filter response. Since the filter has a flat-topped characteristic, the signal remains perfectly intelligible over a

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**Fig. 2—The L.F. branch amplifiers, detector, a.v.c. and switching circuits.**

- **C = 100 μuf.**
- **C1, C2, C3, C4, C5, C6, C7, C8, C9, C10, C11, C12; 0.01 μuf. ceramic.**
- **C13, C14; 1 μuf.**
- **C15, C16; 50 μuf. electrolytic.**
- **C17, C18 = 4 μuf. 600-volt electrolytic.**
- **C19, C20; 1 μuf. 600-volt electrolytic.**
- **R1; 25,000 ohms.**
- **R2, R3, R4; 47,000 ohms.**
- **R5, R6, R7; 1,000 ohms.**

- **E1 = 220 volts.**
- **R1; 10,000 ohms, 1 watt.**
- **R2; 5,000 ohms.**
- **R3; 100 ohms.**
- **R4, R5; 10 megohms.**
- **R6; 2.2 megohms.**
- **R7, R8; 1 megohm.**
- **R9; 0.1 megohm.**
- **R10; 0.01 megohm.**
- **R11, R12; 10,000 ohms, 1 watt.**
- **R13, R14; 10,000 ohms, 1 watt.**

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**SINGLE SIDEBAND FOR**
range of about 2½ kc. of tuning. Heterodyne QRM falling out of one sideband can be completely eliminated by judicious tuning.

For each sight, or for each common recep- tion of a.m. tuning from one sideband to another requires that both the high-frequency oscillator and the i.f., injection oscillator be moved simulta-

ently in order to maintain zero beat. The switching circuit of Fig. 4 is used for this purpose. With just the two switching condensers, C1 and C2, the circuit would not perform properly be-

cause the shift in the high-frequency oscillator would be different at different parts of its tuning range. Compensation capacitor, C3, takes care of this problem. He switch is brought out to the front panel, and a calibration mark is made for each 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. Even if desired to go to the trouble, this condenser might easily be changed with the main tuning.

The audio system includes a standard Select-a-

ject to help with the heterodynes, etc., that are not eliminated by the sideband filter. Following that, it is important that there be plenty of gain, so that within the Select-aject, the QRM stage need operate above its detection limits to produce enough audio output.

A somewhat unusual method is used for disa-

abling the receiver during transmissions. Ap-

plying negative bias to various amplifiers in a com-

mon scheme, but only an arrangement is

usually encountered for the time constants in the

a.m. circuit. In this receiver, a.m. 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. heterodyne stages. The latter cir-

Ctural has only a few 500-µf. r.f. bypass condens-

ers 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 i.f. grid will rectify the incoming signal and produce enough bias to cut off all the other stages and silence the re-

ciever. Meanwhile, the a.m. detector experiences no signal because its bruch amplifier is cut off, and the receiver comes back with full life the in-

stant 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 crystal lined. At the same time it should be a challenge to even the most proficient listener to try building your own "dream" re-

ceiver; you'll enjoy it.

THE RADIO AMATEUR
Variable L.F. Selectivity for the Communications Receiver

One feature of the SX-86 receiver is the wideband-i.f. amplifier that provides six bandpass filters, 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 of the ferrite sleeve is shown in Fig. A. As the frequency is increased, the efficiency of the ferrite decreases and the Q of the tuned circuit is increased. The slug in the ferrite core is adjusted so that the gain of the amplifier remains constant.

![Diagram of the variable-bandwidth i.f. amplifier](A)

**Fig. 1.** (A) Basic circuit of the variable-bandwidth i.f. amplifier. The tuning is increased as C is made smaller, and the Q of the coil is increased. The slug core is held constant, and the slug is tuned by tapping the grid and plate up or down on the coil. (B) The effect of varying C and B is shown. (a) is the passband "tuned" out to a higher frequency, as illustrated here.

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 core was removed from the ferrite core and a means was found for threading the glass- and glass-bonded ferrite. Almost anyone can build a sharp i.f. amplifier if he has a batch of Q = 130 coils, but the SX-86 i.f. has the wide range of bandwidths mentioned earlier. This poses quite a problem, because the frequency must not be changed rapidly with the bandwidth-variations method, and the gain must be held substantially constant. This was accomplished by the Hall-Moore engineers in the general way shown in Fig. 1A. This simplified diagram shows a variable condenser aligned with a variable resistor—in the actual receiver these are switch-position controls. It can be seen that the smaller the capacity of C, the higher will be the coupling between the two tuned circuits, Lc's and Lp's. Furthermore, the larger the value of R is made, the lower becomes the Q of the grid tuned circuit, Lg's. 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 50-kc. i.f. stages has taps on the coil, as represented in Fig. 1A, by the leads to Rg, and this enables the gain of the i.f. amplifier to be held relatively constant over the entire range.

The wideband 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 these conditions ("sharpest," "medium," and "broadest") are shown—the effect is as though the bandwidth "grows" to the higher frequency. It is pointed out here to explain why it sometimes requires some time after the switch is turned to a high-frequency side of the i.f. passband.

**EASY-TO-TUNING**

S.S.B. SIGNALS

The receiver has to be set with an accuracy measured in cycles per revolution of the tuning knob makes tuning a single-bandwidth i.f. setup more difficult than one that covers 20 to 25 kc. per knob rotation.

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**SINGLE SIDEBAND FOR**
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 signal to the low frequency. A circuit for use with a 455-ke. i.f. and a 50-kc. i.f. is shown in Fig. 1. The 455-ke. signal from the receiver is picked up by coupling through a 1- or 2-mfd. capacity from the grid of the second i.f. amplifier to a length of the shielded tube that connects to the 455-ke. input jack of Fig. 1. The position of switch S1 determines which crystal oscillator is in operation. The third position of 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-ke. i.f. chokes in the screen leads of the mixers provide tuning for the oscillator portions, and it may be found necessary to adjust them with capacitors other than those shown in Fig. 1, depending upon the crystals and i.f. chokes that are used.

Other combinations of frequencies can be used, of course. For example, an 800-ke. selective i.f. amplifier working from a 455-ke. amplifier would require oscillator crystals of 370 and 430 kc.

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Just to prove that one really doesn't need to complicate a receiver for a.e. W2NPJF (now W1CCU) cites a QSL card he received from North Carolina, where the SWL was using a 6GFG regenerative receiver!
Selective Sideband with VFO and a Filter-Type Generator

This page, passed along by Jim Freund, W3QMI, 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. W3QMI’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 band is selected, depending upon the sideband to be used in the annular band, and this is used to heterodyne (the low-frequency s.s.b.) signal to the operating frequency. The output of the second mixer stage, of course, is 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 to set 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 “Tuesday side” of filter is always used.

Fig. 1 — Block diagram of a heterodyne method for selecting upper or lower sidebands with s.s.b. filters. s.s.b. = sideband signaling. With a filter type oscillator the crystal frequency is changed to produce the desired sideband. For choosing the modulated-frequency in that the modulated carrier is set up once for just filter action and left there.

Fig. 2 shows a modulated low-frequency s.s.b. generator, using a Collins mechanical filter or a crystal-button 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 6AK5 oscillator plate coil and the 6AK5 amplifier plate coil, for best carrier suppression.

Fig. 2 — A modulator circuit for use with the Collins mechanical filter. T2 is a small universal output transformer.

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SINGLE SIDEBAND FOR
The "Phasing" Method of Generating Single Sideband

DONALD E. NORGAARD, W5KJ

**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. 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 diagrams 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 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.

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**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 obscures the modulating signal in A by 90° when the two signals are combined, and it is this process which produces the desired cancellation, resulting in a single-sideband signal.
Fig. 2 — The system outlined in Fig. 2 becomes practical by using two single channels (2 or 4 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 by translating the input 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 how low or how high these frequencies may be, but, of course, there are practical limits. The phase-shifting networks can be designed to cover a frequency range of 5 octaves, for 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 constant-voltage level, using linear amplifiers to build up the power.

5) Single 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 average of 90° over a frequency range of at least 7 octaves. Of course, the ideal differential phase-shift is exactly 90°, and the excursion of the actual phase-shift curve is ± 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

undesired sideband

desired sideband

is 3:1, and for α = 2°,

\[ \tan \left( \frac{\pi}{2} \right) = 0.874, \text{ or } -33 \text{ db} \]

The symbol α represents the deviation of the actual performance from the ideal 90°, and, in the above example, it 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 2000 c.p.s. There is little to be gained by improvement of this ratio, since subsequent amplifier distortion can introduce serious components in sufficient amount to mask any improvement gained by distorting the phase-shift network characteristics.

A Practical二 二Layout

While the "black box" diagram of Fig. 3 is useful in explaining the principle of generating single-sideband signals, it does not represent a complete single-sideband writer 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 eliminated, 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 25- or 25-meter "phone" and Q5 within three bands of 200 c.p.s. In order that the sidebands of the signal 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-

SINGLE SIDEBAND FOR
time where the grid takes no current (Class AB). If suitable bias is not given to the grid, a conservative 50-watt peak output may be obtained. In either case, the output power is sufficient to drive additional amplifiers of like-sounding ratings or to supply directly a low-power single-sided-plate transmitters.

The functional block diagram (Fig. 5) might appear formidable to 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 circuit and the radio-frequency circuit must work hand-in-hand in order to generate single-sided 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 sur-

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of how "high fidelity"-minded one may be, crowded bands force the operator who listens to the transmission to restrict his reception 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 futile.

2) Elimination of frequencies below 200 p.p.s. removes a large percentage of the high-energy speech components that do not contribute to intelligibility. Such elimination permits the trans-

mitter to concentrate its efforts on only the essential portions of speech power. In practice, this means something like 5 to 8 db. in system effec-
tiveness. Two or three dollars spent on a suitable radio 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 dia-

gram in Fig. 5. The objectives in this portion of the single-sided system are:

1) Very high order of frequency stability.

2) Provision for 90° of phase shift in the ex-

tinction for the two balanced modulators.

3) Ease and stability of adjustment.

4) Absence of r.f. feedback.

5) Low distortion in modulation and subse-

quent amplification.

6) Provision for adjustable carrier level; gen-

eration of a.c., p.p.m., and single-sided audio signal; output-level control.

7) (optional) banding Operation on 15- or 20-

meter bands; max QVW within each band; choice of sideband transmitted.

Obviously, a number of methods exist for ac-

complishing these objectives. Many of the pos-
sible methods that may occur to the designer will satisfy the requirements quite well; some will not.

Others, while technically adequate, may be dif-

ficult to adjust or may be impractical in some other way. Since the handling of radio frequencies is common to 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 banding arrangement to make the balance controls to prevent accidental shifting of their positions.

Balanced Modulators

Fig. 5 illustrates the use of the two balanced modula-
tors. A little explanation might be helpful in un-
derstanding why and how balanced modulators are used.

In a 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 side-

bands to be generated, but to balance out the ca-

rier after it has served its purpose. There are

many forms of balanced modulator; some bal-

ane carrier out in the offset or the two signals sup-

plied; others balance out both input signals.

But none of these 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 rebuilding an amplifier, since the same principle is involved. In the case of the balanced modulator, the per-

etration of balance required is usually quite high, and some methods for satisfying the conditions necessary for balance must be provided. Very few arrangements automatically provide the condi-
tions necessary for perfect balance and frequently those that do are limited to operation at low fre-

quencies, where circuit losses have negligible effect. It has been found practical to "grab the ball by the horn" and use some arrangement where separate phase-and-amplitude 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,
and opposite in phase as is feasible using circuit components of ordinary commercial baluns. The RC circuit between point 2 and point 3 (fig. 3, No. 1 of the first modulator tube) is a 0-SAT variable tone in this example may be designed to provide about 20° phase shift at the operating frequency, for example, 300 mc, and 2,000 mc. The RC circuit in the other grid can be designed to produce whatever phase shift may be required to balance the output of the tetrode amplifier, C. This practice gives a phase correction of about 15°—sufficient to assure perfect phase balance of the signals applied to the tubes. No attempt is made to equalize the amplitudes of the signals in the grid circuits because it is almost too much to expect that a properly-balanced pair of tubes could be found in order to take advantage of balanced amplifiers. Instead, the function of amplitude balance is accomplished by means of a bias-ad- justment 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 plug-in is completed by applying push-pull modulating signals to the No. 3 grid so that the sidebands generated by the separate modulation generator in each tube add together in the common plate circuit. The audio-frequency component balances out in the plate and screen cir- cuits, this being a case of a balanced modulator that balances against each of the input signals. However, slight imbalance 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 fre- quencies.

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 modulates the carrier, the effi- ciency is obtained primarily by the amplification of the audio-frequency components. The efficiency obtained in practice is more like 45% at X-band frequencies to a little under 70% at audio frequencies. A loss of 50% efficiency in the generation of the signal results in a 100% loss of power in the resultant signal, which under these conditions is a low-power level with the modulator operating under conditions that are comparable to those of a tubes amplifier. To the level of the point of operation, the input to the grid of the output stage would be substantially reduced from its normal level.

The audio phase networks are designed in order to avoid any form of distortion that could result from an intermodulation between the modulating and carrier frequencies. This is accomplished by using a balanced modulator and a balanced network with maximum signals of about 1 to 1.5 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 may employ receiving tubes of extremely narrow bandwidth in the grid circuit when the signal levels reach the power- level cases. For instance, the doped portion of Fig. 5, up to the grid circuit of the output stage would be modified in overall magnification and construction for L2, part of an average communi- cations receiver. The versatility of Fig. 5 should make it attractive, since some of the versatility is obtained at the expense of circuit configuration and not fundamentally a part of single-sided operation. This is apparent when com- paring Fig. 5 with Fig. 3.

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Table 1

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Audio phasing networks

The 90-degree audio phasing network, as- cribed in the preceding article is an easy “ac- crib” type (transforming amplifiers) that has since been superseded by simpler “passive” networks employing only resistors and condensers. As an example of the latter in the network desired by H. B. Dunn, W2BAJ, and used by W2RJ in the phasing circuit identified as Fig. 3. A passive network using even fewer parts was incorporated in W2RJ’s “380-SB” (E. H. Huns Xmas, Vol. 5, No. 6) and has since been made available commer- cially as a pre-amplifier-increment unit. These two networks have substantially identical performance, the principal difference in their characteristics being that the Dunn net- work can be terminated in resistance while the Neppard net work must work into an open circuit.
Post-Phasing Distortion

GEORGE GRAMMER, WIDF

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 nonlinearities 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 phase-shift network, when applied to the grids of separate amplifiers that distort each signal to exactly the same way, will result in second fundamental and second harmonic component having the same relative phase with respect to its fundamental.

The second-harmonic component, being exactly 90° out of phase with the fundamental, will result in second-harmonic distortion, which is the same at both sidebands.
middle drawing. The fundamentals \(A_1\) and \(B_0\) have been omitted in this drawing since they are no longer needed and merely follow the usual pattern for single-sideband generation. The sec-
ond harmonic, \(A_2\), of the first audio channel, when applied to the first r.f. channel generates side frequencies \(A_{12}\) and \(B_{20}\), upper and lower

\[
\begin{align*}
A_{12} & \quad B_{20} \\
A_{21} & \quad B_{10} \\
\end{align*}
\]

respectively, spaced twice the fundamental fre-


frequency from the carrier. Similarly for the har-
monic \(B_1\) generated in the second audio channel when applied to the second r.f. channel. When the carrier-tolerance r.f. components are combined in the


balanced modulator the two lower side fre-

cuences add together as shown at \(A_1\) in the low-

erst drawing and the two upper side frequencies

combine into \(U\). These are identical with the


upper and lower side frequencies produced by

double-tone modulation in one balanced modula-

tor — that in, double-sideband suppressed-car-

errier transmission.

All harmonics generated in the circuits under

discussion will be shifted in phase in proportion to

frequency. Thus, if the fundamentals in the

two channels are 90° degrees apart, the second

harmonics will be 180° degrees apart, the third

harmonics 270° degrees, and so on. It has been

shown that the second harmonic is transmitted

as a result of double-sideband signal; this is

true of all even harmonics. The case of the third

harmonic is interesting because a phase differ-

ence of 270° degrees between the two audio chan-

nels is equivalent to a 90° degree difference, but

with the lead and lag reversed as compared with

the fundamentals. Hence, identical third-har-

monic distortions in the two channels will give

single-sideband transmission again, but with the

output in the second sideband. On the other

hand, the 90° shift at the fifth harmonic is

360° degrees, which is identical with the 90°-degree shift at the fundamental; hence, fifth-harmonic distortion gives rise to a composite component in the desired sideband but not in the other. Odd

harmonics give single-sideband frequency, alternat-

ingly from the desired and undesired sideband.

This discussion has been limited to the es-

sential features of certain distortions in each channel, exact 90°-degree r.f. phasing, and complete carrier suppression. For even har-

monics, departures from exact audio phase and

amplitude balance will affect the amplitudes of

the side frequencies to some extent; so will inco-

erante 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 un-

equal amplitudes), and incomplete carrier sup-

pression also will cause phase modulation along

with amplitude.

As a first approximation the spurious compo-

nents 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 dis-

tortion, 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 de-

grading performance.

Fig. 3 — Phase relationships in two channels when the fundamental is in (a), leads the fundamental in the other (b), by 90°. The second harmonic, shown in (b), is repre-

sented with \(A_1\) and \(B_1\) the second harmonic associated with \(B_0\).

Fig. 4 — Single-sideband output when the higher frequency \(A_2\) in the upper drawing is applied to the bal-

anced modulator having \(90°\)-degree r.f. phas-

ing. The r.f. carriers are shown for clar-

ity. Note that there is an additional 180°,

degree phase shift in the lower drawing. When ca-

rriers are eliminated the upper and lower side component, re-

spectively, overlap as shown in the lowest

drawing.

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The article will describe an exciter, built
mostly of junkshack 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 values.
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 over the amateur bands for local
work or to an s.f. amplifier that is adjusted for
linear operation. The operating frequency can be
varied over a wide range without seriously im-
pairing the adjustment. Provision is made for
transmitting either the upper or the lower side-
band.

The Circuit
The circuits were the phasing method of single
sideband generation and is shown in block
manner at 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 90 de-
grees, and the other has an output volt-
age that lags the input voltage by 90
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, that is, two balanced
modulators balance out the quadrature frequen-
cy, so that no carrier appears in the output of
either pair of balanced modulators.
The audio source is fed into a Double-
type phase-shifting network. This net-
work requires push-pull input volt-
age, and delivers two output voltages differ-
ing in phase by 90 degrees. These two voltages are
used to drive separate Class A amplifiers that
serve the dual purpose of amplification and
injection of the network from the audio (the balanced
detectors). Each pair of bal-
anced 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 side-
band is canceled out and the resultant output
is single sideband.

The complete schematic of the exciter is shown
in Fig. 2. Two 6V6 tubes are used as balanced
modulators. The plate circuit of the balanced
modulators use a push-pull-parallel arrange-
ment. The grids of two pair of balanced modu-
ulators are fed through a phase-shift network con-
sisting of a 90° phase 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 90° phase resistor
and a condenser which is adjustable to 2000 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 screens of
each modulator tube is by-passed in ground for

SINGLE SIDEBAND FOR
Fig. 2 - Circuit diagram of the single-ended receiver.

Ra, Rb — 680 ohms, 2 watts.
All resistors 1/8-watt units, specified otherwise.
La — 4 1/2 V, 10-turn No. 28 gauge.
14 Mf. 12 turns No. 30 gauge.
All with close-wound at winding end of slot of
National VR-30 Dual-tune form.
Ls — 3 1/2 Mf. 30-turn 2% watt tank coil with swaging
plug.
Lc — 5 Mf. 75-turn 2% watt tank coil with
swaging plug (Radio 05-0111).
RFG — 2000 ohm, r.f. choke.
Ss — D.P.D.T. switch, preferably with center off position.
Ts, Ts — 10-watt modulation transformer, 10,000 ohm
r.c. to 4000 ohm (Stanley A-252).

A reversing switch, Ss, 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.

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The original W2SNJ exciter used a circuit differing in several details from that given in Fig. 2, but the identi
cal constructional features I am using. This modulator con
sists of a 10-ohm, 2-kilowatt coil connected in series with the output of the audio-oscillator, when it contains only those elements essen

tial to generating r.c. 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 V-830 shown in Fig. 2 is not included in the unit itself, but it serves a purpose on the chassis to accommo

date the plate tank. Table 1 is used in selecting the network com

ponents. The procedure is to select as many re

sistors and condensers as possible with minimal values as indicated in the second column of the chart. Measure all of the condensers first, and select the six resistors whose measured values are closest to the "target" values in the third column. Enter the measured values of these con

densers 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 service

men, and a good ohmmeter should give sufficient accuracy in selecting the network components. Absolute accuracy is not important, if the com

ponents are all in correct proportion to each other. A difference in percentage error between the resistance measurements and the capacit

tance 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 sets of resistors and condensers, and the measuring equipment, as a service to amateurs everywhere.

The B.F. Phasing Inductor

The only other "tricky" component of the ex

citer is the r.f. phasing inductor, L2. This in

ductor is wound on a 2×4×8-form. The coil should resonate, respectively, at the center of the band with a variable condenser set at about 135 ufd. for 3.9 Mc., 75 ufd. for 7.25 Mc., and 27 ufd. 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 sta

tion receiver.


Construction

The exciter is assembled on a 3×10×3-inches chassis. The plate tank timing 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 1⁄4-inch holes drilled through the chassis near each tube-socket plate connection. The 6V6 screen grids are by-passed to ground di

rectly at the sockets. R15, L4, and R13 (all adjustable components) are mounted in a row directly behind the 6V6s. The two 600-ohm amplifiers are mounted at the rear of the chassis, one on each side, with R2 and S2 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 im

pedance will operate satisfactorily for the short

distance usually required. A means for adjusting the r.f. driving power is desirable. A surplus Com

mand set transmitter (DC-506 or T-19-ARC-3), 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 600-ohm output tube in the speech ampli


cifier can be considered for this purpose.

<table>
<thead>
<tr>
<th>Network Design Data</th>
<th>Port</th>
<th>Module</th>
<th>Measured Value</th>
<th>Gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>Port</td>
<td>Module</td>
<td>Measured Value</td>
<td>Gain</td>
<td></td>
</tr>
<tr>
<td>G1</td>
<td>0.05</td>
<td>0.03</td>
<td>125</td>
<td></td>
</tr>
<tr>
<td>G2</td>
<td>0.05</td>
<td>0.10</td>
<td>100</td>
<td></td>
</tr>
<tr>
<td>G3</td>
<td>0.05</td>
<td>0.15</td>
<td>125</td>
<td></td>
</tr>
<tr>
<td>G4</td>
<td>0.05</td>
<td>0.15</td>
<td>100</td>
<td></td>
</tr>
<tr>
<td>G5</td>
<td>0.05</td>
<td>0.25</td>
<td>125</td>
<td></td>
</tr>
<tr>
<td>G6</td>
<td>0.05</td>
<td>0.25</td>
<td>100</td>
<td></td>
</tr>
</tbody>
</table>

All condensers react, and all resistors ±10%.

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for will provide sufficient audio. The modulator tubes should be removed from their sockets, and the center tap of the transformers secondary should be grounded, after removing the bias connection. An alternative method is to use blocking condensers in the audio leads to the single-ended exciter to isolate the modulator bias from the audio phase-split network in the exciter. If some other source of push-pull audio is used, it should have low internal impedance (Clay A triodes, or beam tubes with negative output feedback).

The exciter may be coupled directly to an antenna for use as a low-power transmitter, but most amateurs will want 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 radio operating conditions for the tube under consideration. Grid bias and screen voltages should have very good regulation. For amateur radio operation, tubes may be operated considerably beyond the rating given in the tube manuals, but such operation is beyond the scope of this article. When the r.f. amplifier is operated Class AB or Class B, the grid tank circuit should be bypassed 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 useful carrier present in the output in relation to the sideband output. Also, the exciter will be more stable, and it will remain adjustable longer with lower plate voltages.

A pair of 807s operating Class AB can be driven by the exciter with only 40 volts (at 126 volts) input to the balanced modulator, with the output amplifier also operating at 125 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 modulator requires sufficient r.f. drive to develop 12 volts of grid bias under these operating conditions.

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

- Total balanced-modulator plate current 85 ma.
- VR tube supply current 20 ma.
- Transformer supply current 32 ma.

The total balanced-modulator grid current, measured at the bias terminals, will vary with excitation, but it should be in the range of 15 to 20 ma.

These currents will not change appreciably with varying audio input, and with the exceptions of the grid current, will not change appreciably when the excitation is removed, provided that 45 volts of 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, and transformer bias is used, it is possible to operate the exciter with a transformer bias supply. The disadvantage is that the transformer must be bulky, and a 15 volt bias supply 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 load, apply r.f. exciting to the exciter, feed a source of sine-wave audio into the speech amplifier, and tune the r.f. circuit in the conventional way for maximum output.

Reduce the audio input to zero, and adjust potentiometers R1, R2, and R3 for minimum carrier output. Minimum carrier output may be determined by any sensitive r.f. indicator coupled to the final-amplifier plate circuit. A 6-1 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 uF capacitor. If a null indication cannot be obtained, adjust the amplifier exciter, the transformers, and the circuit as a whole until the desired null is obtained.

A top view of the original exciter. The large switch at the rear selects the selector or line feed, apply r.f. exciting to the exciter, feed a source of sine-wave audio into the speech amplifier, and tune the r.f. circuit in the conventional way for maximum output. Reduce the audio input to zero, and adjust potentiometers R1 and R2 for minimum carrier output. Minimum carrier output may be determined by any sensitive r.f. indicator coupled to the final-amplifier plate circuit. A 6-1 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 uF capacitor. If a null indication cannot be obtained, adjust the amplifier exciter, the transformers, and the circuit as a whole until the desired null is obtained.
be obtained within the range of the potential-
ity, the 6V6 tubes are not evenly matched. Exchanging the positions of the 6V6's may aid in achieving the balance, or other tubes may have to be used.

After the carrier balance is obtained, tune the r.f. source on the detector receiver, and with the antenna terminals shorted and the crystal sensitivity in sharp position, adjust the by-pass tuning to the point where only one sharply-
peaked response is obtained as the receiver is tuned through the signal. Now apply one-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 meter as the receiver is tuned. Set the receiver on the other of the two sidebands and adjust R1 (C) and R3 for mini-
mum sideband strength. If suppression of the other sideband is desired, throw A4 to its other position. A dip obtained with our set of adjust-
ments is not necessarily the minimum. Other conditions should be tried. The final adjust-
ment should give consistent readings for the two sidebands which differ by at least 30 db. The blue voltages on the two pairs of balanced modula-
tions will be approximately equal.

After the adjustments have been completed, the r.f. drive to the carrier 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 aux-
iliary, and an on-the-air test.

If an oscilloscope is available, a simple and more reliable adjustment procedure may be used. Either of the sideband horizontal traces may be used on the oscilloscope. The vertical trace should be caused to output of the transmit-
ter in the same manner as is used for observing modulation amplitude. The zero-volts null-frequency input to the speech amplifier should be any convenient multiple of the oscilloscope sensitivity. A 600-cycle sound-frequency source or a 600-cycle audio frequency are commonly used.

If the exciter is modulated with a single-
sideband audio frequency, the output should be

a single audio frequency. Therefore, the oscillos-
cope should show a straight-edged band across the screen, the same indication as is given by an unmodulated carrier. If carrier output, or un-
wanted sideband output, is present, it will be indicated by "ripples" 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 have heavy modulation. 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 over-
loading of any stage of the transmitter. Overload-
ning is indicated by flattening of the modulation-
peak patterns at the top and bottom.

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Speech Amplifier and Voice Control for the W2UNJ
Exciter

an basic VFO circuit described above can be readily combined with a speech ampli-
ference and voice control circuit to make it a complete carrier in every respect except as an r.f. source. The latter, which is a VFO, is a separate unit in most stations 6000. The additional circuit and the accompanying photographs show the exciter and voice to Figs. 2 and 3 in one unit.

The speech amplifier is designed to accommodate both low and high frequencies, implying only the audio range required for good intelligibility. Its output is coupled to the input of the audio shift network through a transformer with a center-tapped secondary, to provide proper small 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 to the modulation feed-

ers, since interaction may occur that will upset the phase relationship at the output of the two amplifiers. If separate plate voltage sources are not
A rear view of the complete exciter. The two r.f. plating adjustments project from the shield "c". A small relay is mounted in the volume control, the wire-wound transformer is for the power cable.

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available in the output section may be used to isolate the voltage to the speech amplifier.

**Voice Control**

The output of the speech amplifier is taken off through an adjustable voltage-divider circuit, Rs and Rs, and blocking condenser, C10, to the voice-control circuit. Here it is rectified by the diode of the 6SK7, and the resulting d.c. voltage is used to charge C14 negative. An audio choke prevents audio components from appearing across C10. The triode section of the 6SK7 is normally conducting and holding the relay closed, but when the negative voltage appears across C10 the 6SK7 plate current is cut off and the relay opens. When the audio signal is removed, C14 discharges through Rs and the triode again conducts, closing the relay. The built-in voice-operated relay can be used in a number of ways to provide the rapid voice break-in commonly used on V.H.F. a.m. phone. If a good c.w. break-in system is already in use at the station, the voice-controlled relay contacts may be substituted for the key, and no other changes are necessary.

If the lead oscillator in the receiver will key in the plate voltage level satisfactorily, then a single 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 receiver and its associated audio equipment are assembled on a 13 x 7 x 2-inch aluminum chassis. The four 6SK balanced-modulator tubes are arranged in a square pattern toward the front center of the chassis, with the plate tuning condenser and dial off to one side and the 6SK audio amplifier tubes on the other. The two modulation transformers 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 6SK7 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, C4 and C5, 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 6SK7 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 WRJN EXCITR**

**Fig. 4 — Schematic diagram giving circuit arrangement for the commercial phase-shift network in the w.b.s. receiver of Fig. 2.**

**SINGLE SIDEBAND FOR**
Single-Sideband Reception by the Phasing Method

DONALD E. NORGARD, W9UJ

The fact that sidebands—and sidebands alone—provide transmission of intelligence makes single-sideband systems possible. In the case of simplephase 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 phase 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 prefiguration 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.

Selectivity ahead of the detector in a receiver can help prevent misleading effects from strong adjacent-channel signals, but in itself is not a complete solution to the problem of selective-carrier 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 join the detector. This is basically the idea of “excluded-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 inaudible to this one key signal. Reception of c.w. signals has always employed this principle, but it can be applied to “phone carriers.”

Single-sideband reception can be employed on single-sideband-transmitted transmissions (c.w. or c.w.) or on double-sideband transmissions (a.m. or p.m.). Since in the latter case the upper and lower sidebands contain identical information in duplicate, it has been found that the performance of excluded-carrier operation and single-sideband reception is of great benefit in overcoming the

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

**SINGLE SIDEBAND FOR**

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Fig. 2 — A block diagram of a dual-carrier-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.c.c. signal, or it can be used for single-carrier (or double-carrier) 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.

Figure 1 has been simplified a little for purposes of explanation. The oscillator serves as a synchronous 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 l.f. portion of a receiver. The oscillator operates at intermediate frequency, replacing the h.f. and i.f. of the conventional receiver setup.

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 output 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 oscillator-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 realized-carrier demodulator of Figs. 2 and 3 driven phase-shift networks that have the property of 90° differential phase shift over a wide range of audio frequencies. The operation of the system depends upon the transmission properties of the phase-shift networks to realize 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 - A block diagram of a single-channel receiving system incorporating sideband-elimination demodulation.

Fig. 4 - The single-sideband output is obtained by superimposing action of two demodulators and phase-shift networks.

Fig. 5 - The method of connecting the "sum" and "difference" outputs of the phase-shift networks. In positions 1 and 2, the upper sideband output from one sideband or the other, while positions 3 and 4 give demodulated double-sideband outputs.

The outputs of the phase-shift networks as shown in Fig. 6. There are no 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 1/4 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.

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Recevier-System Characteristics

The design of phase-shift networks permits rather good attenuation of an undesired sideband in a sideband mixer 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. 6 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 may be expected is described in the system described in this article is illustrated in Fig. 7A and 7B, which are plan of attenuator versus frequency. An equally powerful sideband of 75 Hz is assumed for the I.F. system of the receiver used as a source signal. The actual response is illustrated in curve A, which might be measured at the output of either channel of Fig. 6. When the symbolic carrier is set at the center of the band as indicated, the apparent I.F. response measured in which position 2 would have the appearance of curves 4-7', while at switch position 2 the response would be as indicated by curve 4-8. This certainly is single-sideband performance, since curves 4-7 and 4-8 overlap one another only an extremely small amount near the carrier. With the I.F. bandwidth of 12 kHz, each sideband is about an octave wider than is desirable for reliable phone communications. A narrower I.F. amplifier will reduce the bandwidth, but somewhat less satisfactory result can be obtained by using 4-9 kHz 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 unacceptable, but the bandwidth is effectively limited to 3 kHz, in each sideband position (curves 4-7' and 7B). Despite sideband rejection in position 3 and 4 the apparent I.F. bandwidth for the same audio response.

It can be appreciated that single-sideband reception of double-sideband signals offers sufficient 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 used. This is not the fault of the receiving system, however. The real fault is that double-sideband transmission are twice as much bandwidth as 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 phase network requires that the signal source have low impedance. This same requirement applies in each case of the input supply for the tube or tubes feeding the network. The output filter condenser of the supply should be large - 40 µF or more.

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Aside from the heterodyne beat notes 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 change to 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., W8QYT, AND JOSÉ MIGUEL NÁZ, XERX

This article describes an improved version of the 'Select-o-jet' 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 grid 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 heterodyne or beat notes in 'phone reception because this 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 short SOJ will be briefly set forth. Fig. 1 shows an ordinary one-stage retubeless-complex-coupled audio amplifier, which could be the first stage of a communications receiver, but whose output terminals the short SOJ is connected to.

Now it is well known that each an R-C amplifier can be represented, insofar as 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 midcircuit. 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 short 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 sign to that of the amplifier's equivalent generator. The a-c. voltages in the circuit will then be replaced for convenience by d-c. voltages, and the generator assigned polarities just as if they were a-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 output 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 is shown in

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phase to cause a null. It might at first seem hard to find a viable 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 ($h$) can be developed across $R_2$ in Fig. 3. Note that $h$ is available outside the SOA. This voltage, properly amplified, gives the correct phase and is fed back to the SOA tube's grid, to serve as the upper-limit-generator voltage $h_2$ and thus the right value to cause $e$ to disappear and produce a null.

Provenal Circuits

Fig. 3 shows how the SOA's tube (V1) is connected to the amplifier in practice. To produce a null, the collet voltage $e_1$ is amplified without phase change and applied between the grid of V1 and ground. This voltage will be called $e$. Since it is desired that the null occur at only one frequency, the amplifier connecting $e_1$ and $e$ must be frequency-sensitive. It is convenient to make the frequency-sensitive portion of this circuit a variable-gain AGC phase-amplifier. A complete schematic of the SOA so connected is shown in Fig. 4. The phase-shifting circuit uses 550K and 5500K parallel trimmers as the frequency-sensitive elements. The 550K resistor limits the highest "reso- nant" frequency to 1000 cycles. The low-frequency limit of response is 1000 cycles. It is obvious that this circuit provides isolation and phase reversion.

When the SOA is used to produce a null, an important point is to keep its insertion loss low by making the plate resistance of V1 as Fig. 4 high, and likewise $R_2$. The SOA then affects the gain of V1 only to a minor extent, at frequencies far from that at which the null occurs.

The select SOA draws approximately 7 ma. at 100 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 $\mu$F) filter condenser directly across its output terminals. This is important.

The performance of the select SOA will vary slightly depending on the external impedances of the circuit to which it is connected. For most receivers this will vary from SOH 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 maintain hum, both from the power supply and by direct pickup from the filament leads, as much as possible. It is suggested that the tube types and component values be altered as at closest as possible. It is desirable to provide some vibration damping in the unit to prevent the vibration of the components and prevent them from close to the chassis.

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Linear Amplifier Design

STYRE G. REQUE, WARSAW

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; we are very delicate circuits, similar to the r.f. and i.f. amplifiers of our receivers. Linear r.f. amplifiers, for instance, are built for a very wide range of soil voltages. 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 amplifier form of linear amplifier (r.f. or audio) is the Class B amplifier, which is used almost without exception throughout our receivers and one of the most important circuits. While its linearity is not quite perfect, it is, unfortunately, quite sufficient. The theoretical limit of efficiency in this case is 50 per cent, while most practical amplifiers run 25 to 35 per cent efficient at full output. At low levels this is not worth cutting away, but when we exceed the 2:1 output level something must be done to improve the 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 among amateur amateurs. Class B amplifiers are theoretically capable of 74.5 per cent efficiency at full output, and practical amplifiers run 60 to 70 per cent efficiency at full output. The same amplifier tuber, with suitable tank circuits and 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 r.f. and linear audio tank circuits, provided, of course, that the tubes are suitable for use in the desired frequency range. In r.f. circuits running Class B only one tube need be used if only half the power is wanted, because the fly wheel action of the tank circuit will smooth out the missing half cycle.

One point is well worth mentioning at this moment. If you back up the Class B r.f. amplifier to the output bus, you will find that the output must be taken as shown to secure the efficiency given in the table.

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order of 31 per cent and not the 40 to 70 per cent quoted above. This arises because the figures given are for a non-adiabatic m.s. system. The efficiency of a Class B amplifier is proportional to the signal voltage i.e., at full output it is 60 to 70 per cent, or at half voltage it is 30 to 35 per cent. In a conventional m.s. system the output 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 31 per cent. This need not deter the amateur running a single-ended circuit with suppressed carrier, since his output circuit can be adjusted to give the correct output level, and he can therefore operate in the linear region without distortion.

Amplifier Design

In most cases the design of a Class B linear amplifier will be simple, since most of the necessary power-supply 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 job is to provide proper tank circuits and drive for the tube. As an example, let us choose a tube of good regeneration as a Class B audio amplifier, such as the 5L4. Typical operating conditions are given in Table 1.

| Table 1 |
|------------------|-------------|
| Class B Audio Amplifier Data | 5L4 Tube |
| Plate voltage | 150 volts |
| Grid voltage | 0 volts |
| Plate-current | 200 ma |
| Plate-current | 200 ma |
| Grid-supply | 150 volts |
| Power-supply | 150 volts |
| Power-supply | 150 volts |

Fig. 1 is a schematic diagram of the usual Class B audio amplifier. Fig. 2 is a diagram of the amplifier changes over from a Class B r.f. amplifier. Our first concern will be the design of the proper tank circuits for the grid and plate circuits. The final point to be considered will be the section on practical adjustment. Let us design the proper plate-tank circuits first. As in all r.f. amplifiers, this tank circuit

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should have a loaded $Q$ of at least 10, if we want to have reasonable efficiency and low harmonic output. The loaded $Q$ is defined as the ratio of the unloaded reactance (equal to the tank-inductance resistance at resonance) and the load resistance by the equation

$$Q = \frac{R_L}{X_C}$$

But we also know that $X_C = 2\pi f L$.

If we choose the 75-meter 'phone band as our example of design, and hence substitute 4 Mc. for $f$ and 600 ohms for $X_C$ in (1), we will find the value for $C$ to be approximately 70 µF. This is the value of the triode across the tank, and we must double it to find the value for each section of our split-capacitance condenser, or 140 µF, 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 2000 volts per section would be clearly indicated. The coils should be chosen or pruned until the proper amount of capacity is arrived 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 C linear amplifier, but is equally true of the Class C, perhaps in 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-306, involves a pair of pentode 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 6F66A) 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 lead 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 rated 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:

1. The load presented to the driving stage must be constant.
2. The voltage at the grids must have good regulation.

Referring now to our original example, the GL-306, 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,

$$R_{eq} = \frac{E_{max}}{P_{grid}}$$

Substituting the proper values of grid-to-grid voltage and grid driving power from the data in Table I gives an equivalent grid loading of 4500 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 2000 ohms. Substituting this value in Equation (2), and the resultant value of $C$ in Equation (3), we find the necessary value of $C$ to be 200 µF across the tank or a single-stage condenser of 450 µF per section. A broadcast receiver condenser of 450 µF per section is readily available and will easily yield the low peak voltages on the grids.

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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 partly a function of the tube chosen. For linear amplifiers it is necessary that the tanks be properly de- signed. If the Class C stage seems to be tolerant of errors in tank design, it is because few of us have been given full consideration to the proper handling of our amplifiers and have been content to operate with the efficiency and the harmonic output found.

**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 rated. If we were to use a tube of this sort we would determine the quantities equivalent to those given in Table 1 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-628-B.

At a first glance, let us suppose that the plate-supply voltage is 360 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 1/4 rated plate dissipation. Bias determined in this way will usually allow better linearity (less distortion) than a bias-chosen to complete cut-off. Since our GL-409-B has a rated dissipation of 20 watts per section, the proper bias will allow 0.7 watts resting dissipa-
tion per section. At 300 volts plate sup-
ply we can use as a resting plate current of 13.5 ma. per section. From the plate characteristic curves of Fig. 3 it will be seen that approximately 18 watts 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 80 per cent, we can now determine the maximum inputs power. The 50 per cent 60-cycle plate current will produce approximately equal the maximum plate dissipa-
tion, which is 40 watts (both sections) for our GL-409-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 maxi-
mum signal plate current. Since the total current will be 115/360 = 0.323 ma. = 300 ma. = 115 ma. per section d.c. plate current at maxi-
mum signal.

A plate-current pushing of each tube of our Class B linear amplifier are, had we a section of a sine wave, such as might have been produced by a half-

1. Approximate values only—used for design purposes.

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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 plate load, 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 W3CKU in Q.S.T. 68, June 1953. 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 ac oscilloscopes.

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 capacity in the filter should be of the order of 40 µF, and values up to 100 µF usually will show considerable 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.

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### Table: 113

<table>
<thead>
<tr>
<th>Type</th>
<th>Class</th>
<th>Peak</th>
<th>Zero-Sig.</th>
<th>Max-Sig.</th>
<th>Peak-Rated</th>
<th>Max-Sig.</th>
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<th>Avg-Quench</th>
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<th>Peak-Rated</th>
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**Exhibit:**
- [Link to exhibit](#)

1. Adjusted to give identical zero-signal plate current.
2. These values are measured under these specified conditions: voltage, wave form, plate current.
3. Values in parenthesis are with two-mart pilot signal.
Distortion in Single-Sideband Linear Amplifiers

WARREN B. BRUENEN, W9TK

If 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 frequency. 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 $3p - 2q$ and $3p - 2q$, where $p$ and $q$ represent any two radio frequencies present in the desired transmission. The fifth-order product frequencies are of order $25$ to $30$ dB. Before recently, the first few years of commercial performance of the order of $25$ to $30$ dB. Recently, however, it has been achieved. Recent developments indicate that even $40$ dB is possible and practical.

In amateur transmitters where only one sideband channel is used, the distortion requirements depend upon the allowable interference to others operating on nearby channels. Practically, the relative amplitude of the signal with distortion to the amplitude of a near-by signal another amateur is trying to receive enters here. Common courtesy on the crowded amateur bands dictates the use of transmitters with no little distortion of the sort the act reasonably permits.

Causes of Distortion and Methods of Reduction

The principal causes of distortion are nonlinear characteristics of the amplifier tubes and midcurrent loading. In order to minimize the generation of distortion substantially to the last stage, all other stages are usually operated Class A. The plate current curve of Class A amplifiers in general can be represented by a single exponential curve as shown in Fig. 2A. The curve is kept low by operating the tube in the most linear portion of its plate current characteristic and by keeping the plate-level low. Fig. 2B shows the nature of the linearity curve of a typical Class A amplifier. The curvature is greatly exaggerated since for $N$ (the ratio of the order of $30$ dB, it cannot be detected visually.

Class A amplifiers usually have a very similar curve. When the linearity characteristics

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of a series of cascaded amplifiers have similar
characteristics, the distortion products generated
by each add together in phase.

Fig. 3 - Effect of grid shorting on linearity.

When amplifier tubes are driven into the grid
current region, the resulting grid-circuit loading
causes the linearity curve to drop at larger 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

smaller than would be expected in a conventional
amplifier. The distortion products are
approximately 180 degrees out of phase with
those previously discussed. A distortion product
is said to be 180 degrees out of phase with another
if the waveforms are 180 degrees out of phase
with each other.

Another way of providing a low driving
impedance is to use a very high resistance drive
circuit, such as a transformer or a resistor, and an
impedance-inverting network. 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 advantage
of this method is that it is difficult to maintain proper coupling
without special adjustment, and these circuits are seldom used in commercial
frequency-overlapping transmitters for this reason. Link

Fig. 4 - Distortion cancellation.

strength. This is because the coupling is in the region
of cancellation. For this reason, the value of dis-
tortion cancellation is not as great as it might appear.

The non-linearity caused by grid current, load-
ing is a function of the operation of the grid
driving source. The regulation of linear amplifiers
with a varying load is poor in general, it is com-
mon practice to use an active terminal in parallel
with the main grid load, and to obtain satisfactory regulation it is usually necessary to

'Green, "Design of Linear Amplifiers for Single-Ended
and March, 1939.'

Fig. 5 - High-gain-
ness driver and impedance-
inverting network.

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off of plate current also depend upon the tube geometry. In all but a few transmitting tubes the plate can swing well below the average voltage before plate emission takes place, and when the plate swings down in this region the plate current drops off quite a bit. If the distortion requirements are not too high, the high plate efficiency realized by using large plate swings can be utilized. Fig. 7 shows a typical linearity curve of a typical tube amplifier, curve of a typical 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 in plate-current limiting takes place.

Estimating Distortion

A means of estimating distortion in a power amplifier is quite useful, and the approximate signal-to-distortion ratio of a two-tone test signal can be obtained from the linearity curve. Equations have been developed for calculating this, and are used to plot the curve in Fig. 8. This curve maps 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 bias elements. It can be substantially eliminated by proper adjustment of bias, screen and plate voltages, so means of estimating distortion from this cause will not be discussed.

Envelope peak flattening which is due to grid current loading and plate current nonlinearity at large plate swings is often the major cause of distortion. The amount of envelope peak flattening due to grid current loading may be easily calculated, see Fig. 9. The equivalent grid load resistance $R_{eq}$ in Fig. 9 is calculated from the grid driving power and the e.i.g. grid swing, $E = V_{g}$, where $E$ is peak e.i.g. grid voltage, and $P_{g}$ is grid driving power = $e_{g}I_{g}$, where $I_{g}$ is d.c. grid current in amperes.

The resistance of the envelope resistor $R_{e}$ is known or can be chosen for the calculation. The equivalent resistance of $R_{eq}$ and $R_{e}$ in parallel is then calculated by:

$$R_{eq} = \frac{R_{e}R_{g}}{R_{e} + R_{g}}$$

If the source impedance looking back at the driver stage is very high compared with $R_{e}$, 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 $R_{eq}$ to $R_{e}$.

Peak flattening = $\frac{E}{R}$ X 100 (per cent).

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

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

Finally the third-order distortion component is at least 6 dB greater than the fifth- or higher-order components, but a sharp break in the line as such might be caused by plate-swinging saturation, as shown in Fig. 10, will contain more fifth- and higher-order components.

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than if it were a smooth curve. This type of non-
linearity is particularly objectionable because of the
wide band over which the distortion prod-
tects appear.

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

\[ V \]
\[ I \]

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
nearly 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-
order distortion components because the third
tends to be cancelled as previously shown in
Fig. 4.

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\[ I \]

Distortion Measurements

Distortion measurements are of particular im-
portance 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 overdriven, 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, and many of measurements in order to
provide a modulation envelope.

There are several different methods of induc-
ing or measuring distortion, and each has a sepa-
rate field of usefulness. The "Linearity Trace" described
below is especially useful for quick
observation of amplifier operation as the effect
of various adjustments can be instantly ob-

\[ V \]
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served. This instrument consists of two

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Fig. 4.
helpful in troubleshooting. By connecting the input envelope detector to the output of the r.f. amplifiers, the overall distortion of the entire 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. 13 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 nonlinearity 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 3000 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.

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 3021B is excellent for this purpose. It has dc amplifiers, which are best when monitoring speech because axis shift is avoided. Good high-frequency characteristics are necessary because the rectified r.f. envelope contains harmonies 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 in a double trace as shown in Fig. 17. This can usually be corrected with an RC network between one detector

tor and the oscilloscope. Such effects are easily remedied and an accurate linearity trace is not difficult to obtain.

The best method of checking the test set-up is to connect the inputs of the envelope detector 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 mixer 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 swing, lower the grid drive, or change plate loading.

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

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

Distortion Checking with a Selective Receiver

A fair idea of the d.c./r-f 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 75-A-3 with

Fig. 18 — Typical linearity traces.
A REGULATED SCREEN SUPPLY

As everyone knows, or soon finds out, tetode linear amplifiers require "off" screen-voltage supplies for lowest distortion. Earl Weaver, W2AW, was one pair of S85s in his output amplifier, and devised the circuit shown here to stabilize the screen voltage. It is a shunt-type regulator that delivers 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 VT tubes extinguish. The output voltage is equal to the sum of the VT drop plus the grid-to-cathode voltage of the 811-A. This grid-to-cathode 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 R4 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 screen current will be drawn. After the adjustment is completed, the meter can be removed from the circuit and the filament resist-up wire directly to ground. Since R4 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 VT tubes may be used to provide a regulated voltage near the desired value. VT tubes with various operating voltages can be connected in series, if the current ratings are the same. Two 811-A can be connected in parallel if higher current capacity is required. The maxi-

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mum current through the 811-A should be such that the manufacturer's max-operation rating is not exceeded. It may be necessary to adjust R4 for a slightly higher current under minimum load than is first expected, to compensate for full-load voltage drops in the high-voltage supply. As W2AW, the MA screen current varies from about 5 to 60 ma., and the shunt regulator holds the screen voltage constant to within 10 or 15 volts.
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. afflute, it has a "clock-sweep" brooch on the other side of the carrier frequency that does not pass 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 parasite in the final and 2.5 kc. filter—it must be the other fellow's receiver, etc., etc. One day I got a letter from a fellow who was doing some testing work for the National Bureau of Standards and some incidentals hence monitoring. The gist of the letter was that he thought he could determine some spatter on either side of the main signal? Sounds nasty, don't it? Stick in the back of my head, and I asked a fellow here with a selective receiver to give me a good going-over of the next time he heard me on the air and see what he could find. He reported spatter, 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 175 from 20 and began looking over the existing s.s.b. signals. The reports he passed out were nothing but grudging. 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 G.E., and who I was to argue with him. Don wrote me an extra-page letter describing types of distortion in linear amplifiers, as a fol-

lowing to a discussion we had over the air. I'm saving it for a day when I can understand malabroon. However, he made some statements and drew some pictures that I could understand, and that was the beginning of my seeing daylight. If I can piece along some of this daylight so that others can understand a few things about linear amplifiers without reacting to side, side, vector analysis, et cetera, that's maybe they can apply the principles to their equipment, as I did at W3ASW, and some of these "double-single-sided" single-sideband signals will be eliminated.

Sources of Distortion

As a start, let's quote Don's letter some of the things which clicked with me. Here's the first part:

DISTORTION IN LINEAR

GRAY HAIR IS ADS, OPEN HEARS

In a linear amplifier, departure from a true linear response fall into several categories:

1. Approximation when, with increased signal level, the gain tends to decrease.

2. Distortion which decreases with increased signal level. This is "feedback," "back-splatter," or

SINGLE SIDEBAND FOR
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 allow between stations. When the thing was about to go open, the sound would be all chopped up and the amplifier screeching about the cut-off point with variation on the same level. On Type 2 distortion, I’ve always called it “saturation,” or “flattening off.” You have heard it splattering all over the hand long before hands began using w.c.b.s. 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 errors of the two types of distortion will result in a series of harmonic imperfections, to be generated. In general, these harmonic outputs are not transmitted to the receiving antenna, since the amplifier is operating at radio frequencies but the harmonic imperfections may be in the output circuit, if the amplifier is placed in the output circuit and gain is large enough to transmit the output circuit. The harmonics generated by the harmonic imperfections may be transmitted to the output circuit of the receiving antenna, since the receiver is designed to operate at radio frequencies. The harmonic imperfections may not be transmitted to the output circuit, if the amplifier is placed in the output circuit and gain is large enough to transmit the output circuit.

When more than one signal is applied to the input terminals of a nonlinear amplifier, harmonic outputs are generated. Many of these will not appear in the output circuit of the harmonic imperfections are transmitted, but theory shows that all will have frequency components that are essentially in the same band of the input signals, and therefore will appear in the output circuit.

There’s nothing to getting a little deep, but the main thing to remember seems to be the fact about only one time applied to a distorting amplifier may not show up in the output circuit as a distorted signal in all cases. This leads somewhat to a general use of the term “amplifier” and a “scope” in adjusting a linear. Notice his standard “less than” now his standard “equal to” or “impressed.” That’s where a two-test tone comes in, the “curve for another way” and you will find that the proper interpretation of what you see with a two-tone test is an easy and simple way to adjust your amplifier. More so on this later.

The next two and one-half pages of the letter contains the most graph-awful holding mathematics and are the real reason the letter was written but they prove statements Don made to me in the extract and which I found hard to believe. I have the papers to several “handy brains,” when I know and they said it is all true, so I believe it and will just Along to you the example given and you can take it from me, it does happen “like that.”

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

THE RADIO AMATEUR
Many a/b transmitters consist of the genera
tor itself (filter or phasing type) followed by one or more linear amplifiers. Low power levels (up to 1 watt) are usually handled by receivers tubes run on Class A amplifiers, so as you have followed the lines in published tables for receiving-type tubes in Class A service, you should have no trouble with these low-power stages. Remember though, no inductively coupled output can be tolerated. This means ade-
quate, or better than adequate shielding, good separation of grid and plate leads, etc., and a fairly good Q in the tuned circuits. Some resis-
tance 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 450-v. am-
plier. This all follows receiver design practices and applies wherever receiver-type tubes are used.

Driver Stages

Whatever type of exciter is used, you even-
tually 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 usually
do some good. Many words have been written on the troubles encountered with 807 stages and Class B driver stages and, in all probability, no two cows will ever respond to the same treat-
ment. 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 814s. While the voltages available would make even an old-type 807 blush, I couldn't find the proper vage for an 807 in Class A service. Note the 807 as just a 6.66 with a top cap, inspection of the ratings of a 6.66 gave a set of voltages that are somewhat nearer here with my 807. Values of 350 plate volts, 250 rectified screen volts, and a 250-ohm cathode resistor will set the tube
up in Class A operation with a loudness of around 3000 volts and an output of from 0 to 10 watts.

Now, with all these voltages applied, my 807 promptly took on its own. I didn't get it to rain down until I applied the v.h.f. choices to operate properly. Then, with neutralization, I began to get a "cold" 807 but with not quite enough drive for the 814s. Some other paper
my experience may vary here with my 807. Values of 350 plate volts, 250 rectified screen, and a 250-ohm cathode resistor will set the tube
up in Class A operation with a loudness of around 3000 volts and an output of from 0 to 10 watts.

From here on, with all these voltages applied, my 807 promptly took on its own. I didn't get it to rain down until I applied the v.h.f. choices to operate properly. Then, with neutralization, I began to get a "cold" 807 but with not quite enough drive for the 814s. Some other paper
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effective capacity of 170 µfd, in the circuit. Com-
pare 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-µfd grid tuning con-
denser just won't do for a µA-7 tube amplifier in Class B where the tubes draw grid current.

What about that varying level? Let's try exacing to steady it a bit. What do I mean by examping? Just this — connect a resistance across the tank circuit, to dissipate some of the excitation and offer a more nearly-
constant steady load. Then, when the tube draws more or less grid current, the overall load
on the drive will not vary as greatly as without it. How much examping? That question has always started a lot of arguments on the air, and each fellow has his own idea about what is cor-
rect. In my case, I started with a rather high value of resistance, to keep the peaks under con-
trol, 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 formula given by Houston and was
near to that used to feed 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 examping resistor across each tank. This cut-and-try may not be the best method of ob-
aining optimum examping 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 eye-
pieces raise and arguments about that last one, but I'll stick by my guns.

Good luck! That's easy; follow the instruction
s for the tube. They will be OK for a starter and may not need changing in the last adjust-
ment, but more on that later.

In figuring the final plate tank, stay with the
Handbook. Use the expected wanted value of plate current at full signal with the plate voltage available and try to get a bit more on the high-frequency side. Usually your tuning con-
derenser won't be anywhere near large enough in capacity. Mine won't, and I had to make up the difference with paddles. Not so fast, vari-
um units from the Common set antenna relay box will work nicely, but don't push, and don't
across the tank from plate to plate. Use one or two
in a "split stage" arrangement, because unchecked harmonic currents will find a better path to ground than with this arrangement. The net
result here was a dual 170-µfd-section vari-
able with each section pushed with a 20-µfd
vacuum potentiometer. 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. Start the exciter uses all moothing-type tubes except the 807 driver, the best arrangement found with one good haiku low-voltage supply with the 807 plate taken off ahead of the regulator. The 807 screen and all other plates are regulated with: Vb-105 and Vb-150 in series, and the dropping
resistor adjusted for an average current through
of them of 20 ma. The plate supply for the final should be as "still" as you can make it. The line
regulation here at W3A7K is very poor, running on a 100-watt load will make quite a disker in the lights. This had me worried and I knew I would do a lot of hand switching 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 Norgard came to the rescue, with the suggestion that I use a large or output capacitor on my final filter as it was possible to use. The plate voltage here averages 1400 volts and the final leads at 30 ma., so I made use of one 10,000-µfd unit along with a 10-henry 500ma choke (sounding out, by the way) as a filter for the final. Use the diode-
input system with all the capacity on the output side. I managed to stay fairly clean under voice
operation, but a steady tone will push the output voltage way down. Incidentally, 60 to 80 v.
of on the exciter plate supply output won't do a lot of harm.

Before getting into the loading, it is assumed that you have good mechanical and panel
suppression in your amplifier. Here again, in-
sibility can't be tolerated! With plate voltage applied, the bias should be adjusted to allow the tubes to draw their maximum rated plate dissi-
pation without peak back-tuning currents. With-
out the antenna or exciter coupled and see if the dial spills over at any point. If it does, it
more work to do. This may be a tough baby to

share, but it is a must. When you are sure the
final behavior, return the bias to normal. My
811s required 45 volts with 7 turns of hook-up
wire awes and found them in each grid lead to
name them.
Testing with a Scope

If you don't have a scope, borrow one and make note of optimum conditions on your final coupling and find plate current, and then try to maintain these.

Fig. 1, 2 and 3 represent what you should see on a scope when a two-tone test is applied to the 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 revealing some flattening in the plate circuit plotted on it. Here the crossover are steep but notice the rounding and flattening vs. 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 -- below zero!"

What is a two-tone test? Just two a.f. signals repeated by about 1000 cycles applied to the

amplifier under test. How do you get it? On a gig rig with a "center insertion" or an "unbalance" control, open up a bit of center 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-taps, the intro-

duction of a tone of about 1000 cycles into the front end of the set and the damping of one balanced modulator will give a similar picture.

Now carefully advance the amount of both signals applied until you either reach the limit of the power-handling capability of your find, 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 find, forget it. But then don't consider the find. This goes for all stages, the way the. A good way to stay within these limits, if you own a scope, is to measure 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 taking. 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 radi-

ated. 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.

As for the tuning type of distortion (Fig. 1), I never have seen it here with annealed brasses. If it should occur, check your bias supply care-

fully 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 S1C's I use three flashlight cells in series, which is about right for my plate voltage.

Before I get using too far, let me say that the above-mentioned two-tone test should be ap-
plied 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 watt input to the S1C's. That is where I operate it.

SINGLE SIDEBOARD FOR
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 c.w. rig.
Curt Smith, W7CM, sends along the dope on the audio-oscillator
scheme he built into the speech amplifier and voice-controlled circuit
of his c.w. transmitter. As shown here, the only components needed to make
this addition are a few resistors and condensers. The switch S1 is
mounted on the 0.25-megohm variable pitch control. With these con-
stants, 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.

THE RADIO AMATEUR
To make your linear amplifier put out a signal that is above criticism you need three things: this article, the preceding one by E. L. WILL, and an oscilloscope. With all three—perhaps almost four—ingredients you really can put into practice whatever the other three tell you, you'll be right on the banks of the roteifiers.

**How To Test and Align a Linear Amplifier**

ROBERT W. EHRLICH, W4CUG, EK6WIN

It was generally 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 AM 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 noise from which the exciter is working so hard to eliminate them. In other words, it can make the test receiver 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 pseudosound 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, every 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) excessive grid and output stage circuits, which yield distortion, splatter, TVI, etc., or (b) underexcitation and too-heavy loading, resulting in inefficiency and loss of output.

The sprintest 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 linear manner and then observe, by means of an oscilloscope, whether the same waveform comes out of the amplifier at maximum signal.

**Test Equipment**

The simplest type of cathode-ray oscilloscope can be used for this test, or tests, as 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, as it might be an added asset to other test equipment tests but also as a permanent monitor during all other authorized operating periods. The scope should be capable of putting down at least that a scope be borrowed to make the line-up checks, whereupon the regular plate and grid meters can serve therefor 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 pickup device and its connections to the oscilloscope are shown in Fig. 1.

**Fig. 1.** The recommended coupling for applying r.f. and applying it to the vertical plates of a scope. The pickup loop is made by plugging the location of the scope where the horizontal plate is connected to the scope, E.L. — Receive to operating frequency. 50—90 Hz. Use or ceramic, 500 volts. R — 474 milliamperes. Replace used direct connection. Necessarily, the pickup loop should be clamped to the dimpling board, antenna tuner, or transmission line — in other words, to a point in the system beyond which any tuning adjustments are to be made.

The only other piece of test equipment will be an audio oscillator. Since only one frequency is involved, 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 tone-out test involves sending through the amplifier the system a pair of r.f. signals of equal amplitude and a thousand cycle or so apart in frequency. The combined envelope of two such signals looks like two sine waves folded on one another. If these waveforms come out of the final, will and good; if not, there is work to do. More about that later.

**SINGLE SIDEBAND FOR**
There are two common-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 Exciter:
1) Tune 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 adjust audio gain until (when the carrier and the sideband are equal) the "scope pattern takes on the appearance of full-"modulation tone; i.e., the curves just meet at the center line. See Chart 1, 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 Exciter:
1) Connect the audio input to the balanced modulator. In the W2UNJ exciter, for example, pull out one 805GT, or in the SSB Jr., place a short from plate to B+ on one section of the 6K7 tube used.
2) Connect the audio oscillator and advance audio gain to get the desired drive. Note that with one balanced modulator set 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 excitors, 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 R-F transmitters will recognize this setup as being the same, except that instead of one trapezoid, this test produces two trapezoids pointing toward each other.

Each individual triangle is subject to the same analysis as the regular trapezoid pattern; i.e., the sweeping ideas 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 audio input, 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 modulator as before. Also, with the latter set-up, the patterns will probably have a loopy appearance at first, and phase correction will be needed to make the figure close up. This can be done either by varying the audio frequency, or by changing the drive in the system.
CHART II—IMPROPER AMPLIFIER OPERATION

(1) Overdrive, indicated by flattening of peaks.

(2) Same as (1), double-trapezoidal test.

(3) Too much bias, causing output tubes to become shaded for
    greater rather than existing straight across output line.

(4) Same as (3), double-trapezoidal test.

(5) Too much gain with v.c.f. potentiometer. Note from dots or
    lines, as unlike ones the function will appear at the peaks.

(6) Too-protection with fundamental frequency parasitics,
    accompanied by overdrive.

(7) Siren, overdrive and parasitics.

(8) Voice pattern showing flattening of peaks due to over-
    driving. Where flattening is appar-
    ent at 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 a little into the microphone, otherwise the meters will always read less. In particular, the average input under two-tone inputs is close to 65 per cent of the actual peak input for a Class B amplifier, about 75 per cent for a Class AB stage, and 80 to 90 per cent for Class A. 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, 55 to 70 per cent for Class AB, and approximately 20 to 60 per cent for Class A.

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 245 watts and, to operate up to this rating, it should be fed up with a linearity test to about 150 watts input. Under normal voice operation, the meter will then read up to around 100 watts, but

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 two tests.

**Chart III—Improper Test Setup**

(12) Two r.f. signals unequal. In Method B, caused by improper matching of either source or output to amplifier. Spurious peaks due to leakage through divided-modulator or unequal impedance due to selective action of some high Q circuit of excellence.

(13) Same as (12), double-trapped test (Method B).

(14) Electrical audio A close to this defect is that successive waves are not identical.

(15) Same distortion as (14), but switched to double-trapped test pattern. Note that current changes regardless of poor audio signal.

(16) Carrier leakage through dividing modulator (Method B only).

(17) Same as (16), double-trapped.

(18) (Note clip (16)) Caused by inadequate suppression of unwanted sideband (Method A) or by r.f. leakage into horizontal circuit of scope.

(19) Doubly trapped with audio phase shift to too wide.
assumptions described previously. They are classified separately so as to show representing correct conditions (Chart I). Finally operation of the r.f. amplifier (Chart III), and various other patents that look irregular but which really represent a peculiarity in the test setup or the exciter but not in the final (Chart IV)

Aside from the problem of resonant, 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

in justice. The following conclusions may be taking:

1) For good efficiency, the final itself must be the limiting element in the power-handling capability of the system.

2) If the power level being driven to its limit, it should be loaded less heavily until such time as

3) If the power level obtained above is less than should be expected, no 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**

300-watt patterns taken at the output of a Class B linear amplifier with equal drive from both drivers. Measured input power is indicated.

(20) 90 watts.
(21) 105 watts.
(22) 150 watts.
(23) 300 watts.
(24) 300 watts.

assumed that the amplifier is not contributing any distortion to the signal as long as the peak power level indicated by the load 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 tube should handle quite easily. In such cases, some attention should be given to the plate loading as foreseen below.

The several patterns of Chart IV were made to show how loading effects 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 all heavier loading the output curve is increased, 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 decrease the final slightly while limiting is apparent, and if proper drive conditions prevail the pattern will improve when the amplifier plate is depressed.

The intermediate and driver gauges will follow the same laws, except that the thing called "limiting" 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 the driver to the succeeding stage is due to a poor match. Just as the Class B audio circuit, 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 arrangement will be surprised to find how effective it is to use the driven stage down on its tank—or otherwise to decouple the system. For example, an 807 driving a pair of

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**SINGLE SIDEBAND FOR**
PEAK LEVEL CONTROL

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, W2XQI, in making the charts for this article.

THE RADIO AMATEUR 131
Coupling to Matched Coaxial Lines

GEORGE GRAMMER, WIDF

The problem of designing a tank circuit to load an amplifier properly is solved quite simply if a suitable variable is adopted toward the solution of coupling. Many circuit engineers look for coupling systems that will, through the use of capacitors, inductors, or other similar accessories, provide the simple design procedure outlined below are followed.

In coupling a circuit to a coaxial transmission line, the impedance of the line must be such that it is matched to the source impedance of the amplifier. The impedance of a coaxial line is given by:

\[ Z_L = \frac{120}{\pi} \times \frac{1}{\tan \theta} \]

where \( Z_L \) is the line impedance in ohms, \( \theta \) is the attenuation in degrees per mile, and \( 120 \) is a constant.

**Resistance**

In a coaxial line, the resistance is given by:

\[ R_L = \frac{120}{\pi} \times \frac{1}{\tan \theta} \]

where \( R_L \) is the resistance in ohms, \( \theta \) is the attenuation in degrees per mile, and \( 120 \) is a constant.

**Load Resistance**

The resistance of the load is given by:

\[ R_L = \frac{120}{\pi} \times \frac{1}{\tan \theta} \]

where \( R_L \) is the resistance in ohms, \( \theta \) is the attenuation in degrees per mile, and \( 120 \) is a constant.

**Coupling the Line**

The design methods outlined below are based on a matched-line circuit - one whose input impedance
Inductive Coupling

The coupling reflects an equivalent resistance, $R_0$, across the plate tank circuit; thus $R_0$, is the load "seen" by the tube.

The $Q$ of the tank or primary circuit is

$$Q = \frac{R_0}{X_C} \cdot \frac{R_0}{2\pi f_0}$$

when $C_0$ is resonant at the operating frequency. If $C_0$ is out of tune to resonance, its $Q$ is

$$Q = \frac{2\pi f_0}{L_0}$$

With these conditions, the coefficient of coupling that will just raise the proper value of $R_0$ to be reflected across the primary is

$$k = \frac{1}{\sqrt{Q_0}}$$

With any smaller value of $k$ the reflected resistance will be too high, and the amplifier cannot be loaded heavily enough to obtain the desired power input.

For purposes 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 is usually convenient to obtain with available coils and condensers. Substituting $\lambda$ for $Q_0$ and rearranging gives

$$Q_0 = \frac{1}{k^2}$$

as the minimum value of $Q_0$ 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 at about a 1-inch coil, and 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 value of coefficient will be obtained when the smaller of the two coils is fairly similar in comparison with the larger, and when it is placed on the outside of the larger coil at its middle.

When the resistive method of coupling is used for coaxial lines the coupling coil usually is larger than the conventional "link" coil, so the coupling coefficient can be expected to run between 0.5 and 0.9, depending on whether the coupling coil is at the end or center of the tank coil. Using these values in the formulas above (for a tank $Q$ of 10) shows that $Q_0$ should lie between about 0.1 and 0.29, the larger value

Based on "Coupling to Coaxial Lines," QST, May, 1934.
being required for the smaller coefficient of coupling.

Applying the figure to a practical case, suppose that the frequency is 3.9 Mc, that $R$ is 4 ohms — that is, a 125-ohm line properly terminated so that the v.c. is 1 e.m.f. and that the coupling coefficient is estimated to be 0.5.

Then, the required Q is 0.5 and the value of inductance needed at $L_2$ is

$$\omega L_2 = \frac{2}{2} \times 3.9 \times 3.9 \times 0.5 \times 1.25 \text{ ab.}$$

The equivalent required at $L_1$ for tuning the secondary circuit to resonance is

$$\frac{\omega L_1}{\omega L_2} = \frac{4 \times 3.9 \times 10.5 \times 0.5}{3 \times 9 \times 0.5} = 1300 \text{ muf.}$$

Although this is a rather large value for $L_1$, it can be obtained by paralleling several broadcast-tube type condensers, since the voltage across the condenser will be relatively low even with fairly high power. Alternatively, a 250- or 300-

The principal disadvantage of using a higher Q than the minimum required is that the sec-

The secondary cannot be operated at exact resonance, which would cause $R$, to be lower than the des-

The tuning of the secondary circuit becomes more critical.

Since the circuits are overcoupled when $Q_1$ is larger than necessary, the reactance reflected into the circuit is far less important when $Q_2$ is determined. There is thus some effect on the setting of the inductance tank constant $C_2$, for resonance.

Although undesirable, a tuning standoff, a moderate amount of such detuning is unimportant for so far as performance of the am-

The $Q_2$ makes a smooth and ope-

The reactance on tank tanks last mentioned can be calculated by using the arrangement shown in Fig. 3. Here the secondary circuit is split into two parts, $L_1$ and $L_2$, and it is shown that there is negligible coupling from $L_2$ to either $L_1$ or $L_2$.

The field difference between line 2 and line 1 that has an important bearing on the

The left side of the figure shows the case where the inductance of $L_2$ exceeds the altered.

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**Single Sideband for**

**General Application of These Circuits.** In Fig. 1, increasing the inductance of $L_2$ also increases the voltage induced in the secondary circuit (even if the coefficient of coupling does not change) because $L_2$ is made larger the mutual inductance between $L_1$ and $L_2$ also increases. In Fig. 3, adding $L_1$ has no such effect, merely the mutual inductance between $L_1$ and $L_2$ remains the same.

For this reason the inductance of $L_1$ does not enter into the value of $Q_2$ when the formulas given earlier relating $Q_1$ and $Q_2$ is used. For purposes of determining coupling, $Q_1$ must always be based on the value of inductance in $L_2$ alone. Hence $L_2$ may have the proper value for suffi-

The $Q_2$ is 200 muf. (as well will inside the range of 250 muf. condenser), then the total secondary inductance required for resonance at 3.9 Mc is

$$\frac{\omega L_2}{\omega L_1} = \frac{4 \times 3.9 \times 10.5 \times 0.5}{3 \times 9 \times 0.5} = 1300 \text{ muf.}$$

Subtracting the inductance of $L_2$ from this gives 700 ab. as the inductance required at $L_1$.

In the circuit of Fig. 3, then, $L_2$ and $C_1$ simply provide a variable reactance by means of which the secondary circuit can be tuned to resonance, and $L_1$ may be varied upon as an easily varied way of extending the effective range of $Q_2$. (For example, the reactance of $L_2$ is just a bit smaller than that of $L_1$, the result 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 $L_2$ 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 $Q_2$. This can be done by making the primary circuit higher. If the C.R. ratio is increased by re-

I give a further improvement in coupling.

The limits of the method on resonating coils are almost irremediably too small, on the lower fre-

Measurements show that the link inductance on some of the medium-power coils for the 3.5-Mc. band is not even as much as 1 ab. As the calculation above shows, this is not large enough for coupling into a 75-ohm line operating at a reasonably low a.e. , unless a plate tank Q considerably higher than the usual 10 or 12 is used. If $L_1$ is 0.75 ab., $Q_1$ becomes

$Q_1 = \frac{2L_1}{R} = \frac{2 \times 3.14 \times 3.9 \times 0.75}{70} = 0.216.$
Assuming the same coefficient of coupling, than the required $Q$ is

$$Q = \frac{1}{4\pi} \times \frac{1}{0.52 \times 0.241} = 16.3.$$  

This assures, of course, that $L_1$ and $C_1$ are used to resonate the secondary circuit. Without such tuning a still higher tank $Q$ would be necessary for sufficient loading.

**Checking Circuit Values**

Cell dimensions can be calculated from the formulas in the Handbook or by means of the Loadling 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 cells to the proper inductance by measurement. A gold-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 $L_0$, the plate tank coil, although either the formula or Calculator is simply necessary for such coils.

**Pi-Network Tanks**

For the special case where a 71 network is not coupled 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 resistance needed when the $Q$ is held at 10. The curves are carried only as far as 140 ohms for the input resistance, since higher values can readily be obtained with a 3000-ohm variable condenser even at 3.4 Mc. When the $Q$ is 10 (or more) the resistance of the coil averages 25 or 30 ohms higher than the reactance of the input condenser. It should be noted that the output resistance 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 2000 ohms. A higher tube load resistance cannot be "matched" unless the $Q$ is increased above 10. Corresponding conditions obtain with the 750-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 one value, a medium value for both 50- and 750-ohm lines, and one that is useful over at least the range 50 to 100 ohms when the inductance is 1000 ohms or more. In this case the curve for the 750-ohm line is extended upward to 1500 ohms.

![Fig. 1 - Pi-network design curves for working into 50- or 75-ohm loads with a Q of 10.](image1)

![Fig. 2 - Pi-network design curves for working into 50- or 75-ohm loads with a fixed output capacitance of 40 ohms.](image2)

A "match" is the best measure of your transmitter 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 before that height with noise. Don't worry if the plate meter doesn't look the way you'd like it to. This slope tells you you're getting all the peak power your amplifier can give, and it's telling the truth.

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The "Single Side-Saddle" Linear

CARL W. ECKHARDT, W2EEK

The Side-Saddle Linear is designed for the low-power station and will operate from a 50-watt 000-volt power supply. 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 noiseless 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 Roque and Long, and using straightforward mechanical and electrical layout. The resulting amplifier is stable and behaves properly, just as the experts say it should.

The receiver has been connected directly to the antenna coax transmission line through an electronic "TV" switch, and although the 807 is not used to cut-off during receive periods (for instance), the linear is perfectly quiet, with no trace of thermal noise.

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In use, the linear amplifier is mounted in a wooden case above the battery, crystal, filter, etc. The batteries are connected across current, grid current and plate current.

The drive requirement is approximately 2 watts. As Fig. 1 shows, the grid is series fed, and 33 volts of bias is supplied by batteries. Since approximately 1 no. 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 2-inch aluminum shield running the depth of the subchassis, near its center, thereby isolating C2 from C3. Further to guard against instability, small 1-ohm nonlinear

tive resistors R5 and R6 are placed in the grid and screen leads directly at the socket. The cathode is grounded to the tube-support chassis with a short wire at the socket. The screen bypass condenser C6, a disk type, is wired directly across the tube socket, keeping leads as short as possible. No additional resistor R5 provides the proper amount of screening.

The voltage given for the grid and plate tank circuits should be adjusted if proper circuit Q's are to be maintained. The high voltages specified must be applied to get the desired L/C ratios. Three turns are required on both L2 and L3. The end limit of L3 35 in. provides 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 L3. Remove a small section of the control covering from the 3rd, 4th, 5th and 6th turns from the ground end of L3 and carefully solder the chassis 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 300 is provided with a fuse shield. Grid and plate tank circuits are near the left and right-hand ends of the chassis as illustrated. You will note that J1 and the rotor of C1 are at -33 volts bias potential and must be insulated from the chassis. This is done by slightly enlarging the mounting hole for C1 and J1, 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, 7½ inches long by 3¼ inches high, with a 1/2-inch lip, should be placed as mentioned earlier.

A pilot light, a power switch, and J3 are mounted on the front center of the chassis as shown. (The pointer meter located below the chassis is not being used and should be ignored.)

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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 connection to the linear, and applying power, the grid current should be 0 and plate current 5-15 ma. The amplifier should be perfectly stable without a trace of self-oscillation or parasitic. In applying grid drive and antenna loading, adjust for the following guides:

<table>
<thead>
<tr>
<th>Grid Current</th>
<th>Plate Current</th>
<th>Screen Volt.</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

Minimum grid, bias: 0 ma, and 0 ma.

Normal grid: 100 ma, and 100 ma.

Normal bias: 0 to 15 ma, and 0 to 15 ma.

Relative readings are suggested for output r.f. current, as this will depend on the impedance of the line and the a.m.r.

Tap up from the ground end of L8 with the antenna tap until the above maximum peaks are reached. Provide just enough excitation to the linear by adjustment of L9 and gain control of the exciter so that the above maximum peaks are achieved. If additional gain or coupling of L8 will not give the peak readings desired, additional coupling to the antenna will be required.

If you are overswinging, 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.

**VAR Switch**

The TR switch in use is shown in Fig. 3, and is a modification of Crown's system. It shows a receiving loss of about 1 decibel on 75 meters over a direct connection, and it protects the receiver to the point where a maximum of 4 volts reaches the receiver audio terminals.

**LINEAR-AMPLIFIER TANK C**

Transcending tank experience in the tuned plate circuit of a linear amplifier can have an adverse effect on the linearity, but if the requirements in this department are no more severe than in r.f. amplifiers used for other purposes. Theoretically, the tank Q should be made with pure Class A operation, with in and more Q needed as the operation passes through AB, and AB 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 wave transformers.

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SINGLE SIDEBAND FOR
The "Little Firecracker" Linear

BEN RUSQ, W2QZ

A

linearization of the output of any low-power neq. s.s.h. signal can be done only by using a linear amplifier. A linear is an amplifier so adjusted that the output voltage is proportional to its input voltage. Use can be made of Class A, A/B, A/B or B. Class A is generally used for very low power levels, as in the earlier output stage.

Not only do we want to amplify our s.s.h. signal, we want to amplify it without adding new and possibly undesirable signals. The unwanted sideband is 40 db down in the signal coming from the oscillator, so we expect the noise (or very close to the same ratio) into the antenna after amplification. We want no intermodulation products added that will either degrade the desired sideband or appear as "noise" outside the sideband. In short, we want a really "high-linearity" amplifier for s.s.h. As in radio work, we 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 tube.
2) High power sensitivity.
3) Multifound operation without opening the cabinet.
4) High-Q LC circuits.
5) Constant-voltage plate, screen and grid supplies.
6) Stability.

Tuses GIP, September, 1953.

The "Little Firecracker" s.s.h. linear sees two 6146's in parallel. It operates on 80 through 10 meters and covers from 3.5 to 30 megacycles. All the supplies are built in the same cabinet as the r.f. components. The two decks at the rear were taken from some war-surplus equipment.

Linearity and Output

It should be recalled that the figure given for audio service in Class A, A/B, A/B or B can be used for r.f. linear amplifiers used with s.s.h., 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 A/B, and A/B ratings of the 6146, an excellent tube with a plate-disipation rating of 25 watts. Slightly more output can be obtained in Class A/B operation, although raising the tubes in Class A/B (no grid current) simplifies the drive problem and greatly reduces the chances for distortion on signal peaks in the grid-current region. In s.s.h. suppressed-carrier operation the maximum screen voltage (260) can be used, resulting in higher power sensitivity and slightly more peak output. The "maximum signal a.c. plate current" is not what your meter reads on peak.

High Power Sensitivity

The 6146 is tailored for the desirable feature. Like all beam tubes, the 6146 requires very little drive, and you can figure practically "line" 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 Million grid-dipper. A pair of 6146's will deliver full rated output when driven by any of

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the commonly-used s.a.h. 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. Strooping is used in both the crystal back circuit and the linear-amp grid-tube tank at W3KX. The amplifier operates Class AB, with 120 watts peak output.

Multi-band Operation

Multi-band operation is not essential to s.a.h., but, as in any other transmitter, it is a nice feature to have, if a shell 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 100-meter band could have been included by switching in an additional inductor in the plate circuit.

High C

The subject of high C (or high C') in tuned circuits has been stressed in many articles relating to s.a.h., as well as in the Handbook. High-C circuits are used in linear amplifiers for two main reasons: for ease of coupling to other circuits, and to minimize the harmonic content. A quick thumbs-up that the s.a.h. 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 HEEL 333-tube turret used in the grid circuit (an red-back assembly with separate link windings for each of the five coils). A 250-sq. mm. variable is used to tune the coil in use.

The Q in the plate circuit can be set, with reasonable accuracy, since a variable is included, and a high-reactance variable condenser is used in a plate circuit. The condenser is a 100-pf. midget, and the shakers are connected together by a ½-inch-wide copper strip. Connected this way, the condenser measures 80 to 300 μf. (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-frequency 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 kilowatt changes are made with the output condenser set at 80. The Q chart in the Handbook was used for working out three settings - the maximum value of d.c. plate current (227 ma) is used in the d.c.-modulator grid.

Regulation of Power Supplies

In no other type of transmitter is the regulation of power supplies as important as in s.a.h.

SINGLE SIDEBAND FOR
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 gives the necessary 250 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' drop, 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 voltage variations, a huge amount of output capacity is recommended, and too many microfarads at this point are not within the realms 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 cur- rent 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 trans- former 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 "swinging." In s.s.b., only one is sufficient, and using only one cuts down another source of series resistance and poor regulation. With 24 ft. of filter condenser and the swinging choke, the regulation is improved and the ripple is only 25; per cent.

TABLE II

<table>
<thead>
<tr>
<th>L. L.</th>
<th>B &amp; W STEL, 36-West Cool Tunnel</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td>1.0</td>
<td>1.0</td>
</tr>
<tr>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>2.0</td>
<td>2.0</td>
</tr>
</tbody>
</table>

A close-up view of the c.f. section shows the smoothing condenser mounted close to the two 1140 plate connectors.
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. Variations manifest themselves as fundamental oscillations or regeneration, grid and v.h.f. penalties. 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-amplifier condition.

Oscillator or regeneration at the operating frequency or evidence of plate-grid feedback and the amplifier was neutralized to offset such coupling. As originally built, no neutralization was included, although every precaution was included to minimize feedback, 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 hazing between 75 and 90 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 (101) in the form of "hazards" across the bond, similar to a transformer. This was an unexpected i.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 probably would have solved 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, Rg, was included in order to create proper warm-up time for the 6X4 and 6146.

Construction
The panel and chassis are aluminum, with an 8X4 X 3-inch panel and a 10 X 17 X 4-inch single sideband for
The large chokes allow 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 console. This spacing also helps to bring the rear agree, with its caps, 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-tune shield box, permitting short lead lengths.

A series of L-inch holes is drilled directly above $R_4$, the voltage bleeder, for better heat dissipation. A current of 40 ma. flows through this resistor when the plate switch is "on."

Condenser $C_{P 0}$ is mounted as close to the output each connector, $J_2$, as possible, and flush against the chassis. It is "grid-dipped" with $L_0$ 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, $C_{P 0}L$ (Fig. 2), is mounted on an insulating washer be-

![Wiring diagram of the power supply.](image)

cause the run is negative and does not ground to chassis. A disk ceramic condenser, $C_{P 0}$, is used to bypass the run to the chassis.

Type t mixers condensers are used in the pi-

network output stage because they lend them-

selves to firm mechanical mounting. Their volt-

age rating is higher than required for the power

used, but they are on hand. The "safety choke," $K_{P 0}$, in Fig. 1, is mounted across the condenser

both sides of the plate-tuning stations are con-

nected together with 14-inch copper straps to

reduce lead inductance and make use of the

condenser's maximum capacitance. The self-

resonance of this condenser, $C_{P 0}$, connected this
tay, measures 120 Mc. — well outside any TV

tand.

The plate choke, $K_{P 0}$, is mounted instantan-

eously to locating the "hot end" directly centered

over the plate range of the 5670a. Here, too, 14-

inch copper strap is used to connect the end of

the choke to the plate blocking condenser, $C_{P 0}$,

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

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move an expected "hole" out of the U-meter hand.

Parts B1, BFC, and C1 are mounted just inside the input tuned shield cage, and directly in back of the 6H46. A lead is brought through close to the bottom of the vertical side of the shield cage, by means of a feed-through molybdenum. This lead connects to the vertical part of the neutralizing condenser, C2. The neutralizing condenser is made of a thin aluminum sheet cut in the shape of the letters "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 diode.

**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. (b) A condenser of this type may 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. (c) If "pumping" of the plate will occur before rated output is reached, meaning proper drive requirements are not 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 i.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 triode tubes, because of their high power selectivity.

With the grid circuit tuned to resonance at the fundamental, the pi-rivet should be tuned starting with the lowest number of turns in the variable inductor. The first resonant point is the desired frequency.

**Results**

This linear has been at the air regularly on 20 and 27 meters, and has been dimly-light 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 as 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 out-gives a load topic.

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 one of the seven local TV stations, was completely unaware of its being in use. Furthermore, at this time, the linear was out of its cabinet and no line-drive filter was used.

For his patient efforts photographing the "Little Firecracker," special thanks to Henry Massey, W2AJN.

There is no magic in s.s.b. that makes it easier than T.V. work. A.M.

If it, 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 solid-state amplifier, of course, but it must be cured at the receiver, in any case.

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SINGLE SIDEBAND FOR
The "Four-In-Line" Linear

GEORGE GRAMMER, W1DF

..."the only exception..." deliver considerably more power...of the order of twice as much, in most cases...when driven into...as in AB...This is not always an...effect, because the savings in other...the number of f...The extra plate dissipation provides a...are as follows...against grid-current operation...1) The extra plate dissipation provides a...battery may be used...This linear amplifier uses four 100-watt...when driven into...reduce the percentage change in...The only important...in twos as much, in most cases...when driven into...the number of f...The extra plate dissipation provides a...The load...The extra plate dissipation provides a...Three important...of the grid circuit...This linear amplifier uses four 100-watt...This linear amplifier uses four 100-watt...
second too long in the mile 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, settled of selecting the power supply to fit the tubes.

Speech waveforms are such that in s.a.b. linear operation the maximum demand—that is, d.c. as indicated by the plate milliammeter—is 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 syllable 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 component, 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 layout of these transformers, in terms of voltage, is the 600-watt 240-volt type. The 240-volt output capacity of such a transformer should be good for peak d.c. inputs of 200 to 240 watts on the above basis. Achieving such peak power inputs to a linear amplifier would not be easy at 600-watts, 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 152As were used in the amplifier described here because they can all be purchased cheaply in surplus, also, 87As can be substituted and are not all expensive.

A plate voltage of 1000 is in excess of any recommended voltage given for these tubes by the manufacturers. However, with Class A/B operation at this voltage the plate current and plate dissipation are well within the normal ratings.

The plate voltage, while higher than the a.v. rating, is not as high as the maximum voltage needed on peaks at the plate-modulation rating.

Optimum Class A/B operation of 152A or 807A 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 fully well four tubes are needed, taking a total peak power input of 288 watts. As stated above, this plate is about twice the maximum demand on the supply, with speech, so the maximum d.c. power is no more than 140 watts. The peak tube output, as taken from the characteristic curve, is approximately 200 watts from the four tubes.

B.F. Circuit

The logical circuit arrangement for four tubes is to use them in push-pull parallel. As shown in Fig. 1, parallel push-pull is used in this amplifier, principally to take the d.c. off the plate-bank coil for safety reasons. The choice originally was the 152A, 240-volt 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 choices will not stand the peak r.f. voltage (continuously). However, the impedance of the 255-A type turned out to be unduly low by 15 meters, as a number of other types were checked. The ones finally used were Mullens type 34140, 2 m.h. these showed good characteristics in all bands.

Tubes in A/B require no driving power except that necessary to overcome circuit losses, but most excitors do have a small amount of power output available. This "static" power can be used to eliminate a tuning control. As shown in Fig. 1, a fixed-bias grid circuit is used. A circuit having a Q of 8 or 9 will have substantially uniform response over 2800 kc. Band centered at 3000 kc, so the L/C ratio of 0.25 is chosen to give approximately this Q in conjunction with the loading resistors, R1, and R2. The values of R1

SINGLE SIDE BAND FOR
Fig. 1 — Circuit of the r.f. portion of the long-wire amplifier. Unless otherwise specified, all components are in µ.

Gm — Copper tube 3/4" wide, app. 3/4" in length, 3/4" overlap.
Cg — 24000 µf at 660 volts, 6/8-inch spacing.
Ca — 20000 µf at 1500 volts, 1/4-inch spacing.
Le — 25 turns No. 22 enameled on 1/4-inch diameter copper tube, 1200-cycle per inch spacing. 6 turns per inch are used at 4500 cycles, 10 turns at 2750 cycles, and 12 turns at 1750 cycles. Le is wound with 1/16-inch No. 20 enameled Cu wire. A ferrite core 1 1/8-inch diam., Lc, and La made from B & W coil stock, C2 a 1/2-inch diam., (1997 and 3999), J4 1/2-inch diam. (1996), all enameled stranded or Milford 9955 plug base.
The grid tuned circuit, embodied by shaded line, is mounted in Miller 1440 chassis base and shield.

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close to a 4-to-1 a.m.r. in 75-kilohm line at the band center. On the 75-meter band the maximum a.m.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.

This method of adjusting coupling is a great convenience, since the center and spanner 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 a.m.r. bridges cannot be used in such a case because the a.m.r. has to be measured at the full driving power level.

Dowistic 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 defying inductances $L_6$ and $L_8$. The other, at about 150 Mc., was the usual type with those tubes and was similarly cured by $L_6$ and $L_8$. Except in cases where resonance at a frequency in a particular TV channel has to be avoided to prevent TVI, nothing very critical about these could be beyond the fact that they must be large enough to put the plate-circuit load resonance below the self-neutralizing frequency. With the constants given, this resonance is at about 35 Mc. This is close to Channel 2, so anyone in a Channel 2 region who is interested

is a similar circuit arrangement would be well advised to increase the inductances of $L_6$ and $L_8$ to move the resonance frequency lower.

The circuit is cross-neutralized by means of $C_5$ and $C_6$. Although the amplifier was wholly staked, instead of self-neutralization at the operating frequency was concerned, without neutralizing the tube 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 milliamperes, so the neutralizing condensers were installed to reduce this reaction in the thought that it would improve linearity.

Power Supply

Tests on a "100-volt-200-ma.» transformer had shown that it could deliver an a.c. output of 200 ma. from the entire high-voltage secondary without undue heating, operating continuously over a period of a few hours. This is an ac. 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 drop in the rectifier tubes and choke, the output voltage is close to 1000 at a load current of 50 ma. of 200 ma.

The entire output current could be available for the amplifier phase if the tubes were used in a "free blinder," but in the circuit of Fig. 2 part of the current is used for the screen grid to avoid the necessity for a separate screen
the driving voltage is higher on one side that the other, the tube's plates on that side will be driven to peak output before them on the other side, and will start saturating or "flattening" before the full output of the amplifier is realized. The con-
densers in the grid tank circuit, C1 and C2, should be matched in capacitance within a per cent or two, and the usual precautions to be taken in maintaining circuit balance should be observed. The r.f. voltage balances can be checked with an r.f. probe and r.f. voltmeter. Another checking method would be to provide individual by-passes in the cold ends of R1 and R2, running out separate d.c. return leads to separate grid-current picks to note whether or not the tubes on both sides of the circuit start taking grid current at the same time.

The oscillation pattern of Fig. 3 show the effect of grid bias on the instability. The peak output was the same in both cases, but in the upper pattern the grid bias was set so that the total grid current of the four tubes was 190 ma. with a small peak current (approximately 20 ma. total plate current). In the lower case, the grid bias was set so that the total grid current of the four tubes was 190 ma. with no significant peak current. While the lower case was the best that could be obtained when the bias was at a peak level. If changes in the self-oscillating circuit from the original to the present form are not substantial, the improvement in linearity resulting from operating at the lowest possible bias is explicable, to say the least. The pronounced curvature in the bottom region of the characteristic is typical of unshielded triodes. In selecting the plate current by adjusting the bias voltage it is advisable to make sure that no tube is over-
loaded. 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 the
**data 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 earphone is low, the optimum condition should be secured with a plate current of 150 to 180 ma, when the driving voltage is just at the point where a trace (a few microseconds) 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 earphone 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 offensomwr plate current should be only 10 ma or so larger than the "in-tune" current. Some sort of ZL output levelmeter such as an antenna ammeter is helpful; the output should start to drop immediately as one sets 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 idea is to give 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-decay capacity for a.m. operation at the same peak output, the linearity also can be checked by the customary a.m. method if the earphone can furnish an a.m. signal. The trap-ended pattern should be used, and a very simple scope such as shown in the Handbook will suffice.

In wave operation using a resting plate current between 100 and 120 ma, the maximum plate current should not be exceeded in the s.w. 14425 circuit.

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**Construction of the plugin grid tanks.** The inductances of the two coils are adjusted for an input impedance of 5 ohms at the center of the band. Grid proving of the grid coil can be for adjusting the spacing of an end turn as in this ZL, assembly. The coil form is measured on a strip of tin-foil material which isstatements on the sides of the plugin input, has current as regulated by the plate meter on voice peaks is about 150 ma.

**Additional Notes**

Fig. 1 shows that each 14425 screen has its own b.gas condenser which, so far as the operating frequency is concerned, simply means that the twin-screen circuit is by-passed with a resistance four times as large since the impedance of the connecting leads is very small. The separate condensers were used in an attempt to prevent the parasitic oscillations involving the screen circuits. Since the small coils in the plate leads set the parasitic question very quickly, no such damping was needed. The condensers may be useful, however, for good suppression of v.h.f. harmonics.

Although the usual practice of shielding wiring with desk by wires was followed as a matter of course, the amplifier was not shielded for v.f.t. Shielding is not necessary for 25 meters, but is likely to be required for i.m. — and perhaps 7.5-Mc. — operation in locations where a harmonic falls directly in a channel having a weak. TV signal. Close AR operation does help — it is only necessary to look at the TV screen while the driving voltage is modulated into the grid-current region to see that — but it is not a complete cure 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 grid 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 ZL output terminals for convenience. C1 and C2 are adjusted by bending the metal tabs from which they are constructed, to vary the spacing.

**SINGLE SIDEBAND FOR**
A Two-Stage Linear

BYRON GOODMAN, WI9X

Preamplifiers may be made to operate satisfactorily as Class B c.f. amplifiers, but the driver problem is greatly reduced if second-tube tubes can be used. When negative-voltage tubes are used, such as any of the tetrodes and quads of the tetrodes, the load on the driver changes as the signal swings in and out of the grid-current region, and the driven's load put out 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 negative" are used across the grid tank to offer a more nearly constant load to the driver. These resistors were driven point; however, and it is apt to be able to avoid using them. The S1-A is our choice for a Class-B stage.

From QST, March, 1941.

FIG. 1 — Wiring diagram of the linear amplifier.

C1 — 100-uf. variable (Millen 24V4).
C2 — Fuse, fuses, mounted on RG-25/U. Active section, 10 filaments.
C3, C4, C5, C6 — ceramic trimmers.
C7 — 400-uf. variable (Millen 2045).
C9 — 0.001-uf. D20-6c.41: 2-5000-ohm wirewound.
C10 — 10-ohm, 10-500-ohm wirewound, with fixed through hole (Heat NC-550).
C11 — Dual variable, 250-pf. per section, 0-475-kc spacing (National RG-250). All resistors are composition, not wirewound.
R1 — 150 ohms, 1/2 watt.
R2 — 480 ohms, 2 watts.
R3 — 2700 ohms, 1 watt (four 2500-ohm in series-parallel).
R4, R5, R6 — 9600, 1 watt.
R7 — 30 kohm, 1 watt.
R8 — Input connector (Bantam 8-H11D).
R9 — Capacitor-tube connector (Cahoon 82FR).
L1, L2 — toroid No. 42, 22 turns, 154-inch-diam. 154-inch long.
M4 — 0-10 milliamperes.
M5 — 0-20 milliamperes.
M6 — 2-5000-ohm, c.f. choke.
M7 — 3-5000-ohm, c.f. choke.
M8 — 3-5000-ohm, c.f. choke.
M9 — 3-5000-ohm, c.f. choke.
V1 — 6.3-volt B-amp, synchronous (Stanley P-6405).
cathode bias. A small resistor, $R_b$, in series with the control grid, and a small plate-leakage condenser, $C_p$, are included to eliminate the tendency to oscillate at various unwanted frequencies, as can be expected of a heterodyne running Class A. The 807 is grid-coupled to the grid circuit of the S11-A to outline the ground returns to their respective stages and to provide a stable means for adjusting the coupling. Since the overall 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 overall feedback.

In the S11-A stage, the chokes $L_1$ and $L_2$ were required to define a v.h.f. parasitic that showed up, and $R_2$ knocked out a low-frequency one. The two resistors, $R_1$ and $R_2$, in the rectifying circuit may seem a little unusual, but they were found necessary to kill a v.h.f. parasitic. Reducing the grid circuit slightly with $R_2$ killed the last traces of instability. If all of these suppression devices make it seem like the amplifier is a hotbed 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 element by-pass capacitors were used because they weren't found to be necessary. The filament winding was done with shielded wire, however, and this helps a little filament-to-ground capacity. All of the non-c.d. leads were made with shielded wire, and were found to be adequate. Although it is probably not necessary in many cases, a shielded wire is a good idea here because of the possibilities for overall feedback.

Construction

The amplifier is built on a 13 X 17 X 3-inch aluminum chassis. A 1N4148-halvishunt diode panel is fastened to the chassis by the socket and two slant 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 RCA No. 5403 3-inch diameter shield cap. The plate coil plugs into a five-pin socket in the same place.

A two-stage linear amplifier for boosting the power level of a code 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, $C_p$. Another 3-inch diameter shield cap protects the 807 plate coil. The plate by-pass condenser, $C_p$, is mounted under the chassis near the 807 socket, and the "cold" lead from $L_2$ and $L_3$ is brought down to it in shielded wire. The coaxial condensers, $C_1$, are made of a length of RG-59/U and drop down through the same chassis hole as does the shielded "cold" lead. The grid coil for the S11-A is shielded by an RCA No. 21942 4 X 5 X 2-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 lip on the top of the cabinet should be bent down to allow more room for the hand that changes coils.

The output tank condenser, $C_p$, is mounted on the chassis with aluminum brackets that also support the jack bar for the output coil, $L_3$. The variable link mount on the jack bar and is connected to the panel control through two flexible couplings and an extension shaft. A B & W 23485 shielded link was used for the output link, but an ordinary link might serve just as well in certain 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 daily current of the tubes. With 600 volts on the plate and 300 on the screen, the plate current will run around 48 ma. If it differs

SINGLE SIDEBAND FOR
At this value, it should be brought back by changing the cathode resistor, $R_v$ or the screen voltage. If the available plate voltage is something other than 600, adjust the plate current for a stable 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 "still" and not vary in output voltage. A low-level emitter 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.

A signal from the emitter is at $J_1$ and, with plate and screen voltage on the 807 but no plate voltage on the 811-As, resonates the circuit for maximum grid current in the 811-As. It should be easy to run this 50 ma, without any change in the 807 plate current. Cut the excitation back to where the 811-As grid current is about 25 ma, and resonate the output stage. You can use the "lick" in the grid current as $C_1$ is tuned through resonance, but a more sensitive indication can be obtained by using a crystal.

### COIL TABLE FOR TWO-STAGE LINEAR AMPLIFIER

<table>
<thead>
<tr>
<th>Band</th>
<th>furn. wires No.</th>
<th>Dia.</th>
<th>Length</th>
<th>ab.</th>
<th>Sparking</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.5</td>
<td>22</td>
<td>22 wires</td>
<td>1</td>
<td>14</td>
<td>19 4 4 1/4</td>
</tr>
<tr>
<td>14</td>
<td>100%</td>
<td>22 wires</td>
<td>1</td>
<td>1/4</td>
<td>2 3 1/4</td>
</tr>
<tr>
<td>3.5</td>
<td>22</td>
<td>22 wires</td>
<td>1</td>
<td>7/4</td>
<td>11.2 6 7/4</td>
</tr>
<tr>
<td>14</td>
<td>11</td>
<td>26 wires</td>
<td>1</td>
<td>7/4</td>
<td>2.5 3 1/2</td>
</tr>
<tr>
<td>3.5</td>
<td>22</td>
<td>22 wires</td>
<td>1</td>
<td>11/4</td>
<td>9.1 8 1/4</td>
</tr>
<tr>
<td>14</td>
<td>12</td>
<td>18 wires</td>
<td>1</td>
<td>13/4</td>
<td>3.3 8 1/4</td>
</tr>
<tr>
<td>3.5</td>
<td>19</td>
<td>14 wires</td>
<td>1</td>
<td>7/4</td>
<td>20 8 1/4</td>
</tr>
<tr>
<td>14</td>
<td>8</td>
<td>6.5 tubing</td>
<td>1</td>
<td>7/4</td>
<td>2.3 3 1/4</td>
</tr>
</tbody>
</table>

When the biasity 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 exact excitation condition as before. You are then in business.

At an exciting having a peak output of about 80 watts, an amplifier must be used. Simply disregard the 807 driver stage and connect the output coil of the exciter to the grid of $J_1$-As, replacing the link coil coupled to $J_2$.

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THE RADIO AMATEUR 153
813s in a High-Power Linear

JOHN J. SIMON, W5RCE

The push-pull 813 linear amplifier shown in the photograph was designed to fulfill a desire for an s.s.b. final that would handle the maximum permissible input, and be driven by a General Electric STA rectifier or its equivalent.

The circuit, shown in Fig. 1, is quite conventional except for the use of the National MB-30E multituned tank in the grid circuit. The choke L2 and L3, wound directly in the plate leads, in conjunction with L4 and L5 in the grid leads, eliminated all traces of c.w. parasitic oscillation. The combination of 2N724s, 6AN7s, and the 1000-ohm resistor in the bias lead similarly takes care of any low-frequency parasitic. The amplifier is rectified by C1 and C2 whose construction is described later. It is believed that the individual filament transformers help in maintaining circuit balance.

Construction

The amplifier is built on a 12 x 13 x 1-inch aluminum chassis. The tank condensers are fastened directly to the chassis, along the rear. The jack for the talkie end, elevated on ceramic pillars, is placed immediately in front of the high condenser. The tubes are mounted symmetrically in position and the high condenser is, of course, shielded completely from all excursions of the cathode power lead. The high condenser is the same as that used in the amplifier A.M. linear.

Fig. 1 — Circuit of the high-power push-pull linear amplifier.

C1, C2 — Neutralizing capacitors — see text.
C3, C4, C5, C6 — 0.001 µfd, 1.5 kV, disc ceramic.
C7 — Full-wave rectifier, 300-µf.d. disc ceramic, 1/2-inch spacing (Johnson 200HD50).
C8 — 0.005 µfd, 1.5 kV, disc ceramic.
C9 — 500-µf.d., 5-kv, disc ceramic.
L1, L2 — 15 turns No. 16 wire, 3/8-inch id., turns close but not tightly (see text).
L3 — 5.5 Mo. — 20 ah — each half 15 turns No. 16 (Johnson 100F1C0).
L4, L5 — 1-turn, No. 16 wire, 3/8-inch id., turns close but not tightly (see text).
L6 — 3.5 Mo. — 25 ah — each half 15 turns No. 16 (Johnson 100F1C0).
L7 — 7 Mo. — 15 ah — each half 8 turns No. 8 (Johnson 100F1C30).
L8 — 14 Mo. — 5 ah — each half 5 turns 5/16-inch tubing (Johnson 100F1C36).
L9 — 27 Mo. — 3 ah — each half 6 turns 5/16-inch tubing (Johnson 100F1C36).
L10 — 30 Mo. — 5 ah — each half 6 turns 5/16-inch tubing (Johnson 100F1C36).
L11, L12 — 300 turns 22-gage, 3/16 inches id., 1-1/4 inch spacing at center.
L13 — 10 turns No. 16 wire, 1/8-inch id., approximately 1/4-inch spacing at center.

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front of the tank coil, and their sockets are submounted to bring the internal shields level with the chassis. Both the tank-condenser lid and the variable-link stubs are driven from the panel controls by means of right-angle gear drives.

The multiband tuner, fitted with National type N-4 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 wired at the bottom and the left in the bottom-view photograph. It was found necessary to enclose these two components with a partition shield, to improve the bad 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 socket up.

Power terminals and a sauce resistor for r.f. input are along the rear edge of the chassis.

Since the multiband tuner precedes below the chassis limits, a cutout was made in the bottom plate, and a 5 x 10 x 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 similarity but, of course, is desirable in the consideration of TVI. The unequal shape, evident from one of the photographs, is a result of the discovery that wiring to meters mounted along the upper portion of the panel, to which the amplifier was eventually attached, was causing anode losses in the r.f. circuit. The indentation at the top of the enclosure still allows the original meter mounting but isolates the meters and wiring 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 peak of the tubes while they are operating.

Originally, capacitance neutralization was used,

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as shown in the diagram. The neutralizing condenser consisted of a pair of ceramic feed-through mountings, 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 multi-band tuner, and 75-turn ribbon was 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-turn 1-Lamp 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 couplings 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. VFT 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 modes of operation.

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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 fetter has good regulation, by using VR tubes to stabilize the voltage dropped. The plate supply should also have good regulation as possible. Use choke input, the lowest value of bleeder resistance the supply will tolerate, and as much output regulation as possible. The proper ruling current for this amplifier is 50 to 100 ma. This ruling 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 e.m. operation with 2000 volts on the plate, 700 volts on the screen, and a biasing voltage of 88 the grid current should be hardly flow 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 800 ma at peaks. The amplifier has been operated with excellent linearity with the following range of applied voltages:

<table>
<thead>
<tr>
<th>Plate</th>
<th>Screen</th>
<th>Bias</th>
<th>Plate</th>
<th>Grid</th>
</tr>
</thead>
<tbody>
<tr>
<td>2000</td>
<td>700</td>
<td>88</td>
<td>1000</td>
<td>50</td>
</tr>
<tr>
<td>1500</td>
<td>600</td>
<td>110</td>
<td>800</td>
<td>40</td>
</tr>
</tbody>
</table>

At 2000 and 1500 volts, it is necessary to operate in the region of grid current, and some screening of the grid circuit is desirable. In this connection, it should be recognized that the 12A0 grid has some screening built into it. Since the amplifier as operated by the voltage just barely breaks the grid-current region, the screening is included in the amplifier.

The amplifier is completely stable, free from parasitics, and has excellent linearity. Operating on 75, 40 and 20 there is absolutely no TVI on Channels 4 and 10 in a fringe area, and only very weak RCI has been noticed. It should be ganged, however, that presumably because of the high Q of the coil this toner on 20 only sufficient power is available using the 12A0 to drive the amplifier to 1000 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

(1) The operator of an amateur station shall transmit the call sign of the station or segment of the station, or a generally accepted identification of the network being reached, or a combination, which is readily recognized and identifiable separately or as a part of any other purpose of a transmission, by means of the medium by which the transmission is made, and shall identify separately or jointly, in the audio channel, or simultaneously in the visual channel, or both, at the beginning and end of each single transmission, and during the transmission of any series of transmissions, or when the duration of the series is less than three minutes, except when the duration of the single transmission is less than three minutes.

(2) At least once every four minutes, or as soon thereafter as possible during a series of transmissions between stations having established communications, the call sign of the station shall be transmitted by the words "this is" or "this is ".

(3) When using teleprinter, phonetic aids to identify the call sign of the station may be employed.
GROUND-GRID LINEAR

with about 1/2-inch spacing between windings and core. Originally, with a 0.5-ohm secondary, it was wound for 80 volts to keep the 880s happy.

The resistor R2 has no particular significance except to limit the d.c. voltage appearing at the end of the cable of the link in disconnected. The 12-volt rail, E1, is part of a conventional BC600 driver. Peak plate input to the driver was around 75 volts, filling current in the 880s is around 80 ma, limiting to 350 to 400 ma on peaks. The linear worked right off, after a minor modification that was killed by the parasitic suppressors in the plate leads, and no neutralizing is required, of course.

Bill also passed along the sketch in Fig. 10, which is the way WIPIX uses his grounded-grid 805-813 linear final. The method takes a higher voltage MAGNET transformer, to make up for the drop in the link and the e.i. choke.

Push-Pull 304-TLs

Ed Brown, W2DGO, has a pair of 304-TLs in his 11-MHz output amplifier driven by push-pull 414-A. The 304s are in the grounded-grid circuit shown in Fig. 2, and Brown says it is about the most breaded amplifier he ever tried. The 260 volts to his is obtained from a Y7-150 and a Y7-50 connected in series, and this holds the filling 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.

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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.a.h. signal is introduced at the "VFO" socket, and the Viking is switched to "C.W." and "VFO." The phasing can be varied by injecting audio from the s.a.h. exciter. W4JMU then tunes into the mixer 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 give the indicated plate current up to about 220 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 W8CPI (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.a.h.'s! What it does mean is that W4JMU found that the bass and frequency response 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 AB2 under the above conditions.

CASCADE DRIVER STAGE

For a number of years the "cascade" circuit has been the exclusive property of the v.h.f. men and the TV set manufacturers, but such is no longer the case.

The cascade circuit is used as a driver by WA8WS for his 887 output stage. A C607 887 is used for power triodes, and for circuit balance.

S.a.h. exponent Art Hulse, WA8WS, finds that it makes an excellent replacement for a 6AG7 amplifier, if you have had trouble taming one of these fussy pentodes. As used at WA8WS, the 6857 cascade follows a 6BQ7 converter that gives the signal from a modified 887 Jr. exciter and a 9-Mc. VFO. The output from the cascade is line-coupled to an 857 amplifier that is a little unusual in that it uses an extra 857 for neutraliz-

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High-Level Converters

T
nually, frequently in need for a good high-level converter to aid in bandchanging at power shows what the small receiving-type mixers will handle, and Norman Bailey, WeEdd, 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 used. WEEED has used a 6UC, 6U4, 6L6 and 6AC7 in the circuit, all with equally good results. As an amplifier, the plate current is around 15 ma. (at 300 volts), with no variation with signal. As a mixer, the plate current rises at around 15 ma. and climbs to 30 ma. on peaks. It requires less than 1 watt of input, and its output is sufficient to overload an 807.

The heterodyning signal has no effect on straight-through operation; WeEdd uses the

\[ V_i = \frac{1}{2} V_o \]

Fig. 1 — This simple high-level mixer/amplifier has been used by WeEdd and others. The heterodyning signal is present at the grid in either amplifier or mixer circuit — the tuning of the output circuit, Lk, determines the frequency. C1 = 100 µfd.

See text for suitable tube types.

Devion straight-through on 75, and a 10-Mc., oscillation and a new end at 42 put the output on 20 meters. Suitable tubes are the same already mentioned include the 807, 2128 and 6A6.

75- and 40-Meter S.B.B. Operation

To amplify two-tube operation with the 45-kilo, crystal-filter oscillator, Ralph Poonen, WE2FL, uses the double-conversion system shown in Fig. 2. A block diagram is shown at A, illustrating the dual operation of the NGO. The output of the high-level mixer is tuned to the desired band.

The circuit of the high-level mixer is shown as

\[ V_o = V_i \]

B, and is self-explanatory. It is essentially the same as the WEEED circuit described earlier. With a system like this, if the lower sideband is obtained on 75 meters it will also be available on 40 meters.

Fig. 2 — While an amplified bandchanging between 75 and 40 with the crystal-filter NGO is a different project, the same NGO can be used to tune the circuit from 75 to 40 with the same C2, L1 and L2 as 5- or 4-meter bands.

BO-457 as a Mixer

The following description of a BO-457 conversion to a mixer is due to D. Miller Sodahl, W4YVP. 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 heterodyne frequencies are mixed to obtain output in 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 experimenter has no provision for switching sidebands, as is sometimes the case in the crystal-filter jobs.

As shown in Fig. 3, the output of a 12AV6 crystal oscillator on 3.3 Mc. is fed to the center.

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REGULATED BIAS SUPPLY

One of the requisites of any linear Class AB or Class B amplifier (except those using zero-bias tubes like the 845-A) is a "stiff" bias source. Although batteries are used in many transmitters, they are really at their best only when the peak grid current is relatively low. Field Manuals, 20x20FLY, worked out a variation of an earlier regulator circuit that he used to give a constant 40 volts of bias, over a grid current range of 0.5 to 80 ma. As shown, the circuit has another advantage in that a 2000-ohm-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 rotating R1. C1 = 50,000-mfd, electrolytic. C2 = Power-supply output condensers. L1 = 0.1-ohm, 150-megohm resistor. R1 = 50,000 ohms. R2 = 15,000 ohms. R3 = 5,000 ohms. R4 = 6.2 ma bias.

THE RADIO AMATEUR
Cutting Down VFO Drift
RICHARD E. LONG, WJSAS

Some years ago the writer went through the
trials of trying to sell a.c. to a bunch who
would come up with almost any reason for
not getting into it. One night a W7 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 me. Knowing that a drift of more
than 50 cycles would throw the stuff into inverted
speech in the other fellow's receiver, I reckoned
that I would have to build something stable.

Mounting Components
The first attempt was the construction of a
several-tuned oscillator in a 3 x 4 y 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 paddles were all
mounted solidly, but I had fastened the paddle
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 bleat
in the signal. One night I tied the rig with the
rear cover of the box removed. The bleat disap-
peared! Put the cover back on, and the frequency
would jump continually. The cure was to mount
the paddle from two adjacent walls of the box,
as shown at B, thus relaxing the compression
and stretch in 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 4 to 8 in the evening, and the temperature
drift for about two hours, after which it set-
tled down. That was just doony! My signal
drilled all evening until I was ready to shut
down! A thermometer placed on the VFO box
showed that it actually required those two hours
from QST, August, 1932.

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change 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, and coil (the two parts of the coil) in a box and ran 500-foot lengths of RG-59-U over to the grid and cathode of the 6AX7 in the rig, as shown in Fig. 2A. The total length of this VFO was something around 200 feet, running entirely from switch and controls, to the set down in about ten minutes and stayed put. Then it was like this! Lucktivity. I used a 4-ply, magnetroid, capacitor (C1) in parallel with the tuning condenser.

Simplication?

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 me new ideas to tinker with. 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 two pieces of coax. One cable is normal and shorter 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 cloned thing would 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 circuit is tuned (tank circuit and apparently has a very detrimental effect on the Q). Worse yet, when I put the rig back up on the air, the gang began again to ask me to get back on frequency. Brother, I'm backing off here! It's not seeminly possible that losses, in feet of coax at 3000 kc, could be sufficient to account for enough heat to account for the drift. My guess is that the increased tank current through the coil, as a result of the lower Q/C ratio, after the coil was pulled to make it shorter, was responsible. Anyway, set that shunt for the single coax line so far as W1ASN is concerned.

Improving the Q

This experience led me to consider ways in which I might improve the Q so that the Q/C ratio could be made as great as possible, still maintaining oscillation. I had noticed at one time during the experiment in which the oscillator behaved a lot differently when one side of the box was removed. The two 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. Chassey Aviation, W2EZD, does a 4 × 5 × 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 001-dvalue attenuators to the back of the box with the tuned circuit, as it was in the first model, I re-
A 5-Mc. Remote-Tuned VFO
GEORGE GRAMMER, W1DF

The remote-tuned VFO circuit shown in Fig. 1 uses the popular series-tuned oscillating 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 outputs in the broadcast band.

The remote-tuned circuit for a variable-frequency oscillator can be used with great advantage in a VFO circuit, if the various parts of the circuit are selected at a fixed frequency in the vicinity of 9 Mc. for 5-5.5 Mc. range permits covering to either the 75 or 20-meter phone bands, taping being carried out by setting the zero beat with either antenna and followed by setting the other antenna to generate a desired frequency.
tuning condenser. The actual band covered is determined by the stroke 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 photograph was adapted as a means of stiffening the assembly and avoiding mechanical strain. 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 in permit exact fitting. C₁ and C₁₁ 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 isolated from the box by their vinyl coverings where they leave it at the rear. C₁₁ is mounted with its shaft downward and is screwdriver adjusted through a hole in the bottom of the box.

The tuning condenser, C₁₁, 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 supposing number of times, and elongating the condenser mounting holes vertically. The slot for gear alignment is both "on center." (It is assumed, of course, that the condenser shaft is sufficiently free in the coupling to permit independent rotation. If R₁ is not, use steel wool on the shaft to give it free, but not loose, motion.)

The tank coil, a length of B & W No. 30-36 Inductos, is connected to a strip of polystyrene mounted on a 1-inch stand-off insulator. By recounting 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 small silver tuning ring that provided by standard vernier dials is highly desirable in actual operation, as the mechanism shown in the photographs was put together. It combines a National type ACN dial with a Type K and a National knife. Since the ACN has a 3.5-inch shaft while the K has a 3-inch bore, it is necessary to put a 1-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 annealing some solder in the end. This combination has a mechanical step-down ratio of about 25 to 1, giving a "light" feel and simple adjustment range for frequency changes of a few cycles.

With the construction shown, 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.

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Existing Exciters

The remote tuned circuit as described above can be directly modulated for the crystal in a crystal-controlled oscillator that is designed to operate in the vicinity of 3 Mc. This has been used successfully in this way with the Central Electronics 10A output, and although this setup arrangement provides no buffering to isolate the oscillator, tests have shown that frequency-modulation effects are negligible at levels up to which the output exciting 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.

Whenever the excitation 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 t.s.o.o. on, the receiver being tuned directly to the oscillator frequency and the signal picked up so that there is no overlapping in the receiver. Single-tone input is best. The power-input 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 connections between the cables and the unit. The ground point should be as close as possible to the crystal socket into which the 300-ohm line plug fits. The best place to ground is at the screw holding the crystal socket, with the near minimum of lead length.

Companion Oscillator Unit

Where there is any doubt about sufficient isolation to eliminate I.M. effects, or if an exciter is under construction, provisions 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 108H, but could logically be made an integral part of a home-built exciter. As shown installed on a 108H, the filament and plate voltages are obtained from the circuit 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 i.c. version of the sets in-tuned oscillator while Fig. 2 shows a

![Fig. 2 - Oscillator/buffer tube circuit.](image)

Capacitance values below 0.099 μf, are in μf, 0.001 and above in μf. All condensers are ceramic type unless indicated otherwise. Resistors are ½-watt composition except where stated.

- 1 - Plate choke (30 or 25 μf, does satisfactorily).
- 2 - Adjustable to 20 μf, 1 inch spacing (5 No. 26 enameled, 3 x ½-inch plug-type plugs from National W.H.).
- 3 - For input No. 24 t.f.c. wound over ground end of L.
- 4 - 3-conductor microphone connector (Armstrong 80-7K27, mating plug for 3-conductor 101-PF Mr. E., R.F.C. 900-μh choke (Millen 33490-560).
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.h.t. plate circuit, particularly when a tuned output circuit is used. The only unusual component values is the r.f. condenser C9 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, L5C5, is used to couple a small coil, L6, to the No. 1 grid of the 6SK7 in the exciter through the socket connection on the rear chassis wall. The number of turns on L5 was adjusted to deliver maximum voltage to the 6SK7. No. 1 grid through a short 7-inch length of ordinary l.c. 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 X 4 X 4 aluminum box. The shelf is mounted on one of the removable sides, and all the wiring except the connections to J1 is done on just those two pieces before the rest of the box is attached. The box portion is neatly made through diagonal corners in short, and J1 is mounted on one of the L-shaped pieces. To insure a good ground connection between J1 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 cables to pass through, and is attached after the wiring is done and the unit has been tested. The second copper 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 in the metal back plate.

Note that the 6SK7 is fitted with a shield. This was in the hope that some control of the high-frequency radiation could be obtained by the use of a shield. The coil is wound to make the shield electrically short-circuited, and it was found that the shield gave some improvement in the high-frequency radiation. However, it is difficult to determine whether this improvement is due to the shield or to the use of a larger number of turns in the coil. In any case, the shield is easy to construct and does not require any special tools or materials.

**THE RADIO AMATEUR**

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 were drilled in the two and three edges of the box, over and under the l.c. board, help keep things cool and thereby maintain frequency drift.

To cover the necessary frequency range - 350 kc - when the e.h.t. generator is 9.5 Mo. - without adjustment a fairly tuning-low-C circuit would be desirable at L5C5. However, a band width of about all that can be 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 L5C5 set at the center of the range, the output, as measured by the rectified voltage at the No. 1 grid of the 6SK7 (using a r.f. voltmeter) drops off at about one-half at each 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 of little critical, since the normal exciter output power can be obtained with identical waveform with no little as 5 or 6 volts.
Simplified Voice Control with a Loudspeaker

WALTER N. HUNTER, W9IB

T 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 the transformer and a tube from the original design. Hank Turkel uses the Novak circuit at W5EML 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 6166 rectifier rectifies the audio signals in the two channels. When a signal comes only from the speech amplifier, the volume at the grid of the 667C goes negative, cutting off the tube and causing the relay to “Fall out.” Signal coming from the receiver tends to bias the 667C grid positive — signals coming through both channels can be made to have light or no effect by adjustment of R4 and R5. 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 altered by changing the value of C1. 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 expectors R4 and R5 tend to shape the lower-channel response in a like manner.

We have used the only control on the front panel in R8, which is used to balance for different operators. His experience has been that, over adjusted, R8 and R9 will need no further attention, and I have also found this to be the case.

To put the unit in operation, first adjust R4 to the point where the relay is held closed with a positive action. While speaking into the microphone, R5 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 R8 to just eliminate this effect.

One word of caution: If your receiver gives a loud “pop” when the relay operates, it will first through the 6SN7 receiver amplifier and may pull the relay off again. The answer to this is use a relay allowing system that will operate without the click. It was done at W9IBM by fitting the output ground connection on the receiver and running it to ground through a set of relay points.

From QST, October, 1945.


SINGLE SIDEBAND FOR
A speech amplifier and operating accessories for S.S.B.

DANA A. GRIFFIN, W2AOE, AND DONALD H. FYRKLUND, W2HLP

A fixed-frequency oscillator is sufficient for tone-up and checking of an a.m. transmitter, and provision is made for such an oscillator in the speech amplifier-voice control circuits shown in Fig. 1. Using an R.C. network that can be switched in and out as desired, the first speech-amplifier stage can be converted into a 1500-cycle audio oscillator.

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 a.m. operators. By means of $R_0$, the voice-control rectifier and amplifier can be used as a v.t. voltmeter for checking both s.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 $S_5$, in the extreme left-hand position.

Fig. 1 — Speech amplifier and operating accessories of the a.m. transmitter shown in the photograph. Capacitor values are in $\mu$F, and resistors are $\frac{1}{2}$ watt, unless otherwise indicated.

- CR — Selenium rectifier.
- $K_1$ — Diode relay (one set of contacts used), 10,000-

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Voice-Controlled Break-In with a Loudspeaker and No Relays

SINGLE-SIDEBAND enthusiasts Ray Berdell, W6JJ, saw the circuit shown in Fig. 1 to give him smooth break-ins 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 485 volts blocking bias for the receiver. These biases exist at either full volume 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 6A6S 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 6AGS just below cut-off. Talking into the microphone on the microphone side of this circuit, no output is seen 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 hold-on until the rectified output of the audio from the speech amplifier decays below the threshold value.

W6JJ adds that if the receiver is to be disabled by applying the -90 volt to the a.v.c. bus, it is recommended that he 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 a.b. reception. The time constants of this diode circuit can be made variable, but the values shown have proven quite satisfactory at W6JJ. Miss, condensors are to be preferred, to stabilize the time constants.

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Fig. 1—This voice-controlled circuit requires no relays and permits automatic break-in. These and other circuits are described in a multi-asooted output developed with the low-impedance, winding substituted for the speaker voice coil.
under various conditions of temp.-nature and humidity.

Lawrence Smith. W7TOJ, has been using the W7G2 switch-control circuit and is pleased with its operation. He made some modifications and had some experiences that we pass along for others who might be interested in the circuit. In his words, "Any short-cut-off pentode r.f. tube can be substituted for the 6AK5 used in the circuit. I ended up with one 6AK5 and a 900I because that's what the job box yielded. However, 6AK5s were tried in both sockets with good operation. Improved testing action may be obtained by altering the values of the common cathode resistor and/or the plate resistor of the right-hand 6AK5. 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 6N27 mixer and 6AX7 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. bias. These three resistors should be changed to the values that gives 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. bias, pick one with a very high back resistance. I had some trouble with a 1N34 diode that is the back resistance decreased after the receiver operated for a period of time. This upset the

"SCOPE INTENSIFIER"

If you don't read the RSGB Bulletin and thus get a chance to see G2ECU's excellent column, "10 Points-Sideways," you will have missed the next trick shown here. It's a dodge for reducing the intensity of the scope trace when there is no signal, and thus reduces the chance of "burning" a line on the screen. It's easy with an a.m. rig, of course, where you're cleaning on and off with all sorts of power supplies, and turning off the scope can be made part of that procedure, but with a.c. 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, all 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's, of course, a simple thing to tie in your 'own switchless'-type oscilloscope circuit, but G2ECU's system becomes self-powered and automatically adjusts itself to the proper level.

"THE RADIO AMATEUR"

G2ECU reports that, in practice, the sensor output inch of vertical deflection is invariable, while the range of deflection is variable but not bright enough to cause a burn.
Break-In with One Antenna

M. E. HINDE, EX-2KSO

When using the transmitting antenna for receiving, the usual procedure is to employ an antenna change-over relay, but this precludes gradual break-in.

In radio, the "TR box," or "duplex," was the solution to this problem. Essentially, it gave the effect of the circuit in Fig. 1. Whenever the transmitter was "on," the quarter-wave

transmitter (the "off") transmitter looked like a high impedance, all of the energy coming down the feed line would go into the receiver. In radio work, the short circuit was obtained by either an open spark gap or use in a "TR tube.

The TR tube system won't work on the amateur bands since 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 radio rigs, it is possible to use a relay to short circuit the line.

"Yeah, but the contacts will burn up or are over or something," you say. Well, let's see.

If your transmission line is matched to the antenna, the voltage on the line is

\[ V = \frac{P}{\sqrt{X}} \]

where \( P \) is the power output of the transmitter and \( X \) is the line impedance. For any standing-wave ratio on the line,

\[ E_{\text{max}} = \sqrt{\frac{2PS}{\pi X^2}} \]

where \( s \) is the voltage S.W.R. To take an extreme case, consider 1 kw. into a 500-ohm line with a 2:1 S.W.R., \( E_{\text{max}} \) works out to be 2000 volts. Hence the voltage across the line at point X might run this high. The current through the

short will be

\[ I = \frac{E_{\text{max}}}{X} \]

where \( X \) is the impedance of the stub line. Assume

\[ X = 500 \text{ ohms} \]

\[ E_{\text{max}} = 2000 \text{ volts} \]

\[ I = \frac{2000}{500} = 4 \text{ amperes} \]

\[ V = \frac{1000 \times 4}{\sqrt{500}} = 200 \text{ volts} \]

Fig. 2 — A practical amateur application of the "TR" system. The short circuit is obtained with relay \( R_{\text{on}} \) and the job is made quite simple by the step-up in impedance. To C wattmeter, in this application, \( C_1 \) is a 0.1 ohm, 1000 amp relay resistor coil combination. The transmitter coil is tuned on and off at will. The relay \( R_{\text{on}} \) a 5000-ohm potentiometer, establishes the receiver gain when the transmitter is on.

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SINGLE SIDE BAND FOR
Send-Receive with One Antenna

During the course of some scatter-sounding experiments at Stanford University, the author (and also Radio Amateurs) in charge of the project devised a 3-way for using the same antenna for transmitting and receiving that is also applicable to 3x. As shown in Fig. 1, the

final amplifier of the transmitter is link-coupled to the receiver through a preamplifier using diodes and neutralized troughs. The diodes prevent 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

tank circuit becomes small and negligible power transfer occurs. The high Q's relative to the triode grid further protect the input tube 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.

Since the shunting capacitances are small, the bias built up discharge in a very short time inversely 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 B. Reedy, W6AAS, has tried the system in his receiver without adding a neutralized triode preamplifier, and he has found that it gives good results with a

THE RADIO AMATEUR

J73
Vacuum Tube Coupler

Another solution to the problem is that devised by Tom Pachuck, W2EKM. The basic principle is to use a low-pass coupling tube between antenna line and receiver, and to pass 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 6AJ1 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 6AJ1 acts as a normal grounded-grid amplifier for the receiver, but during transmit periods it is biased high enough to prevent any appreciable signal leaking the receiver. Additional circuitry can be added that will simultaneously reduce the gain of the receiver.

The physical construction of the 6AJ1 stage should be such that the plate circuit, which forms the receiver input, is completely shielded, so that the only coupling is through the 6AJ1 grid. A metal shield should enclose the 6AJ1 plate terminal and the receiver connection, which should preferably be coaxial. The shield can connect to the 6AJ1 grid directly or to the grounded grid terminals. The lead from the 6AJ1 plate resistor should be by-passed where it leaves the shielded enclosure.

A shielded 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 hum and noise, giving a low-diode voltage at the control unit of more than the 200 volts r.m.s. shot and is the maximum allowable for proper operation. This 200 volts, by the way, corresponds to 500 watts in a matched 50-ohm line, and 570 watts in a 75-ohm line.

It is necessary to feed the 6AJ1 filament in the manner shown in Fig. 2 to keep within the heater voltage ratings. With the National B-3008 chosen, about 35 volts of heater supply voltage was needed because of the drop in the resistance of the filaments.

Automatic Antenna Switching

Although relays can be used for quick switching of antenna from receiver to transmitter when working over a clear break-in, it is much easier to do it automatically. Two circuits used for this purpose by Bill Root, W2ZUM, are shown in Fig. 3. The circuit at A is along the lines that were described by W2OQA (QST, June, 1952). The system at B is presently in use at W2ZUM. The circuit of A is simple 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 24-watt neon will suffice for 75 watts or so. The pilot lamp is a safety factor 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. 2**—In this receiver protection circuit, a 6AJ1 coupling tube is biased back when the transmitter is in use and thus minimizes the power reaching the receiver.

**Fig. 3**—Two TR ("transmit-receive") switch circuits that have been used by W2ZUM. The circuit at A uses two large inductances, L1 and L2, a small condenser, C1, and a scope bulb. The circuit at B uses only one coil and adds a small filament bulb for added protection of the receiver.

C1 = 25-μfad, cap. 600v.
L1 = 50 turns No. 23 on 1 1/2-inch form (3.5 μH).
L2 = 60 turns No. 22, 1 1/2-inch form.
B = 6-61 lamps 150 ma.

The scope bulb can be 1/4 watt with a low-powered r.f. and 2 or 3 watts with a high-powered transmitter.

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For Use With B&W Model 5100 Transmitter

Now, for the first time, you can get really sparkling performance on your SSB, AM phone, or CW. This B&W Single Sideband Generator is set up with the famous Model 5100 Transmitter giving you outstanding SSB operation on all frequency-pair transmitters. Tuning and operation are a breeze. No test equipment is required. Single sideband signals are generated by a single and efficient method perfected after two years of intensive research and testing by B&W engineers. No extra has been left untried to give you such ease in wire repair and push-to-talk control, a speaker desensitizing circuit, TVI suppression, and all internal construction for quick and easy removal of any major section. Completely self-contained, the 51SB requires no more external accessories than a microphone.

Inc. 277 Fairfield Avenue
Upper Darby, Pa.

Combine this Single Sideband Generator with the features of your Model 5100—150 watts peak envelope power input (50 watts peak envelope power output) on SSB; 150 watts on CW; 100 watts on AM phone; VFO or crystal operation; push-to-talk control—and you've got a combination that will turn the head of every amateur.

Write for details.

These are just a few of the hundreds of products super-
curious designed and built by B&W to meet the needs of
the amateur. Others are described in Catalog 2FC
available upon request. Write for your copy.
Q: Why go to Single Sideband?
A: Here are 2 good reasons—the big advantages that have made SSB one of the most talked-about developments in years...
1. With Single Sideband, your back-end rig can provide up to 9 dB gain, or a power increase of 3 times, over conventional double sideband AM.
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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.
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7. Distortion due to selective fading eliminated.
8. Harmonic TVI virtually eliminated through the use of linear amplifiers.

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 going the pro and one 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 items. 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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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.

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Shown above are a few of the E-V microphones designed for effective communication. Shown, lower left, Model 228 with aural and Model 911 crystal. On-off switch. List Price, Model 228, $39.50; Model 911, $69.00. Shown lower right, Model 229 with variable range, with aural and crystal response. On-off switch. List Price, Model 229, $75.00. Shown upper left, Model 125 with range, high quality dynamics, with variable frequency response. On-off switch. List Price, Model 125, $97.00. Shown upper right, Model 126 "Steno" with range, dynamic head. Adjustable frequency. On-off switch. List Price, Model 126, $99.00. Shown lower right, Model 526 with range, high quality dynamics, with controls for variable frequency response. On-off switch. List Price, Model 526, $125.00.

Other E-V microphones for voice and aircraft communications, broadcasting, public address, recording, and phone applications.

For further information, see your E-V Distributor or write for Condensed Catalog No. 116.

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Performance figures achieved with production model VPA 20-3, 3 element beam, in typical house-top installation. Popular commercially built transmitters, receivers and test equipment was used.

Mosley Electronics Inc.
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Page 11
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