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Single Sideband Techniques

Jack N. Brown
W3SHY

Theory

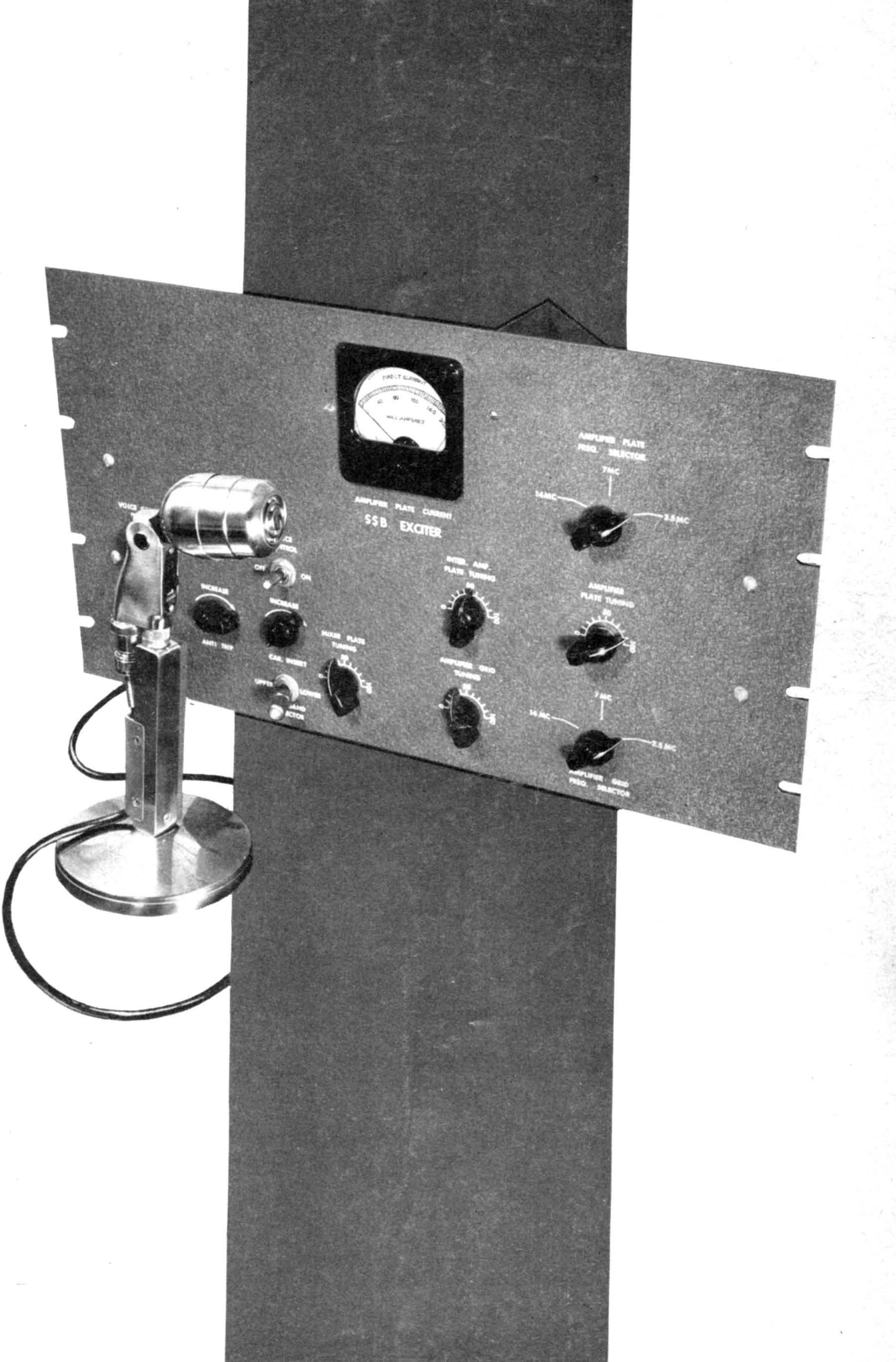
Operation

Construction

CQ

Technical Series

Edited by Oliver P. Ferrell


DIRECT CURRENT
MILLIAMPERES
AMPLIFIER PLATE CURRENT
SSB EXCITER
AMPLIFIER PLATE
FREQ. SELECTOR
14MC
7MC
3.5 MC
INTER. AMP.
PLATE TUNING
AMPLIFIER
PLATE TUNING
INCREASE
ANTI TRIP
INCREASE
CAR. INSERT
OFF
ON
UPPER
LOWER
MIXER PLATE
TUNING
AMPLIFIER GRID
TUNING
14 MC
7 MC
3.5 MC
AMPLIFIER GRID
FREQ. SELECTOR

SINGLE SIDEBAND TECHNIQUES

by

Jack N. Brown, W3SHY

Contributing Editor, CQ

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Contents

FOREWORD

I once had a professor tell me that the easiest way to become an "expert" was to pick a more or less remote subject, read all the material that was available on the matter, and then publish a series of articles or write a book restating the many points that had been covered in the reading material. You therefore automatically become an "expert." I must confess to falling in that same general class of writer. Rather than become an "expert," I prefer to consider myself an out-spoken proponent of single sideband. I have advanced no earth-shaking ideas or inventions, but if I have contributed anything, I hope it is because I have presented the technical considerations in words of one syllable so that they can be more easily understood.

During the period of time that it took me to build my first single sideband exciter, I formulated plans to conduct comprehensive comparative tests of AM versus SSB. Then came the first night when I worked W3ASW with a measured peak output power of two watts. From that date on the filaments of the AM modulators were never lit again. Forgetful? Unscientific? No. The difference was so marked that I didn't need any tests to persuade me how much SSB was superior to AM-double-sideband transmission. The modulator was long since been given away and, I presume, passed on still again because the station to which I'd given the unit is now on single sideband also. After almost six years of operating on single sideband I never cease to be amazed at how well a mere few watts of SSB always seem to be heard. The number of Hams that are on SSB with only a "barefooted" exciter testifies to that fact.

In putting together, under one cover, information on single sideband, I have attempted to give enough information to bring a complete "outsider" up to date on the *techniques*. The reader will find that the material in the first five chapters is an edited and rewritten version of the six-part series, "Getting Started on Single Sideband," that was published in *CQ Magazine* during 1953. The comments from readers was so gratifying and requests for reprints of the material were so numerous that it was felt that this old material would make a good fundamental primer for this handbook. Upon this material is built the other two-thirds of the book.

The author is personally indebted to *Harold Carr, W3JFI,* and *J. Edward Wiley, W3VVV* for the assistance in building the transmitter that is described. Their workmanship and attention to detail relieved me of a very great burden. The following companies generously contributed many of the parts for the SSB transmitter in *Chapter VIII:—Automatic Electric Co., Barker & Williamson Co., Bud Radio, Inc., Collins Radio Co., Chicago Standard Transformer Corp., Electro-voice, Hammarlund Mfg. Co., James Knights Co., James Millen Co., J. W. Miller Co., P. R. Mallory Co., The National Co., The Radio Corporation of America* (Victor Div.), and the *Simpson Electric Co.*

To *Perry Ferrell,* the editor of *CQ* Magazine, has fallen the tremendous job of coordinating and assembling the material for this book. My back still stings from the driving whip lashes of "o.p.f.," but to him goes a special word of thanks. Also a few words of appreciation to *Tom Smith, W2MCJ* who supervised all of the wonderful drafting and to *Ed Kephart, W2SPV* for the photographs in Chapter VIII, IX and X. To the ever-loving XYL, Caroline, who labored long and late over the typewriter, this humble effort is dedicated.

Jack N. Brown, W3SHY

Chapter I

What is SSB?

What makes single-sideband transmission appeal to ever-increasing numbers of amateurs? Is it the hope of gleaning 6 to 9 additional decibels from their equipment, or is it the voice-break-in type of operation that has become attractive after being forced to listen to 10-minute transmissions in old-fashioned AM roundtables? Or, perhaps it is just a matter of "keeping up with the Jones's."

Whatever the incentive, the fact is obvious that increasing numbers of amateurs are changing to single sideband transmission. The usual reaction is "I haven't had such a big thrill since I had my first QSO."

The ability of a few watts of single-sideband signal to be heard even under the most trying band conditions is not to be taken as over-enthusiastic propanganda, but an uncanny truth that has yet to be fully resolved in the minds of many. The apparent superiority of the SSB signal cannot be *fully* explained in terms of the bandwidth occupied, the relative power involved or the receiver used. The "stuff" *just gets through.* The presence on the air of many higher powered (500 to 1000 watts) SSB stations is an indication, *not* that SSB is for those with money to spend, but that a one-kilowatt SSB transmitter is more economical to construct than an AM-kilowatt transmitter.

The "sideband concept" of 'phone transmission may be new to many readers who like the author were brought up to understand that modulation was the process whereby the amplitude of the carrier was "varied" in accordance with the impressed signal. Many amateurs when exposed to single sideband techniques are puzzled as to "what's left" when the carrier is suppressed and the microphone is spoken into. They should not feel self-conscious about these questions for it is a matter of record that one of the early pioneers in the field of communication derived the mathematical expression for the modulation process and then promptly dismissed the two sideband terms as being "imaginary roots" that could not possibly exist! He was taken to task by others in the field, although actually the fued went on for several months in the scientific journals of the day. If one stops to consider, human speech is actually single sideband transmission of the simplest form. It is upper sideband in nature with a carrier of zero frequency.

To better understand SSB we must start with something with which we are familiar and draw our analogies accordingly. Let's consider the ordinary double-sideband with carrier phone signal sketched in *Fig. 1-1-A.* This signal is an idealized phone signal. Only the necessary speech frequencies are transmitted. For good intelligibility, experience has shown that only the speech frequencies between 300 cycles and 3000 cycles need be transmitted. By use of a low-pass filter in the speech equipment, the frequencies above 3000 cycles can be eliminated,

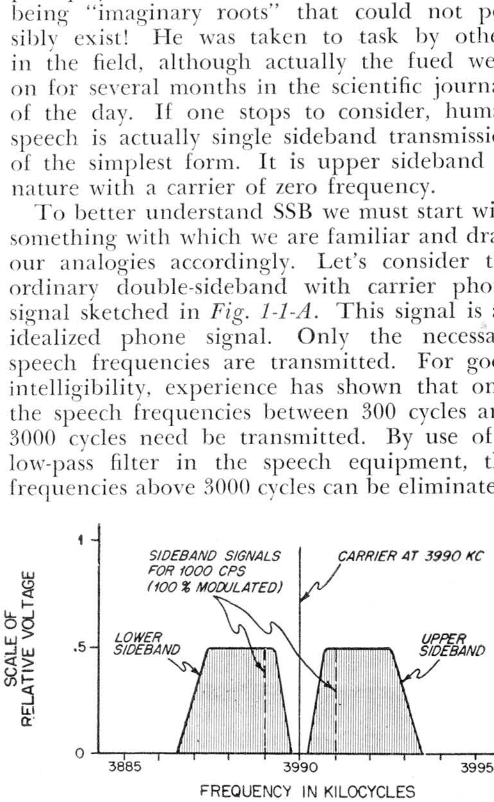


Fig. 1-1-A. Relative amplitude-versus-frequency representation of a double-sideband AM signal. Note the equal distribution of energy (hence intelligence) into two sidebands.

and by proper choice of coupling condensers the power-wasting low frequencies below 300 cycles can be attenuated. The horizontal axis of *Fig. 1-1-A* is a scale of frequency as indicated, and the vertical axis is a relative scale of voltage just for comparison purposes. The shaded portions represent the general area occupied by each sideband. Any speech frequencies transmitted will appear as voltages lying under the boundary of the shaded area (for example the 1000-cycle tone which appears at maximum 0.5 relative amplitude).

If there is more than one tone present in the modulation, as in speech, each frequency as represented by a sideband component under the shaded boundary cannot equal a maximum value of 0.5 relative voltage. Depending on the phase relationships between the speech frequencies modulating the transmitter, the peak voltage of the composite wave form could possibly be equal to several times the amplitude of any one of the individual frequencies. In more direct words the instantaneous sum of all the speech component voltages in one sideband cannot be greater than one-half of the carrier peak voltage. If the sum were to exceed the one-half limit, over-modulation would take place with all of its splatter and consequences.

At this point "actual proof" of the existence of sidebands might well be presented. The photographs of signals shown in this chapter were taken with the cooperation of the *Panoramic Radio Corporation* and *Barker & Williamson, Inc.* Pictures were taken of the oscilloscope presentation of a *Panoramic SB-8A Spectrum Analyzor, Model T-200.* This unit is what might be called a high resolution panadapter. As the cathode-ray beam moves across the face of the tube it indicates the amplitude of the signal present at a frequency which is continuously varying as the beam moves. The sweep rate used for these pictures was one sweep per second. The selectivity of the internal amplifier unit was set for 70 cycles. Thus it was possible to separate the carrier from its sidebands and to determine just what bandwidths were involved for various types of transmission. The picture captions explain in detail just what is being observed.

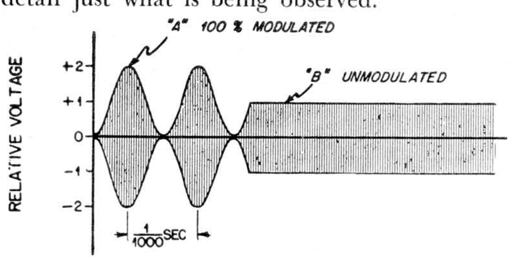


Fig. 1-1-B. This is the common r-f envelope pattern of a double-sideband AM signal with 100% modulation (1000 cycle tone) at the left. The envelope appearance in an unmodulated condition is shown at the right.

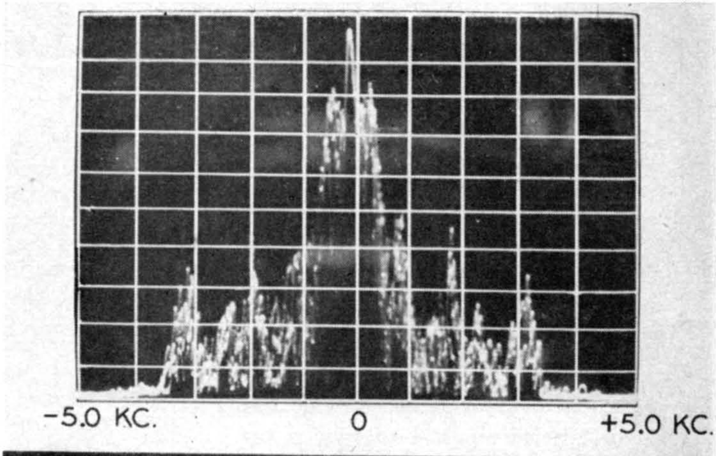
Figure 1-1-B shows the well-known 100% modulation oscilloscope pattern as compared to the unmodulated carrier condition. It can be seen that the instantaneous peak-to-peak voltage at "A" is twice that of the unmodulated carrier at "B." This means that across a dummy load the peak voltage will be twice that of the unmodulated carrier. Since $Power = E^2R$ where R is the dummy load resistance, the peak power at the instant of point "A" in *Fig. 1-1-B* will be *four* times that of the unmodulated condition. All this is leading up to the following point. The final amplifier must be capable of delivering peaks of power four times greater than that of the carrier. Look again at the relative power involved in the carrier and the two sidebands. Take for example a 100-watt phone transmitter. The carrier has a power of 100 watts while each sideband contains only ¼ of that power, or 25 watts. Since each sideband carries the same intelligence and the receiver upon detection combines the audio from each in-phase, the total *useful* sideband power is 50 watts. Can this be true? Unfortunately it is true that to deliver *50 watts* of intelligence carrying power our final amplifier must deliver *400* watts on peaks of modulation. What happens if we get rid of our carrier and generate only the sideband signals? The useful sideband power is still 50 watts and the maximum power required of the final amplifier is *also 50 watts.*

The Useless Carrier

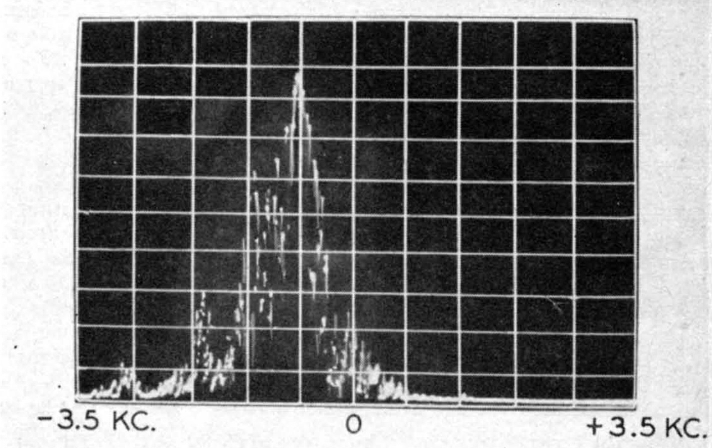
What function does the carrier serve? A very minor one actually: In the second detector of your superhet it mixes (heterodynes, if you prefer) with the two sideband signals to recover the original audio that was fed into the transmitter. Does it do anything else? Nothing else except hold up the needle of your *S*-meter. Then let's arrange the transmitter so that we do not transmit any carrier, but only the *two* sideband signals. Keeping the same peak sideband power (50 watts) take a look at your final tube. Cool as a cucumber, isn't it?

Now, back to the receiver. Ugh! It sounds pretty bad. Let us see if we can copy the double-sideband suppressed carrier (DSSC) signal by inserting a local carrier with the b.f.o. Unfortunately, it is practically impossible to do this. That villain carrier that we just got rid of *did* have one point in its favor. It was of the correct frequency and the correct *phase.* It is this *phase* condition that we cannot meet when we attempt to copy a *double-sideband* suppressed carrier signal by using a b.f.o. We can make a first approximation of the frequency, but never in this world will we be able to maintain the correct phase. We have failed in this attempt.

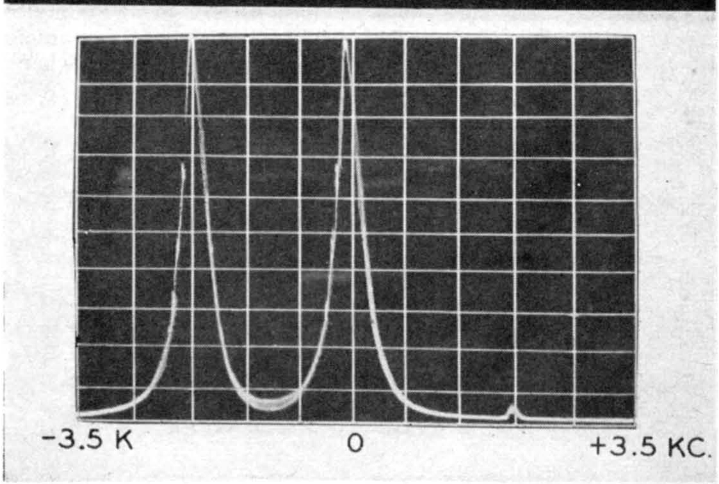
I would like to say at this point that it is possible to detect signals of this nature if a



This is a spectrum analyzor picture of a double-sideband AM signal. A vowel sound is spoken into the microphone. This transmitter was equipped with a 3500-cps low-pass filter in the speech amplifier. Note how the power is concentrated in the lower speech frequencies appearing close to the carrier.



In this spectrum analyzor picture a vowel sound is spoken into the SSB filter-type exciter described in this book. The suppressed carrier is at the center "o" point. Lower sideband is being generated.



A spectrum analysis of a "two-tone" linearity test on the filter-type SSB exciter. The signal at the center of the oscilloscope plot is an "inserted carrier." The other signal is a lower sideband resulting from a 2000-cycle tone. The small "pip" on the upper side of the carrier is 40 db. below the reference signal.

(The vertical scale on these spectrum photographs has been masked out since the grid was linear. The scale values would have been logarithmic with a range of about 40 db.)

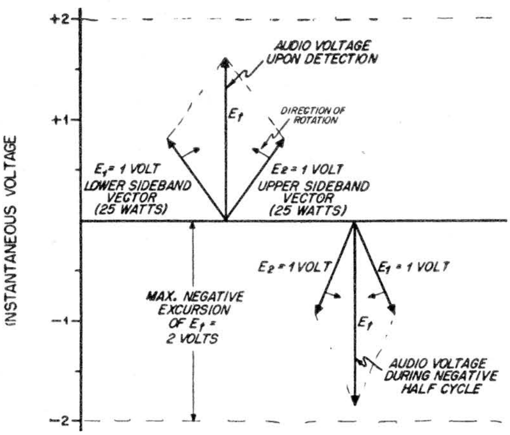


Fig. 1-1-C. Vector presentation showing two examples of how sideband voltages of a double-sideband signal combine in the detector of a receiver.

reduced carrier is transmitted. This carrier must be restored or "exalted" to a high enough level to demodulate the sidebands. For an explanation of DSRC radiotelephony (double sideband with reduced carrier) see Grammer.[20]

Getting back to our little experiment, what would happen if we just eliminated one of the sidebands? Better still we could run the 50 watts peak power in one sideband and still not increase the strain on the transmitter. At the receiver we *can* detect the single-sideband signal successfully by inserting a carrier at the correct frequency with the b.f.o. The phase relationship between the carrier and the sideband has no effect on the audio recovered. However, you will notice that the carrier supplied artificially by the b.f.o. must be very close to the suppressed carrier's frequency. How close? For distortion free copy about 50 cycles and for intelligible copy about 200 to 400 cycles depending on the signal.

Our double-sideband signal mentioned above with 25 watts maximum power in each sideband (total 50 watts maximum) will produce *twice* the *audio power* in our receiver than will the 50-watt, single-sideband signal. This comes about for the following reason. The double-sideband signal produces in the receiver detector circuit the two sideband voltages that can be represented by the two vectors *E1* and *E2* in *Fig. 1-1-C.* The two vectors are equal in amplitude (say for example 1 volt) but they rotate as shown in opposite directions at the rate of the modulating frequency. Example: A 1000-cycle tone modulated signal would have sideband vectors revolving at the rate of 1000 revolutions per second. These two voltages are added vectorially so that the resulting output voltage lies along the vertical line, and, in the case shown, the resultant voltage is *Et.* This *Et* will vary as the vectors whirl around from a maximum value of 2 volts in the positive (upward) direction through a zero value and downward to a maximum value of 2 volts in the negative direction. Thus we have a total change in audio voltage of 4 volts.

The Power Function in the Sidebands

Now let us consider what happens with the 50-watt maximum-power single-sideband signal. We said that we put all of our power in one sideband. To double our power in one sideband, we increase the sideband *voltage* by only a factor of 1.414 since Power $= E^2/R$. [Note: $(1.414^2 = 2)$] So in our detector we have only *one* sideband vector whirling around whose amplitude is 1.414 volts. See *Fig. 1-1-D.* Since it has no other sideband vector to combine with, it swings around from a maximum value of 1.414 volts positive to 1.414 volts negative. This yields a total peak-to-peak audio voltage of 2.828 volts upon detection. Since the double-sideband signal gave 4 peak-to-peak volts and the SSB signal gave 2.828 peak-to-peak volts, the result is 3 db. in favor of the double-sideband signal. In order for our SSB transmitter to produce the same 4 volts at the receiver, we must generate 100 *maximum* (peak, if you prefer) watts power in the one sideband. Norgaard pointed this out in his 1948 article.[26]

Getting back to the transmitter problem at hand, we notice that our maximum capabilities of the amplifier were 100 watts when using double-sideband-with-carrier. We can safely

"What I like about . . . SSB is that . . . it is so nice to tune in . . . an A3 signal afterwards."

(Courtesy **"Short Wave Magazine"** November, 1953)

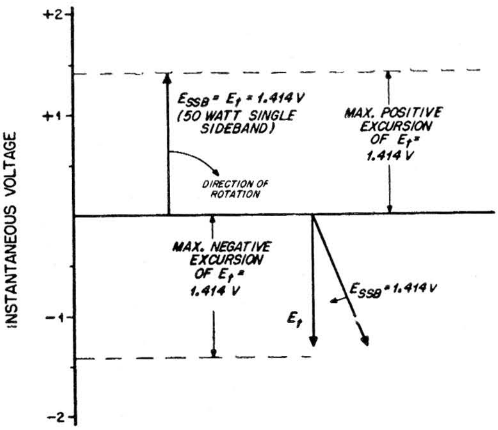


Fig. 1-1-D. This vector diagram illustrates how a 50-watt single-sideband signal yields 70% (3 db.) less audio voltage upon detection than a double-sideband signal with 25 watts in each sideband—see Figure 1-1-C.

increase our single-sideband maximum power to 400 watts which gives us a power gain of 4 over the double-sideband condition. This gain of 4 is an increase of 6 db. in system gain attained at the transmitter.

It must be pointed out here that in order to make our 100-watt AM final amplifier deliver 400 watts on peaks in single-sideband service we will have to double our plate voltage. You can see that in the AM case the modulator furnished the extra voltage upon peaks of the audio wave-form. It is not at all uncommon to find 1200 to 1400 volts being used on a pair of 807's used on SSB. The tubes still operate within rated plate dissipation, but the maximum value of peak power is in the neighborhood of a half-kilowatt!

1.2—The Receiving System

Since our signal is now concentrated in one sideband we can now concern ourselves with possible improvements in our receiving system. If it is possible to reduce the bandwidth of the receiver to match the reduced bandwidth of the transmitted signal, we can also gain another 3 db. in system gain. This comes about for the following reason: For any given bandwidth the signal-to-noise ratio is the limiting factor that determines how weak a signal can be copied successfully. Where receiver thermal noise (hiss noise) is the limiting factor it is also true that the noise power will be reduced by 3 db. if we cut our receiver bandwidth in half. There is considerable difference of opinion whether there is any signal-to-noise improvement in halving the bandwidth in the presence of *impulse-type* noise. The nuisance value of this type of noise is strictly a subjective matter and as far as this writer is concerned has yet to be resolved to everyone's satisfaction. So, rather than be accused of trying to claim 3 db. that might not actually be merited, we will conclude (for now at any rate) that we can gain 3 db. in the case of thermal noise and none in the case of impulse noise. How much of this total of 9 db. is actually realized in actual operation? It all depends.

For the conditions outlined, the first 6 db. are sure-fire available. Since under most amateur operating conditions, signals are not usually marginal—that is—competing with receiver thermal noise, the extra 3 db. are not usually noticed even though they are there. However, there is the more insidious matter of co-channel and adjacent-channel QRM from our brother Hams. This type of interference cannot be evaluated easily in terms of db. gain. Everyone knows that cutting down the receiver bandwidth will do wonders for this too. So you can safely conclude that we can gain 6 db. at the transmitter *plus* an undetermined number of db. at the receiver. The actual receiver advantage will depend on the conditions under which you are operating at the moment.

1.3—Power Economy

There is another way of looking at this double-sideband versus single-sideband comparison. Suppose that we have a one-kilowatt, *average*-power, double-sideband AM transmitter and a one-kilowatt, *peak*-power, single-sideband transmitter. Please take a second look at the italicized words. Without regard to the bandwidths involved and assuming no noise or QRM problems, from the previous discussion we know that the two transmitters will produce exactly the same amount of audio at the receiver output. Again look at the italicized words. The conventional AM transmitter is running 1 kw. *average*-power all of the time when you are not talking, and 4 kw. on *peaks* of 100% modulation. On the other hand, the SSB transmitter probably runs about 300 to 500 watts (depending on the persons voice) *average* power only when you are talking. With no voice input, there is no transmitter output and only a small amount of idling plate power input (usually about ¼ to ½ of the maximum plate dissipation of the final tubes).

In our foregoing comparisons of the AM and SSB operation of a transmitter we have neglected the slight differences in operating efficiencies of the class C (in the AM case) and the class B (SSB case) amplifiers. We feel that this is only worthy of passing notice since the class C efficiency is about 75% while the class B efficiency is about 65%.

There is one further point that should be made at this time. The FCC has interpreted its "one kilowatt input to the final amplifier" for SSB transmitters in this way: The product of the *average* plate current in the final amplifier

	100-watt AM Transmitter	Same Final Amplifier Used on S. S. B.
Maximum power on voice peaks	400 watts	400 watts
Average power	100 watts (continuous)	Approx. 100 to 200 watts (only when talking)
Sideband power	50 watts (2 sidebands)	400 watts (1 sideband)
Bandwidth	6 kc.	3 kc.
Receiver output power (relative)	1	4
D. C. plate voltage	600 volts (example)	1200 volts

SSB and AM Comparison.

and the plate voltage applied should not exceed one kilowatt. This average plate current is the maximum value to which the needle swings on normal voice input. The meter used must have a time constant not to exceed ¼ second. This is the time constant found in practically all milliammeters normally used for amateur work. No additional damping of the meter may be used. What does all this mean? It means simply that the average input power may go to one kilowatt while the peaks of the SSB signal go as high as the linearity of the amplifiers will permit. The peak-to-average ratio in the operator's voice will be the determining factor in this matter. The usual male voice ranges anywhere from 1½ to 1 up to 3 to 1 in peak-to-average ratio. Thus, the maximum power for a "legal 1 kw." SSB transmitter may be anywhere from 1½ kw. to 3 kw. input—and please note: *all* of this power is useful.

1.4—Tuning In SSB Signals

Using a conventional superhet receiver, we have a choice of two systems of tuning in single-sideband signals.

METHOD A: (*Front-end carrier insertion*[39,40]): The receiver is operated in the usual manner for phone reception; that is, a.v.c. turned on, b.f.o. off, and r-f gain full-on. A BC-221 frequency meter, or similar auxiliary oscillator is used in conjunction with the receiver in the following manner: Center the SSB signal in the pass-band of the receiver's i-f stages by tuning for maximum kick on the *S*-meter. Tune the auxiliary oscillator slowly through the frequency occupied by the SSB signal and you will find that the signal will pass through the characteristic "Mickey Mouse" high pitched sound to the guttural "Mortimer Snerd" sound. In between these two extremes lies the narrow margin of intelligible copy. If by chance you have the artificial carrier on the wrong side of the SSB signal, the region between the high and low-pitched sounds will not yield understandable output, so move the oscillator signal over to the other side of the signal. You may have been trying to copy a lower-sideband signal as an upper-sideband signal, or vice versa.

Here is a rule-of-thumb to remember. To tune in a lower-sideband signal the carrier must be on the high frequency side, and to tune in an upper-sideband signal the carrier must be on the low frequency side of the SSB signal. Assuming that you have the carrier correctly placed, you can tune the receiver itself back and forth across the combination of the SSB and oscillator signal for optimum receiving conditions.

The amplitude of the artificial carrier should be adjusted by varying the oscillator coupling to the receiver antenna terminal, so that the maximum audio is recovered from the SSB signal with no signs of distortion due to overmodulation of the carrier. If heavy QRM is encountered, inserting a stronger carrier will prevent "capture effect" of the second detector by other strong carriers.

The crystal filter may be used in a normal manner to notch out interfering heterodynes and to reduce the receiver bandwidth to match the SSB signal bandwidth. You can tell which sideband is being transmitted by determining to which side of the carrier the receiver must be tuned in order to derive the maximum audio output from the signal. The receiver must be reasonably selective to do this, however.

METHOD B: (*b.f.o. carrier insertion*). With the receiver still in the conventional AM position, tune in the SSB signal for maximum *S*-meter kick. Turn off the a.v.c., back off the r-f gain, turn the audio gain up to maximum, and lastly turn on the b-f-o switch. *Slowly* tune the b-f-o pitch control (not the front end of the receiver) so that the b-f-o frequency passes through the band of frequencies occupied by the SSB signal. At one point in its journey the b-f-o will correctly demodulate the SSB signal and the "stuff" will come out as English and not Chinese. Bring up the r-f gain until a comfortable level is available. Advancing the gain too far will produce a garbled signal that cannot be detected no matter where you place the b-f-o carrier. Too much gain causes the SSB signal to overmodulate the b-f-o carrier causing plenty of second-harmonic distortion.

1.5—Adapters For Receiving SSB

There are three prime requirements that must be present in any *good* single sideband receiving system:

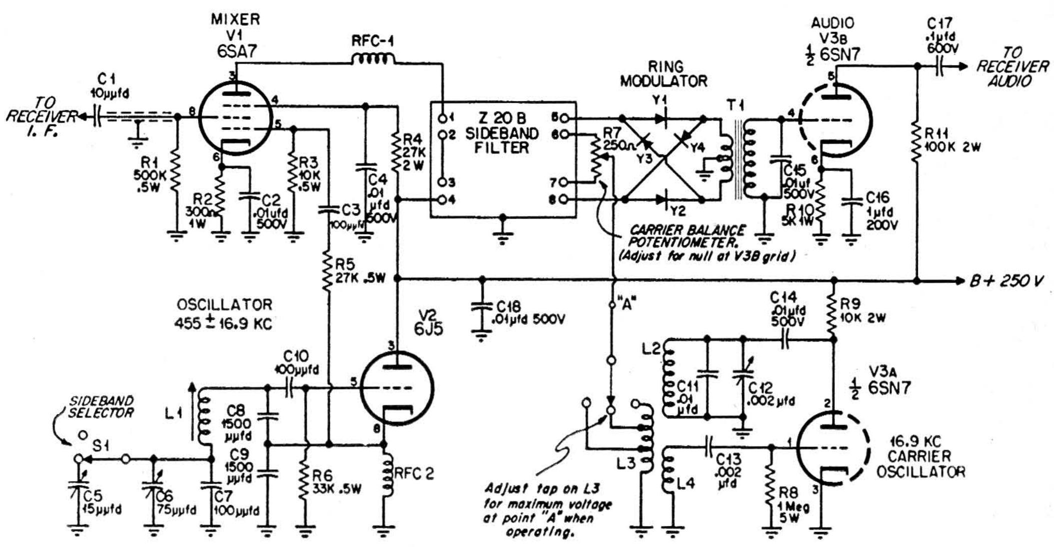
(1) Stability.
(2) Selectivity—3 kc. bandwidth with rejection outside this range, this means good skirt selectivity.
(3) Exalted carrier reception—this means having the b-f-o voltage at least ten to one-hundred times as great as any other signal that is present in the second detector. This reduces the monkey chatter caused by strong QRM-ing signals cross-modulating with each other upon detection.

I don't believe that much need be said about requirement *(1)*, either you have a stable receiver or you haven't and in the latter case you may be forced to do something about it. However, requirement *(2), selectivity,* can be dealt with to good results. The most desirable place to achieve selectivity in a receiver is as close to the front end as possible. This means immediately after the high frequency mixer. Pre-intermediate frequency selectivity is desirable because it prevents strong adjacent channel signals from producing distortion products in the latter i-f amplifiers. These distortion products appear as splatter on the signal to which we are listening while in reality these products are not transmitted by the offending stations. The only means available to us to achieve pre-i.f. selectivity are crystal filters[7,13] of some configuration, or one of the new mechanical filters recently introduced by *Collins Radio Co.*[5]

Post-intermediate frequency selectivity is fairly common in the Ham shack today. The "Lazy Man's Q-Fiver"[14] is probably the most familiar of the double-conversion arrangements. This unit, the BC-453 surplus receiver, is quite effective in increasing the selectivity of a receiver. The b.f.o. of the outboard receiver is used to furnish carrier for SSB signals. The b.f.o. frequency must be off-set either to the high side or the low side of the 85-kc. intermediate frequency to accommodate whichever is being copied—upper or lower sideband signals. The a.v.c. of most receivers is useful for listening to SSB signals when the BC-453 is used for detection.

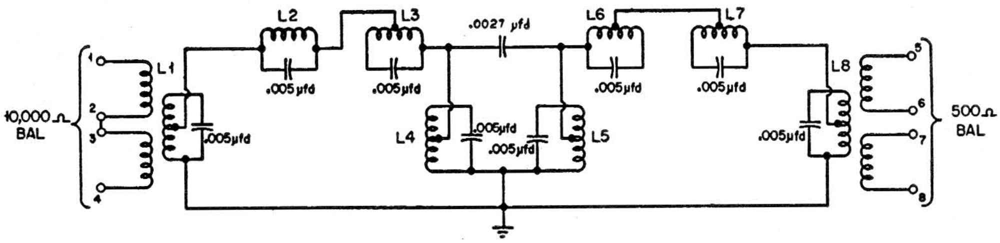
Coil Winding Data

L1—single winding from 455 kc. i-f transformer (slug tuned).
L2—6.8 mh., wound on type P-476930 toroid form.
L3—1/4 number of turns as on L2, wound on same P-476930 form, tapped every 20 turns.
L4—1/5 number of turns as on L2, wound on same P-476930 form.



R1—0.5 megohm, ½w.
R2—300 ohm, 1w.
R3—10,000 ohm, ½w.
R4—27,000 ohm, 2w.
R5—27,000 ohm, ½w.
R6—33,000 ohm, ½w.
R7—250 ohm potentiometer (carbon)
R8—1 megohm, ½w.
R9—10,000 ohm, 2w.
R10—5,000 ohm, 1w.
R11—100,000 ohm, 2w.
C1—10 μμfd. mica
C3, C7, C10—100 μμfd. silver mica
C2, C4, C14, C15, C18—0.01 μfd., disc ceramic, 500v.
C5—15 μμfd. midget variable
C6—75 μμfd. midget variable
C8, C9—1,500 μμfd. silver mica
C11—0.01 μfd. silver mica
C12—0.002 μfd. compression padder
C13—0.002 μfd. mica
C16—1.0 μfd., 200v. metalized paper
C17—0.1 μfd., 600v. metalized paper
S1—s.p.d.t. water switch
T1—Balanced 500 ohm to single grid transformer
RFC1—5 mh. choke
RFC2—2.5 mh. choke
Y1, Y2, Y3, Y4—Copper oxide varistor W.E. type D-98914 (Germanium diodes—1N34 may be used)

Fig. 1-5-A. This is the "modified Single Sider," filter type, SSB receiving adapter.



Winding	Approx. Total Ind. mhy.	Ind. to Tap mhy.	Test Capacity ufd.	Test Freq. - Kc.
L_1 Primary 10,000	16.5 (total)	(Note: - This winding is not critical)		
L_1 Secondary	14.9	4.98	.01937 for tap	18.44
L_2	10.96	no tap	—	21.50
L_3	20.25	14.90	.006798 for tap	15.814
L_4	14.4	2.408	.029883 for tap	18.80
L_5	13.9	2.65	.026235 for tap	19.20
L_6	12.06	4.13	.014524 for tap	20.50
L_7	18.4	5.13	.018366 for tap	16.585
L_8 Primary	14.9	6.6	.011289 for tap	18.44
L_8 Secondary 500	1.09 (total)	c.t.	.068120 for total winding	18.44

Fig. 1-5-B. Sideband filter Z-20-B (courtesy F. M. Berry, WØMNN). Note that the test capacity is used only to adjust the coil inductance to the tap. The total winding must be tuned to the test frequency by its own 0.005 μfd. mica condenser, and then wired into the circuit. Molybdenum Permalloy toroid forms are used with the L1 and L8 forms being type P-476930 and L2 through L7 forms being type P-284395. The toroids may be purchased from the Arnold Engineering Co., Merengo, Ill.

More elegant means may be built to copy SSB signals. Two such adapters are described briefly. It is recommended that the original references be consulted for the whole story.

20 Kc. Filter Adapter

The first of such adapters is the double-conversion 20-kc. filter adapter[2,22] as seen in *Fig. 1-5-A*. Operation is as follows: The i-f signals from the receiver are mixed in the 6SA7 stage, *V1*, with the oscillator signal from the Clapp oscillator, *V2*, to produce i-f signals in the 20-kc. region. The conversion oscillator operates approximately 16.9 kc. higher or lower than the 455-kc. frequency. The signals then pass through the 2.5-kc. wide bandpass filter which selects only the desired signal. The signal is then detected by mixing it with an exalted carrier at 16.9 kc. in the ring-modulator mixer using copper-oxide rectifiers as diodes. Germanium diodes such as the 1N34 may be used. The detected signals are then amplified by a conventional audio amplifier, *V3b*. By leaving the reinserted carrier at 16.9 kc. and switching the first heterodyning oscillator from 16.9 kc. above to 16.9 kc. below the 455-kc. input frequency sideband switch may be accomplished.

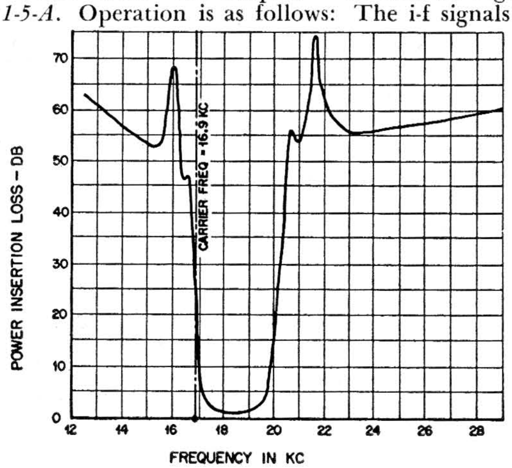


Fig. 1-5-C. Attenuation characteristic of the Z-20-B sideband filter shown in Figure 1-5-B.

Figure 1-5-B shows the schematic of a very fine filter which can be used in the adapter. *Figure 1-5-C* shows the response curve of the aligned filter.* I suggest that you consult an article by Berry[3] for particulars on how to wind toroids and for filter alignment procedures.**

* Also now available from **Barker & Williamson.** See advertisement page.

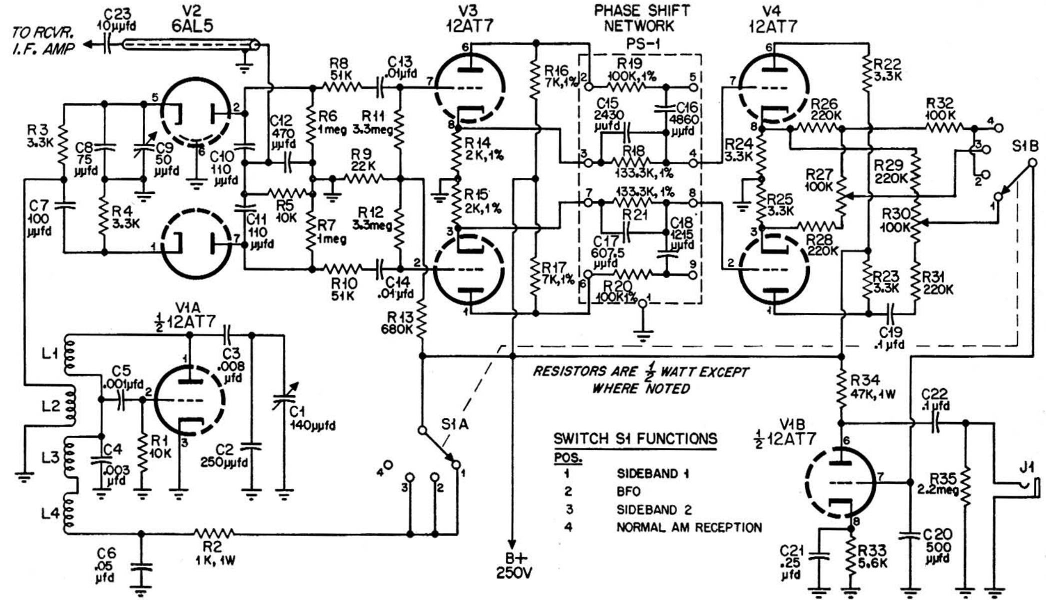
** Or write F. M. Berry, WØMNN, 1200 East 49th Terrace, Kansas City, Mo.

Phasing Type Adapter

The second adapter is the "Signal Slicer" originally described by Norgaard.[31] This unit, *Fig. 1-5-D,* utilizes the phasing principle of detection and post-detection combination to discriminate between sidebands of a received signal. The principle is very similar to that of the phasing-type single-sideband exciter. The incoming i-f signals are detected by the two diodes in the 6AL5, *V2,* by mixing them with the r-f voltage from the 455 kc. oscillator, *V1a.* This r-f voltage is divided into two equal parts and shifted in phase with respect to each other by 90° in the *R/C* networks *R3-C8, C9* and *R4-C7.* The two separate signals after detection are passed through the phase-shift networks designated *PS-1,* and then combined in the output of *V4.* One stage of audio amplification, *V1b,* is included. The net result is that depending on the position of the switch, *S1,* the signals on one side of the 455-kc. oscillator frequency will be combined in such a way as to be cancelled out to a major degree while the signals on the other side of the oscillator will be reinforced and will be heard in the headphones. This unit effectively splits the receiver selectivity characteristic in half—rejecting the one half while receiving the other. Again, I suggest that the original article be consulted for detailed constructional information and for alignment. The phase-shift network, *PS-1,* is available commercially from the *James Millen Co.* The unit is aligned and no further adjustment is needed. The complete adapter is also being marketed either in kit form or in wired form at a reasonable price.

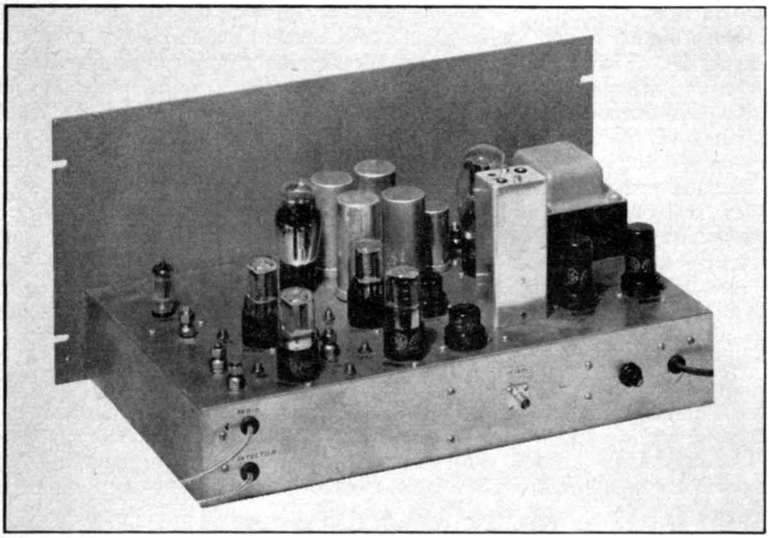
"Apparent" Broadness Of SSB Signals

This subject is just a little "touchy," but must be treated openly for the good of all. The complaint of many phone men of broad



R1—10,000 ohms, ½w.
R2—1000 ohms, 1w.
R3, R4, R22, R23, R24, R25—3,300 ohms, ½w., 5%
R5—10,000 ohms, ½w.
R6, R7—1 megohm, ½w.
R8, R10—51,000 ohms, ½w., 5%
R9—22,000 ohms, ½w.
R11, R12—3.3 megohms, ½w.
R13—680,000 ohms, ½w.
R14, R15—2,000 ohms, ½w., 1% precision
R16, R17—7,000 ohms, ½w., 1% precision
R18, R21—133,300 ohms, ½w., 1% precision
R19, R20—100,000 ohms, ½w., 1% precision
R26, R28, R29, R31—220,000 ohms, ½w.
R27, R30—100,000 ohm pot.
R32—100,000 ohms, ½w.
R33—5,600 ohm, ½w.
R34—47,000 ohms, ½w.
R35—2.2 megohms, ½w.
C1—140 μμfd. variable
C2—250 μμfd. mica
C3—0.008 μfd. mica
C4—0.003 μfd. mica
C5—0.001 μfd. mica
C6—0.05 μfd. paper
C7—100 μμfd., 5% mica
C8—75 μμfd., 5% mica
C9—50 μμfd. variable
C10, C11—110 μμfd. mica (matched within 1%)
C12—470 μμfd. mica
C13, C14—0.01 μfd., mica or paper
C15—2,340 μμfd. (paralleled 0.002 μfd. and 170-780 μμfd. trimmer)
C16—4,860 μμfd. (paralleled 0.0043 μfd. and 170-680 μμfd. trimmer)
C17—607.5 μμfd. (paralleled 500 μμfd. and 9-180 μμfd. trimmer)
C18—1,215 μμfd. (paralleled 0.001 μfd. and 50-380 μμfd. trimmer)
C19—0.1 μfd. paper
C20—500 μμfd. mica
C21—0.25 μfd. paper
C22—0.1 μfd. paper
C23—10 μμfd. mica
J1—open circuit phone jack
L1, L2, L3, L4—National R-100 r-f choke modified so that second pi serves as secondary (L2). The remaining pi's are connected in series as shown.
PS-1—two channel 90° phase shift network.
S1a, S1b—2-pole, 4-position wafer switch.

Fig. 1-5-D. The parts list and wiring diagram of a phasing type single-sideband adapter, modified by W9DYV, from the original design of W2KUJ.



SSB receiving adapter developed by W2KUJ and featured in the "G.E. Ham News" (November/December, 1948). Readers may obtain a useful "SSB Package" containing all the material published in the "G.E. Ham News" by writing to Lighthouse Larry, Tube Department, General Electric, Schenectady 5, New York

(Courtesy **General Electric**)

SSB signals adds up to about this: The receiver used is being operated with a.v.c. turned on and at full r-f gain. When tuned to a channel next to that occupied by a SSB station and copying no signal (or at least a weak one) the receiver is really wide open. The non-existence of a-v-c voltage has raised the gain of the r-f and i-f stages so that the *effective* bandwidth of the receiver is perhaps 15 or 20 kilocycles instead of the 3 kilocycles measured by the manufacturer at the "half-power" points. Naturally, when the adjacent-channel—SSB signal starts up it is going to be heard and quite loudly! The "jumping-up-and-down" of the SSB signal creates quite a terrifying effect on the a-v-c system too. Consider now what the case might have been if the SSB signal had been an unmodulated carrier of the same peak amplitude. The receiver a-v-c voltage would have been raised and held there, the result—reduced receiver sensitivity. The effective receiver bandwidth would have been reduced some and depending on its strength the desired signal might not have been heard at all. In the SSB case, the remedy is to increase the receiver selectivity by using the crystal filter, turn off the a-v-c, and manually adjust the r-f gain for just enough gain to do the job. You may be surprised to find that the SSB signal is not broad after all.

Chapter II

The Filter Method

As in the receiving adapters described in *Chapter I,* there are two methods of generating a single-sideband signal available. The older method in use for many years is the filter method. The other system which has become popular in the last few years is the phasing method. Each has its own points and careful work will produce a satisfactory signal using either system. Conversely sloppy workmanship will yield a sloppy signal with either system.

I would like to spend a little time dealing with each system, outlining what takes place, so that the beginner can better understand what he is trying to do when he tackles an SSB exciter.

2.1—Filter Methods and Designs

As we saw in *Chapter I* an AM signal is composed of a carrier (occupying no bandwith) and two sidebands. The sidebands are like the carrier, in that they are actual r-f signals distributed either side of the carrier in the spectrum. The sideband components corresponding to the lower audio modulating frequencies are close to the carrier, and the higher audio frequencies are proportionately farther away from the carrier. The two sideband signals are identical, in that they carry the same intelligence. The filter-type of SSB exciter must separate these two side-bands, transmitting one and attenuating the other. The matter of carrier suppression is secondary, and can be accomplished with relative ease.

It is not feasible, from a practical standpoint, to accomplish the filtering process at the final operating frequency for these reasons: First, finding circuit components to separate the sidebands, whether coils and condensers, or crystal filters, would be an almost impossible job. Second, if we wished to move frequency, all the circuit components would have to be retuned for the new set of conditions. There is, however, an easier way. Generate the SSB signal at some lower fixed-frequency, and then heterodyne it up with conventional mixers to the desired frequency. This procedure solves both problems for us, in that it makes our filtering problems simpler, and because the VFO can be used as the heterodyning oscillator to put us anywhere in a particular Ham band.

Early Model Filters

The first filters built for this purpose were in the higher audio frequencies, i.e., from about 10 kc. to 25 kc. Direct heterodyning into amateur bands from this low frequency is not practical, because, when two signals are mixed (or heterodyned), the output products are the original signals and the sum and difference of these original signals. If we heterodyned to 4.0 Mc., for example, the tuned circuits would not allow the original 25-kc. signal to appear in the output, *but* the original mixing oscillator near 4.0 Mc. would be present and both the *sum* and *difference* of the two signals would appear in the output. This would give us a steady carrier and an upper-sideband 25 kc. above this carrier and a lower-sideband displaced 25 kc. lower than this carrier. This is undesirable, to say the least. Thus it was necessary to first heterodyne the 25-kc. SSB signal up to about 450 kc., so that normal i-f transformers could discriminate between the various products of mixing, and select the desired one. Then it was necessary to heterodyne once more into the desired amateur band, making this last mixing-oscillator variable for v-f-o control of the output signal. The first transmitters described for amateur use utilized this system.[3,24,25] They were a little complex in adjustment, number of tubes, and tuned circuits used; however, they produced a very high degree of sideband attenuation. The advent of crystal filters for amateur use,[7,11] made possible the generation of SSB signals in the region of 400 to 500 kc. This eliminated the necessity of the extra heterodyning operation.

The sideband filter, whether at 25 kc. and made of toroid coils and condensers, or at 450 kc. and made of quartz crystals, must have certain characteristics to be acceptable for use. The filter, whether bandpass, or high-pass or low-pass in nature, must be capable of separating the sideband components close to the carrier on the high-frequency side from those close to the carrier on the low-frequency side. This dictates a sharp drop-off in the filter response characteristic, or, as the boys say, a good skirt response. We cannot obtain practical filter characteristics with a vertical drop-off, so we must be content with an attenuation of from 30 to 60 db. in about 600 cps. We can help this situation by moving our carrier frequency part way down this slope of the filter characteristic. We accomplish two things by doing this: (1) We attenuate the useless, power-wasting, low speech frequencies below about 400 cps and (2) allow the filter to be more effective on the portion of the undesired sideband nearest the carrier. A good rule of thumb to follow in placing the carrier on the slope of a filter is to put it about 20 db. down on the response of the filter characteristic. This should result in the transmission of few frequencies below 500 cps. If the carrier is so placed and the speech response is not up to maximum until you reach a frequency of 700 or 800 cps, the filter skirt is *not* steep enough, and measures must be taken to correct this. The commercial standard for a *good* sideband filter is 80 db. drop in the filter skirt in 1 kc.! This is a bit more than required for amateur use, but it will give you something to shoot for!

The high-frequency response of the filter characteristic and speech amplifier need only be enough for good speech intelligibility. This means that frequencies above 3000 cps should be attenuated, first of all, to conserve spectrum space, and to result in a pleasant sounding signal. Since we have eliminated the low-frequency speech components, cropping the high frequencies will produce a balanced-sounding signal that will be more intelligible.

The only thing remaining now is to heterodyne the 450-kc. SSB signal up to a desired frequency in an amateur band. Refer to *Fig. 2-1-A* for a block diagram of a simple crystal-filter SSB exciter. You will note that an *upper-sideband* signal is generated in the filter, and when it reaches the 4.0-Mc. band it has changed into a lower-sideband signal. How come? Take a pencil and figure it out for yourself. Take 450 kc. as a suppressed-carrier frequency, and 452 kc. as being the frequency of an upper-sideband resulting from a 2000-cps input tone. If we use the difference-mixture of these frequencies, and a 4350-kc. v-f-o voltage, do the following: (1) Subtract 450 kc. *(carrier)* from 4350 kc. Result—3900 kc. carrier frequency; (2) Subtract 452 *(sideband)* from 4350 kc. Result—3898 kc. sideband frequency. So, in the 75-meter band our sideband is *lower* in frequency than the carrier—a lower sideband.

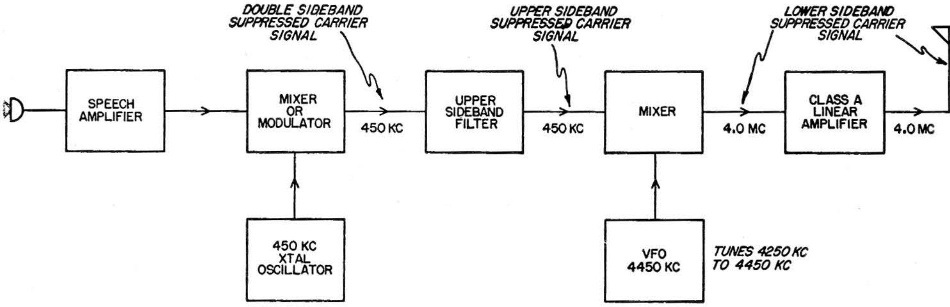


Fig. 2-1-A. Block diagram of a simple crystal-filter type SSB exciter.

2.2—Heterodyning

This section applies to the phasing type of exciters as well as to the filter exciters. Heterodyning, or mixing, as it is more commonly known, is the process whereby two r-f voltages of different frequencies are combined in a non-linear device so that either the sum or difference of the two frequencies is present in the output. Normally, tuned circuits are used to select the mixture product we want. The non-linear device mentioned in the formal sounding definition can be an ordinary vacuum tube, either triode or multi-grid type. Diodes, either vacuum or germanium, also serve as non-linear elements. In the transmitter-exciters to be described, both types are used. In mixing in a non-linear device, the output amplitude of the desired product must be in linear amplitude relation with its corresponding input signal. What does this fancy sounding statement mean? It means that when heterodyning an SSB signal from one frequency to another, the output SSB signal must vary in the same proportion, amplitude-wise, as the input SSB signal. To accomplish this, the mixing oscillator voltage must be several times greater than the SSB signal, the usually accepted ratio being ten to one. For every volt of signal, we will require ten volts of mixing oscillator voltage. This is a safe figure and should eliminate any possibility of signal distortion taking place in the mixer.

2.3—Amplifiers

We have dealt briefly with the filter systems used in generating a SSB signal. This SSB signal has variations in amplitude of its various r-f components, which means that we cannot use conventional class C amplifiers to amplify it. As we must faithfully reproduce the signal as originally generated, the amplifiers must be linear. Linear amplification, in words of one syllable, means that, if the input voltage doubles in amplitude (from 1 to 2 volts, for example), the amplified output voltage must change in the same proportion (20 to 40 volt increase).

There are various classes of linear amplifiers —depending on how much driving power is used and what amount of grid-bias is used. Class A, AB_1, AB_2, or class B amplification may be used depending on the voltage or power level involved. Ordinarily, class A or class AB_1 amplifiers are used at low levels for voltage amplification of generated SSB signals. Class AB_2 and B stages are used for power amplification in the high-level final amplifier stages. Amplifiers will be dealt with in more detail in a later chapter. Here, we will be content with only the voltage amplifier stages—class A and AB_1.

Further discussion will be deferred until we cover the section describing specific exciters.

2.4—A Crystal-Filter SSB Exciter

As representative of the filter-type exciter, the circuit shown in *Fig. 2-4-A* was chosen. Basically, the circuit is that of Edmunds,[11] and has long since been nicknamed the "W1JEO filter." The author felt that the original filter has some inherent disadvantages, namely: The alignment was a compromise one, in which *best* sideband suppression did not occur when *best* pass-band characteristics were realized. The slope of the filter characteristic would not be steep enough when the adjustments for maximum undesired sideband suppression were made. Generally, there would also be a large dip in the middle of the transmitted sideband characteristic. Many of the boys using this type of filter had successfully modified it as follows: Following Edmunds' original alignment procedure with the three-crystal filter, as described in *QST*, careful adjustment was made for best pass-band shape, without worrying too much about the suppressed sideband characteristic. Then the frequencies at which humps appeared in the filter characteristic on the suppressed-sideband part of the curve were noted. Additional FT-241 crystals were chosen so that their *series-resonant* frequencies were at the frequencies of worst attenuation. These extra crystals were then connected in parallel with the input crystal *(Crystal "B" in Fig. 2-4-A)*, with the output crystal *(Crystal "D")*, or both places. As few as two or as many as ten extra crystals have been used. Ken Stiles, W2MTJ, originally conceived these modifications, and suggests that additional crystals may be used to improve the steepness of the filter cut-off characteristic next to the carrier frequency.

Exact Determination of Crystal Frequencies

The series-resonant frequency of a FT-241 low-frequency crystal may be determined in the following manner: Connect the crystal in series with the hot lead of a BC-221 frequency meter (or a signal generator) and the input lead of a scope, sensitive r-f vacuum-tube voltmeter, or some other indicating device that will respond to the frequency range of 400 to 500 kilocycles. Slowly tune the signal generator through the range of the crystal, and a sharp kick will be noted on the VTVM or oscilloscope. This sharp rise indicates the series-resonant frequency of the crystal. Carefully check this peak. Shunt circuit capacity should not move it, so no great care need be taken in circuit arrangement.

One of the SSB gang, W4RL, who tried this modified filter scheme, said that he used two additional crystals. One of the crystals was chosen for the channel between the carrier crystal *(Crystal A)* and the rejection crystal *(Crystal B)*. The other extra crystal was chosen to be very close to the carrier suppression crystal *(Crystal D)*. As in Edmunds' original article, *Crystals A* and *D* are chosen at the carrier frequency. *Crystal C* is the band-pass crystal—used to pass the desired frequencies—and for upper sideband generation is one channel higher than the carrier frequency crystal. *Crystal B* is the lower-sideband suppression crystal and is two channels *below* the carrier frequency crystal.

The above-mentioned modifications will improve the sideband suppression of the W1JEO filter to approximately 40 db., as opposed to the 25 db. available on the unmodified filter.

Coil Winding Data

L1, L2, L7—2.5 mh.

L3, L6—45 turns, #28 formex on National XR-50 coil form.

L4, L5—3 turn link of insulated wire wound on cold end of L3 and L6.

L8—18 turns #18 formex wire on National XR-50 coil form.

L9—8 turns (insulated wire) wound on cold end of L8.

L10—4.0 Mc: 30 turns, #18 on 1½" dia. plug-in form.
7.3 Mc: 15 turns, #18 on 1½" dia. form.

L11—4 turns, #18 wound on cold end of L10.

Circuit Particulars

Certain requirements were outlined for the exciter shown in *Fig. 2-4-A*. These were: (1) Band switching operation on at least two bands, (2) carrier reinsertion for working the unenlightened, (3) voice control, and (4) enough output to drive most of the higher-powered final amplifier tubes without an additional amplifier stage.

Referring to *Fig. 2-4-A,* we will consider the schematic in logical order. The speech amplifier, *V1a* and *V1b,* is conventional, and has sufficient gain for a high impedance microphone. *V2a,* the mixer portion of *V2,* is conventional also and is pretty much as originally described by Edmunds. However, the carrier crystal oscillator, *V2b,* has been changed to a Pierce oscillator because it was found that the grid-plate type of oscillator used originally was very sluggish. By varying condenser *C9* the oscillating frequency of the carrier crystal may be moved a small amount so that it may be moved into the attenuation notch of the carrier rejection crystal of the filter *(Crystal D)*.

Transmitter carrier insertion is accomplished by *V3,* which serves as a conventional amplifier with choke, *L2,* as plate load. Potentiometer, *R15,* controls the amount of carrier insertion.

The filter has been discussed previously. It was felt that sideband switching is not normally required, so only the one sideband filter is shown. You will note that the filter is set up for upper sideband generation. Normally the lower sideband is transmitted on both 75 meters and 40 meters. In order to come out with a lower sideband on the 4.0-Mc band, the v.f.o. must operate *above* the 75-meter band in frequency by an amount equal to the filter carrier-frequency.

This will make our v.f.o. tune the range of approximately 4.25 Mc. to 4.45 Mc. If 40-meter phone is contemplated it will be best to operate the unit in the manner just described. If the filter is set up for lower-sideband generation, and the v.f.o. operated at 3.55 Mc., the second harmonic of the v.f.o. will fall in the 7.0-Mc band. This signal when passed through *V5,* the second mixer used for 40-meter operation, would appear in the output as an undesired c-w signal. The double-tuned transformer, *L3* and *L6,* was used to further attenuate the harmonic output of the mixer and prevent the generation of spurious "birdies." Even if the second harmonic of the v.f.o. is negligible as far as actual radiation is concerned, it would certainly be strong enough to block the local receiver even when not transmitting. It is generally the custom to leave all heterodyning oscillators run when in the stand-by position in order to insure better frequency stability. However, with the v.f.o. at 4.4 Mc., the second-harmonic is at 8.8 Mc. and even the heterodyne-mixture of the fundamental 4.4-Mc. voltage and the 3.3-Mc. mixing voltage of *V5* will fall well outside the 7.0-Mc. band at 7.7 Mc. You can see that this business of heterodyning signals here and there must not be taken too lightly. Always sit down with a pencil and paper and figure out all the possible combinations.

The BC-457 Command transmitter lends itself nicely for v.f.o. use. It is recommended that not over 150 volts regulated plate-supply be used. This will minimize the drift and will still provide plenty of v-f-o signal for the 4.0-Mc. mixer stage.

For 4.0-Mc. operation, the mixer stage employing *V5* is not used, as can be seen by the position of switch *S1* in *Fig. 2-4-A*. 40-meter operation makes necessary the changing of *S1,* as already indicated and the changing of plug-in coil, *L10,* in the output stage, *V6*. The second mixer output circuit will be broad enough to cover the 100 kc. of the 40-meter phone band without re-peaking when QSY-ing.

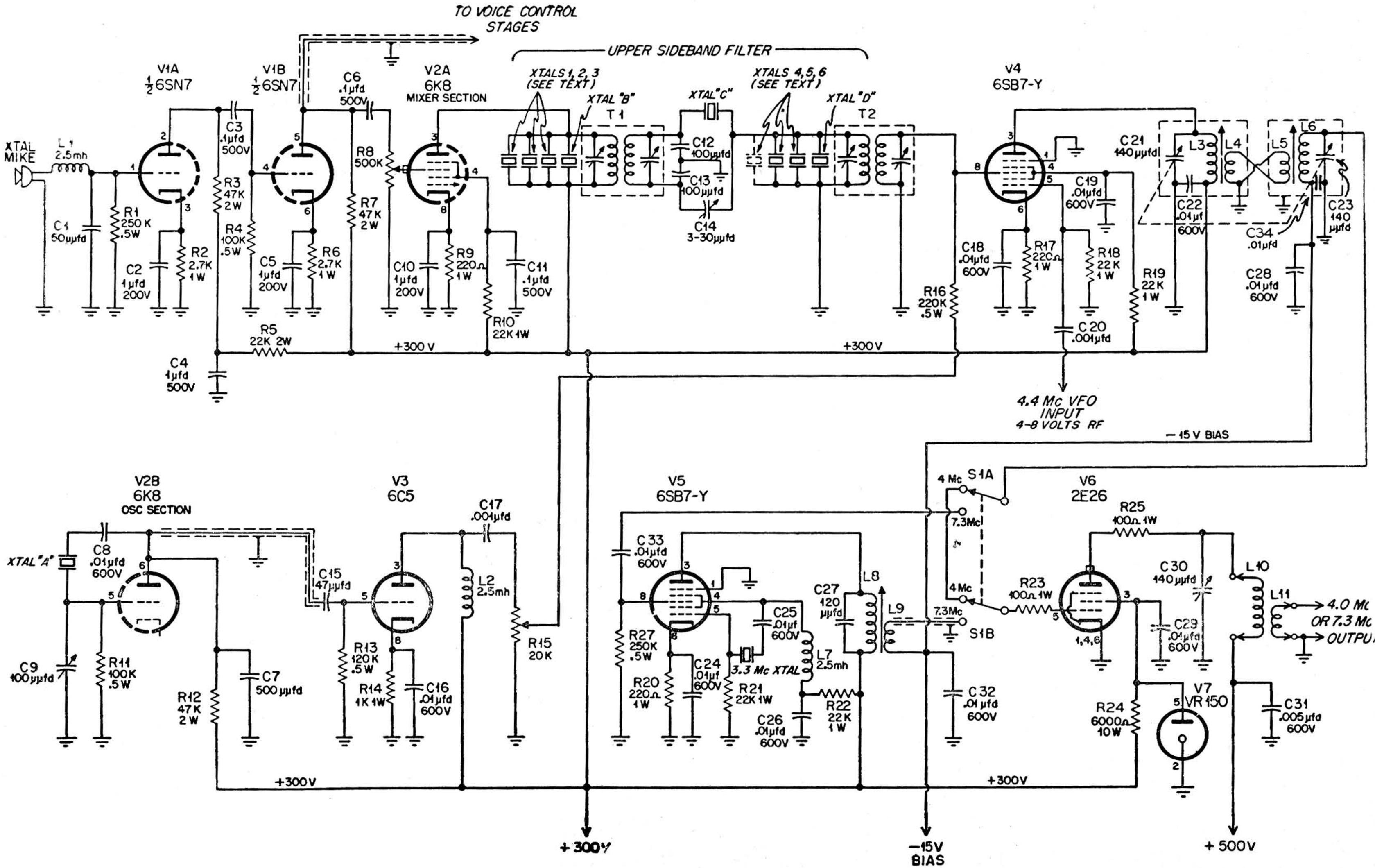
Fig. 2-4-A. Parts list and wiring diagram of the crystal filter exciter.

R1, R27—250,000 ohm, ½w.
R2, R6—2,7000 ohm 1w.
R3, R7, R12—47,000 ohms, 2w.
R4, R11—100,000 ohms, ½w.
R5—22,000 ohms, 2w.
R8—500,000 ohm potentiometer.
R9, R17, R20—220 ohm, 1w.
R10, R18, R19, R21, R22—22,000 ohms, 1w.
R13—120,000 ohms, ½w.
R14—1,000 ohms, 1w.
R15—20,000 ohms potentiometer.
R16—220,000 ohms, ½w.
R23, R25—100 ohms, 1w.
R24—6000 ohms, 10w.
C1—50 μμfd., mica.
C2, C5, C10—1 μfd., 200v., metallized paper.
C3, C6, C11—0.1 μfd., 500v., metallized paper.
C4—1 μfd., 500v., metallized paper.
C7—500 μμfd., mica.
C8, C16, C18, C19, C22, C24, C25, C26, C28, C29, C32, C33, C34—.01 μfd., 600v. disc ceramic.
C9—100 μμfd., trimmer
C12, C13—100 μμfd., mica (matched).
C14—3 to 30 μμfd., compression trimmer.
C15—47 μμfd., mica.
C17, C20—0.001 μfd., mica.
C21, C23—140 μμfd., per section, dual section.
C27—120 μμfd., silver mica.
C30—140 μμfd., air variable.
C31—0.005 μμfd., 600v., feed-through ceramic.
T1, T2—455-kc i-f interstage transformer, J. W. Miller.

S1a, S1b—d.p.d.t. ceramic wafer switch.

Xtal A, B, C, D—see Edmunds, Nov., 1950 QST.

Xtal 1, 2, 3, 4, 5, 6—see text.


TO VOICE CONTROL STAGES
UPPER SIDEBAND FILTER
XTAL MIKE
L1 2.5mh
C1 50μμfd
R1 250 K .5W
V1A ½ 6SN7
C2 1μfd 200V
R2 2.7K 1W
C3 .1μfd 500V
R3 47K 2W
R4 100K .5W
C4 1μfd 500V
R5 22K 2W
V1B ½ 6SN7
C5 1μfd 200V
R6 2.7K 1W
C6 .1μfd 500V
R7 47K 2W
R8 500K
V2A 6K8 MIXER SECTION
C10 1μfd 200V
R9 220Ω 1W
R10 22K 1W
C11 .1μfd 500V
+300 V
XTALS 1, 2, 3 (SEE TEXT)
XTAL "B"
T1
XTAL "C"
C12 100μμfd
C13 100μμfd
C14 3-30μμfd
XTALS 4, 5, 6 (SEE TEXT)
XTAL "D"
T2
+300V
R16 220K .5W
V4 6SB7-Y
C18 .01μfd 600V
R17 220Ω 1W
R18 22K 1W
C19 .01μfd 600V
C20 .001μfd
R19 22K 1W
4.4 Mc VFO INPUT 4-8 VOLTS RF
C21 140μμfd
L3
L4
C22 .01μf 600V
L5
L6
C23 140 μμfd
C34 .01μfd
C28 .01μfd 600V
—15 V BIAS
V2B 6K8 OSC SECTION
XTAL "A"
C8 .01μfd 600V
C9 100μμfd
R11 100K .5W
R12 47K 2W
C7 500 μμfd
C15 47μμfd
V3 6C5
R13 120K .5W
R14 1K 1W
C16 .01μfd 600V
C17 .001μfd
L2 2.5mh
R15 20K
+300V
+ 300V
V5 6SB7-Y
C33 .01μfd 600V
R27 250K .5W
R20 220Ω 1W
C24 .01μf 600V
3.3 Mc XTAL
R21 22K 1W
C25 .01μf 600V
L7 2.5mh
C26 .01μfd 600V
R22 22K 1W
C27 120 μμfd
L8
L9
4 Mc
S1A
7.3Mc
4 Mc
7.3Mc
S1B
C32 .01 μfd 600V
—15V BIAS
+300V
V6 2E26
R25 100Ω 1W
R23 100Ω 1W
1,4,6
C30 140 μμfd
C29 .01μfd 600V
R24 6000Ω 10W
V7 VR150
L10
L11
4.0 Mc OR 7.3 Mc OUTPUT
C31 .005μfd 600V
+ 500V

The 6SB7Y tube was chosen as the mixer because of its very high conversion-transconductance. This makes for a very "hot" mixer and a minimum of stages.

The choice of the 2E26 as output stage was brought about after hearing the complaints of many of the boys who have experienced trouble "taming down" the 6AG7. The excellent isolation provided by the 2E26 was a "natural." Also the fact that the 2E26 was capable of more output at higher plate voltage made it desirable for driving some of the larger tubes directly, without any additional amplification. The fewer amplifier stages you have to use, the less trouble you are going to have keeping the system linear—and linearity means a clean SSB signal.

One note on construction in the 2E26 stage. Keep the grid circuits below the chassis, and the plate tank coil and tuning condenser above the chassis. Use of the ceramic feed-thru condenser, *C31,* is recommended.

Alignment

After the filter is aligned per the original procedure and preceding suggestions, the following should be done: Using the Ham-shack communications receiver set up the v.f.o. on approximately 4.45 Mc. Place *S1* in the 4.0-Mc. position. Feed an audio tone into the microphone input, or insert carrier by advancing *R15.* Tune the receiver to the difference-frequency of the low-frequency carrier crystal (*Crystal A*) and the v.f.o. This should be in the neighborhood of 4.0 Mc. If you are unable to find the signal at this point, lightly couple an insulated wire from the receiver antenna post to the grid of the 2E26, *pin #5.* With the ganged condensers, *C21* and *C23,* set about two-thirds engaged, adjust the slugs of *L3* and *L6* for maximum signal as indicated on the receiver *S*-meter. Disconnect the receiver, use an output indication device (neon bulb or other r-f indicator), and tune the 2E26 plate tank circuit to resonance.

For 40-meter operation, change *S1* to the 7.0-Mc. position and plug the 40-meter coil, *L10,* in the 2E26 output. First, tune the receiver to the frequency of the 3.3-Mc. crystal to make sure it is oscillating. Then, tune the receiver to the sum-frequency of the 4.0-Mc. SSB signal and the frequency of the 3.3-Mc. crystal. This should be in the 7.3-Mc. phone band. Adjust the slug of *L8* for maximum output signal; then again resonate the 2E26 output tank circuit by adjusting *C30.*

You should now be ready for operation. The unit described should give between 10 and 15 watts output "without a sweat." Properly loaded by an antenna or by a following amplifier, the 2E26 should be perfectly stable and free of regenerative tendencies. If such should not be the case, a swamping resistor should be placed across *L10* to discourage self-oscillation. The exact value of this resistor will depend on how bad a case of instability you are afflicted with. Start with a high value and gradually reduce it until the stage tames down. It is better to be conservative in this matter and over-swamp slightly than to suffer periodic seizures of instability. Under normal operation the plate current of the 2E26 should kick upward slightly. No grid current should flow in this stage, however.

Chapter III

The Phasing Method

The heart of the phasing rig is the audio phase-shift network. This formidable sounding gadget is merely a group of resistors and condensers chosen carefully and arranged in a certain way so that if a single tone in the voice frequency range is fed into the input the following takes place: The voltage is immediately divided into two channels and the phase relation of the separated voltages with respect to each other is changed, so that instead of the two voltages being in phase at all frequencies, they differ by 90° when they reach the two sets of output terminals. This 90° phase difference is maintained at all the speech frequencies in which we are interested—normally from 250 cps to 3000 cps. The other requirement of the network is that the two output voltages must have exactly the same amplitude with respect to each other for all frequencies in the speech range. These are quite rigid requirements, and it was only in recent years that designers have been able to come up with practical networks that could be built.[10] You will note that 1% components are used in the network—this accuracy is very necessary, and any deviation from this will adversely affect the degree of sideband suppression.

3.1—R.F. Phase-Shift Networks

There is one more phase-shift network in phasing exciters that must be considered. It is the 90° r-f phase-shift network. This is concerned with only one frequency (not a band of frequencies like the a-f network), and is a cinch to build and to understand. There are various ways of obtaining r-f phase shifts. The easiest method is that of using two lightly-coupled tuned circuits as follows: One of the circuits is detuned on the *high-frequency side* to the 3 db. point. (This is where the voltage across the tuned circuit is 3 db. or 70% less than the voltage when tuned to the carrier frequency.) The other tuned circuit is tuned to the 3 db. point on the *low-frequency side* of the carrier. Under these conditions the voltages existing across the two coupled link windings (see *Fig. 3-2-A*) are 90° apart in the phase relationship.

Now that we understand all about these phase-shift networks (who said that?) let's proceed with the theory of what happens in a phasing exciter. Follow along with the block diagram, *Fig. 3-1-A.* The speech amplifier output is fed into the audio phase-shift network, where, as we mentioned, two equal outputs are

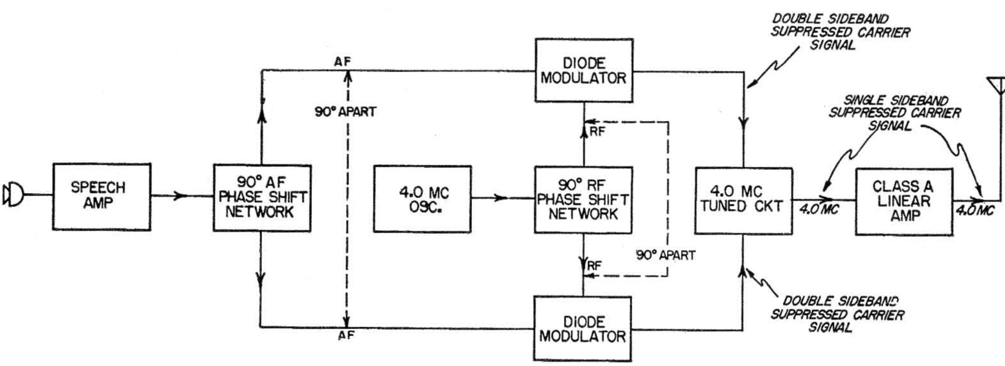
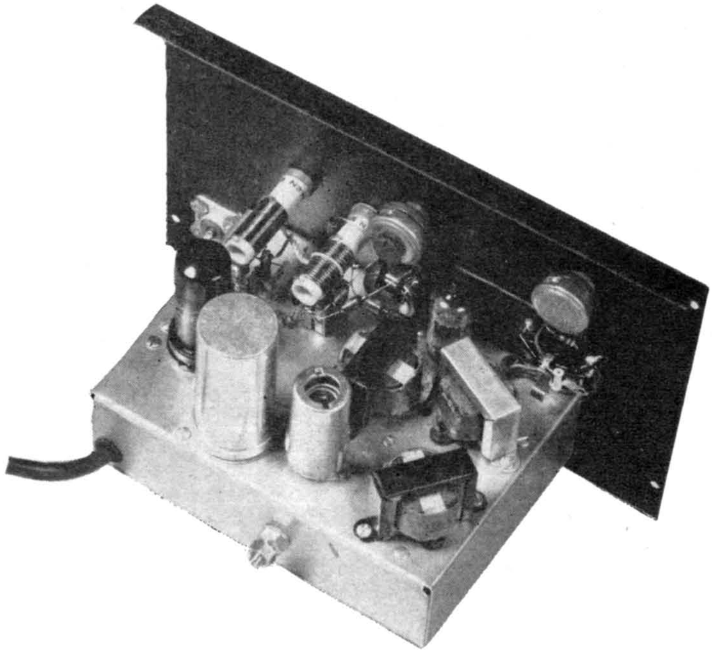


Fig. 3-1-A. This is a block diagram of the "SSB, Jr." shown on the following page. It is a fundamental-frequency type of phasing exciter. Full details in the Nov-Dec., 1950 issue of the "G.E. Ham News."



Above chassis rear view of the "SSB, Jr."
(Courtesy "G.E. Ham News")

obtained. These equal outputs are each fed into separate diode balanced-modulators. These balanced modulators are crystal mixers (mentioned in the section on *heterodyning* in *Chapter II)* that are arranged so that the mixing oscillator voltage is balanced out in the output tank circuit. The r-f carrier is generated in a crystal oscillator and fed into the r-f phase-shift network, where the two outputs are fed into the balanced modulators mentioned above.

Look again at what we have. We have two separate channels where a carrier is combined with a modulating frequency. The carriers have the same frequency, but are 90° out of phase with each other. The modulating frequencies in the two channels are the same amplitude, but also differ by 90° from each other. The individual balanced modulator output consists of a double-sideband suppressed carrier signal. Remember—we balanced out the carriers. These two sets of sidebands are then combined in a common tank circuit.

Sideband Switching

All of this fooling around with phase relationships now pays off. Let's take a specific example: The upper sideband voltages of the two channels will be equal in amplitude, but due to all the phase changes we have purposely made along the line the component voltages of the two upper sideband are exactly 180° out of phase. This means that when combined in a linear device like the tuned tank circuit, the upper-sideband voltages will cancel each other. However, the lower sideband voltages in both channels will not be out-of-phase, but will be exactly in-phase and will add up vectorially to give a lower-sideband voltage twice as large as that existing in either channel. In order to switch sidebands—attenuate the lower and transmit the upper sideband—all that need be done is to reverse the phase of the audio voltage feeding *one* of the balanced modulators. This is done by interchanging any two audio leads going into one of the balanced modulators.

Ordinary double-sideband-with-carrier transmission may be accomplished with a phasing transmitter. To accomplish this, one of the balanced modulators must be disabled, and the carrier of the other balanced modulator unbalanced sufficiently to provide enough carrier to permit proper demodulation of its sidebands at a distant receiver. There is one good point about producing AM this way. If overmodulation takes place, negative peak clipping does not occur, therefore, there are no spurious splatter products generated. However, there is second harmonic distortion present upon detection, but a fair amount of this may be tolerated before the signal becomes unintelligible.

Phase Modulation

Phasing-type exciters can also be made to produce phase modulation. Phase modulation is produced by having a double-sideband suppressed-carrier signal as produced in either of our balanced modulators combined with a carrier that is shifted in phase by 90° from that originally present in the balanced modulator in question. This can be easily accomplished in our exciter by taking the sidebands with no carrier from one balanced modulator and adding a carrier with no sidebands from the other balanced modulator. Switches can be easily provided to do this.

3.2—The Phasing Exciter

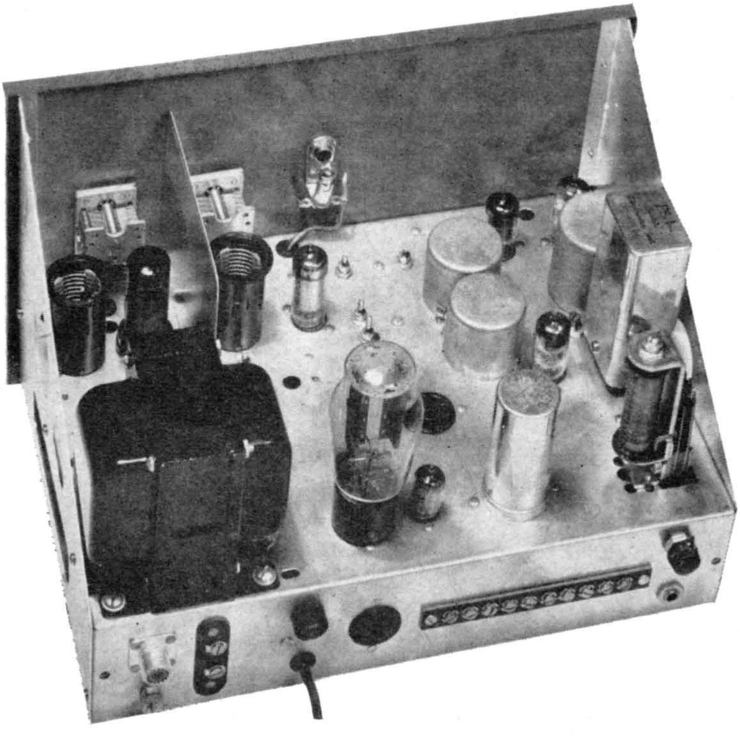
We are grateful to W9DYV of *Central Electronics* for permission to publish a portion of his "Multiphase" exciter schematic. These very popular units are available commercially, either in kit form or completely wired.

The circuit is basically that of the "SSB, Junior," devised by Norgaard,[30] but with improvements that make multi-band operation possible. The original *SSB, Jr.* exciter was a fundamental-frequency operating gadget. It operated fundamentally at 4.0 Mc., and if the operator wished to QSY more than a few kilocycles, he found it necessary to realign the r-f phase-shift network in order to maintain good sideband suppression. Wes, W9DYV, modified the idea by generating the SSB signal at a fixed frequency of 9.0 Mc. and heterodyning into the desired amateur band with a separate mixer stage, just as was done in the filter-type exciter. V.f.o. operation is, of course, possible when using this scheme.

Remembering the foregoing discussion about the phasing method of generating SSB signals, we now refer to *Fig. 3-2-A*. *V1a* and *V1b* are the usual speech preamplifiers. *V2a* is the a-f driver stage which feeds the audio phase-shift network through the transformer, *T1*. Construction by the average Ham, of the a-f phase-shift network, is possible, but a complete aligned unit is available for about the same price for which we could buy the necessary precision stable components.

V3a and *V3b* are the dual-channel amplifiers wherein the audio balance is obtained by adjustment of the cathode resistor, *R18*. The transformers, *T2* and *T3*, are plate-to-low impedance line transformers used to drive the diode balanced modulators. W9DYV indicates that suitable units are not easily obtained on the market so he has special transformers built for his production needs. Switch, *S1*, is the function switch which permits selection of sidebands, or of AM or PM transmission. Studying the switch positions will show that changing from one sideband to the other merely interchanges the connections on the output of *T2*, while switching to the AM and PM positions disconnects the secondary of *T3* from its associated balanced modulator.

The audio in the two separate channels is applied, in series with the r.f. from the link windings on *L1* and *L2*, to the balanced modulators. *Y1* and *Y2* are the diodes of the bal-



Above chassis rear view of the Model 10B "Multiphase." This unit includes a power supply not shown in the general schematic of this exciter on the following pages. The output stage, V5, with plug-in coils is in the upper left in the shielded section. In the center of the chassis is V4. The audio phase shift network, PS-1, is mounted in the rectangular can beside the relay at the lower right.

(Courtesy Central Electronics)

Coil Winding Data

L1, L2—9.0 Mc., tank wound on Cambridge Thermionics slug-tuned form LS-3.

L3—9.0 Mc., balanced mod. tank wound on LS-3 form.

L6, L7—15 Mc. v-f-o harmonic traps wound on form.

L6, L7—15 Mc. v-f-o harmonic trap wound on LS-3 form.

L8, L9—4.0 Mc., 30 turns #18 on 1½" dia. plug-in form.
7.3 Mc., 15 turns #18 on 1½" dia. plug-in form.

Note—L1 through **L7** may be obtained ready-made from Central Electronics.

anced modulator involved with *L1* and *T2*. Likewise *Y3* and *Y4* are associated with *L2* and *T3*.

The R-f Circuitry

V2b is the 9.0-Mc. crystal oscillator, and utilizes *L1* as its plate tank coil. The resonant frequency of the *L1/C11* combination must be higher in frequency than 9.0 Mc., in order for the crystal to oscillate; therefore, to obtain our 90° r-f phase-shift, *L2* and *C13* will have to be tuned *lower* in frequency than 9.0 Mc. You will notice that there is no physical connection from *L2* to the oscillator tank. *L1* and *L2* may be mounted physically within a couple of inches of each other, and the circuit capacity furnishes sufficient coupling to do the job. Condensers *C9* and *C10* prevent r.f. from getting into the audio transformers; yet they must not appear as a low reactance at audio frequencies.

Carrier balance is accomplished by adjusting *R23* and *R24*. Both must be carefully balanced in order to cancel the carrier completely.

The balanced-modulator output transformer *L3* is the point where the outputs of the two previously separate channels are combined. The additional double-tuned transformer, consisting of *L4* and *L5*, is necessary to further attenuate the second-harmonic of the 9.0-Mc. oscillator, generated in the germanium diodes of the balanced-modulator stage.

The mixer stage, *V4*, is conventional with the exception of the trap circuits, *L6* and *L7*, and their associated tuning condensers. They are necessary, when operating in the 14-Mc. amateur band, to attenuate the third harmonic of the 5.0-Mc. mixing voltage. The same v.f.o. may be used for both 4.0- and 14-Mc. operation, for the difference between the 9.0-Mc. SSB signal and the 5.0-Mc. mixing voltage will put the output at 4.0 Mc. If the sum-mixture is selected by the mixer plate tank circuit, *L8* and *C27*, the output will be in the 14.0-Mc. amateur band.

The output amplifier, *V5*, is the inevitable 6AG7. The output is shunt-fed, and obtained by means of a tap on the plate coil.

The swamping resistors, *R32* and *R38*, are usually necessary in order to stabilize the 6AG7 stage. Without these resistors self-oscillation often results. The exact values of these resistors are not shown; the highest value of resistance commensurate with stable operation should be used. *Always* insure that all stages are rock solid before the unit is "buttoned up."

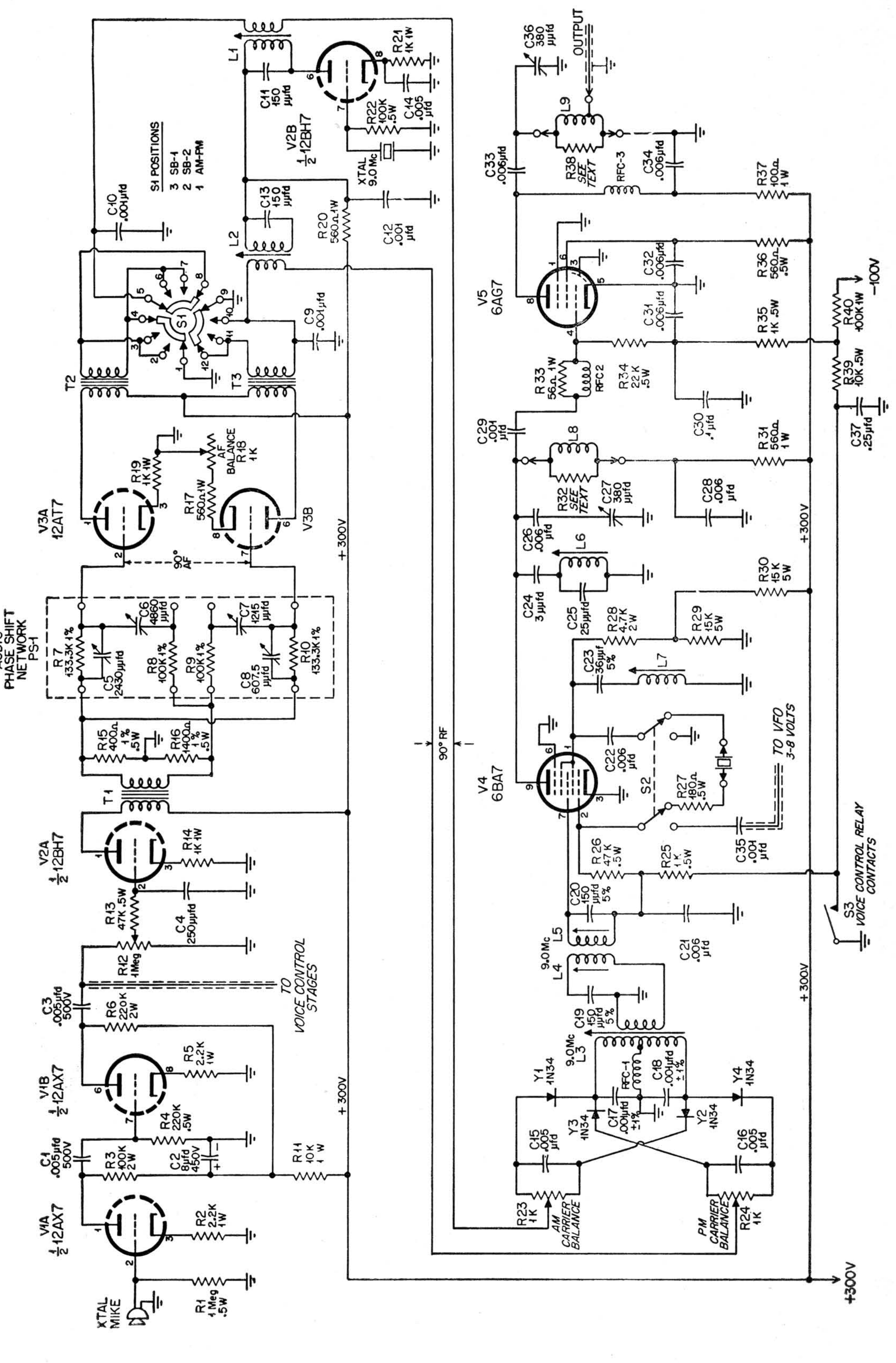
You will note, when in stand-by position *(S3* open), that there is a minus 100 volts of bias applied to the control grid of the 6AG7 and the oscillator grid of the 6BA7-mixer. This thoroughly squelches the output of the exciter. When *S3* is closed, the mixer returns to its normal operating condition, and an operating bias of about minus 10 volts is applied to the control grid of the 6AG7.

Forty-meter operation is also possible with this unit. However, it is *not* recommended that a mixing voltage at 1.8 Mc. be used, because of the various harmonics of this frequency that

R1—1 megohm, ½w.
R2, R5—2,200 ohms, 1w.
R3—100,000 ohms, 2w.
R4—220,000 ohms, ½w.
R6—220,000 ohms, 2w.
R7, R10—133,300 ohms, 1%, ½w.
R11—10,000 ohms, 1w.
R8, R9—100,000 ohms, 1%, ½w.
R12—1 megohm potentiometer.
R13, R26—47,000 ohms, ½w.
R14, R19, R21—1,000 ohms, 1w.
R15—400 ohms, 1%, ½w.
R16—1400 ohms, 1%, ½w.
R17, R20, R31—560 ohms, 1w.
R18—1,000 ohm potentiometer.
R22—100,000 ohms, ½w.
R23, R24—1,000 ohm carbon potentiometer.
R25, R35—1,000 ohms, ½w.
R27—180 ohms, ½w.
R28—4,700 ohms, 2w.
R29, R30—15,000 ohms, 5w.
R32, R38—swamping resistor (see text).
R33—56 ohms, 1w.
R34—22,000 ohms, ½w.
R36—560 ohms, ½w.
R37—100 ohms, 1w.
R39—10,000 ohms, ½w.
R40—100,000 ohms, 1w.
C1, C3—0.005 μfd., 500v., paper.
C2—8 μfd., 450v., electrolytic.
C4—250 μμfd., mica.
C5—2,430 μμfd., (0.002 μfd., mica ± 5% with 170-780 μμfd., trimmer in parallel).
C6—4,860 μμfd., (0.0043 μfd., mica ± 5% with 170-780 μμfd., trimmer in parallel).
C7—1,215 μμfd., (0.001 μfd., mica ± 5% with 50-380 μμfd., trimmer in parallel).
C8—607.5 μμfd., (500 μfd., mica ± 10% with 9-180 μμfd., trimmer in parallel).
C9, C10, C12, C29, C35—0.001 μfd., mica.
C11, C13—150 μμfd., mica.
C14, C15, C16—0.005 μfd., mica.
C17, C18—0.001 μfd., ± 1% mica.
C19, C20—150 μμfd., ± 5% mica.
C21, C22, C26, C28, C31, C32, C33, C34—0.006 μfd., mica or ceramic.
C23—36 μμfd., ± 5% mica.
C24—3 μμfd., ceramic or mica.
C25—25 μμfd., mica.
C27, C36—380 μμfd., air variable.
C30—0.1 μfd., paper.
C37—0.25 μfd., paper.
S1—3 pole, 4 position wafer switch.
S2—d.p.d.t. wafer switch.
S3—relay contacts on voice control relay (see text).
T1—interstage transformer (special). Central Electronics Type 27AM-24.
T2, T3—single plate to voice coil (special). Central Electronics Type 27AO-79
(Note—T1, T2, T3 may be procured from Central Electronics, Chicago, Ill.)
Y1, Y2, Y3, Y4—1N34 Germanium diodes.

Fig. 3-2-A. Partial schematic of the "Multiphase" exciter.

(Courtesy Central Electronics)


XTAL MIKE
V1A ½ 12AX7
V1B ½ 12AX7
V2A ½ 12BH7
V2B ½ 12BH7
V3A 12AT7
V3B
V4 6BA7
V5 6AG7
R1 1 Meg .5W
R2 2.2K 1W
R3 100K 2W
R4 220K .5W
R5 2.2K 1W
R6 220K 2W
R11 10K 1W
R12 1Meg
R13 47K .5W
R14 1K 1W
C1 .005µfd 500V
C2 8µfd 450V
C3 .005µfd 500V
C4 250µµfd
TO VOICE CONTROL STAGES
T1
T2
T3
R15 400Ω 1% .5W
R16 1400Ω 1% .5W
AUDIO PHASE SHIFT NETWORK PS-1
R7 133.3K 1%
C5 2430µµfd
C6 4860 µµfd
R8 100K 1%
R9 100K 1%
C7 1215 µµfd
C8 607.5 µµfd
R10 133.3K 1%
90° AF
R17 560Ω 1W
R18 1K AF BALANCE
R19 1K 1W
S1
S1 POSITIONS
3 SB-1
2 SB-2
1 AM-PM
C9 .001µfd
C10 .001µfd
L1
L2
C11 150 µµfd
C13 150 µµfd
R20 560Ω 1W
C12 .001 µfd
XTAL 9.0 Mc
R21 1K 1W
R22 100K .5W
C14 .005 µfd
+300V
90° RF
Y1 1N34
Y2 1N34
Y3 1N34
Y4 1N34
L3 9.0Mc
RFC-1
C17 .001µfd ±1%
C18 .001µfd ±1%
C15 .005 µfd
C16 .005 µfd
R23 1K AM CARRIER BALANCE
R24 1K PM CARRIER BALANCE
C19 150 µµfd 5%
L4 9.0Mc
L5
C20 150 µµfd 5%
C21 .006 µfd
R25 1K .5W
R26 47K .5W
S3 VOICE CONTROL RELAY CONTACTS
C35 .001 µfd
TO VFO 3-8 VOLTS
R27 180Ω .5W
S2
C22 .006 µfd
C23 36µµf 5%
L7
R28 4.7K 2W
R29 15K 5W
R30 15K 5W
C24 3µµfd
C25 25µµfd
L6
C26 .006 µfd
R32 SEE TEXT
C27 380 µµfd
L8
R31 560Ω 1W
C28 .006 µfd
C29 .001 µfd
R33 56Ω 1W
RFC2
R34 22K .5W
C30 .1µfd
C31 .006µfd
C32 .006µfd
R35 1K .5W
R36 560Ω .5W
R39 10K .5W
R40 100K 1W
C37 .25µfd
-100V
C33 .006µfd
R38 SEE TEXT
L9
RFC-3
C34 .006µfd
R37 100Ω 1W
C36 380 µµfd
OUTPUT

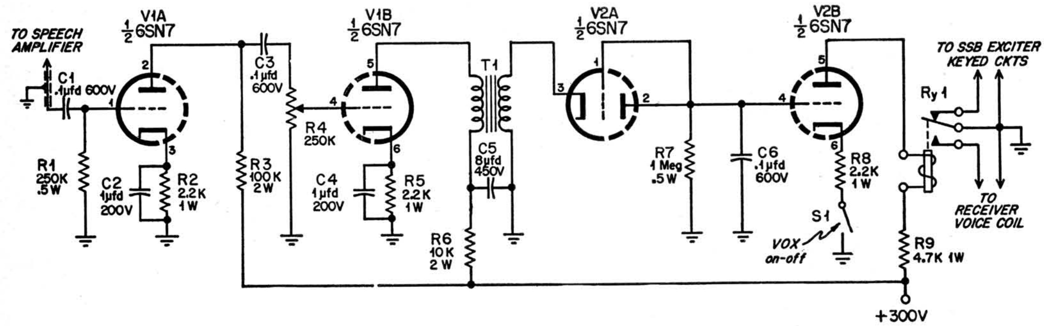


Fig. 3-3-A. Voice control circuit for use in conjunction with the "Multiphase" exciter, or the filter-type exciter described in Chapter II.

R1—250,000 ohm, ½w.
R2, R5, R8—2200 ohm. 1w.
R3—100,000 ohm, 2w.
R4—250,000 ohm potentiometer.
R6—10,000 ohm, 2w.
R7—1 megohm, ½w.
R9—4700 ohm, 1w.
C1, C3, C6—0.1 µfd., 600v., metalized paper.
C2, C4—1.0 µfd., 200v. metalized paper.
C5—8.0 µfd., 450v. electrolytic.
Ry1—5000 or 10,000 ohm vacuum-tube relay, s.p.d.t.
S1—s.p.s.t. toggle.
T1—interstage audio transformer. Single plate to single or push-pull grids.

will fall in or near the 7.2-Mc. phone band. These harmonics are not necessarily present in the v-f-o output, but are generated in the electron stream of the mixer stage. In view of this, the use of a 16.2-Mc. v.f.o. is recommended. This might take a little doing, but is not an impossible task. In any case, a frequency-stable v-f-o voltage of about 3 to 8 volts is needed.

Alignment

Adjust the 9.0-Mc. oscillator tank, *L1*, for oscillation of the crystal. With an appropriate mixing frequency fed into the v-f-o jack, and the receiver tuned to the desired mixture-output frequency, adjust *L3, L4, L5, L8* and *L9* for maximum output with one of the carrier-balance pots slightly off-balance.

Connect a oscilloscope to the output of the 6AG7 amplifier and use a recurrent sweep rate of about 30 per second (see *Chapter X*). Now, carefully balance *R23* and *R24* for as perfect carrier balance as possible. Feed a steady tone of about 1000 cps from an audio oscillator into the microphone jack. Make sure that the audio wave-form is good and that nothing is being overdriven. You will see on the oscilloscope a pattern that resembles a modulated AM envelope. Next, adjust the audio balance control, *R18*, and the r-f phase-shift network, *L2*, for *minimum* ripple (or modulation) on the oscilloscope pattern. The pattern for a properly aligned SSB exciter with single-tone input is a pure c-w envelope with no modulation. The presence of ripple indicates one of three things: (1) Presence of undesired sideband signal; (2) carrier unbalance; or, (3) bad wave-form in the input audio tone or distortion produced by overdriving the audio stages.

3.3—Automatic Voice Control Operation

Voice control operation may be accomplished by using the arrangement shown in *Fig. 3-3-A*. The theory of operation is quite simple. Some of the audio signal is taken from the speech amplifier ahead of the gain control (to insure independence of operation) and amplified up to a relatively high level in *V1a* and *V1b*. This audio is then rectified by the diode-connected half of the second 6SN7, *V2a*, and applied to the grid of the second-half of the tube, *V2b*. The plate current of *V2b* operates a high-impedance relay, which in turn operates the keyed circuits of the exciter.

The filter exciter described in *Chapter II* may be keyed in the cathode circuits of the two mixer stages, plus the cathode of the 2E26. Proper precautions should be taken that the cathode lines are "cold." Good by-passing at the tube socket will insure this. The phasing exciter should be keyed as shown by making *S3* the relay contacts.

Chapter IV

Linear Amplifier Theory

In the foregoing chapters we have shown how it is possible to generate single-sideband signals by either the filter or phasing method. The remaining problem is to amplify these signals to a high enough level so that they can hold their own in the presence of Ham band QRM.

Let us take a quick look at the quality of the SSB signal we have generated in these two exciters. Either one of the two units, when properly aligned and operating conservatively into a proper load, is capable of 40 db. attenuation of the undesired sideband—this is the intelligible stuff that we so painstakingly have filtered or phased out. There are also other sorts of signals that appear not only in the spectrum occupied by the undesired sideband but in the region of the transmitted sideband as well. These signals are the products of distortion in heterodyning and amplification of our SSB signal. These "distortion products" are not intelligible and are just so much garbage as far as conveying any sense to the distant receiver is concerned. The two SSB exciter units will have distortion products that are approximately 50 to 60 db. below the peak value of the transmitted sideband voltage. This is pretty darned clean. If we could just retain these attenuation ratios everything would be just dandy. Unfortunately, we can't and upon trying to do so hangs the tale that follows.

4.1—Linear Amplification

Let us look into the more important aspects of linear amplification. First, what is a linear amplifier? It is one that faithfully reproduces signal in the amplifier output circuit.
all of the amplitude variations of the input

What classes of amplifiers are linear amplifiers? What are some examples? Class A amplifiers are the best known of the lot. The average plate current remains constant, the grid voltage never (but never) swings into the positive region and the distortion products are so low as to be negligible. Examples? The r-f and i-f amplifiers in your receiver are class A. So are the microphone preamp and low-level stages in your modulator. The efficiency is quite low—the peak efficiency being of the order of 25% to 35%.

Class AB_1 amplifiers are somewhat like class A amplifiers in that the grid never swings positive and therefore never draws grid current. However, the average plate current will swing upward on peaks of input signal voltage. The efficiency is higher, maximum is approximately 55%, but the distortion products are worse—not much, but enough to be noticeable. Since the two classes of amplifiers discussed above do not draw grid current there is no grid driving power required. Actually, this is not true because we must supply the grid circuit and coupling circuit losses. Fortunately, this doesn't amount to much—a fraction of a watt, generally. We must, however, furnish a grid voltage swing from our driver sufficient to "get the show on the road."

A class AB_2 amplifier is a cross-breed of the AB_1 type and the full class B animal. The average plate current (as read on a plate current meter), will kick upward and the grid will swing into the positive voltage region for a portion of the excitation cycle and therefore draw grid current. The no-signal plate current is generally higher than that encountered in full class B stages.

The bias on the control grid is usually set so that the idling plate dissipation of the tube is approximately half of the maximum specified for the tube. When signal is applied the input goes up, there is power delivered to the output circuit, and the remaining power not lost in the output tank circuit or coupling circuit appears as the *operating* tube plate dissipation. This *operating* dissipation does not exceed the maximum rating—at least not for long. More on this later.

From the foregoing discussion I believe that you can sense that the class B amplifiers are the ones that get "horsed around" a bit. The plate current is run at a lower idling value by increasing the bias and the grid is usually driven farther into the positive voltage region, drawing more grid current, and requiring more grid driving power. The plate current swings over a greater range than do the previously mentioned classes of amplifiers. Correspondingly, the efficiency is higher (theoretical maximum is 78%) and the possibility of having more distortion products is greater.

A new single sideband operator often wonders just what "enough" sideband suppression might be in terms of decibels. Just how much should he have? Obviously, as much as possible, but practical limits dictate something on the order of the following: With an eye to the future and continued crowded band conditions the signal should be clean enough to permit working on the same carrier frequency but opposite sideband from another group of SSB stations. This necessitates SSB receivers and all. The signal should have *at least* 25 db. suppression at the minimum, and 30 db. should be attained if at all possible. This is not too much to expect of amateur stations and amateur techniques. If single sideband is going to be worth the trouble, it will certainly be worth the trouble to get the signal as clean as possible. After all—20 db. means that 1/10 of your signal voltage is appearing in the adjacent channel—that is 1/100 of the output power, and for a kilowatt that is an appreciable amount of watts. Remember—aim for 30 db. suppression. The main thing to keep in mind is *not* to push the system too hard even though there is a rare DX station on the loose.

Non-linear?—WHO, ME?

This term "distortion products" keeps popping up all the time. Just what are they? In conventional double-sideband AM we would call it *splatter*. Surely, I won't have to draw pictures to explain that particular point. How are they generated in a linear (so-called, that is) amplifier?

A true linear amplifier will produce output signals that are amplified replicas of the input grid signal. If the grid signal varies in amplitude between the limits of 1 to 2 volts, the output signal must vary between the limits of say 100 to 200 volts. As you can see we have a voltage gain of 100. But, suppose the output signal didn't quite make the grade and varies between 100 and 185 volts for the same 1 to 2-volt grid swing mentioned. That, my friend, is non-linearity. Going on with this thinking, assume that we have an r-f linear amplifier with one tube and a parallel-tuned output circuit tuned to the desired operating frequency. The non-linear condition will produce a fundamental-frequency plate current pulse to flow in the tank circuit and *also* harmonic signal currents—the 2nd harmonic, the 3rd, etc. If our tank circuits has a good operating *Q* the harmonic signals will not be transferred to the output circuit. However, if we have an input grid signal made up of more than one signal frequency—say a group of frequencies representing a human voice—something peculiar takes place. Each of these frequencies when passing through our *non*-linear amplifier will have harmonic frequencies generated that will be near 8 Mc., 12 Mc., 16 Mc., and so on if our fundamental operating frequency is at 4 Mc. As you might guess each of these harmonic frequencies will be slightly different from the others. We now are in a position to see how intermodulation distortion comes about.

At this point we had best use specific numbers and see how this works. Assume a suppressed-carrier frequency of 4000 kc. and generate a lower-sideband. Pick two audio input frequencies for convenience—say 1000 cps. and 2000 cps., which will net us two sideband signals at 3999 kc. and 3998 kc. respectively. Passing these through our *non-linear* "linear" amplifier will produce the following. The fundamental signals, 3999 kc. and 3998 kc. will certainly appear in the output.

Will the second harmonic, third harmonic, and so on appear? For all practical purpose, no. Not so fast, now. What are the figures for second harmonics? The second harmonic of 3999 is 7998 kc. and of 3998 is 7996 kc. The third harmonic of 3999 is 11,997 kc. and of 3998 is 11,994 kc. Since our amplifier is non-linear it is capable not only of amplification but also of *heterodyning*. You remember in *Chapter II* we said that any non-linear device could be used as a mixer or heterodyne device.

Keep in mind that any mixtures of the fundamental frequencies and *any* of the harmonic signals or mixtures among the harmonic signals themselves must fall near the 4000 kc. operating frequency to be of concern. All other combinations will be disposed of by the selective properties of the tuned circuit.

Case 1. Mixing the 2nd harmonic of 3999 which is 7998 kc. and fundamental signal 3998 kc.: 7998 plus 3998 = 11,996 kc. (will *not* appear in output).
7998 minus 3998 = 4000 kc. (*will* appear in output).

Case 2. Mixing the 2nd harmonic of 3998 kc. (7996 kc.) and the fundamental signal frequency of 3999 kc.:
7996 plus 3999 = 11,995 kc. (will *not* appear).
7996 minus 3999 = 3997 kc. (*will* appear).

Case 3. Mixing the 3rd harmonic of 3998 (11,994 kc.) and the 2nd harmonic of 3999 (7998 kc.): 11,994 plus 7998—(will *not* appear).
11,994 minus 7998 = 3996 kc. (*will* appear).

Case 4. Mixing the 3rd harmonic of 3999 (11,997 kc.) and the 2nd harmonic of 3998 (7996 kc.) 11,997 plus 7996—(*will not* appear).
11,997 minus 7996 = 4001 kc. (*will* appear).

We have considered only the 2nd and 3rd harmonics of the two signals fed into our amplifier. The 4th, 5th, and so on might be significant also, but what we have done so far will illustrate the point very well. We have fed only two frequencies into the input of the amplifier and *look* at what comes out to the antenna terminals! The original signals, 3999 kc. and 3998 kc., of course, are there. The following signals are also there: 4000 kc. (happens to be at the carrier frequency), 3997 kc., 3996 kc. *and* 4001 kc. All but the last one are at or below the carrier frequency, but the 4001 kc. product is in the *upper sideband*. This is in the region where we have tried so hard to keep things from happening. The ones that fall in the lower sideband where we transmit our intelligence will not really annoy us unless our signal really "stinks to high heaven."

4.2—SSB and TVI

Someone is bound to ask about the harmonic signals that are generated in the above process. Do they cause TVI? The answer is generally *no.* If the output tank circuit has a loaded resonant *Q* of from 12 to 15 there is very little danger of any appreciable harmonic energy being transferred to the antenna circuit.

Actually, the 2nd harmonic currents present in the output tank circuit of a class B amplifier are only about 3 db. less than those in the same amplifier operating in class C. However, the grid driving requirements of the class B case are so much less that it is generally conceded that here lies the secret of the TVI-less operation of the SSB transmitter.

Harmonic TVI, as you know, is the one that is hard to eliminate. The possibilities of front-end over-load are about the same as with any other transmitter of equivalent power. The SSB gang have a little saying concerning harmonics. It is, "If you don't generate 'em, you can't radiate 'em."

4.3—Linearity—How to Get It

The picture has been painted pretty black so far. We have seen what distortion products are and how they come about. The cure isn't really so unpleasant. It is just following good common sense. The following sections will deal with the different localized causes of distortion and how to overcome them.

Grid Circuit Distortion

This is probably where the greatest troubles are caused in the average amateur linear amplifier. These fall into three general classes:

1. *Grid bias troubles.* The bias may be too high causing the tube to operate on the nonlinear "knee" portion of its characteristic curve near cut-off. The bias might also be too low. This causes the idling tube plate dissipation to be excessive. A visual check will usually warn you of this condition. Where the exact value is not known a handy rule-of-thumb to use when adjusting bias on a linear amplifier is to adjust the no-signal plate input so that the plate dissipation of the final tube (or tubes) is *at least* 1/3 to 1/2 of the maximum rated tube plate dissipation. Another grid bias requirement is that the bias voltage *must remain constant* under all operating conditions. This dictates that there must be *no resistance* in the grid circuit whenever any grid current whatsoever is drawn. The bias supply, whether electronic or battery, must have no internal resistance—either use regulator tubes on a supply or a new battery. There is one obvious way to dodge this particular problem—that is to use zero-bias tubes. More about this later.

2. *Grid signal voltage regulation.* If the amplifier being considered never draws any grid current (as in class A or class AB_1) this is of no concern. However, if the grid at some time during the grid excitation cycle swings positive and draws grid current the load on the driver stage increases sharply and causes the grid signal voltage to drop from its otherwise no-load value. This trouble is primarily not in the grid circuit, but is dependent on the so-called "internal resistance" of the driver stage.

Back to our driver stage in the transmitter. To maintain the grid voltage during the periods of heavier loading (during grid current periods), we can lower the "internal resistance" of the stage by raising the tuned circuit *Q* (lower *L* to *C* ratio) along with some swamping of the driver stage plate circuit with a resistor but to keep things going, we will require more power from the driver stage. A healthy attitude to take about this matter is to plan on generating about four times the power you expect to use in driving the final amplifier stage and then swamp the remaining three quarters of the power with a resistor across the driver tank. Sure, this is wasteful, but in the long run is well worth the trouble.

3. *Grid drive.* Obviously, the grid can be over-driven and the stage goes into "saturation" —as some say. This over-driving will cause flattening of the peaks of the output wave-form

and produce distortion in large quantities. Conversely, the grid can be getting too little drive and the driving stage can be called upon to deliver more than it is capable of and distortion will be generated *in the driver, not* in the final.

Plate Circuit Distortion

The plate tank circuit is the gadget that transfers the signal energy from the final tube to the antenna circuit. It also performs another valuable function as we mentioned before. It takes the half-sine waves that the tube furnishes and through the flywheel effect of the resonant circuit supplies the missing half cycle of the r-f waveform. In order to do this the operating Q of the plate tank must be high enough so that the harmonics are attenuated, but not so high that the plate circuit efficiency falls off. The generally accepted limits of the loaded circuit Q are from 12 to 15.

Distortion is created in an amplifier when the loading is maladjusted. If the loading is too light, the amplifier will be driven into saturation much sooner than normal and the output power of course will be considerably reduced. If the loading is too heavy, the stage will not saturate easily, but the output power will be lower than that obtained at optimum coupling. Use some sort of output indicator (an r-f ammeter in the antenna, or field intensity meter), and adjust the coupling for maximum output for some high fixed value of input power.

The plate tank circuit will be discussed later in this chapter in the section dealing with design considerations.

4.4—What Tubes?

There are several schools of thought on the tube matter. There are the two basic divisions, tetrodes versus triodes. Then in the triode class there are again two main groups of adherents: (1) those who prefer the zero bias tubes with their freedom from biasing troubles and severe swamping requirements, and (2) those who prefer to use the low-mu triodes with their high bias requirements so that high power may be obtained without going into the grid current region at all

Each has its advantages and disadvantages. The zero-bias tubes will give higher stage efficiency (about 70%), there is no sharp transition from no-grid current to grid current region since it draws grid current as soon as even a small amount of grid excitation is applied. It will load the driver heavier at higher levels, but the sharp transient is missing and less swamping is needed than where a heavily biased tube is driven into the grid current region. The zero-bias stage is reasonably free of intermodulation distortion when properly operated.

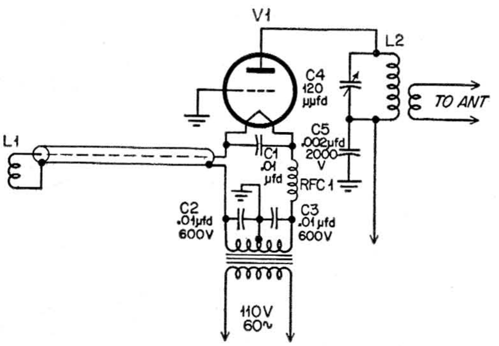
The low-mu tube operating in class AB1 requires practically no driving power, but will require a fairly high grid voltage swing in order to drive the tube to its full capabilities. The distortion products from this type of operation are very low—almost as low as class A operation. The only possibilities for trouble would be in the tube characteristic itself or in improper loading or drive as explained earlier. The stage efficiency is lower (50 to 55% but that is the price you pay for a cleaner signal. It is my own opinion that a clean signal is to be desired even at the expense of total output power.

There is of course the middle ground in this matter, the tube that requires some value of grid bias that is driven into the grid current region for just a portion of the grid excitation cycle. This is the baby that you have to swamp heavily to give good grid signal-voltage regulation. This "middle ground" case unfortunately is not a compromise from the standpoint of distortion. Generally, it has more severe distortion present in the output than either of the previously mentioned triode cases.

This brings us to the next major class of tubes—tetrodes and pentodes. These, too, have their advantages and disadvantages. Their chief advantage is in the low driving power requirements and relatively low grid signal voltage required for full output. One of the chief objections to using the larger tetrodes is the necessity of a stiff regulated screen power supply. Ordinary voltage regulator tubes (VR-150, VR-105, etc.) will be suitable for many tubes while others will require the full treatment. This means that an electronically regulated supply using 6L6's, 6Y6's, or 6AS7's must be constructed, or a shunt regulated arrangement using a small transmitting tube and dropping resistor from the plate supply as suggested by W2AZW.[43] The shape of the tube characteristic curves will probably have more effect on the distortion products in the output than anything else—if the circuit voltages are according to Hoyle. There are some tubes that just aren't suitable for use as linear amplifiers.

4.5—What Circuits?

Here again everyone has their own preference. We shall try to review briefly what the various possibilities are. The two obvious general classes of circuits used are: (1) grounded grid, and (2) grounded cathode amplifiers. The grounded-grid amplifiers have some very attractive features. There is generally no tuned circuit needed for the input of the stage. There is no neutralization needed in the case of triodes, therefore the plate tank circuit can be single-ended, that is, no split-stator condenser and center-tapped coil. If zero-bias tubes are used this further simplifies the problem. See *Fig 4-5-A* for an example of what can be done—



C1, C2, C3—0.01 μfd., 600v., ceramic.
C4—120 μμfd., single section variable, air.
C5—0.002 μfd., 2000v., mica.
RFC1—50 turns #16 enam. on 1" form.
V1—Any zero-bias tube.

Fig. 4-5-A. Grounded-grid linear amplifier, Zero-bias tubes, such as, 811, 805, etc., work out nicely in this arrangement.

note the minimum of parts needed. So much for the advantages, and now for the disadvantages.

Grounded-grid operation will require considerably more drive than use of the same tube in the conventional grounded cathode arrangement. This extra drive is not lost, however. It appears in the output circuit of the final amplifier as useful output. It is possible to get an *output power* from a grounded-grid stage that is greater than the d-c input to the plate circuit of the amplifier. This would make the *apparent* efficiency greater than 100%. As mentioned this is because the *driving* power appears as *output* power. Don't get any ideas now—the FCC has taken care of what you are thinking about with the amateur regulations regarding grounded-grid operation. Another point to keep in mind is that tetrodes cannot be used successfully in grounded-grid service because the presence of the screen grid will tend to make the circuit oscillate at signal frequency. Most pentodes also cannot be used for this reason because the isolation provided by the suppressor grid is not usually enough to prevent self-oscillation. Tetrodes and pentodes *can* be used in a "cathode driven" arrangement by not bypassing the screen and suppressor grids to ground—but to cathode. This is then not strictly speaking "grounded grid operation" but should be considered as a possible way to make use of some of the attractive power tetrodes such as the 4-250A or the 4-400. Still another point—never use a grounded-grid stage as a driver stage where the *driven* stage reflects a changing load back into the grounded-grid stage because this changing load will in turn be reflected back one stage more into the driver's driver. Confusing, isn't it? Confusing or not, the results are bad—more distortion than is healthy.

Again referring to *Fig. 4-5-A,* you will notice that a filament choke is necessary to keep the filament transformer capacity to ground from shunting the r-f driving voltage. The other alternative to using a choke or tuned circuit in the filament wiring is to procure a special low-capacity filament transformer that some of the surplus radar sets were blessed with.

It should be pointed out that the arrangement shown in *Fig. 4-5-A* will operate satisfactorily but the stage efficiency will be improved by using a tuned circuit in the cathode. This is necessary to provide a low-impedance path from ground to cathode for the harmonic components of the tube plate current. The cathode should be tapped down on this tuned circuit for the proper impedance match. See *Fig. 4-5-B* for modification.

Coming now to the more familiar grounded-cathode amplifiers we find circuits that we have been using for years in class C stages. See *Fig. 4-5-C.* There is really nothing new about the actual circuits. It is only the operating voltages on the control grid and the amounts and kind of driving signals that are different in linear amplifiers. The conditions that go to make a *good* class C amplifier are the same that help to make a good linear amplifier. By these I mean good tank circuit *Q*, freedom from oscillation and parasitic oscillations, good mechanical lay-out and construction, and so on. As outlined earlier the stage may be operated in various modes of linear operation. The driving power is generally modest and can be furnished by either of the exciters described in *Chapter II* or *Chapter III.*

4.6—Design Considerations

Much of the work in developing a design procedure has been done by Reque.[34] In this article, W2FZW tells how to design circuits and select operating voltages for tubes which no class B ratings are given.

If you will consult the tube manuals you will generally find that the audio ratings for a tube

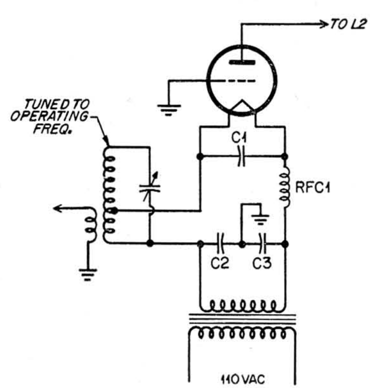
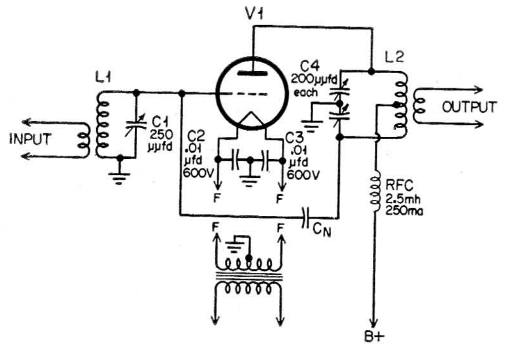


Fig. 4-5-B. Variation of the input coupling method shown in Fig. 4-5-A.



C1—250 $\mu\mu$fd., variable.
C2, C3—0.01 μfd., ceramic, 600v.
C4—200 $\mu\mu$fd. per section split-stator, transmitting type.
Cn—Neutralizing condenser, value depends on circuit and interelectrode tube capacity.
RFC—2.5 mh., 250 ma.
V1—Zero-bias tube, such as, 811, 805, etc.

Fig. 4-5-C. Conventional grounded cathode linear amplifier stage using a zero-bias tube.

are given for class A, AB_1, AB_2, and class B. We merely transfer our operation into the r-f realm and we are on our way—almost! In audio work we are accustomed to ordering a transformer that properly matches the tube grid impedance and likewise the plate impedance to whatever load we are using. We can't be quite as glib as this in r-f service. We have to choose our tuned circuits with discretion in order to accomplish the same end.

Let us take a specific example and work through it and end up with some answers that we can put to work for us. For the low-power boys let us choose the 6146 pentode that has made such a hit. *Table 4-6-A* is extracted from the RCA tube manual for the 6146 in audio service (two tubes).

Table 4-6-A tells quite a story. The triode connected ratings were listed just for your information. You can appreciate the advantages of pentode connected operation without any trouble.

You must now make up your mind as to the details of the final product. These include: number of tubes, series or parallel operation of tubes, class of operation. Let us say for the sake of argument that we want to use two 6146 tubes in some sort of linear amplifier.

Consulting *Table 4-6-A* we see that pentode connection in AB_1 will give use 120 watts output while AB_2 operation will give only 10 watts more—130 watts. Since AB_2 operation involves operating the tubes partially in the grid current region with all of its attendant troubles we would be wise to choose class AB_1 and sacrifice the 10 watts of output. Since we have chosen the class AB_1 mode of operation, the design of the grid tank circuit becomes no problem at all. Our only problem now is to furnish the grids with sufficient voltage to swing them throughout their full range—100 volts peak-to-peak for tubes in push-pull. If the exciter is a little puny on output and you think that getting the class AB_1 mode of operation, the design of approach. Employ link coupling between the exciter tank and the final grid circuit and put a higher *L* to *C* ratio tank circuit in the grid circuit than you have in the exciter output tank. This will give you a voltage step-up as in an ordinary transformer. The resonant impedance of the grid tank will be higher accordingly but this is perfectly o.k. since the grid draws no current and appears as a very high impedance itself. NOTE: The 100 volts grid swing needed was for push-pull operation. This means if only one tube is used the drive needed is 50 volts peak. The same is true if the tubes are operated in parallel—only 50 volts.

Next is the design of the plate tank circuit. Remember, we said that we wanted a loaded *Q* of between 12 and 15—keep this in mind. Every tube for a given set of operating voltages has a plate resistance that depends on the swing of the a-c plate signal voltage and the swing of the plate a-c signal current. We don't have to worry about calculating this value because the tube manufacturers have been kind enough to do it for us. In the case of the 6146 tubes in *push-pull* the plate load resistance is

Table 4-6-A

I C A S Ratings	Class AB1 (triode connected)	Class AB1 (pentode)	Class AB2 (pentode)
d-c plate voltage	400 v.	750 v.	750 v.
d-c grid #1 voltage	-100 v.	- 50 v.	-45 v.
peak AF grid-to-grid v.	200 v.	100 v.	101 v.
d-c grid #2 (screen) v.	-----	200 v.	165 v.
zero signal screen current	-----	1 ma.	0. 6 ma.
max. signal screen current	-----	27. 5 ma.	21 ma.
zero signal d-c plate current	80 ma.	57 ma.	35 ma.
max. signal d-c plate current	136 ma.	227 ma.	240 ma.
effective load (plate-to-plate)	8000 ohms	8000 ohms	8000 ohms
max. signal driving power	0 watts	0 watts	0. 03 watts
max. signal output (approx)	19 watts	120 watts	130 watts

8000 ohms. For one tube in r-f service, the proper plate load is 4000 ohms. Please note however, that the same tube in audio service with an untuned plate load would require a 2000-ohm load impedance. Personally, I don't have any use for the resulting half-sine wave output so let's forget the whole matter!

From these values of load resistance, the desired circuit Q (12 to 15) and the operating frequency we can arrive at the values of tank capacity and tank inductance to use. The formula to use is very simple and is as follows:

$$\text{Reactance (in ohms)} = \frac{\text{plate load resistance required}}{\text{loaded circuit } Q}$$

The reactance is the inductive reactance or capacitive reactance that the coil and condenser will have at resonance. NOTE: Inductive reactance is equal to capacitive reactance at resonance.

For our two tubes in push-pull:

$$\text{Reactance} = \frac{8000}{15} = 533 \text{ ohms}$$

With this value of reactance we substitute it in the following formulas:

$$C \text{ (in } \mu\text{fd.)} = \frac{1}{6.28 \times \text{freq.} \times X_c}$$

$$L \text{ (in microhenries)} = \frac{X_L}{6.28 \times \text{freq.}}$$

where freq. is in *megacycles*
and $X_c = X_L = 533$ ohms (already determined)

For the 4.0 megacycle band:

$$C = \frac{1}{6.28 \times 4.0 \times 533} = 0.000075\ \mu\text{fd.} \quad (\text{or } 75\ \mu\mu\text{fd.})$$

$$L = \frac{533}{6.28 \times 4.0} = 21.2 \text{ microhenries}$$

This means that we will have an *effective* tank capacity of 75 µµfd. or a split-stator condenser with 150 µµfd. per section *in use*. For the practical transmitter we would use a 220 µµfd. or a 260 µµfd. per section split-stator condenser in order to cover the entire band.

Someone is bound to ask "Why not use the tubes in parallel and avoid the use of a split-stator condenser and a center-tapped coil?" This idea is OK, but the appropriate changes must be made in our tank L and C values. Since the plate load resistance of a single 6146 tube is 4000 ohms, two tubes connected in parallel will yield 2000 ohms. You can stick this value in the above formulas, and you will find that the tank circuit capacity has increased by a factor of 4 now being 300 µµfd. instead of the 75 µµfd. for push-pull service.

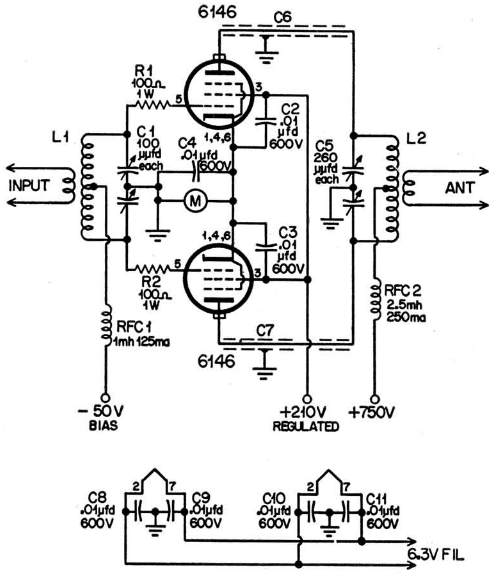
Parallel operation has likewise changed the size of our tank coil. It is now ¼ of its former inductance—now being 5.2 microhenries. This means that if you are in the habit of buying commercial wound coils, for the *parallel-connected* case you should purchase a 40-meter coil and use it on 80 meters. Even in the push-pull case turns must be removed from the standard 80-meter coil in order to attain the proper L to C ratio.

You are now forced to make a decision as to which to use—push-pull or parallel. It is a matter of personal preference—as the man says, "You pays your money and you takes your chances."

This business about the proper L and C values isn't just so much bunk. If the basic rules are followed the tubes will run cooler, the maximum power will be transferred to the antenna, the harmonics will be down—in other words things will be running more efficiently.

4.7—The 6146 Final Amplifier

Using the values we arrived at in the design part of this article, we are now in a position to build an amplifier. *Figure 4-7-A* shows the push-pull arrangement and *Fig. 4-7-B* shows the



C1—100 µµfd., per section split-stator.
C2, C3, C4, C8, C9, C10
C11—0.01 µfd., 600v., ceramic.
C5—260 µµfd., per section split-stator.
C6, C7—Coaxial condensers made of RG-58/U or RG-59/U with the shield grounds.
M—0-300 ma., d-c milliammeter.
R1, R2—100 ohms, 1w.
RFC1—1.0 mh., 125 ma., r-f choke.
RFC2—2.5 mh., 250 ma., r-f choke.

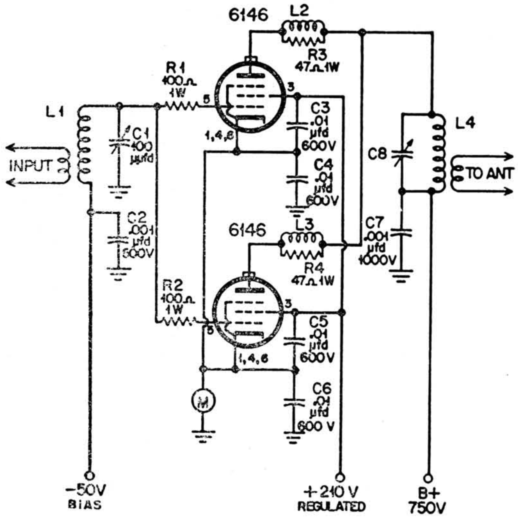
Fig. 4-7-B. Push-pull arrangement for class AB1 using the 6146 tubes. See next page for coil details.

parallel-connected circuit. It is up to the individual preference which circuit is to be used. Care should be taken to isolate the grid and plate circuits. The best policy is to keep the grid circuits below the chassis and the plate circuit components above the chassis. However, if plug-in coils are to be used, the grid tank coil should be isolated by a large chassis shield or totally enclosed in a shield can or box If these precautions are taken, neutralization should not be necessary. Use of plenty of ceramic bypass condensers is recommended wherever possible.

I know some of you are wondering about using the tubes at higher plate voltage, for example 1200 to 1500 volts. As explained in *Chapter I,* this will cause the peak power to be increased and because of the low duty-cycle of human speech, the average plate dissipation of the tubes will not be exceeded.

Using Higher Plate Voltages

There are certain precautions that must be observed when the plate voltage is raised, however. In the particular case in question, the screen voltage should be lowered to 150 volts (regulated, of course), and the grid bias voltage must be raised until the no-signal plate dissipation is again about half of rated maximum. It may be necessary to run the idling plate current above this value in order to get good linearity. If the idling dissipation is believed to be excessively high the amplifier grid bias should be controlled by a set of contacts on the voice con-



C1—100 µµfd., single-section air variable.
C2—0.001 µfd., 500v., mica.
C3, C4, C5, C6—0.01 µfd., 600v., ceramic.
C7—0.001 µfd., 1000v., mica.
C8—325 µµfd., single-section air variable.
R1, R2—100 ohms, 1w.
R3, R4—47 ohms, 1w.
M—0-300 ma., d-c milliammeter.

Fig. 4-7-B. Parallel-connected 6146 class ABI final amplifier. This stage is capable of 120 watts peak output. See box below for coil values.

Coil Table

Fig. 4-7-B

L1—Grid tank, end link.
80 Meters—B&W 80MEL
40 Meters—B&W 40MEL

L2, L3—Parasitic chokes. Ten turns of #22 enam. on R3 and R4.

L4—Plate tank.
80 Meters—B&W 40BEL
40 Meters—B&W 20BEL

Fig. 4-7-A

L1—Grid tank.
80 Meters—B&W 80MCL
40 Meters—B&W 40MCL

L2—Plate tank.
80 Meters—B&W 80BVL with turns removed until C5 is 2/3 meshed at 4.0 Mc.
40 Meters—B&W 40BVL with turns removed until C5 is 1/3 meshed at 7.3 Mc.

trol relay so that during periods of no speech input, the amplifier is biased to cut-off or at least well within safe dissipation limits. When operating the amplifier under these conditions, you will have to be especially careful not to abuse the tubes by whistling into the microphone for more than a *very* short time. Also, if carrier insertion is used, keep the stage operating at greatly reduced continuous levels so that the plates will not blush—not even a little! Two-tone tests cannot be generally made at full input except for very short periods because of the high average power involved. Under normal voice inputs, however, you will find that you can get approximately 400 watts peak sideband power output with 1500 volts on the plates. This is quite a signal for such a small package.

You now have enough information to set up a medium-power SSB station and to operate it. For those who want to go "whole hog" and develop a full kilowatt of single sideband *Chapter V* will give some pointers and work through the necessary steps for "the full treatment." Remember—one man's linear is another man's clear channel.

4.8—Two Tone Tests

You hear SSB operators talking glibly about "two-tone tests." They are referring to the simple test that can be performed on a SSB transmitting system to check the linearity of the amplifiers. We must have some yard stick with which to check our amplifiers that have so "carefully" been designed and constructed.

The idea is briefly this: Feed two steady sine-wave audio signals into the SSB exciter input. You can use two audio oscillators, or you can inject carrier and feed just one tone into the microphone input. The amplitude of these two signals should be kept equal for the tests. Those using the phasing exciter described in *Chapter*

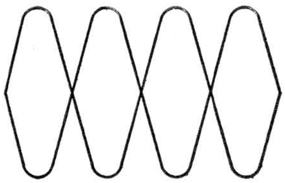


Fig. 4-8-A. Oscilloscope pattern showing the ideal envelope with a two-tone test in progress. No serious non-linearity present here.

III can feed one tone into the microphone input and put the function switch in the AM position and leave the carrier balanced out. This will produce a double-sideband "two-tone" output.

Now, what to look for. Connect a oscilloscope to the output of the last amplifier after

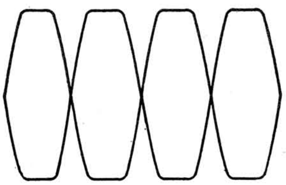


Fig. 4-8-B. In this "two-tone" test pattern there is peak flattening due to either excessive drive, poor regulation on the driver stage or insufficient antenna coupling.

approximating the proper loading with a *dummy load*—never into an antenna except for short tests. Set the sweep rate on the oscilloscope for 20 to 30 sweeps per second. If both signals of our two-tone test are equal we should see a pattern that resembles that in *Fig. 4-8-A.* This is the way the pattern should look if everything is operating properly. Increase the level of the two tones until the system starts to overload—that is evident by flattening of the peaks of the pattern as shown in *Fig. 4-8-B.* If this point of *maximum linear input* is considered less than the final amplifier is capable of, you should determine if the loading on the final is heavy enough, or if the driver stage is over-loading before the final reaches its maximum grid driving requirements. A quick test for light loading is to throw the final plate tank circuit slightly off resonance and watch the pattern. If the flattening disappears as the oscilloscope pattern decreases slightly in amplitude the final is not loaded heavily enough and the proper steps must be taken to increase the antenna coupling. However, if the flattening remains the trouble is in the driver or the coupling arrangement between driver and final. Slightly tighter coupling between stages or at least more efficient coupling must be accomplished.

If your oscilloscope pattern looks like *Fig. 4-8-C,* your trouble is something else. Your grid bias is too high and must be reduced until the two sine wave patterns cross the center-line with perfect sharp "X" patterns. This type of distortion is present at *all levels* of signal input and has been nicknamed "cross-over distortion." Reducing the gain when operating the transmitter on the air will do little to clean up this latter

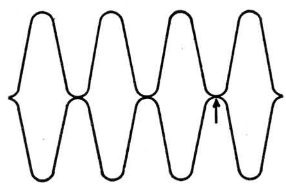


Fig. 4-8-C. Sloppy cross-over characteristics are seen in this "two-tone" test pattern. The grid bias should be reduced until the cross-over (shown at the arrow) is a sharp "X" as seen in Fig. 4-8-A.

type of distortion. Gain reduction would, however, help the peak-flattening distortion mentioned first. Don't reduce your operating bias to the point where your tubes are dissipating more than their ratings call for.

This is by no means the last word on the two-tone test. For more extensive tests and information I recommend that you read the fine pair of articles by Long[21] and Ehrlich.[12]

Hints for Better SSB Operation

An Antenna T-R Switch

A T-R switch is a device that permits using the same antenna for transmitting and receiving without the use of an antenna change-over relay. It must protect the receiver input coil from being burned up by getting too much r.f. from the transmission line when the transmitter is operating. The T-R switch must also not affect the line impedance of the antenna feed line when the transmitter is operating.

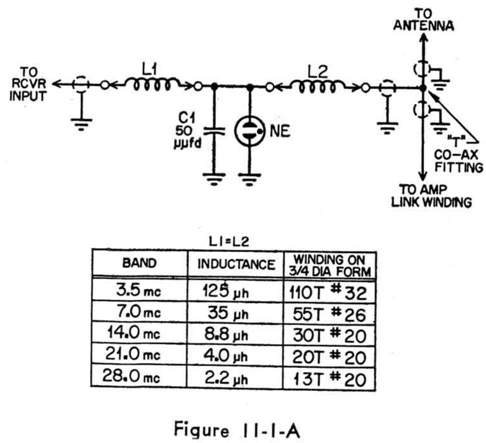
Such a device is shown in *Fig. 11-1-A.* It operates on the principle that a quarter-wave transmission line with a short across one end will appear as an open circuit at the other end. Instead of using actual transmission line for the purpose, an artificial transmission line composed of *L2* in conjunction with *C1* operates as one-quarter wavelength of line while *L1* and *C1* make up another section of quarter-wavelength line. The two coils *L1* and *L2* are identical so that the net result is that there is the equivalent of *one-half* wavelength of artificial line between the antenna and the receiver terminals. One axiom of transmission lines is that whatever impedance appears across the input terminals of a half-wavelength transmission line will also be present at the output terminals of the line. Thus the receiver sees the normal impedance of the antenna transmission line when it is connected to the receiver terminals of the T-R switch. One condition for proper operation has now been met.

When the transmitter operates, the resonant circuit of *L2* and *C1* will produce a high r-f voltage across *C1* because the *L* to *C* ratio of the tuned circuit is very high. If the value of this r-f voltage exceeds the ignition voltage of the neon bulb connected across *C1* the bulb will light. Because of the voltage regulation properties of the gas filled neon bulb the voltage will be limited to about 60 volts maximum value. If the bulb were not present, the peak value of voltage would rise to several hundred volts for a one kilowatt transmitter. The neon bulb creates the effect of a short circuit across *C1* as long as it stays ignited. This short circuit at the end of the one-quarter wavelength line composed of *L2* and *C1* appears as an open circuit when viewed from the antenna transmission line end of the unit. Thus the antenna is unaffected by the operation of the T-R switch. The 60 volts appearing across *C1,* when transformed down through the action of the left-hand quarter-wavelength line composed of *L1* and *C1,* appears as a very low voltage at the receiver terminals.

As already mentioned the *L* to *C* ratio of the system must be very high in order for the voltage appearing across *C1* to be high enough to fire the neon bulb as soon as possible.

The coil data for the amateur bands of 80 through 10 meters is shown in *Fig. 11-1-A.* For all the bands the unit should be peaked by adjusting *C1* for a maximum receiver background noise or for maximum signal when tuned to a station. When changing bands with the receiver it will be noticed that the receiver will have little sensitivity unless the coils in the T-R switch are also changed to the proper set.

The 2-watt neon bulb used must not have a series resistor in its base. If there is any question in the builder's mind regarding this point the base should be removed carefully and the unit examined. A pan of boiling water will usually serve to soften the cement used to fasten the metal base to the glass bulb.



L1=L2

BAND	INDUCTANCE	WINDING ON 3/4 DIA FORM
3.5 mc	125 µh	110T #32
7.0 mc	35 µh	55T #26
14.0 mc	8.8 µh	30T #20
21.0 mc	4.0 µh	20T #20
28.0 mc	2.2 µh	13T #20

Figure 11-1-A

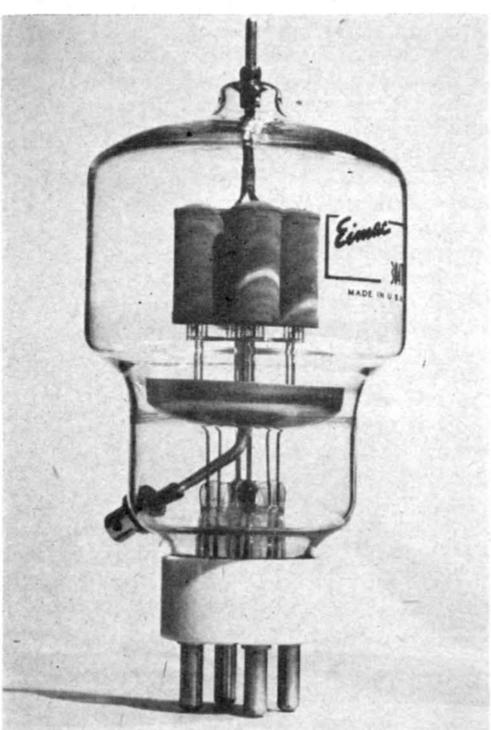
The bulbs appeared to stand up well when using a full kilowatt of SSB, however, it is not advised that the unit be used for a continuous type of service where the average power is in the neighborhood of one kilowatt input, such as AM or radio-teletype. Under these conditions the continuous power will be too high for the 2-watt size bulb. In this type of service a change-over relay is advised since the change from transmit to receive does not take place as frequently as with voice-controlled SSB operation.

Chapter V

High Power Linear Finals

In this chapter I would like to describe some of the aspects of designing high-powered linear amplifiers and also the practical slant on the subject. As it always turns out, the manufacturers never seem to build just the exact coil or condenser you dream up on the design pad. Obviously, they cannot predict all the desired values, so the amateur must make intelligent substitutions in order to preserve the original design ideals.

The amplifier to be described uses the well-known *Eimac* 304TL that gathers dust in many attics. The 304TL is a low-mu, high transconductance triode, that falls into the more desirable class of linear amplifier tubes mentioned in *Chapter IV*. To review briefly: The low-mu tubes require a high d-c grid bias and accordingly a high grid signal voltage swing to afford economical utilization. However, if the positive grid operating region is avoided, no grid driving power is required and at the same time a respectable output is obtained. Also, if no grid current is drawn the foremost cause of distortion products is avoided—poor grid signal voltage regulation.



The EIMAC 304TL is an ideal low-mu triode power amplifier for use in a linear stage.

We must make up our minds before we start just what we want to end up with. Do we want to drive the stage into Class B operation by going into the positive grid current region? Or should we avoid grid current? If we choose to stay out of the grid current region, our grid tank circuit can assume almost any proportions of *L* and *C*, but once we have done this we must never draw grid current or we will really be in hot water because the grid-voltage regulation will be poor. On the other hand if we resign ourselves to operation in the grid current region for at least a portion of the excitation-voltage swing, and design our tank accordingly, we can operate in the no-grid-current region with a clear conscience. In the event of an occasional pulse of grid current the grid-signal voltage will hold up.

My advice? Do the latter—design the tank for grid current operation and play safe. There is a price that we have to pay for this peace of mind. In order to stabilize the grid signal voltage under load changes, we must swamp the grid tank circuit with a resistor which consumes a fair amount of power whether we are in the grid-current region or not.

The material that follows differs considerably from that originally published in *CQ Magazine*.

5.1—Plate Tank Circuit Design

For academic reasons we will list the pertinent information on the 304TL as listed in the *Eimac* characteristic sheet. (See *Table 5-1-A.*)

You will notice that this *Table* is for two-tube operation. The amplifier in question uses only one tube so appropriate changes must be made. The author's high-voltage supply is capable of 3000 volts with good regulation, so the design was worked out with 3000 volts d.c. in mind. Going down *Table 5-1-A* in order of listing, the circuit condition for single-tube operation should be:

Plate Voltage	— 3000 v.
D-c grid voltage	— negative 290 v.
Zero sig. Ip	— 65 ma.
Max. sig. Ip	— 400 ma.
Effective Load	— 4550 ohms
Peak a-f grid voltage	— 390 v.
Max. sig. peak driving pwr.	— 55 w.
Max. sig. plate power output	— 900 w.

You may also note that this extracted tube data differs from that published in *CQ Magazine* in that the above data shows greatly increased grid driving requirements and correspondingly higher output power. At the time of writing the earlier material, the author had an obsolete data sheet. Since then I have been informed by the *Eimac Co.* that the method of

CLASS AB2 AUDIO AMPLIFIER OPERATION 2 TUBES, TYPICAL OPERATION			
DC PLATE VOLTAGE	1500	2000	3000
DC GRID VOLTAGE	−118	−170	−290
ZERO SIG. DC PLATE CURRENT (MA)	270	200	130
MAX SIG. DC PLATE CURRENT (MA)	1140	1000	800
EFFECTIVE LOAD, OHMS, PLATE TO PLATE	2750	4500	9100
PEAK AF GRID INPUT VOLTAGE (PER TUBE)	245	290	390
MAX SIG. PEAK DRIVING POWER (W)	78	87	110
MAX SIG. PLATE POWER OUTPUT (W)	1100	1400	1800

Table 5-1-A

rating the tubes for linear service has been changed. They now indicate maximum signal peak driving power and voltages which are more in line with what is needed for driving impedance calculations.

We will assume a loaded plate tank circuit Q of 15. This Q is the ratio of the load impedance to the reactance of either the tank condenser or tank coil since the two reactances are equal at resonance. In formula form:

$$Q = \frac{Rl}{X} \qquad 1.$$

where we know Q and Rl (equal to 4550 ohms). The thing we want to know is the reactance X so rearranging we obtain

$$X = \frac{Rl}{Q} \qquad 2.$$

If we stick numbers in our formula:

$$X = \frac{4550 \text{ ohms}}{15} = 303 \text{ ohms} \qquad 3.$$

So for a Q of 15 our tank condenser and coil must have reactances equal to 303 ohms no matter what band we are operating on.

Some one will ask, "Why use a Q of 15?" It is like this: If we increase the loaded tank Q we will decrease the tank circuit efficiency so that the tank circuit losses become greater. However, if we decrease the loaded Q the stage efficiency will go up toward the theoretical limit of 78%, but the selective properties of the tank circuit will deteriorate so that harmonics of the fundamental signal will not be suppressed. From experience it has been found that a loaded Q of from 12 to 15 is satisfactory. If you must vary from these limits, my advice is to err toward higher values of loaded Q even at the expense of a few watts output.

Next in our figuring is to get some condenser sizes: Effective plate to cathode capacity

$$C = \frac{1}{6.28 \times f \times X} \qquad 4.$$

where f is the frequency in megacycles, X is the 303 ohms reactance, and the capacity C comes out in microfarads. Again plugging in numbers: (For 3.9 Mc. operation)

$$C = \frac{1}{6.28 \times 3.9 \times 303} \qquad 5.$$

$= 0.000135$ μfd. or 135 $\mu\mu$fd. See *Fig. 5-1-A* for the final arrangement of the plate tank circuit components. The calculated plate-to-cathode capacity is maintained while the circuit is arranged for plate neutralization.

We must now choose our tank coil. From our known value of effective tank capacity (67 $\mu\mu$fd.) and the frequency we can calculate the size of our coil. The formula to use is:

$$L = \frac{1}{39.5 \times f^2 \times C} \qquad 6.$$

where L is in microhenries, f is in megacycles and C is in microfarads. For our 80-meter band case where C is 67 $\mu\mu$fd., the inductance comes out to be approximately 23.3 microhenries. Consulting the catalog, the *Barker & Williamson* Ham-band coils run as follows: The *80TVH* coil has an inductance of approximately 43 microhenries and the *40TVH* coil has approximately 20 microhenries inductance. Our value of 23.3 uh. falls nearer to the 20 uh. value than the other, so we choose the *B&W 40TVH* tank coil. This compromise in choice will raise the loaded tank circuit Q slightly, but not enough to concern ourselves about.*

I'm sure that someone is going to bring up the fact that the *40TVH* coil is only rated at 500 watts. Very true, indeed. Don't forget this rating is for AM or CW transmitter use where the continuous power input might be on kilowatt. In our single sideband case, however, our power input is one kilowatt on peaks of the speech input and the average power (the thing that burns up components) is considerably

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* If the builder desires to use either **E. F. Johnson** or **Bud** plate tank coils the inductance values are as follows:

Bud type VLS80 28.5 μh.
Bud type VLS40 17.2 μh.
Johnson type 500HCS80 45 μh.
Johnson type 500LCS80 27 μh.
Johnson type 500HCS40 19 μh.
Johnson type 500LCS40 10 μh.

These values should also prove helpful in selecting coils for the higher frequency bands.

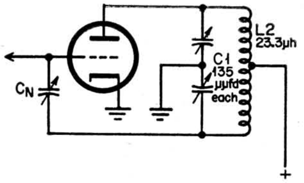


Fig. 5-1-A. Arrangement of tank tuning capacitor and inductance for 80-meter operation. The method of calculating the values is described in this text.

less—usually well below one-half kilowatt. If you make it a practice to operate CW or to insert carrier using this amplifier, remember to keep the maximum continuous input power below one-half kilowatt just so the coil does not melt down. Remember! Reduce the power input by reducing the grid drive not by reducing the antenna coupling. The latter will cause the r-f tank voltage to be abnormally high and hasten the coil's demise. If additional ventilation is provided, this one-half kilowatt limit can be increased somewhat.

How about the coils and condenser settings for the 40-meter and 20-meter bands? Running through the same calculations that we did for 80-meters except substituting the 7.2 Mc. where we had 3.9 Mc. previously, we came up with the following: Split-stator tank condenser setting—65 μμfd. per section; coil inductance required 15 microhenries. Again we are stuck with a compromise—the value desired is closer to the 20 μh. of the *40TVH* coil than the approximately 6 μh. of the *20TVH* coil, so again we can use the *40TVH* coil for 40-meter operation. This will lower our loaded *Q* to our self-imposed lower limit of approximately 12—which is still a safe value.

AMPLIFIER PLATE TANK CIRCUIT		
AMATEUR BAND	EFFECTIVE TANK CAPACITY (PER SECTION)	RECOMMENDED B & W COIL
80	135 μμfd & PADDERS (SEE TEXT)	40 TVH
40	65 μμfd	40 TVH
20	33 μμfd	20 TVH

Table 5-1-B

The 20-meter band parameters come out as follows: Per section tank capacity 33 μμfd.; coil inductance required—7.6 μh. The *20TVH* coil has an inductance of 6 μh., so that is the one to use. Again our loaded *Q* is just a little above what we decided upon, but within reason. *Table 5-1-B* gives these figures in a little more available form.

Grid Tank Circuit Design

Now that we have the plate tank circuit out of the way we can concentrate on the grid tank circuit. The author's particular line-up used an 829B in the output of the exciter to drive the final. The 829B had 500 volts plate supply voltage and is capable of delivering about 100 watts peak power output. This is obviously a lot more than necessary to drive the 304TL amplifier under even the worst of conditions.

The philosophy of generating more driving power than is actually required is one that I recommend all SSB neophytes follow. This philosophy is not only good in SSB work but in CW and AM as well, but the fruits are more apparent on the distant receiving end in the case of SSB. Cleaner, distortion-free signals will be the result.

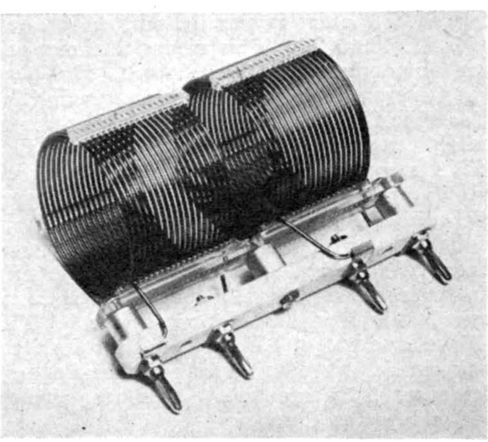
Consulting our extracted tube-sheet (*Table 5-1-A*) characteristics again, we see that there are three items that concern our grid circuit: the d-c grid bias, peak grid voltage, and grid driving power. For the chosen plate voltage, the grid bias is -290 volts. This can be furnished from any negative supply with 300 volts of voltage-regulator tubes hung across its output. This is what was done in the transmitter described and is a rock solid supply. The author finds that the 304TL tubes vary considerably from tube to tube and on some tubes it is necessary to decrease the d-c grid bias to 255 volts to properly adjust the idling plate current to a value where good linearity is obtained. Don't be disturbed by this; just plug in another type of VR-tube and forget it.

To get to the actual tuned circuit we proceed along the same general lines that we did in the design of the plate tank circuit.

We must find what resistance is represented by the grid when it takes driving power and swings into the grid-current region of operation. The equivalent grid resistance is:

$$R(eq) = \frac{\textit{Peak Grid Voltage}}{\textit{Peak Fundamental Grid I}} \qquad 7.$$

where the *Peak Grid Voltage* is 390 volts (from *Table 5-1-A*) and the *Peak Fundamental Grid Current* must be determined graphically from the tube curves. *Eimac* furnishes transparent overlays and brief instructions on how this is



An example of a coil that will handle the peak powers involved in the circuits described in this chapter is the E. F. Johnson 500HCS80.

done. A few minutes spent with this procedure reveals that this value of fundamental-frequency peak-current is approximately 50 ma. This is *not* to be confused with what is to be expected on the grid current meter while operating. Substituting numbers in our little formula we find that *R(eq)* is equal to 7500 ohms. This is the equivalent grid driving resistance for the circuit conditions set up. Unfortunately, this does not remain constant during the complete driving voltage cycle because there is a fair share of the time when the tube is drawing no grid current at all. When no grid current is drawn the tube grid looks like a resistor with almost infinite resistance. As soon as grid current is drawn this infinite resistor lowers its value to as low as the calculated 7500 ohms. This is quite a large relative change. In order to bring this resistance excursion more within reasonable limits, we must lower the no-grid current resistance by placing a grid swamping resistor across the grid tank coil. The exact amount to put across the coil is a matter of debate among many of the SSB

AMATEUR BAND (METERS)	GRID TUNING CAP. (μμfds)	B &W COIL IN USE
80	204 μμfd	40 MEL
40	110 μμfd	20 MEL
20	56 μμfd	10 MEL

Table 5-1-C

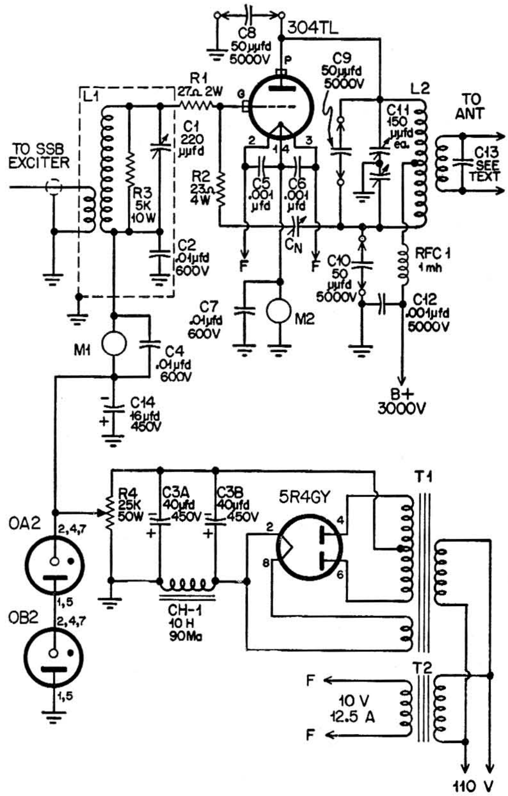
operators. Being on the ultra-conservative side, I take the attitude that the resistor should be lowered in value until you notice a shortage of driving power. With plenty of driving power available in the first place this usually ends up by being quite a low value. In the case of this amplifier, a value of 5000 ohms was chosen.

Let us look again at what happens when the tube swings into the grid current region. The grid resistance for no grid current is 5000 ohms and for the maximum grid current condition is 7500 ohms in parallel with 5000 ohms. This is 3000 ohms. This means that our grid resistance swings from 5000 down to 3000 ohms for the extreme conditions encountered. This is considerably better than swinging from near infinity down to 7500 ohms.

We must now design our tuned circuit arund this 3000-ohm value of grid resistance. We again choose circuit *Q* of 15 (values as high as 20 or 25 are not out of line for grid tanks). As before, the reactance of the grid tank components for all frequencies is calculated by dividing the resistance by the *Q*. This will be equal to 200 ohms. Just as for the plate tank circuit, calculate the size of the grid tank condenser by plugging the numbers already known into the formula:

$$C = \frac{1}{6.28 \times f \times X} \qquad 8.$$

In this case the *X* is 200 ohms and the *f* is the operating frequency in megacycles and the *C* is in microfarads. For 3.9 Mc. operation the required tank capacity comes out to 204 μμfd.; for 7.2-Mc. band operation—110 μμfd.; and 14.2-Mc. operation—56 μμfd. Consulting the information available on coils, the *B&W 40MEL* coil would be suitable for tuning to the 80-meter band, and the *20MEL* coil for 40-meter operation and the *10MEL* for 20-meter operation. *Table 5-1-C* gives this information in a



R1—27 ohms, 2w.
R2—23 ohms, 4w (2 parallel 47-ohm 2w.
R3—5000 ohm, 10w. non-inductive.
R4—25,000 ohm, 50w, with slider.
C1—250 μμfd., single section, air.
C2, C4, C7—0.01 μfd., 600v., disc ceramic.
C3a, C3b—dual, 40 μfd., 450v., electrolytic (insulated from chassis).
C5, C6—0.001 μfd., 600v., mica.
C8, C9, C10—50 μμfd., 5000v., vacuum padder (used only on 80 meters).
C11—150 μμfd., per section, split-stator transmitting, air, variable.
C12—0.001 μfd., 5000v., mica.
C13—600v mica; for 80 meters—1200 μμfd., for 40 meters—600 μμfd., for 20 meters—300 μμfd.
C14—16 μfd., 450v., electrolytic.
CN—neutralizing condenser, 12 μμfd., maximum capacity.
RFC1—1.0 mh., 500 ma. r-f choke.
CH1—10 h., 90 ma., filter choke.
M1—0-15 milliammeter.
M2—0-500 milliammeter.
T1—350-0-350v at 90 ma., 5.0v at 3 amp.

Fig. 5-1-B. Kilowatt linear amplifier with bias power supply schematic and parts list.

more orderly fashion. It would appear that a 250 $\mu\mu$fd. single-section condenser would be suitable for use in the grid circuit.

Circuit Details

Refer to *Fig. 5-1-B* for the following comments. The circuit is conventional in all but a very few respects. The filament circuit is wired for 10-volt operation. This was done because the shack transformer stock furnished a 10-volt, 13-ampere transformer instead of a 5-volt, 25-ampere unit. The center-tap of the tube-filament connections was used as the cathode return point. Both sides of the filament voltage line were bypassed to ground *(C5* and *C6)* at the tube socket with good quality mica condensers. The plate-current meter, or more strictly, the cathode-current meter is connected from the tube-filament center-tap point to ground with appropriate r-f bypassing. The parasitic resistors *R1* and *R2* are in the grid circuit and the grid-neutralizing leads respectively. These values are not critical but non-inductive resistors are mandatory.

The 23-ohm *R2* resistor should be at least four-watts in size for the following reason: During tune-up periods if the antenna loading should be light and the final running at resonance with grid excitation, the circulating r-f currents in the output tank circuit and the neutralizing path back to the grid run very high. This condition doesn't have to exist long before resistor *R2* goes up in smoke. You can not get along without these resistors. An amplifier without these will take off and be over the hill before you can even think about hitting the plate-voltage switch. As one fellow puts it: "One reason I think SSB is free of TVI is because you are compelled to clean up all traces of parasitic oscillations and fundamental instability before you can ever get the transmitter on the air." I suspect there is more truth than poetry in that statement.

The plate tank circuit is conventional with possibly the exception that on the 80-meter band vacuum padder condensers had to be added to reach the necessary capacity to enable tuning the *B&W 40TVH* coil to resonance over the entire 3.5 to 4.0 Mc. band. It was necessary to use two 50-$\mu\mu$fd. condensers connected in "split-stator"—that is, from each end of the coil to ground *(C8-C10)*, and one additional 50-$\mu\mu$fd. padder across the entire coil *(C9)*. If it is possible to place all padding condensers in balanced fashion (split-stator) it should be done, but for the rather obvious reason of parts shortage it was necessary to do as shown. With the padding shown it is possible to cover the entire 3.5 to 4.0-Mc. band with the variable 150-$\mu\mu$fd. per-section tank condenser. The 50-$\mu\mu$fd. padders used were the 5-kilovolt vacuum padders that have been available on surplus in the antenna tuning units for *ARC-5* series aircraft transmitters.

The output circuit link coil you will notice in *Fig. 5-1-B* has a fixed mica condenser, *C13*, across it. This has been found necessary in order to tune the link to resonance at the particular frequency in use. It is a well-known fact that even though the transmission line is perfectly "flat" difficulty can be encountered in properly loading a given final amplifier. For proper loading without the additional capacitor across the link, the link-coil inductive reactance should approximate the characteristic impedance of the transmission line. This is rarely true, however, so a corrective measure must be taken. Ordinary 600-volt "postage-stamp" mica condensers work well in this place. One precaution—*never* operate the amplifier with the antenna disconnected and the condenser *C13* fastened across the link. It will pop before you know what is happening. The same tuning result can be accomplished by a small variable condenser in series with the center lead of the co-ax line at the link coil. About 50 $\mu\mu$fd. to 100 $\mu\mu$fd. maximum capacity would be required.

The Plate Power Supply

Building plate-power supplies for variable-load applications such as single sideband, c-w operation, or for class-B modulator stages is always a problem. The sudden application of load upon a high-voltage supply will usually create a transient response effect upon the d-c voltage. The transient will take the shape of a damped sine wave whose resonant frequency is determined by the *L* and *C* combination of the power-supply filter components. Norgaard has discussed this problem at some length in the 1954 issues of *G.E. Ham News*. He has shown with oscillograms just what takes place when average values of filter capacitance are used.

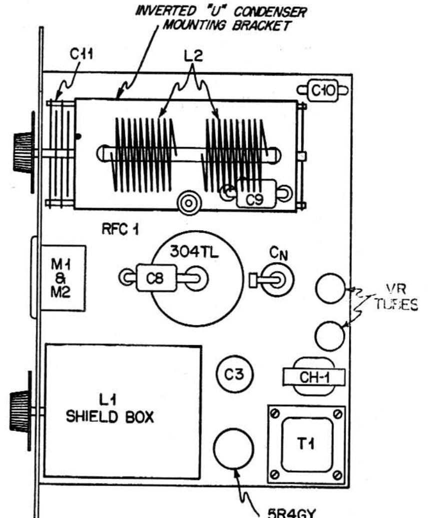


Fig. 5-1-C. Suggested layout plan for the 304TL stage.

Unfortunately, Norgaard's solution to the problem is not the most economical one that might be desired nor does he consider power supplies with voltage outputs in the region of 2500 to 3000 volts. He suggests using banks of series-parallel connected standard can-type 40-μfd. electrolytic condensers with equalizing resistors connected across each condenser unit. He recommended that a total output capacity of 90 μfd. be used. When considered from the standpoint of a 3000-volt supply, the cost and bulk are prohibitive. The author feels that a better way to handle the situation could be evolved—perhaps by using a saturable reactor in the primary of the plate transformer whose saturation characteristic would be controlled by the d-c load current. Possibly the response-time of such an arrangement could be made short enough to handle the transient as well as the steady-state regulation condition.

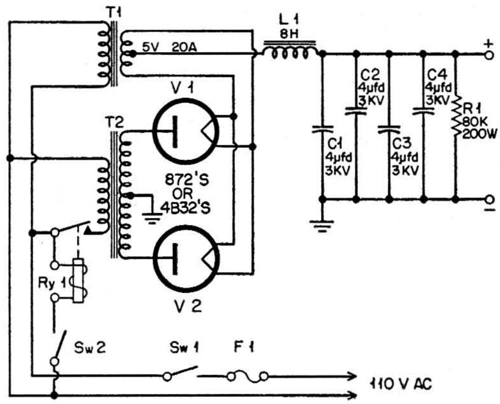
Since it is believed that most Hams will not be able to afford the condenser-swamping idea, the power supply shown in *Fig. 5-1-D* is recommended as a reasonable compromise for use with the amplifier shown in *Fig. 5-1-B*. If it is possible for the user to provide more output capacity in the filter than shown in *Fig. 5-1-D* the operation will be correspondingly better. The author's own plate supply uses 17 microfarads on the output filter condenser position and provides quite usable dynamic regulation.

Constructional Details

The following comments are intended to serve as a guide. *Figure 5-1-C* shows a sketch of the general lay-out of the amplifier. Use it, but be influenced by whatever components you have available and your own particular constructional preferences.

You will notice that the grid tank circuit is enclosed in a box. The author's experience with a previous 100TH linear amplifier prompted this measure. When grid current is drawn the sharp transients generated by crossing from the no-grid current to the grid-current condition cause harmonic frequencies to be generated capable of causing a mild case of TVI. As many Hams have found out, their class C finals cause TVI when only the final grids are being driven—no plate voltage necessary; in fact, the TVI usually decreases when the plate voltages is applied. This was the only TVI counter-measure that was taken in the amplifier—no low-pass filter, no shielding of the cabinet—and the TV set operates only ten feet away.

The plate coil, *(L2)*, was mounted on an inverted "U" shaped channel made out of light-gauge aluminum. This was fastened by the condenser frame screws to the variable tank capacitor so that the mounting channel was at ground potential. The coil jack-bar was then supported by ceramic stand-off insulators on the "U" shaped channel. It is a simple arrangement with plenty of mechanical strength.



C1, C2, C3, C4—4.0 μfd., 3000v, Plasticon AOC3M4.
F1—20 amp. fuse.
L1—8.0 henries, 500 ma., CTC Type R-105.
R1—80,000 ohms, 200 watts wirewound.
Ry1—Plate contactor, Potter & Brumfield PR-1A.
Sw1—Filament switch, SPST, 30 amp.
Sw2—High voltages switch, SPST, 3 amp. contacts.
T1—Filament transformer, 5.0v @ 20 amp., 10,000-volt insulation CTC Type F-520HB.
T2—Plate transformer, 3500-0-3500 volts @ 500 ma., CTC 3025.
V1, V2—Either 872 or 4B32.

Fig. 5-1-D. Suggested high voltage power supply for use with the linear final amplifier.

One of the vacuum padding condensers *(C8)* was mounted so that it connected directly to the heat-radiating plate cap of the tube. It was supported by the plate cap and a pillar of one-inch diameter "dural." This provides a short path for any stray harmonic currents that might exist.

The filament transformer was mounted under the chassis so that the high-current leads could be as short as possible.

Neutralization and Tune-Up

Once the amplifier is built the problem of adjustment arises. The stage operates like a triode amplifier. The neutralization procedure is conventional in that the initial adjustment should be made with the plate voltage off and the grid excitation applied. Tune the plate tank to resonance as indicated by an output indicator (lamp and loop, field intensity meter, etc). and adjust the neutralizing condenser *Cn* for minimum deflection on the indicator, meanwhile keeping the plate tank tuned to resonance after each adjustment of the neutralizing capacitor.

When this is completed, reconnect the plate voltage and using a dummy load, recheck the neutralizing by the "dynamic method." When the plate tank condenser is swung through resonance the plate current should reach a minimum reading at the same dial setting that the grid current maximum occurs. A little practice is necessary to be able to watch two meters simultaneously. The final adjustment shouldn't be far from that arrived at by the no-plate voltage adjustment. PRECAUTION! DON'T ADJUST THE NEUTRALIZING CONDENSER WHEN THE PLATE VOLTAGE IS ON!

Operation

You are now on your own. The operation is similar to most class C amplifiers in tune-up and shouldn't cause any raised eyebrows. A check on proper loading should be made with an antenna current meter or antenna field intensity meter to make sure that optimum coupling is obtained. For a given plate input power the link adjustment should be made for maximum power output. This adjustment should be made at some power level near maximum capabilities of the amplifier. I suggest that a linearity test be made using the two-tone test described in *Chapter IV*. I would also suggest that a oscilloscope be used to monitor the output of the amplifier to make sure that no "peak-limiting" distortion is present while operating. The amplifier is capable of delivering one-kilowatt "meter-average" on SSB and the peaks swing up in the neighborhood of two kilowatts. As explained in an earlier chapter this is perfectly legal—just keep your meters reading below one kilowatt.

A check was made of the output power and on 4.0 megacycles with a steady-state kilowatt input, the output power measured 550 watts into a dummy load. This is an efficiency of 55% which is below the 65% usually predicted for this type of service, but remember we have loaded our tank *Q* pretty high, so this probably accounts for the slight lowering of the output power. No grid current is drawn until the input exceeds 750 watts, so you can go a long way before you need be concerned with grid current.

The power required to drive the amplifier with the indicated grid swamping is between 15 and 20 watts peak. A swamping resistor of higher value could be used with appropriate changes in the grid task if less driving power is available. Remember, it's wise to generate more than you actually need—you won't be sorry.

Hints for Better SSB Operation

An Oscilloscope Monitoring Scheme

It is always a good idea to monitor the r-f output of a single sideband transmitter while operating. The most reliable method is to use an oscilloscope. The most common method in use is to put the r-f voltage on the vertical plates and use the oscilloscope internal sweep circuits for a horizontal signal. This gives an indication of the amplitude characteristic of the transmitter output. While this will give a reasonably good idea of what is going on, the ever-changing shape of output wave-form with human speech is a bit hard to watch.

The following scheme is presented as an alternate method of monitoring the output signal as related to the signal at some earlier point in the system. Since most of the distortion in a properly operated SSB system is generated in the high-powered output amplifier stage, the place to watch for non-linearity is obviously in the output amplifier.

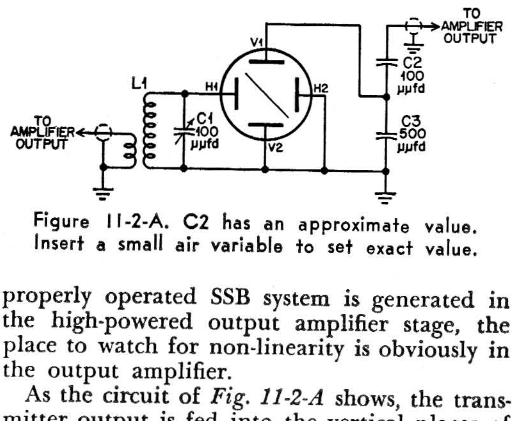


Figure 11-2-A. C2 has an approximate value. Insert a small air variable to set exact value.

As the circuit of *Fig. 11-2-A* shows, the transmitter output is fed into the vertical places of the oscilloscope through a capacitive voltage divider made up of *C2* and *C3*. Condenser *C3* can be of the ordinary postage-stamp mica type (600-volt rating) while *C2* is a variable air condenser. The variable condenser is adjusted to give the proper amount of deflection voltage to the vertical plates. The horizontal signal is derived from the final amplifier input circuit. The easiest method is to connect the link line of *L1* in parallel with the grid input coaxial line of the amplifier. *L1* and *C1* are obviously tuned to the operating frequency.

The patterns to be expected are shown in *Fig. 11-2-B*. For no input signal to the amplifier the idling oscilloscope picture is just the spot in the middle of the tube. For a perfect linear relation between the amplifier output and its input the pattern of *Fig. 11-2-B (a)* will be seen. This will be true for a single tone, inserted carrier, or for a complex waveform such as speech. *Figure 11-2-B (b)* shows the pattern if the grid bias is too high and the amplifier tubes are operating on the knee of the operating curve. The corrective measure is to decrease the grid bias voltage and run a higher value of idling current.

Figure 11-2-B (c) shows the pattern to be expected if peak limiting is taking place. This might be caused by insufficient amplifier loading or too much driving voltage to the amplifier grids. The corrective measures are obvious.

Figure 11-2-B (d) is the combination of the two troubles just described—grid bias too high and peak limiting. In all probability the pattern shown in *Fig. 11-2-B (e)* will be encountered when setting up the circuit. This is not a sign of improper transmitter operation, but that there is a phase shift between the two signals fed to the oscilloscope plates other than that of 180° normally encountered between the grid and plate voltages of an amplifier. Juggling the tuning capacitors will probably clean this trouble up, or the operator may find that it is necessary to place a resistor across one set of plates of the oscilloscope in order to arrive at a straight line pattern. A little tinkering will bring the desired result.

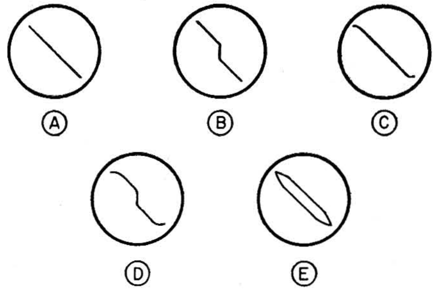


Figure 11-2-B

The author has tried this means of monitoring the transmitter output and found that it is more convenient to watch than the usual envelope pattern. It is also a more sensitive indication of amplifier non-linearity then the envelope presentation method.

Chapter VI

Balanced Modulators

Many amateurs when slightly exposed to some of the circuitry of single sideband exciters often complain that the circuits are too complex. They look at a schematic of an exciter and are appalled by such terms as filters, phase shift networks, and balanced modulators. There is nothing really so mysterious or difficult about the operation of any of these circuits that cannot be understood by anyone capable of passing the General Class of amateur examination.

If each of the stages in any transmitter is considered from the standpoint of its function and behavior the overall operation of the transmitter becomes much simpler. This applies to AM transmitters as well as SSB transmitters. Thus it is proposed that each of these new types of stages be treated separately, and in some detail, so that a more elementary understanding of their operation may be had. The first of these to be treated is the general class of mixers called balanced modulators.

We learned in *Chapter 2* that there are four terms that are synonymous. These were: heterodyne, mix, convert and modulate. Keep in mind that when one term is used that it means just the same as the other three. Just what is meant by mixing, heterodyning, conversion, and modulating? Mixing is the process whereby two signals of different frequencies are combined in some manner so as to produce new signals of still different frequencies from those of the originals.

6.1—Push-Pull Balanced Modulators

It would be best at this point if a specific example were taken and followed through. Consider the form of the balanced modulator shown in *Fig. 6-1-A.* At first glance this would appear to be a push-pull amplifier with some form of modulation applied to the center-tap of the grid coil. It may be considered as such except that the effects are slightly different than those normally encountered. The notations *f(m)* and *f(c)* denote the two frequencies that are fed into the balanced modulator stage—*f(c)* being what we shall call the carrier frequency and *f(m)* the modulating frequency.

For this example, assume that the modulating frequency, *f(m)*, is the range of speech frequencies from 0 to 3000 cycles per second and the carrier frequency, *f(c)*, is 1000 kilocycles. If the push-pull tank circuit in the plates of the triodes is tuned to a frequency of 1000 kilocycles the following conditions might exist: If no speech modulation signal, *f(m)*, were present and the two halves of the push-pull circuit were perfectly balanced, there would be no signal present in the output circuit of the stage. It can be seen that the carrier voltage, *f(c)*, is applied to the center-tap of the

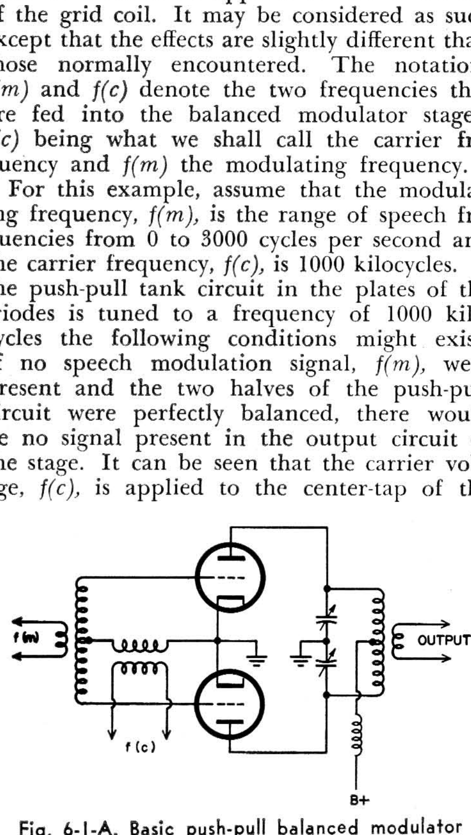
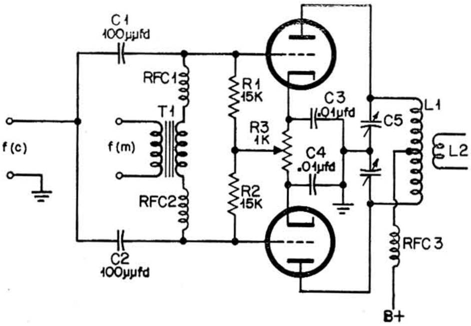


Fig. 6-1-A. Basic push-pull balanced modulator circuit. A practical adaptation of this idea is shown in Fig. 6-1-B.



C1, C2—100 μμfd., mica.
C3, C4—0.1 μfd., mica.
C5—Dual 140 μμfd., air variable.
L1, L2—Plate tank tuned to f(c).
R1, R2—15,000 ohms, ½w.
R3—1000-ohm, carrier balancing potentiometer.
RFC1, RFC2—0.5 mh., r-f choke.
RFC3—2.5 mh., r-f choke.
T1—Plate-to-grid audio transformer.

Fig. 6-1-B. A practical push-pull balanced modulator.

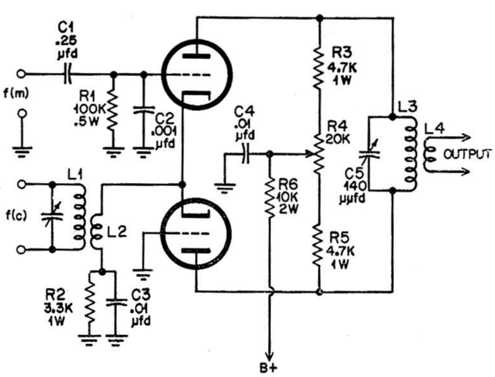
grid transformer and thus drives the two grids in parallel. If both grids swing positive simultaneously the individual plate currents of the two tubes will increase simultaneously. The effect in the plate tank circuit is that both pulses of plate current cancel each other out making the net result in the output link winding zero. Thus any variation in grid signal voltage that is applied to both tube grids in parallel will cancel out in the plate circuit due to the push-pull connection.

Consider now what would be the case if *only* the modulation frequency, *f(m)*, were applied to the stage and the carrier frequency, *f(c)*, were turned off for the moment. The speech signal is applied to the grids in push-pull connection and thus causes the tube plate currents to vary in a push-pull manner. Thus the plate current pulses of the triodes feed the tank circuit in the approved fashion of conventional push-pull amplifiers. There is one point, however, that changes the picture radically. The tuned plate circuit is tuned to 1000 kilocycles and not to 0 to 3000 cycles. Therefore, the plate tank circuit does not present any load impedance to the speech frequencies and the pulses of plate current merely scurry on to the B-plus supply causing no signal voltage to appear at the terminals of the output link coil. What has been accomplished thus far? By careful "balancing" we have eliminated the carrier signal in the output circuit, and by proper choice of tuned circuits used as plate loads we have also prevented the modulating frequency from appearing at the output terminals. Thus, when either signal is present in the stage by itself, there is no output signal.

What happens when both signals are applied to the stage at the same time? The carrier, *f(c)*, again feeds the grids in parallel, but the push-pull feed of the modulation causes one of the grids to swing less positive than the other by an amount equal to the instantaneous peak voltage of the speech modulating voltage. The stage is no longer perfectly balanced and a signal voltage appears across the 1000-kilocycle tuned circuit. One might think that this signal voltage is made up of pulses of carrier signal and nothing else. However, if the signal at the output terminals is observed on a spectrum analyzor, it can be seen that the output consists of signals grouped either side of the carrier frequency and that no carrier signal is present. We have thus generated new signals whose frequencies are equal to the carrier frequency *plus and minus* the modulating frequency. In other words we have created two sidebands (one on each side of 1000 kc.) and no carrier.

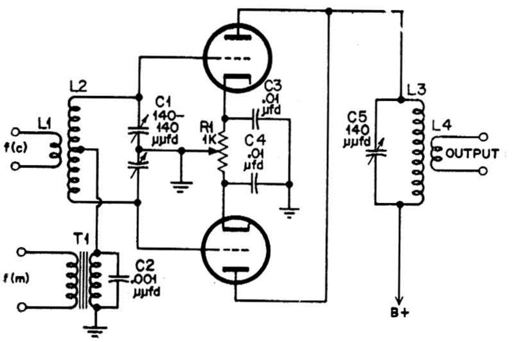
The amplitude relation of the sidebands generated to the amplitudes of the two parent signals, *f(c)* and *f(m)*, is such that the output signal is proportional to the product of the two input signals. In somewhat simpler terms, if the modulating frequency signal doubled in amplitude, the output signal would be twice as large; however, if both the carrier and the modulating signal doubled in their individual amplitudes the output signal would be two-times-two or four times as great as the original. In the applications that we will deal with, the carrier signal will remain constant and the modulating signal will vary in amplitude causing the output signal to be linearly related to the modulating signal's amplitude.

There is one circuit condition that has not been mentioned up to this point. The tubes



C1—0.25 μfd., paper.
C2—0.001 μfd., mica.
C3, C4—0.01 μfd., mica.
C5—140 μμfd., variable
L1/L2—Carrier oscillator tank and link winding.
L3/L4—Output tank circuit tuned to desired frequency.
R1—100,000 ohms, ½w.
R2—3300 ohms, 1w.
R3, R5—4700 ohms, 1w.
R4—20,000-ohm potentiometer.
R6—10,000 ohms, 2w.

Fig. 6-2-A. Modified push-pull balanced modulator.



C1—140 µµfd., dual air variable.
C2—0.001 µfd., mica.
C3, C4—0.01 µfd., mica.
C5—140 µµfd., variable.
L1/L2—Center tapped grid tank with link winding. L2 is tuned to f(c).
L3/L4—Plate tank and link tuned to f(c).
R1—1000-ohm carrier balancing potentiometer.
T1—Plate-to-grid audio transformer.

Fig. 6-3-A. A push-push balanced modulator with both the grid and tank circuits tuned to the same carrier frequency.

used in the balanced modulator must be operated on the non-linear portion of their grid voltage-plate current curve. This is quite contrary to the usual push-pull audio amplifier operation. Without this non-linear condition, the stage could not operate as described. Generally, in order to produce satisfactory results the carrier signal must be several times the amplitude of the modulating signal. This condition assures that the output signal will be linearly related to the modulation signal. The ratio of ten-to-one in carrier-to-signal amplitudes is generally accepted as the minimum value that will produce desirable results. Higher ratios are often desirable.

The circuit shown in *Fig. 6-1-A* is the basic circuit described. This circuit does not have any provisions for balancing the "almost-always" existing tube and circuit unbalanced conditions. *Figure 6-1-B* is the practical working circuit that should be used. The 1000-ohm cathode potentiometer, *R3,* provides the balancing adjustment necessary by adjusting the cathode biases on the two tubes. The modulating signal, *f(m),* is isolated from the carrier signal, *f(c),* by the pair of r-f chokes, *RFC1* and *RFC2*. This prevents the audio transformer, *T1* from loading down the source of r-f signal.

6.2—The Modified Push-Pull Circuit

Figure 6-2-A shows an interesting modification that is possible on the push-pull arrangement. The carrier signal, *f(c),* is fed into the two cathodes in parallel feed causing cancellation in the push-pull connected plate circuit. The two triodes are thus operated in grounded-grid operation for the carrier signal. The grid of the upper triode is at r-f ground potential because the reactance of condenser *C2* is low at the carrier frequency, *f(c)*.

The audio signal, *f(m),* is not fed into the two grids in push-pull as in *Fig. 6-1-A* but is fed to only one grid by an R/C coupled circuit. This causes only one grid to be unbalanced when speech is applied, but this is all that is necessary to accomplish the desired end This circuit also eliminates the need for the interstage audio transformer of *Fig. 6-1-B*.

6.3—The Push-Push Connected Circuit

Figure 6-3-A is very similar to the push-push frequency doubler that has been used for years. The plate tank circuit, however, is not tuned to twice the grid input frequency but is tuned to the fundamental of the grid r-f signal. It will be noted that the r-f carrier signal, *f(c),* is fed to the grids in push-pull while the modulating signal, *f(m),* is fed to the two grids in parallel through the grid-coil center tap.

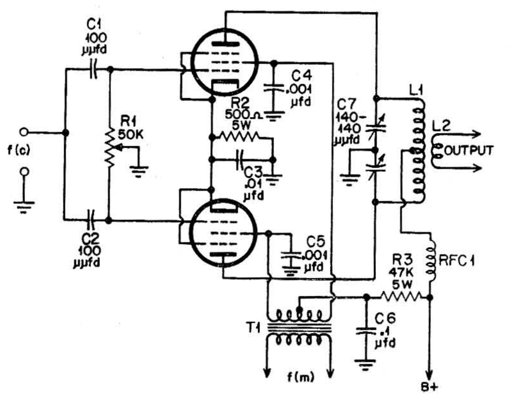
The fundamental frequency of *f(c)* will cancel out in the tank circuit because each pulse from the parallel-connected plates will oppose the "filled-in" half-cycle of the r-f wave form furnished by the fly-wheel effect of the tuned tank circuit. Thus, if the two tubes are perfectly balanced, no carrier signal will appear in the output circuit. It should be mentioned here that due to the push-push connection, the even harmonics (i.e., the 2nd, 4th, 6th, etc.) will tend to be accentuated and their appearance in the output will depend on the tuned circuit Q. (See the section on "Spurious Signal Generation.")

Since the modulating voltage is connected to the grids in parallel the plate current of the two parallel-connected plates will respond directly. However, again as in *Fig. 6-1-A* the plate tuned circuit will not respond to *f(m)* so it will not appear in the output. When *f(m)* and *f(c)* are simultaneously applied to the stage the result is the same as for the first type described, that is, a double-sideband, suppressed-carrier signal is generated in the plate circuit.

6.4—Multi-Grid Tubes in Balanced Modulator Circuits

The circuits described so far have made use of triodes only. The multi-grid tubes may also be used and the isolation offered by the separated grids makes them attractive for some applications.

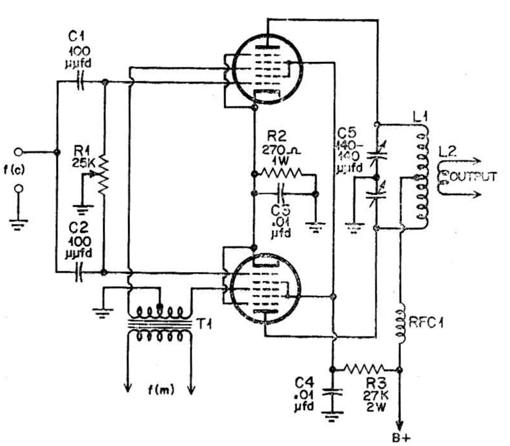
Basically the operation is the same as for the types already described. The r-f carrier is applied in parallel to the control grids in the case of the pentodes and to the oscillator injection grids in the case of the pentagrid converter tubes. When the stage is balanced the carrier is non-existent in the plate circuit. The modulating voltage is applied to the screen grid



C1, C2—100 μμfd., mica.
C3—0.01 μfd., mica.
C4, C5—0.001 μfd., mica.
C6—0.01 μfd., paper.
C7—Dual 140 μμfd., air variable.
L1/L2—Plate tank (center tapped) tuned to f(c).
R1—50,000-ohm carrier balancing potentiometer.
R2—500 ohms, 5w., wire wound.
R3—47,000 ohms, 5w., wire wound.
T1—Plate-to-5000 ohm load audio transformer.

Fig. 6-4-A. Pentode balanced modulator.

of the pentodes and to the signal grid of the pentagrid converter. Application of modulation disturbs the balance of the system by making one plate current vary in a different manner than the other. The balancing potentiometer control operates by varying the proportion of the excitation voltage that is applied to each grid. Since both ends of the potentiometer are at the r-f carrier potential, it must be located near the grid terminals of the tube. This may prove a disadvantage in some instances. The cathode balancing circuit of *Fig. 6-1-B* is always at zero r-f potential and may be located remotely to the balanced modulator stage.



C1, C2—100 μμfd., mica.
C3, C4—0.01 μfd., mica.
C5—Dual 140 μμfd., air variable.
L1/L2—Plate tank (center tapped) and link winding all tuned to f(c).
R1—25,000-ohm, balancing potentiometer.
R2—270 ohms, 1w.
R3—27,000 ohms, 2w.
RFC1—2.5 mh., r-f choke.
T1—Plate-to-push/pull grid audio transformer.

Fig. 6-4-B. Pentagrid converter balanced modulator.

Other Vacuum Tube Circuits

There are other balanced modulator circuits that can be built that utilize vacuum tubes. One principle must be kept in mind in order to achieve mixing action. This principle is: If the carrier signal is fed into the circuit in push-pull, the modulation must be fed in parallel connected, and if the carrier is in parallel the modulation must be in push-pull. This condition must be maintained if the signals are fed into the same pair of grids, as in *Fig. 6-1-A*, or into separate grids as *Fig. 6-4-A and 6-4-B*. Connecting both signals in the same mode will net zero mixing in the plate circuit. Actually, a slight amount of mixing might take place because of dissimilar dynamic tube characteristics.

6.5—Use of Varistors

The term "varistor" may be new to some readers. To dispel any confusion that might exist, a varistor is a group of usually two or four identical rectifying units. The individual units may be the familiar 1N34 germanium diode, a copper-oxide rectifier, a silicon diode, or even a selenium rectifier. The "varistor" group may be sealed in a shield can or glass envelope to protect them from damage.

The individual rectifier units of a varistor group are usually selected so that their characteristics are as closely matched to the other rectifier units in the varistor as possible. Thus a type 1N35 varistor is nothing more than a pair of carefully selected 1N34 germanium diodes.

There are three basic types of circuits in which varistors can be used as balanced modulators. These are: shunt, series, or the ring type. In order to achieve balancing action the varistors can be connected in various ways using either 2 or 4 diodes in the varistor unit. It is the various combinations of circuitry and varistor configurations that are of interest to us.

Shunt Connected Varistor Modulators

Figure 6-5-A shows the two possible connections using a four-unit varistor and a two-unit varistor. *Figure 6-5-A(a)*, the four-unit varistor in shunt connection, has the advantage of requiring no center-tapped windings for the carrier supply, *f(c)*, for the modulation voltage, *f(m)*, or for the output transformer. It has the disadvantage of being awkward to balance. There is no really convenient place to put a balancing resistor. The situation usually boils down to putting a small value of variable re-

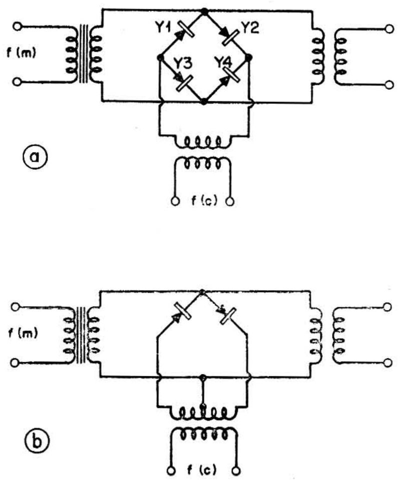


Fig. 6-5-A. Shunt type varistor balanced modulators. In these basic circuits; (a) is a four-unit varistor and (b) is a two-unit varistor.

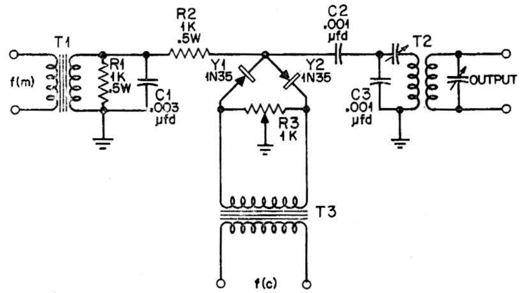
sistance in series with one of the diodes. Finding the right diode is a matter of cut and try. *Figure 6-5-A(b)* shows the basic circuit of the two-diode shunt varistor which produces a balancing action for *f(c)* if the voltages either side of the r-f transformer center tap are equal. This is usually not the case in the practical circuit so the schematic of *Fig. 6-5-B* was evolved. The balancing function is now performed by the 1000-ohm potentiometer. The need for a center-tapped r-f winding is now eliminated.

There is another matter that must be considered in the shunt-connected circuit in order to realize efficient operation. This is the matter of the input and output impedances. The impedance of the *f(m)* source (the audio transformer secondary) must be low at the audio frequencies and must appear as a high impedance at the carrier frequency. The impedance of the output circuit should be a low impedance at the carrier frequency and a high impedance at the audio frequencies. The latter condition is easy to realize by the use of a series tuned circuit in the output resonant at the desired r-f output frequency. Accomplishing the impedance condition for the audio frequency input circuit is a matter of compromise. The transformer winding has an impedance of 500 ohms at audio frequencies and the R/C network *(R1, R2,* and *C1)* helps to bring about the desired condition at the higher carrier frequency. This matter of having the correct impedances available for the input and output circuits often makes the use of the series-type of modulator more desirable.

The Series-Type Varistor Modulator

We now come to the series-connected varistor modulator covered in *Fig. 6-5-C*. This is the four-varistor circuit in which the carrier, *f(c)*, cannot appear in the output circuit when the varistor is perfectly balanced. Application of modulating signal, *f(m)*, disturbs the balanced condition and upper and lower sideband signals about the carrier frequency are generated. Balancing this four-diode varistor series circuit is as hard to achieve as in the shunt counterpart. For this reason, the circuit of *Fig. 6-5-D* is preferred in which *R1* is used as a balancing control.

The input and output impedances of the modulator must now be considered. The impedance of the audio source must be high at audio frequencies and low at the r-f carrier



C1—0.003 μfd., mica.
C2, C3—0.001 μfd., mica.
R1, R2—1000 ohms, ½w.
R3—1000-ohm potentiometer.
T1—Plate-to-500 ohm line audio transformer.
T2—I.f. transformer with series tuned primary.
T3—Carrier oscillator transformer, secondary has low impedance—approximately 500 ohms.
Y1, Y2—1N35 germanium varistor.

Fig. 6-5-B. Practical adaptation of Fig. 6-5-A(b).

frequency. This is accomplished by using a high-impedance secondary on the audio transformer and shunting it with *C1* which appears as a low reactance at the r-f carrier frequency but whose reactance is negligible (with respect to the transformer secondary impedance) at the highest audio frequency used, 3000 cps. The other impedance condition that must be met is that the output impedance of the modulator must be high at the r-f output frequency and low at the audio frequencies. This condition is most easily met by a parallel-tuned circuit.

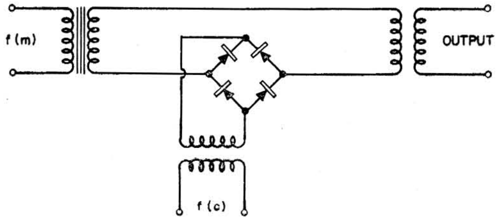


Fig. 6-5-C. Basic circuit for the series balanced modulator using a four-unit varistor.

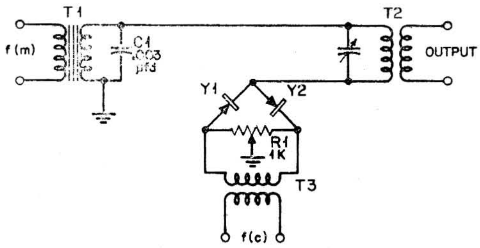


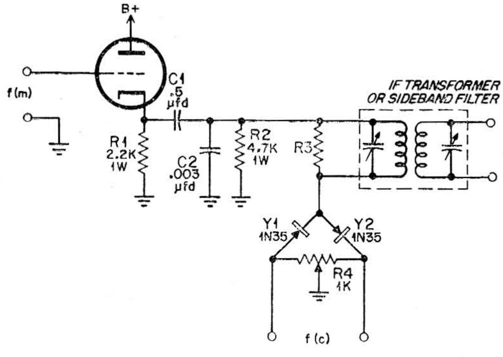
Fig. 6-5-D. A series type balanced modulator employing a varistor arrangement.

This circuit can be just an ordinary L/C tuned circuit or it can be the input circuit of a filter whose input impedance is high in the filter passband and low outside the passband limits.

A modification of this two-diode modulator was used in the SSB exciter described in *Chapter VIII.* This stage, shown in *Fig. 6-5-E,* uses one of the *Collins* 455-kc. mechanical filters which has an input impedance of 16,000 ohms in the passband region. The input circuit is a parallel tuned circuit which will fall to a low value of impedance at audio frequencies. The cathode follower stage was substituted for the input transformer of *Fig. 6-5-D.* The cathode follower load resistor, *R1,* is high impedance when compared to the filter impedance for the audio frequencies. The shunting condenser *C2* is a low reactance at 455 kc. when compared to the 16,000 ohms filter characteristic impedance, yet *C2* has negligible shunting effect at 3000 cps. The blocking condenser *C1* prevents the cathode bias across *R1* from biasing the diodes and disturbing operation.

The Ring Modulator

Figure 6-5-F is the classical circuit used by the telephone companies for the land-line SSB generators. This same circuit has been used by many Hams using filters in the 10 to 50-kc. region. The input and output impedances of this circuit are low—about 600 ohms. The balancing potentiometer *R1* is placed between the two halves of the winding of the output transformer primary. The impedance of the carrier-voltage feed is very low and is most easily matched by using a link winding on the carrier voltage supply. This type of circuit is seldom used above a frequency of about 100 kilocycles.



C1—0.5 μfd., paper.
C2—0.003 μfd., mica.
R1—2200 ohms, 1w.
R2—4700 ohms, 1w.
R3—Filter termination (if needed).
R4—1000-ohm potentiometer.
Y1, Y2—1N35 germanium varistor.

Fig. 6-5-E. Modified series balanced modulator.

The Modified Ring Modulator

The circuit shown in *Fig. 6-5-C* is one form of modified ring modulator that has proven very useful in SSB exciters. This basic circuit is used in the "SSB-Jr." described by W2KUJ and in the popular commercial "Multi-Phase" exciters. The two signals are connected in series to the common feed-point of *Y1* and *Y2* which is the slider of the balance potentiometer, *R1.* The secondary of *T1* is a low impedance winding—usually a voice coil winding—and *L1* is a link coil wound on the carrier supply oscillator coil. The push-pull tank circuit must resonate at the frequency of the desired output.

Capacitive Balance

In the discussion thus far, resistive balance has been covered, but no mention has been made of correction for unequal tube and circuit stray capacities. Quite frequently no capacitive balance circuit is needed. The resistive balancing potentiometer will in most instances give quite satisfactory balancing action.

If capacitive balance is also needed, a small variable condenser of the air variable or compression mica type should be added to either one of the two possible points in the balanced modulator circuit. Finding the correct point to connect the correcting condenser is a matter of experimentation. This added condenser should be connected between ground and the point in question. If the capacitive unbalance is too great for the variable range available on the correcting condenser additional fixed mica condenser should be added to bring the balance point within range.

The capacitive balance condenser should be connected between ground and one of the grids of *Fig. 6-1-B.* A 100 μμfd. compression mica condenser is usually adequate. Either plate of *Fig. 6-2-A* is used for balancing. The balance condenser should be placed from one grid to ground in *Fig. 6-3-A.* One control grid of *Fig. 6-4-A* is used while one of the oscillator injection grids (grid #1) of *Fig. 6-5-A* is the correct place for the balance condenser.

The series and shunt types of varistor balanced modulators (shown in *Figs. 6-5-B, 6-5-C* and *6-5-D*) are usually handled by connecting the balance condenser to one side or the other of the carrier supply source. If an isolation

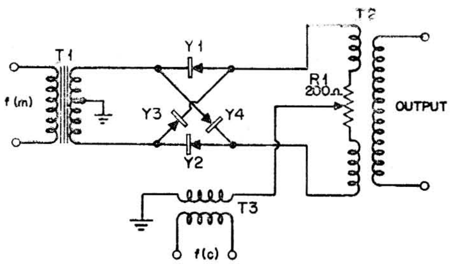


Fig. 6-5-F. The ring modulator using four copper oxide rectifiers as a varistor.

transformer is used between the carrier oscillator and the varistor, the balance condenser should be connected to one of the secondary windings.

The ring modulator of *Fig. 6-5-F* is best corrected for capacitive unbalance by placing the small balancing condenser from one of the outer ends of the output transformer primary to ground. The modified ring modulator of *Fig. 6-5-G* is also balanced by connecting the balancing condenser from one side of the tuned output coil to ground.

6.6—Spurious Signal Generation

Use of the balanced modulators described in this chapter has been restricted to modulating an r-f carrier frequency signal with a speech frequency signal (0 to 3000 cps). Practically any of the circuits described can be used to convert a single-sideband signal of one frequency to another frequency where it can be used. For example, if a filter-type exciter is used and the sideband is generated at 455 kilocycles, we must convert this into the 3.9-Mc. band so that we may make use of it. The first thought that might come to mind would be to use an ordinary penta-grid mixer tube and not use a balanced modulator at all. This is often the simplest solution. There are occasionally instances where spurious signals are generated in the mixing process that make the more complex balanced modulator circuits attractive.

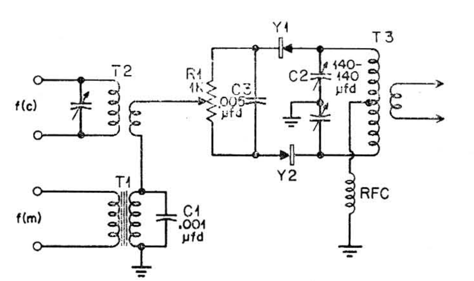
The spurious signals are formed in the same manner as intermodulation distortion products in a linear amplifier. The output tuned circuit for a frequency converter will not generally be tuned to either of the two input signal frequencies *f(1)* or *f(2)* but will be tuned to either *f(1)* + *f(2)* or *f(1)* — *f(2)* so that a new frequency *f(3)* is formed.

The spurious signals we are concerned with will be those that fall near *f(3)* in the frequency spectrum. The proximity of danger will be dependent upon the selectivity of the tuned circuits that follow. The actual generation takes place because of the mixing action that will take place between harmonics of *f(1)* and harmonics of *f(2)*. For example, the 4th harmonic of *f(1)* might mix with the 7th harmonic of *f(2)* to produce a "difference" mixture that falls near *f(3)* in frequency. Even if the wave shapes of both *f(1)* and *f(2)* signals are free of harmonic content, the mixing process generates harmonic-frequency signals that will cause trouble.

The ordinary variety of mixer stage where both *f(1)* and *f(2)* signals are fed into the tube in a "single-ended" arrangement and the output circuit is also single-ended is probably the worst offender. This circuit will form both the odd and even harmonics of both input signals *f(1)* and *f(2)*. The picture may appear darker than really need be for the situation can be helped in the single-ended mixer circuit by keeping a high ratio between the amplitude of the heterodyning oscillator signal and that of the SSB signal to be converted. The desirable ratio might be as high as 100 to 1.

If, however, trouble is encountered, one of the various forms of balanced modulators should be tried. Before starting it should be determined by some scratch-pad calculations just what harmonics of the *f(1)* and *f(2)* signals are causing trouble. It is an even harmoni of *f(1)* mixing with an odd harmonic of *f* the problem is to eliminate either one of offending harmonics being generated. It necessary to eliminate both. If one is elin. nated the spurious mixture product will be eliminated.

The best "all-around" solution would be to use the push-pull type of balanced modulator *Fig. 6-1-B* because the parallel feed of the heterodyning oscillator signal *f(c)* and the push-pull connection of the output circuit tend to cancel *all* harmonics of the *f(c)* signal. This is fortunate if the related circuits that can be used permit the configuration shown. If it is not possible to use such an arrangement then an intelligent choice must be made among the other types of circuits.



C1—0.001 µfd., mica.
C2—Dual 140 µµfd., air variable.
C3—0.005 µµfd., mica.
R1—1000-ohm potentiometer.
T1—Plate-to-4 ohm voice coil transformer.
T2—Oscillator tank circuit and link winding.
T3—Output tank tuned to desired frequency.
Y1, Y2—

Fig. 6-5-G. Modified ring modulator.

Hints for Better SSB Operation

Transistor Audio Oscillator

The reasonable price of the *Raytheon CK-722* junction transistor tempted the author to build a small compact audio oscillator that would always be available to plug into the audio input jack of the SSB transmitter for tune-up and two-tone tests. The extreme low-current drain of the transistor made the battery life limited to practically its shelf life.

The Circuit

The oscillator circuit is the transistor equivalent of the familiar Colpitts circuit. The tuned circuit of *Fig. 11-3-A* is made up of a 3-henry toroid coil *L1* and the pair of condensers, *C2* and *C3*. The toroid has a molybdenum permalloy core. Such units are available commercially from the *Arnold Engineering Co.,* Merengo, Ill. Ordinary iron-core filter chokes will not be satisfactory because of the saturation characteristics of the core material and the *Q* of the coil will not be high enough for reliable oscillation to take place.

The battery supply for the oscillator is made up of three pen-light cells. Since the total battery drain is approximately 30 microamperes it was decided not to include an *ON-OFF* switch in the unit because the operating life of the batteries should be about one year with continuous operation.

The values of the two resistors *R1* and *R2* may have to be adjusted for the particular transistor used. There is some variation between different CK-722 units and the resistors should be adjusted for best output waveform as viewed on an oscilloscope. For a zero resistance at *R2,* the waveform of the output will be distorted showing signs of clipping of one half of the sine-wave. The waveform improves rapidly for an increasing value of *R2* and its value should be set for the best wave shape and still have sufficient output voltage to drive the equipment. Since this oscillator feeds into the microphone jack, the voltage can be a very small fraction of a volt and still be usable.

The oscillating frequency is approximately 1000 cycles and the output voltage of the unit described was 0.7 volts rms. The output impedance of the oscillator is high and should not be loaded down by any circuit in the transmitter input. Loading will cause a change of frequency and possibly stop oscillation.

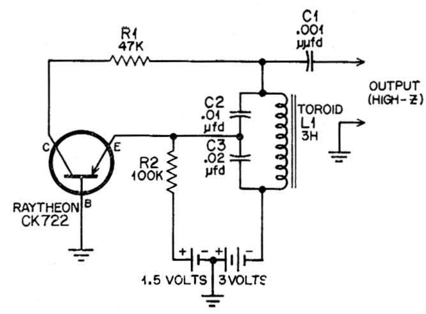


Figure 11-3-A.

For simplicity and convenient operation the unit was mounted in a small home-made aluminum box that was fitted to the end of a shielded phone plug. The unit is completely self-contained and has to be just picked up and plugged into the exciter. The specification sheet put out by *Raytheon* shows the connections of the small transistor unit. One precaution that should be kept in mind is to connect the battery voltage to the emitter element of the transistor before connecting the voltage to the collector element. The emitter is the element represented by the arrow symbol.

Chapter VII

Phase Shift Networks

In the phasing method of generating a SSB signal we must furnish both an audio signal and and r-f carrier signal to each of two balanced modulator stages. As described in *Chapter III* the audio signals furnished the two balanced modulator stages must be identical in every respect except that the phase relationship of one of the audio voltages must differ from the other by ninety degrees. The ability of the equipment to maintain the ninety-degree phase difference and also amplitude equality will determine to what extent the unwanted sideband is suppressed. This discussion is concerned with the intelligible sideband products. The intermodulation distortion products will be forgotten for the moment.

7.1—Effect of Phase Shift Errors

Several writers[10,27,28] have shown that if the two audio voltages differ in phase by an angle A and assuming that all other circuit conditions are perfect that the suppression of the unwanted sideband with relation to the wanted sideband in decibels is:

$$\text{Sideband Suppression in db.} = 20 \times \text{Log} \times \text{Cot } A/2$$

If actual values for angle A are assumed and substituted in the above formula, we find that a phase error of 1 degree will result in a sideband suppression of 40 db., an error of 2 degrees = 35 db., and an error of 3.5 degrees = 30 db. suppression. This will give the reader an idea of how much an error in the phase-shift network will affect the output signal. Again, this was assuming that the amplitude relation was perfect and also that the r-f phase shift network, discussed later, is perfect.

7.2—Effect of Amplitude Errors

Errors in amplitude equality of the two audio signal voltages will also affect the amount of sideband suppression. The amount of effect is expressed by the fromula:

$$\text{Sideband Suppression in db.} = 20 \text{ Log} \frac{200 + E}{E}$$

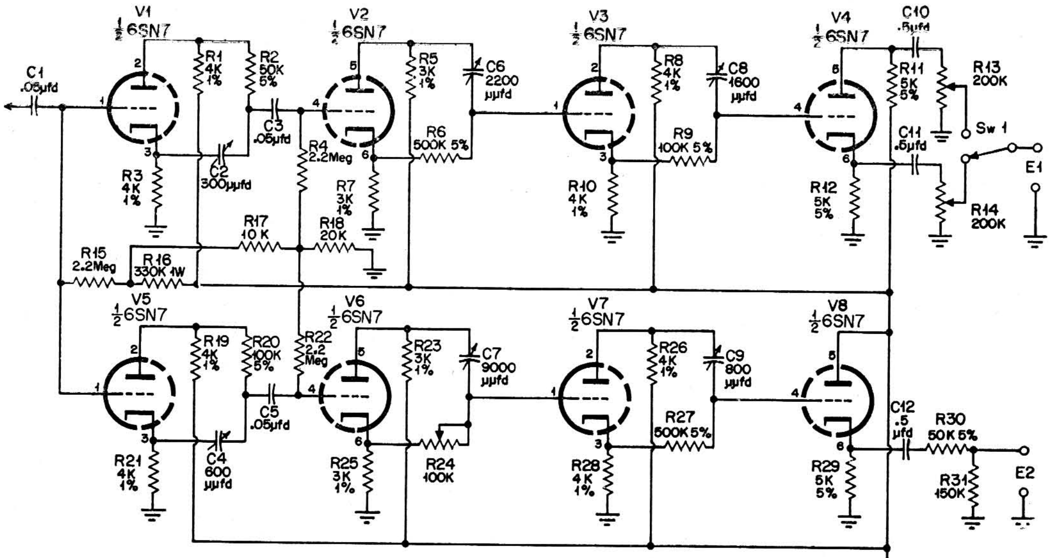
where E is the difference in the two voltages expressed in *per cent*. Thus a difference between the two voltages of 1% would of itself result in a sideband suppression of 45 db., an error of 2% would yield 40 db., 4% results in 34 db. suppression and so on.

7.3—Practical Circuits

The idea of generating single-sideband signals by the phasing method was seen to be feasible many years before it became a practical reality. The difficulty arose in being able to design a phase shift network that would be effective over a band of frequencies—namely those used in the transmission of speech. R. B. Dome[10] revised the first widely used circuits. He showed that it is possible to build circuits using RL and C combinations, but also showed how the results could be obtained by using only RC circuits. This is considered to be the easiest approach because of the difficulty involved in making an inductance that has a low distributed capacity and low resistance. It is easier to build a pure resistance and pure capacitor.

The Dome Vacuum Tube Circuit

Dome[10] also showed how it was possible to use vaccuum tubes to couple the various sections of the RC network. (See *Fig. 7-3-A*) It was this circuit that Norgaard, W2KUJ used in his first exciter[28] described in *CQ* and also in his YRS-1 single-sideband receiving adapter.[27] This network successfully maintained the amplitude and phase relationships necessary over the voice



frequency range from 50 to 5500 cps. This more than provides the minimum speech frequency spectrum necessary to carry on satisfactory communication.

C1, C3, C5—0.05 µfd., 400v.
C2—300 µµfd., compression mica.
C4—600 µµfd., adj. (500 µµfd. mica +100 µµfd. compression mica).
C6—2200 µµfd., adj. (2000 µµfd. mica +200 µµfd. compression mica).
C7—9000 µµfd., adj. (8200 µµfd. mica +800 µµfd. compression mica).
C8—1600 µµfd., adj. (1250 µµfd., mica +350 µµfd. compression mica).
C9—800 µµfd., adj. (500 µµfd., mica +300 µµfd. compression mica).
C10, C11, C12—0.5 µfd., 400v.
R1, R3, R8, R10, R19, R21, R26, R28—4000 ohms, ±1%, ½w.
R2, R30—50,000 ohms, ½w.
R4, R15, R22—2.2 meg-ohms, ½w.
R5, R7, R23, R25—3000 ohms, ±1%, ½w.
R6, R27—500,000 ohms, ½w.
R9, R20—100,000 ohms, ½w.
R11, R12, R29—5000 ohms, ±5%, ½w.
R13, R14—200,000-ohm potentiometer, linear taper.
R16—330,000 ohms, 1w.
R17—10,000 ohms, ½w.
R18—20,000 ohms, ½w.
R24—100,000-ohm potentiometer, linear taper.
R31—150,000 ohms, ½w.
NOTE: All resistors are composition or deposited carbon types—no wire wound resistors may be used.

Fig. 7-3-A. "Dome" audio phase shift network with an effective range of 50-5500 cps. For alignment information see reference footnotes and text material.

The Passive Network

It is also possible to build audio phase-shift networks using only *R* and *C* in a passive network—no tubes. There are various configurations that work successfully. Probably the most simple circuit is shown in *Fig. 7-3-B.* The example shown was designed to work between 300 and 3000 cps. This is considered an adequate speech spectrum to transmit good intelligible speech. The complete phase-shift network is shown in *Fig. 7-3-C* with component values. From the design equations the following equality of ratios is found for the assumed band of speech frequencies:

$$\frac{R2}{R1} = \frac{R4}{R3} = \frac{C1}{C2} = \frac{C3}{C4} = 1.58$$

$$\frac{1}{R1 \times C1} = \frac{1}{R2 \times C2} = 3015$$

$$\frac{1}{R3 \times C3} = \frac{1}{R4 \times C4} = 11{,}780$$

Using the above ratios different component values may be used merely by arbitaraily choosing values for one "R" in each branch of the circuit—*R1* and *R3* for example. Instead of assigning values to two "R's", values could have been chosen for a "C" in each branch—*C1* and *C3* for example. For the network values of *Fig. 7-3-C, C1* and *C3* were assigned the value of 680 µµfd. and the other values calculated from the above ratios.

This particular network* was used by L. J. Russo,** W3JTU, in several phasing exciters and found to work satisfactorily. The networks were constructed using 2% tolerance condensers and 1% tolerance resistors. The maximum phase error between the two output voltages was found to be not greater than 1.5 degrees. This would account for a maximum possible sideband suppression of 37.5 db. This figure would probably be degraded slightly by the possibilities of error in the r-f phase-shift network or inequalities in the amplitudes of

* Available from B&W as their Model 350, Type 2Q4.
** From a thesis for M.S. at the University of Pennsylvania.

the audio or r f signals fed to the balanced modulators.

The circuit configuration of *Fig. 7-3-C* is the same as the audio phase-shift network used by Norgaard in his "SSB, Jr."[30] and in the *Central Electronics* "Multiphase" exciter described in *Chapter III*. This network is shown in *Fig. 7-3-D* for comparison purposes. The circuit values are different but the performance is the same.

Network Input and Output Impedances

The networks shown in *Fig. 7-3-C* and *Fig. 7-3-D* must be properly matched in order to perform properly. The input impedance must

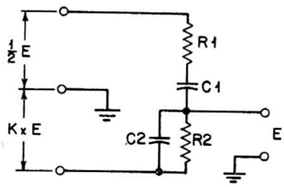
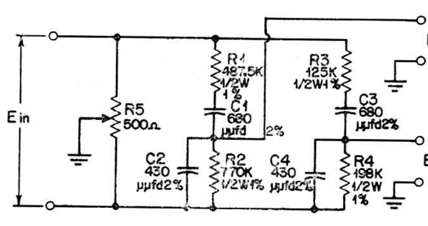


Fig. 7-3-B. Basic element of passive R/C audio phase-shift network.

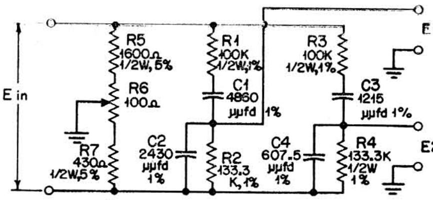
be kept low—500 ohms or less while the output impedance must be infinitely high—in practice a pair of class-A amplifier grids. If these conditions are not met no guarantee of the network performance can be made.

Note that the 500-ohm potentiometer *(R5)* used on the input to the phase-shift network of *Fig. 7-3-C* will not be in the midpoint position when properly adjusted. This control should be set up at a frequency of 1000 cycles so that the output voltages are at 90° relation, as shown by a circular pattern, when the two signals *E1* and *E2* are connected to the horizontal and vertical amplifier terminals and displayed on an oscilloscope. The pattern may not be circular even when the network is properly adjusted due to unequal channel gain settings and in this case the two axes of the



C1, C3—680 μμfd., ±2%, silver mica.
C2, C4—430 μμfd., ±2%, silver mica.
R1—487,500 ohms, ±1%, deposited carbon, ½w., IRC.
R2—770,000 ohms, ±1%, deposited carbon, ½w., IRC.
R3—125,000 ohms, ±1%, deposited carbon, ½w., IRC.
R4—198,000 ohms, ±1%, deposited carbon, ½w., IRC.
R5—500-ohm, potentiometer, carbon.

Fig. 7-3-C. Complete passive audio phase-shift network with an effective audio range of 300-3000 cps.



C1—4860 μμfd., ±1%, silver mica.
C2—2430 μμfd., ±1%, silver mica.
C3—1215 μμfd., ±1%, silver mica.
C4—607.5 μμfd., ±1%, silver mica.
R1, R3—100,000 ohms, ±1%, ½w.
R2, R4—133,300 ohms, ±1%, ½w.
R5—1600 ohms, ±5%, ½w.
R6—100-ohm potentiometer.
R7—430 ohms, ±5%, ½w.

Fig. 7-3-D. Passive audio phase-shift network used in the "SSB, Jr." and the "Multiphase."

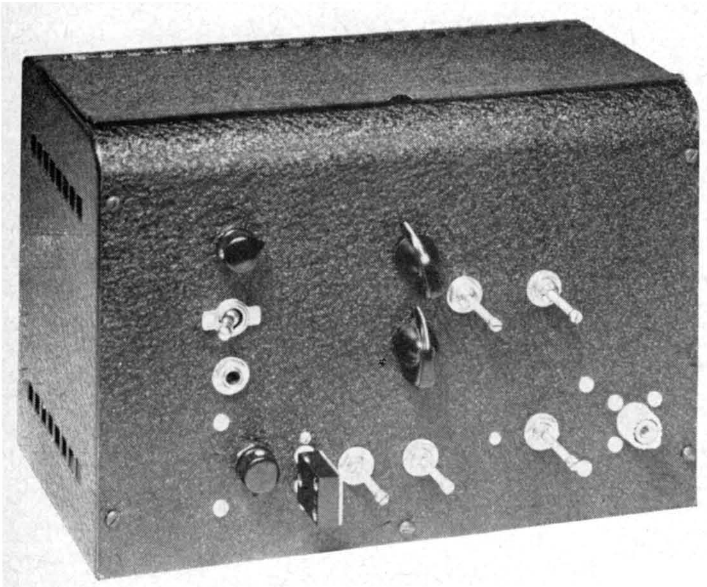
ellipse pattern should be made to be at right angles to each other. Once this control is set there is no further need for adjustment. The network as used in the "Multiphase" exciter has fixed values of voltage dividing resistors and no adjustment is necessary.

Restricting the Speech Spectrum

Since the two passive networks are phase accurate over a 10-to-1 speech frequency range (300 to 3000 cps) measures should be taken to restrict the speech frequencies to that range. Any speech frequencies outside the 300 to 3000 cps limits that pass through the phase-shift network will come to the dual output terminals without the necessary 90-degree relationship. This means that unnecessary spectrum space is being used both on the wanted sideband side of the carrier frequency and on the *unwanted* sideband side of the carrier over 3000 cycles distant from the carrier. Thus, on the suppressed sideband side of the carrier there will probably be satisfactory suppression of the sideband frequencies from 300 to 3000 cycles from the carrier and unsatisfactory suppression from 3000 cycles on out to the limit of the transmitted speech spectrum. The high-frequency sibulant sounds of human speech extend well beyond 3000 cps. Harmonic distortion in the speech amplification stages also produces audio signals higher than 3000 cps.

Therefore, it is necessary to insert a low-pass audio filter in the speech system preceding the phase-shift network. This filter must effectively attenuate the speech frequencies above 3000 cps. This can usually be accomplished by a single-section balanced *pi* filter of the type shown in *Fig. 7-3-E*. This filter must have a low characteristic impedance so that it can work into the low input impedance of the network shown in *Fig. 7-3-C*.

The 26 mh. coils of *Fig. 7-3-E (L1* and *L2)* can be toroid-wound inductances or can be a pair of slug-tuned TV-width adjusting coils



The popular "SSB, Jr." designed by W2KUJ is an excellent example of an exciter employing "fundamental phasing," where the sideband is generated on the same frequency as the output signal.

(Courtesy **General Electric**)

that are available in all television parts stores. The *Stancor WC-5* coil has an inductance range of 4 to 39 mh. and should be adjusted to the specified 26 mh. value by using an impedance bridge or Q-meter. If such an instrument is not available the slug setting can be made by resonating the coil with a 0.1 µfd. condenser at 3120 cps. The layout shown in *Fig. 7-3-F* can be used for adjusting the slug. The oscilloscope or VTVM indication should be maximum for the series-resonant condition. The two coils should be mounted so that there is minimum mutual coupling between the windings. If toroidal coils are used, they may be mounted on the same mounting screw because a toroid has no external field. The characteristic impedance of this filter is 500 ohms.

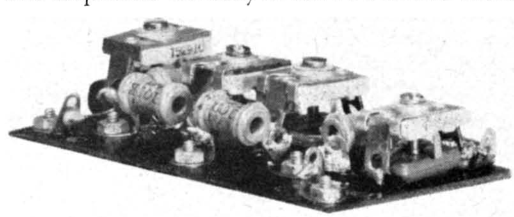
Other Types of Audio Phase-Shift Networks

There are other more elaborate networks with correspondingly more accurate phase and amplitude characteristics. These usually have wider audio bandwiths and small phase error—an example of one would be a bandwidth of from 300 to 20,000 cps and a maximum error of 0.2 degrees. This particular network used 12 resistors of the precision type and 10 precision condensers.

The networks described have sufficient bandwidth and phase accuracy to be satisfactory for most amateur applications.

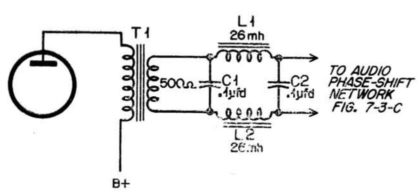
7.4—The R-F Phase Shift Network

As mentioned previously the r-f signals that are fed to the two balanced modulators in a phasing exciter must have equal amplitudes and a 90-degree phase relationship. The phase and amplitude accuracy of the r-f network affect the sideband suppression in exactly the same manner that the audio phase-shift network does. The same formulas that reveal the sideband suppression for given errors in phase relationship and amplitude relationship for the audio network also can be used for the r-f phase shift network. Thus if there is an error of 1° in phase the sideband suppression will be 40 db. and so on as discussed earlier in this chapter.



Detail view of the audio phase-shift network in the "SSB, Jr." The resistors are precision values and the trimmers are in parallel with fixed values to "tune" the network.

(Courtesy **General Electric**)



C1, C2—0.1 µfd., 200v.
L1, L2—26 mh., toroid wound, or use a TV width coil as described in the text.

T1—Plate-to-500 ohms line audio transformer.

Fig. 7-3-E. Speech frequency low-pass filter with a cut-off at about 3000 cycles.

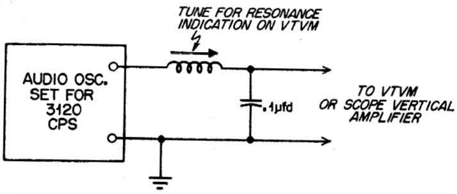


Fig. 7-3-F. Test set-up to be used in adjusting a TV width coil for a 26 mh. inductance value for use in the filter shown in Fig. 7-3-E.

There are two general philosophies in building phasing type SSB exciters and these are: (1) Build a fixed frequency phasing-type sideband generator and heterodyne the single sideband signal into the amateur bands of interest and; (2) Build a "fundamental phasing" type of exciter in which the sideband is generated at the same frequency as that desired for the output signal.

The "Multiphase" exciter is an example of the first type and as such generates the single sideband at a frequency of 9.0 megacycles. Thus, the r-f phase-shift network must be aligned to operate at only 9.0 Mc. For such fixed frequency service it is relatively simple to design

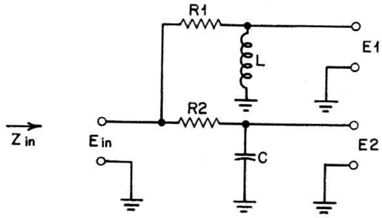


Fig. 7-4-A. R/L/C radio frequency phase-shift network. (See Table 7-4-A for values)

an r-f network that will perform quite well and give little error in phase and amplitude of the two output voltages.

However, with the "SSB, Jr."[30] the situation is different. The "SSB, Jr." is an example of the second type of signal sideband exciter and generates the sideband on the fundamental frequency of operation. The so-called "W2UNJ exciter"[35] is also a fundamental-type phasing exciter. These units take a v-f-o signal at the output frequency and pass it through an r-f phase shift network and combine it with the dual-channel audio signal as already discussed. What happens to the phase and amplitude relationships of the r-f network output signals when the v-f-o frequency is changed must be of primary concern.

f (Mc)	R1, R2(OHMS)	L(μh)	C(μμfd)
3.9	50	2.04	817
7.25	50	1.10	439
14.25	50	.56	223
21.2	50	.37	150
29.0	50	.274	110

Table 7-4-A: Values for use with the network shown in Fig. 7-4-A. The input impedance is 50-ohms.

The R-L-C Network

Figure 7-4-A shows the schematic of a simple r-f phase shift network which will perform satisfactorily if the components are chosen properly. The phase relationship is always 90° regardless of the frequency if the component sizes are chosen to satisfy the following relationships:

$$R1 = R2 = R$$

$$C = \frac{1}{2\pi F_o R} \ ; \ L = \frac{R}{2\pi F_o}$$

where F_o is the center frequency of each amateur band.

However, the amplitude characteristic of the network will not be constant. It is found that if the frequency is moved *n* per cent from the center design frequency, the amplitude difference between the two output voltages will be *n* per cent. Knowing the possible values of *n* we can estimate the worst sideband suppression that is due to the r-f phase-shift network. The 75-meter band is the widest amateur band percentage-wise, that is, the 200-kc. width divided by the 3900-kc. mid-band frequency. Assuming a design center frequency of 3900 kc., the maximum deviation for the v.f.o. to cover the whole phone band would be 100 kc. Thus $100/3900 \times 100 = 2.56\%$. So for 2.56% change in frequency there will be a difference in amplitude between the two output voltages of 2.56% also. This will cause the sideband suppression to be degraded to 37 db. if all other conditions are perfect.

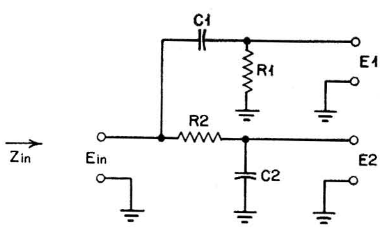


Fig. 7-4-B. R/C radio frequency phase-shift network. (See Table 7-4-B for values)

The input impedance of the network in *Fig. 7-4-A* is resistive and is equal to *R*. The output impedance is very high and should not be loaded down by grid resistors or should not be operated into tube grids that draw grid current over a portion of the r-f cycle.

If a value is assumed for the two equal resistors such that we could match a v-f-o source —say a 50-ohm line, it is then possible to calculate the values of *L* and *C*. *Table 7-4-A* shows

$Z_{in} = \frac{1}{2}(50 - j50)$ AT CENTER FREQ.		
f (Mc)	R1, R2 (OHMS)	C1, C2 (μμfd)
3.9	50	817
7.25	50	439
14.25	50	223
21.2	50	150
29.0	50	110

Table 7-4-B. Values for use with the network shown in Fig. 7-4-B.

values that have been calculated for each amateur band from 80 through 10 meters. Component values for other assumed values of *R* can be worked out by the interested reader.

The effect of nominal tube input capacitances shunting the inductive branch of the network is very slight and for all practical purposes may be neglected. An attempt should be made to hold any shunting capacity to a minimum, meaning no shielded leads on the output side of the network.

The R-C Network

Figure 7-4-B shows another r-f phase shift network that might be used. For each design center frequency, the capacitive reactance of the two condensers should be made equal to the value chosen for the two resistances.

The input impedance of this network is *not* a pure resistance and varies considerably with frequency. This might possibly give trouble by causing a varying load on the v.f.o. as the frequency is changed. Otherwise the characteristics of the output are similar to those of the *R-L-C* circuit shown in *Fig. 7-4-A. Table 7-4-B* shows values calculated for assumed values for the *R* values of 50 ohms.

The PI Network

Figure 7-4-C is shown for informational purposes. The over-all performance of this network is poorer than those of *Fig. 7-4-A* and

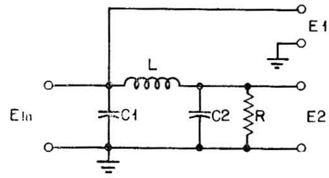


Fig. 7-4-C. The "pi" r-f phase-shift network.

7-4-B because a change of the variable oscillation frequency causes both a change of phase and a change of amplitude so that the overall exciter performance is considerably poorer than with the other two types of networks. Also since there is a slight loss in the *pi* section there is a certain amount of initial attenuation in the *E2* signal that must be corrected by inserting a small attenuation in the *E1* circuit.

Chapter VIII

A 35-Watt SSB Transmitter

The recent advent of the *Collins* "mechanical filter" has made available to the experimenter selectivity that heretofore had been only dreamed of. The package was also very small. The literature has described the performance characteristics[5,36,42] and a number of adapter units have been described that permit use of the mechanical filters in existing communications receivers.[1,6,9] To date, little has been mentioned about using the mechanical filter in a transmitter-exciter for single sideband use. The market price makes the 455-kc. units attractive for SSB generation. It is the purpose of this chapter to describe an exciter using the 3-kc. bandwidth *Collins filter* in complete detail.

When it was decided to build an exciter specifically for publication in this book, it was thought that one would be designed that had everything *including* the "kitchen sink." We recognize that the desires and requirements of the individual vary, so the end result has been to include as many features in the exciter that appear to be practical. The individual can delete the features that he deems unnecessary for his particular application.

To get to specific examples: The exciter includes the following features: (1) CARRIER INSERTION for tune-up or working stations unable to copy SSB, (2) SIDEBAND SWITCHING—permitting either upper or lower sideband transmission, (3) VOICE CONTROL operation which permits virtually full duplex type of operation, (4) RECEIVER ANTI-TRIP feature which permits loudspeaker operation while operating voice control, (5) V-F-O control, (6) provisions for "patching out" to an external band-changing heterodyne unit so that operation may be had on other than the fundamental 75-meter band, and (7) 6146 tetrode output stage that gives approximately 35 watts of peak output power.

Any of the above features may be eliminated without destroying the purpose of the basic exciter. The appropriate tube or tubes can be eliminated to make the unit smaller and less complex. The philosophy behind this exciter was *not* to make it as simple as possible, but to make it as sure-fire and conservative in operation as possible. All of the circuits used in the exciter have been used in at least one and in most cases several exciters built by the author.

The Block Diagram

To successfully build and operate any transmitter the user must have a clear understanding of the function of each stage in the unit. Thus a block diagram (see *Fig. 8-1-A*) is in order to best explain the various inter-related functions of the single sideband exciter described in this chapter.

Starting on the left of the block diagram, the microphone of the usual high impedance crystal variety feeds a two-stage SPEECH AMPLIFIER (6SN7). This provides more than adequate speech amplification. This stage, strictly speaking, is the only essential audio equipment needed in the exciter. A portion of the speech voltage is fed into the VOICE CONTROL STAGES which will be discussed in a moment. The normal output of the speech amplifier feeds into the CATHODE FOLLOWER stage preceding the BALANCED MODULATOR. The CATHODE FOLLOWER serves as an impedance matching and isolating device between the speech equipment and the r-f BALANCED MODULATOR.

The BALANCED MODULATOR stage is the two-diode series-type modulator discussed in *Chapter VI.* In essence, this stage combines the audio signals from the speech equipment and the r-f carrier signal from the block labeled L.F. XTAL. OSC. The output of the BALANCED MODULATOR block consists of a double-sideband suppressed-carrier signal in the region of 455 kilocycles.

The L.F. Xtal. Osc. block consists of a Pierce crystal oscillator in which one of two crystals is chosen for sideband switching plus a resistive-coupled phase-inverter stage furnishing push-pull output voltage from the single-ended Pierce oscillator. This push-pull voltage is necessary for operation of the Balanced Modulator stage already mentioned.

The block labelled Carrier Insertion is a cathode follower isolation stage whose output is adjustable with a potentiometer so that variable amounts of carrier may be shunted around the Balanced Modulator and Mechanical Filter blocks to the input of the I-F Amplifier. This permits the operator to insert any amount of carrier into his signal that he might desire without disturbing the critical carrier balance controls of the Balanced Modulator. Carrier insertion is convenient for transmitter tune-up, two-tone tests, and for working stations unable to copy carrierless SSB signals.

The block labelled Collins Filter rejects one of the sideband signals generated in the Balanced Modulator and allows the other to pass through to the I-F Amplifier.

The block labelled I-F Amplifier is just what the name implies. It is exactly like the i-f amplifier stages encountered in many run-of-the-mill receivers. The cathode resistance is variable making the gain also variable. The signal that this stage receives from the filter is a single-sideband suppressed-carrier signal and thus the stage must operate class A to faithfully reproduce the signal. This stage is necessary to make up the insertion loss of the mechanical filter and provide sufficient signal for the next stage.

The next stage is the Mixer and this is again very similar to the mixer stages used in many receivers. The Mixer combines the 455-kc. SSB signal and the c-w signal from the VFO. The output circuit of the Mixer stage is tuned to the additive mixture of the 455-kc. SSB signal and the 3350 to 3550 kc. v-f-o signal. You can see that this results in an SSB signal that falls in the 3800 to 4000-kc. amateur 'phone band.

Our SSB signal is now at a useful frequency. Thus our next stage is the 4-Mc. Amplifier which operates in class A and provides voltage amplification of our SSB signal. The output of this stage, you will note, goes to a chassis output terminal and is immediately connected back into another terminal that leads to the output stage, the Power Amplifier. The reason for this round-about path for the signal is to provide multi-band operation of the exciter. For fundamental operation in the 4.0-Mc. band, the signal should be patched back in as shown by the dotted line between the inter-stage terminals. For operation on other amateur bands, the signal is patched through a coaxial line to a separate mixer unit that again heterodynes the SSB signal from 4.0 Mc. to any other amateur band of interest.

The signal received at the input of the Power Amplifier is either from the preceding 4-Mc. Amplifier at a frequency in the 75-meter band or from the external heterodyne unit at a frequency in the 7.2 or 14.2- megacycle bands. Thus the input circuit must be capable of tuning to any amateur band. A band-switching

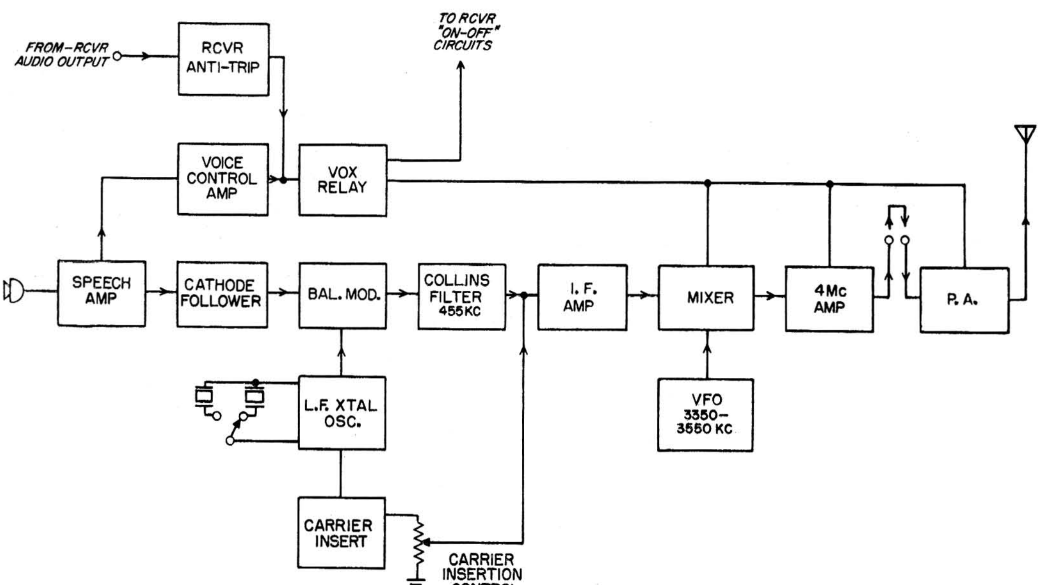


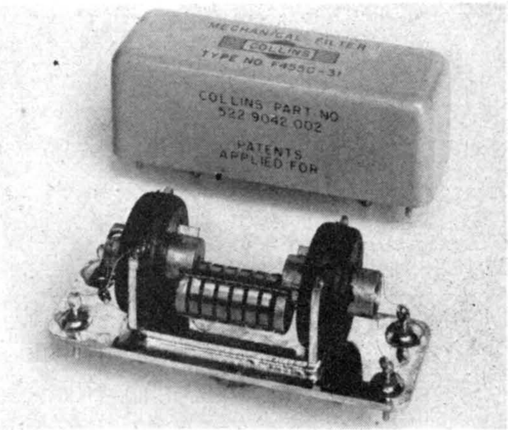
Fig. 8-1-A. Block diagram of the SSB transmitter using the mechanical filter.

arrangement was used. The POWER AMPLIFIER stage is a power tetrode (type 6146) operated in class AB_1. The output is in the neighborhood of 35 watts peak power and is at the frequency of the input signal of the stage, in other words, the stage always runs "straight through." Frequency multiplication is taboo in single sideband power amplifiers. The output circuit is again a band-switching turret with link coupling to the antenna or high-power amplifier circuit.

Voice Operation

We have thus far covered all the essential operational features of the exciter except the VOICE CONTROL AMPLIFIER and RCVR. ANTI-TRIP blocks. A portion of the speech amplifier signal is shunted into the block labelled VOICE CONTROL AMP. This two-stage amplifier boosts the amplitude of the audio signal to a high level and feeds it to the voice control relay circuit, VOX RELAY. Here, the speech signal, when present, operates the 3-pole, double-throw relay. Thus when the microphone is spoken into the relay is operated. When the relay is operated the plate and screen voltages are applied to the MIXER stage and to the 4-MC. AMP. stage. Another pair of contacts on the relay applies the screen voltage to the POWER AMP. stage.

The block labelled RCVR. ANTI-TRIP is interlocked with the VOICE CONTROL AMP. circuits so that any signal that comes out of the receiver loudspeaker and enters the microphone circuits will not trip the VOX RELAY. This is accomplished by taking a portion of the voice coil receiver signal and applying it to the relay control circuits so as to oppose the action of the regular VOICE CONTROL AMP. signal. A detailed discussion of the operation and circuitry is covered under the next section.



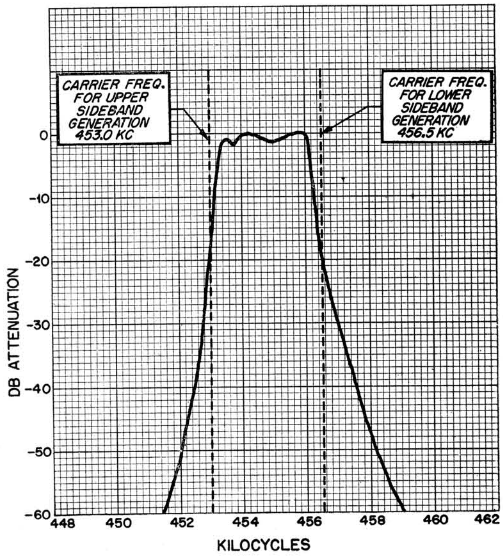


Fig. 8-2-B. Inside view of the Collins mechanical filter used in the transmitter described in this chapter. The band-pass curve shows the carrier frequencies and their relationship to the curve. See text on page 68 for additional details.

8.2—The Circuitry

The Speech Amplifier

Refer to *Fig. 8-2-A* for all discussion in this section. The microphone input circuit is for the usual variety of crystal microphone. Choke, *Ch-1*, and *C7* form a simple L-section, low-pass filter that prevents r.f. from getting into the speech system. The complete speech amplifier consists of the two sections of *V1*, a type 6SN7. The speech channel gain control is *R17*. The output of the speech section feeds into the grid of the cathode follower, *V2a*, which in turn is the coupling device to the balanced modulator circuit.

The Voice Control System

The principle of any voice control system is to take a portion of the speech signal and amplify it to a high level regardless of the amount of distortion in the wave shape that is produced. In fact, limiting of the speech wave shape is desirable so that a more positive voice control action takes place. The high value of speech voltage is then rectified and used as cut-off bias on a relay control tube. Most of the limiting action mentioned takes place in the cut-off characteristic of the relay tube, i.e., once the value of cut-off grid voltage is reached, application of more negative bias will accomplish no more than does just the bare essential for plate current cut-off.

The rectified voice control voltage is applied to the relay tube through an *R/C* time-constant circuit which prevents the bias voltage from falling below the cut-off value between syllables and words of speech. The usual *R/C* time constant required for good amateur practice is between one-tenth and one-quarter of a second. To calculate the time constant, multiply the value of the resistance in megohms by the value

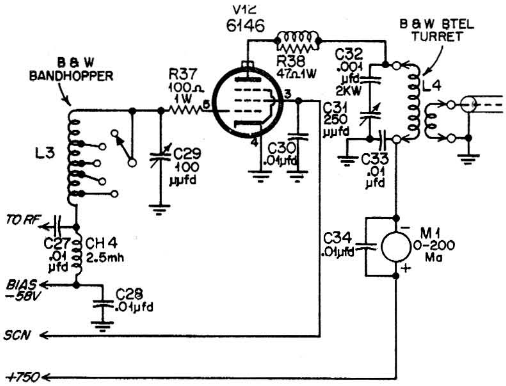


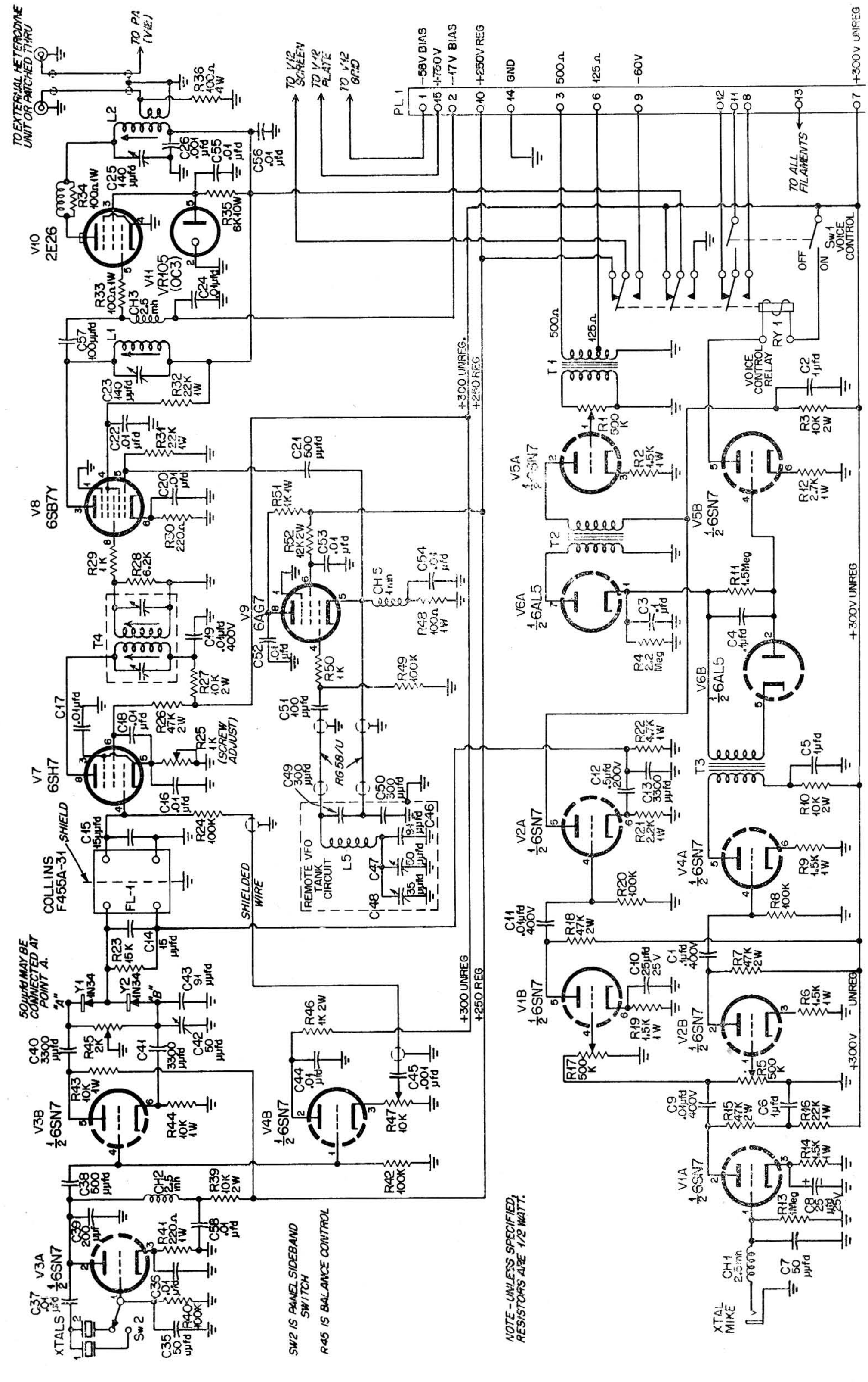
Fig. 8-2-A. Wiring diagram of the power amplifier stage. The main schematic is on the facing page. The complete parts list for both schematics is below.

of the associated condenser in microfarads and the answer will be in seconds. The time in seconds will be the time it takes the resistor to discharge the condenser so that the voltage appearing across the condenser is down to approximately one-third of the starting fully-charged voltage. It can be seen that the actual time the voice control system will stay energized will depend not only on the *R/C* time constant but also on the amount of the peak rectified d-c voice control voltage.

The input to the voice control channel is connected in parallel with the input to the second speech amplifier stage, the grid of *V1b.* The two potentiometers, *R17* and *R5,* are connected in parallel with their sliders being connected to the second speech amplifier stage, *V1b,* and to the input of the first voice control stage, *V2b,* respectively. This allows the voice control gain to be independent of the speech amplifier gain setting. This arrangement thus gives the voice control system three stages of amplification counting the first stage in the speech amplifier.

The transformer, *T3,* on the output of the last voice control stage, *V4a,* is a step-up transformer that gives still additional voltage for the VOX (voice control) system. The a-c voltage measured across the secondary of *T3* should be between 200 and 300 volts rms for a good loud whistle in the microphone with the voice control gain pretty well opened. The rectified d-c voltage appearing across the diode load resistor, *R11,* should be in the neighborhood of 200 volts with the same conditions. This is more than enough to cut the relay tube, *V5b,* off. Under average voice input conditions the

C1, C3, C4—0.1 μfd., 400v.
C2, C5, C6—1.0 μfd., 400v.
C7—50 μμfd., 100v., mica.
C8, C10—25 μfd., 50v.
C9, C11, C16, C17, C18, C19, C20, C22, C24, C27, C28, C30, C33, C34, C36, C37, C44, C52, C53, C54, C55, C56, C58—0.01 μfd., 600v. **Centralab** D6-103.
C12—0.5 μfd., 200v.
C13, C40, C41—3300 μμfd., 600v., **Centralab** D6-332.
C14, C15—15 μμfd., 600v., **Centralab** D6-150.
C51, C57—100 μμfd., 600v., **Centralab** D6-101.
C23—140 μμfd., variable, **Hammarlund** type APC-140.
C25—140 μμfd., variable, **Hammarlund** type MC-140S.
C26, C45—0.001 μfd., paper, 400v.
C29—100 μμfd., variable, **Hammarlund** type HF-100.
C32—0.001 μfd., 2500v., mica.
C31—250 μμfd., variable, **Hammarlund** type MC-250M
C35—50 μμfd., 600v., **Centralab** D6-500.
C21, C38—500 μμfd., 600v., **Centralab** D6-501.
C39—200 μμfd., 600v., **Centralab** D6-201.
C42, C47—50 μμfd., variable, **Hammarlund** type APC-50.
C43, C46—91 μμfd., 600v., Silvered mica.
C48—35 μμfd., variable, **Hammarlund** type MC-35S.
C49, C50—300 μμfd., 600v., Silvered mica.
Ch1, Ch2, Ch3, Ch4—R-f choke, 2.5 mh., **Millen** No. 34102.
Ch5—R-f choke, 1.0 mh., **Millen** No. 34107.
FL-1—Collins Mechanical filter, type F455A-31.
L1—35 turns, #28 enam., wound on **National** XR-50.
L2—24 turns, #22 enam., wound on **National** XR-50. 4 turns #22 formex linked around cold end.
L3—B&W "Band Hopper" type 2A, Stock #3121.
L4—B&W "BTEL" turret with 3.5 and 28 Mc. coils removed.
L5—27 turns, #22 enam., close wound in center of **National** XR-16 form.
M1—Meter, 0-200 ma., **Simpson** type 27.
R1, R5, R17—500,000-ohm potentiometer, **IRC** type Q.
R2—1500 ohms, 1w.
R3, R10, R27, R39—10,000 ohms, 2w.
R4—2.2 megohms, ½w.
R6, R9, R14, R19—1500 ohms, 1w.
R7, R15, R26, R18—47,000 ohms, 2w.
R8, R20, R24, R40, R42, R49—100,000 ohms, ½w.
R11—1.5 megohms, ½w.
R12—2700 ohms, 1w.
R13—1.0 megohms, ½w.
R16—22,000 ohms, 1w.
R21—2200 ohms, 1w.
R22—4700 ohms, 1w.
R23—15,000 ohms, ½w.
R25—1000-ohm potentiometer.
R28—6200 ohms, ½w.
R29, R50—1000 ohms, ½w.
R30—220 ohms, ½w.
R31, R32—22,000 ohms, 1w.
R33, R37, R48—100 ohms, 1w.
R34—100 ohms, 1w. Wind 6 turns of #22 enam. around the body of the resistor.
R35—6000 ohms, 10w., wire wound.
R36—100 ohms, 4w. (two 200-ohm, 2w. resistors in parallel).
R38—47 ohms, 1w. Wind 6 turns of #22 enam. around the body of the resistor.
R41—220 ohms, 1w.
R43, R44—10,000 ohms, 1w.
R46—1000 ohms, 2w.
R47—10,000-ohm potentiometer.
R45—2000-ohm potentiometer.
R51—1000 ohms, 1w.
R52—12,000 ohms, 2w.
Ry1—Relay, 3PDT, 10,000-ohm coil. **Automatic Elec. Mfg.** type R-45L.
Sw1—SPDT toggle.
Sw2—SPDT toggle
PL1—Plug, male chassis, 15 contacts, **Jones** No. P-315-EB.
T1—Input transformer, 500/125 ohms pri., 100,000 ohms sec. **Merit** No. A2924.
T2, T3—Interstage transformer, 1:3 ratio. **Stancor** A-63-C.
T4—I-f output, 455 kc., **Miller** No. 12-C2.
Xtal 1—Crystal 453 kc., **Knight** type H 7.
Xtal 2—Crystal 456.5 kc., **Knight** type H17.
Y1, Y2—Diodes, IN34A.
V1, V2, V3, V4, 5—6SN7GT, **RCA.**
V6—6AL5, **RCA.**
V7—6SH7, **RCA.**
V8—6SB7Y, **RCA.**
V9—6AG7, **RCA.**
V10—2E26, **RCA.**
V11—VR-105, **RCA.**
V12—6146, **RCA.**
ALSO REQUIERED:
1—Receptacle to match above plug, **Jones** S-315-CCT.
11—Octal sockets.
1—7-prong miniature socket.
2—Crystal sockets, **Millen** No. 33302.
5—Co-ax receptacles, SO-239.
5—Co-ax plugs, PL-259.
2—Coil shields, **National** type RO.
1—Dial, **Millen** 10039.
1—Chassis, **BUD** #AC-417.
1—Panel, **BUD** #PA-1106.
1—Chassis mounting brackets, steel, **BUD** MB459.
1—VFO box, **BUD** 4"x5"x6".


TO EXTERNAL HETERODYNE UNIT OR PATCHED THRU
TO PA (V12)
V10 2E26
V11 VR105 (OC3)
V8 6SB7Y
V9 6AG7
V7 6SH7
COLLINS F455A-31
FL-1
SHIELD
SHIELDED WIRE
REMOTE VFO TANK CIRCUIT
RG 58/U
V3A 1/2 6SN7
V3B 1/2 6SN7
V4B 1/2 6SN7
XTALS
Sw 2
50 μμfd MAY BE CONNECTED AT POINT A.
SW2 IS PANEL SIDEBAND SWITCH
R45 IS BALANCE CONTROL
NOTE - UNLESS SPECIFIED RESISTORS ARE 1/2 WATT.
TO V12 SCREEN
TO V12 PLATE
TO V12 GRID
PL 1
-58V BIAS
+750 V
-47V BIAS
+250V REG
GND
500 Ω
125 Ω
-60V
+300V UNREG
TO ALL FILAMENTS
Sw 1 VOICE CONTROL
OFF
ON
RY 1
VOICE CONTROL RELAY
T1
T2
T3
T4
V5A 1/2 6SN7
V5B 1/2 6SN7
V6A 1/2 6AL5
V6B 1/2 6AL5
V2A 1/2 6SN7
V2B 1/2 6SN7
V4A 1/2 6SN7
V1A 1/2 6SN7
V1B 1/2 6SN7
XTAL MIKE
CH1 2.5mh
+300 UNREG
+250 REG

d-c voltage should run somewhere in the region of 60 or 70 volts. This assures that the relay will drop out and operate the various exciter and receiver circuits without chopping off the first syllable of the first spoken word. The VOX gain control should be advanced far enough so that the system holds in between normally spoken words and syllables. A VOX system with insufficient gain or too short a time constant will produce very choppy sounding speech at the receiving end of the circuit. However, the time constant should not be so long that the circuit will not drop out during natural pauses in speech permitting the other station to break in for comment.

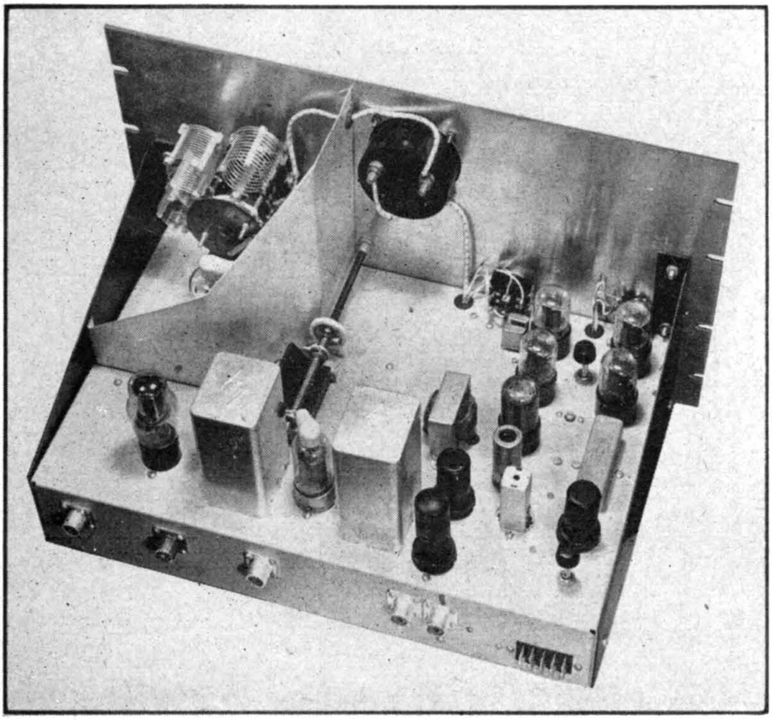
The diode load, *R11*, for the VOX channel rectifier, *V6b* is not between ground and the grid of the relay tube, *V5b*, but is in series with another diode load and associated time-constant condenser which is part of the receiver anti-trip circuit.

Receiver Anti-Trip Circuit

This circuit is a refinement that has become very popular with many amateurs that prefer to operate using a loudspeaker with VOX operation. Without the anti-trip circuit, any sound coming out of the loudspeaker would cause the VOX system to operate and give a false "operate" signal to all transmitter and receiver circuits. Since the VOX system is tied into the receiver circuits, as soon as the VOX system operates, the receiver is disabled and the actuating loud speaker sound is interrupted. This causes the VOX system to "unoperate" once again energizing the receiver and so on. The net result is a motor-boating action with an oscillating rate determined by the loudness of the receiver signal and the system time constant.

One solution to the problem without going to the anti-trip system is to use headphones so that the receiver signal cannot get back into the speech channel of the transmitter. Another solution to this problem is to use two microphones—one for the normal speech channel and the other (a throat microphone) connected into the voice control channel. This has the obvious disadvantage of requiring the operator to assume the trappings of a jet pilot. However, this does have the advantage that the operator is the only person who can possibly operate the system—obviously a very desirable feature for those operators with noisy junior op's or other sources of high background noise. The author actually used this latter system for almost a year and found it to be quite workable even though a good natured ribbing had to be taken occasionally from the rest of the SSB brethren.

The principle behind the *anti-trip* circuit is to take some of the receiver output audio signal from the voice coil circuit, amplify it, rectify it and apply it to the same relay control tube



Above chassis view of filter exciter showing general arrangement of parts. The various tubes and components may be identified by referring to page 73.

associated with the transmitter voice-control system. This *anti-trip* voltage must be applied with its polarity opposite to that of the negative voice-control voltage. Thus the *anti-trip* voltage is connected so as to apply a *positive* voltage to the relay control tube, *V5b*. Transformer *T1* of *Fig. 8-2-A* is a line-to-grid transformer. This transformer matches the receiver 500-ohm output impedance to the grid of tube *V5a*.

It can be seen that the *anti-trip* system has only one stage of amplification. The gain control for this section is *R1*. Transformer *T2* which couples to the diode rectifier is identical to *T3* used in the VOX channel. The two diode load resistors for the *anti-trip* and VOX systems are connected in series so that the voltage appearing between ground and grid of the relay tube, *V5b*, is the algebraic sum of the negative voltage of the VOX system and the positive voltage of the *anti-trip* system.

The theory of operation is simple and can be explained as follows: When the operator speaks into the microphone a negative voltage is generated in the VOX channel but the operators speech does *not* generate a positive voltage in the *anti-trip* channel. This causes the relay tube to be cut off and operate the transmitter and receiver circuits in a normal fashion. If, however, a signal is coming out of the station loud-speaker, it will enter the microphone and cause a negative voltage to be generated as before. However, simultaneously the electrically-connected circuit from the voice coil to the *anti-trip* circuit will generate a positive voltage resulting from the same signal coming out of the loud speaker. These two opposite-polarity d-c voltages appear across their respective load resistors and depending on which is greater the relay tube will operate or not operate the VOX relay. If the *anti-trip* gain control, *R1*, is set high enough the positive voltage appearing across the diode load of *V6a (R4)* will be slightly greater than the negative voltage appearing across *R11* thus causing the grid of the relay tube, *V5b*, to swing slightly positive. The high values of the series connected diode load resistors prevent the tube from drawing excessive grid current and resultant high plate current by acting as grid current limiting resistors for the positive grid operating region of *V5b*.

The Relay Circuit

The contacts of the 3-pole, double-throw voice-control relay, an *Automatic Electric R45-L*, perform all the switching functions of the voice-control system. One set of contacts apply and remove the plate and screen voltages to the mixer stage, *V8*, and to the 2E26 stage, *V10*. The second set of contacts apply the +250 volts to the screen grid of the power-amplifier 6146 stage when the voice control circuit is actuated. When in the non-operate condition, a −60 volts is applied to the screen grid of the 6146 tube. This feature of swinging the screen minus during stand-by periods may not be necessary for use of the exciter by itself, but is quite necessary if the exciter is followed by a high-gain amplifier. In this case, the very slight residual plate current of the 6146 will create electron shot-noise in its output circuit which will be amplified by the following stage. After amplification this noise (identified as a rushing sound in the receiver) may have sufficient strength to block the shack receiver when the exciter is in the stand-by position.

The third set of contacts on the relay are used to control the station receiver by whatever means available. The particular wiring arrangement shown in *Fig. 8-2-A* is for shorting the voice coil of the station loud speaker. Thus the second section of VOICE CONTROL ON-OFF switch, *Sw1*, opens the armature circuit when the switch is thrown over to the "OFF" position (or *manual* operation). This permits turning the exciter on to "zero in" on another SSB station. To do this the station receiver must be active and this would be impossible without the additional switch section. If non-*VOX* operation of the exciter is desired for an actual QSO, the station receiver must be disabled manually by the operator. This is a rare occurrence, the author has found. If some other method of controlling the receiver is desired, such as opening the cathodes of the i-f amplifiers or r-f amplifiers, or removing the B+ from some of the stages (always leave the B+ on the h-f oscillator) the following circuit changes must be made. The connections to *Sw1* must be changed so that the appropriate switch section *(Fig. 8-2-A)* is no longer in series with the armature lead to the external circuits. The arm of this section of *Sw1* must be connected across the armature and the relay contact used to control the receiver circuits. This is a simple modification, but has been pointed out for clarity of operational functions.

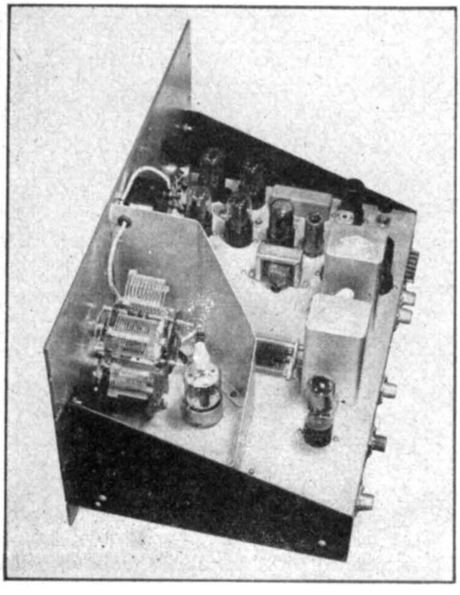
The Balanced Modulator

The balanced modulator is the point at which the first step of making a single-sideband suppressed-carrier signal takes place. That is, the suppressed carrier part of the operation is produced. The output of the balanced modulator that exists across the input terminals of the mechanical filter is a double-sideband suppressed-carrier signal. The operation of this type of modulator circuit was covered in detail in *Chapter VI*.

The cathode follower, *V2a*, isolates the audio stages from the modulator and simultaneously provides a proper impedance transformation. The condenser, *C13*, appears as a low reactance at the 455-kc. carrier frequency and thus creates the appearance that the audio cathode follower stage has a low-impedance at the carrier frequency. However, the same cathode follower stage must appear as a relatively high impedance at audio frequencies. This takes place because the reactance of *C13* is high in the audio-frequency spectrum. Condenser *C12* acts as a d-c blocking condenser so that the cathode bias developed across *R21* will not disturb the operation of the two diode rectifiers. The audio voltage appearing across *R21* should not exceed about 0.5 volts peak value.

The balancing controls, *R45* and *C42,* are necessary to correct for both resistive and capacitive unbalance in the modulator circuit. The potentiometer corrects for inequalities in the output voltage of the phase-inverter stage, *V3b,* and any difference that there might be in the characteristics of the two germanium diodes, *Y1* and *Y2.* The air variable condenser, *C42,* in parallel with *C43* corrects for the differences in the output capacities of the phase inverter stage and the stray circuit capacity.

Because two different carrier frequencies are used (the sideband switching feature) it may be necessary to rebalance the carrier suppression controls when switching sidebands. This was thought to be no great disadvantage since the sideband switch is generally operated only when changing bands. In the exciter described, the carrier did not rise to an annoying level when the sideband switch was thrown. It might be possible to arrive at a compromise setting of the balance controls so that a usable suppression of the carrier could be realized for both sidebands. If the carrier is suppressed greater than 20 db., the amplifiers will not be burdened by the "useless carrier." A suppression of 30, or more, db. is desirable so that the heterodyne to adjacent channel stations will not be annoying. With the exciter shown, after warm-up, a stable suppression of 60 db. was possible using the arrangement shown in *Fig. 8-2-A.*



Above chassis view showing P.A. stage and shield.

It may be necessary to change the capacitive balance condensers to the other end of the balance potentiometer, *R45,* in order to accomplish perfect balance. The opposite position should be tried if there is any question about getting a satisfactory balance. The two balance controls must be adjusted alternately for minimum carrier at the exciter output.

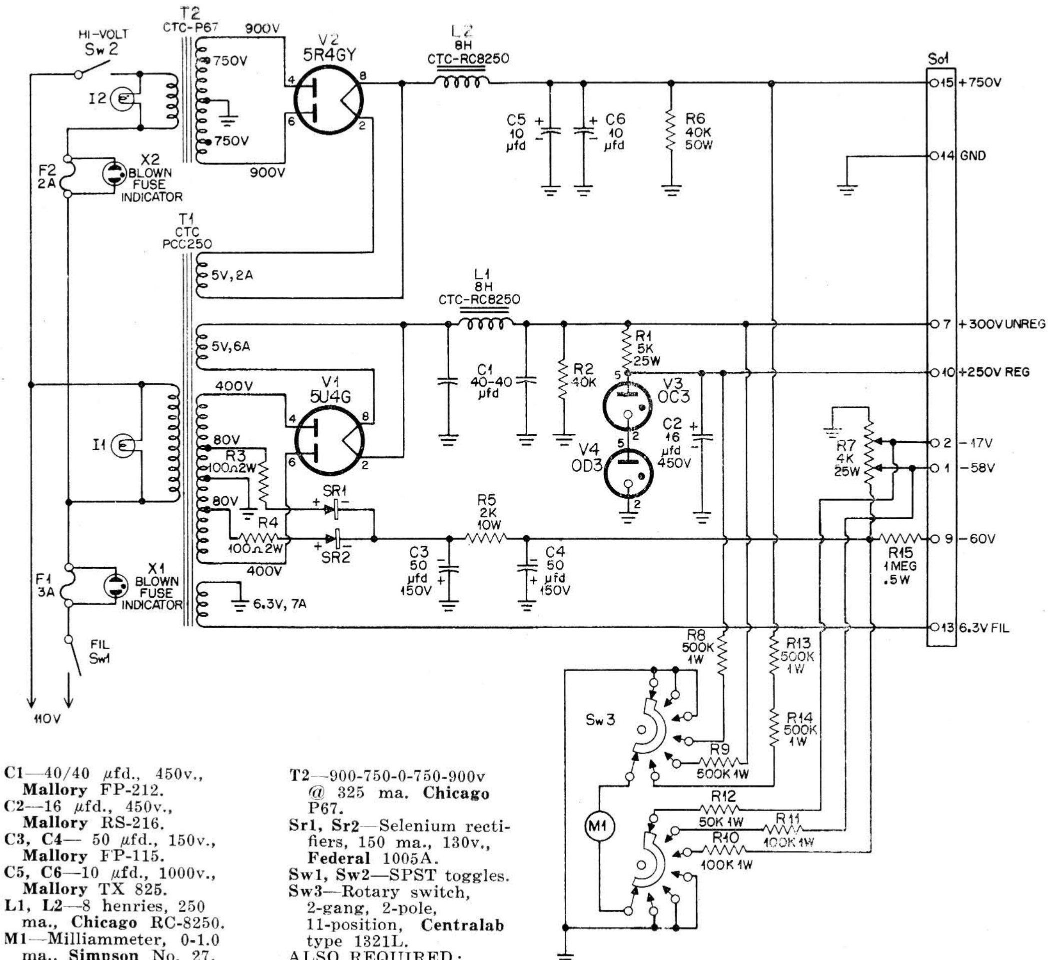
The Low Frequency Crystal Oscillator

This stage uses the conventional Pierce oscillator circuit. The only precaution that the constructor must take is to provide enough feedback to allow the low-frequency crystals to oscillate, and yet not too much feedback so that the crystals are damaged. The circuit parameters shown in *Fig. 8-2-A* have proven satisfactory for several different crystals of the war surplus and new commercial types. Condensers, *C35* and *C39,* determine the amount of plate-to-grid feedback. A portion of the plate-to-ground r-f voltage is fed back in series through the crystal and appears between the control grid and ground. If the capacity of condenser *C35* is increased, the impedance of the grid-to-ground circuit is decreased and thus the feedback voltage is also lowered. The reverse is of course true. If the crystal used is a normal active unit, it should oscillate with the values shown and the plate voltage indicated. The r-f voltage at the plate of the oscillator should be between 10 and 20 volts.

The selection of the two low-frequency crystals is dictated by the shape of the response curve of the *Collins* mechanical filter. See *Fig. 8-2-B* for the response curve plotted from the filter used and the proper frequency for the two crystals. It can be seen that the proper placement of the carrier frequency should be approximately 20 or 25 db. down on the slope of the filter response curve. This indicates that frequencies of 453 kc and 456.5 kc would be the proper selection for the indicated conditions. When this curve is compared with published *Collins* typical filter curves it was found that they fell close to what the company considers to be the average characteristic. However, each filter used should be checked to determine at what frequencies the "20 db. down" points occur. The crystals actually used were the H-17 type from the *James Knights Company,* Sandwich, Illinois. Appropriate crystals of the surplus FT-241 type may be used if available.

The Phase Inverter

The function of this stage is to furnish a push-pull output voltage to drive the balanced modulator. This "double-ended cathode follower," as some people call it, furnishes two equal output voltages which are 180 degrees out-of-phase with each other. The gain of the



C1—40/40 μfd., 450v., **Mallory** FP-212.
C2—16 μfd., 450v., **Mallory** RS-216.
C3, C4— 50 μfd., 150v., **Mallory** FP-115.
C5, C6—10 μfd., 1000v., **Mallory** TX 825.
L1, L2—8 henries, 250 ma., **Chicago** RC-8250.
M1—Milliammeter, 0-1.0 ma., **Simpson** No. 27.
R1—5000 ohms, 25w., wire wound, adjustable.
R2—40,000 ohms, 10w., wire wound.
R3, R4—100 ohms, 2w.
R5—2000 ohms, 10w., wire wound.
R6—40,000 ohms, 50w., wire wound.
R7—4000 ohms, 25w., wire wound, adjustable.
R8, R9, R13, R14—500,000 ohms, 1w., see text.
R10, R11—100,000 ohms, 1w., see text.
R12—50,000 ohms, 1w., see text.
R15—1.0 megohm, ½w.
T1—Transformer, 400-80-0-80-400v. @ 250 ma., 5.0v. @ 6 amps., 5.0v. @ 2 amps. 6.3v. @ 1 amp. **Chicago** PCC-250.
T2—900-750-0-750-900v @ 325 ma. **Chicago** P67.
Sr1, Sr2—Selenium rectifiers, 150 ma., 130v., **Federal** 1005A.
Sw1, Sw2—SPST toggles.
Sw3—Rotary switch, 2-gang, 2-pole, 11-position, **Centralab** type 1321L.

ALSO REQUIRED:
2—Dial light assemblies, one red, one green, **Dialco** #75.
2—Blown fuse indicators, **Dialco** series 95408 with NE-51 neon bulbs.
1—Chassis, 10"x17"x3", **ICA** #29017.
1—Panel, 10½"x19"x⅛", **BUD**, PA-1106.
1—Pair, steel chassis mounting brackets, **BUD**, MB448.
4—Octal sockets.
1—Receptacle, a-c power, **Amphenol** No. 61-F11.
1—Socket, 15 contact, female, **Jones** types S315AB.
1—Plug, 15 contact, male to match above.
2—Fuse holders, **Littelfuse** 342001.

Fig. 8-5-A. Power supply wiring schematic and essential parts list

stage is less than unity, so with the balanced modulator load removed (*C40* and *C41* disconnected) the r-f voltages from cathode and plate to ground of *V3b* should be almost equal and should be slightly less than the r-f voltage from control grid to ground. This unloaded voltage is usually about 10 to 15 volts. When the balanced modulator is connected to the two output terminals, the voltage will drop considerably—usually to about 2 or 3 volts rms. This will depend to a great extent upon the position of the balancing potentiometer arm, *R45*. If the arm is in the middle of the range, the voltages should be as indicated.

The Filter

Little more can be said about the mechanical filter unit. Fortunately all the hard work and misery of alignment has been taken care of by the manufacturer. It should be mentioned that isolation of the input and output circuits of the filter must be realized or the selective characteristic of the filter will be ruined by "leak-around." This creates the effect of filter failure causing the suppressed sideband to reappear. A small metal shield was placed between the input and output terminals and care was taken to route signal and power wiring so that stray pick-up by these leads would not act as a signal-shunt around the filter.

Condenser, *C14* and *C15,* that are across the input and output terminals of the mechanical filter respectively are necessary to properly resonate the input and output tuned circuits of the filter. The actual capacity sizes were those recommended by *Collins* for the particular filter used. If a different model filter is used, consult the *Collins* literature for the correct tuning capacities.

The Carrier Insertion Amplifier

The term "amplifier" is strictly not valid here because *V4b* operates as a cathode follower and thus has a stage gain of less than unity. However, it does provide isolation between the crystal oscillator stage and the grid of the i-f amplifier where the carrier signal is fed. The potentiometer, *R47,* provides for sufficient carrier voltage so that full exciter output can be realized for tune-up or CW operation.

The isolation resistor, *R24,* prevents the relatively low output impedance of the cathode follower from loading down the grid circuit of the i-f amplifier or loading down the filter output impedance and changing its selective properties.

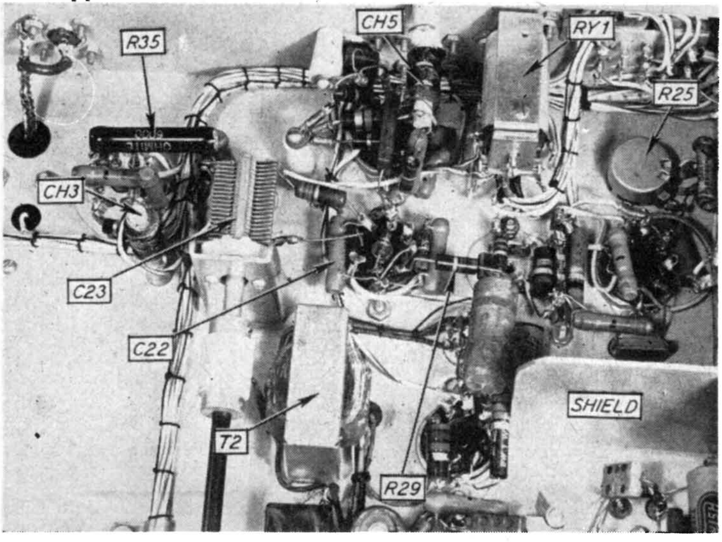
The I-F Amplifier

This stage is very conventional and little need be said concerning its construction and lay-out. The normal precautions were taken to use plenty of good low-inductance bypassing (ceramic condensers) and to keep the input and output circuits isolated. The stage as shown gave no signs of instability for any setting of cathode gain control, *R25.* This stage is necessary to bring the SSB signal that is present at the mechanical filter output terminals up to a useful level for heterodyning. The r-f voltage that appears at the filter output terminals is in the vicinity of 0.05 volts and must be amplified so that the next stage can function properly. The full gain capabilities of the 6SH7 tube, *V7,* are not needed. However, it was found that the most linear operating condition for the i-f stage was near full gain setting of *R25,* so the swamping resistor, *R28,* across the secondary of i-f transformer, *T4* was needed to bring the signal voltage at the mixer down to a safe value.

The Mixer Stage

The familiar 6SB7Y tube is used in the mixer stage that heterodynes the 455-kc. signal up into the 75-meter phone band. This is accomplished by mixing the SSB signal with the v-f-o signal furnished by the Clapp oscillator, *V9.* The circuit in the plate of the mixer tube is tuned to the sum of the v-f-o and 455-kc. signals. This circuit is tuned from the front panel and should be peaked when greatly altering frequency in the 200-kc. wide band. The slug in the plate coil is set so that the air variable condenser, *C23,* is turned to maximum capacity when the exciter is operating on 3800 kilocycles. This was done because if more tuning capacity were available on the control, it might be possible to tune the plate tank circuit to the frequency of the VFO which is *not* the signal desired at the output of the transmitter. The successful attenuation of the v-f-o signal depends solely on the selective properties of the tuned circuits from the mixer plate circuit on through to the exciter output.

The oscillator injection voltage fed to grid #1 *(pin #5)* of *V8* should be in the region of 6 to 10 volts rms as measured on an r-f vacuum tube voltmeter. The signal voltage into signal grid *(pin #8)* should not exceed 0.25 volts peak. As mentioned previously, *R28* was placed across the secondary of *T4* in order to limit the 455-kc. signal level to a safe value. If the value of resistance shown is not low enough to keep



Under chassis view showing the part of the chassns occupied by the i-f amplifier, mixer and 4-Mc. amplifier.

the SSB signal voltage within the limits mentioned, the resistance should be lowered still more. Resistor *R29* is a parasitic suppressor that has been found necessary in similar applications.

The plate voltage and screen voltage to this stage are keyed by the voice control relay. As tried originally, the plate and screen were swung negative during voice control stand-by periods. The mixer tube did not function properly under these conditions for when full plate and screen voltage were reapplied, when the relay was actuated, the stage did not unblock for two or three seconds. Thus the first few words were not transmitted. When the negative voltage feature was removed during stand-by periods, the stage functioned normally and responded quickly to any input signal. The blocking action was apparently a space charge effect in the tube because there were no *R/C* circuits present that could account for the long delay observed.

The Variable Frequency Oscillator

The first requirement for v.f.o. for a single sideband transmitter is stability. The stability normally encountered in the usual AM-transmitter v.f.o. with the tolerant tuning and bandwidth of the shack receiver just isn't good enough for SSB operation. The ever present cry of the SSB roundtables is, "Who's off frequency?," or "Hey, Joe, you're drifting, come back on frequency."

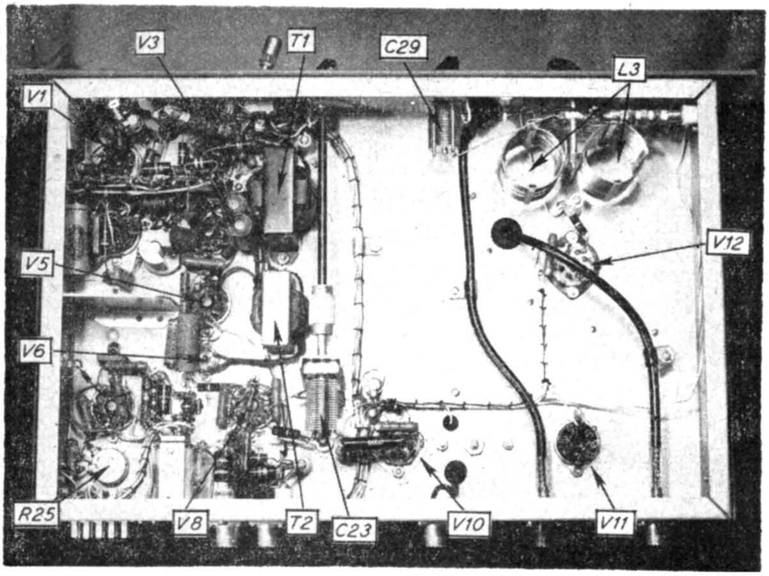
The Clapp oscillator with its freedom from plate voltage and loading effects is a natural solution to the problem. There is also the question of temperature effects on the tuned circuit to be considered. It was decided to take the easy way out of this last situation and remove the tuned circuit and its temperature sensitive components from any source of heat. As Long[23] pointed out, it is possible to remote the Clapp tuned circuit by use of coaxial cables. The cable capacity adds to that of the voltage dividing condensers, *C49* and *C50*. The cable length used with the exciter shown was 12 feet so that the exciter could be mounted in the main station rack and the v-f-o box mounted on the operating table near the receiver. The cables can be made any convenient length. The change in cable capacity with changes in length will have to be compensated for by the bandset capacitor, *C47*. CAUTION! Do not defeat the purpose of remoting the tuned circuit by placing the v-f-o box on top of the receiver or near any other heat radiating piece of equipment.

Resistor, *R50*, is a parasitic suppressor that prevents the stage from oscillating at usually a low frequency dependent upon the r-f choke, *Ch-5*, and the circuit capacities. If such an oscillation takes place, it can be recognized by the multitude of rough-sounding spurious "birdie" signals that appear either side of the fundamental v-f-o signal. Increasing the size of *R50* or shunting *Ch-5* with a resistor of a few thousand ohms will usually cure this trouble.

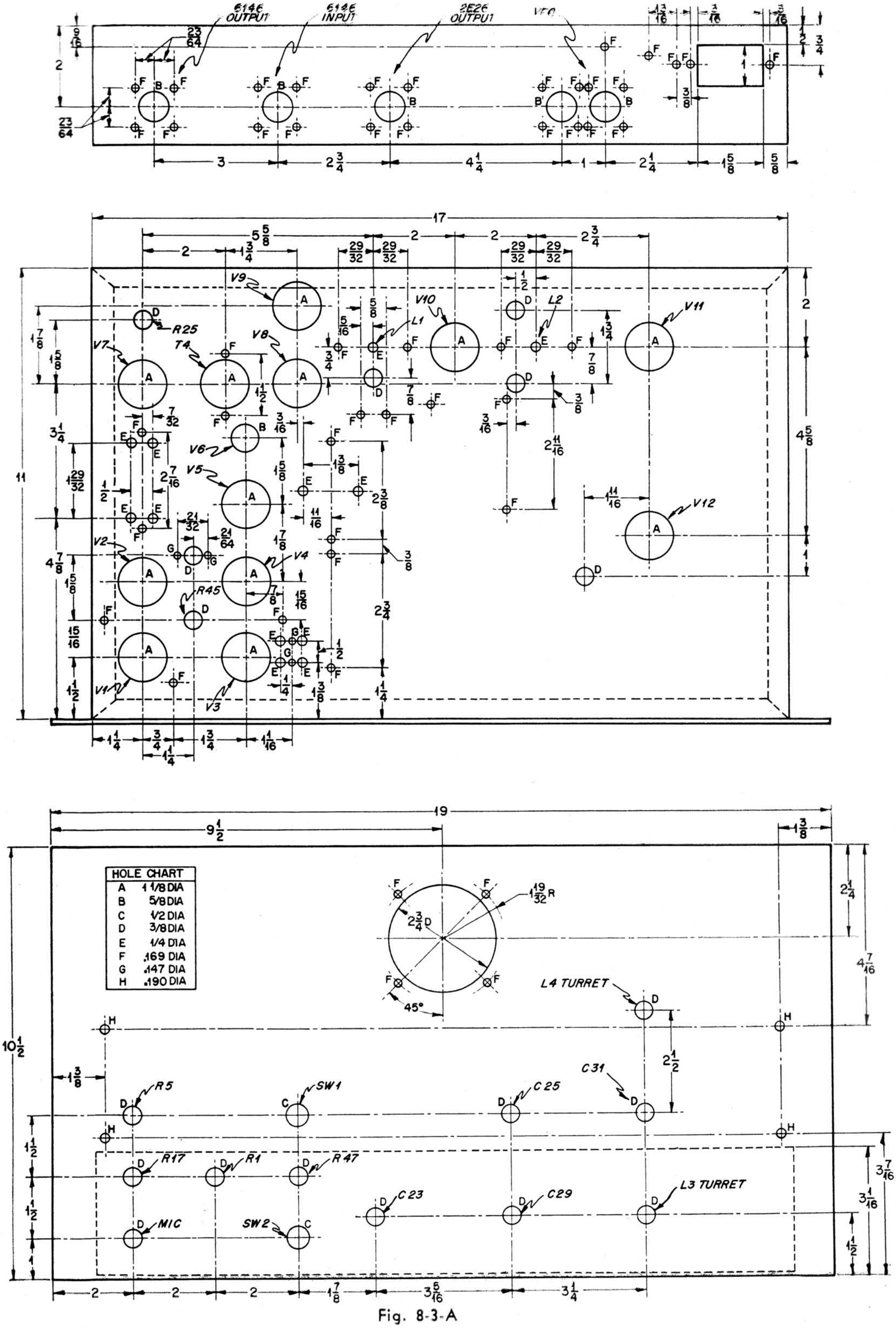
The plate supply for the v.f.o. is the 250-volt regulated source from the power supply chassis. This v.f.o. performed very well and resulted in excellent stability for SSB operation.

The 4-Mc. Amplifier

A type 2E26 tube is used in this stage of linear amplification. This particular tube was chosen in preference to the 6AG7 and others of similar characteristics because it has exhibited more stable operating tendencies than the other types tried. The fact that the plate connection came out the top of the envelope making it more convenient to isolate plate and



Full under chassis view showing location of main components.



HOLE CHART	
A	1 1/8 DIA
B	5/8 DIA
C	1/2 DIA
D	3/8 DIA
E	1/4 DIA
F	.169 DIA
G	.147 DIA
H	.190 DIA

Fig. 8-3-A

MATCHMASTER

Models 650 and 651

A Dummy Load, R-F Watt Meter, SWR Bridge, All in One

Here's the instrument you asked for. And once you've tried it, you'll wonder how you ever got along without it. It provides, in one completely self-contained cabinet, 6" x 8" x 8",—

A Dummy Load—for all kinds of tests on your transmitter without putting a signal on the air. Maximum SWR 1 to 1.2 over a frequency range of 300kc to 30mc.

A Direct-Reading R-F Watt Meter—for precise adjustments of all r-f stages up to 125 watts, and even higher powers by sampling. Excellent repeat accuracy over full 125 watt scale.

Integral SWR Bridge—for matching antennas and other loads to your transmitter, giving you precise adjustment of beam antennas, antenna tuning networks, and mobile whip antennas.

Controls — including a 3-position function switch, and a meter adjusting knob—are conveniently grouped on the attractive, silk-screen-gray front panel, which also contains a 3-inch calibrated meter, and Type SO239 input and output connections. The ventilated steel cabinet is finished in attractive blue Hammertone. Two types are available;

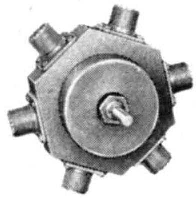
Model 650: 52-ohm line–Model 651: 73-ohm line

For details, write for descriptive Bulletin 650.

AUDIO PHASE SHIFT NETWORK

Type 2Q4—Model 350

This octal based, audio phase shift network provides a constant 90° phase shift, ± 1.5°, over the audio range of 300 to 3000 cycles, yet requires no more space than a 6J5 tube. Designed especially for single sideband receiving and transmitting applications.



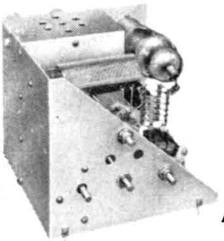
MULTI-POSITION COAXIAL SWITCH

Model 550

Takes The Mess Out of Switching Circuits

At last you can have an inexpensive, multi-position coaxial type switch—for selecting antennas . . . transmitters . . . exciters . . . receivers . . . and other r-f generating devices using 52-75 ohm coaxial line—without fumbling or breaking your back trying to screw and unscrew connections. This B&W Model 550 is equipped with six SO239 type connections for selecting any one of five 52 or 75 ohm lines. It will handle 1kw of modulated power with a maximum crosstalk of —45db at 30mc. Housed in a heavy, 4" diameter aluminum case, the Model 550 is made for single hole mounting.

ON THE



MULTI-BAND FREQUENCY MULTIPLIER

Model 504C

Gives You Any Band At The Flip of a Switch

Here is a newly conceived and designed exciter unit that makes transmission on any band available at the flip of a switch. Compact in its 8" x 7" x 9½" size, the Model 504C covers the 80 through 10 meter bands with a nominal power output of 25 watts from the 807 amplifier stage through a flexible pi-network output circuit. Its broad band type amplifiers require no tuning, and the unit comes equipped with four 6AQ5's that make up its multiplier string. An external VFO or crystal oscillator (80 meter fundamental) is required, as well as a suitable power supply. Sturdily constructed of heavy gauge frosted aluminum, the Model 504C also makes an ideal basic mobile foundation unit for multi-band operation.

PRECISION TOROIDAL TYPE SSB BANDPASS FILTER

Model 360 and 361

Here is a precision bandpass filter valuable for use in heterodyne type sideband generation. Containing eight toroidal type coils in an LC type filter, it is designed to pass the frequencies 16.9 to 20kc. Extreme skirt attenuation. Two types are available: a receiving type (Model 360) for 20,000 ohm input and output; and a universal transmitting or receiving type (Model 361), for 20,000 ohms input and an output of 20,000 ohms unbalanced, plus two 500 ohm balanced outputs. Both types are precision adjusted and housed in hermetically sealed, tinned steel cases measuring 2⅝" x 2¼" x 3¾", exclusive of mounting studs and terminals. Write for Bulletin 360.

BARKER & WILLIAMSON,

AIR WITH B&W

SINGLE SIDEBAND GENERATOR — Model 51SB

For Use With B&W Model 5100 Transmitter

Now, for the first time, you can get really sparkling performance on either SSB, AM phone, or CW. This B&W Single Sideband Generator teamed up with the famous Model 5100 Transmitter gives you outstanding SSB operation on all frequencies provided in the 5100. Tuning and operation are a breeze. No test equipment is required. Single sideband signal is generated by a simple and efficient method perfected after two years of extensive research and testing by B&W engineers. No stone has been left unturned to give you such extras as voice operated and push-to-talk controls, a speaker deactivating circuit, TVI suppression, and unitized construction for quick and easy removal of any major section. Completely self-contained, the 51SB requires no more external accessories than a microphone.

Combine this Single Sideband Generator with the features of your Model 5100—150 watts input on SSB and CW, 135 watts on AM phone; VFO or crystal operation; pi-network final—and you've got a combination that will flutter the heart of the most critical operator. The 51SB cabinet is made to bolt right onto the 5100 cabinet, extending the 22-inch length to 32 inches. Distinctive panel styling and appointments are the same for both. Easy to install, the 51SB comes factory wired and tested, complete with tubes and all necessary components to convert your Model 5100 Transmitter to SSB. This combination provides a superlative driver for *any* hi-powered linear amplifier. Write for Bulletin.

Inc. 237 Fairfield Avenue
Upper Darby, Pa.

These are just a few of the hundreds of products especially designed and built by B&W to meet the needs of the radio amateur. Others are described in Catalog 2PC available upon request. Write for your copy.

grid circuits was thought to be a good recommendation. The maximum allowable plate voltage and dissipation also made it attractive in case it became necessary to call upon it for a little more output signal to drive the external heterodyne unit.

The grid is capacitively coupled to the preceding mixer stage through *C57*. A small amount of fixed grid bias (minus 17 volts) is furnished from the exciter bias supply on the power supply chassis. It was found that the stage operates slightly into the class AB_1 region and therefore ordinary cathode resistor bias could not be used. The varying plate current under signal conditions would cause the grid bias to vary as would the stage's operating point on the tube characteristic curves. For this reason it was also necessary to regulate the screen voltage with a separate voltage regulator tube. The stage idling plate current should be about 25 ma.

The plate tank circuit is tuned to the 4.0-Mc. phone band. A four-turn link winding is placed around the "cold" end of the tank coil for output to the external heterodyne unit or to the grid tank of the 6146 output stage. Resistor, *R36* (two, 200-ohm, 2-watt carbon resistors in parallel), was found to be necessary to swamp a tendency toward oscillation in the stage. Since this stage is capable of over two watts output, the power lost in the swamping resistor does not seriously affect the drive to the 6146 tube. There is still plenty of grid voltage swing available to operate the tube conservatively.

The link winding is connected to a coaxial fitting on the back of the chassis so that it may be either patched to the companion fitting beside it for straight-through operation on 4.0 Mc. or connected to the input terminal of the external heterodyne unit for multi-band operation.

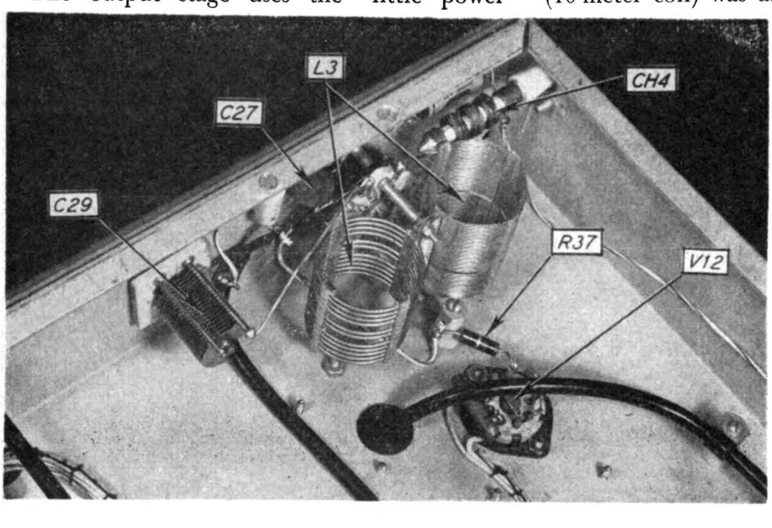
The Power Amplifier

The output stage uses the "little powerhouse" 6146 tetrode in class AB_1 Strangely enough no difficulty was encountered in this stage. The stage was completely stable and showed no signs of parasitic oscillation.

The grid bias is again furnished from the minus supply in the power supply chassis. The plate current is adjusted for an idling value of 35 ma. by adjusting the slider on resistor *R7* *(Fig. 8-5-A)* in the power supply chassis. For the two or three tubes tried, the average value of grid bias came out to be 58 volts for a d-c plate voltage of 750 volts with 250 volts on the screen grid.

The grid tank circuit utilizes a *Barker & Williamson* band-switching unit called the *"Band-Hopper," Type 2A*. This tapped inductance unit tunes all bands from 80 through 10 meters. It was found that it was necessary to use the band switch in the next higher-frequency band position than that of the operating frequency so that the correct *L/C* ratio could be realized. The coupling to the grid tank is accomplished by running the low-impedance line input cable in series with the bottom of the grid coil to r-f ground. Condenser, *C27*, prevents the d-c path from being shorted to ground by the link while *RFC4* allows the grid bias to be applied without loading down the input circuit. This system of coupling worked quite well and did not appear to reflect any tuning changes back into the preceding tank circuit as the grid tuning condenser *C29* was rotated through resonance.

The plate tank circuit uses a *B&W type BTEL* coil turret. The correct plate load impedance for a single 6146 requires that about 110 μμfd. be used on the 75-meter band. This is accomplished when the BTEL unit is switched to the 40-meter position and tuned to resonance at 75 meters. The 80-meter coil was removed from the bandswitch for space reasons. Since operation was not contemplated above 20 meters, the highest frequency coil (10-meter coil) was also removed from the tur-



Under chassis view showing mounting arrangement and placement of P.A. grid circuit components.

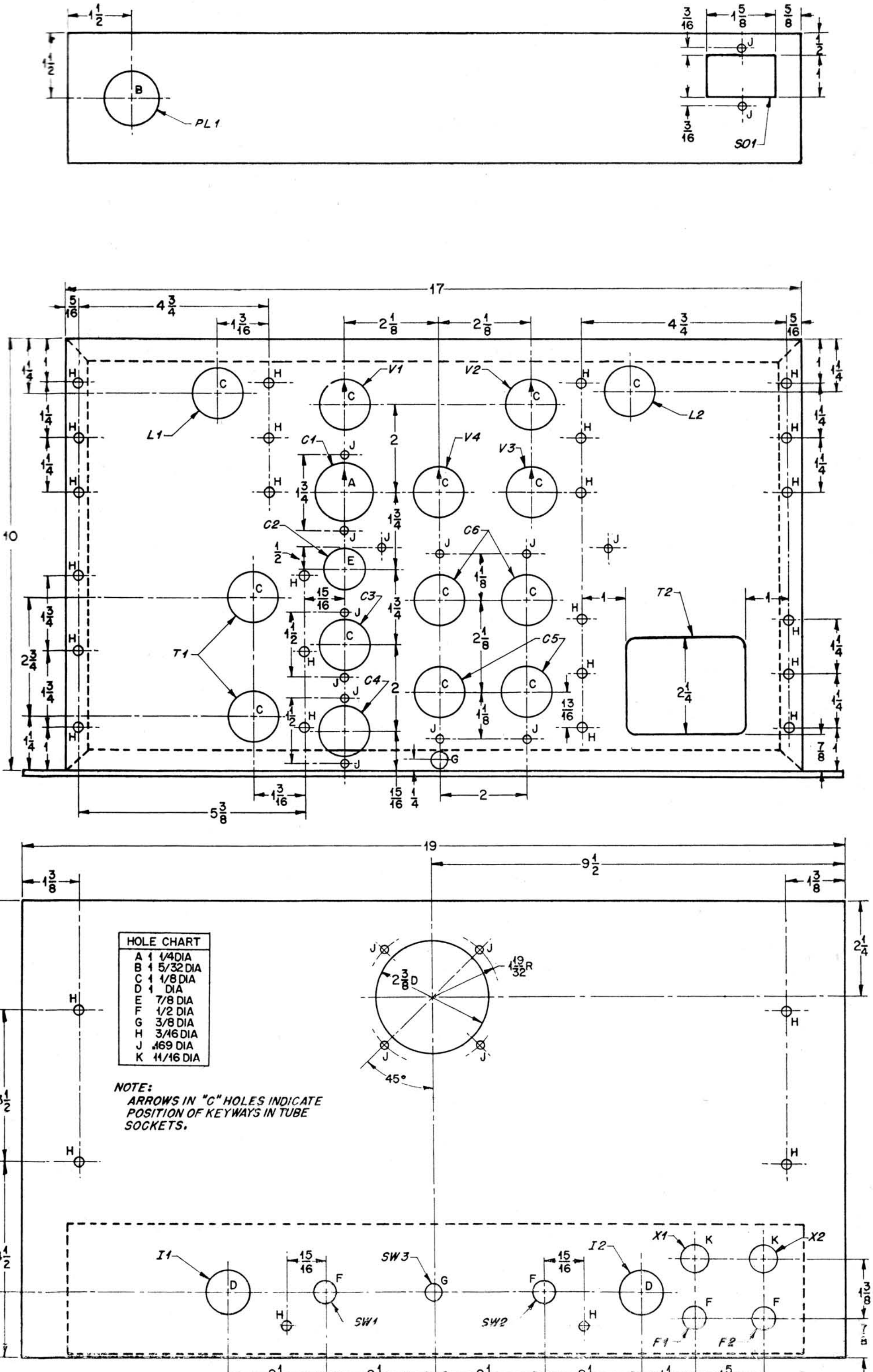
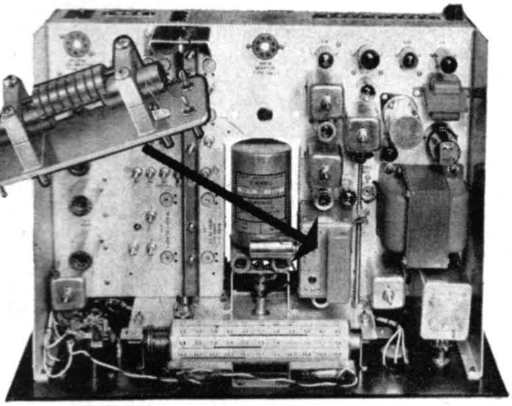

HOLE CHART
A 1 1/4 DIA
B 1 5/32 DIA
C 1 1/8 DIA
D 1 DIA
E 7/8 DIA
F 1/2 DIA
G 3/8 DIA
H 3/16 DIA
J .169 DIA
K 11/16 DIA
NOTE:
ARROWS IN "C" HOLES INDICATE
POSITION OF KEYWAYS IN TUBE
SOCKETS.
PL1
SO1
L1
L2
V1
V2
V3
V4
C1
C2
C3
C4
C5
C6
T1
T2
I1
I2
SW1
SW2
SW3
X1
X2
F1
F2

Fig. 8-5-B

FILTERS in SSB

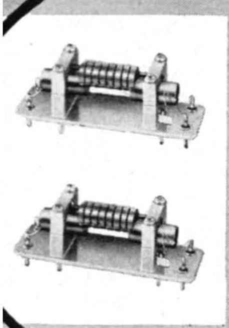


SSB RECEPTION WITH 75A-3 RECEIVER

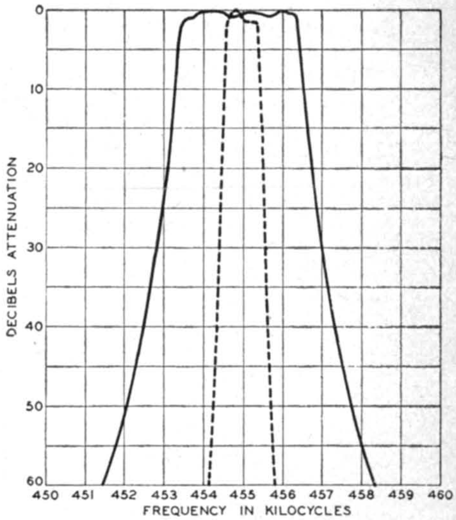
Satisfactory SSB reception requires high stability, and the 75A-3 has it; quartz crystals, a different one for each band, control the h-f oscillator while the low-frequency injection voltage is supplied by the famous Collins permeability-tuned oscillator.

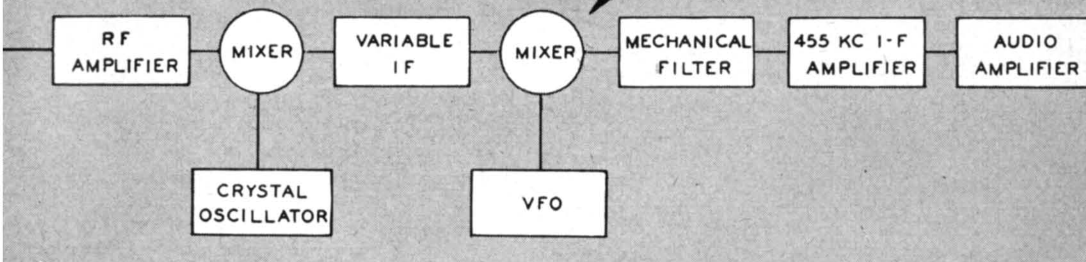
But stability alone would be inadequate. Stability *plus selectivity* is required, and the selectivity curve in the 75A-3 is ideally suited to SSB reception.

With the 75A-3, you don't have to nurse SSB signals in with the b.f.o. Just set the b.f.o. knob for your choice of either sideband, then use only the main tuning dial.



Selectivity curves of 455 kc Mechanical Filters with nominal 0.8 kc (dotted line) and 3.1 kc (solid line) bandwidths at -6 db. Also available are Filters for 6.0 and 1.2 kc bandwidths.





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Collins Radio Company of Canada, Ltd., 74 Sparks St., OTTAWA, ONTARIO ret. This left the 40, 20, and 15-meter inductances used on 80, 40, and 20 meters respectively. Condenser, *C32,* is a d-c blocking condenser to prevent the plate voltage from appearing across the plates of the tank tuning condenser, *C31.* It was found in tests that the peak r-f voltage plus the d-c voltage was enough to cause arc-overs in the tuning condenser especially if the loading was a little light.

Since the *B&W BTEL* turret has a fixed link arrangement it would be advisable to use some form of antenna coupling device to insure a proper amount of loading of the 6146 output stage. If the unit is to be used as a driver for a high-power final amplifier, the variable loading feature can be taken care of in the next tuned circuit which would be the final grid tank. This could be accomplished by using a variable link coil in the high-power final grid circuit.

8.3—Exciter Mechanical Lay-Out

Figure 8-3-A shows the detailed sketch of the chassis and panel lay-out. While many of the features are self-explanatory, a few comments on certain items will be outlined to emphasize or clarify their importance.

The lay-out plan used resulted in a grouping of all the audio stages in the left-front corner of the chassis (as viewed from the front panel). The rather large number of components resulted in a congested array of circuitry. This part of the chassis was nicknamed "confusion corner" for rather obvious reasons. The underside view of the exciter chassis will give the reader a fair idea of how crowded things came out. The insulated tie-strip which runs parallel to the front panel about two inches back from the front chassis wall is mounted above the tube sockets on 1½-inch bushings. This tie-strip supports all of the plate load resistors and plate decoupling resistors for tubes *V1, V2, V4,* and *V5.* If reasonable care is taken in wiring in the various parts, a relatively neat arrangement can be made despite the number of components per square inch that have to be installed. No difficulty with hum or r-f pickup was experienced with the lay-out used.

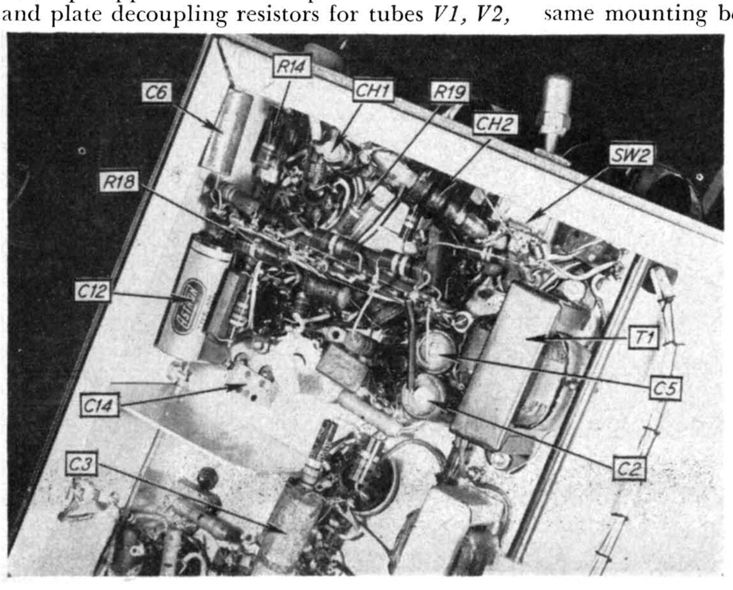
A small shield must be installed separating the input and output terminals of the *Collins* mechanical filter, *Fl-1.* This should be about 3 inches in length and should extend for the full depth of the chassis—also 3 inches. As pointed out earlier, this is to prevent leakage around the filter of any stray signal or carrier voltages.

An effort must be made to keep all bypassing lead-lengths as short as possible. Tubular ceramic r-f bypass condensers were used throughout.

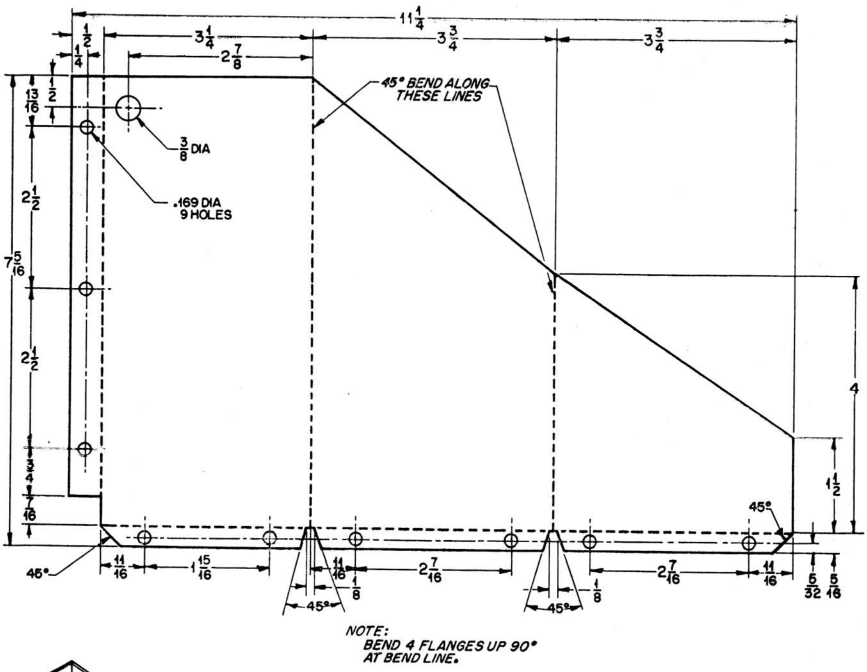
The connections between the sideband switch, *Sw-2,* and the two low-frequency-crystal holders *(Millen #33302)* should be kept as short and direct as possible. These leads should not be cabled up with other power and audio leads. This will save trouble in arriving at a satisfactory carrier balance and also keeps r.f. out of the speech equipment.

The shielded lead from the carrier insertion potentiometer, *R47,* to the grid of the i-f amplifier, *V7* will logically run near and past the mechanical filter position. For this reason it was necessary to shield this wire. In the exciter, a B+ wire also was routed past the filter and was also shielded to prevent leakage around the filter. Both of these shielded conductors were run along the bottom flange of the chassis to keep them as far away from the filter terminals as possible.

Audio transformers *T2* and *T3* are physically mounted one over the other. One of the transformers is mounted above the chassis and the other is mounted below the chassis using the same mounting bolts. It is unimportant which transformer occupies which position as the leads for either position are about the same length.



Under chassis view showing arrangement of parts in the speech amplifier and voice control stages.



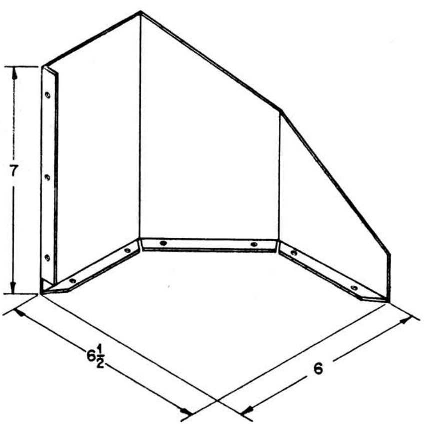


Fig. 8-3-B. Layout and bending of the inter-stage shield used to isolate the 2E26 from the 6146 stage.

The orientation of the octal-tube base key positions were found to be the most satisfactory for direct wiring and shortest leads.

The small letters that appear in each hole outline are the reference letters (or key) for the hole diameter shown in the tabulated box for *Fig. 8-3-A*.

An effort was made to keep critical grid and plate circuits isolated from each other. The policy followed was to keep one circuit above the chassis and the other below the chassis. This was accomplished in all cases with the exception of the placement of the 2E26 plate circuit and the 6146 plate circuit. Since the 2E26 plate tank is coupled to the grid tank of the 6146 stage (on 75 meters) this in effect puts both the grid and plate circuits of the 6146 above the chassis. An inter-stage shield was constructed that successfully isolates the two stages. See *Fig. 8-3-B* for a detailed sketch of the shield used.

The front panel lay-out of controls is shown in *Fig. 8-3-A* in which the symbol identification as referred to *Fig. 8-2-A* is used. A different control arrangement may be used if care is used in keeping the r-f switching and control circuit wiring as short as possible. Whatever panel lay-out is used, it will always improve the appearance of the unit if control identification marking is used. Unfortunately, all the word-markings needed for a single-sideband exciter are not usually available in the commercial kits, but with a little patience, and not a little manual dexterity, individual letters may be assembled to form the desired words.

The large cans used for shields around coils *L1* and *L2* are *National Co. Type RO* units. The screw adjustments for the two *National XR-50* slug-tuned coil forms come out underneath the chassis. Once these slugs are set no further adjustment is necessary, therefore there is no need to make them available above the chassis.

The back wall of the chassis has five coaxial cable sockets and a multi-terminal power socket visible. The two right-hand co-ax sockets (as viewed from the rear) are for the two coaxial cables of the remote-tuned Clapp v-f-o circuit. A multi-pin single-plug and socket arrangement could have been used just as well. The middle coaxial socket is the 2E26 output terminal. This will be connected to the socket immediately to the left for 75-meter operation or connected through a coax line to the heterodyne unit described in *Chapter IX*. The co-ax socket next to the 2E26 output is the 6146 grid input terminal. The far left socket is the 6146 output socket.

A bottom chassis cover plate is recommended particulatly if the unit is mounted in the same rack with a high-power final amplifier where there might be a chance of r-f feedback into the low-level, high gain stages.

8.4—The Remote VFO Box

The box used to house the remote tuned circuits of the Clapp oscillator was a standard 4″x5″x6″ unit. The coil form used was a *National XR-16* mounted on sturdy bushings so as to be physically mounted in the center of the box. The bandspread condenser is mounted so that it aligns with the coupling of the *Millen #10039* dial assembly. The photo of the inside of the v-f-o box shows these details clearly (see *Fig. 8-4-A*). All mechanical mountings must be sturdy as any mechanical vibration will seriously affect the frequency stability. The two co-ax lines that come in the back of the box must be anchored firmly so that any movement of the external lengths of the two cables will not move anything inside the box and cause a frequency shift.

8.5—The Power Supplies

Figure 8-5-A shows the schematic for the power supply unit for the exciter.

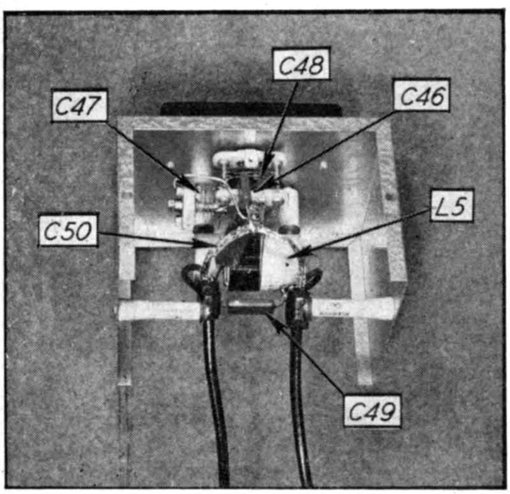


Fig. 8-4-A. Inside view of the remote v.f.o. control box. Details on the box and dial appear in the text above.

The power supply chassis contains three power supplies: (1) the low voltage supply (350 volts unregulated and 250 volts regulated), (2) the bias supply (minus 60 volts for the 6146 and minus 17 volts for the 2E26) and (3) the high voltage supply (plus 750 volts).

The Low-Voltage Supply

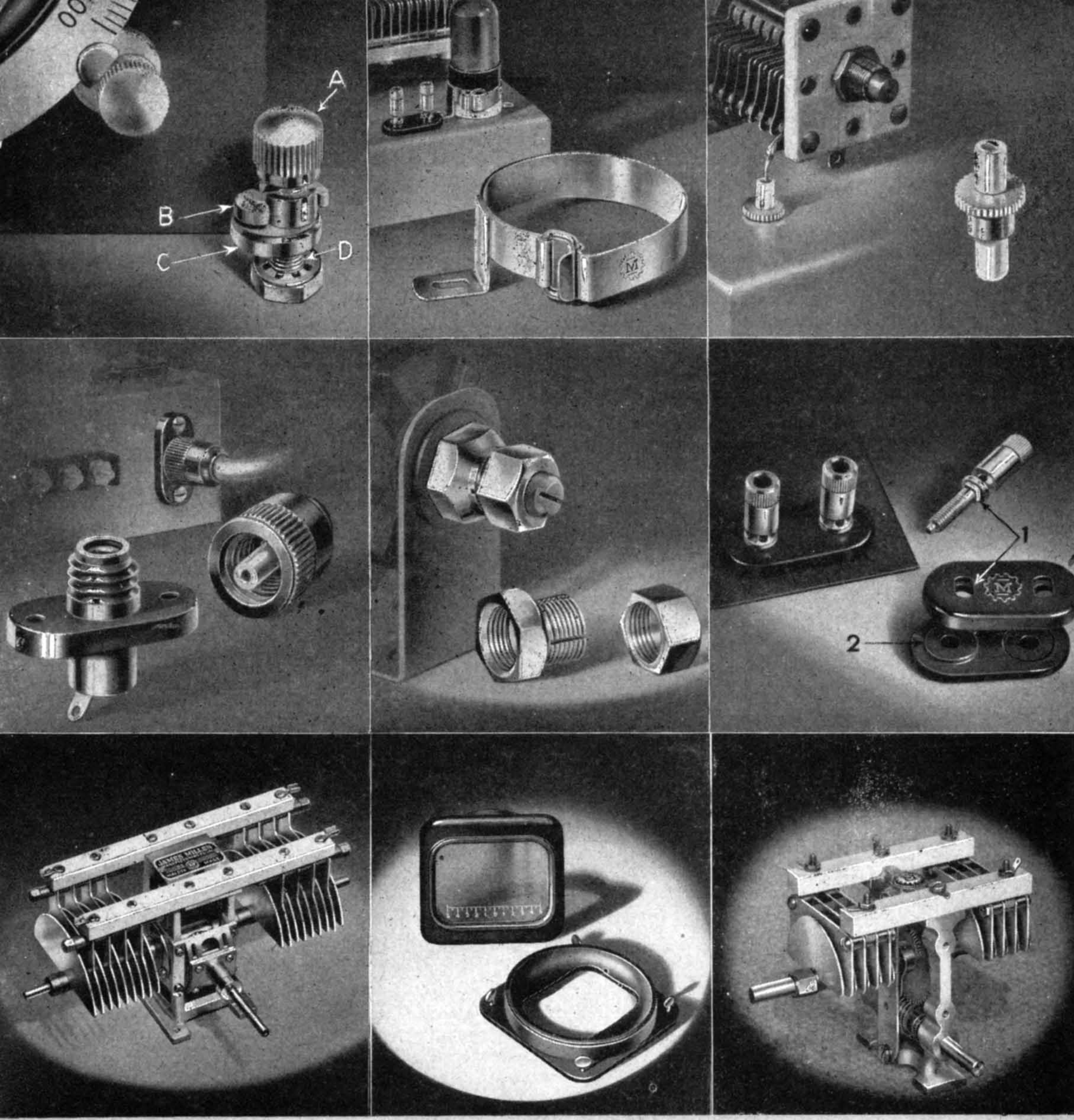
The circuitry of the low voltage supply is quite conventional and little need be said. The dual 40 μfd. filter condenser and the 8-henry filter choke were found to be an adequate filter for the SSB exciter. One thing that should be kept in mind is that the low-voltage unit should not be operated for very long periods of time with the power cable to the exciter chassis unplugged because the no-load voltage of the supply goes up to a value that exceeds the safe operating voltage of the electrolytic can-type filter condenser. The "goo" inside the can will cook out and the unit will lose capacitance and the filtering action will deteriorate. With the exciter cable plugged into the unit the normal idling load of the low-level stages in the exciter will provide sufficient load to keep the supply voltage within safe limits.

The Bias Supply

The *Chicago Transformer Corp.* transformer *(PCC-250)* used for the low-voltage supplies fortunately had taps for a bias supply at 80 volts either side of the secondary center-tap. Selenium rectifiers were used in a full-wave supply. The two 100-ohm resistors, *R3* and *R4*, are current limiting resistors to prevent damage to the selenium rectifiers. The bias supply filter is the *R/C* type in which a 2000-ohm resistor is used in place of the conventional filter choke. The bias supply filter condensers, *C3* and *C4*, are the can-type and thus must be insulated from the chassis on the phenolic mounting wafers provided with them. This places the cans about 100 volts above ground and contact with them and the chassis will provide a mild shock—BEWARE! The tapped resistor, *R7*, has two sliders so that the two bias voltages can be set to the correct value for the 2E26 and 6146 stages.

The High Voltage Supply

The plate transformer for the high voltage supply *(Chicago type P-67)* is rated conservatively and more than covers the requirements for the two exciters. The secondary is tapped for 900 volts and 750 volts either side of the center-tap. The use of choke input on the supply permitted using the 900-volt taps so that the loaded output voltage of the supply was between 750 and 800 volts d.c. The output filter condensers are 10 μfd. at 1000 volts providing a total output capacity of 20 μfd. This is considered adequate filter for the load drawn. The low-voltage power transformer had



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an extra 5-volt filament winding that is used for the high-voltage rectifier. This permits the high-voltage supply to be controlled in the 117-volt primary circuit. The bleeder resistor, *R6*, provides a token bleeding action and serves to discharge the 20-μfd. filter condensers when the supply is shut off.

Switching and Fusing

Both the low-voltage and high-voltage supplies are fused individually and the fuses are connected in parallel with blown-fuse indicators which are nothing more than neon lamps. When the fuse blows, the line voltage appears across the neon lamp and indicates which circuit is open. This feature provides convenient indication when trouble is encountered.

Metering the Output Voltages

Another feature that was included which might be considered unnecessary by many builders was the panel-mounted d-c voltmeter and circuit selector switch. This six-position switch permits monitoring the six voltages supplied by the power supply unit. The feature appealed to the author because of the ease with which trouble could be narrowed down in the event of trouble shooting. The metering of the 6146 bias voltage also provides an indication when the tube goes into the grid current region (shown by an increase in bias voltage). The gain can then be reduced so that the tube operates in class AB_1 with no grid current being drawn.

The resistors, *R8, R9, R10, R11, R12, R13* and *R14,* were selected from a group of resistors so that they had a value as close as possible to the value indicated. This directly affects the accuracy of the voltmeter readings as these resistors are the voltmeter multipliers.

The two series-connected multiplier resistors, *R13* and *R14,* are each 0.5 megohm instead of being a single 1.0 megohm unit because the voltage limit per resistor cannot be exceeded if the resistors are expected to maintain their value. Composition resistors will change their value if exposed to too high a potential. This may be a gradual process affecting the meter accuracy over a long period of time. The voltmeter itself is an 0 to 1 ma. full-scale meter.

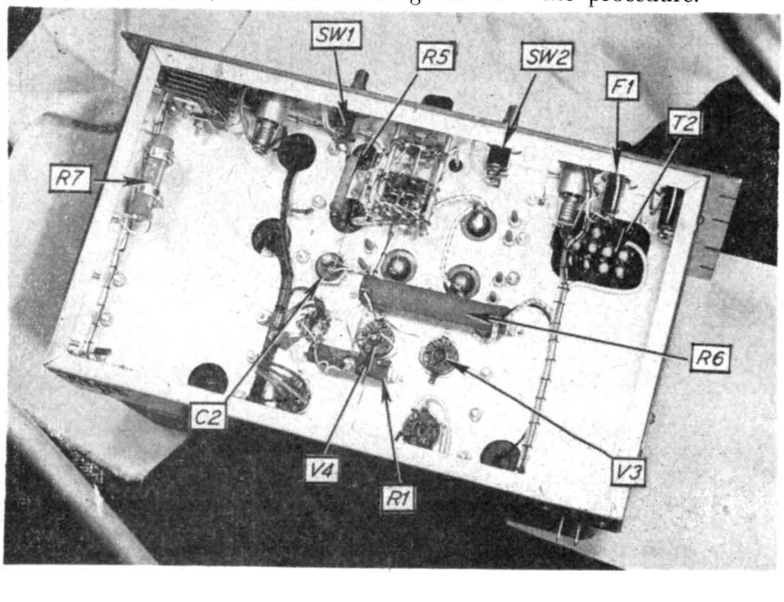
The Power Supply Mechanical Lay-Out

Figure 8-5-B shows the power supply chassis and panel mechanical arrangement. The photographs also give a good idea of just how the unit is arranged. The chassis was given a more or less symmetrical lay-out more for asthetic reasons than for practical ones. Changes in the basic lay-out can certainly be made without affecting the basic operation of the unit.

8.6—Exciter Initial Alignment

Once the exciter is wired and given a careful check for wiring errors, the power supply should be connected through its power cable to the exciter chassis. The 2E26 and 6146 tubes should be removed from their tube sockets for the first phases of the alignment procedure.

A BC-221 frequency meter would be very convenient if available. A receiver should also be available for monitoring purposes. Other test equipment that will be needed are a 10,000 ohm-per-volt d-c voltmeter, and an r-f vacuum tube voltmeter or an r-f probe that could be used in conjunction with the standard d-c voltmeter already mentioned. An oscilloscope would be desirable, but not necessary for the initial phases of the alignment. Actually the exciter has been successfully aligned without an oscilloscope being used for any step of the procedure.



Under chassis view of the power supply unit.

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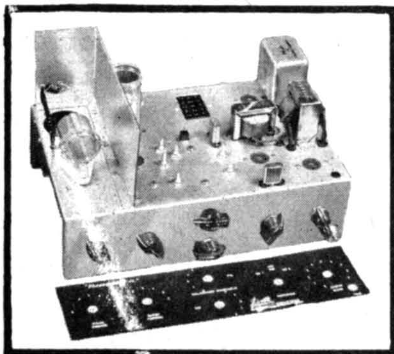

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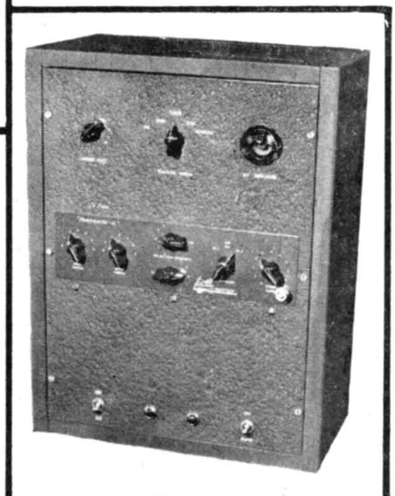

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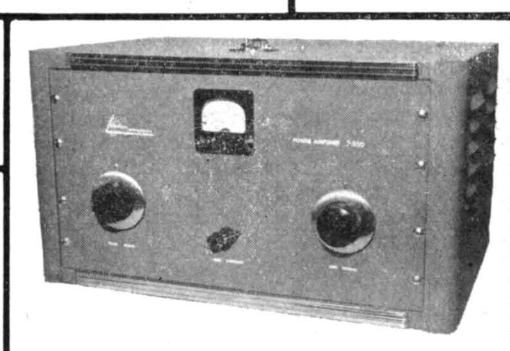

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The first thing that is to be done is to determine if all circuit voltages are as they should be. The power supply output voltages are easily measured with the voltmeter on the power supply unit front panel. The voltmeter should be used to check the voltage at each of the tube plates. The plate voltage measured at the plates of tubes *V1* through *V5* should be between 100 and 150 volts for normal operation.

The Low-Frequency Oscillator

The next step is to determine if the low-frequency crystal oscillator is operating. Using the BC-221, loosely coupled to the plate of the Pierce oscillator, *V3a,* and determine if the oscillator is operating at the marked frequency of the crystal in use. The r-f VTVM may also be used for oscillation indication. If no oscillation is present with either crystal, there apparently is insufficient feedback if the circuit is wired correctly. Lowering the value of *C39* somewhat will increase the feedback to the oscillator grid circuit. Lower this value just enough to give satisfactory operation—too much feedback will damage the crystal unit. Once oscillation is obtained, the frequency of operation should be determined. The BC-221 frequency meter comes in handy here. If surplus crystal units are used they should be carefully checked as many of them will oscillate off-frequency. The r-f voltage from *pin #2* of *V3a* to ground should be between 10 and 15 volts rms. This same voltage should appear at *pin #4* of *V3b* and *pin #1* of *V4b.*

The Audio System

Next determine if the speech channel is operating by monitoring the audio with a pair of high-impedance earphones the signal from *pin #4* of *V2a* to ground. The signal should be clean and free of hum and should have sufficient volume to be heard with the gain control advanced toward maximum. The audio voltage across *R21* should not run over 0.25 volts for normal operation.

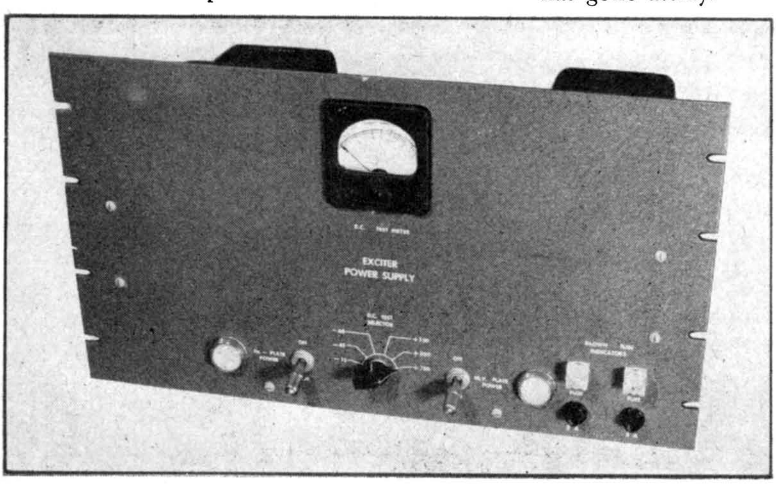
The Balanced Modulator

The only alignment to be made on the balanced modulator is the carrier balance. When the signal is monitored in the BC-221 frequency meter, controls *R45* and *C42* should be adjusted alternately for a minimum of carrier signal as heard in the head 'phones. The capacitive balance control, *C42,* may have to be shifted to opposite end of the balance potentiometer than that shown in *Fig. 8-2-A.* More or less capacity may have to be connected in parallel with the capacitive balance condenser in order to realize a perfect balance. Final balance of the circuit before operating should be made only after the unit has come up to operating temperature.

The I-F Amplifier

Next transfer the BC-221 coupling wire (in series with a 10 μμfd. condenser) to the grid of *V8 (pin #8).* With the BC-221 tuned to 455 kilocycles advance the carrier insertion knob slightly until a weak heterodyne is heard in the headphones. Align the slugs of the primary and secondary windings of the i-f transformer, *T4.* Each of these slugs should pass through a definite maximum signal output condition. Next set the cathode gain control, *R25,* to give a maximum intensity of signal. It will be noted that if the gain control is advanced too far a sharp decrease in gain will result. The gain should be set slightly below this position.

At this point the quality of the generated single sideband signal can be observed. Turn the carrier insertion control to zero and talk into the microphone. The signal may be tuned in using the BC-221. This will give you an idea just how you are progressing. If the signal sounds satisfactory when properly tuned in, you may proceed to the next stage. If the signal is distorted and you are sure that it is not overloading the BC-221 and is not a function of tuning, you should look back over the preceding steps and determine where the signal has gone astray.



Front panel view of power supply unit.

The V.F.O.

The Clapp oscillator is the next order of business. The r-f VTVM should be connected from *pin #5* of *V8* (the mixer) to ground. This is the injection grid of the frequency conversion stage. Chances are that the oscillator is operating. An r-f voltage between 6 and 10 volts should be present at the mixer injection grid. If the oscillator is not operating the cause is generally insufficient feedback. If the circuit oscillates when a screwdriver is used to short the plates of *C48*, the *Q* of the oscillator coil is too low or the feedback as divided across *C49* and *C50* is insufficient to cause the tube to oscillate. The solution is to decrease the size of the two voltage dividers, or at least the size of *C49*. Once oscillation is obtained the station receiver should be used to determine the frequency. If the band-set condenser, *C47*, does not have sufficient range to permit adjusting the oscillator between 3350 and 3550 kc. then slightly modify the coil inductance. Additional fixed capacity may be added across *C46* to pull the frequencies within range. Any additional fixed capacity added to the circuit should consist of silvered mica capacitors so that the unit will remain relatively free of ambient temperature changes.

The bandspread condenser and band-set condenser of the v.f.o. should be now adjusted. With the bandspread condenser, *C48*, set near maximum, adjust the band-set condenser, *C47*, so that the v.f.o. is at 3350 kc. Next determine at what frequency the oscillator operates when the bandspread condenser is set near minimum capacity. This frequency should be above 3550 kc. for complete 75-meter band coverage. In the unit described, one plate was removed from the *Hammarlund* 35-μμfd. bandspread condenser so that the 75-meter band could be spread out over most of the 180 degree coverage of the *Millen #10039* dial.

The Mixer

The mixer stage is the next to be aligned. With the *Voice Control* switch (Sw-1) in the "off" position make the following adjustments. (Turning the *VOX* switch to the "off" position disables the relay and applies normal screen and plate voltage to *V8.)* The station receiver should now be tuned to the "sum-frequency" of the v.f.o. and 455-kc. SSB signal. The mixer output tank circuit should than be peaked to this 75-meter SSB signal frequency. Carrier should be inserted by advancing *R47* about half way. As measured on the r-f VTVM the voltage from *pin #5* of *V10* should be between 5 and 10 volts when the carrier is inserted. With the carrier insertion control at zero a residual meter reading will be noted. This is the 3.5-Mc. v-f-o signal coming through.

The selective properties of the one tuned circuit are not great to completely discriminate against the v-f-o signal. As mentioned previously, the slug of coil *L1* should be adjusted so that condenser *C23* is set at almost maximum when the SSB signal is at the 3800-kc. end of the 75-meter band. This prevents accidental tuning of the mixer tank to the frequency of the v.f.o.

The 2E26 Stage

Now plug the 2E26 tube *(V10)* into its tube socket. If the precautions indicated in the diagram and in the preceding text have been taken, the stage should not oscillate. Oscillation can be detected by tuning the 2E26 output tank circuit condenser while listening in on the appropriate frequency range with the receiver. If a strong rough "birdie" is heard, oscillation is present. Additional bypassing on the screen and plate B+ circuits should be tried. If this fails, the value of *R36* should be lowered sufficiently to keep the stage stable. Once the stage is stabilized, the plate tank circuit should be tuned to resonance. Again the slug of *L2* should be set so that it is impossible to tune the tank circuit to the frequency of the v.f.o. The r-f voltage across *R36* will be between 10 and 15 volts for a loud whistle in the microphone. The bias voltage on this stage should be set so that the idling plate current is between 25 and 30 ma.

The 6146 Stage

Make sure that the high-voltage power supply switch is turned off and then plug the 6146 tube into its socket, *V12*. The first thing is to determine if the stage is stable. If the stage is violently regenerative, there will be no question left in the builder's mind concerning the matter. However, if the stage oscillates only for certain combinations of the grid and plate tuning controls, the following procedure should be followed in looking for instability. First, connect a 50 or 70-ohm dummy load on the output connector. NEVER OPERATE THE STAGE UNLOADED! With plate and screen voltage applied go through the various band-switch positions keeping the grid and plate turrets on the same band, and slowly tune the grid tuning condenser, *C29*, through its full 180-degree movement while rapidly tuning the plate tank condenser back and forth through its range. This procedure should give all possible combinations of the grid and plate tuning controls with a minimum of effort. Watch the plate current meter, *M1*, for a slight change (upward *or* downward) in reading. If a change is noted, investigate more thoroughly. These tests are made without a driving signal from the preceding stages.

The grid bias should now be set so that the idling plate current of the 6146 is between 35 and 40 ma. as read on *M1*.

Assuming the stage is stable, select the proper coils in the grid and plate turrets for 75-meter operation (the nominal "40-meter" coils in the *B&W* turrets) and apply excitation by turning

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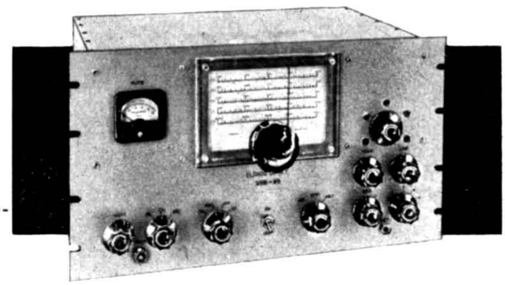

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the *VOX* switch "off" and turning the high-voltage switch "on," meanwhile insert carrier with the *Carrier Insertion* control. The plate current meter of the 6146 stage should go to a high value. Swing the plate tank condenser to resonance as indicated by a plate current dip as in any familiar class C stage. With full carrier inserted, the stage at resonance under full-load conditions should draw about 120 ma. from the power supply. For these conditions of input the power measured into the dummy load should be between 35 and 45 watts.

If operation on other than 75 meters is desired, the heterodyne unit described in *Chapter X* must be used. In this case all the stages through the 2E26 amplifier will be operated as described, and only the 6146 output amplifier will have to be bandswitched to the new amateur band. If operation on 40 meters is desired the band-switches should select the coils in the *B&W* turrets that are the nominal 20-meter coils on the unmodified turret. If operation on 20 meters is desired, the 15-meter coils should be used. When the heterodyne unit is properly patched into the exciter chassis the 6146 stage is tuned as before. When a test tone or carrier is fed through the system, the plate current is dipped as described before.

The Voice Control System

A couple of preliminary checks should be made on the voice control operation. The VTVM should be attached across the secondary of *T3* and the microphone whistled into. The meter should swing up to between 200 and 300 volts rms. Next connect the d-c voltmeter from *pin #4* of *V5b* to ground. Under the "loud whistle condition" the voltage existing at this point should be about minus 200 volts. When the microphone is spoken into in a normal tone for a lower setting of the *VOX* gain control the reading should be about minus 70 volts. All of these measurements must be taken with the *anti-trip* amplifier gain control *R1,* turned to zero. When *VOX* switch *Sw1* is in the "on" position, the relay should drop out as soon as the microphone is spoken into in a normal tone. If not, advance the *VOX* gain control until it does. The relay should remain the "operate" position for a few tenths of a second after speech is stopped. The relay should not "pull in" between normally spoken words of a sentence, but should pull in for normal pauses in conversation. This gives the other station a chance to break in.

With the primary of *T1* connected to the audio output of the station receiver, and the *anti-trip* gain control set to zero, the sounds coming out of the loud speaker should trip the *VOX* system of the exciter. With the receiver audio set for usual confortable audio volume, advance the *anti-trip* gain control, *R1,* until the loud-speaker audio no longer trips the *VOX* system.

Operation

When the great day arrives that the unit is to be "test flown" the operator will be faced with several new problems if he has never previously operated single sideband.

There is the problem of properly "zeroing in" on a station to be worked. Single sideband operation has generally assumed the custom of multi-way round tables, that is, many stations on one frequency all using voice control. This requires that all stations maintain the same frequency and that they be capable of aligning to that frequency before breaking in. One way to do this, if the operator is using the BFO in his receiver to copy the SSB signals. is to carefully tune in the station to be "sat on" and turn on the low-level exciter stages while inserting a very small amount of carrier. Then zero-beat to the same frequency as monitored in the receiver loudspeaker.

This is probably the most simple way of doing the job. If the receiver uses a filter-type SSB receiving adapter, as described in *Chapter I,* the zero-beat point cannot be heard because the carrier oscillator is 20 or 25 db. down the slope of the bandpass filter. Thus it is necessary to "talk yourself on to the correct frequency." This generally requires the use of phones so that feedback will be avoided. The *VOX* switch *Sw1* is turned to the "off" position and the signal monitored while the v.f.o. is changed in frequency. The microphone should be "hello'd" into while listening and when the speech sounds intelligible, the exciter has been aligned to the frequency of the receiver. As the operator becomes accustomed to the ways of SSB operation, he will develop his own pet way of tuning to the frequency.

One thought that must be borne in mind is that the exciter capabilities should not be overtaxed. By this is meant that the linear limits of the amplifiers should not be exceeded. Experience has proved that a conservatively operated SSB system is definitely desirable. If all the trouble of generating a single sideband signal whose undesired sideband is suppressed 40 or 50 db. is worth while, it therefore, appears ridiculous to generate distortion products that are only 20 db. below the intelligible signal. It is economically unwise, as the user might as well have stayed on double sideband and had two sidebands that made sense, as to be on SSB and occupy the same spectrum space and have only sideband that says anything intelligible.

With the particular exciter just described the plate current of the 6146 stage should not kick up over 70 or 80 ma. This is with an average male voice and the plate loading up to the level described under *Alignment.* It is always a good idea to use an oscilloscope to monitor the exciter.

Chapter IX

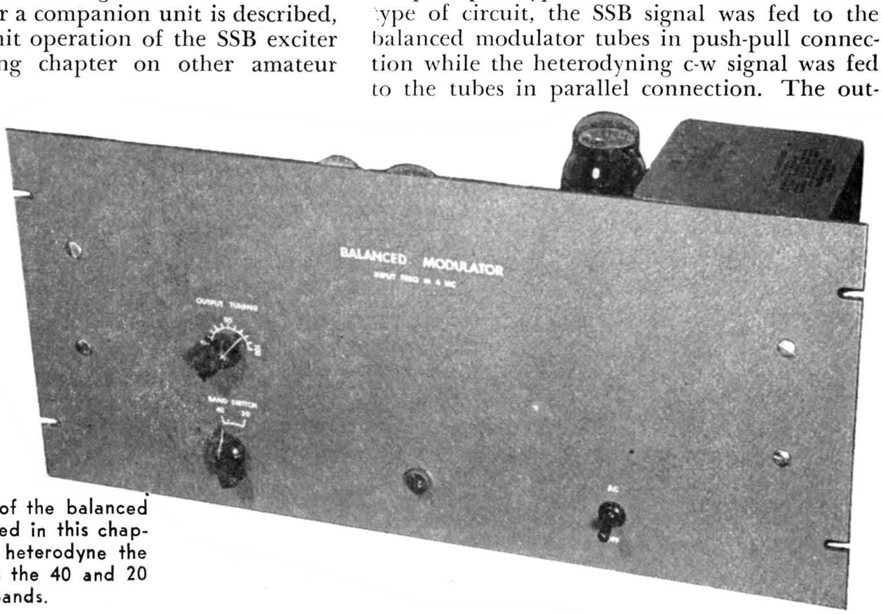
High Level Heterodyne Unit

All-band operation with single sideband equipment is not the simple matter that it is with c.w. or double-sideband transmission. Once the single sideband is generated, band changing by frequency-multiplication cannot be used. The SSB signal must be *heterodyned* to the new desired frequency. There is the alternative method of generating the sideband on the fundamental frequency of operation by the phasing method. This has been dealt with in detail in *Chapters III* and *VII* with the advantages and disadvantages outlined.

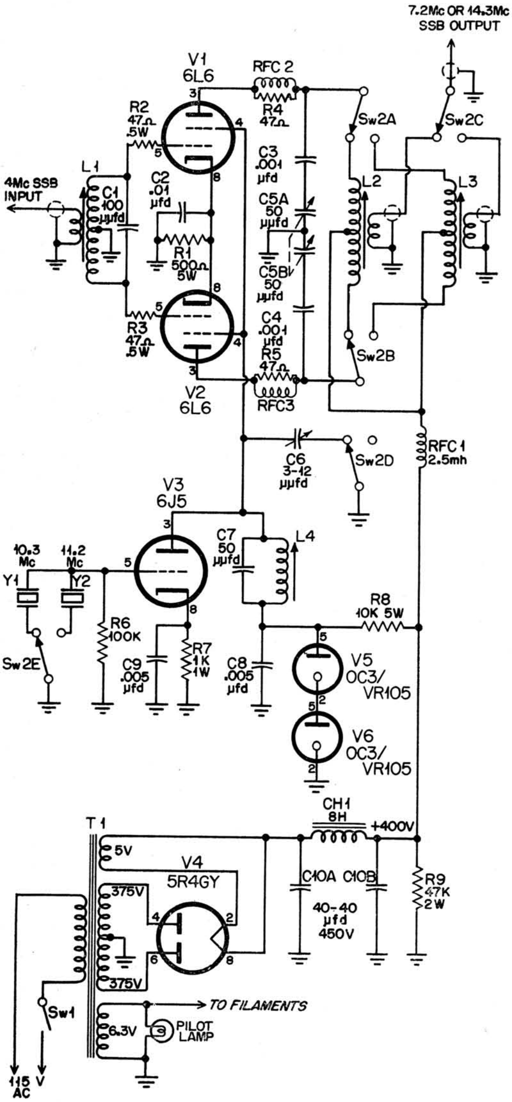
In this chapter a companion unit is described, which will permit operation of the SSB exciter in the preceding chapter on other amateur bands.

9.1—The Balanced Modulator

The general subject of heterodyning with balanced modulators has already been discussed in *Chapter VI* and it is suggested that the reader consult this information for a more complete discussion of the subject. From the discussion in *Chapter VI* it was apparent that there was one type of balanced modulator that possessed characteristics that were more attractive than most of the others. This circuit was that of the push-pull type balanced modulator. In this type of circuit, the SSB signal was fed to the balanced modulator tubes in push-pull connection while the heterodyning c-w signal was fed to the tubes in parallel connection. The out-



Front panel view of the balanced modulator described in this chapter. It is used to heterodyne the 75-meter signal to the 40 and 20 meter bands.



C1—100 μμfd., 500v.
C2—0.01 μfd., 500v., disc ceramic.
C3, C4—0.001 μfd., 600v.
C5a/C5b—50 μμfd., split-stator variable, **Hammarlund HFD-50.**
C6—3-12 /μfd., compression mica, or ceramic concentric.
C7—50 μμfd., 500v.
C8, C9—0.005 μfd., 600v., disc ceramic.
C10a/C10b—Dual 40 μfd., 450v., electrolytic, **Mallory** FP-238.
L1—4-Mc. input coil. 34 turns #26 formex wire, bifilar wound, see text. 2-turn link (all on **National** XR-50).
L2—7.2-Mc. output coil. 32 turns #26 formex wire, bifilar wound, see text. 4-turn link (all on **National** XR-50).
L3—14.3-Mc. output coil. 16 turns #20 formex wire, bifilar wound, see text. 3-turn link (all on **National** XR-50).
L4—10 or 11 Mc. oscillator coil, 15 turns #18 formex wire, single winding, on **National XR-50.**
R1—500 ohms, 5w., wire wound.
R2, R3—47 ohms, ½w.
R4, R5—47 ohms, 1w.
R6—100,000 ohms, ½w.
R7—1000 ohms, 1w.
R8—10,000 ohms, 5w., wire wound.
R9—47,000 ohms, 2w.
RFC1—2.5 mh., r-f choke, **National** R-100.
RFC2, RFC3—6 turns #26 wire wound on R4 and R5.
Ch1—8 henries, 120 ma., choke.
Sw1—SPST toggle.
Sw2,b,c,d,e—5-pole.
Sw2a,b,c,d,e—5-pole, ceramic switch (see text).
T1—Power transformer. 375-0-375v. at 120 ma., 5.0v. at 3 amp., 6.3v. at 4 amp. **Chicago** type PCC-120.
Y1, Y2—Crystals, see text.

Fig. 9-1-A. Wiring schematic and parts list of the heterodyne unit.

put circuit is push-pull connected so that the fundamental and all harmonics of the c-w heterodyning signal are effectively cancelled out in the plate tank circuit. The even harmonics of the SSB signal are attenuated in the output circuit because of the inability of a push-pull circuit to act as an efficient frequency multiplier for the even harmonics.

The schematic as shown in *Fig. 9-1-A* is a modification of the pentode circuit described in *Chapter VI* and illustrated by *Fig. 6-4-A* of that chapter. The change that was adopted for use in this unit was the interchanging of the feed points of the two signals to be mixed. In this case the heterodyning signal is fed into the screen grids of the balanced modulator tubes in parallel while the SSB signal is fed into the control grids in push-pull connection. The reason that this method was adopted was that screen-feeding could be realized without any coupling devices whatsoever—not even a coupling condenser.

The problem involved in this heterodyne unit is that of frequency-converting a 3.8 to 4.0 megacycle SSB signal into other amateur bands. The two bands chosen as examples by the author were the 7.2 to 7.3-Mc. and 14.2 to 14.3-Mc. phone bands. Conversion to other amateur bands with the unit is also possible by making two appropriate changes. Since the basic 4.0 megacycle exciter has a v.f.o. as a part of its equipment, the heterodyne unit therefore need include only fixed-frequency heterodyning. Choosing the right crystals for the job is quite simple. It was decided to use the "difference-mixture" for the 40-meter band, and the "sum-mixture" for working the 20-meter band. Thus a crystal frequency of 11.2 Mc. was chosen for the 40-meter heterodyning oscillator. Subtracting 4.0 Mc. from 11.2 Mc. gives 7.2 Mc.—the lower band edge for the 40-meter band. Moving the v.f.o. so that the SSB signal starts at 3.9 Mc., the 40-meter signal will be at 7.3 Mc.—the high-frequency band edge. Using the difference-mixture as in this case where the heterodyning frequency is higher in frequency than the SSB frequency will invert the sideband relation of the output signal to that of the original 4.0 megacycle-band signal. By this is meant that when operating on 40 meters, the "upper" and "lower" sideband designations of the sideband switch will be reversed. This should be kept in mind. This is no handicap as long as the operator is aware of the situation.

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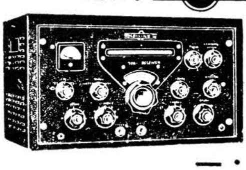

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For 20-meter band operation, since the "sum-mixture" was chosen, a crystal frequency of 10.3 Mc. is necessary to cover the 14.2 to 14.3-Mc. phone band. Thus when the input signal is at 4.0 Mc. the 20-meter signal is at 14.3 Mc., and when the input signal is at 3.9 Mc. the output signal is at 14.2 megacycles. In this case, the sideband relation between upper and lower sidebands stays the same as that of the original 4.0-Mc. signal.

9.2—Circuitry

Refer to *Fig. 9-1-A.* The first version of this unit that was built used two pentagrid converter tubes in a balanced modulator circuit. This arrangement did not furnish the necessary output to drive the 6146 output stage of the exciter described in *Chapter VIII.* The present model was then built using a pair of 6L6 pentodes. A gain of almost unity was realized on both 40 and 20 meters. Since the 2E26 driver stage had more drive than was actually needed by the final output stage, a gain of slightly less than one could be tolerated.

The Bifilar Coils

The only feature that might be unfamiliar to the constructor is the use of bifilar-wound input and output tank coils. These coils are not difficult to make and the sketch of *Fig. 9-2-A* shows how the winding is connected. The use of bifilar coils was decided upon to attempt to establish a balance on the input and output circuits. The author, however, believes that the unit would perform satisfactorily using ordinary-wound coils for the inductances. Since the coils were made up and the unit worked without any "bugs," it was felt that the bifilar coils should be retained.

National XR-50 slug-tuned coil forms were used for all coils in the unit. Extra terminals were put on the phenolic coil forms by drilling and tapping the form to take short 3/16-inch number *2-56* brass screws. Care must be taken not to screw the brass terminals too far into the tapped hole so that they "short out" to the powdered iron slug. This accidentally happened to the unit built and a B plus to ground short took place with the usual results—an overheated r-f choke. The brass screws may be used for the tie-points for the extra winding of the bifilar coils.

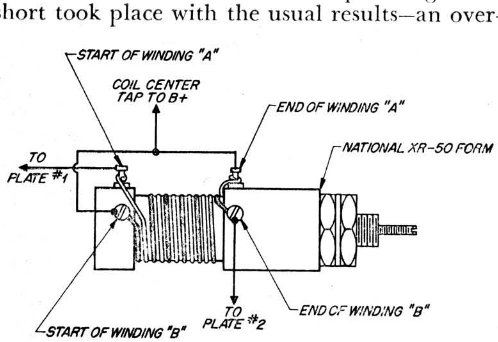


Fig. 9-2-A. Method of winding a bifilar coil on a National XR-50 coil form. Two small #2-56 brass screws were added to the form as terminals for winding "B." See text.

Winding a bifilar coil is most easily accomplished by taking two parallel strands of wire and carefully winding them simultaneously on the coil form making sure that there are no twists and that the winding is smooth and symmetrical. When the winding is completed, and the two separate windings are soldered to their respective tie points, an ohmmeter should be used to find out which ends of the wires belong together. *Figure 9-2-A* shows that the finishing end of winding *A* should be connected to the starting end of winding *B* in order to establish the electrical center-tap of the coil. The remaining two ends are then the top and bottom terminals of the coil. The notation of the number of turns in *Fig. 9-1-A* lists the *total* number of turns and the number of turns put on the coil when winding the two parallel wires on should be *half* of this indicated number. The crystal-oscillator coil, *L4* is not a bifilar-wound coil and is a conventional single-layer solenoid.

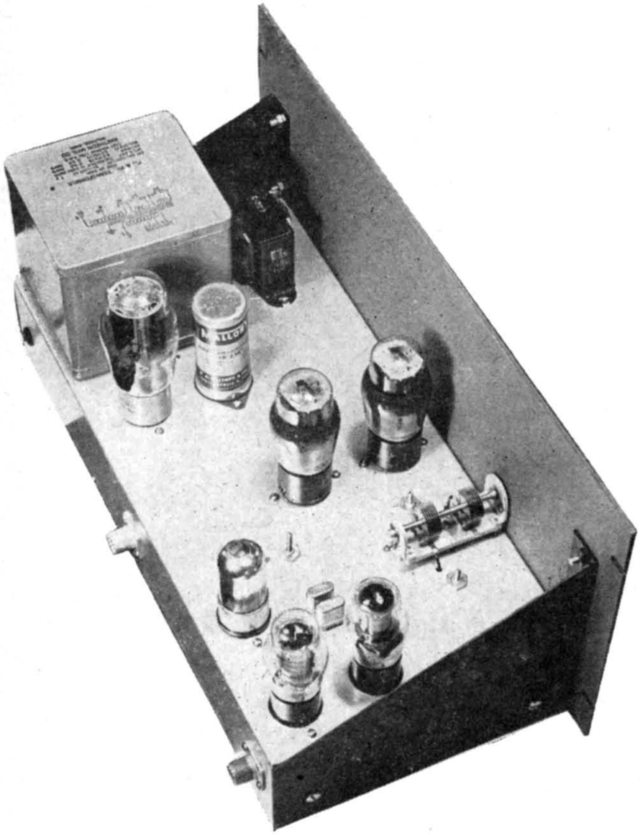
The Crystal Oscillator

The crystal oscillator tube is a 6J5 or any other similar triode. The availability of good crystals in the frequency range required make this stage simple to build and to adjust. The crystals used were the *type H-173* of *James Knights Co.* The two crystals used were excellent oscillating units and gave absolutely no trouble in starting. A conventional tuned plate triode circuit was used. The plate voltage was regulated with two VR-105 tubes because it was found that the screen current of the 6L6 tubes varied considerably under signal input and this caused a varying plate voltage due to the drop across *R8.* This change in plate voltage caused a slight amount of frequency-modulation of the crystal frequency. This would result in f.m. of the output signal which would be strictly taboo for a good SSB signal. The condenser *C6* is used to correct the oscillator tuning when the low-frequency crystal is switched into the circuit by *Sw2e.* For adjustment of this circuit see the section under *Alignment.*

The Mixer Circuit

As mentioned previously the oscillator signal is fed into the parallel connected screen directly from the crystal oscillator tank circuit without benefit of coupling condensers. The plate voltage dropping resistor for the oscillator stage also serves as the screen dropping resistor for the 6L6 tubes. The r-f voltage present at the screen terminals of the mixer tubes *(pin #4)* should be between 40 and 60 volts rms. This should be measured with an r-f VTVM.

A balancing potentiometer was originally placed in the cathode circuit, however, this was found to be unnecessary and was eliminated for the sake of simplicity. A common cathode resistor, *R1,* is used instead.



Above chassis view of the balanced modulator.

The usual precautions against parasitic oscillation were taken by installing parasitic resistors in the grid circuits and *R/L* parasitic chokes in the plate circuits. No trouble with either fundamental or parasitic frequency instability was encountered.

Condensers *C3* and *C4* are used as d-c blocking condensers to keep the plate voltage off the stator sections of the tank tuning condenser, *C5*. The tank condenser used was a *Hammarlund HFD-50.*

The bandswitch used in the unit was one made up of two wafers from available switches. It was thought that it would be desirable to have some physical separation between the switch wafer used in the oscillator circuits and the wafer used to switch the output tank circuits. The switch wafers were thus moved apart by extending the switch bushings and mounting screws. The shaft also had to be extended. This is most easily done by using a standard ¼-inch shaft coupling (non-insulated type). A separation of about 5 inches was used in the unit shown. The switch sections labelled *Sw2a, Sw2b,* and *Sw2c* are on the front wafer section next to the front panel while *Sw2d* and *Sw2e* are on the rear wafer near the crystal oscillator.

The Power Supply

This circuit is conventional and little need be said. The transformer, *T1,* used in the author's unit was one that happened to be available in the shack. There are a number of commercial transformers that would be suitable for the job—such as the *Chicago type PCC-120* unit. A dual 40-μfd. filter condenser provides suitable filtering when used in conjunction with an 8-henry choke.

Alignment

Alignment of the unit is simple and should proceed as follows: The first step is to adjust the crystal oscillator circuit for oscillation. Connect the r-f probe of a VTVM from the screen terminal of *V1* or *V2 (pin #4)* to ground. Move the bandswitch to the 40-meter position —this switches the 11.2 Mc. crystal into the circuit. Adjust the slug of *L4* for oscillation as indicated by a reading on the VTVM. Adjust the slug so that a high reading of voltage is read, however, set the control conservatively so that when the bandswitch is thrown back and forth the crystal starts oscillating easily. Check the oscillation frequency by checking on the shack receiver, or a BC-221 frequency meter. Next, throw the bandswitch to the 20-meter position. Indication of oscillation will probably cease. Leave the slug of *L4* in the present position and now adjust condenser *C6* until oscillation is again present on the new crystal frequency. Now switch back and forth several times between the two band position and make sure that the respective crystals oscillate reliably each time. The r-f voltage from the screen terminals to ground should be between 40 and 60 volts rms.

Connect the SSB exciter to the input terminals and adjust the SSB excitation so that it is near 4.0 megacycles. Turn the bandswitch to the 40-meter position and using the station receiver monitor the mixture-frequency near 7.2 megacycles. Insert a small amount of carrier and peak the output tuning condenser *C5* for a maximum signal in the receiver as indicated by a maximum reading of the *S*-meter. If the condenser will not quite tune to the correct frequency adjust the slug of *L2* until the tuning range of the condenser will hit the output frequency. The input tuned circuit, *L1,* should

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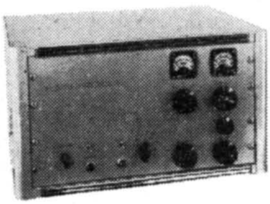

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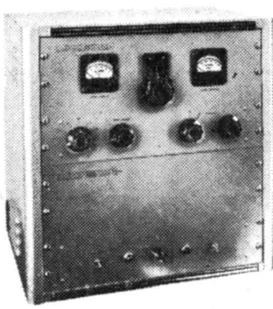

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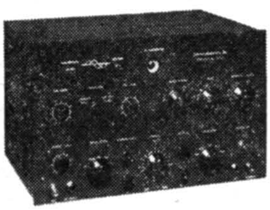

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Switch *Sw2* to the 20-meter position and go through the same adjustments of the 14-Mc. output circuit as described above. Adjustment of the input circuit need not be made again since it holds for either band.

Some reflected tuning changes were noted between the exciter final grid circuit and the output tuned circuit of the heterodyne unit. Alternate peaking of each condenser will usually yield an optimum setting for both controls. This is apparently an impedance matching process that takes place because of the manner of driving the final grid coil from the low-impedance line.

It is generally a good policy to always use the station receiver when tuning up the heterodyne unit for operation. This is to prevent tuning up on a spurious mixture or harmonic signal. This is a safety precaution that may not be necessary once the operator is familiar with the tuning position of the controls. The author noticed no spurious signals in the unit output that could *normally* cause trouble. All harmonics of the crystal oscillator frequencies are far from the amateur bands concerned.

Mechanical Lay-Out

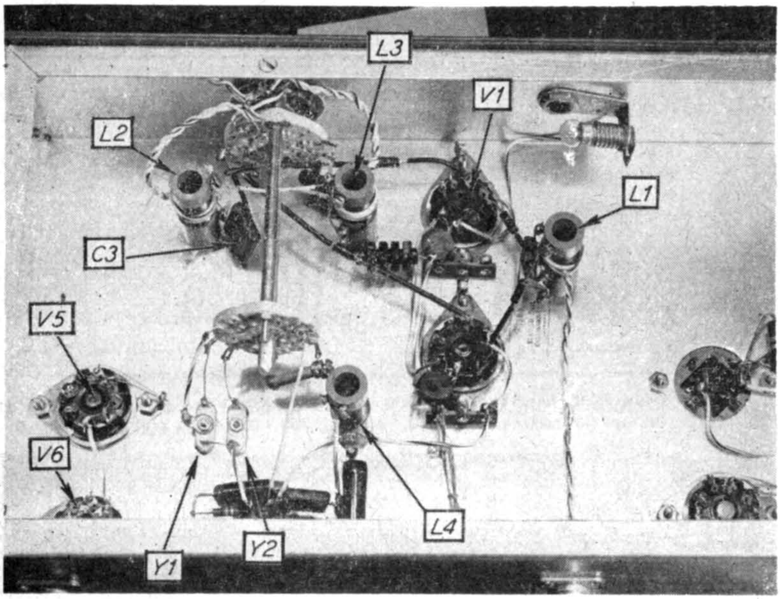
No lay-out drawing is presented for this unit since the physical arrangement of the circuits is not critical. The photographs show the arrangement used by the author, and if the constructor desires, this general plan may be followed. Other mechanical mounting schemes will suggest themselves. The dish-type of chassis mounting offers an interesting means of constructing a simple unit such as this. Since the stage input and output circuits are tuned to widely divergent frequencies, there is little danger of oscillation taking place even in the most awkward of coil placements.

Operation

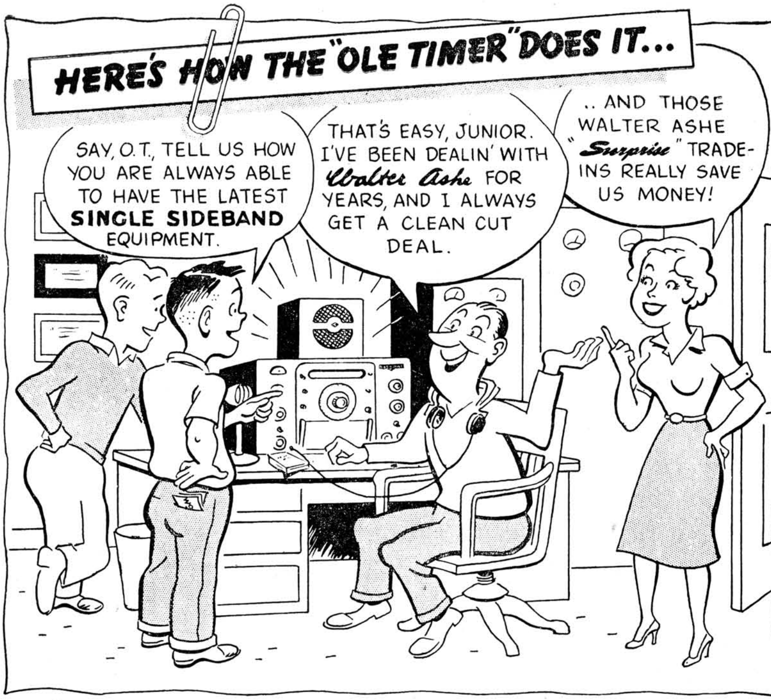
Operation of the unit once tuned up is exceedingly simple—just leave it alone and make all operational adjustments on the exciter proper. The output tuning might have to be peaked up when moving from one extreme band edge to the other, but normally the circuit bandwidth is sufficient to accommodate normal frequency changes. The unit built did not create any TVI in the author's TV set, but in some instances this might not be the case. The offender would most likely be harmonics of the crystal oscillator stage that would be radiated directly from the unit. The usual precautions of shielding and filtering the input and output lines should then be taken.

Using the Unit on Other Amateur Bands

While operation on 40 and 20 meters has been covered in this discussion, it is apparent that there is no restriction in adding other bandswitch positions and appropriate crystals and coils to cover 15 and 10 meters. Choosing the correct crystal is merely a matter of subtracting 4.0 megacycles from the phone band upper frequency limit. The output coil should tune to the desired output frequency with the tuning capacity available.



Under the chassis of the balanced modulator. The parts layout is not critical.



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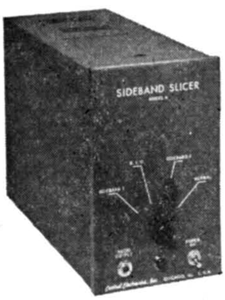

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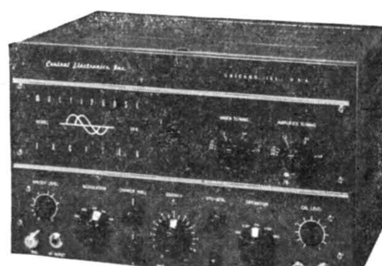

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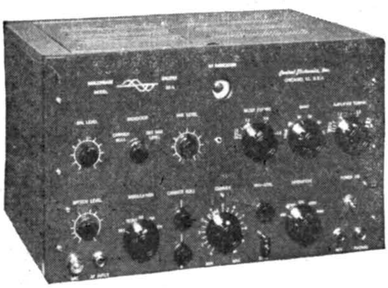

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Chapter XI

Filters and Filter Alignment

Little has been said in the preceding chapters of the different types of filters commonly used by SSB operators. Although it would be impossible in the space available to cover the subject of filters completely, it is felt that a brief discussion of the different types of crystal filters that can be constructed by the oft-mentioned "average amateur" would be in order. The radio amateur literature has had frequent articles on crystal filters that use the surplus FT-241 type low-frequency crystals. These CT-cut crystals have been plentiful and relatively cheap for a number of years and are in the hands of many amateurs. The general run-of-the-mill war surplus crystal may or may not be good. The author's experience with about two-thousand of the FT-241 crystals has been that approximately one-quarter of the lot are either inoperative or very inactive and sluggish. Obviously, these do not make good filter crystals.

Checking Surplus FT-241 Crystals

A signal-generator or BC-221 frequency-meter can be used in conjunction with an oscilloscope or r-f vacuum-tube voltmeter to determine the relative activity of a crystal and also its series resonant frequency. The crystal should be connected in series with the "hot" lead from the signal generator to the oscilloscope. Slowly tuning through the frequency range of the crystal with the signal generator will produce a sharp increase in indication on the oscilloscope or VTVM. Carefully tune to this peak and determine its frequency. This is the series resonant frequency of the crystal. About 200 cycles higher in frequency experimenters may be fortunate enough to detect the parallel-resonant frequency as the frequency at which the oscilloscope gives a sharp null indication. The height of the peak and the sharpness of the null is a rough indication of the crystal's activity and quality. When surplus crystals are being selected for use in a filter, a number of them should be checked and the most active ones selected for use.

Matching crystals for filters sometimes requires that pairs of crystals must be selected so that their frequencies are identical, or nearly so. This is most easily done by using them as oscillating elements in a Pierce oscillator circuit. The oscillating frequency can be checked by a BC-221 or similar device.

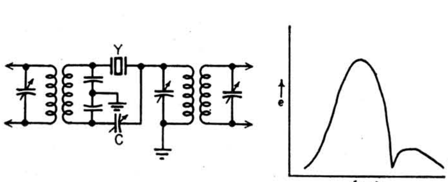


Fig. 11-1-A. The common crystal arrangement as used in most modern communications receivers.

11.1—Filter Circuitry

Let us now consider some of the basic types of filter configurations that are in common use. The most common filter circuit in use is shown in *(Fig. 11-1-A)*. This is the single-crystal arrangement that is standard equipment in most modern communications receivers. The response curve for a typical setting of the phasing condenser *C* is shown at the right of the filter schematic. Varying *C* will change the shape of the main passband lobe considerably, but the frequency of the main response peak will remain fixed. However, the sharp null or "notch" will be moved as the value of *C* is changed. When *C* is equal exactly to the shunt crystal-holder capacity the null will *not* be present in the response curve shown. At this setting of *C*, the passband will have its sharpest passband characteristic. As *C* is varied either above or below this balancing value, the notch will appear either on one side of the center frequency or the other. As *C* is varied still farther from its balanced setting the notch will get closer to the center frequency of the filter

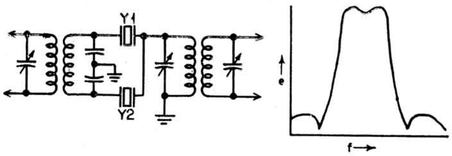


Fig. 11-3-A. The popular half-lattice crystal circuit.

response. The notch cannot come any closer to the center frequency than about 200 or 500 cycles. This accounts for the often noticed inability to null out heterodynes that are close to the desired signal frequency. The notch of the filter while quite deep is sometimes broad across the top of the slot. The variation in passband shape is also annoying at times.

11.2—The Q-Multiplier

For the reasons outlined above many Hams are substituting one of the various forms of the "Q-Multiplier" circuits[8] devised by O. G. Villard, W6UYQ.[38] This filter, while not a crystal filter, is superior in several respects to the conventional single-crystal filter. The operator has the choice of using the device as either a "boost" or a "reject" filter. In the "boost" position a very narrow band of frequencies is selected and all others rejected; while in the "reject" position a very narrow band of frequencies is rejected leaving the remainder of the normal i-f receiver response curve unchanged. It is suggested that the original literature be consulted for theory of operation and constructional details.

11.3—The Half-Lattice

Figure 11-3-A is the increasingly popular "half-lattice" circuit that many of the SSB operators are using. It is a derivation of the full-lattice shown in *Fig. 11-3-B*. Both filters are bandpass filters and are ideal for generating SSB signals or as receiver i-f selectivity filters. The half-lattice filter theoretically has the same response curve as the full-lattice. The frequencies selected for the crystals *Y1* and *Y2* must be 2 to 3 kilocycles apart. The exact difference is not important. Good active crystals should be used as already mentioned.

Once aligned the bandwidth of the filter will be approximately 1.2 times the frequency

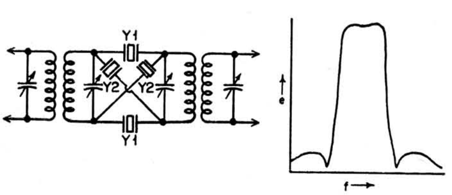
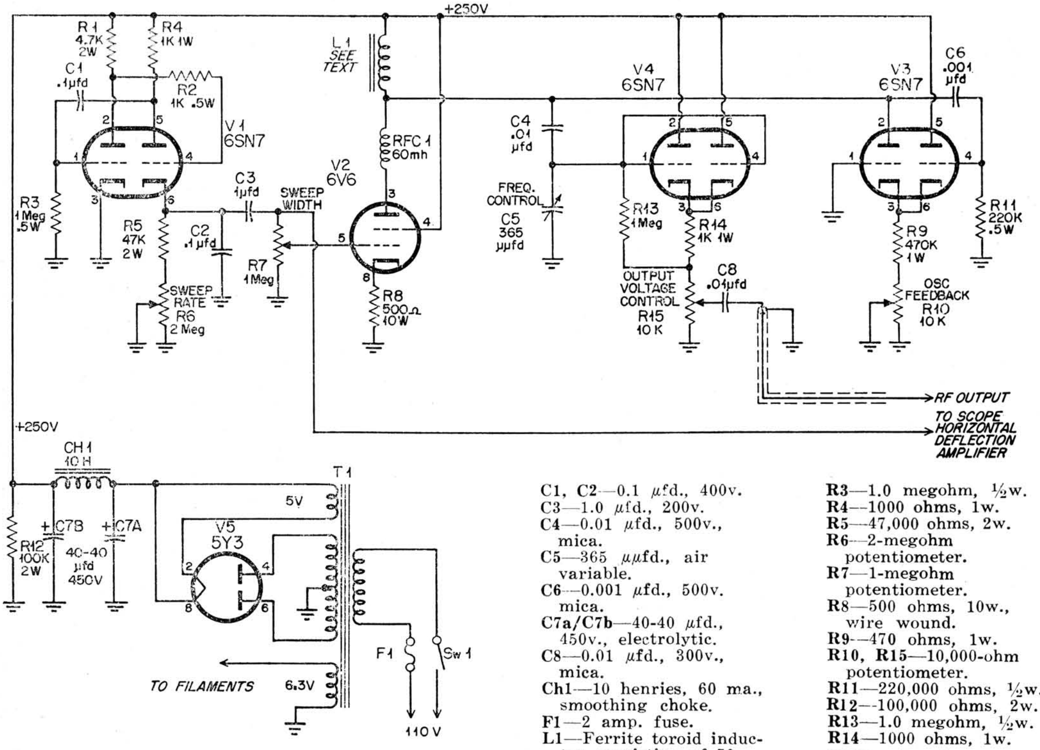


Fig. 11-3-B. A full-lattice crystal arrangement.



C1, C2—0.1 μfd., 400v.
C3—1.0 μfd., 200v.
C4—0.01 μfd., 500v., mica.
C5—365 μμfd., air variable.
C6—0.001 μfd., 500v. mica.
C7a/C7b—40-40 μfd., 450v., electrolytic.
C8—0.01 μfd., 300v., mica.
Ch1—10 henries, 60 ma., smoothing choke.
F1—2 amp. fuse.
L1—Ferrite toroid inductor consisting of 50 turns #18 DSC on a Crowley toroid, type CR70, Part #5391.
R1—4700 ohms, 2w.
R2—1000 ohms, ½w.
R3—1.0 megohm, ½w.
R4—1000 ohms, 1w.
R5—47,000 ohms, 2w.
R6—2-megohm potentiometer.
R7—1-megohm potentiometer.
R8—500 ohms, 10w., wire wound.
R9—470 ohms, 1w.
R10, R15—10,000-ohm potentiometer.
R11—220,000 ohms, ½w.
R12—100,000 ohms, 2w.
R13—1.0 megohm, ½w.
R14—1000 ohms, 1w.
RFC1—60 mh., iron choke, J. W. Miller.
Sw1—SPST toggle.
T1—Power transformer, Thordarson T13R13.

Fig. 11-5-A. Parts list and wiring details of the "Ferri-Sweeper."

separation of *Y1* and *Y2*. The sharp notches either side of the passband are quite deep (about 60 db. for a single filter section). The filter should be aligned so that the drop-off outside the passband is sharp while at the same time the extra response lobes beyond the deep notches are kept as far down as possible. The actual alignment usually ends up as a compromise because steepness of skirt response and low "outer lobe" response do not generally occur simultaneously.

As it turns out the correct impedance for the two crystals to look into at the primary winding of the output i-f transformer is just half of that presented by the *secondary* of the *input* transformer. Thus it is often advantageous to place a resistor across the terminals of the primary of the output transformer in each section of the half-lattice filter. The actual value must be determined by experiment and is in the order of a few hundred thousand ohms. This will tend to give a better over-all response curve shape. Two sections of a half-lattice filter are usually necessary to obtain enough selectivity for SSB generation or reception.

11.4—The Full-Lattice

Figure 11-3-B as already mentioned is the "full-lattice" filter. The same comments apply to this filter as applied to the half-lattice filter with the exception that the crystals labeled *Y1* must be as close to the same frequency as possible; the same applies to the *Y2* crystals. These crystals should be matched to within at least 50 cps—closer if possible. The frequency difference between *Y1* and *Y2* groups should be 2 to 4 kc. as in the half-lattices. Two tandem-connected filter sections will give approximately 60 db. attenuation of all out-of-passband signals.[7]

Filter Combinations

Figure 11-4-A shows one of the many possible combinations of crystal filters that may be used to good advantage. The basic filter is the half-lattice of *Fig. 11-3-A* plus two additional crys-

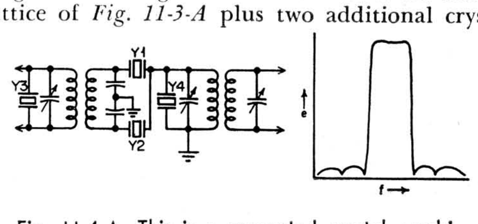


Fig. 11-4-A. This is a suggested crystal combination with a half-lattice and two shunt crystals across the windings of the i-f transformer.

Front panel view of the "Ferri-Sweeper" developed to enable alignment of SSB filters. The sweep rate is variable between 2 and 20 sweeps per second.

tals connected across the i-f transformer windings. The extra crystals are selected so that their series resonant frequencies fall in the middle of the outer lobes that are undesirable in the basic filter characteristic. The extra crystals need not be restricted to two in number, but as many as six or eight crystals can be used if necessary.

Another combination that suggests itself is one section of full-lattice followed by a section of half-lattice. The serious experimenter can try any combination that looks logical and satisfy his curiosity as to just what happens.

11.5—The "Ferri-Sweeper"

There comes a times in every filter SSB-man's life when he wishes he had a sweep frequency generator to simplify the alignment of a sideband generating filter. If he is lucky, he manages to borrow one from a friend, but is usually dismayed to find that the sweep recurrence rate is usually tied directly into the 60-cycle line frequency. In order to align filters with rapid skirt attenuation such as the filters just described, the sweep rate must be slowed down to between five and ten sweeps per second, and the total sweep width should not be more than fifteen or twenty kilocycles. Generally, modification of an existing f-m signal generator is not too successful—it usually ends up as a mediocre filter alignment generator and unusable for its original purpose.

A sweep-oscillator unit nicknamed the "Ferri-Sweeper" has been used by the author for some time and has been successfully employed to align various kinds of filters—both crystal and L/C types. The sweep rate is adjustable from two or three up to about twenty sweeps per second. The sweep width is continuously adjustable from zero up to about 100 kilocycles in the i-f range. The output amplitude is adjustable and is isolated from the oscillator by a cathode follower stage. The circuit was originally suggested by Peter Sulzer, W3HFW, as a result of his experiments with one of the ferrite (hence the name "Ferri-Sweeper") toroidal cores used in the oscillator circuit.

Circuitry and Theory of Operation

Refer to *Fig. 11-5-A* for the circuit details. The two halves of *V1* serve as the sweep generator acting as a free-running multivibrator. The waveform of the voltage across the cathode resistors, *R5* and *R6*, of *V1B* is a sawtooth. The recurrence rate is adjustable with the 2-megohm potentiometer, *R6*. The sawtooth voltage is coupled to the sweep-amplifier stage, *V2*, through the 1.0 μfd. coupling capacitor, *C3*. The large value of coupling capacity is necessary because of the low sweep rates used. This value should not be decreased if satisfactory operation is desired at low sweep recurrent rates.

The sweep voltage (sawtooth) is also taken to a terminal on the rear of the chassis so that it can be used as the horizontal sweep voltage for the oscilloscope used in the filter alignment process. This permits the 'scope to be synchronized with the sweep generator.

The plate current pulse of the sweep amplifier tube is also sawtooth in shape and it is this current that causes the oscillator to sweep in frequency. Note that the plate current from the sweep amplifier passes through the 60-millihenry r-f choke, *RFC1* and then through the actual oscillator winding on the ferrite toroid coil, *L2*, and thence to B plus. The r-f choke, *RFC1* serves to keep the plate of the sweep amplifier from loading down the oscillator tank circuit and stopping oscillation. At the same time the sawtooth-shaped current can pass through the choke to the toroidal coil.

Ferrites have the property of having their permeability changed when they are under the influence of a magnetizing field. If a linearly varying direct current is used as a magnetizing current, the frequency of the oscillating tuned circuit will vary linearly over a maximum range equal to almost half the value of the center frequency of oscillation.

The oscillator circuit is a two-terminal oscillator using cathode coupled triodes as oscillator tubes. This oscillator was described by Sulzer[4] in 1950. The cathode potentiometer, *R10,* is used to adjust the amount of feedback for optimum conditions. It should be set so that reliable oscillation takes place over the working range of the oscillator. The oscillator tank circuit is composed of the ferrite-toroid coil, *L1,* and the single-section broadcast condenser, *C5.* The 0.01 μfd. condenser, *C4,* is used to keep the B plus voltage from appearing on the plates of the tuning condenser.

The output cathode follower is conventional and provides variable output voltage with complete isolation from loading effects upon the oscillator. The output impedance is relatively low being in the region of a few hundred ohms. The maximum output voltage is in the neighborhood of three volts r.m.s.

Construction

This unit was built strictly as a functional piece of test equipment and the physical shape of the lay-out is not critical. Anyone of several methods of construction may come to mind. The conventional mechanical lay-out used by the author is shown in *Fig. 11-5-B* merely to serve as a guide.

The power supply was included as an integral part of the unit. Any small power transformer that can furnish 60 or 70 ma. of plate current at 200 to 300 volts d.c. will be satisfactory. Good filtering must be provided so that any ripple present in the B plus supply will not have a tendency to synchronize the multivibrator sweep generator at sub-harmonics of the 120-cycle ripple frequency. The dual 40-μfd. electrolytic condenser, *C7A* and *B,* and single choke, *Ch1,* were found to be satisfactory.

The only non-standard component in the whole piece of equipment is the toroid core on which the oscillator tank coil, *L1,* is wound. A winding of 50 turns of #18 double cotton covered wire was found to cover the frequency range from 300 kc. to 800 kc. with the tuning condenser indicated. The toroidal core itself costs only about ten cents each in small quantities and can be obtained from *Henry L. Crowley & Co.** The type number and part num-

* West Orange, New Jersey

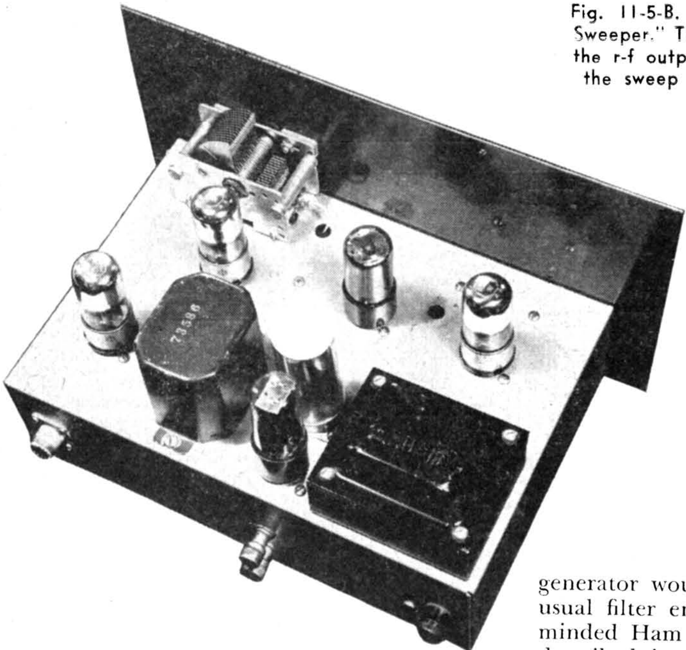


Fig. 11-5-B. Rear chassis view of the "Ferri-Sweeper." The coaxial socket on the skirt is for the r-f output. The two binding posts are for the sweep signal output to the oscilloscope.

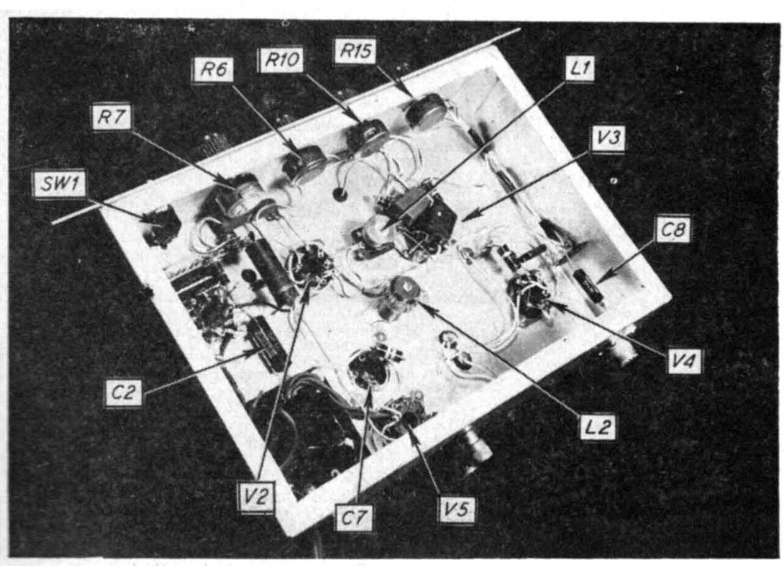
ber are given in the list accompanying *Fig. 11-5-A.* Other manufacturers also make ferrite toroids that would certainly perform satisfactorily in this oscillator.

As the photos of the constructed unit indicate, there are five front panel controls (plus the ON-OFF switch). These are: sweep rate *(R6)*, sweep width *(R7)*, oscillator feedback *(R10)*, output voltage *(R15)*, and frequency control *(C5)*.

Using the Sweep Generator

A few pointers on using the sweep frequency generator would be proper at this point. The usual filter encountered by a single sideband-minded Ham will be one of the general types described in the first of this chapter (see *Fig. 11-1-A, etc.*). Filters such as those shown are usually high impedance in character in the passband region. This means that the mixers or amplifiers that are used with the filters must not load down the filters by appearing as low impedances across the input or output terminals. This is usually accomplished by using pentode class A stages before and after the filter.

When we use the sweep generator to align the filter we must also remember not to load the filter with the generator output impedance. We can avoid this by feeding the sweep generator signal into the grid of the stage preced-



Under chassis view showing the placement of some of the most obvious and outstanding parts.

ing the filter, or by using a very small coupling condenser between the generator output and the "hot" lead of the filter input. This condenser should usually not be over 4 $\mu\mu$fd. in value. This constitutes quite a drop in signal voltage between the generator and the filter but is the price that must be paid for impedance matching. It must be remembered that the shunt capacity on the input and output of the filter, in the test set-up, will usually not be exactly like that encountered in the actual operating circuit. Thus, when installed in the final transmitter-exciter or receiver i-f strip, the input and output trimmer condensers will probably have to be retuned slightly to compensate for the change. If it is possible to align the filter in the actual circuit in which it is to be used—do so by all means. In this way, the problem will be avoided.

In filter alignment a broadly tuned amplifier following the filter is necessary to make up the insertion loss of the filter and to bring the signal up to a high enough level to be presented on an oscilloscope or measured by a

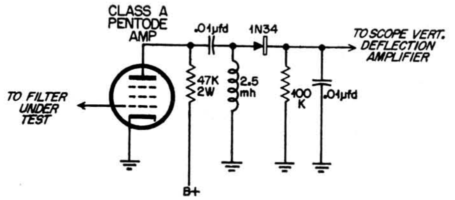


Fig. 11-5-C. Envelope detector for use with the "Ferri-Sweeper."

VTVM. Most oscilloscopes will respond satisfactorily to signals in the i-f region. However, a simple diode detector may be used to rectify the r-f signal so that only the envelope of the resulting pulse is seen on the screen of the oscilloscope. Such a simple diode detector is shown in *Fig. 11-5-C*.

In aligning filters with sweep frequency generators one is hampered by seeing only part of the amplitude response instead of the full 40 or 60 db. that may be of interest. The "top 20 db." of the response curve is about all that can be observed because the amplitude response of the oscilloscope, and the amplifiers before and after the filter, is linear.

Index

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About the Author

Mr. Jack N. Brown was born in Franklin, Pennsylvania in 1922. His interest in electrical engineering and radio started when he was a teen-ager—resulting in a Ham license in 1938 under the call of W8SHY. Mr. Brown's formal study at Penn State College was interrupted by a 42-month stint with the U.S. Signal Corps ending in 1946. During the latter portion of his Armed Service duty a familiar call on both 10 and 20 meters was W8SHY/J9.

Mr. Brown returned to Penn State and received his engineering degree in 1948. He then joined the staff of the *Central Radio Propagation Laboratory, National Bureau of Standards,* Washington, D.C. While at the *CRPL,* Mr. Brown contributed mostly to the study of the propagation of very low-frequency radio waves. He is the author, and co-author, of a large number of scientific papers in this field.

In 1953. Mr. Brown transferred to *Barker & Williamson, Inc.* as a consultant on equipment development. He is currently employed at their Bristol, Pa. plant.

Mr. Brown became interested in 1948 in SSB equipment and operation. During the spring, summer and early fall of 1953, *CQ* Magazine published his very well-known series, "Getting Started in Single Sideband." This book is the result the response afforded that series of articles. Since moving back to Pennsylvania, Mr. Brown has relinquished his W4OLL call and is now known as W3SHY.

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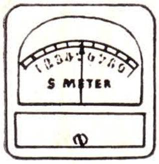

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