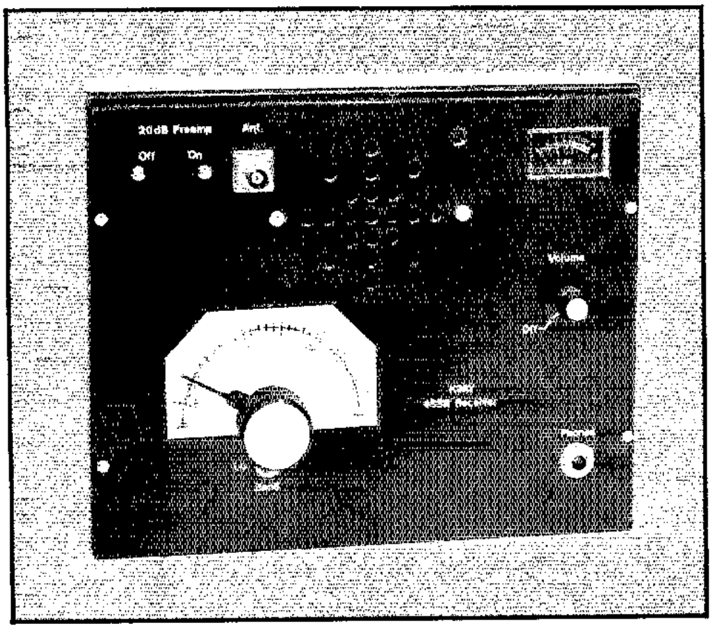

A New Breed of Receiver

Build a direct-conversion SSB receiver? Why not— it's easier to do than you might think.

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The phasing method for generating and receiving SSB has been around for a long time. The first amateur 14-MHz SSB transmissions were made in October 1947. In July 1948 *QST*, Donald E. Norgaard, W2KUJ, described the phasing method for SSB reception. In September 1969 *QST*, Richard S. Taylor, W1DAX, described a direct-conversion (D-C) SSB receiver. That article does not cover all the circuitry that comprises the receiver, however.

Modern amateur SSB equipment uses the filter method. Thanks to high-quality, low-cost filters, the phasing method seems to have disappeared.

My recent experiments, as demonstrated in this receiver, suggest that the phasing method is no dinosaur! Recent developments in design techniques and components have made it possible to build a D-C SSB receiver with surprising performance. The block diagram of this receiver is shown in Fig 1. (See the sidebar for an explanation of how the phasing method of SSB reception works.)

Phasing Circuits

My interest in the phasing method began with the accidental discovery of 90° phase-shift network design tables. In a book on filter design, Arthur Williams has taken the classic work of Bedrosian and updated it for modern op amp all-pass network design.[1,2] The audio phase shift network used in this receiver is shown in Fig 2. Each op-amp stage provides a phase shift centered on a particular frequency. Cascading op-amp stages provides a stable phase shift characteristic over a band of frequencies, much like using stagger-tuned filters to get a flat passband over a desired frequency span.

Each network uses TL084 FET-input op amps, and all resistors and capacitors have 1% tolerance. Standard resistor values were placed in series to obtain the design resistance. These 1%-tolerance parts are inexpensive and easy to obtain, quite a change from those early days of SSB![3]

The road to RF phasing was more roundabout. I tried several experimental configurations (see Fig 3). My first attempt was an effort to design a "proper" network, using a 0° hybrid power divider to isolate the output ports and the ± 45° L-network phase-shift elements. The circuit worked well, but was not particularly wideband, covering about 200 kHz at 7 MHz with better than 1° accuracy. This circuit is shown in Fig 3A.

In another breadboard, I used LR and CR networks without paying attention to port-to-port isolation (Fig 3B). This simple plus and minus 45° network worked every bit as well as the more complex design tried earlier. My final choice for a circuit is shown in Fig 3C. It is an RC network with variable capacitors for adjustment.

This circuit performs well when termi-

[1]Notes appear on page 23.

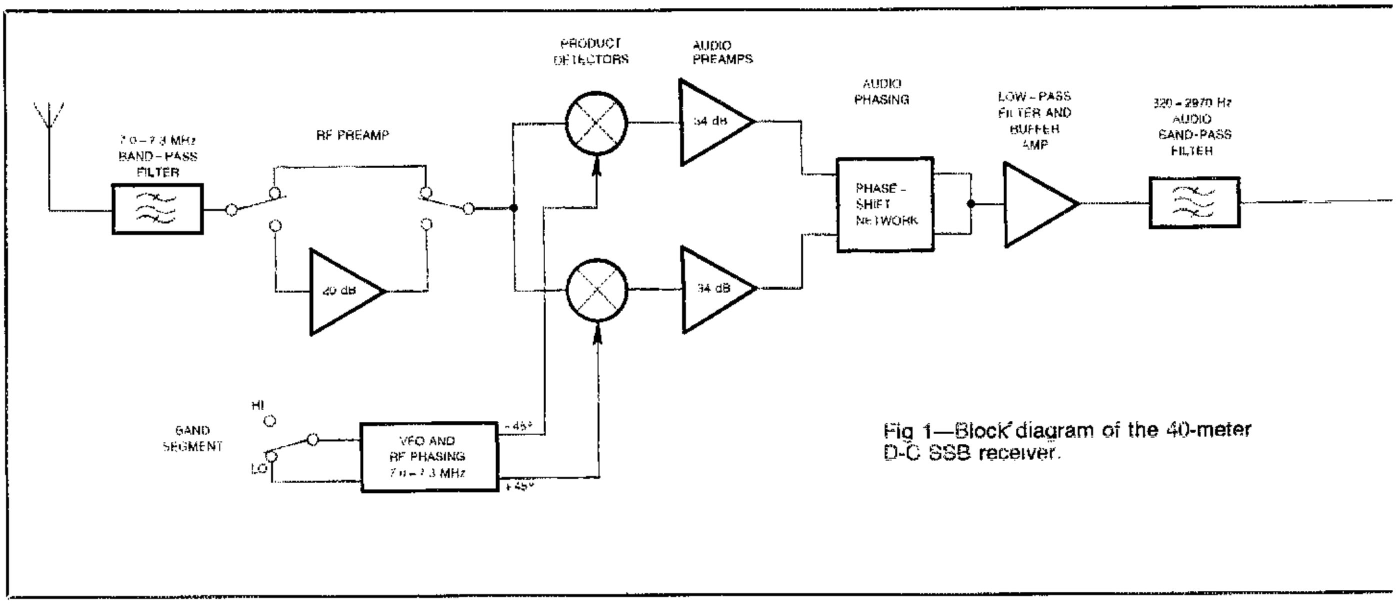


Fig 1—Block diagram of the 40-meter D-C SSB receiver.

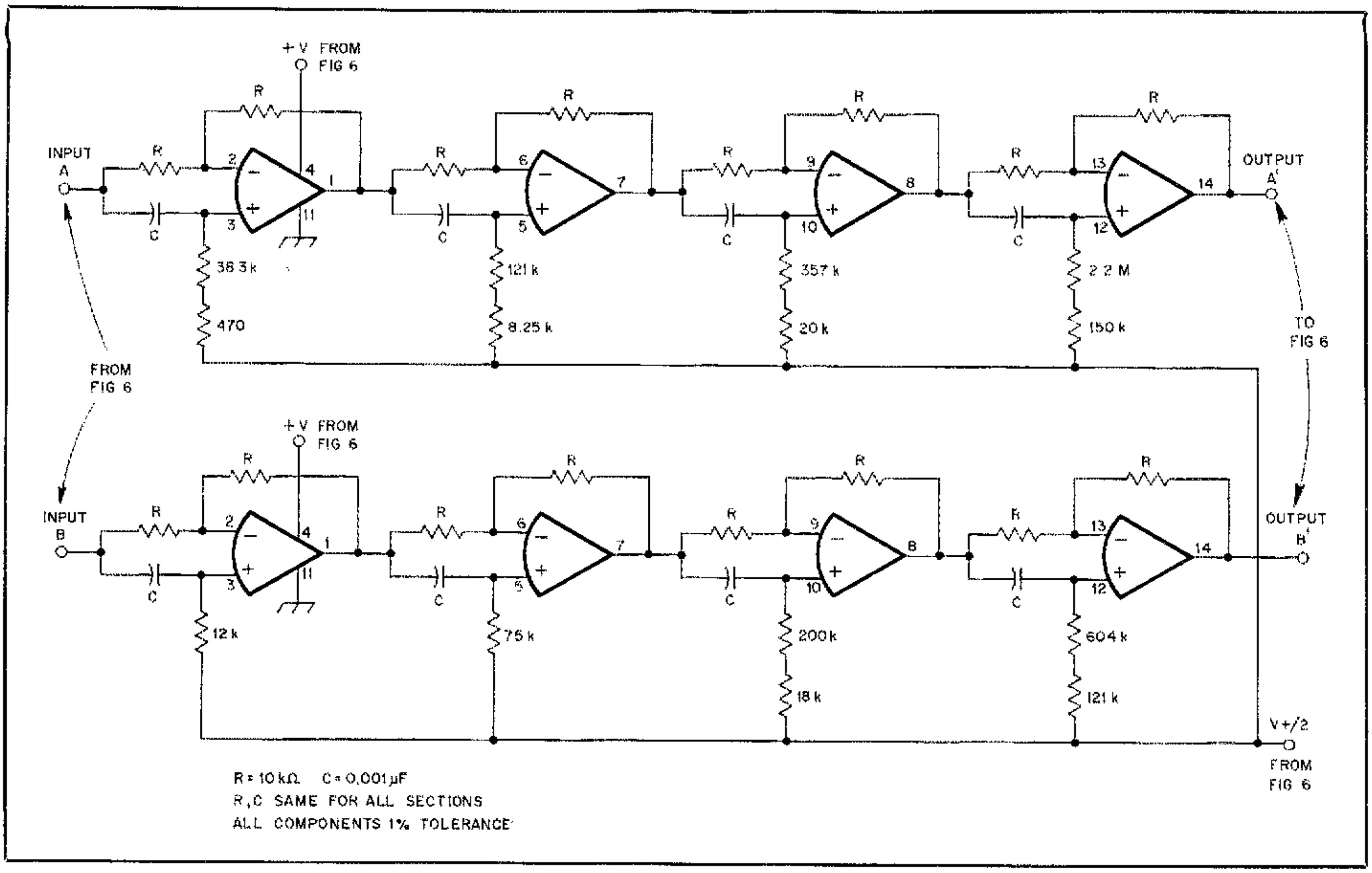


Fig 2—The 90° audio phase-shift network used in the receiver. Each leg consists of cascaded all-pass sections.

nated in 50-ohm resistors, but when connected to the product detectors, it does not provide the desired phase shift. There is enough reactance in the doubly balanced mixers to cause a significant phase variation. Through trial-and-error substitution I determined the final values shown in Fig 4. If you use this circuit in another application, start with the proper theoretical value capacitors (442 pF). Substitute capacitance values if balance cannot be obtained within the range of the trimmers.

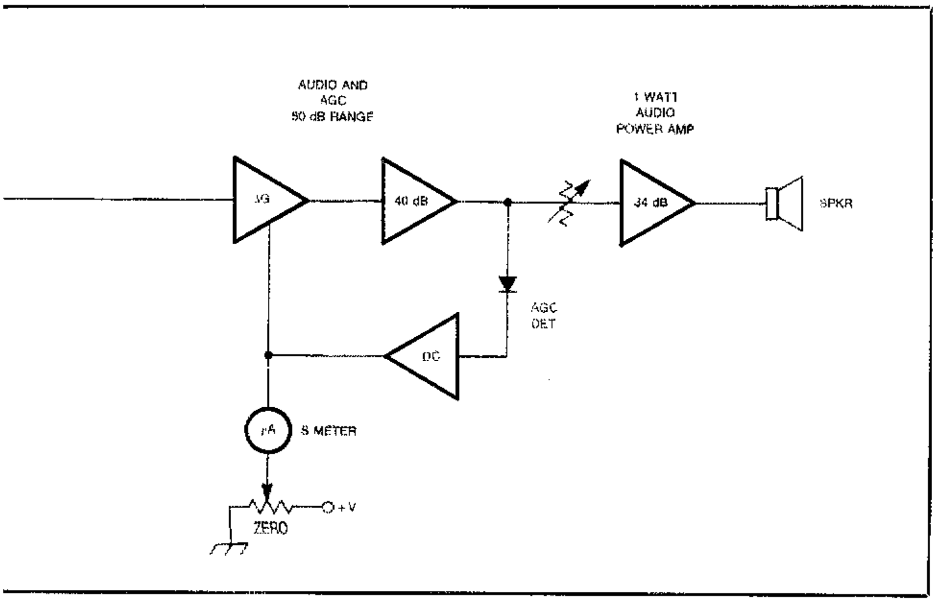
The phasing network follows a "standard" VFO, variations of which have been published in several ARRL publications. As shown in Fig 4, the series-tuned VFO has the tuning capacitor placed across the feedback capacitors, to obtain better tuning linearity. The only unusual feature of my application is band switch S1, which is used to get full 40-meter band coverage with the capacitor and reduction drive I had on hand.

The SSB Detector

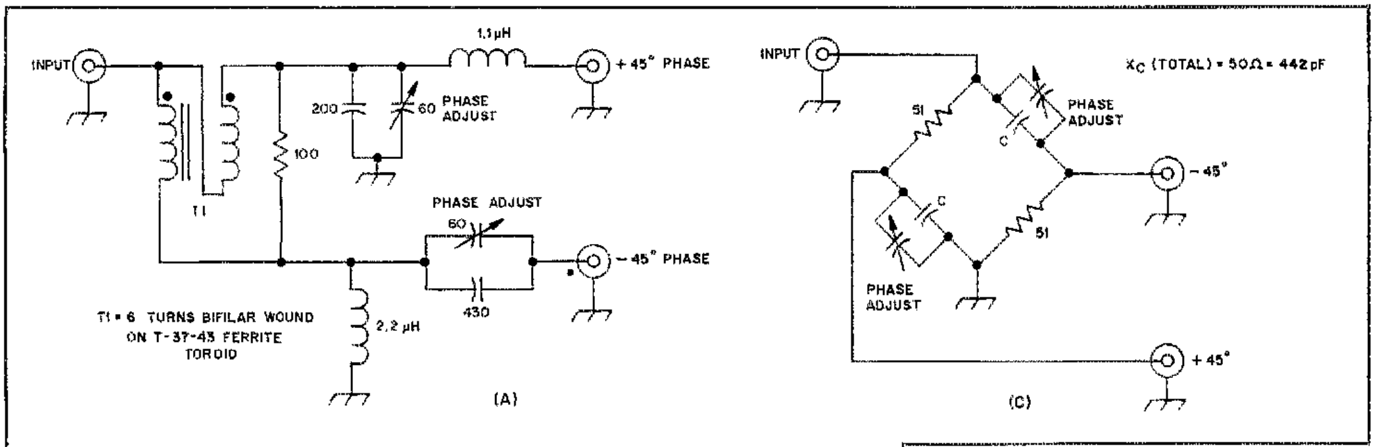
With both RF- and audio-phasing networks determined, I could proceed with the project. Fig 5 shows the front-end filter and RF preamplifier used in the receiver. A double-tuned, top-coupled filter provides adequate selectivity, with band edges down about 3-4 dB from the band center. The switchable RF preamp is another example of how recently developed components simplify receiver design. For the active device, I used an NE5205, which has 20 dB gain with matched 50-ohm inputs and outputs, in an 8-pin package. All that is needed is coupling capacitors and power supply bypassing. RF design was never easier!

The voltage regulator circuit is "kluged" by raising the ground pin of the 7805 above ground. This provides the desired 6.5 volts for the NE5205. Rather than use this method, you could use an LM317 adjustable regulator, or you could simply use a series connected dropping resistor of approximately 180 ohms in place of the regulator.

The phasing detector (Fig 6) starts with two SBL-1 diode doubly balanced mixers as product detectors. The outputs are terminated in 51 ohm resistors, and the audio voltage across them fed via RFCs to NE5534 low-noise op amps. I set the gain of the audio preamps at 33 dB after I discovered that higher gain caused audio clipping before the front end experienced overload. The final gain distribution of the audio stages now has a clipping threshold about equal to the overload threshold of the product detectors.



At the output of the phasing network,



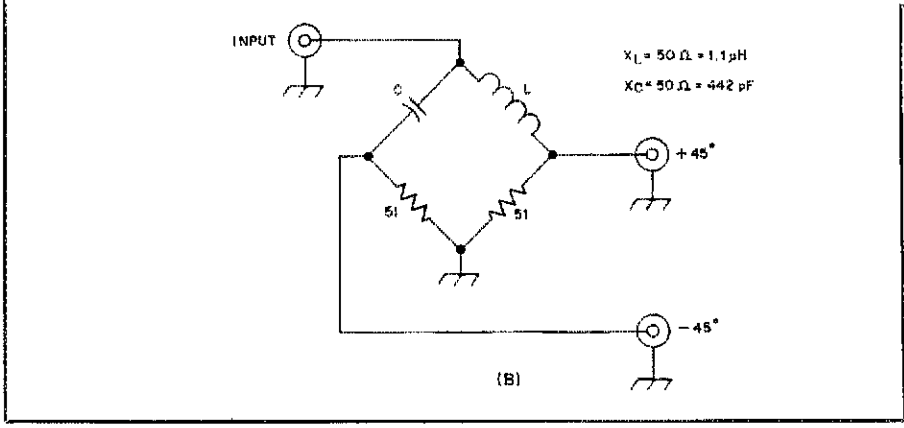


Fig 3—RF 90° phase-shift networks evaluated for the receiver. At A, an RF phasing network using a 0° hybrid and two LC phase-shift networks. At B, a simple RL and RC phase shift network. At C, a dual RC version of the above network shown at B; this is the network used in the receiver.

two signal paths are summed through a BALANCE potentiometer that compensates for amplitude differences in the two audio chains. A single TL081 op amp functions as a buffer stage, and adds two poles of low-pass filtering. Note that there is only one control to adjust in this circuit, in addition to the two trimmer capacitors in the RF phasing network.

Audio Filters

Using the phasing method only eliminates part of a crystal filter's function: removing the opposite sideband. The other filter function is to establish the overall band-pass characteristic. In a phasing rig, band-pass filtering is accomplished at audio, where circuit layout is not critical and common components may be used.

This filter circuit has four cascaded

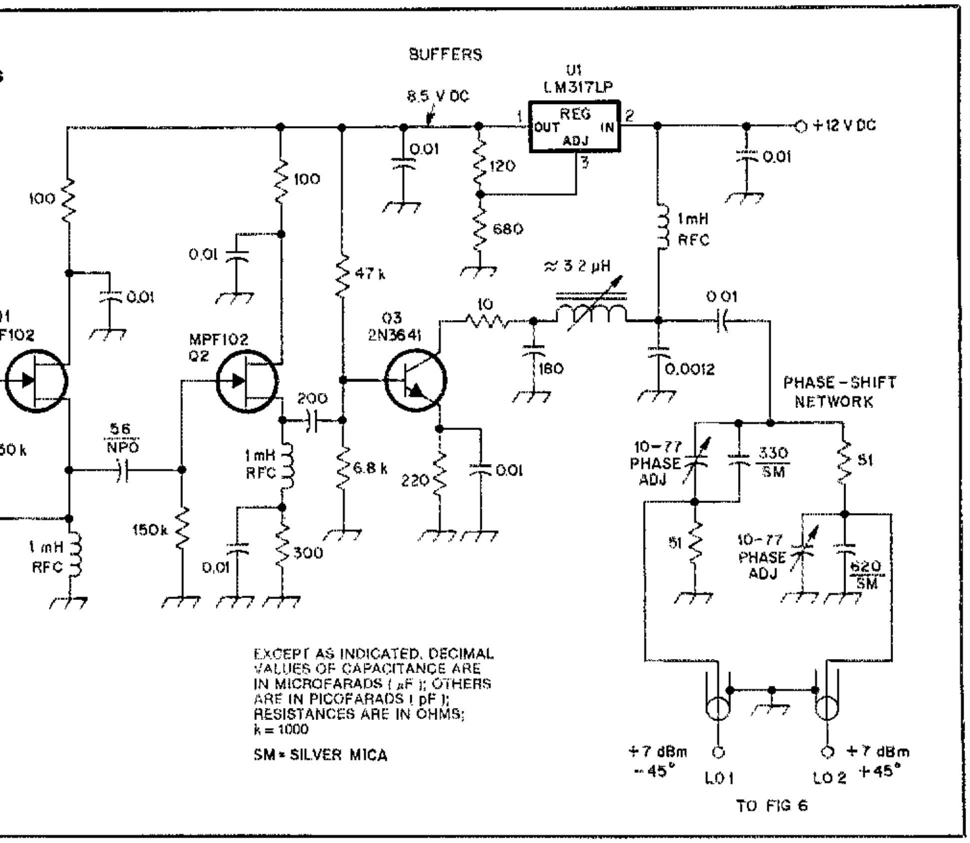


Fig 4—VFO and RF phase shift network. The 77-pF trimmer capacitors control RF phase shift.

C1—250 pF. Approximate—built from a larger value capacitor with plates removed to provide the desired band spread.

L1—35 turns no. 28 enamel on ½-in ceramic form.

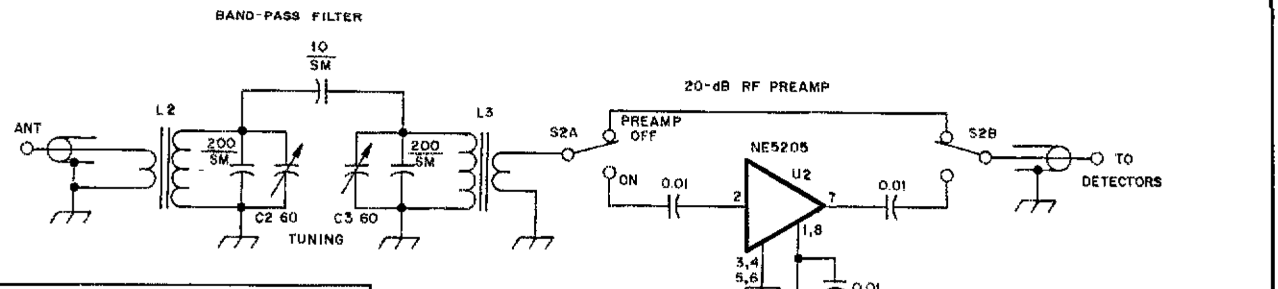
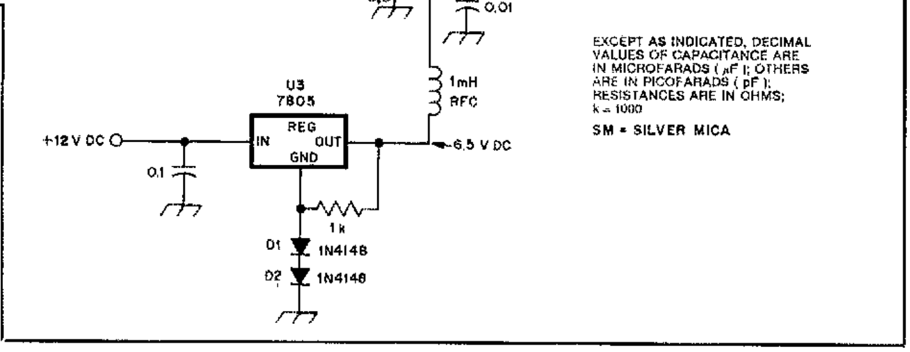


Fig 5—Front-end filter and RF preamplifier circuit. The 60-pF trimmer capacitors, C2 and C3, are adjusted for a flat response from 7.0 to 7.3 MHz.

L2, L3—20 turns no. 28 enamel on a T-50-2 toroid core. Link is 3 turns no. 28 enamel wire.



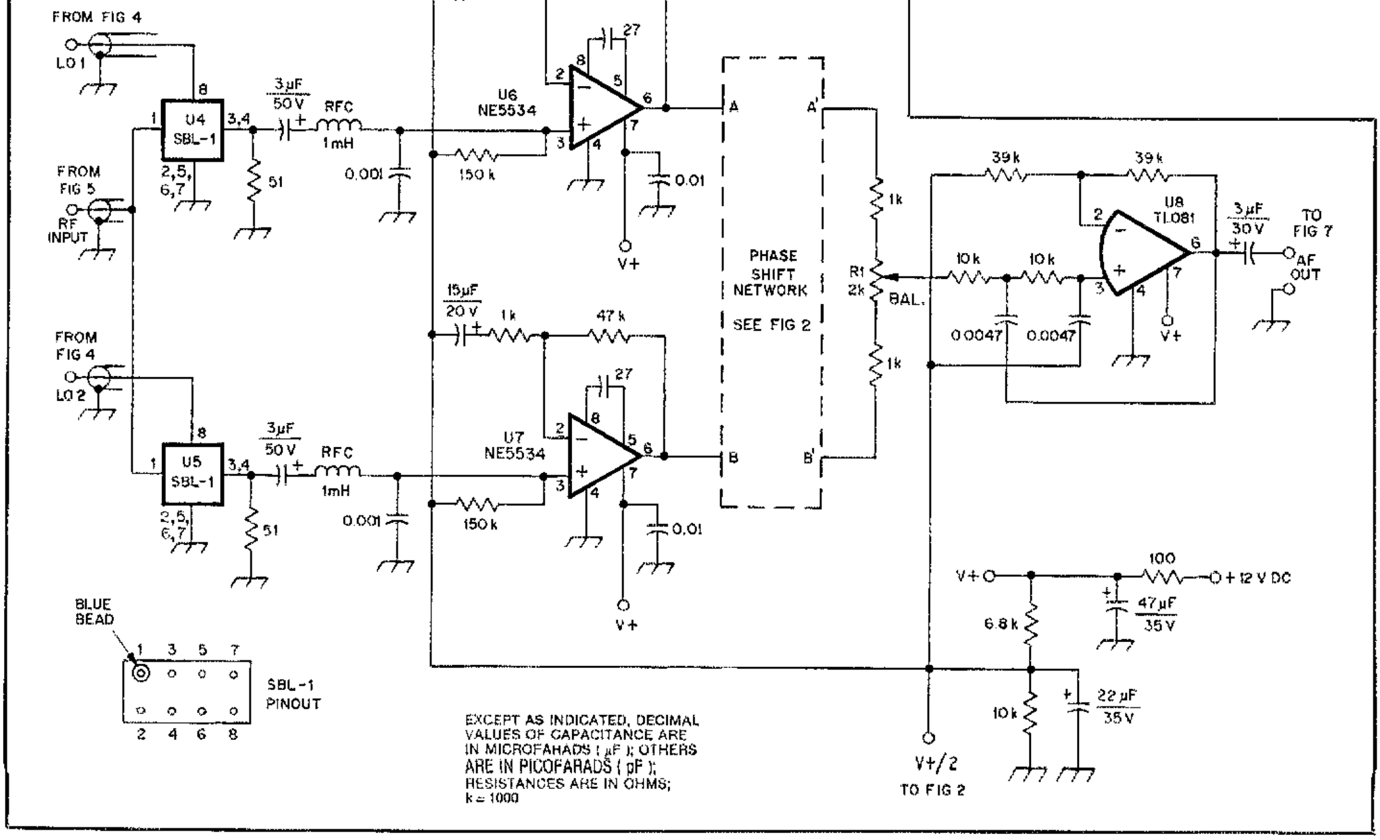
sections, as can be seen in Fig 7. The first section is a non-ideal elliptic low-pass filter, designed for maximum attenuation of frequencies immediately above the cutoff frequency. The filter uses standard 88-mH telephone toroids, with the capacitor values and driving impedances determined by the fixed inductor values and cutoff frequency. (One advantage of op amps is that an arbitrary impedance can be accommodated just by using proper value resistors.) Response of this filter section is shown in Fig 8A (solid line).

The second filter section is a four-pole active high-pass design, with a 1-dB-ripple Chebyshev characteristic. Component values for this filter, like those in the buffer/filter on the detector board, were taken from published design tables.[4]

The third section is another low-pass filter using 88-mH inductors, but using a 0.5-dB-ripple Chebyshev design. Again, the driving impedance and capacitor values were determined by the fixed characteristics of the passband and the inductor value. This section's response is shown in Fig 8A (dashed line). The fourth filter section is identical to the second—a four-pole high-pass active filter.

Some trial-and-error experimentation was used to optimize the final LC filter response, since 10% tolerance capacitors

Fig 6—The phasing SSB detector circuit. Phase shift network circuit is shown in Fig 2. R1—2-kΩ, 10 turn, linear taper, PC mount.



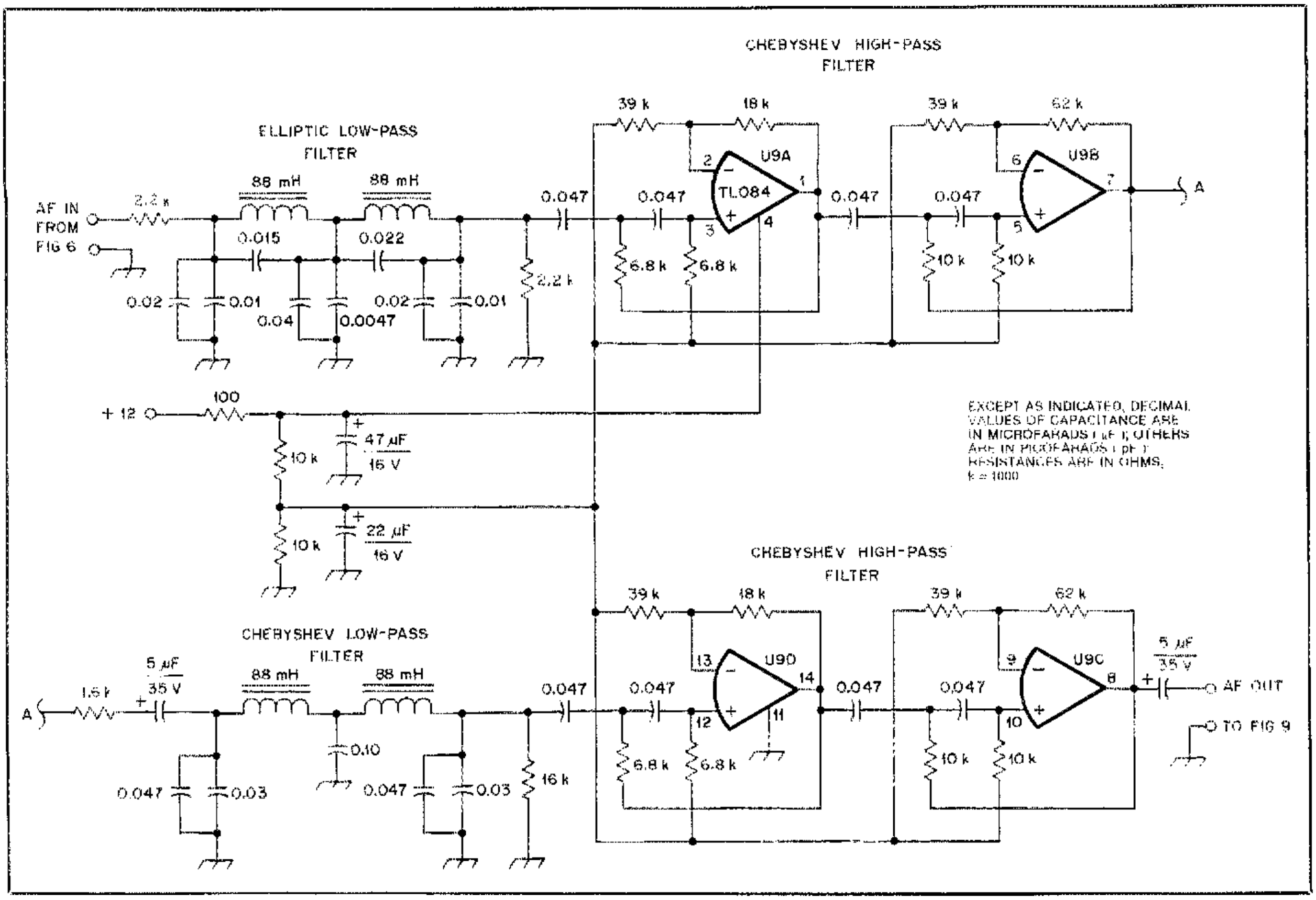


Fig 7—Circuit diagram of the four-section audio filter.

are used, and the inductors are of unknown accuracy. Adequate performance will result if you don't bother with the tweaking. Overall passband response of the cascaded filter assembly is shown in Fig 8B. The shape factor between −6 and −60 dB is 1.27, surpassing all but the best (and most expensive) crystal filters.

Audio AGC and Power Amplifier

To make the receiver more useful for casual listening, I included an AGC stage (Fig 9). The biggest drawback of any D-C design is that AGC must be accomplished at audio. The slow variations of an audio waveform limit the AGC attack time. Getting a simple circuit to work without pops or blasts is pretty tough. The circuit used here is a compromise; it has about 50 dB of AGC range and adequate listening characteristics. The attack time is fast, with some popping evident when using headphones. Popping is not noticeable using the speaker, except on very strong signals.

The circuit uses an MC3340P audio attenuator IC as the gain control element, followed by a 40-dB-gain op amp stage for loop gain. A sample of the audio output is rectified and used to develop the AGC control voltage. The attack time is set by series resistor R3; decay time is determined by the parallel combination of R2 and C4. Another op amp section provides dc offset and some dc gain before driving the MC3340P control pin.

A 200-µA meter indicates the swing of the AGC control voltage, serving as a useful S meter. The meter that I used is a CB-style unit that I found at a local surplus store. Potentiometer R6 provides dc balance for zeroing the meter. Another potentiometer, R5, is a full-scale SENSITIVITY control. The meter zero changes with power supply voltage and is a handy battery condition indicator! Zero moves up scale as the voltage drops.

Audio power is provided by an LM380 amplifier IC. More than enough audio output is available. A three-inch speaker provides plenty of volume.

Alignment

Alignment of the receiver begins with the VFO, the only place where really sensitive adjustments have to be made. With S1 (see Fig 4) in the HI position, the 15-pF trimmer is adjusted for resonance at 7150 kHz when C1 is fully meshed. Use a frequency counter or receiver to check the VFO frequency when C1 is rotated to its fully open position. Remove plates from the capacitor as needed to get the desired 7150 to 7300 kHz coverage. The trimmer capacitor will need to be readjusted each time plates are removed from C1.

Once the right band-spread is reached, tune the VFO to 7150 kHz and change S1 to the LO position. Adjust the 12-pF trimmer for 7000 kHz output. Return S1 to the HI position (operating frequency of 7150 kHz), and adjust the slug tuned coil in the Q3 collector circuit for maximum output.

The next adjustments are in the AGC circuit (refer to Fig 9). With no signal at the receiver input, set the output of U11A to 2.5 volts dc by using R4. Next, set the meter to zero with R6. Later, when the receiver is in normal operation, R5 will be used to set the meter sensitivity.

At this point, signals may be heard when the receiver is connected to an antenna. A signal generator, crystal calibrator or other stable signal source will be needed for the remaining adjustments. First, the input band-pass filter (see Fig 5) is peaked at 7150 kHz using trimmers C2 and C3. Because this is an over-coupled filter, it is normal for this peak to be broad, or to show a peak at two different settings of the trimmers. After getting the best possible peak at 7150, tune to 7100 and adjust C3 for a peak. Now tune to 7200 and tune for a peak with C2. Repeat these last two adjustments. The filter is now relatively flat over the entire 40-meter band.

The following phasing adjustments are all made near 7150 kHz. First, tune in the signal source on the lower sideband as

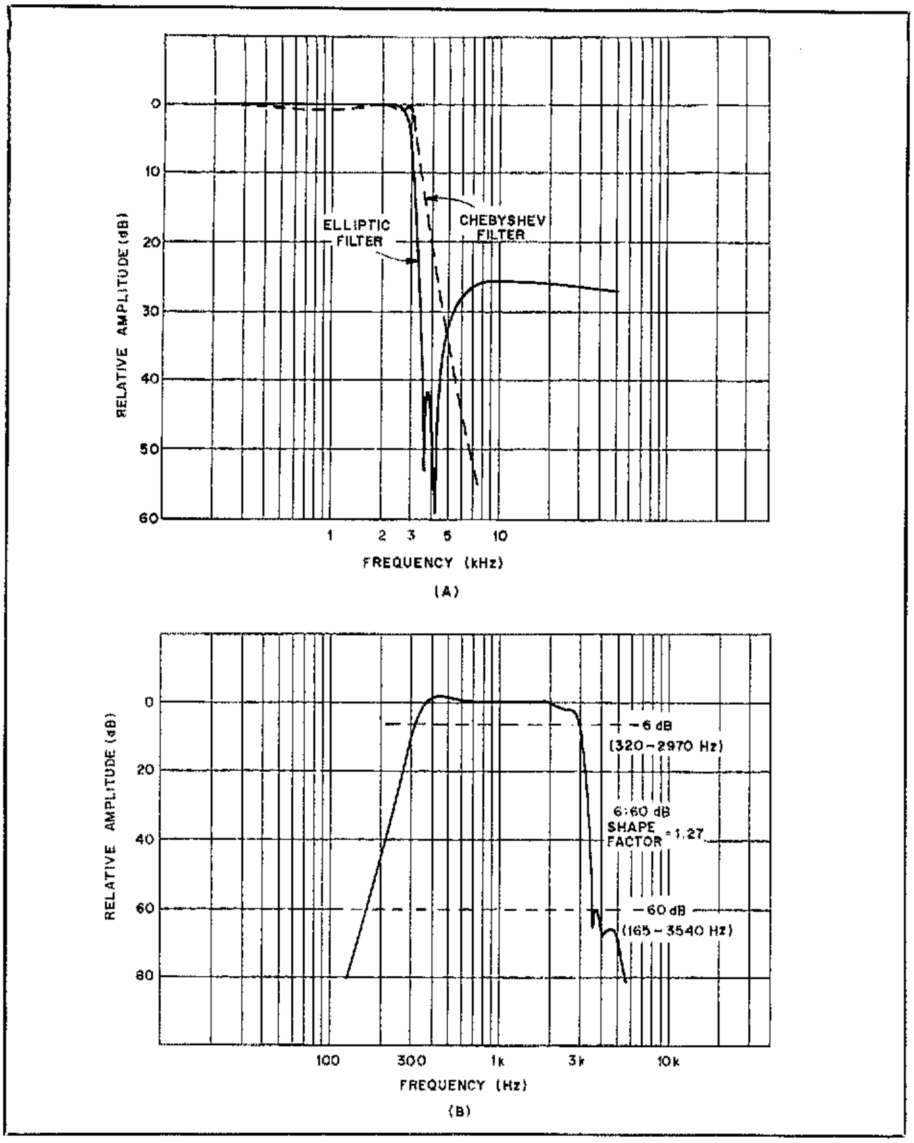


Fig 8—Band-pass characteristics of the two low-pass filter sections presented in Fig 7 are shown at A. Overall audio filter response is shown at B.

Table 1

Unwanted Sideband Rejection as a Function of Audio Frequency and Operating Frequency

Audio Freq	*Operating Frequency (kHz)*				
	7000	*7075*	*7150*	*7225*	*7300*
350	36	44	42	38	34
500	40	35	39	37	34
750	42	41	44	45	41
1000	41	47	65	50	45
1500	41	53	45	55	50
2000	40	50	42	46	54
2500	41	49	42	45	53
2900	43	50	42	47	52

Opposite Sideband Rejection (dB)
(Balanced at 7150 kHz, 1-kHz audio)

verified by listening to SSB stations. Tune through zero-beat to the upper sideband, which should be 10 dB or more lower in level than the lower sideband. If it is stronger, reverse the phasing by switching the local oscillator connections to the two product detectors.

Now tune the receiver for a 1 kHz audio note on the upper sideband. Adjust the BALANCE potentiometer, R1 (Fig 6), for minimum output. The change should be quite noticeable. Next, alternately adjust the two trimmer capacitors in the RF phase shift network (Fig 4) for best null. If a definite null cannot be found, some changes in capacitor values may be needed, as noted before. Once all is in order, you should get a very distinct and deep null in the audio output. Repeat the adjustment of the RF phasing and audio balance controls until the best null is achieved. You should be able to make the upper sideband virtually disappear. Tuning back through zero-beat should make the lower sideband magically reappear! All that remains is to set the S meter sensitivity as you wish.

Performance

This receiver was constructed as an experiment. I set out to see what level of performance could be achieved using the phasing method. I am definitely not disappointed.

First, unlike most D-C designs, this receiver has absolutely no microphonics. At each board, power supply connections are decoupled with 100-ohm resistors and large electrolytic capacitors. In addition, the op amps have internal power-supply isolation.

Hum is only a minor problem with this receiver. This is not true with many D-C designs. Magnetic coupling to nearby power transformers is one small difficulty. (I can't use this receiver within about 18 inches of the main rig's power supply.) Hum from local oscillator radiation, 60-Hz modulation and reception of the modulated LO signal is minimized by shielding the VFO and using metal-can mixer/detectors. With the RF preamp in the circuit, reverse isolation reduces LO radiation even more. There is significant hum when using the combination of an ac-operated power supply and a random-wire antenna attached directly to the antenna connector. With an outdoor antenna, there is no difference in performance with batteries or ac-operated power supply.

Outstanding sensitivity and dynamic-range performance were not major goals of this project. Demonstration of the phasing method was the primary objective, but RF performance is certainly adequate. I estimate the noise floor of the receiver to be about −100 dBm without the preamp and −117 dBm with it. This kind of performance is adequate on 40 meters, and I can hear every signal on this receiver that I can hear on my Heathkit SB-104A transceiver.

The onset of third-order intermodulation distortion from two signals spaced 20 kHz apart was measured at a very respectable −32 dBm without the preamp and −49 dBm with the preamp. Spurious-free dynamic range is then 68 dB. My feeling is that this estimate of the noise floor may be too conservative, but I don't have access to the proper test equipment for minimum discernible signal (MDS) measurement.

The best news comes last: Unwanted sideband rejection is better than expected. Table 1 lists the rejection levels versus audio frequency at several places in the 40-meter band. Note that only at the lowest audio frequencies does the rejection ever fall below 40 dB! The simple RF phasing network and the one-adjustment audio phasing combine for fine performance. Why is it so good? Can it be even better?

First, there is an element of luck in the performance over the wide (4.2% bandwidth) 40-meter band. It seems that there is some compensation between the variations in the audio and RF networks that works to my advantage by maintaining good rejection out to the band edges.

Next, the output amplitude of the audio phase shift network varies somewhat from

Understanding SSB Reception by the Phasing Method

Disadvantages of phasing-type SSB reception:

- Requires RF and audio phase shift networks.
- More difficult to tune and adjust.
- Unwanted signal rejection is better with a good filter-type receiver.

Advantages of phasing-type SSB reception:

- Does not require an expensive filter.
- Useful over a range of frequencies (filters are not).
- Ideal for a D-C receiver.

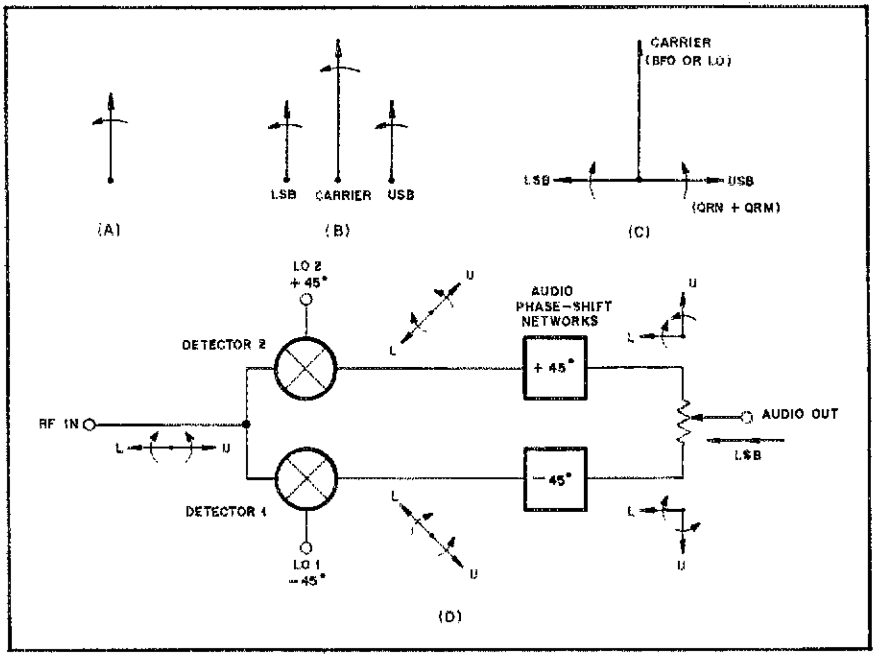
How it works

Any radio signal can be represented as a rotating vector (see Fig A). The rotational speed of the vector corresponds to the signal frequency. For example, the vector that represents a 7.2 MHz signal rotates 7.2 million times per second. The vector length corresponds to the signal amplitude.

Fig B shows three vectors. They represent a carrier and upper (USB) and lower (LSB) sidebands. The USB, being higher in frequency, rotates more rapidly than the carrier, which rotates more rapidly than the LSB. In Fig C, the phase relationship between the three, at an arbitrary instant, is shown with the carrier used as a reference. Vector rotation for USB and LSB, in this case, is shown as the difference between their actual rotation and the rotation of the carrier vector. In the receiver built by Gary Breed, the LSB is the desired signal. The USB consists of noise, interference or unwanted signals. The carrier appears as the LO signals that are used in the product detectors.

A block diagram of a phasing-type SSB receiver similar to Gary's is shown in Fig D. Assume that the same phase relationship that appears in Fig C exists at the receiver input. The local oscillator (LO) represents the carrier (reference). In Detector 1 the carrier is delayed by 45°, and in Detector 2 it is advanced by 45°. This shifts the phase relationship between sidebands in the detector outputs as shown in the diagram. The audio phase shift networks cause a phase shift difference of 90° between the two. In this case, we assume that the upper network has a +45° shift and that the lower network has a −45° shift. This means that the vectors representing signals in the upper path move forward 1/8 of a revolution, while vectors in the lower path move backward the same amount. The result is that LSB energy from the two paths will add, and USB energy will cancel. For a USB receiver, you can swap the LO connections between the detectors, or swap the inputs to the audio phase-shift network.

It is *very* important that a 90° phase difference exists between LO1 and LO2, and that the two audio phase-shift networks exhibit a 90° phase difference. There is nothing "special" about ±45° phase shifts. For example, 20° and 110° would do fine. You can use vector analysis, as shown here, to prove that to yourself.—*Chuck Hutchinson, K8CH*



At A—a signal can be represented as a vector (arrow). The curved arrow shows the direction of vector rotation. At B—vector representations of a carrier with sidebands. The USB rotates faster than the carrier, and the LSB rotates slower than the carrier. At C—vector representation of the carrier and sidebands. In this case, the carrier appears as a fixed reference. Indicated rotation is relative to the carrier vector rotation. At D—block diagram of a phasing-type SSB receiver showing phase relationships at an arbitrary instant.

high to low frequencies. I found that every audio frequency could be balanced to better that 50-dB rejection just by adjusting the BALANCE potentiometer after the audio networks. If this variation can be removed, truly high performance could be achieved.

The most encouraging result is that the phase response of the audio network is outstanding—better than 0.5°. Also, the performance of the simple RF phasing is surprisingly good, once the external influences are compensated for. I doubt that this level of performance could be maintained if switching between LSB and USB is needed. Further optimization of the circuitry would be required, but that's what experimental designs are all about.

What about subjective performance observations? When I demonstrate the receiver to my friends, the first comment is usually, "It sounds so good!" It's true—this is a clean-sounding receiver. Because my receiver has a little wider passband than commercial rigs, it gives an impression of better audio fidelity. Less than 1% distortion in the audio stages also adds to this effect.

Most of all, this receiver has the distinctive D-C clarity. This is not an illusion—it is lack of distortion of the received signal. Crystal filters have a definite time-domain response (ringing), particularly when driven with fast-rise-time pulses such as impulse noise (QRN). Audio filters exhibit such responses to only a minor degree, and then only at the transition frequencies near cutoff. This "clean" characteristic will exist in any phasing receiver, even a superheterodyne, as long as there is no crystal or mechanical filter.

The phasing method can provide economical SSB performance. A little more

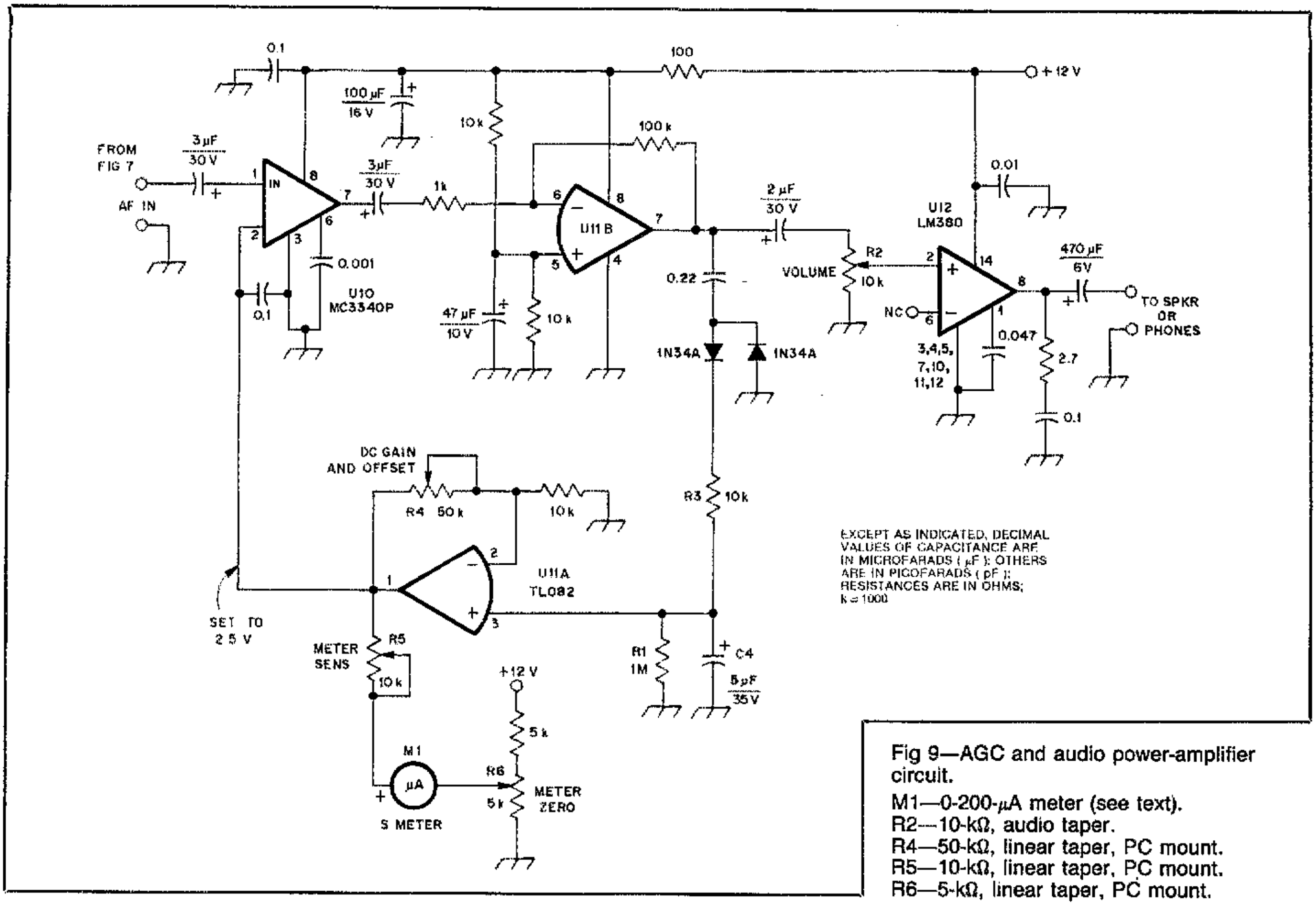
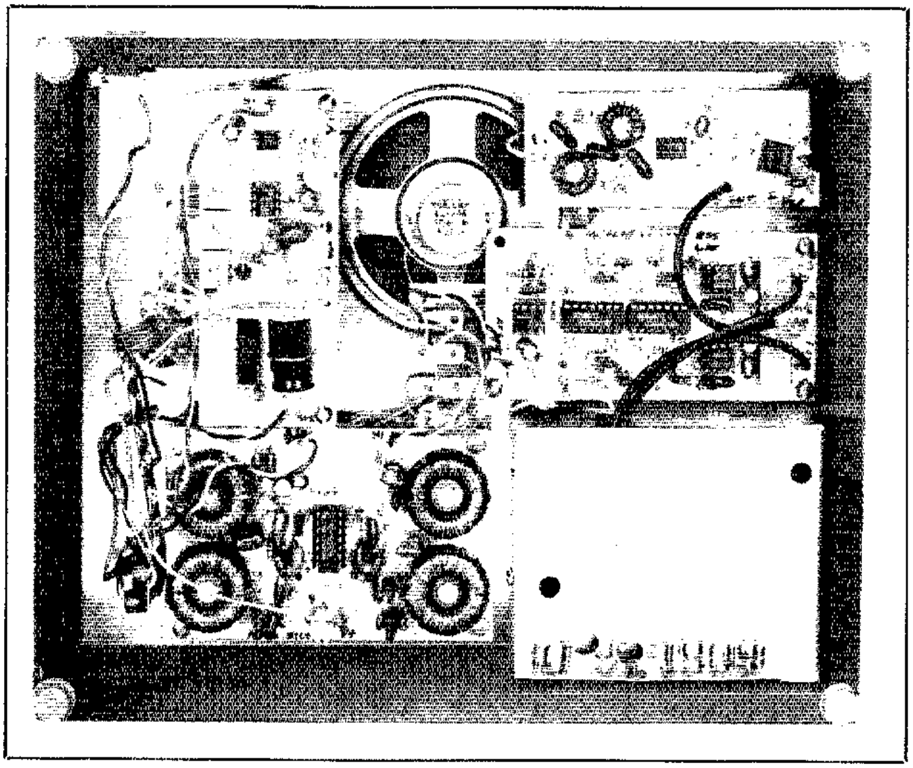


Fig 9—AGC and audio power-amplifier circuit.

M1—0-200-µA meter (see text).
R2—10-kΩ, audio taper.
R4—50-kΩ, linear taper, PC mount.
R5—10-kΩ, linear taper, PC mount.
R6—5-kΩ, linear taper, PC mount.



The receiver consists of five modules. Clockwise, from the upper right, they are: input filter and RF preamp, phasing detector, VFO and RF phasing (in shielded box), audio filter and AGC and audio power amplifier.

precision in AF and RF phasing networks is all that's needed. This receiver demonstrates that good phasing SSB is indeed within reach.

Gary A. Breed was licensed in October 1961 as WN9AYP. He upgraded to Technician in 1962, and that same year moved up to General class. Gary became an Extra Class ham in 1971, and received his current call sign, K9AY, in 1977. He prefers 40 meters, but occasionally operates on the other bands. Gary spent 15 years in broadcasting, eight of them as radio and television station chief engineer, and seven as a consultant. He's now the editor of RF Design *magazine, a technical journal boasting 34,000 readers—12,000 of them ham radio operators.*

Printed Circuit Boards for this receiver are available from A & A Engineering, 2521 W. LaPalma Ave Unit K, Anaheim, CA 92801, tel 714-952-2114.

Notes

[1]Arthur B. Williams, *Electronic Filter Design Handbook* (New York: McGraw-Hill, Inc 1981), pp 7-10 through 7-34.
[2]S. D. Bedrosian, "Normalized Design of 90° Phase-Difference Networks," *IRE Transactions on Circuit Theory*, June 1960, pp 128-136.
[3]One-percent-tolerance resistors and polyester film capacitors are available from Mouser Electronics, 2401 Highway 287 North, Mansfield, TX 76063.
[4]Don Lancaster, *Active Filter Cookbook* (Indianapolis: Howard W. Sams & Co, Inc, 1975), Chapters 4 and 8.