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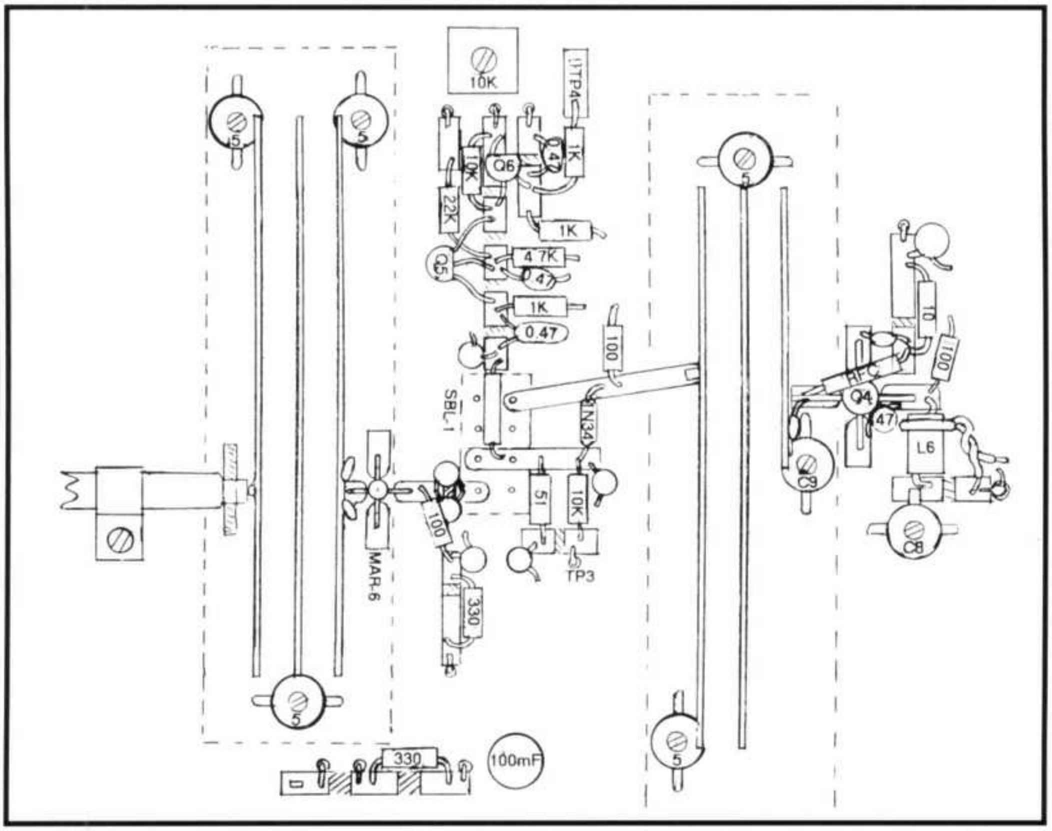
ARRL Experimenter's Exchange

April 1994



Homebrew Your PACSAT Receiver

QEX: The ARRL
Experimenter's Exchange
American Radio Relay League
225 Main Street
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QEX

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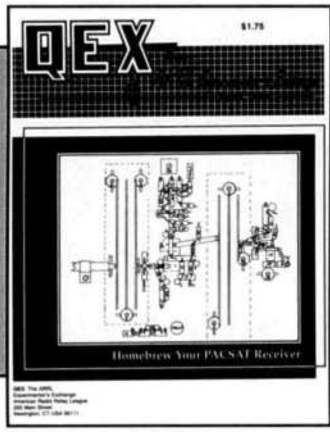

About the Cover:
W6IOJ's homebrew PACSAT receiver is built "dead bug" style on unetched PC board.

ISSUE NO. 146

Features

April 1994 QEX Advertising Index

THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interests in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

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1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio field
3) support efforts to advance the state of the Amateur Radio art

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Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking

TAPR Looks Forward

The folks at Tucson Amateur Packet Radio (TAPR) are at it again—trying to drag the rest of packet radio kicking and screaming into the future. At the 1994 TAPR Annual Meeting, held in Tucson on March 4-6, new TAPR projects were described, and talks and papers were presented that show that fresh directions in packet radio are still possible—and being pursued.

Among the highlights of the meeting were a talk by Bob Stricklin, N5BRG, about the TAPR DSP-93 project, now nearing beta test, and a progress report from Phil Karn, KA9Q, on his efforts to devise a new, robust, link-layer protocol that provides superior performance (compared to AX.25) by use of forward error correction (FEC) and improved collision avoidance. Also reported at the meeting, by Lyle Johnson, WA7GXD, was an exciting opportunity, supported by TAPR, to include a wideband digital transponder experiment on the upcoming AMSAT Phase 3D satellite. And a paper in the proceedings from Kantronics, Inc, introduced their new G-TOR HF data communications protocol for the KAM, which features error correction, data compression and adaptive speed selection.

News items from the meeting include the announcement that next year, for the first time, the annual meeting will be held elsewhere than in Tucson—specifically, in St. Louis. TAPR also took the opportunity to award well-deserved recognition to several amateurs who have worked mightily to make TAPR the vibrant organization it is, including Bob Nielsen, W6SWE, TAPR officer and board member from 1989-1994; Jerry Crawford, K7UPJ, board member from 1991-1994; Dan Morrison, KV7B, long-time contributor and board member from 1983-1994, and Heather Johnson, N7DZU, who served as TAPR's office manager from 1989 to 1994, and from 1983-1985 as an officer.

In a sad note, TAPR President Greg Jones, WD5IVD, noted the passing of former TAPR President Andy Freeborn, NØCCZ. TAPR is accepting donations in Andy's name for the American Cancer Society. (Donations can be sent to TAPR, 8987-309 E. Tanque Verde Rd #337, Tucson, AZ 85749-9399, or call 817 383-0000.)

The Saturday conference had presentations which included, along with those mentioned above: "Preliminary Test and Evaluation of Metcon-1," by Ron Bates, AG7H; "219 MHz Report," from Jon Bloom, KE3Z; "RF Path Analysis for Digital Links," by Jim Wortham, W7GNP; "WINNOS: A Graphical Concept," by Lew Shannon, KØRR; "Internet Gateways, Opportunities and Problems," by Remi Hutin, W5/F6CNB; and "New Network Services, Technologies and Requirements," by Tom McDermott, N5EG.

Papers in the proceedings that were not presented at the meeting include: "TUC52 Circuit Description," by Paul Newland, AD7I; "What is TNOS?" by Brain Lantz, KO4KS; and "G-TOR: The New Faster HF Digital Mode," by Phil Anderson, WØXI, et al.

The 53-page *Proceedings of the 1994 TAPR Annual Meeting* is available from TAPR for $6, with a 10% discount for TAPR members.

Kick and scream as we might, aren't we glad to have TAPR doing the dragging?

This Month in *QEX*

Receivers—especially UHF receivers—are just too complicated for the home builder, right? Wrong! John Reed, W6IOJ, proves that with "A Simple Junkbox Satellite Receiver."

Your editor contributes "A CVSD Codec System for Your PC," to get you started using digital voice communications.

Analog phasing SSB generation is a time-honored technique that has always suffered because audio phase-shift networks didn't provide the expected performance. Find out why by performing "Phase-Shift Network Analysis and Optimization," as described by Kevin Schmidt, W9CF.

"Estimating the Length of Coaxial Feeder"; no problem when you're dealing with 10-foot lengths, but how about those 100-plus-foot rolls? George Brown, G1VCY, shows us an elegant way of measuring the length with a simple circuit.

Connecting your VHF transceiver to a microwave transverter requires a step-down in power levels, something most easily done by "Building VHF Power Attenuators," says Paul Wade, N1BWT.

This month, ARRL staffer Larry Wolfgang, WR1B, reviews a handy pocket reference book for the electronics enthusiast.

Pleading the pressure of work, Harold Price, NK6K, defers his next "Digital Communications" column to the June issue. It'll be twice as good as usual, we're sure!—*KE3Z, email: jbloom@arrl.org.*

A Simple Junkbox Satellite Receiver

Use this receiver to monitor the 9600-baud 70-cm PACSATs

By John Reed, W6IOJ

Single-conversion receivers—including direct-conversion designs—have received a good deal of attention recently, primarily because of the excellent article by Rick Campbell, KK7B, in August 1992 *QST*. The receiver described here has been configured for monitoring the 70-cm polar-orbiting PACSATs, and has demonstrated very good performance in spite of the limitations of single-conversion designs. And it has retained the simplicity, compactness and versatility of single conversion. The receiver can operate in any part of the 70-cm band. The output is an IF signal having a 250-Hz to 2-MHz band pass. Fig 1 is a block diagram of the receiver. This diagram shows one particular PACSAT configuration where the receiver is used with a 50-kHz IF filter/amplifier/FM discriminator, a 9600-baud modem, and a TNC/computer.

General Description

The receiver's 3×5¼×5⅞-inch metal

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Santa Barbara, CA 93111

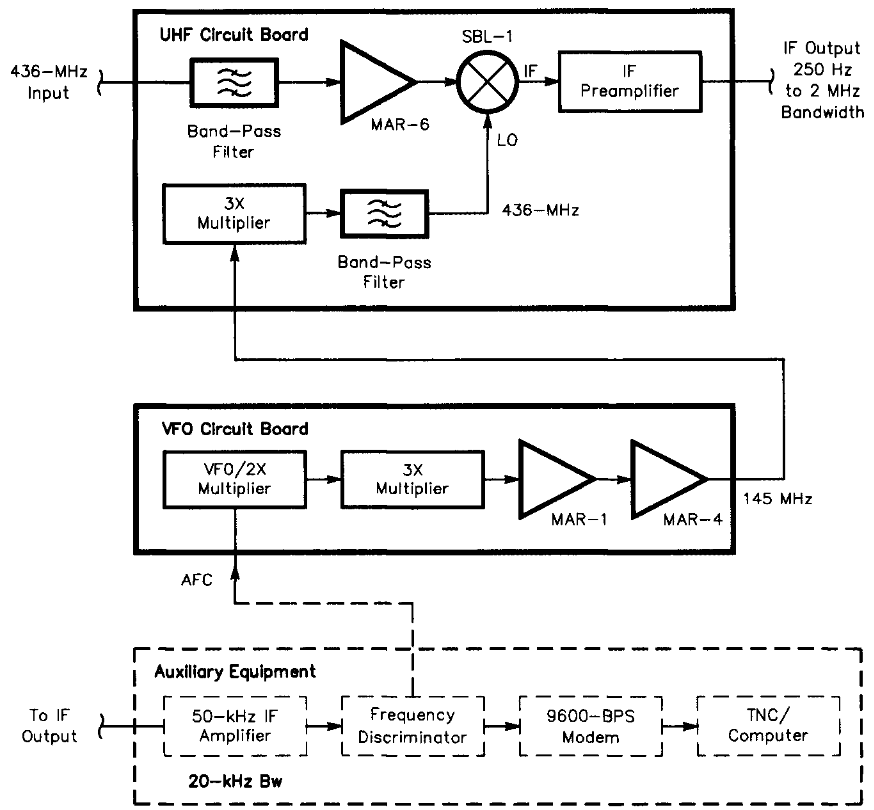

Fig 1—Block diagram of the receiver. The dotted line section indicates a particular configuration for monitoring 9600-baud PACSATs.

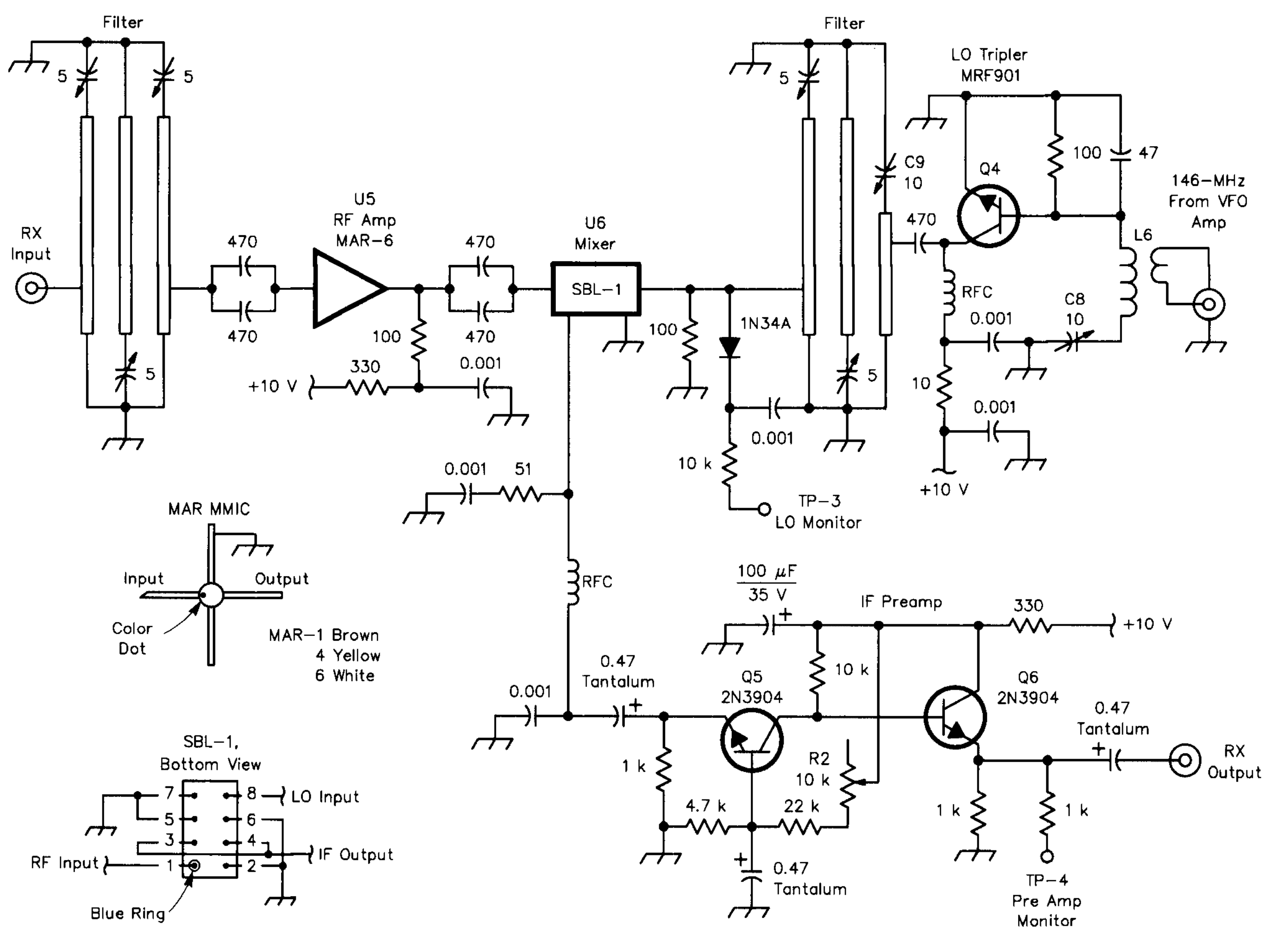

Fig 2—Schematic of the UHF circuit board. See Fig 3 for detail of the UHF filters. All capacitors are 50-V disc ceramics unless otherwise noted. All RFCs are 20 turns of #26 wire having an ID of 1/16 inch. C8 and C9 are 10-pF FILMTRIMS (Sprague-Goodman part #GYA10000). L6 is 7 turns of #18 wire, 1/4-inch ID, 3/8-inch long with a 1-turn link coupling coil that is connected to the VFO amplifier output coupling cable.

cabinet contains two 4¾×5-inch circuit boards. One is the UHF circuit board that has a 70-cm input filter, low-noise monolithic preamplifier, double-balanced mixer and IF preamplifier. There is also an LO driver consisting of a tripler/filter arrangement operating from a 145-MHz, 6-mW source. The VFO circuit board has a 24-MHz varactor-tuned VFO and a 145-MHz frequency multiplier followed by a two-stage monolithic amplifier. This amplifier output is the LO driver input. The VFO circuit board includes a 10-V regulator operating from a 12- to 20-V external source.

On the front panel there is a 12-position band selector switch covering 3.6-MHz of the 70-cm band in 300-kHz steps. Fine tuning within these steps is accomplished with a potentiometer covering a 500-kHz spread. The selector range can be placed in any part of the 70-cm band by trimmer adjustments located on the circuit boards. The back panel contains a BNC connector for the 70-cm input and jacks for the receiver output and the AFC input.

Limitations

Single-conversion receivers lack discrimination of the unused sideband. As an example, in my PACSAT application, with a 50-kHz IF, there will be an image frequency 100 kHz from the received signal. In actual operation I have never experienced interference from a signal at this image frequency while monitoring PACSATs, and in the rare case where there may be interference, you can tune to the opposite sideband, placing the image at a different frequency. Probably of more importance is that unattenuated noise at the image frequency causes a 3-dB S/N degradation. Although this is clearly not optimal, my observation has been that typical signal variations of polar-orbiting satellites are so large that this loss does not represent a major compromise. A second possible limitation is 1/f noise originating from the diode mixer. But practically all 1/f noise is below 10 kHz. Therefore my PACSAT application, with its 50-kHz IF, is not affected. Even in applications requiring the use of lower frequencies, the receiver's RF amplifier will largely override the 1/f noise.

The IF preamplifier has been left wideband simply as a versatility feature. Although the dynamic ranges of both the mixer and IF preamplifier start to roll over at about the same

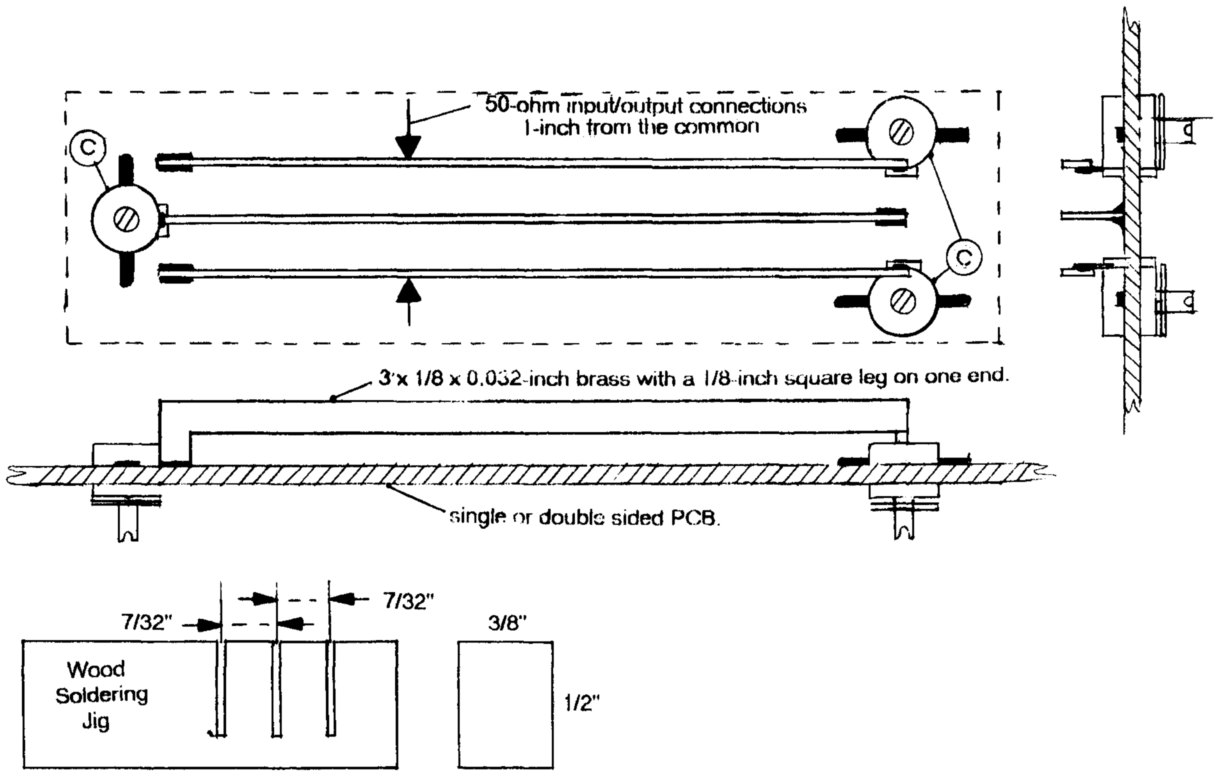

Fig 3—Detail of the UHF filter. The filter is easily assembled using the wood soldering jig to hold the striplines at the proper spacing while soldering. I use a ⅛-inch drill placed between the striplines and PC board to ensure proper height above the mounting surface while assembling. The jig slots are made with a hacksaw, which makes the desired 0.032-inch slot width.

The devices marked C are 1.6 to 6-pF FILMTRIMs, a Sprague-Goodman plastic dielectric capacitor (part #GYA5R000). Also, surplus 2.4- to 9-pF ceramic trimmers worked well.

The local oscillator filter is the same except the input stripline is 1.5-inches long rather than 3-inches long, and the related capacitor, C9, is increased to 10 pF (#GYA10000).

input levels, a filter between the mixer and preamplifier will help to avoid possible overloading effects of unwanted signals that pass through the input filter but are outside the useful passband. As an example, a 50-kHz filter (20-kHz bandwidth) will improve the performance of my PACSAT configuration during some interference conditions.

Frequency stability is a major consideration of simple 70-cm local-oscillator design. One influencing factor in this case is that polar-orbiting satellites have total Doppler shifts of up to 20 kHz. This alone requires automatic frequency control—unless you are willing to keep one hand on the tuner! Of course, if AFC can compensate for Doppler, it can also compensate for some drift in the local oscillator. I arbitrarily selected a tolerable drift of 2-kHz during a 20-minute orbit period, and no more than 20-kHz drift in a 6-hour continuous operating period that includes a 10° (F) ambient temperature shift.

Circuit Boards

The circuit boards are assembled using a glue-down strip-line technique that holds the components and acts as conducting RF lines. Using this method, the printed-circuit board foil remains a solid groundplane, making it appropriate to use a single-sided board. A second feature is that the glued-down pads can be easily removed to accommodate layout changes. The component striplines are about ⅛-inch wide, and the lengths are determined by how many connections are desirable in a single line. The connecting pads are separated by foil notches made with a hacksaw, about 3⁄16-inch apart or longer, depending upon layout convenience. The 50-Ω RF conducting lines are made 3⁄32-inch wide, assuming the use of standard glass-epoxy 0.059-inch material. The width is different from conventional etched striplines due to the raised glue-down stripline edge effect. In this practice, in this particular application, the critical RF lines are so short the type of PC board material is of little consequence. I use Elmer's Clear Houshold Cement for fastening the striplines. The cement sets up enough to use the pad in a few minutes. Removal of a pad becomes a little difficult after setting-up for several weeks or more.

You may notice the VHF interstage link coupling. This technique may be outdated, but it's an effective and forgiving method that allows stages to be separated by one inch or one foot without a significant performance tradeoff. The link provides a common return

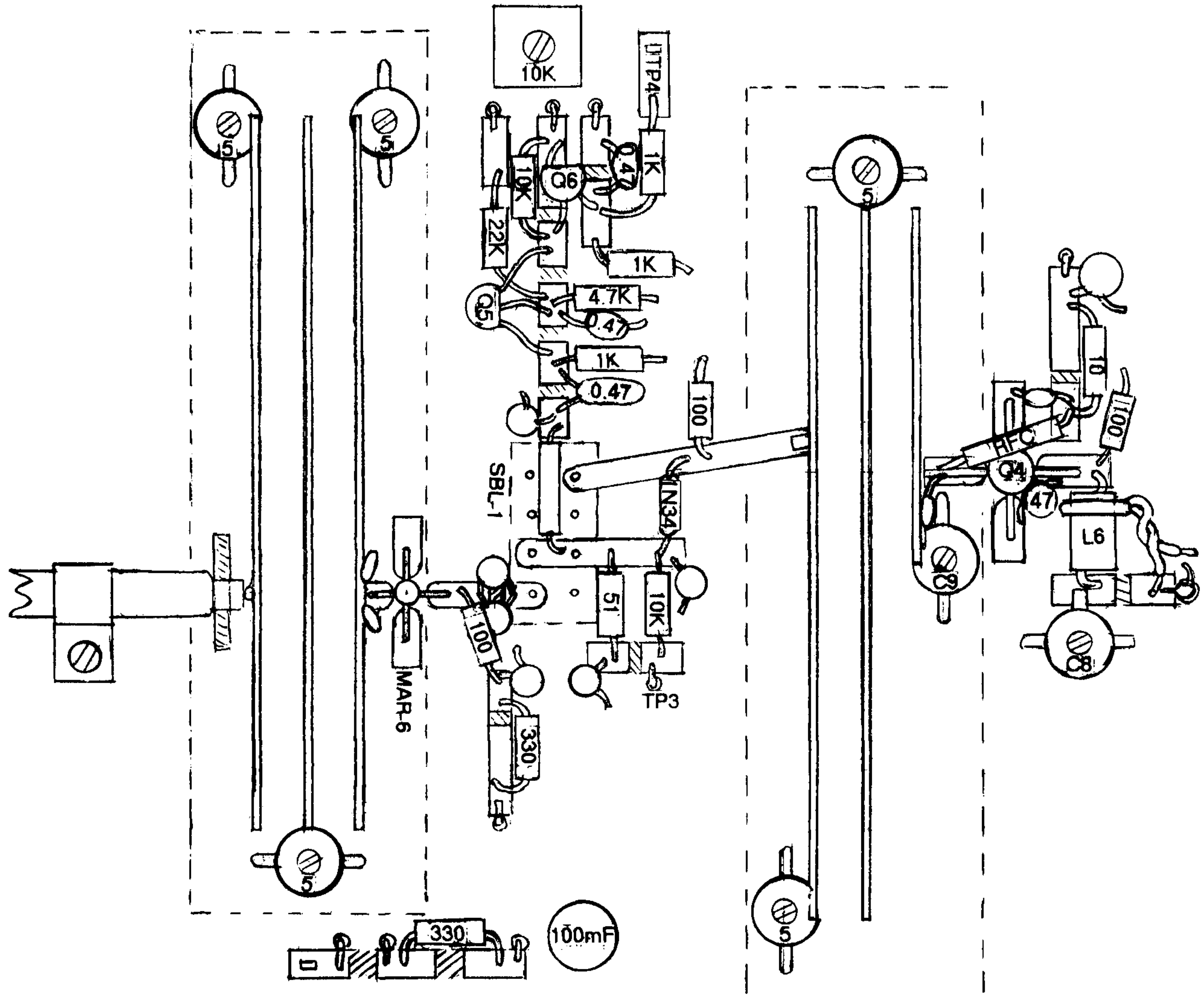

Fig 4—Layout of the UHF circuit board.

that ensures optimum isolation between stages.

UHF Circuit Board

The possibility of overloading of the preamplifier by off-frequency interference is minimized by first passing the input signals through a three-section stripline filter. It has an insertion loss of 0.7 dB. As shown in Fig 3, the filter is easy to build from materials that are not difficult to find. Although the diagram specifies a Sprague capacitor, which is available from many sources (Digi-Key as an example), there are inexpensive surplus miniature ceramic trimmers that will work fine as long as the minimum capacitance is 2.5 pF or less.

The MAR-6 preamplifier MMIC has a typical gain of 18 dB with a noise figure of 3.0 dB. The critical operating characteristic is the 3.5-V bias voltage (measured at the MMIC output terminal). A 10-V V_{cc}, together with 430-Ω series resistors, sets the proper bias to allow the MMIC to operate near its nominal 16-mA current specification. Although chip coupling capacitors are recommended, I use standard disc ceramics with little performance compromise, and for someone with poor eyesight like myself they are much easier to use. Two capacitors in parallel reduce possible compromising inductance. The two MMIC ground leads are raised above the board surface using strips of 1/16-inch thick brass to make them level with the input and output leads, which connect to the glue-down strip lines.

The SBL 1 mixer is mounted in the conventional manner, except I use a drill to slightly ream each of the eight pin holes to avoid pin contact with the foil. This allows proper connection to the glue-down striplines. Grounded pins are soldered to the PC board with a small piece of soldering braid over the pin. This permits a relatively easy desoldering procedure if for some reason it becomes necessary to remove the mixer. The LO level into the mixer is monitored by a diode peak-reading detector. The nominal level is 5 mW, or about 0.8 V at TP3.

The LO driver filter is like the input filter except the input stripline is made shorter and used with a larger value capacitor. This optimizes loading of Q4, the MRF 901 tripler. The LO driver has a maximum output of 20 mW. The output is reduced to the

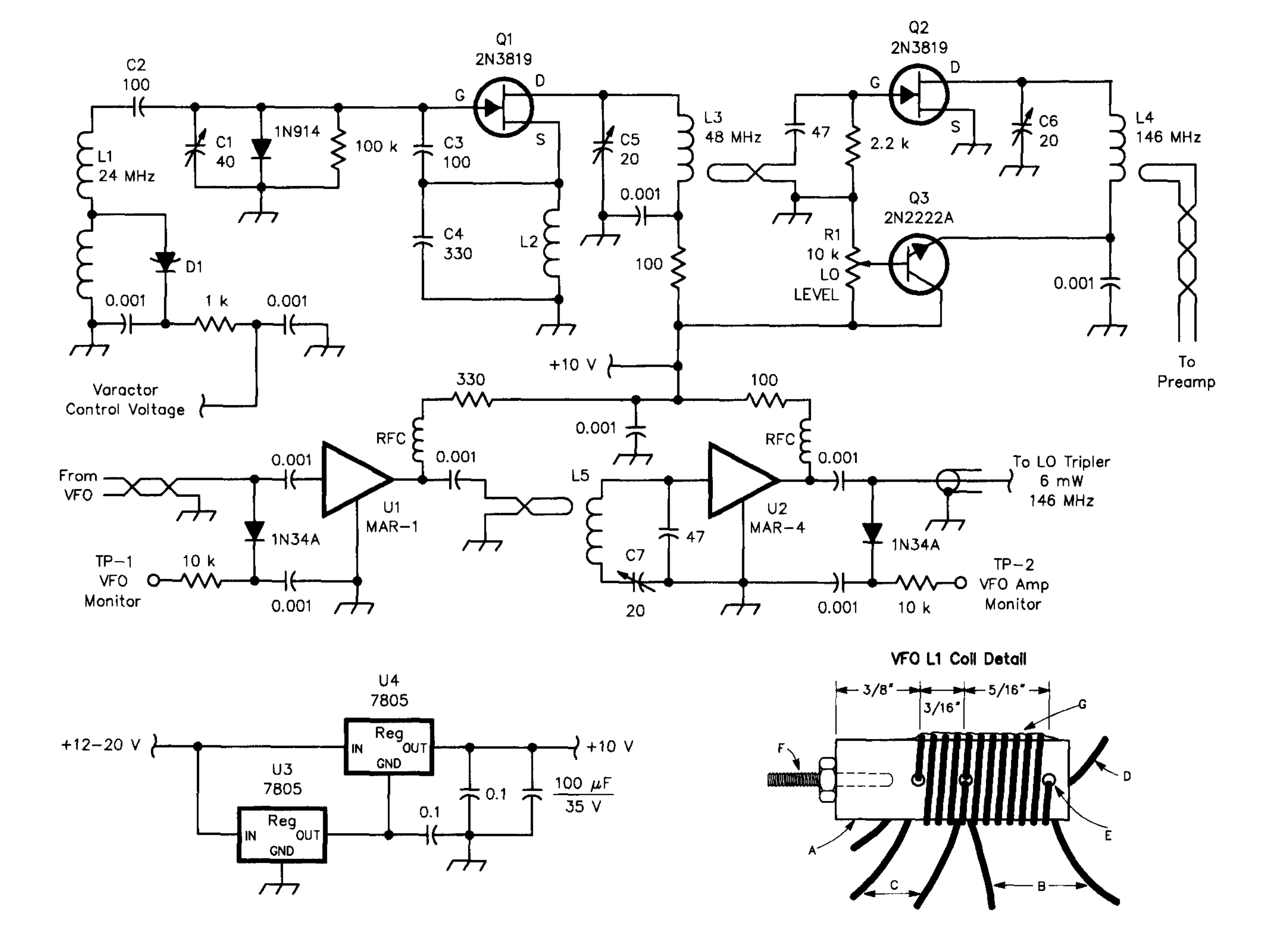

Fig 5—Schematic of the VFO circuit board.

C1—40-pF FILMTRIM (Sprague-Goodman #GYC40000)
C2—100-pF silver mica
C3—100-pF polypropylene (Panasonic #ECQ-P1H101JZ)
C4—330-pF silver mica.
D1—1N4742, 12-V Zener diode (used as a varactor).
L2—10 turns #26 wire wound on a 1/4-inch dia form (wood dowel).
L3—10 turns #26 wire wound on a 1/4-inch dia form with a 2-turn link coupling coil made from #30 wire-wrapping wire. The twisted pair is about 2 inches long.
L4—5 turns #18 wire, 1/4-inch ID, 5/16-inch long with a 1-turn link coupling coil made from #22 hook-up wire. The twisted pair is about 1-inch long.
L5—7 turns #18 wire, 1/4-inch ID, 3/8-inch long with a 1-turn link coupling coil made from #22 hook-up wire. The twisted pair is about 1.5-inches long.
RFC—20 turns #26 wire, 1/16-inch ID.
Output coupling to the tripler—About 1-foot of miniature microphone cable (RS 278-510).

A—3/8-inch dia Plexiglas rod.
B—7 turns coil #26 gauge wire.
C—4 turns coil #26 gauge wire.
D—0.010-inch dia carpet thread wound with the coils for uniform turn spacing.
E—3 1/16-inch dia holes to hold the coil.
F—4-40 coil mounting stud. It is cemented or threaded into the Plexiglas form.
G—Place several lines of clear cement along the coil length. (Elmer's Clear Household Cement).

desired level by the drive control, R1, located on the VFO circuit board.

The IF preamplifier is similar to the one decribed in Campbell's *QST* article. The grounded-base stage, Q5, provides a 50-Ω load to the mixer and approximately 40 dB of gain. It is followed by an emitter follower to supply a low output impedance.

VFO Circuit Board

The VFO, Q1, is a 24-MHz JFET Colpitts oscillator. It is tuned by a 12-V Zener diode connected to operate like a varactor. It produces approximately a 10-pF capacitance change that provides the desired 436-MHz LO shift of 4 MHz (222-kHz shift at the VFO). The output is taken from the FET drain with a 48-MHz tuned circuit. This doubles the frequency while providing reasonable isolation from the VFO.

As mentioned earlier, the one critical characteristic of this simple LO source is frequency stability. From past experience I found that polystyrene capacitors together with a coil wound on a thin-wall ceramic form provide excellent stability. The small positive temperature coefficient of the ceramic coil-form inductance is cancelled by the similar negative coefficient of the capacitors. It can be tuned with either a capacitor or a variable ferrite core. The problem is that both the coil form and the capacitors may be hard to find. For that reason, I investigated the use of alternative parts. Specifically, a wood dowel or Plexiglas coil form, silver mica or polypropylene capacitors, and an air or plastic trimmer for the band-set adjustment. For the coil, the Plexiglas form was chosen because there was some evidence of minor long-term mechanical instability of the wood dowel assembly. A combination of mica and polypropylene capacitors proved to be a reasonable choice because of their opposite temperature coefficients. Both air and plastic trimmers performed well for the band-set function, although the plastic trimmer was a bit difficult to adjust. Each configuration was tested by turning it on at 8:00 AM and periodically monitoring the results as the day progressed. Typically, the shop temperature would start at about 70° (F) and rise to 80° or more during the afternoon. The results, shown in Fig 6, indicate relative component temperature coefficients as applied in the circuit. Fig 6 also describes two acceptable component combinations. The dominant temperature-sensitive components are the inductor (L1) and the tuned circuit capacitors (C2,3,4). The band-set trimmers (both the capacitors and the variable ferrite), and the varactor are less sensitive. The two recommended combinations have an initial turn-on frequency shift of about 10 kHz during the first 15 minutes of operation. After that the shift is less than 10 kHz. An uncompensated configuration using all mica capacitors together with a Plexiglas coil form has a frequency shift of about 15 kHz per degree change of ambient temperature.

A resistive divider network, shown in Fig 7, provides the varactor tuning voltage and an input for the AFC or scanner function. A 12-position switch selects voltage divider resistors having values that compensate for the varactor nonlinearity, resulting in a 300-kHz frequency shift for each position. The voltage for the varactor will vary from 3.7 to 6.6 V or 3.4 to 6.0 V depending upon the position of the fine-tuning potentiometer. The

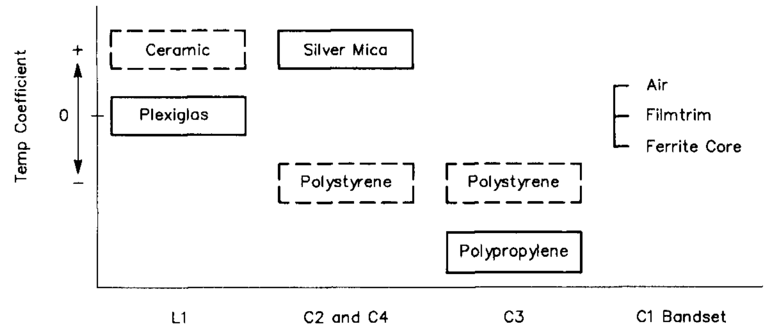

Fig 6—Relative temperature coefficients of the VFO components as observed during circuit operation. The solid lines indicate a configuration using a solid Plexiglas coil form and the dotted lines a thin-wall ceramic form. Both configurations will operate continuously for six hours together with a 10° F ambient temperature change with less than a 20-kHz shift in frequency. There is little performance difference between the three types of bandset methods as a function of temperature change.

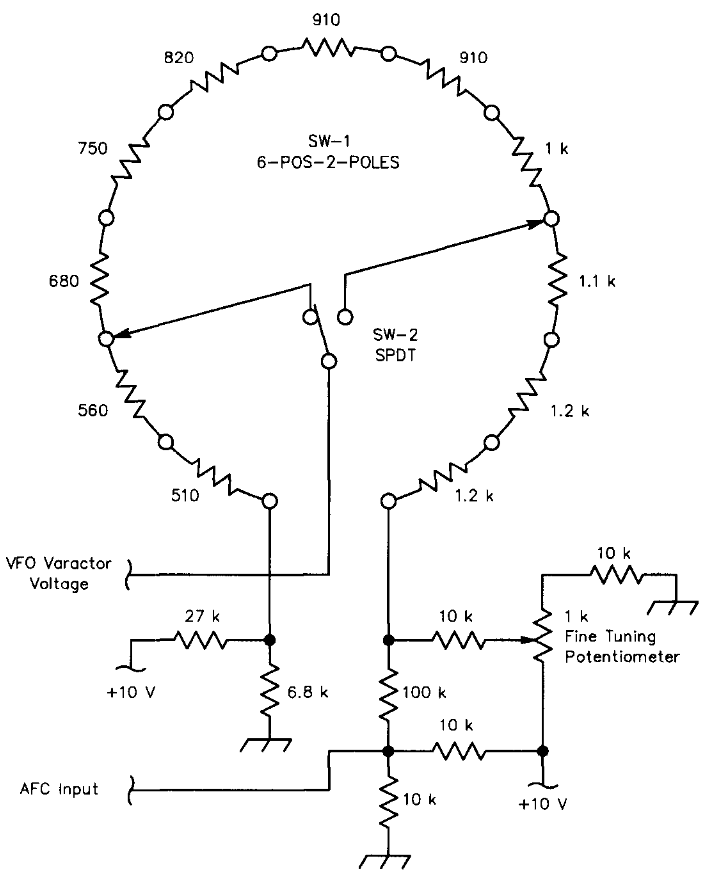

Fig 7—Schematic of the varactor voltage controller. There is a 300-kHz frequency shift between each SW-1 position. The fine tuning frequency shift is 500 kHz. The total tuning range is 4 MHz. The AFC input can shift the frequency up to ± 200 kHz. It is biased to +5 V by a 5-kΩ resistive divider. Maximum shift occurs when it is forced to 0 or +10 V

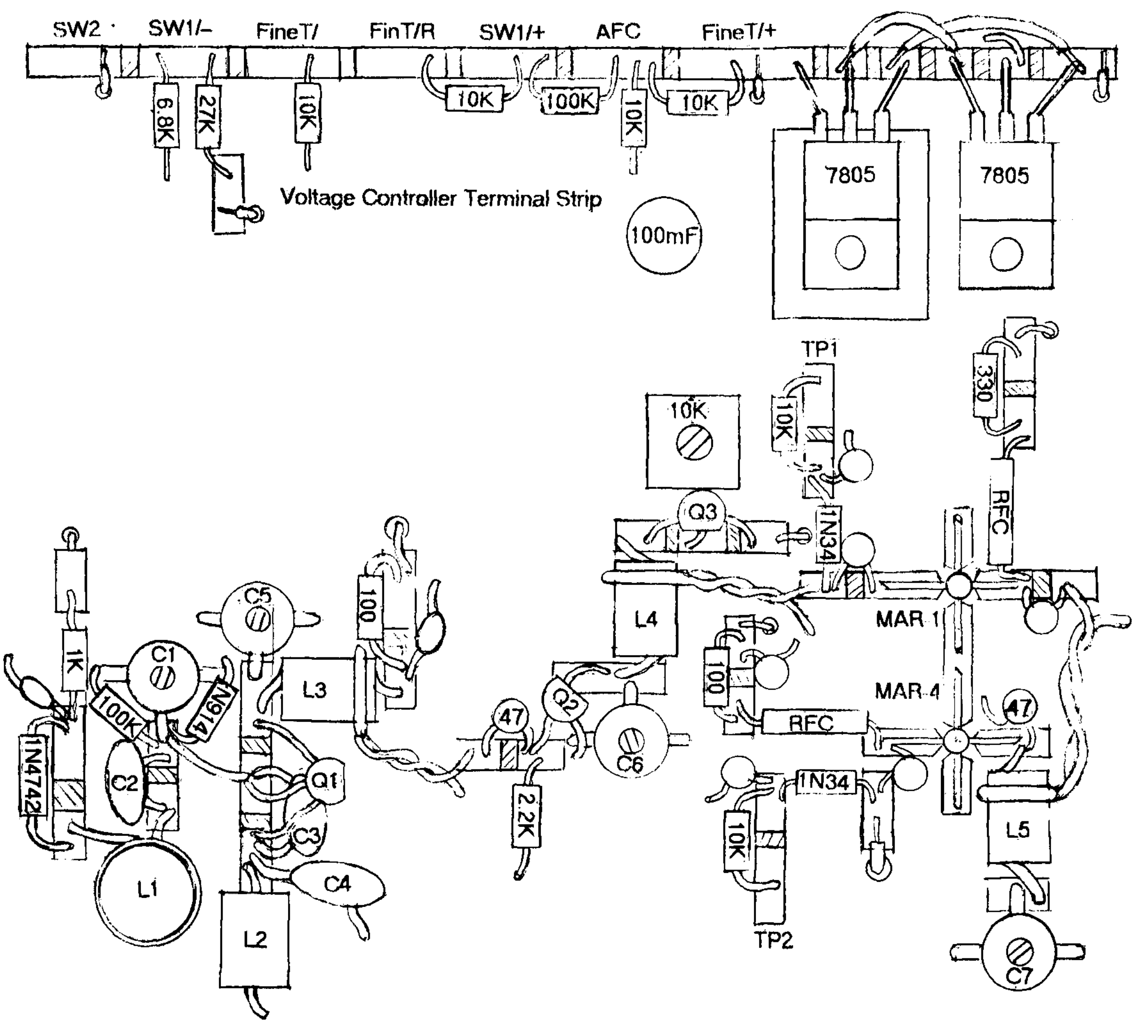

Fig 8—Layout of the VFO circuit board.

AFC/scanner input is nominally biased to 5 V. The frequency can be shifted up to 200 kHz either plus or minus by forcing the bias to 0 or 10 V. The bias resistance is 5 k.

Two 10-turn potentiometers, one for bandset and the other for fine tuning, are an alternative method. Frequency calibration is performed by simply monitoring the varactor input voltage with a meter. I used this method for about six months and it worked very well. It didn't compensate for the nonlinear varactor, but the tuning was fine enough to make that unimportant.

The VFO is followed by an FET tripler and a two-stage MMIC amplifier. The 145-MHz tripler has a maximum output of about 0.5 mW. This level is controlled by varying the FET drain voltage with the potentiometer R1, and monitored by a diode peak detector (TP1). The first MMIC MAR1 stage has a gain of about 18 dB and an output capability of +7 dBm in the VHF region. Its nominal operating current is 17 mA with a bias of 5 V. Although this output is probably enough to drive the LO tripler, the MAR4 was added as a safety factor to allow for circuit performance variations. It has a gain of about 8 dB with an output of +13 dBm. Its nominal operating current is 50 mA with a bias voltage of 6 V. The bias resistors, 300 and 100 Ω, together with the 10-V source, operate the MMICs close to their nominal values. There is a relatively high-Q tuned circuit used in the interstage coupling between the two MMICs for filtering out unwanted VFO responses. There are three tuned circuits for this purpose, L4/C6, L5/C7 and L6/C8, to ensure a reasonably pure waveform for driving the LO tripler. Output of the VFO amplifier is monitored by a peak voltmeter at TP2. It is also used to initially align L6/C8. At resonance, TP2 will read a minimum value due to lowering of the load impedance. About one foot of miniature microphone cable is used to connect the amplifier RF output to the UHF circuit board. There is a bit of loss in using this cable, but there is power to spare and the cable flexibility is an important consideration.

Alignment

The receiver can be aligned using nothing more than a multimeter and a 2-m transceiver. The 2-m receiver per-

Table 1—Procedures for Final Alignment

1. Adjust all capacitors to minimum capacity except C1, C5 and C8. Set these three at maximum capacity. Set R1 for maximum VFO output.
2. Turn the tuner controls for midband response. SW1: pos 6, SW2: low pos. Fine tuning potentiometer: midway point.
3. Set the 2-m transceiver to 145.33 MHz and arrange conditions such that a rubber ducky, or some other pick-up device, can be placed near L1.
4. Tune the VFO to 24 MHz by decreasing C1 until the VFO is heard on the 2-m receiver.
5. Tune L3/C5 to 48 MHz by decreasing C5 while peaking the 2-m receiver response.
6. Peak the 146-MHz multiplier response by decreasing C6 while monitoring TP1; it should read about 0.3 V.
7. Peak the VFO amp response by increasing C7 while monitoring TP2; it should read greater than 1 V.
8. Peak L6/C8 to 146 MHz by increasing C8 while monitoring TP2; it will null to about 0.8 V.
9. Peak the LO filter by decreasing C9 while monitoring TP3, then trim the remaining two 5-pF capacitors. TP3 should read about 1.2 V.
10. Correct the LO mixer input by adjusting R1 on the VFO circuit board for 0.8 V at TP3.
11. Normalize the IF preamp operation by adjusting R2 to make the dc voltage at TP4 2 V.
12. Peak the three input filter 5-pF capacitors for maximum response to the 2-m transmitter's third harmonic.
13. Optimize the S/N by tuning to a marginal input signal that is about +10 dB S/N (transmitter harmonic, noise, etc) and peaking the three filter 5-pF capacitors. S/N alignment requires the receiver input be terminated with a 50-Ω load.

Parts Suppliers

MMICs—Down East Microwave, RR1 Box 2310, Troy, ME 04987, tel 207 948-3741.
SBL1—Oak Hills Research, 20879 Madison St, Big Rapids, MI 49307, tel 616 796-0920.
Panasonic P-Series Polypropylene capacitors and Sprague-Goodman FILMTRIMs—Digi-Key Corp, 701 Brooks Ave S, PO Box 67, Thief River Falls, MN 56702-0677, tel 800 862-5432.
Silver mica capacitors and ceramic trimmers—All Electronics, PO Box 567, Van Nuys, CA 91408-0567, tel 800 862-5432.
Plexiglas—I found two local sources in the Yellow Pages under Plastics.
Brass—Hobby shops usually stock small sheets of brass in various thicknesses.

mits initial frequency set of the VFO, and the third harmonic of the 2-m transmitter provides a signal for initial test and alignment of the input filter and RF amplifier. The required alignment steps are detailed in Table 1.

Performance of the compact stripline filter is far from outstanding. However, it does allow me to monitor PACSATs while transmitting into the adjacent 2-m uplink antenna without interfering with the received signal (for the PACSAT full-duplex mode). Actually my only interference is from the terrible radar signals on 70 cm, and in this situation I find that an additional external high-performance filter results in little or no difference in damage to the processed data. As a result, my conclusion is that a more complex internal filter would not substantially improve the receiver's performance.

The receiver has been used to copy FM 9600-baud packet from UO22 and KO23, and 1200-baud PSK from AO16. It is a simple assembly that is a pleasure to use. □□

A CVSD Codec System for Your PC

Get started using digital voice with this simple PC add-on card.

by Jon Bloom, KE3Z

One of the challenges that Amateur Radio faces in the coming years is how to adapt to the digital communications revolution. While amateurs have embraced digital data communications, largely through packet radio, there still is little use by amateurs of digital voice communications. We seem to be stuck in the analog world of FM repeaters and SSB weak-signal stations. This can't last if amateur communications technology is to progress apace with changes in commonly available commercial technology.

Of course, there's not one "right way" to delve into digital voice communications. The major commercial digital voice service—the upcoming generation of cellular telephone systems—is in the midst of a shakeout between radically different digital voice technologies. But it's time—past time—for amateurs to get their feet wet in the development and deployment of digital voice technology. In that spirit, here's a circuit you can use to generate one common type of digital voice encoding for experimentation...or for use on the air.

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Background

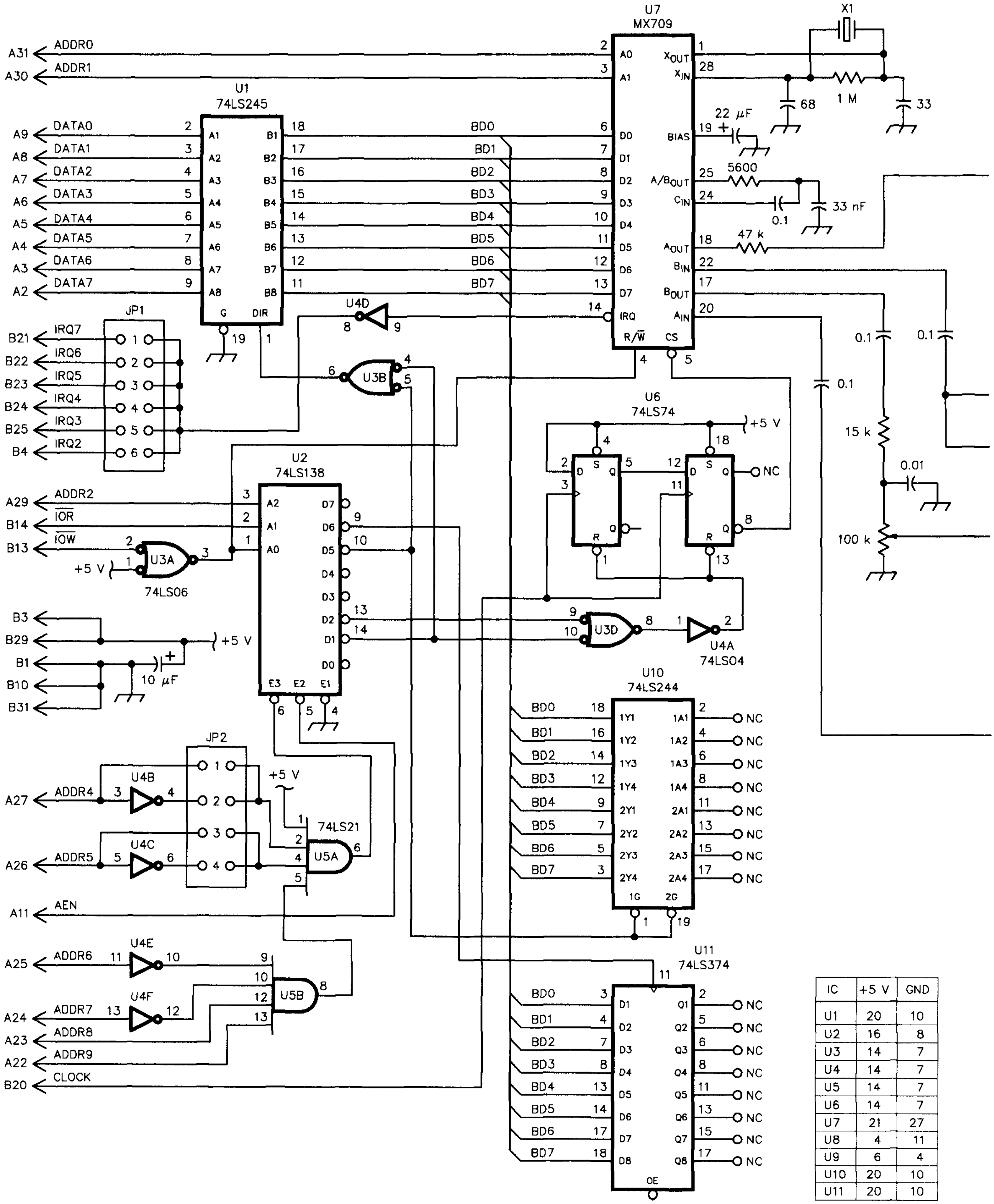
The classic way of generating a digital bit stream from an analog signal is direct digitization, or pulse-code modulation (PCM). A PCM system generates a fixed-size binary word that represents the amplitude of the input signal at the instant it was sampled. Successive samples, taken at a regular rate, give successive amplitudes of the signal. If you transmit these digital words through, say, a packet link, then feed them into a D/A converter, at the same rate at which they were sampled, you get a reproduction of the original signal. (Subject to a couple of niggling details, that is, like ensuring that the sample rate is more than twice the highest frequency present in the input signal.)

PCM works well, and is used extensively. For example, your PC's sound card works this way. But when the object is to send the resulting digital data over a radio link, PCM has one major failing: it can be severely degraded by any bit errors that occur in the system—something that is quite likely when transmitting data via radio. This can, of course, be overcome by use of various error-correcting techniques. But these add overhead, in the form of additional bits that need to be transmitted. That adds to the required bit rate, which adds to the required signal bandwidth—something to be avoided in these days of ever-increasing pressure on the spectrum.

Because of PCM's high bit rates, a number of techniques have been developed over the years to reduce the required number of bits per second, while still achieving acceptable quality when transmitting speech. One such technique is called *continuously variable slope delta modulation*, or CVSD.[1] CVSD is a particularly good choice for radio links, due to its tolerance of bit errors. (See the sidebar: "How CVSD Works.")

A CVSD codec (coder/decoder), the device that produces a bit stream from an analog signal, and vice versa, is a moderately complex animal. Fortunately, as with most useful complex circuits, someone has taken the

[1] Rabiner, L. R. and Schafer, R. W., *Digital Processing of Speech Signals*, Prentice-Hall, Englewood Cliffs, NJ (1978).



IC	+5 V	GND
U1	20	10
U2	16	8
U3	14	7
U4	14	7
U5	14	7
U6	14	7
U7	21	27
U8	4	11
U9	6	4
U10	20	10
U11	20	10

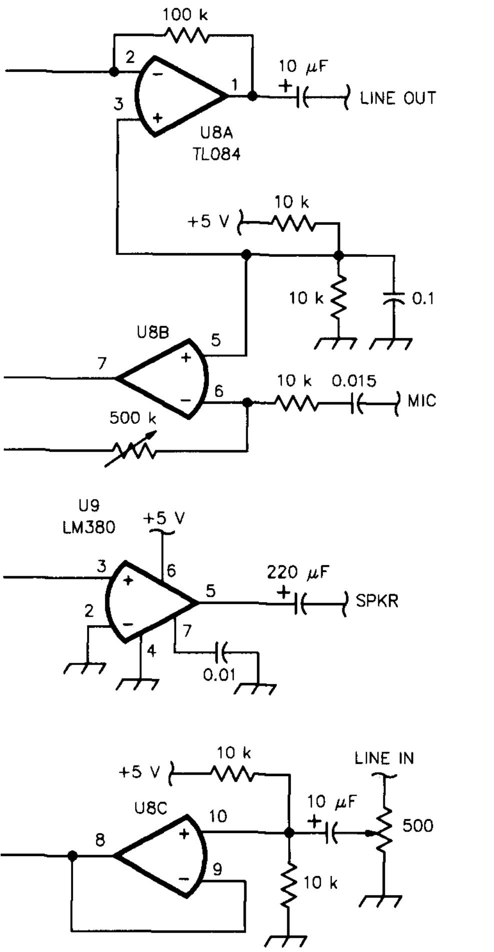


Fig 1—Schematic diagram of the CVSD codec board.
U1—74LS245 bus transceiver IC.
U2—74LS138 3-line to 8-line decoder IC.
U3—74LS08 quad 2-input NAND gate IC.
U4—74LS04 hex inverter IC.
U5—74LS21 dual 4-input AND gate.
U6—74LS74 dual D flip-flop IC.
U7—MX-COM MX709J CVSD codec IC.
U8—TL084 quad op-amp IC.
U9—LM386 audio amplifier IC.
U10—74LS244 octal bus driver IC.
U11—74LS374 octal D latch IC.
X1—2-MHz crystal.
Not shown are 0.1-µF capacitors from each IC's V_{cc} pin to ground.

trouble to develop an IC that contains all of the required functions. In this case, that someone is MX-COM, Inc. The device in question is the MX709 Voice Store Retrieve CVSD codec. (The name gives some indication of what MX-COM thinks is its main application.) One of the nice things about MX-COM devices is that you can get them; you can order small quantities—one, even—directly from MX-COM and pay for it with plastic money (Visa or MasterCard). Try *that* with the big boys!

The CVSD Circuit

For experimental purposes, one useful implementation of a CVSD system is as an add-on to a PC. The circuit presented here is an 8-bit PC plug-in card that contains an MX709 and the circuitry necessary to interface it to the PC, on the one hand, and to audio signals, on the other. The MX709 supports two channels of input and output, switching between the channels under command of the controlling microprocessor (the PC, in this case). On this card, I've implemented one of these channels as a channel with a low-level (microphone) input and a high-level (speaker) output. The other channel uses line-level (about 0.5-V pk-pk) input and output signals.

The interface to the PC is via I/O ports, with an interrupt used to signal the computer when the codec is ready for more data, either input or output. The MX709 contains two instruction registers that control operation of the device, a status register, input and output data registers, and a power-measurement register, about which more later. Each of these registers is mapped to a PC I/O-port address by the circuitry on the board. In total, the board occupies 16 I/O addresses, although only four addresses actually contain registers. (The disparity exists to keep the amount of decoding circuitry reasonable.) I also included 8-bit parallel (TTL) input and output ports, on the assumption that any useful experimentation will need to do more than just encode and decode bit streams! You can leave these off if you don't need them.

I built the circuit on a PC plug-in wire-wrap board. The circuitry is fairly simple, so construction isn't difficult. No special layout considerations are needed, except possibly to keep the analog circuitry near the back edge of the card—the one next to the back wall of the computer—for convenience sake.

The Circuit

The design, shown in Fig 1, is straightforward. U2 decodes the I/O address. JP2 can be used to select the address range of 300H-30FH, 310H-31FH, 320H-32FH or 330H-33FH. The outputs of U2 provide enabling signals for the MX709 and parallel port, resulting in the address map shown in Table 1.

Table 1—CVSD Codec Card Registers

Address (Hex)	*R/W*	*Register*
3x0	R	MX709 status register
3x1	R	MX709 power register
3x2	R	MX709 encoder data
3x3	R	N/A
3x4	R	Parallel input port
3x0	W	MX709 instruction register A
3x1	W	MX709 instruction register B
3x2	W	MX709 decoder data
3x3	W	N/A
3x4	W	Parallel output port

'x' can be 0, 1, 2 or 3, depending on the jumpers used on JP2:

Jumpers	*x*
2,4	0
1,4	1
2,3	2
1,3	3

The MX709 requires a 150-ns data set-up time, meaning the data must be stable on the inputs for that period before the chip is enabled. U6 ensures this requirement is met by inserting a short delay between the time the card is selected and the time the MX709 is enabled.

U1 buffers the data between the PC bus and the card circuits, providing an internal data bus, signals BD0-BD7, that connect to the MX709 and the parallel port chips, U10 and U11.

The analog circuitry is equally straightforward, with each of the two inputs, A and B, fed from an op amp. Channel B's op amp, U8B, can be adjusted for a gain of up to 50, allowing use of microphone levels at that input, while channel A can give, at best, a gain of 1 via U8C. The output of channel B is fed to an LM386 audio power amp, U9, in order to drive a speaker. While the LM386 is not the world's best audio amp, its performance substantially exceeds that of the codec, so

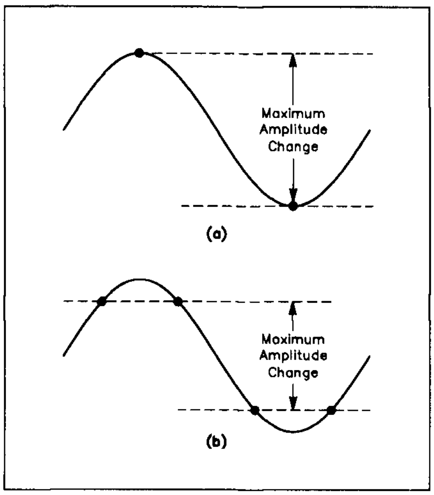
How CVSD Works

Considering that a 64-kbit/s CVSD bit stream gives worse signal-to-noise performance than a 64-kbit/s set of 8-bit PCM data, it's tempting to wonder why CVSD is worth bothering with. To understand why, it's best to begin by finding out what's wrong with using PCM.

In "straight" PCM, each binary word represents the amplitude of the input signal at the sample time. Say you are using an 8-bit system. You have 256 possible amplitude levels representable by 8 bits. The value of the least-significant bit represents the smallest possible change in amplitude, and the value of the most-significant bit is 128 times that. Suppose that when you send a particular byte, one of the bits is received incorrectly. If the bit that gets corrupted is the most-significant bit, the sample amplitude the receiver thinks was sent is wildly different from the amplitude that *was* sent. This creates a large error in the decoded audio signal, making it useless.

CVSD is a form of *delta modulation* (DM). This method of encoding a speech signal takes advantage of this fact: the faster you sample the signal, the less change in amplitude there is between samples. Fig A shows this effect. A signal sampled at just twice the input frequency (a) can change much more between samples than a signal sampled at four times the input frequency (b). Similarly, if you sample at much higher rates, the signal can only change a tiny amount between samples. The benefit of this is that now, instead of sending 8 bits to represent the sample, you only have to send as many bits as it takes to represent the maximum amplitude change possible between samples. That will be fewer than 8 bits. And if you sample fast enough, you can get it right down to *one* bit.

Fig A—At (a), the maximum amplitude change between samples of a signal sampled at twice the input frequency is equal to the peak-to-peak amplitude of the signal. For a signal sampled at four times the input frequency (b), the maximum change between samples is reduced.



One advantage of the one-bit conversion technique is that all of the bits have the same, very small, value. If a bit gets corrupted in transmission, it can only cause a slight error in the received signal, adding slightly to the received noise, but not corrupting the signal to the point of unusability—unless a lot of bits are corrupted.

The further refinement CVSD adds is *adaptive quantization.* Adaptive quantization, which can also be applied to non-DM schemes, such as PCM itself, recognizes that a particular fixed LSB weight isn't always going to be the best choice. In the case of 8-bit PCM, consider what happens when you have a signal that swings over the full range of 256 values: the S/N ratio is about 50 dB. But when the input signal drops in amplitude, the S/N drops as well, since the noise level depends on the resolution of the LSB. What if you could detect the change in signal level and change the weighting of the LSB to make it less? You don't need the same full range of amplitudes you needed for the high-level signal, so you can use your 8 bits to quantize the lower-level signal by reducing the amplitude represented by a one-LSB change.

Of course, if you do that, you need some means of communicating the change to the receiving end. CVSD provides an elegant way of doing that. Recall that we are using a one-bit quantization change for CVSD. If the amplitude of the input signal is greater than that of the previous (quantized) level, we transmit a 0. If it's less, we send a 1. At the receiving end, a received 0 tells the decoder to increase the output voltage by the quantizing value, and a received 1 tells it to decrease the output voltage by the quantizing value. So, if the input signal at the transmitting (encoding) end is constant (0 volts, say) we end up sending 101010101... ad infinitum. As the signal begins to rise, we start sending a string of zeroes; if the signal decreases, we send ones. The way CVSD adapts its quantization value is by monitoring the number of consecutive ones (or zeroes). If a number (usually 3) of ones appear consecutively, the encoder or decoder increases the quantization value. That is, any time the current bit and the previous two bits are all the same value (0 or 1), the quantization value is increased. Once that is no longer the case, the encoder or decoder begins to decrease the quantizing value back to its original minimum value. This scheme causes the quantizing value to adapt to the amount of change in the input signal.

Doing that means that some of the transmitted bits now represent a greater amplitude change than others, so the signal has become a little more susceptible to bit errors. It's not nearly as susceptible as PCM, though, and the improvement in S/N ratio that comes with adaptive quantization is well worth the increase in susceptibility to errors.

For purposes of sending digital voice serially, as we typically would do via a radio link, instead of in parallel words, as a computer bus does, the one-bit quantization of CVSD has a significant advantage over PCM: there is no need to synchronize the receiver to the word boundaries of the data because there *are* no word boundaries; it's just a serial stream of bits. (Or, another way of looking at it: each word is one bit wide.) Although the codec used here passes data to the computer in 8-bit chunks, that's just for the convenience of the computer. The data is still a series of single bits.

it's adequate for this use. The channel A output is amplified with a gain of 2 in U8A.

There is a third analog signal input on the MX709: input C. This input is used strictly for power measurement. The idea is to provide a way of determining the average power in the input signal, which makes implementation of VOX easy. In the MX709, there is also independent measurement of the power of the currently selected input

(A or B). But the power measurement is not constant with frequency, so you will get different readings from a high-pitched voice than from a low-pitched voice at the same amplitude. In this circuit, I've connected the A/B output, which is a filtered copy of the selected input signal, to the C input through a small R-C filter that provides some frequency compensation for the power measurement. Thus you can obtain either the compensated measurement or the uncompensated measurement; they both appear in the power register.

The MX709 requires a clock, provided here by a crystal. You can use one of a variety of crystal frequencies. I've used 2 MHz, which gives available bit rates of 62.5, 50, 31.25 and 25 kbit/s.

Table 2—MX709 Instruction Register A

Bit (LSB=0)	*Name*	*State*	*Action*
0	Encoder Idle	0	Encoder encodes input signal
		1	Force encoder to 1010101...pattern
1	Decoder Source	0	Data from PC
		1	Encoder output (digital loopback)
2-4	Decode Clock Divider	010	62.5 kbit/s, f_{max}=3.32 kHz
		011	31.25 kbit/s, f_{max}=3.32 kHz
		110	50 kbit/s, f_{max}=2.66 kHz
		111	25 kbit/s, f_{max}=2.66 kHz
5-7	Encode Clock Divider		(same as bits 2-4)

Programming the MX709

Operation of the MX709 is controlled by two instruction registers, as shown in Tables 2 and 3. I've given the clock divider settings only for the 2-MHz clock I used; you'll have to refer to the MX-COM data sheet if you use a different clock frequency. Table 4 shows the bits of the MX709 status register, while Table 5 describes the power register.

The MX709 is easy to program. You only need to set the appropriate control bits and begin reading and/or writing the data. You can either use the status register bits to determine, in a polled mode, when the codec is ready for a data byte transfer, or you can use interrupts. The latter case simply requires that you enable the PC's interrupt line and begin by transferring the first data byte.

There is only one interrupt line, but there are three possible interrupting conditions: encode data ready, decode data ready, and page (power data) ready. If you fail to service any one of these conditions, *all* of the interrupts are disabled in the codec. Therefore, your interrupt service routine should check each of the three "ready" bits in the status register and service whichever data registers need service. When an overspill occurs, which indicates that the interrupts have been disabled, the corresponding ready bit is set to 0. Therefore, your interrupt service routine should also check the overspill bits; if an overspill bit is set, you need to service the corresponding data register to get interrupts started again.

Instruction register B contains bits that control the page size, which can be set to specific values between 32 and 256. The power measurement reported in the power register is the average power of the measured signal over the last page of data. Thus selecting a particular page size determines the averaging time constant of the power measurement and the rate at which new measurements are available. The measured power is a 4-bit number that represents the signal level at the codec input over a range of +6 dBm to –24 dBm. (Subtract 12 dB from these numbers if instruction register B, bit 7 is set to 0. Watch out for overrange of the power reading in this case; there is no indication it has occurred.)

Table 3—MX709 Instruction Register B

Bit (LSB=0)	*Name*	*State*	*Action*
0-2	Page Size	000	32 bytes
		001	64 bytes
		010	96 bytes
		011	128 bytes
		100	160 bytes
		101	192 bytes
		110	224 bytes
		111	256 bytes
3	A/B Encode	0	Encoder input is channel A
		1	Encoder input is channel B
4	A Output Select	0	Decoder output
		1	A input (analog loopback)
5	B Output Select	0	Decoder output
		1	B input (analog loopback)
6	Power Save	0	Circuit enabled
		1	Circuit disabled
7	Power Sensitivity	0	+12 dB
		1	0 dB

Software

I've written a set of C language functions to assist in programming the card. These functions provide an easy-to-use interface to the card. The source code for the functions, along with some test programs, is available for downloading from the ARRL BBS (203 666-0578) and via Internet from ftp.cs.buffalo.edu (in the /pub/ham-radio directory), in file QEXCVSD.ZIP.

Performance

The performance you achieve from this system depends largely on the encoding rate you select. At 64 kbit/s, the codec is capable of providing a 35-dB S/N ratio if the input signal uses the full available input range. The S/N drops to about 27 dB at 32 kbit/s, and to a noisy 17 dB at 16 kbit/s. These are hardly "broadcast quality," but when you consider that a 20-dB quieting signal is considered near-full-quieting on an FM repeater, it's clear that these numbers are within the range of acceptability for voice communications.

An advantage of CVSD over PCM, as explained in the sidebar, is that errors in the received bit stream have relatively little effect. This is important for digital voice, as it means that slow, overhead-laden protocols (such as AX.25) are not needed; a few bad bits are hardly noticeable. So you can avoid the overhead of ARQ or FEC, keeping the bit rate reasonable. (Well, perhaps a weak FEC would be worth looking into. *There's* a good subject for experimentation.)

Of course, you still have to find a way of transmitting and receiving the bits. One possibility is the WA4DSY 56-kbaud modem, described in the *6th ARRL Computer Networking Conference Proceedings*. This circuit can probably be pushed to 62.5 kbit/s, and can easily handle 50 kbit/s. So the radio hardware to support this kind of operation is out there.

Conclusion

I'd be interested in hearing from anyone who uses this circuit (or *any* circuit, for that matter) to implement digital voice over Amateur Radio. The time is now. □□

Feedback

I was going over my article, "The Multipurpose Morse Code Generator Circuit," February 1994 *QEX*, and noticed an error on the third schematic (the beacon schematic). The part number of the hex inverter should be 74HC14, not 74HC04 as shown.—*Nick Ciarallo, VE2HOT*

Table 4—MX709 Status Register

Bit (LSB=0)	*Name*	*State*	*Action*
0	Encode Data Ready	0	No data ready, or overspill
		1	Data byte ready
1	Decode Data Ready	0	Not ready for data, or overspill
		1	Ready for input data
2	Page Ready	0	Not ready, or overspill
		1	Page encoded, power data valid
3	Encode Overspill	0	Normal (after data read)
		1	Overspill, data lost
4	Decode Overspill	0	Normal (after data written)
		1	Overspill, idle pattern sent
5	Page Overspill	0	Normal
		1	Overspill, power register not read

Table 5—MX709 Power Register

Bit (LSB=0)	*Name*	
0-3	A/B Power	Average signal level at A or B input over last page of data
4-7	C Power	Average signal level at C input over last page of data

Phase-Shift Network Analysis and Optimization

Analyzing a phase-shift network shows how its performance varies with component tolerances.

by Kevin Schmidt, W9CF

Introduction

The phasing method of single-sideband generation or detection requires two signals with a 90° relative phase shift over the audio frequency range. The phasing method has never been very popular, particularly once relatively inexpensive filters became available. In the future, presumably, digital signal-processing techniques will perform the necessary audio phase shifting or directly generate the radio frequency single-sideband signal. Why then should you be interested in audio phase-shift networks? Perhaps because they are relatively low cost, easy to build and are fun to play with. In addition, the techniques that I describe here are useful for efficient analysis of other cascaded networks.

For many years, the *ARRL Handbook* has included a circuit for an audio phase-shift network designed by HA5WH.[1] I have not located the original reference for this network. The *Handbook* claims that the circuit gives approximately 60 dB of opposite sideband suppression using 10% tolerance components. This flies in the face of the usual result that you need 1% components to get around 40-dB suppression. In this article, I will analyze and give design equations for this type of network. Unfortunately, this analysis shows that using 10% tolerance components *can* lead to poor sideband suppression. With ideal components the network can give excellent performance, and by using either high-tolerance components or well-matched lower-tolerance components, the network still can give good performance.

Notes appear on page 23.

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In the following sections I give the general formula for the sideband suppression in terms of the phase and amplitude errors in the phasing network; derive an efficient method of analyzing a general network of the HA5WH type; give the analysis of an ideal realization of the network; describe the optimization of the network in terms of easily calculated elliptic functions; and, give the effects of component tolerances. The result is a set of simple design equations for the ideal network and an estimate of the sensitivity to component tolerances. A set of FORTRAN programs that implement the methods described are available for downloading.

The Effects of Phasing Errors on Sideband Suppression

The phasing method generates a single-sideband signal, given mathematically as $\cos((\omega_c \pm \omega_a)t)$, where the + (or −) sign gives the upper (or lower) sideband, and $\omega_c = 2\pi f_c$ where f_c is the carrier frequency. Similarly, $\omega_a = 2\pi f_a$ where f_a is the audio modulating frequency. The cosine can be written as

$$\cos\left((\omega_c \pm \omega_a)t\right) = \cos\left(\omega_c t\right)\cos\left(\omega_a t\right) \mp \sin(\omega_c t)\sin(\omega_a t) \qquad \text{Eq 1}$$

the basic equation of the phasing method. The multiplications on the right-hand side are accomplished using balanced modulators, and the two audio frequencies (as well as the two radio frequencies) must be 90° out of phase and

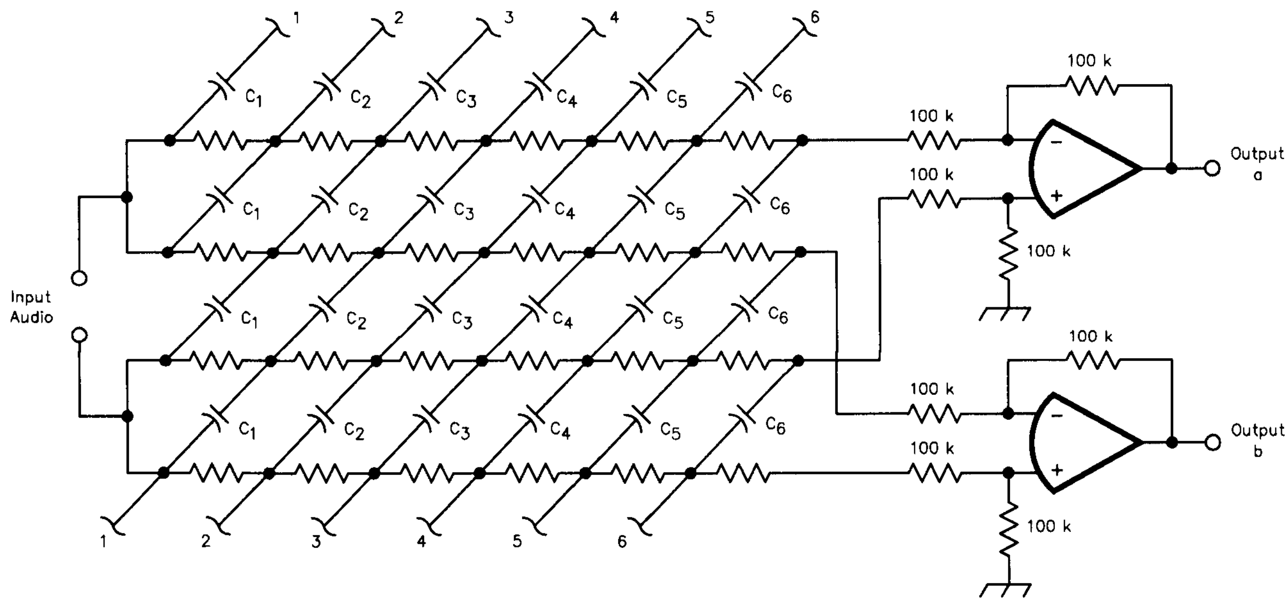

Fig 1—The schematic diagram of the HA5WH wide-band phase-shift network.

of equal amplitude. I will assume that the radio frequencies are exactly 90° out of phase, and of equal amplitude. Using the usual complex notation with $V_A e^{j\omega_a t}$ to be one audio signal, and $V_B e^{j\omega_a t}$ to be the other, the result of using a nonideal phasing network will be

$$\mathrm{Re}\left[\cos(\omega_c t)V_A e^{j\omega_a t} + \sin(\omega_c t)V_B e^{j\omega_a t}\right] =$$
$$\frac{1}{2}\mathrm{Re}\left[e^{j(\omega_c+\omega_a)t}(V_A - jV_B) + e^{-j(\omega_c-\omega_a)t}(V_A + jV_B)\right] \qquad \text{Eq 2}$$

and the sideband suppression (or enhancement) is given by

$$20\log_{10}\left|\frac{V_A + jV_B}{V_A - jV_B}\right| \qquad \text{Eq 3}$$

Notice if $|V_A|$ equals $|V_B|$, that is if the two signals have equal amplitude then for a phase error of δ, the suppression in dB is simply,

$$-20\log_{10}\left|\tan\left(\frac{\delta}{2}\right)\right| \qquad \text{Eq 4}$$

Analyzing the HA5WH Network

Fig 1 gives the circuit diagram of the HA5WH network as shown in the *ARRL Handbook*. Given this circuit, it is easy to analyze the network numerically using a mesh or nodal analysis. The disadvantage of this brute force approach is that it gives no insight into why the network works, or how changes in the network affect its performance. I will therefore describe a method that is both more efficient numerically, and, by using the symmetry of the ideal network, leads to simple design equations.

The network consists of six sections each with four input connections and four output connections. One of these sections is shown in Fig 2. I have labeled the input voltages and currents V_1, V_2, V_3, V_4, I_1, I_2, I_3, I_4. The corresponding output voltages and currents are labeled V'_1, V'_2, V'_3, V'_4, I'_1, I'_2, I'_3, I'_4. A straightforward nodal analysis of this network gives the eight linear equations represented by the matrix equation

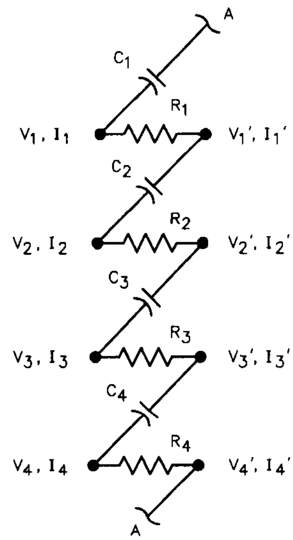

Fig 2—The schematic diagram of one section of an HA5WH network.

$$\begin{pmatrix} I \\ I' \end{pmatrix} = \begin{pmatrix} M_{11} & M_{12} \\ M_{21} & M_{22} \end{pmatrix}\begin{pmatrix} V \\ V' \end{pmatrix} \qquad \text{Eq 5}$$

where V, V', I, I' are length 4 vectors, and the M_{ij} are 4-by-4 matrices. Eq 5 compactly represents the eight equations that are the requirements of current conservation at each of the nodes of the network section. The M_{ij} matrices are

$$
M_{11} = \begin{pmatrix} \frac{1}{R_1} + j\omega C_1 & 0 & 0 & 0 \\ 0 & \frac{1}{R_2} + j\omega C_2 & 0 & 0 \\ 0 & 0 & \frac{1}{R_3} + j\omega C_3 & 0 \\ 0 & 0 & 0 & \frac{1}{R_4} + j\omega C_4 \end{pmatrix}
$$

$$
M_{12} + \begin{pmatrix} -\frac{1}{R_1} & 0 & 0 & -j\omega C_1 \\ -j\omega C_2 & -\frac{1}{R_2} & 0 & 0 \\ 0 & -j\omega C_3 & -\frac{1}{R_3} & 0 \\ 0 & 0 & -j\omega C_4 & -\frac{1}{R_4} \end{pmatrix}
$$

$$
M_{21} = \begin{pmatrix} \frac{1}{R_1} & j\omega C_2 & 0 & 0 \\ 0 & \frac{1}{R_2} & j\omega C_3 & 0 \\ 0 & 0 & \frac{1}{R_3} & j\omega C_4 \\ j\omega C_1 & 0 & 0 & \frac{1}{R_4} \end{pmatrix}
$$

$$
M_{22} = \begin{pmatrix} -\frac{1}{R_1} - j\omega C_2 & 0 & 0 & 0 \\ 0 & -\frac{1}{R_2} - j\omega C_3 & 0 & 0 \\ 0 & 0 & -\frac{1}{R_3} - j\omega C_4 & 0 \\ 0 & 0 & 0 & -\frac{1}{R_4} - j\omega C_1 \end{pmatrix}
$$

Eq 6

In exact analogy with cascading two-port networks using *ABCD* matrices, to cascade these network sections I define a new matrix equation,

$$
\begin{pmatrix} V' \\ I' \end{pmatrix} = \begin{pmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{pmatrix} \begin{pmatrix} V \\ I \end{pmatrix} \qquad \text{Eq 7}
$$

Solving for the A_{ij} matrices gives,

$$
\begin{aligned} A_{11} &= -M_{12}{}^{-1} M_{11} \\ A_{12} &= M_{12}{}^{-1} \\ A_{21} &= M_{21} - M_{22} M_{12}{}^{-1} M_{11} \\ A_{22} &= M_{22} M_{12}{}^{-1} \end{aligned} \qquad \text{Eq 8}
$$

where $M_{12}{}^{-1}$ is the inverse of the matrix M_{12}.

Labeling the 8-by-8 matrices for each of the n sections of the network by $A^{(1)}$, $A^{(2)}$, ... $A^{(n)}$, the matrix relating the input to the output of the full network is $\tilde{A}$ made up of the four 4-by-4 matrices $\tilde{A}_{ij}$,

$$
\begin{pmatrix} V_{out} \\ I_{out} \end{pmatrix} = \begin{pmatrix} \tilde{A}_{11} & \tilde{A}_{12} \\ \tilde{A}_{21} & \tilde{A}_{22} \end{pmatrix} \begin{pmatrix} V_{in} \\ I_{in} \end{pmatrix} \qquad \text{Eq 9}
$$

where $\tilde{A}$ is the matrix product $A^{(1)}A^{(2)}A^{(3)} \ldots A^{(n)}$.

The *Handbook* circuit drives four resistors on the four output connections. Labeling these as $R_1{}^{(out)} R_2{}^{(out)} R_3{}^{(out)}$, $R_4{}^{(out)}$, and defining a 4-by-4 load matrix L,

$$
L = \begin{pmatrix} \frac{1}{R_1^{(out)}} & 0 & 0 & 0 \\ 0 & \frac{1}{R_2^{(out)}} & 0 & 0 \\ 0 & 0 & \frac{1}{R_3^{(out)}} & 0 \\ 0 & 0 & 0 & \frac{1}{R_4^{(out)}} \end{pmatrix} \qquad \text{Eq 10}
$$

I can write the relationship between the output voltage and current as,

$$
(I_{out}) = L(V_{out}) \qquad \text{Eq 11}
$$

Solving for I_{out} and back substituting gives the final network matrix equation relating the four output voltages to the four input voltages,

$$
(V_{out}) = \left(1 - \tilde{A}_{12}\tilde{A}_{22}^{-1}L\right)^{-1}\left(\tilde{A}_{11} - \tilde{A}_{12}\tilde{A}_{22}^{-1}\tilde{A}_{21}\right)(V_{in}) \qquad \text{Eq 12}
$$

where 1 in Eq 12 stands for the unit matrix

$$
\begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{pmatrix} \qquad \text{Eq 13}
$$

If the output resistors are large compared to the other circuit impedances, L can be taken to be zero. In that case the equations simplify to,

$$
(V_{out}) = \left(\tilde{A}_{11} - \tilde{A}_{12}\tilde{A}_{22}^{-1}\tilde{A}_{21}\right)(V_{in}) \qquad \text{Eq 14}
$$

The *Handbook* network has (V_{in}) proportional to

$$
(V_{in}) \propto \begin{pmatrix} 1 \\ 1 \\ -1 \\ -1 \end{pmatrix} \qquad \text{Eq 15}
$$

After calculating the $\tilde{A}$ and L matrices from the circuit values, the output signals need to be combined as,

$$
\begin{aligned} V_{out,1} - V_{out,3} &= V_A \\ V_{out,2} - V_{out,4} &= V_B \end{aligned} \qquad \text{Eq 16}
$$

and the sideband suppression is given by Eq 3. The relative amplitude and phase of the signals can also be calculated. Most phase-shift networks are based on all-pass networks so that the amplitude of all signals is equally attenuated. The HA5WH network is not an all-pass network. Ideally, we want both good sideband suppression and we want V_A and V_B to be constant in amplitude and phase across the passband of the audio circuit.

I have written a FORTRAN program to implement the analysis of this section. It is available for download from the ARRL BBS (203 666-1578) and via Internet FTP from ftp.cs.buffalo.edu, in the \pub\ham-radio directory. The file name is pshift.zip. If the matrices that are inverted become singular, the above analysis breaks down at the singular points. For example, M_{12} becomes singular when

$$\omega^4 = \frac{1}{R_1 R_2 R_3 R_4 C_1 C_2 C_3 C_4} \qquad \text{Eq 17}$$

Near these points, roundoff error in the calculations will be large. For the analysis done here, this is not a big problem. However, analysis on networks with many sections or near singular points will require more numerical care than I have taken in the FORTRAN program, or the use of a standard formulation where the full set of network equations are solved at once.

Analysis of the Ideal Cyclic Network

The design of the HA5WH network, as shown in the *Handbook*, has four identical resistors and four identical capacitors in each of the six network sections. This means the network is invariant under a cyclic interchange of the ordering of its ports. That is, if we were to relabel the ports by letting 1 become 2, 2 become 3, 3 become 4, and 4 become 1, we would obtain exactly the same equations describing the network. Such invariances are treated generally using the mathematics of group theory, which greatly simplifies the study of the system with symmetries.[2] The ideal HA5WH network has what is known as cyclic 4 or C_4 symmetry. The network equations can be analyzed using group theory. Analysis of the character of the matrix that represents the cyclic operator shows that each of the four irreducible representation of C_4 appears once. These therefore correspond to the four eigenvectors of the $\tilde{A}$ matrices, which can then be written down immediately.

Most hams probably are unfamiliar with group theory, however, the results can be easily verified without using group theory. The right eigenvectors $\psi^{(m)}$ and the eigenvalues λ_m of a matrix M are defined by finding the solutions to the equations,

$$M\psi^{(m)} = \lambda_m \psi^{(m)} \qquad \text{Eq 18}$$

That is, multiplying the eigenvector by the matrix gives the same eigenvector back as the result, simply multiplied by the eigenvalue. The effect of multiplying a matrix times one of its eigenvectors is to simply multiply the eigenvector by the eigenvalue.

The cyclic eigenvectors in our basis are those that change by a constant phase between the elements, with the same phase change between the last and first elements. This gives,

$$\begin{pmatrix}1\\1\\1\\1\end{pmatrix}, \begin{pmatrix}1\\-1\\1\\-1\end{pmatrix}, \begin{pmatrix}1\\j\\-1\\-j\end{pmatrix}, \begin{pmatrix}1\\-j\\-1\\j\end{pmatrix} \qquad \text{Eq 19}$$

By direct matrix multiplication, it is easily verified that these are the eigenvectors of *all* the M matrices if all the R and C values are the same in a network section. This is a direct consequence of the cyclic 4 symmetry. Further, since the $\tilde{A}$ matrices are combinations of products of the M matrices, these same vectors are the eigenvectors of the $\tilde{A}$ matrices. Since V_{out} is given as a combination of $\tilde{A}$ matrices times V_{in}, if we express V_{in} as a linear combination of the four eigenvectors, V_{out} will be given by taking this same linear combination and multiplying each term by an appropriate eigenvalue. The network must then be designed to produce a 90° relative phase shift.

The input to the HA5WH network contains only the last two eigenvectors written above. That is

$$V_{in} = \begin{pmatrix}1\\1\\-1\\-1\end{pmatrix} = \frac{1-j}{2}\begin{pmatrix}1\\j\\-1\\-j\end{pmatrix} + \frac{1+j}{2}\begin{pmatrix}1\\j\\-1\\-j\end{pmatrix} \equiv \frac{1-j}{2}\psi_a + \frac{1+j}{2}\psi_b \qquad \text{Eq 20}$$

where the last line defines the relevant eigenvectors as ψ_a and ψ_b. Further, the output is also not sensitive to the first two eigenvectors if the output impedances are identical and the operational amplifiers have good common-mode rejection. Having both of the conditions will be helpful if the cyclic symmetry is broken because of component tolerances.

With the input as in Eq 20, the output will in general be

$$V_{out} = \frac{1-j}{2} g_a \psi_a + \frac{1+j}{2} g_b \psi_b \qquad \text{Eq 21}$$

and the two outputs to the balanced modulators will be

$$\begin{aligned} V_A &= (1-j)g_a + (1+j)g_b \\ V_B &= (1-j)jg_a - (1+j)jg_b \end{aligned} \qquad \text{Eq 22}$$

the suppression in dB is found using Eq 3,

$$20\log_{10}\left|\frac{g_a}{g_b}\right| \qquad \text{Eq 23}$$

So to design a good network, we must eliminate one of these final two eigenvectors.

The analysis so far shows how the HA5WH network can be motivated. The C_4 eigenvectors have equal amplitudes for the four voltages and have phase shifts between adjacent ports of 0°, +90°, 180° and 270°. This last is equivalent to a phase shift of –90°. We want to choose the network drive, connections and component values to select one of the two 90° phase-shifted eigenvectors. As an aside, the same ideas could be used to design a 60° relative phase shift by using a network invariant under the group C_6, or a 45° shift from C_8, etc.

The first step in selecting the component values is to calculate the eigenvalues of the four M matrices. By direct multiplication, I get

$$\begin{aligned} \lambda_{11}^a &= \lambda_{11}^b = \frac{1}{R} + j\omega C \\ \lambda_{12}^a &= -\frac{1}{R} - \omega C, \lambda_{12}^b = -\frac{1}{R} + \omega C \\ \lambda_{21}^a &= \frac{1}{R} - \omega C, \lambda_{21}^b = \frac{1}{R} + \omega C \\ \lambda_{22}^a &= \lambda_{22}^b = -\frac{1}{R} - j\omega C \end{aligned} \qquad \text{Eq 24}$$

where the superscript a or b indicates the eigenvalue corresponds to the eigenvector ψ_a or ψ_b, respectively.

The effect of one of the A matrices, when a single eigenvector is input, is given by replacing the M matrices in

Eq 8 by their eigenvalues. After a little algebra, I get,

$$A^a = \frac{1}{1+\omega RC}\begin{pmatrix} 1+j\omega RC & -R \\ -2j\omega C & 1+j\omega RC \end{pmatrix}$$
$$A^b = \frac{1}{1-\omega RC}\begin{pmatrix} 1+j\omega RC & -R \\ -2j\omega C & 1+j\omega RC \end{pmatrix} \qquad \text{Eq 25}$$

The A^b matrix is proportional to A^a. If we feed the section of the network with a linear combination of ψ_a and ψ_b, the section suppresses ψ_a relative to ψ_b by a factor of

$$\frac{1-\omega RC}{1+\omega RC} \qquad \text{Eq 26}$$

The HA5WH network has the properties that the magnitude of the ratio given in Eq 26 is always less than one for positive frequencies and is exactly zero for $\omega=1/(RC)$. The first property says that additional network sections can only improve the relative 90° phase shift of the outputs. The second says that we can set the frequencies of exact 90° phase shift by selecting the R-C values of single network sections. These two properties greatly simplify the design and optimization of the network.

The sideband suppression at a single frequency is given for an n section network, with R-C values in section i given by R_i and C_i, as

$$\text{Suppression} = 20\sum_{i=1}^{n}\log_{10}\left|\frac{1-\omega R_iC_i}{1+\omega R_iC_i}\right| \qquad \text{Eq 27}$$

A simple method of picking the R-C values for each section is to use a computer to plot the above result, and adjust n and R_iC_i to achieve the required suppression. This is, in fact, the obvious technique to use if you are trying to design with a set of parts already in your junk box. However, the form of the suppression makes it easy to select optimum values, as seen in the next section.

Optimizing the Sideband Suppression

The optimum values of R_iC_i can be easily calculated using elliptic functions. Typically, we want the worst-case suppression to be the highest possible. This leads us to the equal ripple or Chebychev approximation. The mathematics are straightforward and given in detail by Saraga.[3] For an upper and lower frequency of f_u and f_l respectively, the R_iC_i values for an n-section network are,

$$R_iC_i = \frac{\mathrm{dn}\left(\frac{2i-1}{2n}\mathrm{K}(k),k\right)}{2\pi f_l} \qquad \text{Eq 28}$$

where

$$k=\sqrt{1-(f_l/f_u)^2},\ \mathrm{K}(k)$$

is the complete elliptic integral of the first kind, and $\mathrm{dn}(u,k)$ is a Jacobi elliptic function.[4,5]

One of the FORTRAN programs calculates the R_iC_i values given the upper and lower frequencies and the n value. In Table 1, I give some calculated values for some networks of interest to hams, and their theoretical sideband suppression. These theoretical results will, of course, be best cases assuming perfect components.

In passing, I note that Saraga's Taylor approximation is given by simply choosing all the R_iC_i values to be the same and equal to

$$\frac{1}{2\pi\sqrt{f_uf_l}}$$

(see note 3). Also, if maximum suppression is needed at a particular frequency (for example if you wanted to use audio tones in a single-sideband transmitter to produce frequency shift keying), it is simple to select R_iC_i values appropriate for these frequencies and then optimize the other network sections.

Effects of Amplitude Variations and Component Tolerances

So far, I have only looked at the relative phase shift of the two outputs. To have a high-quality audio signal, the network must have a flat amplitude output. Usually, this is handled by constructing, respectively, an all-pass network. Since the HA5WH network is not an all-pass form, we must examine its attenuation as a function of frequency. In Figs 3, 4, and 5, I have plotted the sideband suppression and the amplitude and phase variations of one of the output signals, for the optimal 4, 6, and 8-section filters designed for the frequency range 300 to 3000 Hz with equal value resistors. The network sections are ordered from largest RC value to smallest, as in the original HA5WH design. As shown, the amplitude variations are less than ±1 dB, the phase variation is smooth, and the sideband suppression is of the equal ripple form—as expected.

One of the main selling points given in the *Handbook* description of this network is the claim that low-tolerance components can be used to obtain a high-performance network. From the analysis in the previous section, if cyclic symmetry is maintained, the network will perform

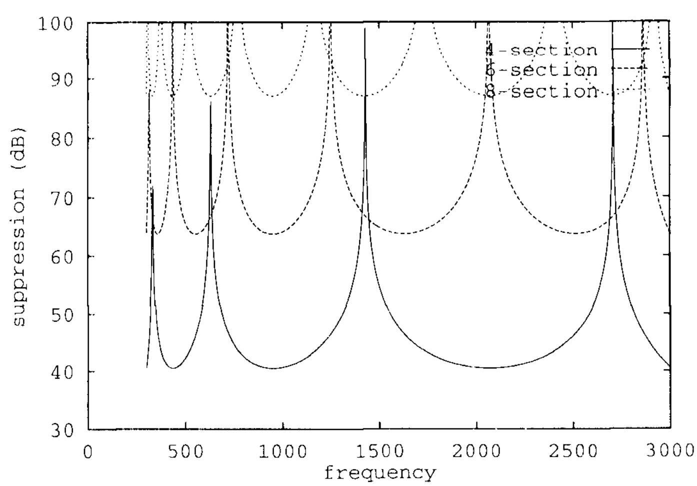

Fig 3—Ratio of the magnitude of the unwanted to wanted sideband for the 4, 6, and 8-section optimal Chebychev networks for the frequency range 300 to 3000 Hz.

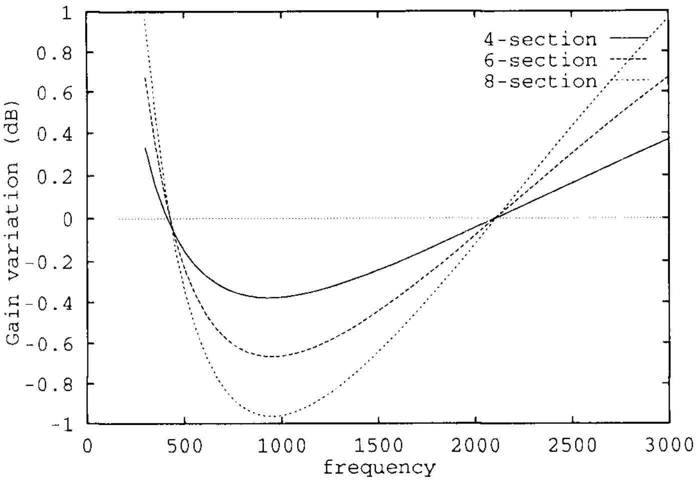

Fig 4—Relative amplitude of one output signal for the 4, 6 and 8-section optimal Chebychev networks for the frequency range 300 to 3000 Hz.

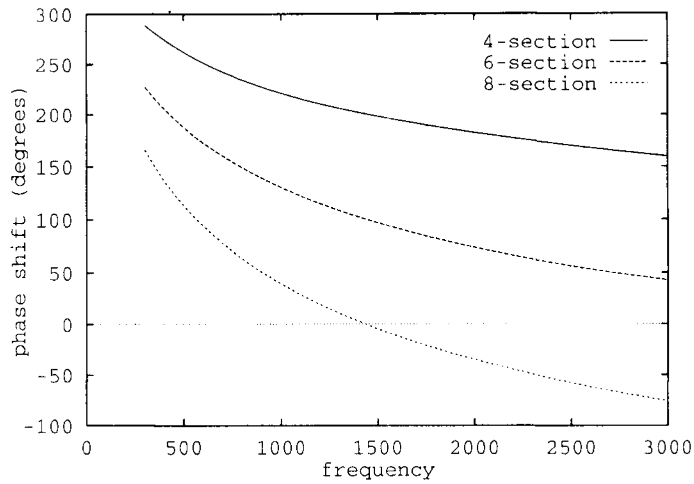

Fig 5—The phase shift of one output signal for the 4, 6, and 8-section optimal Chebychev networks for the frequency range 300 to 3000 Hz.

perfectly at the n selected frequencies corresponding to $f=1/(2\pi RC)$ for each network section. Since matching components is generally easier than measuring their values accurately, I examine the effect of a change of these node frequencies caused by perfectly matched, but low-tolerance components. Since both the resistors and capacitors can vary, using 10% components can vary the nodal frequency values by approximately 20% if both components change value in the same direction. A worst-case condition would be for all the sections to have too high or too low of a frequency by 20%. This simply shifts the network center frequency by 20%. For the optimal 6-section filter from 300 to 3000 Hz, this changes the sideband suppression from over 60 dB to about 42 dB. If a 10% variation of network node frequencies is assumed, that is 55 components, and again all the frequency changes are assumed to be in the same direction, the suppression is nearly 50 dB. This shows that relatively low-tolerance but *well-matched* components can give excellent results. Eq 27 can be used to predict the effect of changing the R-C values of each filter section due to component tolerances when the components are perfectly matched in each section. The case where unmatched R and C values in each section are used is of course the one with the most practical interest. Here, we can get an idea of what the worst-case possibilities are by looking at the cross terms between ψ_a and ψ_b when the M matrices are no longer cyclic. Typical terms give contributions like,

$$\left(\frac{1}{R_1}+\frac{1}{R_3}\right)-\left(\frac{1}{R_2}+\frac{1}{R_4}\right) \qquad \text{Eq 29}$$

or

$$\omega(C_1+C_3)-\omega(C_2+C_4) \qquad \text{Eq 30}$$

where here the subscripts 1, 2, 3 and 4 indicate the position in the network section as in Fig 2. This indicates that a single section with a tolerance t (t = 0.1 would be 10% tolerance) can reduce the overall suppression to roughly 20 $\log_{10}(t)$ dB. That is, 10% components could give suppressions as low as 20 dB, and 1% components as low as 40 dB if the components in a network section are not matched. Notice that to be sure to obtain 60-dB opposite sideband attenuation, components with short- and long-term tolerances of 0.1% would need to be used.

As a concrete example of this sensitivity to unmatched components, I calculated the suppression of the original HA5WH 6-section filter for the case where only the resistors in the last section have been changed by 10%. R_1 and R_3 have been raised by 10%, and R_2 and R_4 have been lowered by 10%. For ideal components, the suppression is greater than 57 dB, but changing just the resistors in the last section reduces the suppression to 26 dB, in rough agreement with the simple calculation above. Using these results to try to cook up a near worst case, I tried changing all the resistors in exactly the same manner in each section. In addition I changed all the capacitors by raising the C_2 and C_4 values by 10% and lowering the C_1 and C_3 values by 10%. The result was to further lower the unwanted sideband suppression to about 17 dB. Clearly, 10% components and bad luck will produce unacceptable sideband suppression.

One last comment on the *Handbook* circuit is the design of the operational amplifier circuit for the output. One section of this circuit is shown in Fig 6. All the resistors have the same value in the *Handbook* circuit. This does not give a balanced output and would be another source of phasing errors. If I assume that the operational amplifier input impedances are very large, the input impedance at point 2 is clearly $2R_2$. The voltage at the noninverting input is therefore $V_2/2$. The current drawn from input 1 is therefore

$$\frac{V_1 - V_2/2}{R_1}$$

and since $V_1 = -V_2$ with perfect phasing, the impedance seen at input 1 is $1.5R_1$. So $2R_2$ should be equal to $1.5R_1$, and in addition, dc balancing of the operational amplifiers may be required to compensate for input bias current. For the *Handbook* circuit, the unbalanced output resistance

Table 1

The optimal Chebychev values for some ideal HA5WH type phasing networks. f_l and f_u are the upper and lower frequencies, n is the order of the network, and f_i, where i is 1 through n, are the frequencies of exact 90° phase shift. The corresponding R-C values are $1/(2\pi f_i)$. Sup is the minimum sideband suppression over the network range in dB.

f_l	f_u	n	Sup(dB)	f_1	f_2	f_3	f_4	f_5	f_6	f_7	f_8
300	3000	4	40.5	332.2	629.8	1429.0	2709.0				
300	3000	5	52.1	320.5	500.7	948.7	1797.6	2808.1			
300	3000	6	63.7	314.2	435.5	720.3	1249.5	2066.8	2864.5		
300	3000	7	75.4	310.4	397.8	595.3	948.7	1511.8	2262.4	2899.4	
300	3000	8	87.0	308.0	374.0	519.4	771.2	1167.0	1732.7	2406.2	2922.5
200	4000	5	42.9	219.5	398.4	894.4	2008.1	3645.0			
200	4000	6	52.7	213.5	332.1	633.1	1263.6	2408.9	3747.8		
200	4000	7	62.5	209.9	294.6	497.5	894.4	1608.2	2715.5	3812.0	
200	4000	8	72.2	207.5	271.2	417.8	689.9	1159.6	1915.0	2949.6	3854.8
150	6000	6	44.7	163.6	287.7	628.9	1431.1	3128.3	5500.9		
150	6000	7	53.1	160.0	247.7	471.0	948.7	1910.7	3633.0	5626.4	
150	6000	8	61.5	157.6	223.1	381.3	696.7	1291.9	2360.2	4033.2	5710.4

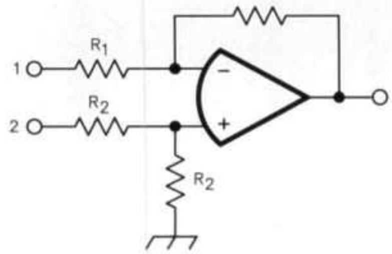

Fig 6—The schematic diagram of one operational amplifier section.

reduces the sideband suppression even in the ideal component case to about 35 dB.

Conclusion

The HA5WH network takes advantage of cyclic symmetry to give simple design equations and excellent sideband suppression with ideal components. If the cyclic symmetry is maintained, the network is not very sensitive to component tolerances. This means that the components in each of the network sections should be carefully matched. Breaking the cyclic symmetry by using unmatched components can drastically affect the performance of the network.

I have given a set of formulas and FORTRAN programs to design the optimum ideal networks and to analyze both the ideal and nonideal cases. Analyses other than the cases that I have described here can be easily done with these methods and programs.

Notes

[1] *The ARRL Handbook for the Radio Amateur*, (American Radio Relay League, Newington, 1993), and many previous editions.

[2] See for example, Cracknell, A.P., *Applied Group Theory*, (Pergaman, Oxford, 1968) for an introduction to group theory, with reprints of selected original papers.

[3] Saraga, W., "The Design of Wide-Band Phase Splitting Networks," *Proc IRE*, Vol 38, p 754 (1950).

[4] Cayley, A., *An Elementary Treatise on Elliptic Functions*, (Dover, New York, 1961).

[5] Abramowitz, M., and Stegun, I., *Handbook of Mathematical Functions with Formulas, Graphs, and Mathematical Tables*, National Bureau of Standards, Applied Mathematics Series, (US Government Printing Office, Washington, DC 1964). □□

Estimating the Length of Coaxial Feeder

How many parts does it take to build a cable-length measuring circuit? Would you believe—one?

By George Brown, G1VCY

How many times has the average amateur or SWL been faced with the question, "Have I enough cable on this drum to reach between transceiver and antenna?". Yes, there is one sure way to find out, but measuring the length of tens of meters of cable is by no means easy, bearing in mind the recalcitrant behavior of long lengths of feeder, whether it is RG58 or the semi-rigid variety. This article goes about things in a slightly unorthodox manner, but one which is guaranteed to produce a value for the length which can be trusted, while never unwinding the cable from the drum.

Introduction

When a length of coaxial cable is needed to stretch from the transceiver to the antenna, wherever it may be, a reasonable estimate can be made of the distance involved, knowing the approximate dimensions of rooms, garden, etc. The problem really begins, however, when sitting staring at a partially full drum of a favorite feeder, wondering whether there is enough for the job. There are three obvious means of solving the problem: (1) use the feeder as a short- or open-circuited line, measuring its resonant frequency and hence its length; (2) send a series of fast rectangular pulses down the cable, and use an equally fast double-beam oscilloscope to measure the delay between one end of the cable and the other; (3) unwind the whole drum in the garden and measure it! The third method has major advantages—it requires no test equipment and is very accurate. It does, however, require a fine day, much patience and the muscular ability to wrestle with that long, black snake that doesn't want to be coiled up again—very much a case of trying to get the genie back into the bottle! The first method implies that the experimenter has access to a wide-range RF source, a means of detecting resonance in the cable and a way of sorting out the possible ambiguity in the result. The second technique is fraught with termination and matching problems and, for smaller lengths of cable, the delays will be indistinct and *very* short.

The method described here involves using the signal velocity in the cable to produce a delay, and using that delay to control the frequency of an oscillator based upon a single logic gate. The period can be read from the screen of an oscilloscope (by counting several cycles of the wave and averaging). If the period involved is too small for direct measurement, it can be changed to a convenient value using a chain of binary dividers, measured, and the resulting value rescaled by the same factor. I've measured feeder lengths between 100 m and 5 m, and the values obtained were always within 1% (and nearer 0.1% for long lengths) of the correct value. The use of this technique is not necessary on lengths less than about 10 m, but even so, it has been proved to work on such short lengths.

Theory

Electromagnetic waves travel through free space with a speed of 2.998×10^8 meters per second. Whenever they travel through a different medium, their speed is reduced (some modes of wave-guide propagation excepted). In a coaxial feeder, the center conductor needs support. This is done with a plastic material whose characteristics are

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carefully chosen to give the cable the correct impedance and loss figures. Sometimes the plastic material (or *dielectric*) is solid, sometimes it is a helical strand around the center conductor, and sometimes its bulk is reduced by the presence of "strands" of air running through an otherwise solid material. All these differences give rise to slight variations in the speed with which a radio wave is propagated along the line; the more air there is, the faster it travels.

Many readers will recognize the formula for the phase velocity of an electromagnetic wave as

$$v_p = \frac{1}{\sqrt{\mu\varepsilon}} \qquad \text{Eq 1}$$

where μ is the permeability of the medium, and

ε is the permittivity of the medium.

For any medium, $\mu=\mu_0\mu_r$ and $\varepsilon=\varepsilon_0\varepsilon_r$ where

μ_0 is the permeability of free space;

μ_r is the *relative permeability* of the medium;

ε_0 is the permittivity of free space; and

ε_r is the *relative permittivity* of the medium (otherwise known as the *dielectric constant*).

In free space (and air, for all practical purposes), the values of μ_0 and ε_0 are as follows:

$$\mu_0 = 4\pi \times 10^{-7}\,\text{Hm}^{-1}, \quad \text{and}$$

$$\varepsilon_0 = \frac{1}{36_\pi} \times 10^{-9}\,\text{Fm}^{-1}$$

Eq 1 can now be written

$$v_p = \frac{1}{\sqrt{\mu_0\varepsilon_0}} \times \frac{1}{\sqrt{\mu_r\varepsilon_r}} \qquad \text{Eq 2}$$

But the free-space velocity is the velocity of light, c, and is given by

$$c = \frac{1}{\sqrt{\mu_0\varepsilon_0}} \qquad \text{Eq 3}$$

Therefore, Eq 2 can be written as

$$v_p = \frac{c}{\sqrt{\mu_r\varepsilon_r}}$$

Now, the velocity factor, *VF*, of a cable, is the propagation velocity in the cable expressed as a fraction of the free-space velocity, ie,

$$VF = \frac{v_p}{c}$$

or,

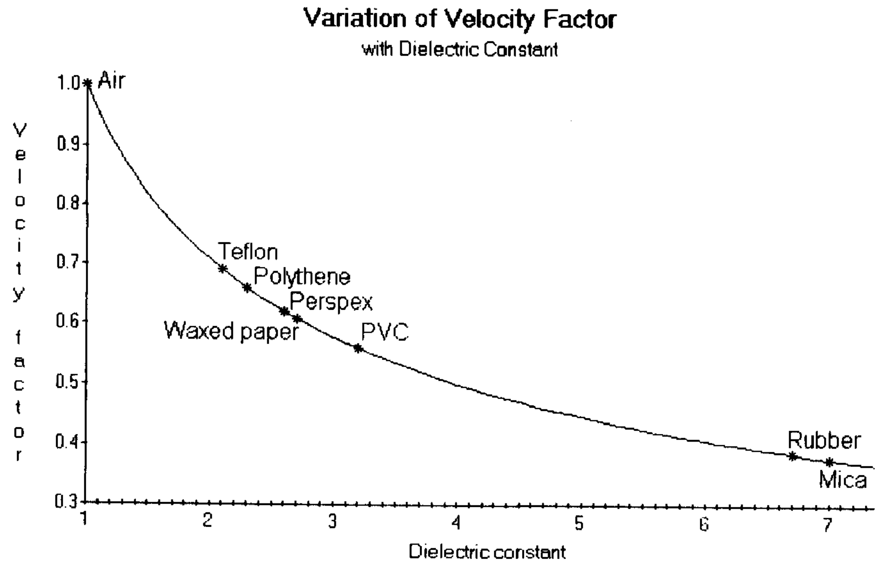


Fig 1—The positions of some common materials on the velocity factor curve.

$$VF = \frac{1}{\sqrt{\mu_r\varepsilon_r}} \qquad \text{Eq 4}$$

Provided that the medium is non-magnetic (which is true), $\mu_r = 1$, in which case Eq 4 reduces to:

$$VF = \frac{1}{\sqrt{\varepsilon_r}} \qquad \text{Eq 5}$$

So, reduced to the simplest terms, the velocity factor of a feeder cable is equal to the reciprocal of the square root of the effective dielectric constant of the material between the center conductor and the braid. Fig 1 shows the variation of velocity factor with dielectric constant (relative permittivity), with some common materials (from air to mica) marked on the curve. It is this velocity factor that makes the following technique possible.

Circuit

Many readers will recognize Fig 2 as being a very simple relaxation oscillator based upon a Schmitt trigger inverter. The period of oscillation (ie, the time taken for one complete cycle) is determined by the time the capacitor, C, takes to charge and discharge through the resistor, R, between the two input voltage thresholds of the Schmitt trigger. As with all astable circuits, the period is the result of a simple delay circuit, in this case the resistor/capacitor combination.

Fig 3a shows the circuit of the delay-line oscillator used for measuring cable lengths. A friend called it "the

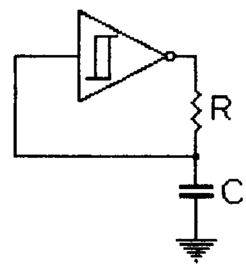


Fig 2—A simple relaxation oscillator.

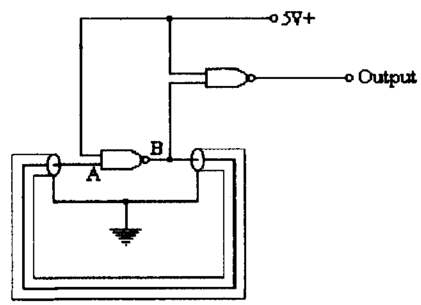


Fig 3a—The delay-line oscillator used to measure feeder length, (for details of the logic gates, see text).

circuit with no components"—a slight over-simplification! It is based upon a single TTL inverter, a NAND gate with one input tied to logic 1, followed by a second gate to buffer the output from the first gate. Provided that the propagation delay in the logic gate is appreciably smaller than the cable delays to be measured, it will cause no serious impairment of the measurement. For cables down to 5 m in length, a 74LS00 is only just acceptable, and a 74AC00 is recommended. CMOS logic will *not* work in this circuit.

The NAND gate is being used as an inverter so, ignoring the propagation delay in the gate itself, whenever A is logic 1, B must be logic 0, and *vice versa*. Take the initial condition A=1; the circuit is stable and will remain so unless disturbed. (This is the case during the period **ab** in Fig 3b). Now suppose a signal arrives at A which takes it down to logic 0; B rises immediately to logic 1, this is at time **b**. The signal corresponding to this sudden change of level is propagated down the cable, during which time the circuit is stable again. This is period **bc** on the diagram. **bc** corresponds to a time, τ, the propagation delay of the cable (ie, the time taken for a signal to traverse the length of the cable). After this time, the rising edge from 0 to 1 reaches A (at time **c**), and the inverter changes state again. This process is repeated indefinitely, and the circuit becomes an oscillator whose period of oscillation is 2τ, as can be seen from the diagram of Fig 3b. If the period can be measured, the cable delay is given by:

Delay = period / 2

If the reader is contemplating using an oscilloscope to measure the period directly, he should bear in mind that, for cable lengths in the 5-10 m range, the oscilloscope should have a time-base speed faster than 50 ns per centimeter, and a vertical-amplifier frequency response of around 20 MHz. To avoid this constraint, and to make the technique usable with low-frequency oscilloscopes, the frequency must be divided to produce a usable figure.

Fig 4 shows a version of the circuit that can be used with an oscilloscope of very modest specifications, provided that its time base is reasonably well calibrated. For cable lengths between 5 m and 100 m, the approximate frequencies of oscillation are 20 MHz and 1 MHz, respectively. To bring these frequencies down to convenient figures, the oscillator frequency is divided by a factor of 256 in two 4-bit binary counters. By this method, the frequency extremes become 78 kHz and 3.9 kHz approximately, which are within the range of the most modest oscilloscopes. The new technique involves the same counting and averaging method as before, except that the period obtained needs to be divided by 256 to bring it back to the true value.

It is also possible to achieve the same measurement accuracy using a calibrated audio frequency generator.

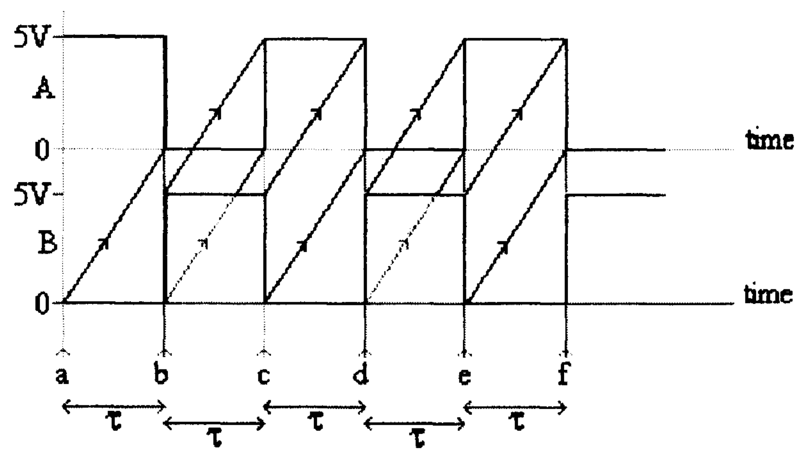


Fig 3b—Timing diagram for the operation of the oscillator.

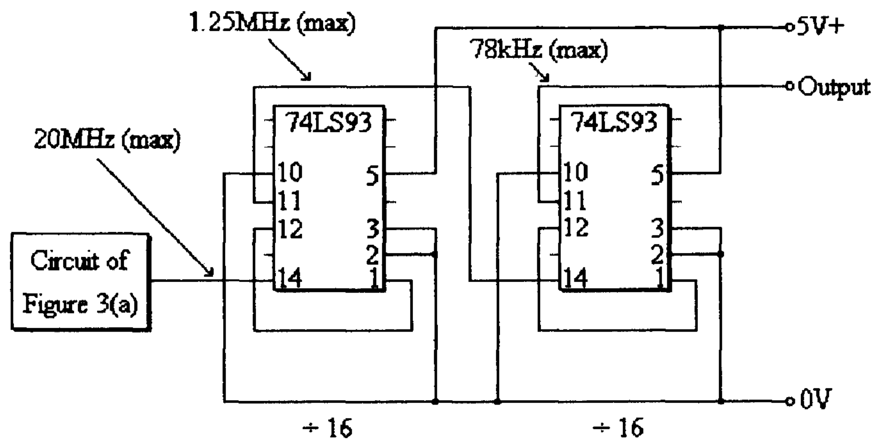


Fig 4—Frequency divider circuit (÷256).

Table 1—Characteristics of Some Well-Known Coaxial Feeders.

Cable Type	*Impedance (Ω)*	*Velocity Factor*	*Approx UR Equivalent*
RG58/U	53.5	0.659	UR43
RG213/U	50.0	0.66	UR67
RG58C/U	50.0	0.659	UR76
—	50.0	0.96	UR83

The frequency produced by the delay-line oscillator can be compared with that of the frequency generator using the standard method of "beats." The zero-beat frequency can be converted to period using the relation:

period = 1 / frequency

at which point the timing information is the same as that obtained from the oscilloscope.

Before any calculations can be performed, the velocity factor of the feeder in use must be known. Table 1 is a table of such values for some of the more common cables.[1]

Notice that the solid dielectric types

[1]Jessop, G. R., *Radio Data Reference Book*, 5th Edition, RSGB 1985.

have velocity factors which are basically the same. UR83 has a helical membrane, which explains the high value.

Fig 3b showed how the delay in the cable was equal to one-half of the period of oscillation. Suppose the delay is calculated to be d seconds, and the velocity factor is VF. Use the relation:

$$\text{Time} = \text{Distance} / \text{Speed} \qquad \text{Eq 6}$$

and rearrange it to give

$$\text{Distance} = \text{Speed} \times \text{Time} \qquad \text{Eq 7}$$

The speed, c', of the signal in the cable is given by

$$c' = VF \times c \ \mathrm{ms^{-1}} \qquad \text{Eq 8}$$

where c is the velocity of light in free space (see Eq 3).

The time taken for the signal to travel the length of the cable is d seconds, as measured. Thus, the length, L, of cable that would produce this delay is:

$$L = c' \times d \qquad \text{Eq 9}$$

or

$$L = VF \times c \times d \text{ meters} \qquad \text{Eq 10}$$

For example, suppose a delay of 33 ns (corresponding to a frequency of 15.15 MHz) was measured in a length of UR67 cable. First, the delay must be expressed in seconds, so this becomes $d = 33 \times 10^{-9}$ s. Using the velocity of light as $c = 3 \times 10^{8}\ \mathrm{ms^{-1}}$, the expression for the length becomes:

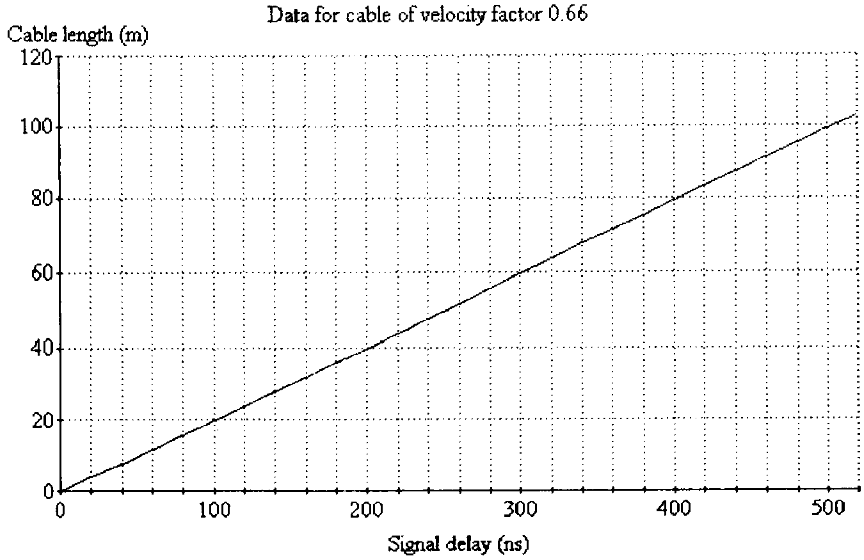
$$L = 0.66 \times (33 \times 10^{-9}) \times (3 \times 10^{8})$$
$$= 6.5 \text{ meters.}$$

If you think mathematics smacks of black magic, a graph is given in Fig 5 by which the cable length can be read directly from a knowledge of the delay, for cables of velocity factor 0.66. The accuracy is obviously not as great as by direct evaluation, but it may serve as a check on any calculation.

The method of construction of the circuit is not critical; the prototype was built on a plug-in type breadboard and worked satisfactorily on cable lengths from 10 meters upwards.

Conclusion

The circuit itself is so simple that users may find it the sort of test equipment that is built when needed, then dismantled until the next time, thus saving precious space in the shack! So, build yourself this little circuit, *prove* that it does what is claimed, and prevent yourself from being in that awkward and embarrassing position when you have rolled out your drum of expensive feeder, one end connected to the antenna feed point, and find that it ends about a foot from your rig! □□



Building VHF Power Attenuators

Getting transmitter output levels down to transverter input levels the easy way.

By Paul Wade, N1BWT

An attenuator, or pad, is frequently needed in ham equipment to reduce power, gain or signal levels. Tables of resistor values are available in most handbooks, so design is not difficult.[1,2] However, if significant power handling is required, power resistors suitable for RF frequencies can be difficult to locate. The tables below and the computer program that generated them can be used to make do with available components.

For example, all my microwave transverters are designed to be driven with the 2-W output of an old ICOM IC-202, which is ideal for portable operation. When I wish to use them at home with a larger transceiver, or a friend wants to use one with a more modern rig, much more power is available. We could push the low-power button, adjust the output, and hope that we don't forget next time...

I prefer to make things fool-resistant (nothing is foolproof!) and avoid smoke. So a resistive attenuator is needed. A typical 10-W transceiver for two meters delivers about 14 W of output power at 13.6 V, so about 8 dB of attenuation is necessary, capable of dissipating 14 − 2 = 12 W.

Notes appear on page 29.

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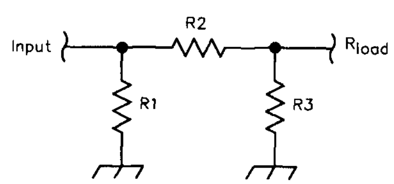


Fig 1

The largest resistor that works well at two meters is a 2-W carbon composition type, but these aren't readily available anymore. A survey of the junkbox and the local surplus emporium yielded only a few values of 1- and 2-W resistors, so I had to design around these values.

Next, I had to figure out how much power is dissipated in each resistor. If we examine a pi attenuator, Fig 1, we can readily determine the voltage at each end from the attenuation:

$$V_{out} = V_{in} \times 10^{-dB/20}$$

Since all the resistors are connected to the ends or to ground, we know the voltage across each resistor, and power is just

$$Power = \frac{V^2}{R}$$

The powers tabulated in Table 1 list the power dissipated in each resistor as a percentage of the input power—anything left over is the output power. These powers are correct only if the input and output impedances are close to the design value (usually 50Ω), since reflected power from mismatches must also be dissipated.

For a T attenuator, Fig 2, we perform the same sort of calculation using the current in each resistor, but only a couple of calculations are necessary before we notice that the power in R1, R2 and R3 is the same for pi and T attenuators of the same attenuation. Thus, there is only one set of power numbers in Table 1.

Getting back to our example, in order to dissipate 12 W in 1- and 2-W

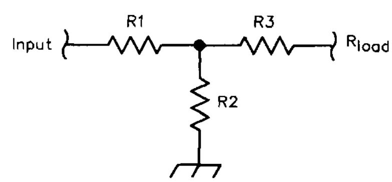


Fig 2

Table 1—Resistance and Power Dissipation for T and Pi Attenuators with 50-Ω Input and Output Impedance.

Loss	*—T—*		*—pi—*		*Power dissipation*		
dB	*R1,R3*	*R2*	*R1,R3*	*R2*	*R1*	*R2*	*R3*
1	2.88	433.34	869.55	5.77	5.8%	10.2%	4.6%
.5	4.31	288.1	580.5	8.68	8.6%	14.5%	6.1%
2	5.73	215.24	436.21	11.61	11.5%	18.2%	7.2%
2.5	7.15	171.34	349.83	14.59	14.3%	21.4%	8%
3	8.55	141.93	292.4	17.61	17.1%	24.2%	8.6%
3.5	9.94	120.79	251.52	20.7	19.9%	26.6%	8.9%
4	11.31	104.83	220.97	23.85	22.6%	28.6%	9%
4.5	12.67	92.32	197.32	27.08	25.3%	30.2%	9%
5	14.01	82.24	178.49	30.4	28%	31.5%	8.9%
5.5	15.32	73.92	163.17	33.82	30.6%	32.5%	8.6%
6	16.61	66.93	150.48	37.35	33.2%	33.3%	8.3%
6.5	17.88	60.96	139.81	41.01	35.8%	33.8%	8%
7	19.12	55.8	130.73	44.8	38.2%	34.2%	7.6%
7.5	20.34	51.29	122.92	48.74	40.7%	34.3%	7.2%
8	21.53	47.31	116.14	52.84	43.1%	34.3%	6.8%
9	23.81	40.59	104.99	61.59	47.6%	33.8%	6%
10	25.97	35.14	96.25	71.15	51.9%	32.9%	5.2%
11	28.01	30.62	89.24	81.66	56%	31.6%	4.5%
12	29.92	26.81	83.54	93.25	59.8%	30.1%	3.8%
13	31.71	23.57	78.84	106.07	63.4%	28.4%	3.2%
14	33.37	20.78	74.93	120.31	66.7%	26.6%	2.7%
15	34.9	18.36	71.63	136.14	69.8%	24.8%	2.2%
16	36.32	16.26	68.83	153.78	72.6%	23%	1.8%
17	37.62	14.41	66.45	173.46	75.2%	21.3%	1.5%
18	38.82	12.79	64.4	195.43	77.6%	19.5%	1.2%
19	39.91	11.36	62.64	220.01	79.8%	17.9%	1%
20	40.91	10.1	61.11	247.5	81.8%	16.4%	0.8%
21	41.82	8.98	59.78	278.28	83.6%	14.9%	0.7%
22	42.64	7.99	58.63	312.75	85.3%	13.5%	5%
23	43.39	7.12	57.62	351.36	86.8%	12.3%	4%

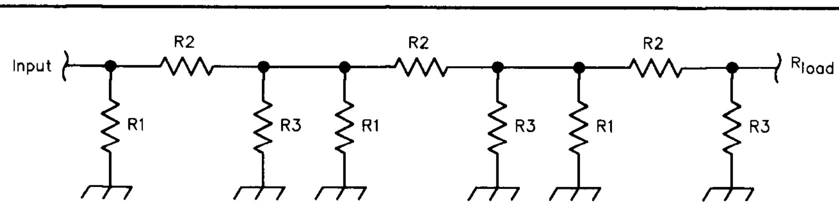


Fig 3

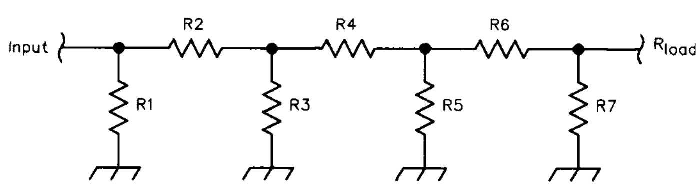


Fig 4—

Resistor	*R(Ω)*	*Power(W)*	*Implementation*
R1	**870**	**0.9**	**Two 1.8 k, ½ W (parallel)**
R2	**6**	**1.56**	**Three 10 Ω, 1 W (parallel)**
R3	**220**	**2.76**	**Two 120 Ω, 2 W (series)**
R4	**18**	**2.55**	**Three 56 Ω, 1 W (parallel)**
R5	**126**	**2.36**	**220 Ω , 2 W parallel with 300 Ω, 1 W**
R6	**24**	**1.24**	**Two 56 Ω, 1 W (parallel)**
R7	**220**	**0.54**	**220 Ω, 1 W**

resistors we must find series and parallel combinations that equal the required resistance and can handle the power. For an 8-dB attenuator, R1 must dissipate 43% of the input power, or about 6 W.

On the other hand, we could put together a series of small attenuators that added up to 8 dB, each dissipating a part of the power. For instance, a 1-dB attenuator only dissipates 20.5% of the input power, or about 3-W total, of which about 1.5 W is in R1. Obviously, we could stack up eight 1-dB attenuators, or succeeding ones, which only have to handle the remaining power, could have higher attenuation.

What I did was to look at Table 1 and mark all the resistances for which I had something close. Then I marked the values I could approximate by paralleling two (half the resistance) or three (one-third) identical resistors, or two identical ones in series (twice the resistance). Now I had an idea which attenuators I could make; a few more calculations gave me an idea how much power each could handle. The final configuration was 1 + 3 + 4 dB, all of the pi type, as shown in Fig 3. The next step was to combine the end resistors of adjacent sections as shown in Fig 4, with the actual resistor combinations I used. Note that this combination is not bilateral—if the ends are reversed, smoke may result!

I built this unit in a small metal box with two coax connectors from a recent hamfest. The measured attenuation at two meters was 8.7 dB, with a VSWR of about 1.15. The output power was a bit less than I wanted, so I made small adjustments at the output end (so the VSWR was not affected much), ending up with the final values shown in Fig 4. Now the output power is exactly 2 W, and the resistors are barely warm after several minutes with key down.

Conclusion

Using Table 1 and a hand calculator, you can quickly design an attenuator for any needed attenuation and power level, using available components. The program PAD.EXE may be used to calculate other attenuations, attenuators with input and output impedances other than 50 Ω, and values for bridged-T type attenuators. For those inclined to computer programming, the source code is available (for further improvement) for download from the ARRL BBS (203 666-1578) and via Internet FPT from ftp.sc.buffalo.edu ithe\pub\ham-radio directory. The file name is QEXPAD.ZIP.

Notes

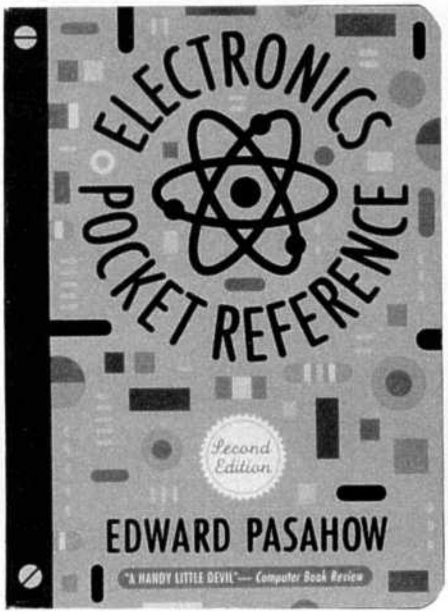
[1] *ARRL Handbook for Radio Amateurs*, ARRL, 1992, p 25-39.

[2] *Reference Data for Engineers: Radio, Electronics, Computer, and Communications*, Seventh Edition, Sams, 1990, pp 11-3 to 11-7. □□

Book Review

Electronics Pocket Reference, Second Edition, ISBN: 0-07-048737-5. By Edward Pasahow. 1994, McGraw-Hill Inc, 1221 Avenue of the Americas, New York, NY 10020. Tel 800 262-4729. Paperbound, 4 × 5½ inches, 528 pp. B&W illus. $24.95.

Reviewed by Larry Wolfgang, WR1B, ARRL Senior Assistant Technical Editor



The first edition of this book was titled *Electronics Ready Reference Manual*. Although not quite small enough to fit in a shirt pocket, the *Electronics Pocket Reference* would slip easily into a jacket pocket, briefcase or purse. The book is about ¾ inch thick.

In the back of the book we can learn a little about the author. Edward Pasahow is an electronics instructor at San Diego Community College. He holds a BS degree in Electrical Engineering from the University of Washington and an MS degree in Management Science from the United States International University. Previously he was technical director at PRC, Inc, and a senior member of the technical staff of Ball Systems Engineering Division.

The Table of Contents is quite detailed, taking up seven pages in the front of the book. The 16 chapter titles reveal something of the scope of this little book: **Definitions and Equations, Passive Components, Active Components, Linear Circuits, Filters, Power Supply and Regulation, Electronic Measurement, Communications, Digital Circuits, Microprocessors and Computers, Electronics Mathematics, Mathematical Tables and Formulas, Symbols, Conversion Formulas and Tables, Properties of Materials**, and **Safety and First Aid**. The main topics under each chapter are also listed here, making it a fairly simple task to locate the appropriate text for most topics.

Twelve pages of index entries also provide a good way to locate the text about any topic covered in the book. The cross references seem adequate, and the detail is good.

In the front of the book, there is a list of figures that fills six pages and a list of tables that takes up another five pages. These lists are really handy. Many times I've recalled seeing a particular figure or table in a book, and then spent too much time trying to locate it later for reference. With the figures list in *Electronics Pocket Reference*, I can quickly locate the diagram showing bandwidth and half-power points in a resonant circuit on page 12, or the 555 timer pin assignments on page 98. The list of tables will help quickly locate the table of international time and frequency stations on page 185. I even discovered a table listing the frequencies of the notes on a piano keyboard by scanning this list!

What will you find inside this book? Very little text to read, for one thing, but lots of information. The first chapter is **Definitions and Equations**. The information here is brief and to the point. The first listing is for the ampere (A). "One ampere is the constant current flowing in two parallel conductors one meter apart that would produce a force of 2×10^{-7} newtons per meter of length." There are lots of equations, many shown with alternate forms solved for the quantities of interest, such as three versions of Ohm's Law, or the resonance formula solved for resonant frequency, inductance or capacitance. You will find the basic time constant equations for R-C and R-L circuits, and a table of values for charging and discharging the circuit at various time-constant multiples.

There are diagrams and graphs to illustrate the basic concepts, along with simple examples. You won't find much text explanation for the mathematics and little or no explanation of the significance of a given topic nor how the electronics really works, however. The book isn't intended to *teach* you about electronics, it is intended to *remind* you of the proper equations and provide a ready reference for various data.

In the **Active Components** chapter you will find information about LEDs, a summary of transistor amplifier equations and a table of common TTL IC types and drawings of pin assignments, among other things. There is even a handy listing of semiconductor letter symbols and their meanings. So if you are seeing repeated references to I_{FXM}, and are unsure what this group of letters represents, you can check the *Electronics Pocket Reference* to learn that it refers to the "peak forward blocking current" of a thyristor.

The **Microprocessors and Computers** chapter includes a wealth of information about computer systems. In addition to instruction-set information for several microprocessors, there are block diagrams and pin-out assignments for several versions of the Motorola 68040 processor. Other tables include those with connector information for PC edge cards, the SCSI bus interface, the IEEE 488 bus interface and the RS-232 interface.

I found the **Electronics Mathematics** chapter to be particularly useful. There are reminders of many basic algebra rules, plane and solid geometry formulas and an extensive section on trigonometry. **Mathematical Tables and Formulas** is another handy chapter. It includes a table of mathematical constants that are sometimes required for calculations, but are difficult to remember, such as the value for *e* (2.7182818285). The section about decibels gives the important equations along with a useful table of decibel values and power, voltage and current ratios. I'm not sure when I would ever need a table of 2500 five-digit random numbers, but if I ever do, I can find it in the *Electronics Pocket Reference*!

The **Symbols** chapter includes a listing of the upper and lower-case characters of the Greek alphabet. There is also an extensive table of electronics schematic-diagram symbols. It is hard to imagine that I will ever need to perform a unit conversion that isn't covered somewhere in the 25 pages of conversion factors. There is a two-page temperature conversion table that I found especially interesting. Read a known value in bold type on this chart. The value to the left is the Celsius equivalent of a Fahrenheit value or the one to the right is the Fahrenheit equivalent of a Celsius temperature. It is a very easy to use chart, with values given for every degree from 0 to 100, and every ten degrees down to −100 and up to 500. (This means the table covers a range between −148° and 932°F or −73.3° to 260°C.)

I believe many amateurs would find the *Electronics Pocket Reference* to be a valuable addition to their technical library. Although the nearly $25 price tag may seem high for such a small book, I am sure it will be referenced often. The binding seems to be of good quality, and the paper cover has a plastic lamination to increase its durability. I would have preferred a binding method that would allow the book to lay open on my desk, but that is one of the few complaints I might make about this book. □□