QEX

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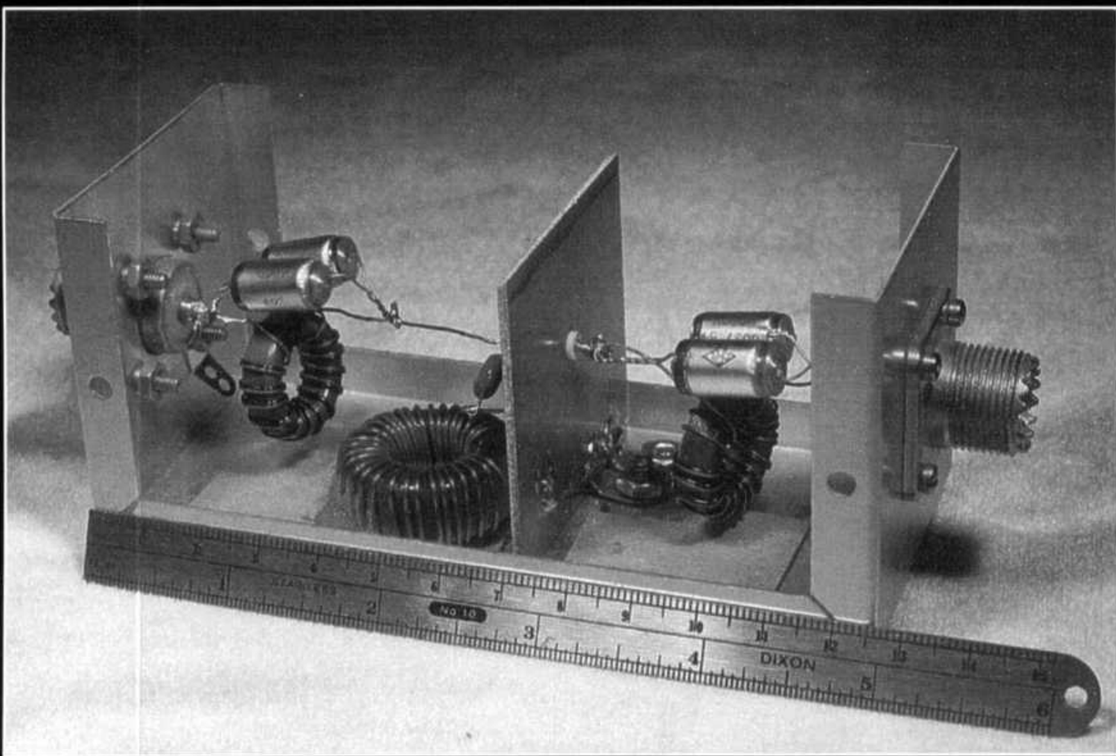
ARRL Experimenter's Exchange

November 1995



Effective AM-BCB Suppression Filters

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David Sumner, K1ZZ
Publisher

Jon Bloom, KE3Z
Editor

Lori Weinberg
Assistant Editor

Harold Price, NK6K
Zack Lau, KH6CP
Contributing Editors

Production Department

Mark J. Wilson, AA2Z
Publications Manager

Michelle Bloom, WB1ENT
Production Supervisor

Sue Fagan
Graphic Design Supervisor

Joseph Costa
Technical Illustrator

Joe Shea
Production Assistant

Advertising Information Contact:

Brad Thomas, KC1EX, *Advertising Manager*
American Radio Relay League
860-667-2494 direct
860-594-0200 ARRL
860-594-0259 fax

Circulation Department

Debra Jahnke, *Manager*
Kathy Fay, N1GZO, *Deputy Manager*
Cathy Stepina, *QEX Circulation*

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Telex: 650215-5052 MCI
FAX: 860-594-0259 (24 hour direct line)
Electronic Mail: MCIMAILID: 215-5052
Internet:qex@arrl.org

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About the Cover

Keeping the local AM station out of your 160-m receiver can be a challenge. Band-stop filters from W3NQN meet that challenge.

ISSUE NO. 165

Features

Columns

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Empirically Speaking

Need Technical Information?

Most *QEX* readers are no doubt aware of the several ARRL publications that serve up technical information. Aside from *QEX*, there is *QST*, ARRL's official journal, which carries several technical articles each month. And ARRL publishes a number of technical books, of which the annual *Handbook* is the "flagship," but including specialty books of diverse technical subjects, among them RF design, digital communication, antennas and spread spectrum. The question this month is: What, in your opinion, are we really missing? What aspects of communication technology, as it applies to Amateur Radio, are you having trouble finding information about? Where in your amateur efforts are you blocked by not having ready access to technical information?

Having diverse publications available in which to cover technical subjects gives us flexibility to cover a subject in the most fitting manner. For subjects that are narrowly focused, *QEX* or *QST* treatment may be best. Those areas that have broader application may be better described in the *Handbook*, while a yet broader subject, of wide appeal, may even warrant a book of its own.

Of course, we here at ARRL, with the aid of our members, try to track developing technology and get it in front of our readers. In fact, that's pretty much the mission of *QEX*. But in an effort such as this, you can't have too much input. We're open to suggestions; what are yours?

ARRL Electronic Sources

Speaking of information available from ARRL, this might be a good time to recap the various ways in which you can get information electronically from ARRL HQ.

The oldest electronic service of ARRL is the ARRL BBS, which has been operating for more than 10 years. Currently, the BBS handles 4 lines using V.34 (28.8-kbit/s) modems. Operated by the Field Services Department at ARRL HQ in Newington, the BBS includes hundreds of downloadable files, those from *QEX* among them. The telephone number of the ARRL BBS is 860-594-0306.

ARRL also provides several Internet-based services. A few files—*QEX* files and informational files about the ARRL and Amateur Radio—are available via anonymous FTP from ftp.arrl.org. A great deal of text information is available from the ARRL INFO server, an automatic email server at ARRL HQ. To use it, you simply send an Internet message containing a few simple commands. The server responds by mailing back one or more messages containing the information you requested. To get information about use of the INFO server, send a message to info@arrl.org containing the text lines:

HELP
QUIT

You need not supply a subject line for the message.

The files available via the ARRL BBS and those from the INFO server are also available via anonymous FTP at oak.oakland.edu in the /pub/hamradio/arrl directory.

The newest ARRL Internet information service is *ARRLWeb*, the ARRL's World Wide Web site, which has come on-line only within the past few months. Available using the URL http://www.arrl.org/, *ARRLWeb* provides links to ARRL file archives, up-to-the-minute ARRL bulletins, information about ARRL publications and software, and more.

This Month in *QEX*

Is that local AM broadcast station overloading your 160-m receiver? Try using "Band-Stop Filters for Attenuating High-Level Broadcast-Band Signals," as described by Ed Wetherhold, W3NQN.

Maybe homebrewing an entire radio is too much to fit into your busy schedule. If so, try "Homebrewing Black-Box Style." Hugh Duff, VA3TO, describes how to build an FM transceiver from a Kenwood plug-in module.

Modern commercial receivers usually include a variable-bandwidth IF. That's not the case with many homebrew units, but "A Variable IF Selectivity Unit," by A. R. Thomson, GM3AHR, (reprinted from September 1995 *Radio Communication*) might just drop into the 9-MHz IF strip of your latest homebrew goody.

In this month's "RF" column, Zack Lau, KH6CP/1, describes a buildable transition from 10-GHz waveguide to an SMA connector. You can stop scouring the flea markets for those items!—*KE3Z, email: jbloom@arrl.org*

Band-Stop Filters for Attenuating High-Level Broadcast-Band Signals

Eliminating AM BC interference can be a problem for the 160-meter operator. Here's a solution.

By Ed Wetherhold, W3NQN

Introduction and Background

This article was prompted by a request from Paul Elliott, W5DM, of Hobbs, New Mexico, for a filter to attenuate the signal of a nearby 1.48-MHz broadcast-band station that was causing receiver overload and preventing reception in the 160-m band. An examination of *The 1996 ARRL Handbook* showed that the only filter design intended to attenuate broadcast-band signals was not suitable for this particular application because it provided only about 24 dB of attenuation at 1.48 MHz when more than 60 dB is needed.[1]

A request from a *QST* reader in "The Doctor is IN" column asking what to do about a receiver overload problem caused by a nearby BC-band transmitter also indicates the need for a band-stop filter designed specifically for attenuating a particular broadcast-band signal.[2] One of the Doctor's suggestions was to use the *Handbook* high-pass rejection filter. However, this filter will solve the overload problem only if the BC-station frequency is below 1.3 MHz, where the filter provides at least 60 dB of attenuation.

It appears that a tabulation of band-stop filters (BSF) covering the broadcast-band frequency range and using only standard-value capacitors would be a useful addition to the tables of standard-value-capacitor low-pass and high-pass filter designs currently in the *ARRL Handbook*.

This article discusses the procedures involved in designing and testing BSFs. A table of 25 computer-calculated broadcast-band BSFs is presented in which only standard-value capacitors are used, to simply the filter construction. An example of the construction and testing of a BSF is given, and sufficient background information is provided so the design principles may be applied to other BSF applications.

Design Approach

Before starting the filter design, you should first confirm that the interfering signal is entering the receiver via the antenna and not by other paths. This can be done by disconnecting the antenna from the receiver and placing a 50-Ω resistor across the antenna terminals. If the interference disappears, the interfering signal path is probably via the antenna, and a BSF placed between the antenna and the receiver input terminals may solve the overload problem. Also, the amount of filter insertion loss needed to prevent

[1]Notes appear on page 12.

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1426 Catlyn Place
Annapolis MD 21401-4208

overload can be estimated by placing a resistive attenuator having about 60 dB of loss between the antenna and receiver. If receiver overload is eliminated, 60 dB of filter attenuation will be sufficient.

The usual procedure in designing a band-stop filter is to first design a high-pass prototype having a cutoff frequency equal to the desired bandwidth (BW_{Ap}) of the BSF and also having an appropriate reflection coefficient. The high-pass prototype is then transformed into a BSF by resonating the high-pass elements to the desired center frequency of the BSF.

For generating a table of BSFs, this procedure is slightly modified by using several pairs of standard-value capacitors to evaluate designs for a particular BSF center frequency and then picking one design that has suitable parameters for the final tabulation. The filter impedance level is fixed at 50 Ω because this is the impedance level of most receiving systems. For these designs, a total BSF attenuation of about 60 dB is assumed to be sufficient, and a third-order high-pass filter (to be transformed into a three-resonator BSF) is used.

Fig 1 shows the proposed 3-resonator filter in a T-network configuration. Because it needs only one ground connection, the T-network was chosen over the alternate pi-network. High-level circulating ground currents are anticipated, and there might be a problem in finding a "clean" ground; consequently, it is advisable to minimize the number of ground connections.

Of the two choices between the Butterworth and Chebyshev filter types, the Chebyshev was chosen because of its better selectivity and because it permits several different pairs of capacitor values to be considered, whereas the Butterworth has only one pair of values. Fig 2 shows the typical insertion- and return-loss responses of a lossless 3-resonator Chebyshev BSF having a reflection coefficient (RC) of about 6% and with an arbitrarily chosen center frequency

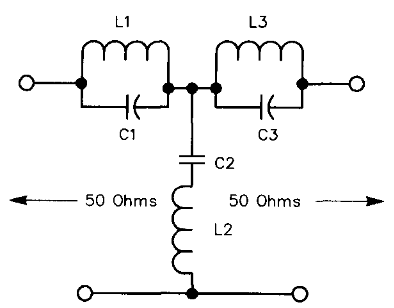


Fig 1—Schematic diagram of the 50-Ω, 3-resonator band-stop filter.

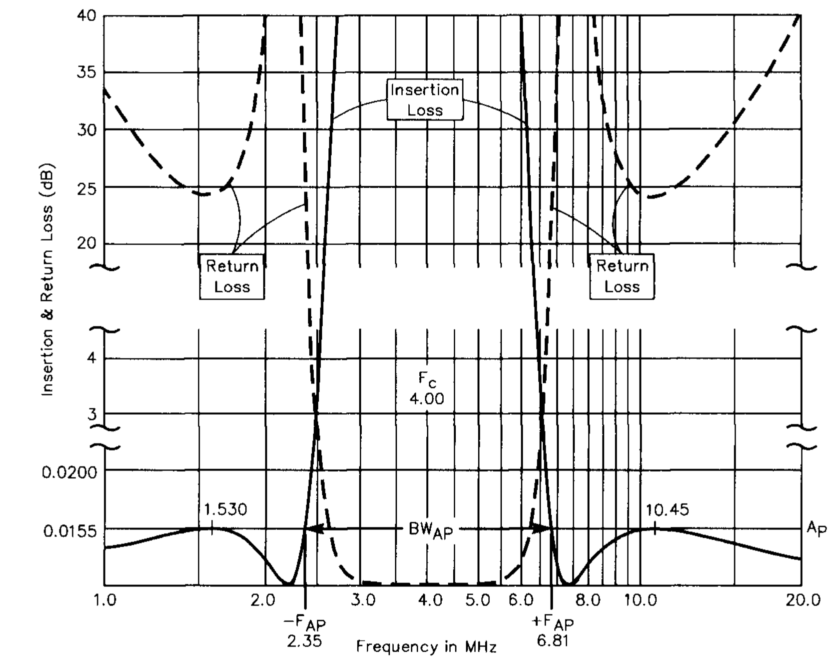


Fig 2—Typical insertion- and return-loss responses of a 3-resonator Chebyshev band-stop filter with the lower and upper ripple cut-off frequencies indicated by $-F_{AP}$ and $+F_{AP}$, respectively.

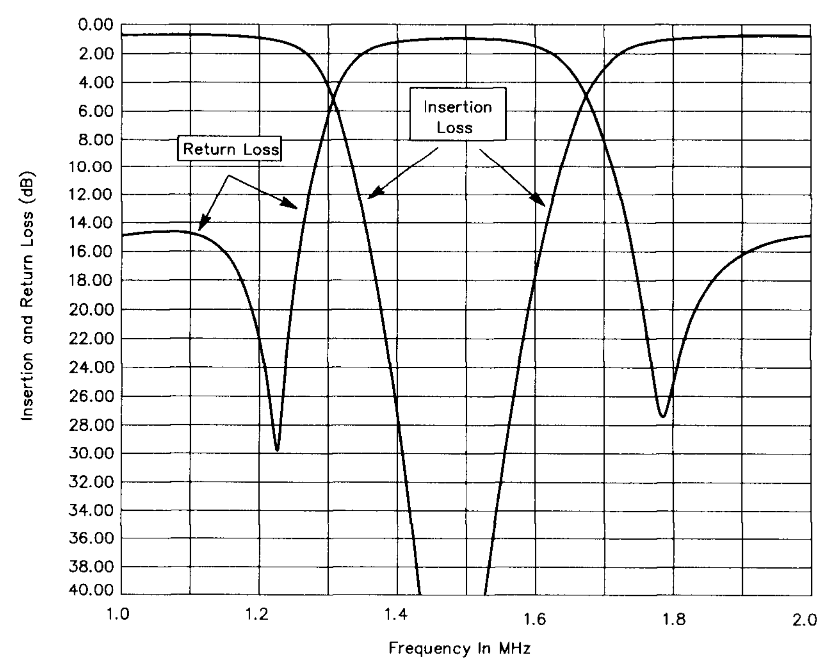


Fig 3—Computer-calculated insertion and return-loss responses of the 1.48-MHz band-stop filter of design #2 in Table 1. The inductor and capacitor Qs were specified as 78 and 500.

of 4 MHz. The significant points on the insertion-loss response curve are the upper and lower ripple cutoff frequencies ($+F_{Ap}$ and $-F_{Ap}$), the upper and lower 3-dB frequencies and the center frequency (F_c). Note that the upper and lower insertion-loss and return-loss responses intersect at the 3-dB level when the filter components are lossless. The ripple and 3-dB bandwidths are determined from their corresponding upper and lower ripple and 3-dB cutoff frequencies. The filter stop band (BW_{Ap}) is defined as existing between the upper and lower ripple cutoff frequencies on the response curve, and it is this ripple bandwidth that corresponds to the ripple cutoff frequency of the high-pass prototype from which the BSF is derived.

The lower and upper F_{Ap} frequencies are those frequencies where the insertion loss first exceeds the maximum passband ripple amplitude of A_p= 0.0155 dB. Also note that the frequencies of maximum passband ripple amplitude (at 1.53 and 10.45 MHz in Fig 2) are also the frequencies of minimum return loss, maximum RC and maximum VSWR.

The maximum ripple amplitude, A_p, is directly related to the filter RC and VSWR and inversely related to the return loss. Those designs having the lowest possible RC were preferred for the BSF tabulation.

Having selected the filter parameters of order, network configuration, impedance level and different pairs of standard-capacitor values, several Chebyshev BSF designs were computer calculated based on previously published Chebyshev design equations.[3] The designs were then compared, and the one having the lowest RC, and also having an upper ripple cutoff frequency ($+F_{Ap}$) less than 1.8 MHz, was selected for inclusion in a tabulation of designs ranging over a center frequency of 1.48 MHz to 0.56 MHz in decrements of 40 kHz.

Table 1—Computer-calculated designs for 50-Ω 3-resonator band-stop filters for center frequencies between 1.48 and 0.56 MHz in decrements of 40 kHz.

No	*Ctr Freq (MHz)*	BW(A_p) *(MHz)*	BW(3 dB) *(MHz)*	*Max* +F_{Ap} *(MHz)*	*Min Refl Coef (%)*	*Ret Loss (dB)*	*C1,3 (pF)*	*C2 (pF)*	*L1,3 (µH)*	*L2 (µH)*
1	1.48	0.4993	0.342	1.751	12.320	18.19	6800	820	1.701	14.103
2	1.48	0.4887	0.372	1.744	19.165	14.35	5600	820	2.065	14.103
3	1.44	0.6132	0.372	1.779	7.717	22.25	6800	1000	1.796	12.215
4	1.40	0.6708	0.434	1.775	9.839	20.14	5600	1200	2.308	10.770
5	1.36	0.6200	0.420	1.705	11.734	18.61	5600	1200	2.446	11.412
6	1.32	0.7938	0.462	0.775	6.614	23.59	5600	1500	2.596	9.692
7	1.28	0.8629	0.533	1.782	8.239	21.68	4700	1800	3.289	8.589
8	1.24	0.8878	0.479	1.761	5.012	26.00	5600	1800	2.942	9.152
9	1.20	0.9762	0.557	1.784	6.122	24.26	4700	2200	3.743	7.996
10	1.16	0.8681	0.534	1.673	8.123	21.81	4700	2200	4.005	8.557
11	1.12	0.8824	0.478	1.645	5.090	25.87	5600	2200	3.606	9.179
12	1.08	0.9655	0.555	1.666	6.291	24.03	4700	2700	4.621	8.043
13	1.04	1.0470	0.644	1.688	8.108	21.82	3900	3300	6.005	7.097
14	1.00	1.2262	0.680	1.786	5.525	25.15	3900	3900	6.495	6.495
15	0.96	1.0588	0.647	1.626	7.896	22.05	3900	3900	7.047	7.047
16	0.92	1.0600	0.571	1.592	4.986	26.05	4700	3900	6.367	7.674
17	0.88	0.8970	0.541	1.436	7.517	22.48	4700	3900	6.959	8.387
18	0.84	1.0706	0.573	1.531	4.861	26.26	4700	4700	7.638	7.638
19	0.80	0.8910	0.540	1.361	7.638	22.34	4700	4700	8.421	8.421
20	0.76	0.8531	0.473	1.298	5.539	25.13	5600	4700	7.831	9.331
21	0.72	0.7067	0.443	1.155	8.725	21.18	5600	4700	8.725	10.396
22	0.68	0.6654	0.383	1.090	6.336	23.96	6800	4700	8.056	11.655
23	0.64	0.6312	0.331	1.029	4.526	26.89	8200	4700	7.542	13.158
24	0.60	0.4954	0.306	0.897	8.209	21.71	8200	4700	8.581	14.970
25	0.56	0.4503	0.260	0.829	6.412	23.86	10000	4700	8.077	17.185

1. The BW(Ap) and BW(3 dB) column headings refer to the computer-calculated ripple and 3-dB bandwidths, which are based on perfect components. The $+F_{Ap}$ frequency is the calculated upper limit of the filter stopband and is designed to be less than 1.8 MHz.
2. For typical inductor and capacitor Qs, the actual bandwidths will be about 8% greater than the listed bandwidths. For example, the anticipated actual 3-dB bandwidth of design #2 is about 400 kHz for L and C Qs of 100 and 500, respectively.
3. All BSF designs were based on the listed standard-value capacitors and a return loss of preferably greater than 18 dB for a $+F_{Ap}$ frequency less than 1.8 MHz. As the center frequency decreased, it was possible to calculate designs having higher levels of return loss as compared to designs with center frequencies nearer 1.8 MHz.

Table 1 shows the 25 BSF designs that were selected. The BASIC computer program used to calculate and tabulate these designs was modified to calculate individual designs based on the center frequency and capacitor values specified by the user. The modified program appears in the "BASIC Program" sidebar.

Those BSFs with center frequencies near 1.8 MHz were designed with reflection coefficients greater than 12% to assure that their upper bandstop would not intrude into the lower edge of the 160-m band. For example, see designs 1 and 2 in Table 1 where the RC levels are relatively high, at 12.3 and 19.2%. Those BSF designs having center frequencies further removed from 1.8 MHz, such as designs 11 through 25, can have a wider stopband with a correspondingly lower RC.

To attenuate a high-level BC-band signal having a frequency between two of the listed frequencies in Table 1, use the capacitor values of the nearest listed design and tune the inductors to resonate the three capacitors to the frequency of the BC-band signal. The resulting BSF response will be similar to that of the nearest design. For example, for a desired center frequency F of 1.34 MHz, use the capacitor values of 5600 and 1200 pF or 5600 and 1500 pF and resonate the capacitors to 1.34 MHz with appropriate inductors. The resonating inductance is calculated from the equation: $L = 25330.3/(CF^2)$ where L, C and F are in µH, pF and MHz, respectively. The resulting design may be confirmed by using the BASIC program.

The G2/G1 ratio (RA in line 50 of the BASIC program) is equal to $39.4784 \times C1 \times C2 \times (F \times R)^2 \times 10^{-12}$, where C is in pF, F is in MHz and R is the system impedance level in ohms. For example, for C1=5600 pF and C2=1200 pF, F=1.34 MHz and R=50, the G2/G1 ratio is 1.19091. If the G2/G1 ratio is greater than 1.17 and less than 1.59, the corresponding reflection coefficient will be between 13% and 3.96%, respectively. (This chosen range of reflection coefficient is arbitrary. Similar values are equally satisfactory as long as the RC does not become too high, with a correspondingly high VSWR and greater sensitivity to component and termination impedance tolerance.)

For a center frequency of 1.34 MHz and C1=5600 pF and C2=1500 pF, the G2/G1 ratio is 1.4886 and the corresponding RC is 5.758%. Any BSF design within the 3.96 to 13% RC range will be satisfactory for a center frequency less than 1.4 MHz. For those frequencies above 1.4 MHz, which are closer to the lower edge of the 160-m band, BSFs with a higher reflection coefficient must be used in order to obtain a narrower bandwidth so any significant loss will not extend into the 160-m band. However, since the noise level in the 160-m band is usually high, a few dB of filter loss in the 160-m band will probably not be noticeable.

How to Confirm the Computer-Calculated BSF Designs

If you are using an unfamiliar computer-operated filter design program, you should always check at least one design by manual calculation to confirm the computer program is operating correctly. If just one design can be demonstrated as being correct, then probably all the computer-calculated designs are correct as the same program was used to calculate

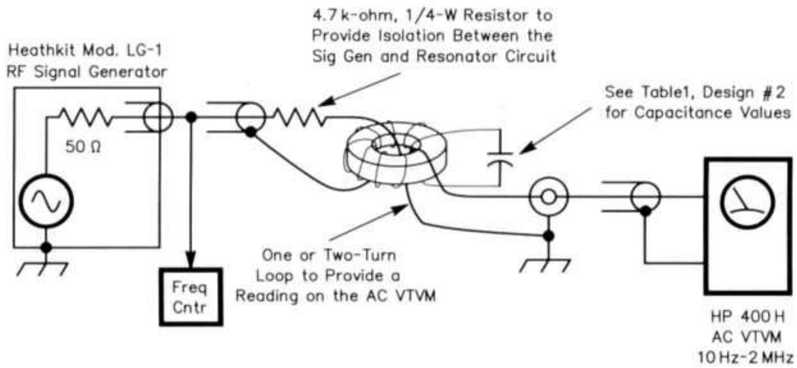


Fig 4—Equipment set-up and procedure used to tune each filter resonator to the center frequency of the band-stop filter.
Notes:
1. Turns are removed or added to the inductor core and then squeezed together or spread apart until the circuit resonates at the desired center frequency.
2. Resonance is indicated by a peak reading on the ac VTVM.
3. The LG-1 signal generator output across a 4.7-kΩ resistor is about 500 mV. The signal level coupled into the ac VTVM by the 1 or 2-turn loop is about 300 µV.

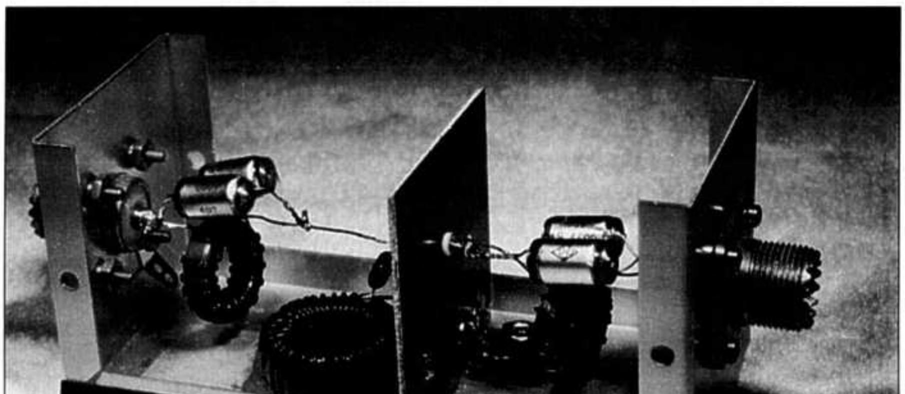


Fig 5—Photo of the assembled 1.48-MHz band-stop filter. Inductors L1 and L3 and capacitors C1, C2 and C3 are light enough to be supported by just their leads. The largest inductor, L2, is secured to a cardboard pad on the bottom of the aluminum box with RTV.

the rest of the designs.

The values of $BW(A_p)$ in Table 1 are listed so the reader can independently confirm each design. This can be done by referring to Table 2 ("The Normalized L and C Values"), where the normalized C and L values are listed for each design in Table 1. The normalized C and L values corresponding to the selected RC level are used to calculate the final C and L component values using the procedure explained in the *Handbook*.[4] For example, use the following procedure to confirm design #2, for F_c=1.48 MHz and RC=19.165%.

1. Refer to Table 2 and select the low-pass L-in/out configuration for transformation into a high-pass prototype. For the design with RC=19.165%, the corresponding normalized low-pass values of L1,3 and C2 are 1.16314 and 1.15467, respectively.

2. Transform the LP prototype into a high-pass by replacing the two inductors with capacitors and the one capacitor with an inductor. The normalized value of the two capacitors will now be the reciprocal of the former inductor value and the normalized inductor value will now be the reciprocal of the former capacitor value. Thus, C1,3=1/1.16314 =0.85974 F and L2=1/1.15467=0.86605 H.

3. Design a 3-element C-in/out high-pass prototype filter having a RC of 19.165%, an impedance of 50 Ω and a cutoff frequency equal to the ripple bandwidth (0.4887 MHz) of the BSF. The 0.4887-MHz bandwidth was obtained from the $BW(A_p)$ tabulation for design #2 in Table 1.

(a) Calculate the inductance and capacitance scaling factors, L_s and C_s:

$$L_s = \frac{R}{2\pi \cdot BW_{Ap}} = \frac{50}{2\pi \cdot 0.4887 \times 10^6} = 16.2835 \times 10^{-6}$$

$$C_s = \frac{1}{R \cdot 2\pi \cdot BW_{Ap}} = \frac{1}{50 \cdot 2\pi \cdot 0.4887 \times 10^6} = 6513.4 \times 10^{-12}$$

(b) Using the inductance and capacitance scaling factors, calculate L2 and C1 and C3 of the high-pass filter prototype (see Fig 1) by multiplying the normalized *L* and *C* values from step 2 by the L_s and C_s scaling factors:

$$\begin{aligned} \mathrm{L2} &= \mathrm{L_s} \times 0.86605\ \mathrm{H} \\ &= 16.2835 \times 10^{-6} \times 0.86605 \\ &= 14.10\ \mu\mathrm{H} \\ \mathrm{C1,3} &= \mathrm{C_s} \times 0.85974\ \mathrm{F} \\ &= 6513.4 \times 10^{-12} \times 0.85974 \\ &= 5600\ \mathrm{pF} \end{aligned}$$

These manually calculated values of C1,3 and L2 confirm the values given in Table 1 for design #2. Next, convert the 3-element high-pass prototype into a BSF design by resonating each C and L at the center frequency of 1.48 MHz:

4. Use the equations:
$C=25330.3/(F^2L)$ and $L=25330.3/(F^2C)$ to find C2 and L1,3 where *F* is the BSF center frequency in MHz and *C* and *L* are in pF and μH.

(a) Calculate C2=25330.3/(1.48^2× 14.10)=820 pF

(b) Calculate L1,3=25330.3/(1.48^2× 5600)=2.065 μH

These manually calculated values of C2 and L1,3 also confirm the computer-calculated values listed in Table 1 for design #2. The other computer-calculated designs may therefore be viewed with confidence because the same BASIC program used to calculate design #2 was also used to calculate the other designs.

Further confirmation of BSF design #2 was obtained using the *ELSIE Filter Design and Analysis* software.[5] The plot of insertion loss and return loss over the 1 to 2-MHz frequency range is shown in Fig 3. (Note: This computer-generated plot is upside down from the plot of Fig 2 because the programmer of the filter-analysis software prefers this orientation. In contrast, I prefer to show increasing values of attenuation going up.)

Fig 3 shows that the Trinity Software calculated minimum return-loss levels at 1.08 and 2.00 MHz are in agreement with the minimum return loss value (14.35 dB) of design #2 listed in Table 1. In particular, note that the calculated return loss over the 160-m band (1.8-2.0 MHz) is greater than 14.35 dB. This level of return loss indicates that only a small amount of the passband signal is reflected at the filter input, and most of the signal power over the filter passband goes into the filter and receiver.

If the filter of design #2 is lossless and the receiver has a perfect 50-Ω input impedance, the percentage of signal power going through the filter and into the receiver at the frequency of maximum VSWR can be approximated by squaring the reflection coefficient of the filter at this frequency (0.19165^2=0.03673), changing it to percent (3.673%), and then subtracting it from 100 (100−3.673) to find that about 96% of the input power is getting to the receiver. In comparison, at the center of the BSF filter at 1.48 MHz, the return loss is less than 1 dB, indicating that virtually all of the signal power is reflected, with virtually none going into the receiver.

The Importance of Insertion and Return Loss

The design and evaluation of any filter is not complete until both the insertion loss and return loss are evaluated. Usually in amateur-radio applications, only the filter insertion loss is plotted; the equally important return-loss measurement is ignored. In the past, I have been guilty of this omission, and it is only in the last few years that I have come to appreciate the importance of having both the insertion- and return-loss plots of a filter to provide confir-

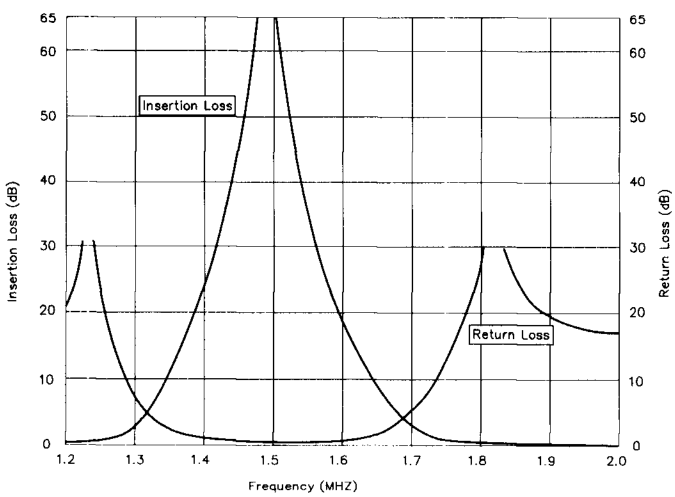


Fig 6—Measured insertion- and return-loss responses of the 1.48-MHz band-stop filter of design #2 in Table 1. The measured L1,3 and L2 inductor Qs are 110 and 160, respectively.

mation that the filter was properly designed and assembled. For example, it is possible for improperly calculated component values to produce what appears to be a satisfactory BSF insertion-loss response where the typical attenuation peak is observed. However, if the corresponding return-loss response does not have a sufficiently high level in the filter passband (preferably greater than 14 dB), it is likely that the design or assembly was incorrectly performed and unnecessary passband signal loss will occur. As you can see from Table 1, almost all of the BSF designs have minimum return-loss levels greater than 18 dB.

Filter return loss is not difficult to measure. The *Handbook* provides circuits for both audio-frequency and radio-frequency return loss bridges (RLB); for example, see pages 26.38 and 26.39, where a 100-Hz to 100-kHz RLB and an RF RLB are described.[6]

BASIC Program

This BASIC program calculates and prints to screen the parameters and component values of a 50-Ω, 3-resonator BC-band reject filter based on the center frequency and C1 and C2 capacitor values given by the user. Designs are restricted to those having reflection coefficients between 3.8 and 27.2%. This program can be downloaded from the ARRL BBS (860-594-0306) or via the Internet from http://www.arrl.org/qexfiles or ftp://ftp.arrl.org/pub/qex. The file name is qexbsf.zip.

```
10 REM FILE NAME: 'QEXBSF.ASC' FOR USE IN 11/95 QEX BSF ARTICLE. 01 NOV 95.
20 REM Program writes RC%, RET LOSS & OTHER PARAMETERS OF 3rd-ORDER BSF.
30 PI=4*ATN(1) : Z=50 : N=3          : REM  Z = SOURCE & LOAD IMPEDANCE IN OHMS.
40 PRINT: INPUT "ENTER FC, C1 & C2 in MHz and pF";FC,C1,C2 : PRINT
50 RA = C1*C2*((2*PI*FC*Z)^2)*1E-12 : REM RA=LOWPASS G2/G1 RATIO FOR RC CALCS.
60 IF RA>1.6 GOTO 370 : IF RA<.8   GOTO 370 : REM  OMITS OUT-OF-RANGE RATIOS.
70 FOR C=1 TO 4        : REM START OF CALCS TO FIND %R.C. FOR CALC'D RA.
80 A(C) = SIN(.5*PI*(2*C-1)/N) : B(C)=SIN(PI*C/N)^2 : NEXT C
90 K2 =A(2)/(RA*A(1))            : REM  RA IS LOWPASS G2/G1 RATIO FOR %RC CALC.
100  M =SQR(B(1)/(K2-1))         : X =M+SQR((M^2)+1)
110 AP=8.68589*LOG(((X^(2*N))+1)/(((X^(2*N))-1))) : REM AP = Ap(dB)
120 R1=100*SQR(1-(.1^(.1*AP)))         : REM  REFLECTION COEFF (%)
130 RC = R1/100 : VS = (1+RC)/(1-RC) : REM VS = VSWR BASED ON RC (REFL COEFF).
140 RL = -20*(1/LOG(10))*LOG(R1/100) : REM RL = RETURN LOSS IN dB.
150 REM  START OF CALCS TO FIND G1 AND G2 BASED ON PREVIOUSLY CALC'D R.C.
160 A = AP/17.3718 : B=LOG((EXP(A)+EXP(-A))/(EXP(A)-EXP(-A)))
170 D = (EXP(B/(2*N))-EXP(-B/(2*N)))/2
180 FOR K=1 TO N   : A(K)=SIN((((2*K)-1)*PI)/(2*N))
190 B(K) = ((D)^2)+(1-COS((2*K*PI)/N))/2 : NEXT K
200 G(1) = 2*A(1)/D                : REM G(1) = NORMALIZED VALUE OF FIRST ELEMENT
210 FOR K=2 TO N : G(K) = 4*A(K-1)*A(K)/(B(K-1)*G(K-1)) : NEXT K
220 G1 =G(1) : G2 =G(2)  : REM  LP G-VALUES (USE RECIPROCALS FOR HP G-VALUES)
230 L2 = 25330.3/(FC^2*C2)         : REM BASED ON GIVEN FC & C2.
240 L1 = 25330.3/(FC^2*C1)         : REM BASED ON GIVEN FC & C1.
250 E=SQR(RC^2/(1-(RC^2)))         : REM RIPPLE FACTOR 'E' BASED ON RC.
260 V=1/N*LOG(1/E+SQR(1/E^2-1)) : F3=(EXP(V)+EXP(-V))/2 : REM F3=F3/FAp RATIO
270 BW =1000000!/(C1*G1*Z*2*PI) : B3=BW/F3 : REM BASED ON C1 AND CALC'D G1.
280 BL =SQR(FC^2 + (BW^2)/4) -BW/2   : REM CALCS LOWER BAND EDGE FOR FC & BW.
290 BU =BL + BW        : REM CALCS UPPER BAND EDGE BASED ON BL AND BW
300 PRINT "F-CENTR BW(Ap) BW(3dB) + FAp  R.C.  R.L.  C1,3  C2   L1,3   L2   LP
G2/G1"
310 PRINT " (MHz)  (MHz)   (MHz)  (MHz)   (%)   (dB)  (pF)  (pF)   (uH)   (uH)
RATIO"
320 PRINT USING " #.## ";FC;  : PRINT USING "  #.####";BW; : PRINT USING "   #.###
";B3;
330 PRINT USING " #.### ";BU; : PRINT USING " ##.##";R1;    : PRINT USING "
##.#";RL;
340 PRINT USING " ##### ";C1; : PRINT USING " ####";C2;
350 PRINT USING "   ##.## ";L1;: PRINT USING " ##.##";L2;  : PRINT USING "
#.#####";G2/G1
360 PRINT : PRINT "     END OF RUN." : END
370 PRINT "RA=";RA; : PRINT "    G2/G1 RATIO (RA) MUST BE BETWEEN .8 & 1.6."
380 PRINT "                 TRY AGAIN WITH DIFFERENT CAP VALUES." : GOTO 40
```

Example of program output to screen for FC, C1 and C2 = 1.48, 5600 and 820.

F-CENTR	BW(Ap)	BW(3dB)	+FAp	RC	RL	C1,3	C2	L1,3	L2	LP G2/G1
(MHz)	(MHz)	(MHz)	(MHz)	(%)	(dB)	(pF)	(pF)	(µH)	(µH)	RATIO
1.48	0.4887	0.372	1.744	19.16	14.3	5600	820	2.07	14.10	0.99272

Procedures for measuring both the RF insertion loss and return loss of a filter are explained in "Insertion-Loss and Return-Loss Measurement Procedures."

Filter Construction

For attenuating the 1.48-MHz signal of the nearby broadcast-band station, design #2 of Table 1 was selected for assembly. Design #1 would have been equally suitable, as the parameters of both these designs are satisfactory. In particular, note that the ratio of the larger to smaller inductance and capacitance values is less than ten to one, and the reactances of L1,3 and L2 are about 19 and 131 Ω at 1.48 MHz. As a general rule, it is desirable to minimize the component value spread and also to keep the inductor and capacitor reactance values at resonance between 5 and 500 Ω in a 50-Ω system. When the inductor or capacitor reactance falls below 5 Ω, either the equivalent series resistance of the inductor, or the series inductance of the capacitor, becomes significant and degrades the circuit Q. And when the inductive reactance becomes greater than 500 Ω, the inductor is very likely approaching self-resonance and the circuit Q is again degraded.

Capacitors C1 and C3 were each realized by paralleling 2700 and 3000-pF polystyrene capacitors that were measured with a digital capacitance meter and selected from a group of capacitors to get within ½ percent of the 5600-pF design value. Single capacitors of the desired 5600-pF value were not available at the time, otherwise they would have been used. Capacitor C2 was selected from a group of 820-pF polypropylene capacitors that were on hand. Both capacitor types have a low dissipation factor (0.1% or less) and are appropriate for low-power RF filtering applications such as this.

Inductors L1 and L3 were wound on Micrometals T80-17 (blue/yellow) powdered-iron cores, using between 18 and 20 turns of #22 magnet wire. The A_L of this core is 2.2 nH/N^2. The −17 mix was developed by Micrometals as a temperature-stable alternative to the older −12 mix, and this newer mix is recommended for all new designs. Although this −17 mix is suggested for use in the 20-200 MHz range, it is preferred for this lower-frequency application instead of the higher permeability −7 mix. The greater number of turns required by the −17 mix as compared to the −7 mix permits a finer adjustment of tuning while still providing excellent Q at 1.48 MHz. L2 was wound with 32 turns of #20 magnet wire on a Micrometals T106-7 (white) core. The A_L for this core is 13.3 nH/N^2. The larger inductance of L2 requires a larger core with a higher

Table 2—Normalized L and C Values

Element values of 3rd-order Chebyshev low-pass filters in Fig A normalized for a ripple cutoff frequency (F_{-Ap}) of one radian/sec and 1-Ω terminations. Use the top column headings for the low-pass C-in/out configuration and the bottom column headings for the low-pass L-in/out configuration.

No	*RC (%)*	*F3/F-Ap Ratio*	*C1, 3 (F)*	*L2 (H)*	*G2/G1 RATIO*
1	12.320	1.4594	0.937474	1.13007	1.20544
2	19.165	1.3133	1.16314	1.15467	0.99272
3	7.717	1.6467	0.763425	1.06244	1.39168
4	9.839	1.5449	0.847377	1.10154	1.29994
5	11.734	1.4773	0.916784	1.12463	1.22672
6	6.614	0.7167	0.716017	1.03433	1.44455
7	8.239	1.6182	0.784846	1.07367	1.36800
8	5.012	1.8539	0.640239	0.979386	1.52972
9	6.122	1.7534	0.693752	1.01950	1.46954
10	8.123	1.6243	0.780137	1.07128	1.37319
11	5.090	1.8459	0.644181	0.982562	1.52529
12	6.291	1.7404	0.701485	1.02477	1.46086
13	8.108	1.6251	0.779530	1.07097	1.37386
14	5.525	1.8040	0.665605	0.999203	1.50119
15	7.896	1.6366	0.770836	1.06643	1.38347
16	4.985	1.8568	0.638883	0.978285	1.53124
17	7.517	1.6583	0.755053	1.05781	1.40097
18	4.861	1.8700	0.760127	1.06063	1.39534
20	5.539	1.8027	0.666290	0.999717	1.50042
21	7.873	1.6379	0.769891	1.06593	1.38452
22	6.336	1.7370	0.703536	1.02614	1.45855
23	4.526	1.9083	0.615023	0.958211	1.55801
24	8.209	1.6198	0.783635	1.07306	1.36933
25	6.412	1.7313	0.706981	1.02844	1.45469
26	4.796	1.8772	0.629196	0.970296	1.54212
No	*RC (%)*	*F3/F-Ap Ratio*	*L1, 3 (H)*	*C2 (F)*	*G2/G1 RATIO*

Notes:

1. The normalized values of nos 1 to 25 were used in calculating the twenty-five band-stop filters listed in Table 1. The no. 26 values are included to allow them to be confirmed by comparing them with the published values in Table 16.2 of *The 1996 ARRL Handbook* for N=3.
2. The G2/G1 ratios were obtained by dividing L2 by C1.

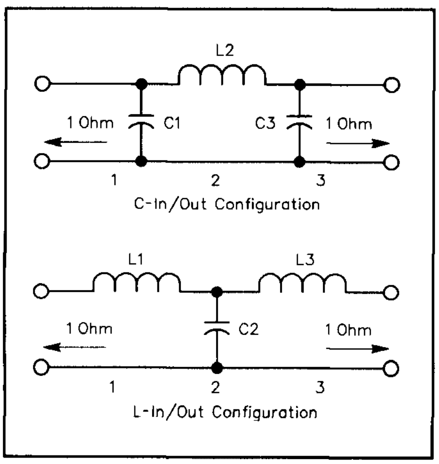


Fig A. Schematic diagrams of 3rd-order Chebyshev low-pass filters.

A_L than that used for L1 and L3.

The three resonator circuits were individually tuned to 1.48 MHz using the test circuit and procedure described in Fig 4. Tuning was accomplished by first removing or adding turns to the inductors and then by squeezing or spreading the turns so resonance occurred at exactly 1.48 MHz, as indicated by a digital frequency meter. Resonance was indicated when a peak response was obtained on the ac VTVM. The Qs for L1,3 and L2 were measured in the Fig 4 circuit as 110 and 160, respectively. The Qs were calculated by dividing 1.48 MHz by the 3-dB bandwidth of the tuned circuit. These Q values appear to be on the low side, which is probably due to the loading effect caused by the 50-Ω signal-generator impedance plus the 4.7-kΩ resistor and the ac VTVM.

After all resonators were tuned, they were installed in a 2¼x2¼x5-inch aluminum minibox in accordance with the schematic diagram of Fig 1. A 1⅞-inch square PC board was placed in the approximate center of the box, and the foil was grounded to provide shielding between the input and output resonators. A small Teflon feedthrough on the PC board was used to provide a connection between the input and output resonator circuits. UHF-type RF connectors provide input and output connections to the filter. Fig 5 is a photograph of the completed filter.

Filter Testing

The completed band-stop filter was tested for insertion and return loss over the 1.2 to 2.0-MHz frequency range using the test procedures explained in "Insertion-Loss and Return-Loss Measurement Procedures." Fig 6 shows the measured insertion- and return-loss responses of the filter in a 50-Ω test system. The insertion-loss response shows a loss of more than

Insertion-Loss and Return-Loss Measurement Procedures

Insertion Loss

Measurement of insertion loss can be performed using the test set-up of Fig B, as follows:

1. For a given frequency and with both coax switches set to the FLTR position, adjust the signal generator amplitude and tune the receiver until the signal is received as indicated by the deflection of the receiver S-meter.
2. Optimize the receiver controls for a maximum S-meter deflection and adjust the signal generator level until a suitable reference indication is obtained on the S-meter. For example, the +20 dB-over-S9 scale marker can serve as a convenient reference indication.
3. Change the coax switches to the ATTN position and add or remove attenuation until the same S-meter deflection is obtained as with the filter. Do not change any receiver control settings during this part of the measurement procedure.
4. Record the dB setting on the step attenuator and the corresponding frequency. When it is not possible to obtain the exact same reference level indication on the receiver S-meter scale because a 1-dB step change is either too large or too small, take the lower dB setting and add ½ dB to obtain the closest estimation of the attenuation.
5. Change the signal generator to the next test frequency and repeat steps 1 to 4.

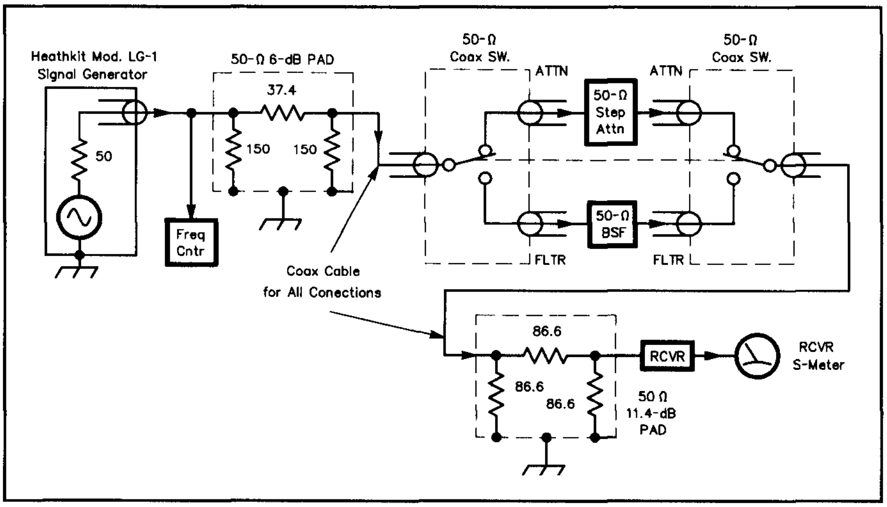


Fig B—Test set-up used to measure the insertion loss of the band-stop filter in a 50-Ω test system.

Notes:

1. All coaxial cables are 50-Ω type RG-58/U.
2. All resistors are ¼-W, 1% metal film.
3. The 50-Ω pads are used to stabilize the impedance level.
4. See the *1995 Handbook* for a description of the 50-Ω step attenuator.†

65 dB at 1.48 MHz. The single-peak response is the result of carefully tuning each resonator to 1.48 MHz.

The return loss over the 160-m band is also satisfactory, staying above 20 dB from 1.8 to 1.89 MHz and above 17 dB up to 2.0 MHz. Although the minimum return loss was calculated to be 14.35 dB, the measured value is about 2.6 dB greater than the calculated value. The reason for this difference is not known. The measured return loss over the filter stopband (from about 1.3 to 1.7 MHz) is very low, as expected. Below 1.3 MHz, the return loss again rises to another peak similar to that in the upper frequency range. The return loss above 2.0 MHz was not measured, but a measurement of input impedance versus frequency using a Hewlett-Packard Model 4193A vector impedance meter showed that when the filter was terminated in a 50-Ω load, its input impedance varied less than 2 Ω relative to 54 Ω over the 3.5 to 30-MHz range. The corresponding phase angle gradually increased from –7 degrees at 3.5 MHz to +10 degrees at 30 MHz. This means that the BSF can be left connected to the receiver over the 160 to 10-m band range with no significant effect on the receiver performance.

The satisfactory measured insertion- and return-loss responses indicate the filter was properly designed and assembled and should perform as expected when inserted between the antenna and receiver. An earlier version of the filter, shown in Fig 5, was sent to W5DM, and he reported that his receiver overload problem was eliminated after the filter was installed.

Summary

Paul Elliott, W5DM, was experiencing a receiver overload problem due to a nearby high-power broadcast-band transmitter at 1.48 MHz. The overload problem was solved by inserting a spe-

Return Loss

Use the set-up of Fig C to measure return loss:

1. Switch the RLB "?Z" port to the input of the terminated filter with the step attenuator set to 0 dB.
2. Set the signal generator to the desired frequency and tune the receiver to pick up the signal.
3. Adjust the signal generator level and optimize the receiver controls until a reference level of +20 dB over S9 is obtained on the S-meter.
4. Switch the RLB "?Z" port to the open-circuit position and note that the S-meter indication will increase if the terminated filter has some return loss.
5. Insert attenuation with the step attenuator until the S-meter returns to the previous reference level. Do not change any receiver control settings during this part of the measurement procedure.
6. Read the return loss from the step attenuator and record it with the frequency. When it is not possible to obtain the exact same reference level indication on the receiver S-meter scale because a 1-dB step change is either too large or too small, take the lower dB setting and add ½ dB to obtain the closest estimation of the return loss.
7. Change the signal generator to the next test frequency and repeat steps 1 to 6.

Note: To check the accuracy of the return-loss measurement procedure, replace the terminated filter first with a grounded 75-Ω resistor, then with a grounded 100-Ω resistor. The corresponding measured return-loss levels should be 14 and 9.5 dB, respectively.

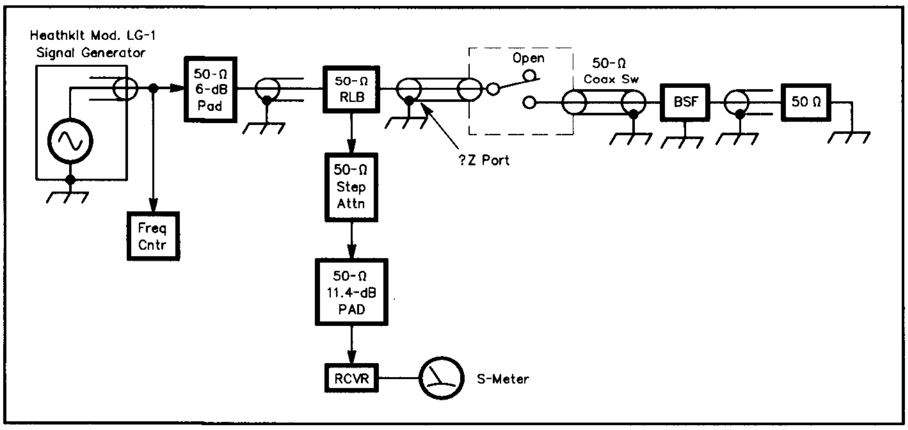


Fig C—Test set-up used to measure the return loss of the band-stop filter when it is terminated in 50 Ω.

Notes:

1. All coaxial cables are 50-Ω type RG58/U.
2. The 50-Ω pads are used to stabilize the impedance level.
3. See Fig B for the 50-Ω pad resistor values.
4. See the *1995 Handbook* for a description of the RF return loss bridge.‡
5. See the *1995 Handbook* for a description of the 50-Ω step attenuator.†

† *The 1995 ARRL Handbook*, (Newington: ARRL) "Low-Power Step Attenuators," p 26.40-41.
‡ *The 1996 ARRL Handbook*, (Newington: ARRL) "An RF Return-Loss Bridge," p 26.39, Fig C.

cially designed 3-resonator band-stop filter in front of his receiver. The design and analysis of the filter was accomplished using a DOS-based filter design and analysis program called *Elsie* from Trinity Software. The filter was assembled and tested for both insertion-loss and return-loss in a 50-Ω system, and the measured responses were in good agreement with the computer-calculated values.

Because of the usefulness of this particular design, a number of similar designs covering the 0.56 to 1.48-MHz broadcast-band range were also calculated along with their associated design parameters and component values. A 38-line BASIC program is included so the interested reader can calculate other 3-resonator band-stop designs based on the exact capacitor values on hand. A method was explained how the reader could confirm the correctness of the computer-calculated filter designs, and procedures were explained for making insertion-loss and return-loss measurements on the completed filters.

The amateur now has complete and convenient band-stop filter-design information that previously was not available. This design information can be applied in the design of any 3-resonator band-stop filter for center frequencies less than 30 MHz and where lumped L-C components are feasible.

Acknowledgments

I gratefully acknowledge the assistance of Wes Hayward, W7ZOI, Rex Cox and Heyward Preacher for their review and comments on this article.

Notes

[1] *The 1996 ARRL Handbook*, (Newington: ARRL) "A BC-Band Energy-Rejection Filter," p 16.36, Fig 16.68.

[2] "The Doctor is IN" column, *QST*, July 1995, p 56.

[3] Wetherhold, Edward E., W3NQN, "Calculate 5- and 7-Element Filter Components," *QEX*, May 1987, p 4.

[4] *The 1996 ARRL Handbook*, (Newington: ARRL) "Band-Pass Filters," pp 16.15-16.17.

[5] Trinity Software, 3526 Highway 66, Suite B-174, Rowlett, TX 75088.

[6] *The 1996 ARRL Handbook*, (Newington: ARRL) "Return Loss Bridges," pp 26.38 and 26.39.

[7] See the Micrometals catalog #3, Issue D. Contact Micrometals at 1190 Hawk Circle, Anaheim CA 92807; tel: 800-356-5977 or 714-630-7420.

Upcoming Technical Conferences

The Central States VHF Society Conference

The 30th Annual Conference of the Central States VHF Society will be held July 26-28, 1996 at the Thunderbird Hotel & Convention Center in Bloomington, Minnesota. The Thunderbird is adjacent to the Minneapolis-St Paul International Airport and the Mall of America.

The program will feature technical presentations, antenna gain measurements, noise figure testing, a flea market and the premier opportunity to meet VHFers from across North America and beyond.

We will also have a full family program, with both organized group activities and suggestions for young families who may prefer to do things on their own. We are within walking distance of the fabulous Mall of America, with entertainment and shopping to interest almost anyone.

In response to my survey at Colorado Springs, we will have the use of a large hospitality suite for the family program and we will offer babysitting services. The hotel also has beautiful indoor and outdoor pools and a unique Native American theme.

The Thunderbird is ready to take reservations at 800-328-1931. We have a large block of rooms at a special rate of $79 (plus tax) until July 1, 1996. Be aware that, because of its proximity to the Mall, the hotel will be fully booked in advance for these nights. Please make your reservations early!

For those interested in extending their vacation in Minnesota, the Office of Tourism has a very nice Web page at: http://tccn.com/mn.tourism/mnhome.html.

Resort bookings should also be made during the winter to avoid disappointment.

More information will be posted as it becomes available. The Northern Lights Radio Society and I look forward to greeting many of you next summer!—*Paul Husby WØUC, 1462 Midway Parkway, St. Paul, MN 55108. We also now have a Web page up for the Conference, at: http://www.umn.edu/nlhome/m042/liebe009/#CONTEST.*

Feedback

There were a couple of pin-out errors in the schematics of my August 1995 *QEX* article, "A Simple Frequency Deviation Meter Using Time-Domain Techniques." In Fig 2, the rail polarities should be reversed as shown on U5D, and in Fig 3, the negative rail should go to pin 11 (not pin 8) as shown on U6B. Sorry they weren't caught before publication.—*Cary T. Isley, Jr., W7KIM*

Some serious technical inaccuracies appeared in the May 1995 *QEX* article, "Graphic Method for Calculating Z" (which was reprinted from the February 1995 issue of *Radio Communication*).

1. The equation on page 15, column 1, giving the values of the normalized resistance and reactance corresponding to the center of the SWR circle should read:

$$R_0 = \frac{S + \frac{1}{S}}{2} \text{ and } X_{max} = \frac{S - \frac{1}{S}}{2}$$

2. In the caption to the diagram contained in "Why the Impedance Varies Along a Transmission Line," the portion of the reflected power should be p^2 and not simply *p* since this power is proportional to the product of the magnet and electric components of the reflected wave.

3. At the end of "Why the Impedance Varies Along a Transmission Line," the equation for the reactive component of the input impedance should read:

$$X_i = \frac{(X_t + T)(1 - X_t T)^2 - TR_t^2}{(1 - X_t T)^2 + (TR_t)^2}$$

This is of particular importance for people trying to turn the equation into a BASIC program or use them in a spreadsheet—a technique I find extremely useful for solving transmission line problems.

It might also be worth noting, since it has come up in correspondence, that to refer the impedance forward from the measurement point to a point nearer the aerial, this last formula (number 3) applies, but *T* is given a negative value.—*A. J. Harwood, G4HHZ*

Homebrewing Black-Box Style

Take advantage of tried-and-true technology at a higher level. Kenwood's TM-741/742 FM-band modules make a great start for a home-brew FM transceiver project.

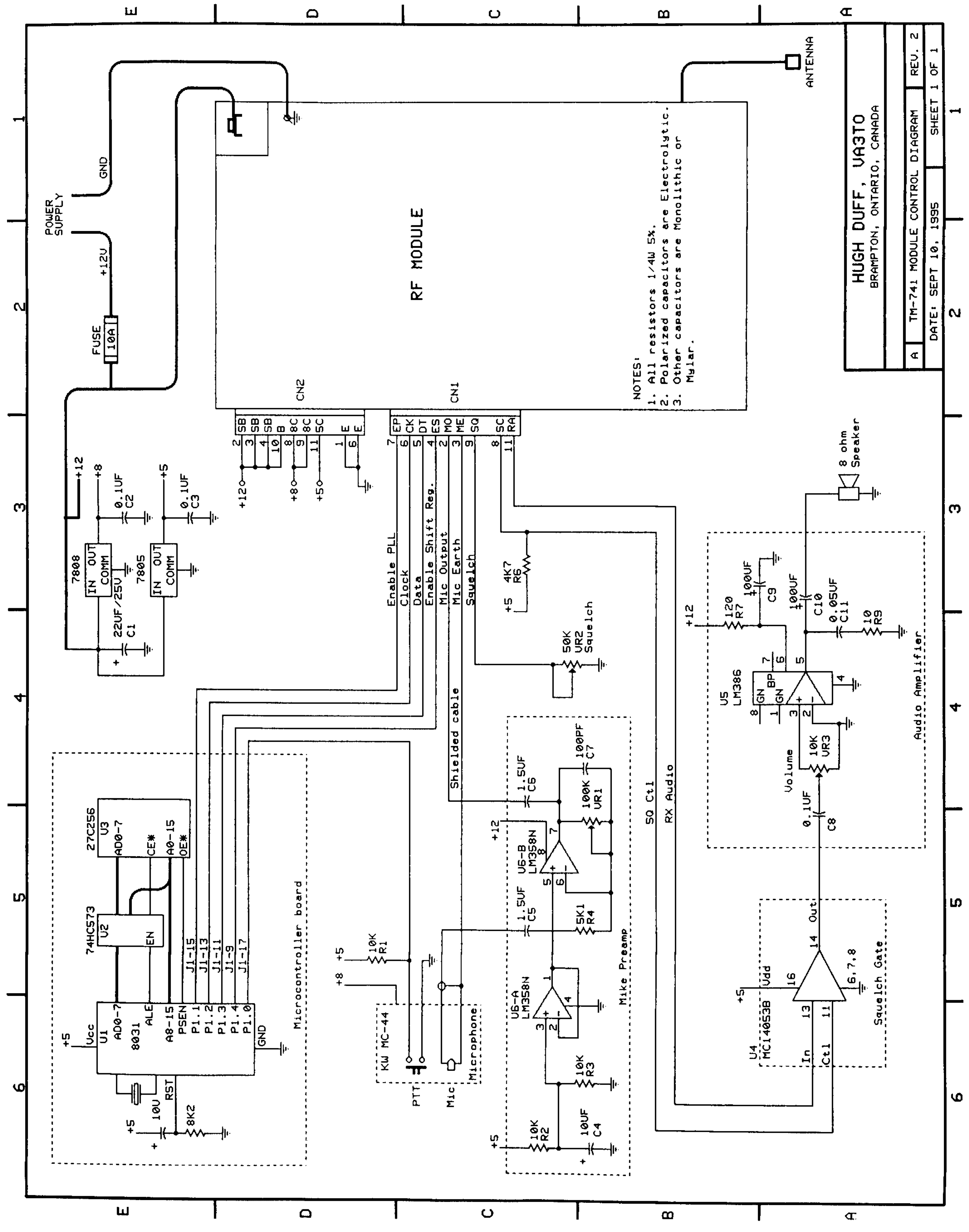
By Hugh Duff, VA3TO (ex-VE3OYH)

Designing and building a stable, broad-banded, digitally tuned transceiver can be quite a challenge. It takes months for teams of engineers to come up with a working prototype of a modern commercial transceiver. The days of homebrewing a rig to match the reliability and performance of these rigs are closing. Not many hams have the RF and digital expertise or equipment to accomplish this on their own. However, you can take advantage of the proven technology of readily available components by using them as building blocks. For example, the optional RF-band modules used in some multiband transceivers can be put to use in a home-brew transceiver. Harnessing the capabilities of these RF black-boxes still presents us with a challenge. The difficulty lies in providing the necessary input signals and "gluing" the additional hardware together to form a working transceiver. This article describes the findings of my experimentation with such a project and will hopefully give others a helpful head start.

Kenwood's innovative TM-741 and TM-742 modular triband FM transcievers are fine examples of modern RF-meets-digital technology in the world of Amateur Radio. Sold as dual-band 2 m/70 cm radios with space for an optional third band, these units have four other band modules available: 28 MHz, 50 MHz, 220 MHz and 1.2 GHz. Any combination of three band modules can be installed at one time. The modules perform all of the RF functions. All other functions, including power regulation and distribution, frequency control and display and audio amplification and squelch gating are performed by the main chassis. For anyone willing to do some experimenting, one of these 4×6×0.5-inch modules can be assembled into a monoband FM transceiver. With the proper considerations, a nice repeater could be assembled using two of these modules. Used modules often turn up at flea markets and on swap-net listings at pretty reasonable prices. I picked up the 2-m module from a friend who replaced it with another module in his TM-741. Just for fun, I decided to see if I could control it independently of the main chassis. By studying the service manual and schematics, and with the aid of a logic analyzer, I was able to construct a test platform that performs some of the basic functions required to get the module on the air.[1]

136 Baronwood Court
Brampton, ON L6V 3H8 Canada
e-mail: hduff@humnet.humberc.on.ca

[1]Notes appear on page17.



HUGH DUFF, VA3TO
BRAMPTON, ONTARIO, CANADA
TM-741 MODULE CONTROL DIAGRAM
REV. 2
DATE: SEPT 10, 1995
SHEET 1 OF 1
RF MODULE
NOTES:
1. All resistors 1/4W 5%.
2. Polarized capacitors are Electrolytic.
3. Other capacitors are Monolithic or Mylar.
POWER SUPPLY
GND
+12V
FUSE 10A
ANTENNA
CN2
CN1
7808
7805
22UF/25V C1
0.1UF C2
0.1UF C3
Enable PLL
Clock
Data
Enable Shift Reg.
Mic Output
Mic Earth
Squelch
Shielded cable
4K7 R6
50K VR2 Squelch
SQ Ctl
RX Audio
U5 LM386
Audio Amplifier
8 ohm Speaker
120 R7
100UF C9
100UF C10
0.05UF C11
10 R9
10K VR3 Volume
0.1UF C8
U4 MC14053B
Squelch Gate
Mike Preamp
U6-A LM358N
U6-B LM358N
1.5UF C6
100PF C7
100K VR1
1.5UF C5
5K1 R4
10K R3
10K R2
10UF C4
10K R1
KW MC-44
PTT
Mic
Microphone
Microcontroller board
27C256 U3
74HC573 U2
U1 8031
10U
8K2
RST

Fig 1—(left) Interconnections and additional circuity used to interface to and control a Kenwood 2-m RF module.

The following information describes the basic hardware and programming required to use one of these modules as a stand-alone FM transceiver.

The Heart and Soul

The nucleus of the TM-741 and TM-742 main chassis is the microcontoller. It is the microcontroller that offers all of the bells and whistles found in most modern transceivers. The Kenwood RF module uses a serial shift register to control its power-level adjustment and the TX/RX switching. It also uses a serially programmed PLL unit for frequency control. The main chassis sends digital data to the module to control these functions. On my test platform, the RF module is controlled by an L. S. Electronics ELC-31, an 8031-based microcontroller.[2] An on-board EPROM was programmed to send the correct data to the module.

Four parallel I/O lines are connected to the enable shift register (ES), clock (CK), data (DT) and enable PLL (EP) lines on connector CN-1 of the module. These lines are toggled appropriately by firmware, as shown in "Programming Word Description." To send data to the shift register, 8 bits are clocked in using CK and DT, one bit at a time. Then the ES line is toggled to latch the new data into the output stage of the shift register. The PLL requires two 21-bit words. The *reference* word sets up the step size, and the *comparison* word controls the frequency. To program the PLL, the EP line must be enabled, then the first 21-bit word is clocked in one bit at a time. The EP line must be disabled and enabled again before the second 21-bit word is clocked in. Then the EP line must be disabled to terminate the transfer. The data must be sent from the microcontroller to the RF module at start-up and any time a function is altered, such as a change in frequency, or whenever PTT is pressed or released. Note the logic levels of the programming lines at rest in "Programming Word Description." The 8031 firmware (Listing 1) is well commented and may help to explain the method of programming the module. It can be downloaded from the ARRL BBS (860-594-0306) or via the Internet at http://www.arrl.org/qexfiles or ftp://ftp.arrl.org/pub/qex as file blkbox.zip.

The Bloodlines

Refer to the schematic diagram, Fig 1. This shows the hardware functions that must be provided to support the RF module. Three dc voltages are required: 12 V, 8 V and 5 V. The main external supply is 12 V. 8 V and 5 V are derived using 78xx 3-pin regulators that are supplied from the main 12 V. Ground (or earth) must be connected to pins 1 and 6 of the module, the (E) lines of connector CN-2. Ground should also be attached to the chassis of the module through the use of a heavy gauge wire with an eyelet terminal. Pins 2, 3, 4 (SB) and 10 (B) of CN-2 require 12-V dc. The exposed terminal near the back of the module should also be supplied with 12 V through a heavy gauge wire. This supplies power to the RF amplifier brick within the module. Pins 8 and 9 (8C) of connector CN-2 require 8 V. Pin 11 (5C) of connector CN-2 requires 5 V. The microcontroller will also require 5 V. The low-level receiver audio comes from the RA line of connector CN-1. This line must be gated through a 4053 CMOS analog switch IC to provide squelch muting. The analog switch is controlled by the SC line. The SQ line must be connected to the wiper and high side of a 50-kΩ potentiometer. The low side of the potentiometer should be connected to ground. This is the squelch control. Gated audio goes into a simple LM386 amplifier that drives an 8-Ω speaker.

I use a Kenwood MC-44 amplified condenser microphone that's wired into the module through an 8-pin chassis-mount connector. An LM358 op amp is used to amplify the TX audio since the mic audio alone proved to be too low. A trimmer is used to adjust for optimum transmitted audio. The output of this amplifier is connected to the MO (mic output) and ME (mic earth) pins on connector CN-1. The PTT line of the mic gets grounded on transmit. This is sensed by the microcontroller, which in turn sends the appropriate serial data to the shift register to put the module into transmit mode. The microphone used here has a built-in DTMF pad that requires an 8-V supply. The remaining lines of CN-1 and CN-2, such as the SM line (S-meter output), are unused at this time and may be left open.

It's Alive!

I encountered RF feedback once I got the module to transmit. Proper shielding and the liberal use of bypass capacitors on all dc lines is required. I fabricated a metal cover and attached it to the bottom of the module since it is normally exposed. This reduced my RFI problems significantly.

Consideration must be given to overheating since the 2-m module is capable of transmitting 50 W of power on the high setting. The die-cast housing of the module is not sufficient to dissipate the amount of heat generated by the power amplifier module. The TM-741 and TM-742 transceivers use the other installed band modules as additional heat sinking, as well as a small fan to keep the active band module cool. An appropriate heat sink should be attached to the stand-alone module, especially if running on the high power setting.

The RF performance of the project is as expected. Signal and audio reports have been good! The only problem I had was the poor quality of audio coming out of the 2-inch PC speaker that I originally used (what was I expecting?). A proper extension speaker is now used, and the audio quality is superb.

The scope of this article is to demonstrate how to calculate and load the registers to bring the module to life, so the 8031 firmware as listed performs only the most rudimentary of functions to get the module up and running on a single frequency. The next step will be to add frequency agility and display, memory channels, multiple VFOs and other bells and whistles through further programming of the microcontroller and additional hardware. Of course, a 68HC11 or PIC16Cxx, or any other microcontroller that you may be familiar with, can be used instead of the 8031—as long as the basic programming core is followed. Although the register data in this article is specific to the 2-m module, the programming concept and hardware remain the same for all of the others. Consult the service manual.

Other Tidbits

Mating header connectors are used at CN1 and CN2 to avoid soldering directly to the main PC board of the module. These were obtained from Kenwood Parts under part number E40-5452-05. Curiosity had me wondering if the PLL chips used in these modules are generic parts or custom to Kenwood. I could not determine the actual part number or manufacturer of the chip since the PLL and VCO are sealed into a functional block. The service manual refers to this block by a

Programming Word Description

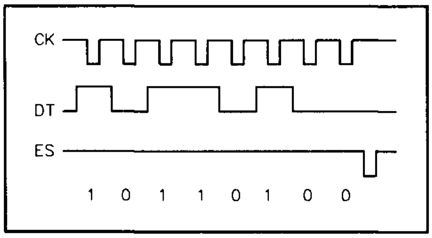
The serial *shift register* inside the module controls power level settings and TX/RX switching. A surface-mount 4094 CMOS part is used. Data is sent MSB first.

SHIFT REGISTER FORMAT
1011XXXX
^ ^
Bit 8 Bit 1

Note that this data gets complemented (inverted) inside the module before reaching the 4094 shift register.
Bit 1: TX/RX switching: 0=RX, 1=TX
Bits 2 and 3: Output power selection: 00=low, 10=medium, 01=high (bit3, bit2)
Bit 4: Special function register. Selects AM or FM mode on receive for the 2-m module: AM mode below 136 MHz=1, FM mode 136 MHz and above=0.
Bit 5: =1
Bit 6: =1
Bit 7: =0
Bit 8: =1
These bits not used or function unknown. Set to specified levels for normal loperation.

Example: To set the module to FM receive mode with the power setting set to medium, send the following data: 10110100

Waveforms:



The *reference register* sets the frequency step size of the PLL. Data is sent MSB first.

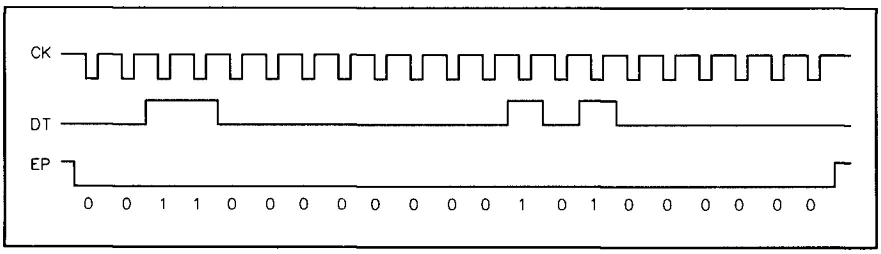
REFERENCE REGISTER FORMAT
0011 XXXXXXXXXXXXXXXXX
^ ^
Bit 21 Bit 1

Bits 1-17: PLL reference oscillator frequency division ratio. Sets the step size. 17-bit word length.
Bit 18: Binary 1 identifies this stream as the reference register data.
Bit 19: =1
Bit 20: =0
Bit 21: =0
These bits not specified or function unknown. Set to specified levels for normal operation.

The following equation determines the binary value to be loaded into the reference oscillator divider to set up the step size.

Division ratio = F_{osc}/(N <prescalar> x desired step size <kHz>)
For a 5-kHz step size:
Ref = 12800/(8 x 5)
= 320
= 00000000101000000 binary

Waveforms:



The *comparison register* comprises the A and N counters, which determine the operating frequency of the PLL. Data is sent MSB first.

COMPARISON REGISTER FORMAT
0010 XXXXXXXXXX XXXXXXX
^ ^
Bit 21 Bit 1

Bits 1-7: PLL A counter register. 7-bit word length.
Bits 8-17: PLL N counter register. 10-bit word length.
Bit 18: Binary 0 identifies this stream as the comparison register data.
Bit 19: =1
Bit 20: =0
Bit 21: =0
These bits not used or function unknown. Set to specified levels for normal operation.

The following equations determine the binary values to be loaded into the N and A registers for a particular operating frequency (all frequencies in kHz).

For RX: $F_{vco}=F_{rx}-10700$ (IF) For TX: $F_{vco}=F_{tx}$

N register value=$(F_{vco}/5)/128$ (whole number only)

A register value=$(F_{vco}/5)-(N \times 128)$

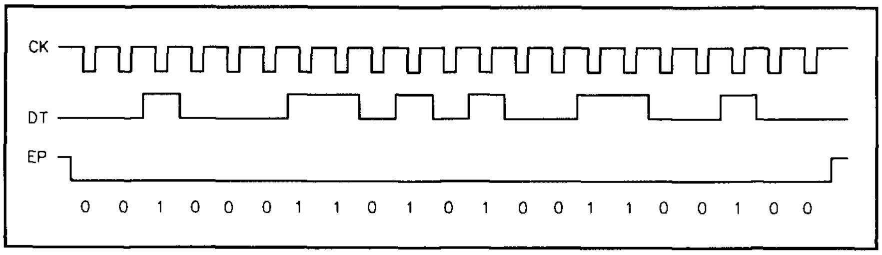
For example, to receive 146.88 MHz;

$F_{vco}=146880-10700=136180$ (kHz):

N=(136180/5)/128
= 212 (whole number only)
= 0011010100 binary

A=(136180/5)–(212 x 128)
=27236–27136=100
= 1100100 binary (first 7 bits only)

Waveforms:



Kenwood part number. However, the programming method is similar to that used on some Motorola PLL chips.[3]

An alternate use for these modules could be to use two of them as a repeater. Both modules could run off a single microcontroller and power supply. A decent repeater controller could take care of the switching and audio functions. Of course, a large heat sink would have to be fixed to the TX module to support the heavy duty cycle. This could make for an affordable alternative to buying a commercially available repeater.

The information in this article should help get you on your way to running one of the Kenwood modules as a monoband FM transceiver on any ham band from 28 MHz to 1.2 GHz. I hope it encourages some experimentation and homebrewing from a different perspective. My thanks to Francis, VE3TDL, and Steve, VE3TZC, for their assistance. For further discussion, comments or questions I can be reached via Internet (address above), packet radio: VA3TO@VA3BBS. #SON.ON.CAN.NA or by snail-mail (replies by SASE only).

Notes

[1]Familiarization with these modules through the use of the TM-741 or TM-742 service manual is recommended.

[2]L. S. Electronics, 2280 Camilla Rd, Mississauga, ON L5A 2J8 Canada, tel: 905-277-4893. LSE carries a wide range of MCS-51-based microcontroller products, including a microcontroller beginners book and application notes.

[3]For a greater understanding of phase-locked-loop circuity, I suggest reading the "Radio Frequency Oscillators and Synthesizers" section in a late edition of *The ARRL Handbook*.

Listing 1

```
; Programs the PLL and Shift registers of a Kenwood TM-741/742 2 meter
; module to control the frequency and TX/RX control.
; This program runs on an L.S.Electronics ELC-31 8031 microcontroller.
; Clock xtal = 11.0592 MHz.
; Assembled using the PseudoSam(TM) 51 Assembler V1.6.02
;
; By Hugh Duff VA3TO April 1995 Brampton, Ontario, Canada
;
;**************************************************************
; Port pin definitions

      .equu PTT,p1.0 ; Push-To-Talk input...Lo=TX, Hi=RX
      .equ  EP,p1.1  ; PLL enable output
      .equ  CK,p1.2  ; Clock output
      .equ  DT,p1.3  ; Data output
      .equ  ES,p1.4  ; Shift Register enable output

;*************************************************************
; Start of program..micro initialization
; PSW (Program Status Word) BANK 0 contains the RX data, PSW BANK 1
; contains the TX data.
; R7 to R5 contain the Reference register word
; R4 to R2 contain the Comparsion register word
; Both words are 21 bits in length. The top 3 bits of R7 and R4 are
; unused. These get purged at the "bit_5" subroutine.

INIT:  mov   SP,#0x2F ; setup stack
       clr   PSW.3    ; set PSW to register to point at bank 0
       clr   PSW.4    ;         "

; Load PSW BANK0 with RX step,frequency and keying data
       mov   R7,#0x06 ; REF data for 5kHz step,   1st byte
       mov   R6,#0x01 ;    "             2nd byte
       mov   R5,#0x40 ;    "             3rd byte
       mov   R4,#0x04 ; COMP data for 146.88 MHz RX, 1st byte
       mov   R3,#0x6A ;    "             2nd byte
       mov   R2,#0x64 ;    "             3rd byte
       mov   R1,#0xB4 ; Shift Reg. data for RX

; Load PSW BANK1 with TX step,frequency and keying data
       setb  PSW.3    ; select TX bank in PSW
       mov   R7,#0x06 ; REF data for 5 kHz step,   1st byte
       mov   R6,#0x01 ;    "             2nd byte
       mov   R5,#0x40 ;    "             3rd byte
       mov   R4,#0x08 ; COMP data for 146.28 MHz TX, 1st byte
       mov   R3,#0x72 ;    "             2nd byte
       mov   R2,#0x48 ;    "             3rd byte
       mov   R1,#0xB5 ; Shift Reg. data for TX

; Set module control lines to appropriate resting levels
       setb  EP       ; PLL enable line normally HI
       setb  CK       ; PLL clock line normally HI
       clr   DT       ; PLL data line normally LO
       setb  ES       ; Shift Register enable normally HI
```

```
;*************************************************************
; Main routine loads receive data,waits for a PTT then loads
; transmit data and waits for PTT to drop...

MAIN:  clr   PSW.3   ; select RX bank in PSW
       acall LD_SHFT ; load RX shift register data
       acall LD_REF  ; load RX Reference data into PLL
       acall LD_COMP ; load RX Comparison data into PLL
RX_IDLE:jb   PTT,*   ; wait for PTT to ground
       acall TX      ; go load TX data and return
TX_IDLE:jnb  PTT,*   ; wait for PTT release
       acall RX      ; go load RX data and return
       ajmp  RX_IDLE ; repeat this loop

;*************************************************************
; Selects receive databank and calls load routines

RX:    clr   PSW.3   ; select receive data bank
       acall LD_COMP ; load com.freq. data to PLL
       acall LD_SHFT ; disable TX in shift reg.
       ret           ; return

;*************************************************************
; Selects transmit databank and calls load routines

TX:    setb  PSW.3   ; select transmit data bank
       acall LD_COMP ; load comp.freq. data to PLL
       acall LD_SHFT ; enable TX in shift reg.
       ret           ; return

;*************************************************************
; Loads PLL Reference register

LD_REF: clr  EP      ; enable PLL
       acall DELAY   ; wait
       mov   A,R7    ; get 1st PLL REF data byte
       acall bit_5   ; send it out
       mov   A,R6    ; get 2nd PLL REF data byte
       acall bit_8   ; send it out
       mov   A,R5    ; get 3rd PLL REF data byte
       acall bit_8   ; send it out
       acall DELAY   ; wait
       setb  EP      ; disable PLL
       ret           ;

;*************************************************************
; Loads PLL Comparison register

LD_COMP:clr  EP      ; enable PLL
       acall DELAY   ; wait
       mov   A,R4    ; get 1st PLL COMP data byte
       acall bit_5   ; send it out
       mov   A,R3    ; get 2nd PLL COMP data byte
       acall bit_8   ; send it out
       mov   A,R2    ; get 3rd PLL COMP data byte
       acall bit_8   ; send it out
```

```
      acall  DELAY    ; wait
       setb  EP       ; disable PLL
       ret            ;

;***************************************************************
; Loads Shift Register

LD_SHFT:mov  A,R1     ; get data
       acall bit_8    ; send it out
       clr   ES       ; strobe data into shift register
       acall DELAY    ; wait
       setb  ES       ; return ES to rest level
       ret            ;

;***************************************************************
; Rotates 8 bits of data in reg.A out to DATA line

BIT_8: mov   B,#0X08 ; setup reg.B as an 8 bit counter
LOOP8: rlc   A        ; rotate left 1 bit thru carry
       mov   DT,C     ; move carry status to DATA port line
CLK8:  acall CLOCK    ; clock the data
       djnz  B,LOOP8 ; keep looping until 8 bits sent
       ret            ;

;***************************************************************
; Rotates 5 bits of data in reg.A out to DATA line
; The top 3 bits of R7 and R4 are not sent since the registers
; are only 21 bits in length.

BIT_5: mov   B,#0X05 ; setup reg.B as a 5 bit counter
       rlc   A        ; waste unused MSB
       rlc   A        ;     "
       rlc   A        ;     "
LOOP5: rlc   A        ; rotate left 1 bit thru carry
       mov   DT,C     ; move carry status to DATA port line
CLK5:  acall CLOCK    ; clock the data
       djnz  B,LOOP5 ; keep looping until 5 bits sent
       ret            ;

;***************************************************************
; Toggles the clock line

CLOCK: clr   CK       ; Clock line LO
       acall DELAY    ; wait
       setb  CK       ; Clock line HI
       acall DELAY    ; wait
       ret            ;

;***************************************************************
; Delay 200 us based on 11.0592 MHz xtal

DELAY: mov   R0,#0X5C; load count
DLY1:  djnz  R0,DLY1 ; kill time
       ret

      .end
```

A Variable IF Selectivity Unit

Reprinted from the September 1995 issue of Radio Communication.

By A. R. Thomson, GM3AHR

The unit described provides variable IF selectivity for older commercial or home-brew transceivers with a 9-MHz IF. It was originally designed as an add-on for the G3TSO transceiver, but it can be used in other designs and with other IF frequencies with a change in the value of some of the components. It operates by superimposing a second filter response over that of the receiver IF filter. By altering the position of the second response, relative to the first, the effective selectivity can be made variable. No originality can be claimed for the design, which is a derivation of circuits found in some commercial and home-brew amateur-radio equipment.

Meadow Rise
4 Lawview Gardens
Bonnybank, Levin, Fife KY8 5SW
England

Theory

A block diagram of the unit is shown in Fig 1. The output of the receiver SSB filter is mixed with the output of a variable 19.7-MHz crystal oscillator and the output of the mixer passed through a second filter of 10.7 MHz. For best results, this filter should have the same bandwidth and shape as the SSB filter.

An attempt was made to construct this second filter from 10.7-MHz crystals using published data for ladder crystal filters, but the results were disappointing due probably to the shape and the high passband ripple which was evident. A commercial crystal filter was obtained at a reasonable price and has proved entirely satisfactory.

The output of the filter is mixed with the 19.7-MHz oscillator resulting in the original 9-MHz signal. By varying the oscillator above and below 19.7 MHz, ie, ±2.5 kHz, the signal applied to the second filter will be outside its passband and effectively blocked from the IF stages. Varying the oscillator by smaller amounts, interfering signals can be eliminated and still retain sufficient intelligibility from the wanted signal, as well as SSB signals, CW reception is much improved.

The circuit is given in Fig 2. A design for a PC board is not included, but a suggested board layout is given since it was felt that individual layouts of receivers would require different PC board sizes and shapes. My own PC board, which is double sided to aid stability, measures 3×2 inches, and the layout basically follows the circuit diagram. An 8-V stabilized supply is required for the oscillator to ensure frequency stability.

Frequency shift is obtained by the varicap diode, D1. All transformers are the same type. An 82-pF capacitor is connected internally, and it may be necessary to connect an extra 10 pF across T4 externally to bring it into tune. The output of the second mixer contains signals at 9 and 10.7 MHz, and the latter must be removed before it reaches the receiver IF stages. Ideally, a filter with a bandwidth of about 10 to 15 kHz at 9 MHz should be used, but could not be obtained at a reasonable price.

By using loosely coupled T4 and a further stage of amplification and tuned circuit, T5, satisfactory results were obtained.

If other IF frequencies are considered, it would be advisable to retain the 10.7-MHz filter since it is readily obtainable. The 19.7-MHz xtal would need to be changed for one that would give the required 10.7 MHz from the first mixer. All the transformers are of the same type—KAC6148As. They have parallel 82-pF capacitors fitted across the high-impedance winding inside the can as standard and are designed for 10.7 MHz. T1 and T2, used in the 10.7-MHz section of the circuit, can use these transformers unmodified.

T3, T4 and T5 are used in the 9-MHz section of the circuit and the manufacturer claims that they will tune to this frequency. However, additional capacitance may be required to tune the transformers to 9 MHz, particularly T4 because it is very lightly coupled; see "Setting Up."

The supply to the board and the connections to the panel potentiometer, VR2, are by plug and socket (shown in the photo), and the input and output connections are by soldered pins at each end of the board.

Setting Up

Connect the unit to the receiver with the system fitted between the output of the SSB filter and the following IF stages. Ensure that the supply is between 12 and 14 V.

Use a receiver or frequency meter and check that the 19.7-MHz oscillator is working.

Set the panel control midway and adjust the preset control until signals are resolved normally. Peak all transformers for maximum signal. Varying the panel control, RV1, should result in signals disappearing about halfway from the center position in each direction.

The panel control should have a reasonably stiff action to prevent accidental movement of this control and a sudden dead receiver!

The value of R14 was selected so that the full range of frequency shift of the second filter equals 180° rotation of the potentiometer, ie, 90° each side of center. There is no reason why the travel could not be made greater to give greater resolution as long as the signal disappears before the end of travel at each end. With the value of R14 given, the signal does disappear as $\pm 90°$; I find this is not too coarse.

The PC board should be fitted as near as possible to the receiver IF board and short coax connections made between the two. The panel control wiring may need screening, but I found this to be unnecessary.

The dc supply to the PC board is taken from the receiver 13-V line and, in my own case, this supply is disconnected and grounded on transmit. The

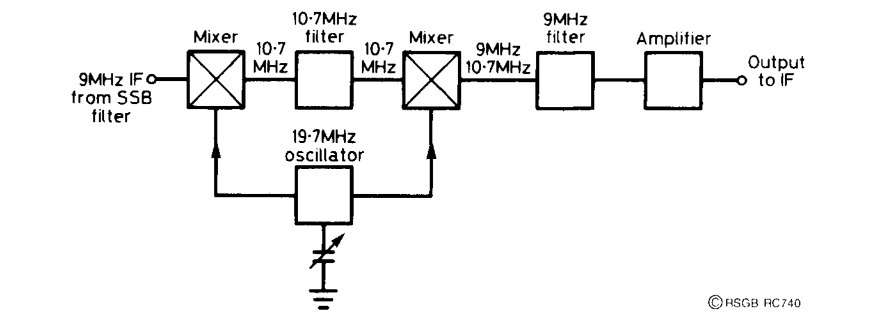


Fig 1—Block diagram of the variable IF selectivity unit.

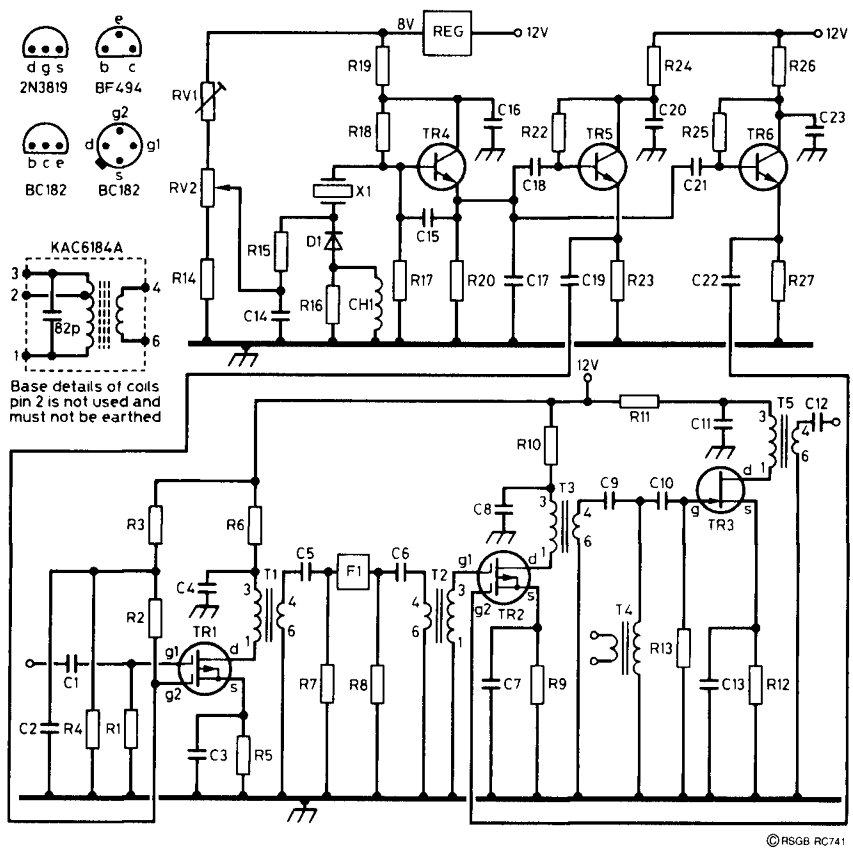
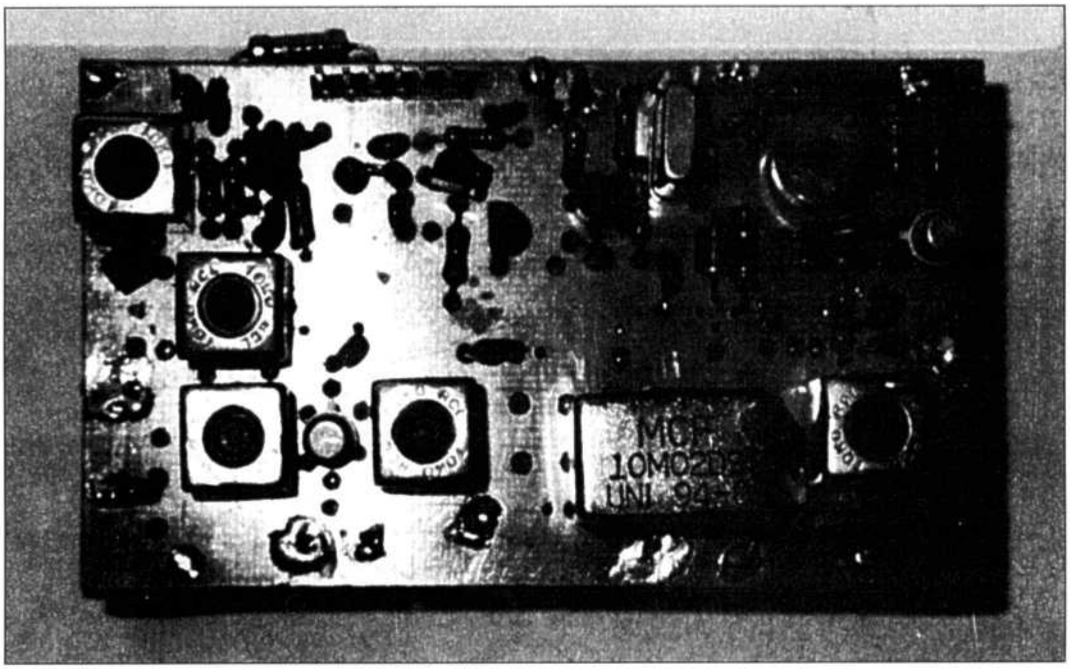


Fig 2—Circuit diagram of the variable IF selectivity unit.



PC board layout of the variable IF selectivity unit.

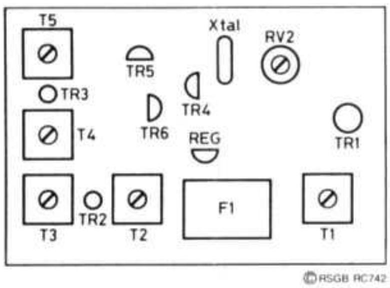


Fig 3—Component layout of the variable IF selectivity unit.

Component List

Resistors

R1, R2, R9, R12, R23, R27	1 kΩ
R3, R15	47 kΩ
R4	2.7 kΩ
R5, R19	100 Ω
R6, R10, R11, R24, R26	200 Ω
R7, R8	560 Ω
R13, R22, R25	180 kΩ
R14, R16, R21	10 kΩ
R17, R18	15 kΩ
R20	330 Ω
RV2	10-kΩ linear
RV1	10-kΩ preset

Capacitors

C9, C10	2 pF
C12	1 nF
C1, C2, C4, C7, C13, C14, C16, C19, C20, C22, C23	10 nF
C3	47 nF
C8, C11	0.1 µF
C5, C6, C15	100 pF
C18, C21	33 pF
C17	39 pF

Inductors

RFC	3 turns on FX1115 bead

Semiconductors

REG	78L08
TR1, TR2	MFE20
TR3	2N3819
TR4	BF494
TR5, TR6	BC182
D1	BB105

Additional items

F1	10MO2DS
CH1	5.6 µH
X1	19.7 MHz
T1, T2, T3, T4, T5	KAC6184A

Components are available from JAB Electronics Components, 1180 Aldridge Road, Great Barr, Birmingham B44 8PB England.

use of switching diodes is not therefore necessary to isolate the new board, but this may be necessary in some cases.

On tuning up, once the potentiometer has been set midway and VR1 adjusted to give a normal signal, the transformers can be peaked for maximum signal. This can be done by injecting a 9-MHz signal at the input from a signal generator, but the frequency must be set accurately to 9 MHz. In my own case, I connected the new circuit into my receiver and observed the output on the S-meter. T1 and T2 should be peaked first and then the other three, repeating the process. It should be possible to find two tuning positions for each coil, one with the slug well into the coil and the other with the slug beginning to come out of the coil. This occurs because the tuning slug is moving out of the coil in the other direction.

With regard to using a GDO as the signal source, difficulty may be experienced with drift. The best method for tuning is the receiver method, which uses the IF frequency of the unit it is to be used with. The GDO can be tuned to any desired frequency the receiver covers and drift is easily corrected by the receiver tuning control. If the receiver has an RF gain control, this should be reduced until the S-meter is about half scale. Alternatively, move the GDO further away from the antenna input. □

RF

By Zack Lau, KH6CP/1

10-GHz SMA to WR-90 Transition

Connecting 10-GHz waveguide to SMA connections requires a precision-constructed transition assembly. This design is meant to be easy to duplicate. Despite the compromises made to simplify construction, you can get a return loss of around 20 dB at 10.368 GHz without tuning. Instead of using expensive test equipment, a relatively inexpensive dial caliper can be used to achieve tolerances of perhaps five thousandths of an inch.

I recommend that you use two- or four-hole flange SMA connectors with captivated center contacts. The captivated contacts are needed to keep the probe in place. Two-hole connectors work best, but they tend to be more difficult to find. The mounting flange of a four-hole connector extends past the capped end of the waveguide because the connector must be placed so close to that end (see Fig 1). The four-hole connectors can be used by filing the mounting flange after you solder the shorting back plate to the waveguide.

You can identify captivated contacts by inspecting the connector and looking for a small circle of glue on the shield next to the 1/4-36 threads or shell. You could also try gently moving the center pin to see if it's locked in place, but don't blame me if you break the connector!

The probe diameter of 3/32 inch (Fig 2) was chosen for convenience; you can make the probe out of hobby brass tubing. A study of the literature indicates that the usual 50-mil center-pin diameter would work better, but connectors with long center pins tend to be harder to find. It's much easier to file pieces of tubing to the right length than to file center pins. If you aren't careful, you can dislodge, bend or even break center pins. A bonus of filing cheap tubing is that you can toss the mistakes till you finally get the right length of 0.270 inches. The tubing is easily measured with a 6-inch dial caliper.

Fig 3 shows the waveguide flange, made from sheet brass, that completes the assembly. Making your own flange is a cheap—and easy—alternative.

I found it important to keep the matching section—the low-impedance transmission line formed by the thick probe passing through a 0.166-inch hole in the waveguide wall. I didn't have much luck getting a no-tune design when I kept the impedance at 50 Ω.

You may add tuning screws to improve performance. Normally, people put a couple of screws along the center line, spaced either an eighth or a quarter of a guide wavelength apart. However, professional transitions have a tuning post off to one side. You might look into this for more advanced designs.

The problem with tuning screws is properly adjusting them. It is relatively straightforward with a precision directional coupler. Ordinary couplers will have much less directivity, perhaps only 12 to 15 dB. This makes them almost useless for SWR measurements. Such a low directivity results in a rather large region of uncertainty. Even a 26-dB directivity coupler can result in noticeable errors.

A simplified but useful approximation for estimating the error appears on an ancient HP slide rule. It assumes that the effective source match is equal to the directivity, and that there is no calibration error.

$$uncertainty = A + A \times rho^2$$

where A is the directivity and rho is the measured value of reflection coefficient.

225 Main Street
Newington, CT 06111
email: zlau@arrl.org

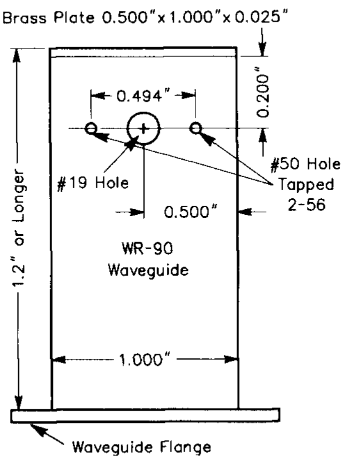


Fig 1—Top view of the WR-90 to SMA transition without the SMA connector in place.

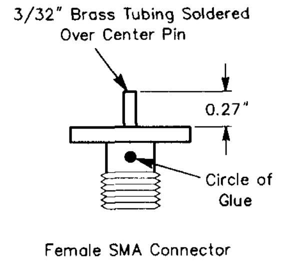


Fig 2—SMA connector and probe details.

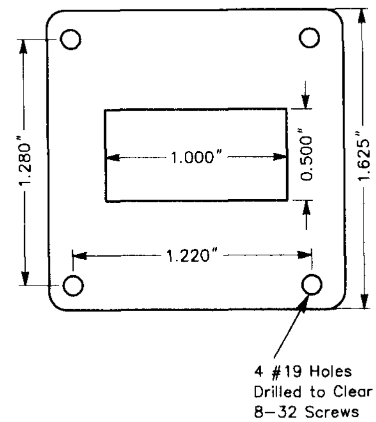


Fig 3—Dimensions for making cheap waveguide flanges for WR-90.

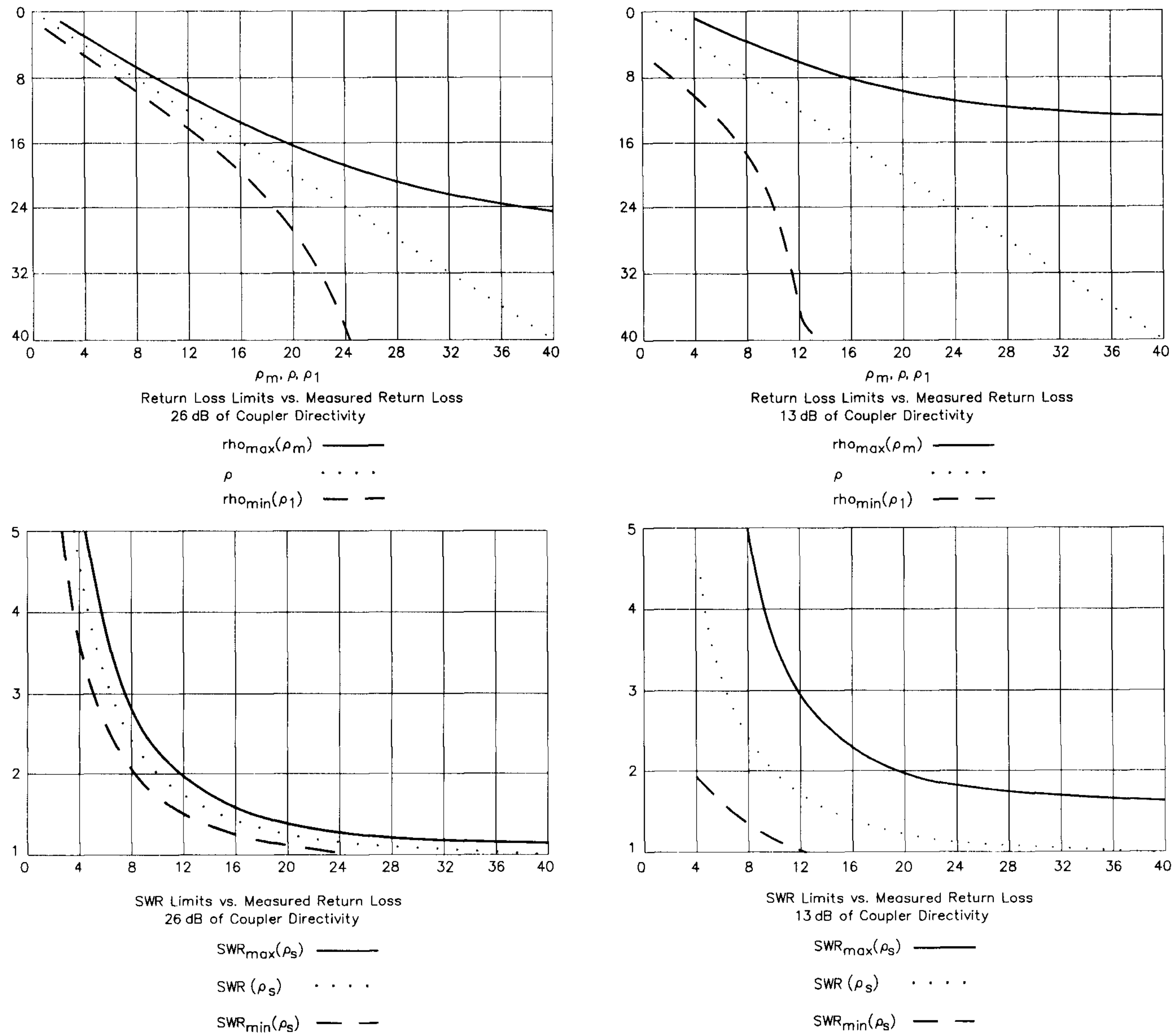


Fig 4—Graphs showing the possible error of return-loss and SWR measurements made using directional couplers. The greater the coupler's directivity, the more certain the measurement.

If A is 0.05 (–26 dB) and *rho* is 0.1 (–20 dB), we get an uncertainty of 0.0505. Thus, the actual return loss could be between 0.1505 and 0.0495, and the measured return loss of 20 dB with such a coupler could actually be between 16 dB and 26 dB (1.35 to 1.1 SWR). Using a coupler with only 15 dB of directivity, the actual SWR could be anywhere between 1.78 and 1.1, for that same measurement. I've included some graphs in Fig 4 to show this a bit more clearly.

Construction

I begin by getting the necessary materials. Normally, I get waveguide and SMA connectors at hamfests, where they can be found inexpensively. The brass tubing and sheet stock are available from Small Parts.[1] They also sell 2-56 taps and stainless-steel hardware. The thickness of the brass sheet for the shorting plate that seals off the back of the waveguide isn't important—I normally use anywhere from 0.015 to 0.032-inch sheet stock. I use a shear to cut it into 0.500×1.000-inch rectangles. I usually make the waveguide at least 1.2 inches long. You might make it even longer if you intend to add tuning screws.

I next prepare the waveguide by marking it with a dial caliper. I get excellent results by buying an inexpensive caliper and using the jaws as a scribe. I save so much time with this technique that it would be worthwhile even if the jaws wore out every few years. (I've not noticed any wear with the soft materials I mark.) A dial caliper works much better than an expensive digital one for this purpose—I find it easier to count off the increments needed for parts like SMA connectors. For instance, I'll mark the center line of the waveguide, then decrement by 247 mils for the mounting holes. With an oxidized waveguide surface, the scribe lines are quite visible. If you polished it up to a bright shiny finish, it might be necessary to make the lines more visible! Professional machinists use layout dye. I've found that a permanent marker is an acceptable substitute. A good source of low-cost machinists' supplies is Enco Manufacturing Company.[2]

The dimensions shown are for standard SMA connectors. I've not seen nonstandard two-hole flange connectors, but I have purchased unusual four-hole flange connectors. These are easily identified by the flange dimensions: they aren't the usual 0.5-inch rounded squares.

Three holes are drilled for the SMA connector. (See Fig 1.) A no. 19 bit makes the 0.166-inch hole for the probe. The 0.070-inch holes for the 2-56 tapped holes are drilled with a no. 50 bit. The holes are then deburred. The outside holes are easily done with a large drill bit or commercial deburring tool. I use a file to smooth the inside surface of the guide. I tap the holes after deburring them. I find that using tapping fluid significantly reduces the number of taps that break. The *Microwave Handbook, Volume 2, Construction and Testing*, has useful information on this sort of metal work.[3]

The 0.50×1.00-inch shorting plate is soldered to the waveguide with a 100-W soldering iron while being held in place with a C clamp. Other people use propane torches to solder waveguide, but I've found the iron sufficient. You can make the plate slightly smaller if you wish. This can make it easier to file the surface of the waveguide flat. This step is needed if you want to attach a four-hole flange SMA connector. This connector will stick out over the edge, so it's not as aesthetically pleasing as a two-hole flange connector. But four-hole flange connectors are often significantly cheaper.

Attaching the waveguide flange completes the soldering job. To save on weight, you might consider making your own flanges out of 25 or 32-mil brass. Thicker material is difficult to work with, while thinner material isn't quite sturdy enough. I highly recommend you compare your flange layout against a known good flange before drilling those asymmetrical holes.

The probes are easily cut with a hacksaw. They can be filed or sanded to precisely the right length. It's normal for them to fit loosely over the center pin, so it may take a few tries to solder them with the proper alignment. Finally, the SMA connectors are attached with 2-56 screws to the waveguide. A minor problem is that the ideal length for the screw shafts is about 100 mils. Of course, the nearest standard-size screw is 1/8 inch, or 125 mils. I've found hex nuts to be precisely the right thickness to space the screws, if you are willing to use Loctite instead of lock washers. Alternately, you could use a pair of lock washers. This is just a little too thin, but it seems to work just fine, even with a little bit of stainless sticking inside the waveguide.

Notes

[1]Small Parts Inc, 13980 NW 58th Court, PO Box 4650, Miami Lakes, FL 33014-0650. Tel: (800) 220-4242 and (305) 557-8222; fax: (800) 423-9009.

[2]Enco Manufacturing Company, 5000 W Bloomingdale Avenue, Chicago, IL 69639. Tel: (800) USE-ENCO [(800) 873-3626] and (312) 745-1520.

[3]*Microwave Handbook.* This 3-volume set is published by the Radio Society of Great Britain. Edited by M. W. Dixon, G3PFR, and sold in the US by the ARRL.