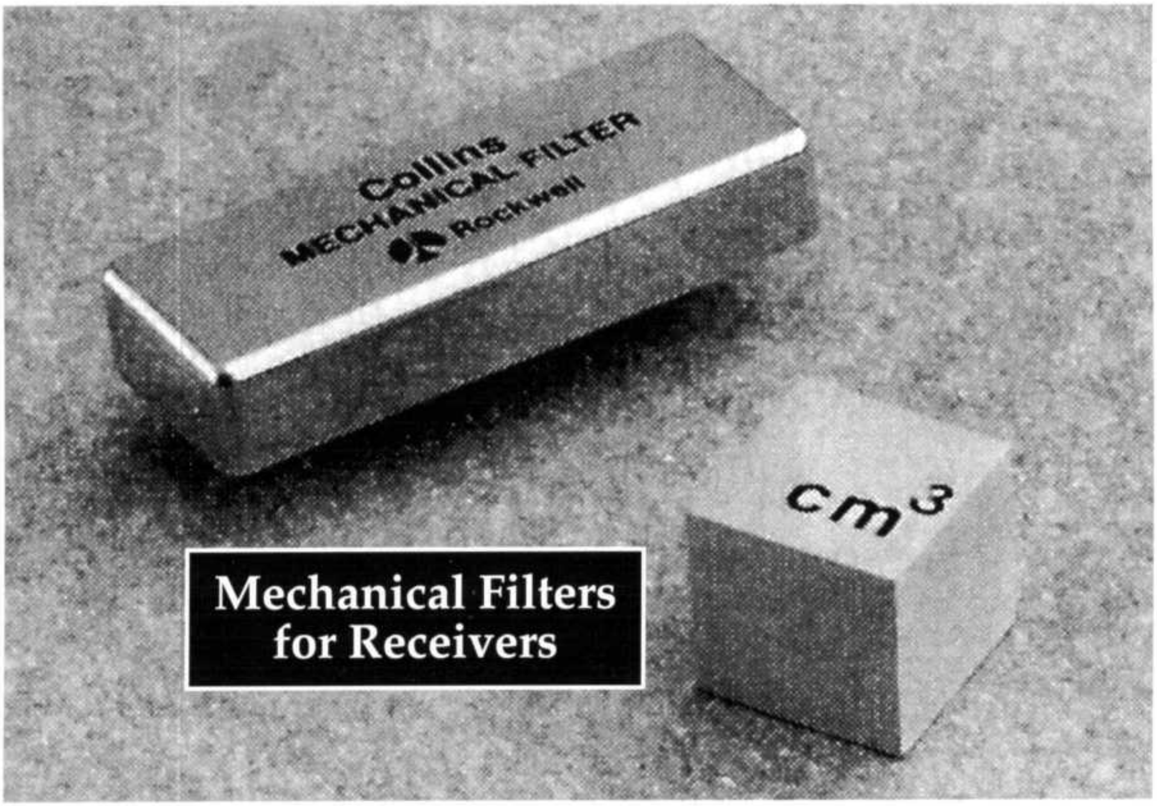
QEX

$1.75

ARRL Experimenter's Exchange

March 1996



Mechanical Filters for Receivers

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QEX

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About the Cover

Low-cost units mark a resurgence of mechanical filters in receiver IF chains.

ISSUE NO. 170

Features

Columns

March 1996 QEX Advertising Index

Empirically Speaking

Bigger: Not Always Better

Seems that every few years we need to upgrade our program-development packages. Aside from the fact that the user community keeps migrating to new operating-system environments (*Windows 95* being the latest), the tools available for programming keep getting more and more capable.

Recently, Borland came out with *Borland C++ 5.0*. This significant upgrade to their previous 4.5 version includes 32-bit Windows support (for *Windows 95* and *Windows NT*), as well as a slew of new tools, such as a version-control package and a *Windows 95* install packager. The problem is, BC++ 5.0 is *big*. It's not so bad that a full install takes upwards of 200 Mbytes of hard disk—200 Mbytes of drive space is pretty cheap these days. No, the problem is that it takes scads of RAM to run BC++ 5.0 efficiently. And RAM prices never seem to drop much, lately. Borland calls for 16 Mbytes of RAM minimum, 24 Mbytes recommended. And some users are reporting that having even more memory makes a big difference. Users with 32 Mbytes are finding that a noticeable improvement in speed of the BC++ 5.0 development environment occurs with the addition of RAM to the 48- or 64-Mbyte level! Of course, particular system configurations will vary these numbers.

Now, part of this increased resource requirement can be chalked up to the fact that BC++ 5.0 can do more than BC++ 4.5. But not *that* much more! And the fact that the BC++ 5.0 development environment is 32-bit hosted, whereas 4.5 was 16-bit, may make a difference, too. But the bottom line is that to usefully use this new package, you need RAM, and lots of it.

This isn't unique to BC++ 5.0, and we don't mean to pick on Borland. Other recently released software packages, including Microsoft's latest release of *Visual C++*, aren't much less resource hungry. And the new features of BC++ 5.0 are, we think, worth the cost in resources. Still, we can't help but wonder if these new features couldn't have been accomplished using at least a little less in the way of resources.

So, what has this got to do with *QEX* and ham radio? Simply this: A few months ago we suggested in this space that developers of software for amateur applications consider carefully the target environment. We described this need in the context of choosing to make a DOS application or a Windows application. Among the many responses we received were a number from readers whose reaction was that they would be reluctant to buy a Windows application because of the resource overhead of Windows. Well, guess what. It's only going to get worse. As faster computers with more RAM and disk space become available, software developers will find ways to make them run slowly! They'll do so, of course, by adding features and capabilities to applications, and we'll get used to those features and demand them. So you might as well plan that CPU/RAM/disk upgrade now. You're going to need it.

On the other hand, this 90-MHz Pentium with 16 Mbytes of RAM and 1.2-Gbyte hard drive sure runs those old DOS applications fast!

This Month in *QEX*

Adjusting a resonant antenna's length based on the conductor diameter is normally done using the *Handbook* formula. But if you're modeling an antenna system, you may need a more exact approach. "Calibrating *K* to *NEC*," by L. B. Cebik, W4RNL, provides one.

Need a simple, low-distortion audio or low-frequency oscillator? Maybe you should consider the venerable Wein bridge. "Designing a Wein-Bridge Oscillator," can be done quite simply, as Parker R. Cope, W2GOM, shows us.

New, low-cost mechanical filters bring this oldie-but-goodie technology back to the forefront of receiver design. Learn to use them effectively in, "The Mechanical Filter in HF Receiver Design," by William E. Sabin, WØIYH.

Rainer Bertelsmeier, DJ9BV, gives us yet another of his microwave circuits, this time a "PHEMT Preamp for 6 cm."

Finally, in this month's "RF" column, Zack Lau, KH6CP/1, presents an overview of current preamp technology for above-VHF use. Zack also supplies a complete design for a 13-cm preamp.
—*KE3Z, email: jbloom@arrl.org.*

Calibrating K to NEC

The Handbook *rule-of-thumb for determining electrical antenna length may not be adequate for computer modeling of antennas.*

By L. B. Cebik, W4RNL

K is the factor by which you shorten the physical length of an antenna to achieve the correct electrical length. A half-wave dipole in free space or a vertical quarter-wave monopole over perfect ground will, at resonance, be physically shorter than a half or a quarter of a wavelength, respectively, as shown in Fig 1. The amount of shortening is regular; that is, it is a function of the antenna and not of the surroundings (which will also have an effect upon the length of a resonant antenna). Nor does the shortening depend upon deformations, usually loops, at the ends of wire antennas, which simply add to the end effect.

For most practical purposes, hams cut half-wavelength wire antennas to a length in feet approximating the constant 468 divided by the frequency in MHz. This standard formulation has been in *ARRL Handbooks* since at least 1930. In 1947, the *Handbook* introduced a graph relating the ratio of the antenna element diameter and a half wavelength in free space to a multiplier, *K*. *K* is always less than 1 and approximates the amount by which the antenna is shortened relative to a full half wavelength by virtue of its diameter. *The Antenna Book* has carried the graph continuously since 1949.

Unfortunately, the 1947 discussion refers to end effect only in terms of the additional system capacitance added by antenna end deformation.[1] In point of fact, deformation is an additional end effect. The value of *K* dependent upon antenna diameter is also an end effect to the degree that it is applicable only to the end quarter-wavelength sections of a dipole, however many wavelengths long it may be. Likewise, it is applicable to the final quarter-wavelength section of a vertical antenna over perfect ground, no matter how many wavelengths long the monopole might be. Other sections of such antennas (each a half wavelength for dipoles or a quarter wavelength for monopoles) will be roughly (but not exactly) a full half or quarter wavelength, respectively.

Explanations of the shortening effect are ordinarily traceable to accounts of the end capacitance of an antenna element, as if the end consisted of a spherical cap with the wire's radius.[2] Within the antenna-transmission-line analogy upon which classical mathematical treatments of antennas rest, the shortening effect is reducible for practical work to a constant, often called *K*, corresponding to the diameter-to-half-wavelength ratio at the frequency of interest.[3]

Working values of *K*, the decimal multiplier used to account for regular

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[1]Notes appear on page 8.

end-effect shortening, ordinarily come from graphs of one or another sort. Such values often suffice for ordinary construction purposes where other variables intervene in determining the final antenna element length. However, some calculations, especially those related to antenna modeling by method-of-moments techniques as implemented in such programs as *MININEC* and *NEC-2*, may require more precise—or at least different—values for K than those provided by traditional graphical representations of K.

Calibrating *K* to *NEC-2*

The widespread use of antenna modeling programs, such *as NEC-2* and *MININEC*, suggests an alternative method of calculating both K and the length of antennas. It is possible to correlate values of K with *NEC-2* models, to separate end effects from material and other uniform effects upon antenna elements, and then to calculate predictively, with good accuracy relative to *NEC-2*, the length of single- and multiple-quarter and half-wavelength antenna elements. A small *GWBASIC* program suffices for the calculations, once the calibration is accomplished.

The methods described here produce accurate results in terms of *NEC-2* models for finding K and the length of antenna elements up to seven quarter and half wavelengths for the frequency range of 3 to 30 MHz and for selected antenna materials ranging from #18 AWG copper wire through 2-inch aluminum tubing. Limitations on that accuracy will be described at each stage of the program development.

Let us call K_T the total antenna shortening factor, K_M the portion of the shortening factor attributable to material losses and any other factor uniformly affecting every increment of length along the antenna, and K_E the end effect or the shortening factor applicable to the end quarter wavelength of a vertical over perfect ground or the end quarter wavelengths of a dipole in free space. The technique for calculating K_T requires the selection of an element diameter and the derivation of values of K_T for two frequencies, F_L and F_H, at the lower and upper limits of a range of frequency interest. HF antenna designers, for example, may be interested in the range of 3 to 30 MHz. For any selected element diameter, K_T will vary regularly across the range from K_L to K_H, where K_L will, of course, be the larger value at the lower frequency limit.

Deriving K_L and K_H from antenna models requires the construction of a suitable basic antenna model, for example, a quarter-wave vertical over perfect ground. For a selected element diameter, the antenna is resonated. Technically, resonance is a condition at which the antenna (or, in this case, the model) shows only resistive impedance, with the reactance equal to zero. In modeling practice, anything close to zero is of little practical consequence for a real wire antenna. In this case, however, reactance must be nulled to some arbitrarily small value. For purposes of correlating equations and modeling results, I have used a reactance less than ±0.01 Ω as sufficiently precise to permit calculation to four significant figures. Again, in practice, this degree of precision is spurious, but the aim here is the correlation of two calculational schemes, and the added precision is useful in tracing curves unambiguously.

The choice of *NEC-2* implementations may affect the consistency of modeling results. In addition, the selection of the number of segments per quarter wavelength will also affect the results, since the calculated feedpoint impedance will vary slightly with the number of segments per unit of an-

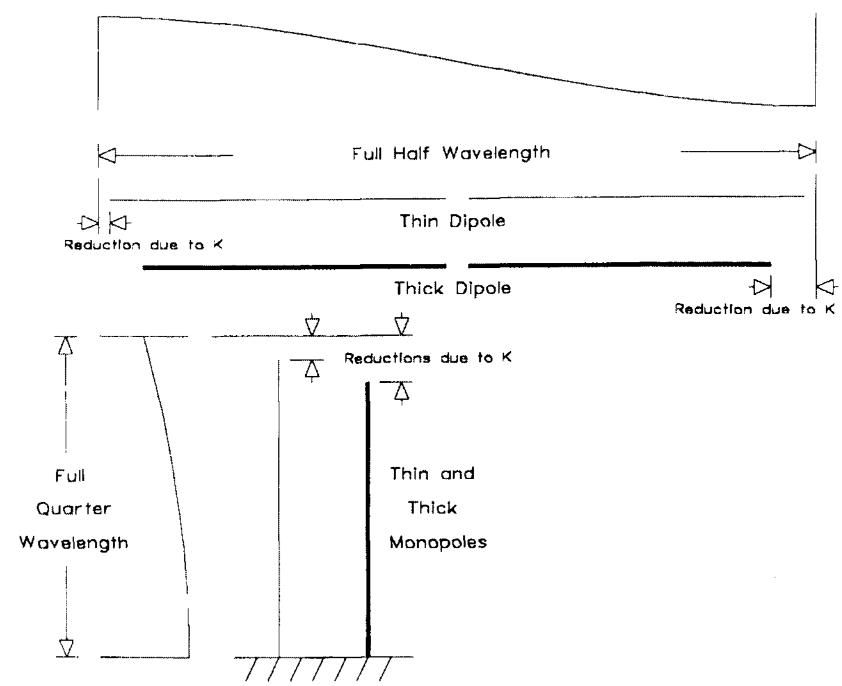

Fig 1—Reduction in the resonant length of dipoles and monopoles due to the "K-factor."

Table 1—Variation of Resonance in *NEC-2* for a Sample Quarter-Wavelength Antenna of #10 AWG Copper Wire at 3 MHz.

Number of Segments	*Variation in jX from Standard (Ω)*	*Resonant Length of Antenna (in)*
5	-0.180	958.202 @ -0.001 Ω *jX*
10	-0.057	958.066 @ -0.001
15	-0.023	958.030 @ -0.001
20	-0.009	958.015 @ -0.000
*25	-0.001	958.004 @ -0.001
30	-0.008	958.012 @ +0.001
35	-0.019	958.013 @ +0.001

*Standard: #10 AWG copper wire quarter-wavelength vertical over perfect ground at 25 segments per quarter wavelength.

tenna length. Table 1 shows a representative sample of the variation in question at 3 MHz using #10 copper wire. Let us use 25 segments per quarter wavelength as the standard, resonating the quarter wavelength antenna to ±0.001 Ω jX. The second column shows the reactance variation as the number of segments is varied, while the third column shows the length of the antenna if resonated within the same limits as the standard. The figure of 25 segments per quarter wavelength was selected because results converged within ±0.01 Ω jX between 20 and 30 segments per quarter wavelength.

The length in familiar units of a quarter-wavelength of a radio wave in free space at a given frequency accords with the familiar equations,

$$L_{(\mathrm{ft})} = 245.8928 / F_{(\mathrm{MHz})} \text{ or}$$
$$L_{(\mathrm{in})} = 2950.7136 / F_{(\mathrm{MHz})} \quad \text{Eq 1}$$

where $L_{(\mathrm{ft})}$ is the quarter wavelength in feet, $L_{(\mathrm{in})}$ is the length in inches, and $F_{(\mathrm{MHz})}$ is the frequency in MHz. Constants are based on the value of the speed of electromagnetic radiation used in implementations of *NEC-2*. Calculation of the length of a half wavelength of a radio wave in free space would use numeric constants exactly twice the values in Eq 1.

Models of resonant real antennas composed of elements, ordinarily with a circular cross section of measurable dimension, will be shorter. The shortening factor, K_T, for single quarter wavelength verticals over perfect ground is simply

$$K_T = L_{(\mathrm{Model})} / L_{(\mathrm{Ideal})} \quad \text{Eq 2}$$

where $L_{(\mathrm{Model})}$ is the resonant length of the modeled antenna and $L_{(\mathrm{Ideal})}$ is the length of the ideal antenna calculated in Eq 1. Apart from any other method of obtaining a value for K_T, we can always find a value for K_T at a specific frequency for a specific antenna element material in this manner. Since modeled antennas ordinarily specify the material from which the antenna is constructed, K_T includes the combined effects of both end effects and of material effects.

Resonating antenna models to a suitable degree of precision can be a tedious task, especially if one surveys numerous materials at many frequencies. There is an alternative. First, for a given antenna element material, derive from the modeling program values of K_H and K_L at the frequency extremes of the range of interest. Then, for any specific frequency within the range F_H and F_L, corresponding to the limiting values K_H and K_L, calculate K as follows:

$$K_T = K_H + \left[\left(\log\frac{F_H}{F}\right)^{EE} \bullet (K_L - K_H)\right] \quad \text{Eq 3}$$

where K_T is the antenna shortening factor, K_H is the value of K of a given diameter element at its highest frequency, K_L is the value of K of a given diameter antenna element at its lowest frequency, F_H is the highest frequency (for K_H), F_L is the lowest frequency (for K_L), and F is any frequency in the range of interest.

The exponent, EE, also varies with frequency.

$$EE = 0.0333[(F/3) - 1] + 0.61 \quad \text{Eq 4}$$

for the frequency range 3 to 30 MHz. The value of EE is approximately linear from 3 to 30 MHz. It ranges from 0.61 to 0.91. Eq 4 simply calculates proportional parts for the value of the exponent.

The values of K_T returned by the equations are quite precise, relative to *NEC-2* models of quarter-wavelength monopoles over perfect ground, using 25 segments and brought to resonance. The maximum variance from values derived from the models is about 0.03% with the largest diameter materials, and closer with all other antenna element sizes. This variance sets the practical limits of accuracy of this calibration scheme in terms of element diameters.

Table 2—Sample Calculated Lengths of 7-Half-Wavelength Dipoles at 3 MHz.

Antenna element diameter (in)	*Length of a 7-Half-Wavelength Dipole (feet)*
Copper wire antennas:	
#18: 0.0403	1141.48
#16: 0.0508	1141.76
#14: 0.0641	1141.94
#12: 0.0808	1142.07
#10: 0.1019	1142.14
Aluminum antennas:	
0.125	1141.83
0.25	1141.89
0.5	1141.67
0.75	1141.43
1.0	1141.24
1.25	1141.03
1.5	1140.88
1.75	1140.71
2.00	1140.58

K_M, K_E, *and Long Antennas*

Separating the factors K_M and K_E from the overall value of K_T is straightforward. By choosing resonated antennas of multiple quarter wavelengths, one may derive another set of values for the overall shortening effect. Let us distinguish the two values of K_T by calling the quarter wave value K_{QW} and the multi-quarter wave value K_{TQ}. Let us also assume that the end effect appears only on the last quarter wavelength section, while the material effect appears on every quarter wavelength section. We may now derive the values of K_M and K_E by solving the resultant simultaneous equations, which reduce to the following (where n is the number of quarter wavelengths in the longer antenna):

$$K_E = \frac{(n-1) K_{QW} L_{(\mathrm{Ideal})}}{K_{TQ}\left(n L_{(\mathrm{Ideal})}\right) - K_{QW} L_{(\mathrm{Ideal})}} \quad \text{Eq 5}$$

and

$$K_M = \frac{K_{QW}}{K_E} \quad \text{Eq 6}$$

We may now use these values of K_M and K_E to calculate the lengths of multiple quarter-wavelength antennas (where n is the number of quarter wavelengths):

$$L = (n-1) K_M L_{(\mathrm{Ideal})} + K_E K_M L_{(\mathrm{Ideal})} \quad \text{Eq 7}$$

The length, L, will be in the units of choice, usually dependent upon the choice of units for the length of the modeled antennas at 3 and 30 MHz.

Dipoles in free space will be exactly twice the length of the corresponding verticals over perfect ground.[4] Values of K_T, K_E, and K_M will be the same, but the value of $L_{(\mathrm{Ideal})}$ will be twice the value used for quarter wavelength antennas in Eq 7. Alternatively, one may recast the equations to reflect preferred basic antennas and units of measure.

The initial choice of n, the number of wavelengths in the longer antenna up-on which the calculation of K_E and K_M are based in Eqs 5 and 6, will influence the direction of error in the final calculation of long antennas. If n is small, for example, 3, then calculated values of antennas with more quarter or half wavelengths will be short. If n is large, say, 7, then calculated values of antennas with fewer quarter or half wavelengths will be long. Using each of these values of n and checking antennas with five quarter or half wavelengths yields errors under 0.05% using either technique, relative to *NEC-2* models.

K_M reflects predominantly material

Program Listing

```
10 REM     file "KNEC.BAS"
20 COLOR 11,1,3:CLS:X$=STRING$(79,32)
25 REM     Option selection page
30 LOCATE 1,16:PRINT"Calculation of K, the Antenna Shortening Factor,":LOCATE 2,18:PRINT"and
Vertical and Horizontal Antenna Lengths":LOCATE 3,30:PRINT"L. B. Cebik, W4RNL":PRINT
40 PRINT " This program calculates values of the antenna shortening factor and antenna
lengths, including quarter-wave verticals over perfect ground, half-wavelength dipoles
in free space, and long wire vertical and horizontal antennas. The"
50 PRINT " frequency limits are 3 to 30 MHz. All dimensions are calibrated to NEC-2
antenna models. Available materials are AWG #18 to AWG #10 copper wire and aluminum rod or tubing
from 0.125 to 2.0 inch diameters.":PRINT
60 PRINT " Select one of the following two options:":PRINT:PRINT " <A> A table of antenna
lengths from 1 to 7 half wavelengths.":PRINT
70 PRINT " <B> A table of values of K, along with the lengths of quarter-wavelength vertical
and half-wavelength horizontal antennas."
80 Z$=INKEY$:IF Z$="a" OR Z$="A" THEN 90 ELSE IF Z$="b" OR Z$="B" THEN 470 ELSE 80
85 REM     Option A calculations
90 CLS:LOCATE 1,8:PRINT "Tables of Horizontal Antenna Lengths from 1 to 7 Half
Wavelengths":PRINT:VV$="###.##":WW$="####.##":YY$=STRING$(3,32)
100 PRINT " AWG Wire sizes are copper; decimal wire diameters are aluminum. Quarter-wave
vertical antennas over perfect ground are 1/2 the values shown where Xn = the number of quarter
wavelengths.
110 LOCATE 6,1:INPUT " Enter the frequency of interest in MHz   ",F
120 IF F<3 OR F>30 THEN LOCATE 6,1:PRINT X$:GOTO 500
130 WLF=983.5712/F
140 LOCATE 6,1:PRINT X$:LOCATE 6,1:PRINT" Frequency:";F;"MHz";:LOCATE 6,58:PRINT"Wavelength:
";:PRINT USING "###.##";WLF;:PRINT" ft"
150 LOCATE 7,25:PRINT "Number of half-wavelengths; Length in feet"
160 PRINT " Wire size";:PRINT YY$;:PRINT " X 1   X 2   X 3   X 4   X 5    X 6    X7"
170 FOR A=1 TO 14
180 IF A=1 THEN 200 ELSE IF A=2 THEN 210 ELSE IF A=3 THEN 220 ELSE IF A=4 THEN 230 ELSE IF A=5
THEN 240 ELSE IF A=6 THEN 250 ELSE IF A=7 THEN 260 ELSE IF A=8 THEN 270 ELSE 190
190 IF A=9 THEN 280 ELSE IF A=10 THEN 290 ELSE IF A=11 THEN 300 ELSE IF A=12 THEN 310 ELSE IF
A=13 THEN 320 ELSE IF A=14 THEN 330
200 W$=" #18-0.0403":LQL=959.435:LQH=95.335:LTL=6848.87:LTH=684.82:GOTO 340
210 W$=" #16-0.0508":LQL=959.183:LQH=95.252:LTL=6850.53:LTH=684.8:GOTO 340
220 W$=" #14-0.0641":LQL=958.885:LQH=95.154:LTL=6851.67:LTH=684.768:GOTO 340
230 W$=" #12-0.0808":LQL=958.478:LQH=95.04799:LTL=6852.43:LTH=684.699:GOTO 340
240 W$=" #10-0.1019":LQL=958.001:LQH=94.931:LTL=6852.81:LTH=684.618:GOTO 340
250 W$=" 0.125 in. ":LQL=957.22:LQH=94.807:LTL=6850.95:LTH=684.45:GOTO 340
260 W$=" 0.25 in. ":LQL=955.36:LQH=94.35:LTL=6851.31:LTH=684.082:GOTO 340
270 W$=" 0.50 in. ":LQL=952.85:LQH=93.734:LTL=6850.05:LTH=683.55:GOTO 340
280 W$=" 0.75 in. ":LQL=951.03:LQH=93.275:LTL=6848.59:LTH=683.144:GOTO 340
290 W$=" 1.00 in. ":LQL=949.58:LQH=92.898:LTL=6847.46:LTH=682.82:GOTO 340
300 W$=" 1.25 in. ":LQL=948.31:LQH=92.575:LTL=6846.2:LTH=682.555:GOTO 340
310 W$=" 1.50 in. ":LQL=947.22:LQH=92.292:LTL=6845.25:LTH=682.332:GOTO 340
320 W$=" 1.75 in. ":LQL=946.22:LQH=92.038:LTL=6844.28:LTH=682.14:GOTO 340
330 W$=" 2.00 in. ":LQL=945.3:LQH=91.81199:LTL=6843.45:LTH=681.983:GOTO 340
340     Q=2950.7136#:LQW=Q/F:LQWH=Q/30:LQWL=Q/3:KQH=LQH/LQWH:KQL=LQL/LQWL:KTH=LTH/
(3*LQWH):KTL=LTL/(3*LQWL)
350 EE=(((F/3)-1)*.0333333)+.61:KQW=KQH+((.4343*LOG(30/F))^EE)*(KQL-KQH)
360 KTQ=KTH+((.4343*LOG(30/F))^EE)*(KTL-KTH)
370 LQ=KQW*LQW:LT=KTQ*(3*LQW):KE=(6*LQ)/(LT-LQ):KM=KQW/KE:PRINT W$;:PRINT   YY$;
380 FOR B=1 TO 7
390 BB=B-1:LD=(2*LQW)/12:LL=((BB*LD)*KM)+(KQW*LD)
400 IF B<6 THEN PRINT USING VV$;LL;:PRINT YY$;
410 IF B>5 AND B<7 THEN PRINT USING WW$;LL;:PRINT YY$;
420 IF B=7 THEN PRINT USING WW$;LL
430 NEXT
440 NEXT
450 PRINT:PRINT " <Print Screen> for hard copy";YY$;YY$;"<A>nother run, <V>alues of K, or
<Q>uit";
```

```
460 Z$=INKEY$:IF Z$="a" OR Z$="A" THEN 90 ELSE IF Z$="v" OR Z$="V" THEN 470 ELSE IF Z$="Q" OR
Z$="q" THEN 810 ELSE 460
465 REM         Option B Calculation
470 CLS:LOCATE 1,16:PRINT "Calculation of K, the Antenna Shortening Factor"
480 LOCATE 3,1:PRINT" KT is the total shortening factor. KM is the shortening factor due to
element material. KE is the shortening factor due to end effect. Values calibrated to"
490 LOCATE 5,1:PRINT" NEC-2 models for 3-30 MHz. AWG sizes are copper; decimal sizes are
aluminum."
500 LOCATE 7,1:INPUT " Enter the frequency of interest in MHz   ",F
510 IF F<3 OR F>30 THEN LOCATE 7,1:PRINT X$:GOTO 500
520 WLF=983.5712/F
530 LOCATE 7,1:PRINT X$:LOCATE 7,1:PRINT" Freq:";F;"MHz";:LOCATE 7,20:PRINT
"Wavelength:";:PRINT USING "###.##";WLF;:PRINT" ft":LOCATE 7,48:PRINT"1/4-WL Vertical";:LOCATE
7,65:PRINT"1/2-WL Dipole"
540 LOCATE 8,2:PRINT"Wire Size":LOCATE 8,20:PRINT"KT":LOCATE 8,30:PRINT "KM":LOCATE
8,40:PRINT"KE":LOCATE 8,52:PRINT"L (ft)":LOCATE 8,67:PRINT"L (ft)"
550 FOR A=1 TO 14
560 IF A=1 THEN 580 ELSE IF A=2 THEN 590 ELSE IF A=3 THEN 600 ELSE IF A=4 THEN 610 ELSE IF A=5
THEN 620 ELSE IF A=6 THEN 630 ELSE IF A=7 THEN 640 ELSE IF A=8 THEN 650 ELSE 570
570 IF A=9 THEN 660 ELSE IF A=10 THEN 670 ELSE IF A=11 THEN 680 ELSE IF A=12 THEN 690 ELSE IF
A=13 THEN 700 ELSE IF A=14 THEN 710
580 W$=" #18-0.0403":LQL=959.435:LQH=95.335:LTL=6848.87:LTH=684.82:GOTO 720
590 W$=" #16-0.0508":LQL=959.183:LQH=95.252:LTL=6850.53:LTH=684.8:GOTO 720
600 W$=" #14-0.0641":LQL=958.885:LQH=95.154:LTL=6851.67:LTH=684.768:GOTO 720
610 W$=" #12-0.0808":LQL=958.478:LQH=95.04799:LTL=6852.43:LTH=684.699:GOTO 720
620 W$=" #10-0.1019":LQL=958.001:LQH=94.931:LTL=6852.81:LTH=684.618:GOTO 720
630 W$=" 0.125 in. ":LQL=957.22:LQH=94.807:LTL=6850.95:LTH=684.45:GOTO 720
640 W$=" 0.25 in. ":LQL=955.36:LQH=94.35:LTL=6851.31:LTH=684.082:GOTO 720
650 W$=" 0.50 in. ":LQL=952.85:LQH=93.734:LTL=6850.05:LTH=683.55:GOTO 720
660 W$=" 0.75 in. ":LQL=951.03:LQH=93.275:LTL=6848.59:LTH=683.144:GOTO 720
670 W$=" 1.00 in. ":LQL=949.58:LQH=92.898:LTL=6847.46:LTH=682.82:GOTO 720
680 W$=" 1.25 in. ":LQL=948.31:LQH=92.575:LTL=6846.2:LTH=682.555:GOTO 720
690 W$=" 1.50 in. ":LQL=947.22:LQH=92.292:LTL=6845.25:LTH=682.332:GOTO 720
700 W$=" 1.75 in. ":LQL=946.22:LQH=92.038:LTL=6844.28:LTH=682.14:GOTO 720
710 W$=" 2.00 in. ":LQL=945.3:LQH=91.81199:LTL=6843.45:LTH=681.983:GOTO 720
720 Q=2950.7136#:LQW=Q/F:LQWH=Q/30:LQWL=Q/3:KQH=LQH/LQWH:KQL=LQL/LQWL:KTH=LTH/
(3*LQWH):KTL=LTL/(3*LQWL)
730 EE=(((F/3)-1)*.0333333)+.61:KQW=KQH+((.4343*LOG(30/F))^EE)*(KQL-KQH)
740 KTQ=KTH+((.4343*LOG(30/F))^EE)*(KTL-KTH)
750 LQ=KQW*LQW:LT=KTQ*(3*LQW):KE=(6*LQ)/(LT-LQ):KM=KQW/KE:IF KM>.9999 THEN KM=.9999
760 V=KQW*(245.8928/F):D=V*2:Y$=STRING$(5,32):V$=STRING$(9,32):U$=".####":
T$="###.##":S$=STRING$(6,32)
770 PRINT W$;:PRINT S$;:PRINT USING U$;KQW;:PRINT Y$;:PRINT USING U$;KM;:PRINT Y$;:PRINT USING
U$;KE;:PRINT V$;:PRINT USING T$;V;:PRINT V$;:PRINT USING T$;D
780 NEXT
790 PRINT:PRINT " <Print Screen> for hard copy";V$;"<A>nother run, <W>ire lengths, or
<Q>uit";
800 Z$=INKEY$:IF Z$="a" OR Z$="A" THEN 470 ELSE IF Z$="Q" OR Z$="q" THEN 810 ELSE IF Z$="w"
OR Z$="W" THEN 90 ELSE 800
810 END
```

loss affects on antennas, which are in part dependent upon the surface area (and hence, the diameter) of the element. However, figures derived from modeled values for lossless elements show a residual value for K_M which appears to increase with antenna element diameter. However, the amount by which the lossless K_M is less than 1.0 is very small. Consequently, the source of the residual K_M cannot be definitively traced to antenna factors. Nor can it be definitely ascribed either to the limits of the modeling program or to the limits of this calculation scheme.

Those unfamiliar with the variation of length with various other antenna properties may find a surprise in Table 2. The Table lists the calculated lengths (confirmed by *NEC-2* models) of seven half-wavelength dipoles in free space for common copper wire and aluminum sizes. Although it is unlikely that anyone will build such an antenna from 2-inch aluminum tubing, the list of values is instructive with respect to the interaction of material losses and end effect upon antenna length. Note that the modeled antenna length increases with the copper wire size and even with the first two sizes of aluminum. Although the table has intentionally used a frequency where the length variation curve is most dramatic, a similar, if lesser, variation occurs throughout the HF region.

The *GWBASIC* Program

The utility program in *GWBASIC* provides values of K_T, K_M and K_E in the 3- to 30-MHz frequency range for a selection of common amateur antenna materials ranging from #18 AWG wire to 2-inch diameter tubing. In addition, the program provides the lengths of dipole antennas in free space from one to seven half wavelengths long. (Vertical antennas over perfect ground will be exactly half the lengths of the dipoles for a number of quarter wavelengths equal to the number of half wavelengths in the corresponding dipole.) The program makes use of Eqs 1 through 7. Key to the program is the development of limiting values of resonant antenna lengths from *NEC* models at 3 and 30 MHz for each material. For the sake of linear programming, the limiting values appear in the program listing twice, a set within the calculation FOR-NEXT loops for each of the output options.

Program output is a pair of tables, each for 14 antenna materials. If a particular material is missing from the list, you can easily modify one or more of the information lines to include it. Or, you can interpolate with quite reasonable accuracy values for intermediate materials. Option A lists the lengths of short to long dipoles. Option B lists values of K_T, K_M and K_E for each material. For reference (and because screen space was available), corresponding quarter-wave monopole (over perfect ground) and half-wavelength dipole (in free space) lengths are given for each material. For wire smaller than #18, the longest given length will very likely suffice for all antenna construction purposes.

Verticals and dipoles over real ground, of course, will vary in resonant length according to ground conditions and (for dipoles) height above ground. Table 3 lists the modeled resonant lengths for an aluminum 0.5-inch diameter dipole at 28.5 MHz at 4-foot increments from 16 to 36 feet above medium ground. This same type of data has often been shown as variations in resistance and reactance of a fixed length dipole as height is increased. Displaying it as "effective K_T" is simply another useful perspective on the data.[5]

Fortunately for antenna builders, these basic antennas exhibit acceptably low reactances across a reasonably wide frequency range. Hence, a knowledge of precise values of K_T is not needed for successful antenna building. However, for some investigations, the values of K_T yielded by the utility program may prove useful in sorting promising from unpromising formulations of experimental or modeling trends.

Table 3—The Effective K_T of a 0.5-inch Diameter Dipole Over Real Ground at Heights from Less Than 0.5 Wavelength to Greater Than 1 Wavelength.

Values of "Effective K_T" for Dipoles Over Real Ground

Height (feet)	*Resonant Antenna Length (feet)*	*Effective K_T*	*Feedpoint Resistance (Ω)*
Free Space	16.46	0.9535	72.0
16	16.64	0.9640	74.9
20	16.48	0.9546	63.4
24	16.34	0.9465	68.3
28	16.42	0.9509	77.3
32	16.54	0.9582	76.3
36	16.52	0.9567	68.9

Notes:
1. All lengths and feedpoint resistances are rounded for ease of reading. Antenna lengths were varied until resonance (±0.01 Ω reactance) was achieved.
2. Feedpoint resistance is listed to demonstrate the absence of a simple direct correlation with antenna length or K_T.

The goal of this project is neither to supplant traditional graphical representations of *K* nor to force upon antenna constructors an unnecessary level of precision. As noted earlier, the functions of the level of precision used in the development of the BASIC program were to trace value curves accurately and to compare values produced by the modeling and the calibrated computational scheme. For most wire antennas with loops at the ends, the traditional 5% total end-effect reduction will continue to perform adequately for real antennas amid the host of surrounding objects within which we raise them.

For many, this program may be only a curiosity, especially in view of the fact that it does not itself evaluate the accuracy of *NEC-2* models relative to real antennas. However, it can shorten the trial and error time required to zero in on resonant antenna models of various lengths. In the absence of a modeling program such as *NEC-2*, the BASIC program yields equally usable resonant antenna lengths for free space or over perfect ground, as may be apt. Moreover, it gives some further insight into antenna length as a combined function of end effects and material losses. The frequency range and the selection of materials used in the program may be modified by the user. Indeed, the program reference values can be calibrated to any present or future version or implementation of *NEC*. Finally, the program may also serve as one kernel within a larger program within which values of *K*, the antenna shortening factor, may play a significant role.

The program file (knec.bas) can be downloaded from the ARRL BBS (860-594-0306) or via the Internet from **http://www.arrl.org/files/qex** or **ftp://ftp.arrl.org/pub/qex**.

Notes

[1] *The Radio Amateur's Handbook*, 24th Ed (West Hartford: ARRL, 1947), pp 194-195. I am indebted to Michael Tracy, KC1SX, of the ARRL Technical Information Service, who added his efforts to my own in the search for the source of the *Handbook* graph. Nothing in *QST* or other League publications has shown itself. If anyone has knowledge of an authoritative source for the graph, I would appreciate correspondence or email.

[2] Schelkunoff, S. A., *Electromagnetic Waves* (New York: Van Nostrand, 1943), p 465. For example, see the treatment of end "caps." At best, one can draw some inferences from sources such as this, but they do not show the definitive development of the Handbook graph. See Note 1. The designation of the shortening factor as "K" is also problematic in connection with transmission-line analogy calculations, which often use *K* to designate the characteristic impedance of a biconical antenna and sometimes of its cylindrical counterparts.

[3] For information on experimental work refining the antenna shortening factor, see Belrose, John S., "VLF, LF, and MF Antennas," *The Handbook of Antenna Design, Vol 2*, Editors Rudge, Milne, Oliver and Knight (London: Peter Peregrinus, 1983), pp 564-565.

[4] The exactitude is a function of the manner in which *NEC* calculates vertical antennas over perfect ground.

[5] All models were constructed on *EZNEC* 1.06, available from Roy Lewallen, W7EL. Although the work could have been done on virtually any version of *NEC-2*, the wire dimension manipulation features of *EZNEC* made it especially apt to the reiterative nature of this task. I recommend that, if the attached program is to be used regularly in conjunction with a version of *NEC*, then the program should be calibrated for the specific version of *NEC* used by confirming the values of, or developing replacement values for, lines 200-330 and 580-710 of the program. □□

Designing a Wein-Bridge Oscillator

A tried-and-true low-distortion sine-wave oscillator made simple.

By Parker R. Cope, W2GOM

The venerable Wein bridge has been the heart of low-frequency frequency meters for decades and continues to be found in low-frequency applications. Its popularity derives from its simplicity and its use of common resistors and capacitors to determine the tuned frequency. These attributes lead to using the Wein bridge in an oscillator that can be designed by a knowledgeable technician or hobbyist. The procedures and equations used in designing the oscillator are given here. While the illustrative example described is for a three-decade audio oscillator, the design procedures can be used to design fixed-frequency subaudible tones for RF transmitter control or a source for an ultrasonic cleaner or solderer.

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The following theoretical discussions provide a basis for understanding the underlying mechanisms of the oscillator and selection of the parts used, but an empirical design is quite simple using the circuit shown in Fig 3:

1. Select a compatible lamp/BJT combination.
2. Build the amplifier with R1 of the feedback network disconnected from the collector.
3. Measure the voltage across the lamp and the current through it, and calculate the lamp's resistance, $R=V_{lamp}/I_{lamp}$.
4. Use a variable resistor for R_C that can be varied over a range of 2 to 5 times the lamp resistance.
5. Connect the feedback network and adjust R_C to produce an acceptable output.
6. The BJT should operate in its linear region, that is, not cut off or saturated. An oscilloscope can confirm linear operation. The amplifier can be assumed to be in its linear range when the peak instantaneous collector voltage ($dc+1.4{\times}V_{rms}$) is a volt or more less than the supply voltage.

In the oscillator described, oscillations occur at the frequency that causes the Wein bridge to be balanced. A bridge is said to be balanced when the voltage between two opposing arms is zero. The Wein bridge in Fig 1 is balanced when the voltage at points C and D are equal in magnitude and phase. That is, when voltage difference between C and D is zero. Balance exists when $X_{C1}=X_{C2}=R1=R2$ and $R_C=2R_S$. When $C1=C2=C$ and $R1=R2=R$, the frequency at which the bridge balances is expressed as:

$$f_o = \frac{1}{(6.28 \times C \times R)} \qquad \text{Eq 1}$$

where f_o is in Hz, C is in farads and R is in ohms. At f_o, the voltage E_C at point C is

$$E_C = \frac{V_{in} \times Z_2}{Z_1 + Z_2} = \frac{1}{3}\angle 0^\circ \qquad \text{Eq 2}$$

where

$$Z_1 = R_1 - jX_{C1} = R - jR = 1.414R\angle 45^\circ$$

$$Z_2 = \frac{1}{Y_2} = \frac{1}{G_2 + jB_2} = \frac{1}{G + jG} = \frac{1}{1.414G} = 0.707G\angle 45^\circ$$

$$\frac{Z_2}{Z_2 + Z_1} = \frac{0.707\angle 45^\circ}{(0.707 + 1.414)\angle 45^\circ} = \frac{1}{3}\angle 0^\circ$$

$$\frac{E_C}{V_{in}} = \frac{1}{3}\angle 0^\circ$$

For the voltage at D to equal the voltage at C, R_C must be $2R_S$.

The voltage at C is maximum when $X_C = R$, but more important for an oscillator is the phase change at f_o. At frequencies below f_o, the voltage at C lags the voltage at D; at frequencies above f_o, the voltage at C leads the voltage at D. The rate of change of phase with frequency is maximum at f_o.

If a dual potentiometer is used for R1 and R2, the bridge can easily tune over the entire audio range from 20 Hz to 20 kHz. Of course, with a three-decade tuning range, setting to a particular frequency will be a bit touchy. Adding fixed resistors in series with the pots to limit the resistance variation to 10:1 restricts the tuning range to a decade, so setting a particular frequency is much easier. However, covering the three decades in three ranges does require switching three sets of capacitors.

Selecting the components for the frequency-selective network shown in Fig 2 is straightforward. Choose convenient values for the pairs of capacitors C1A, C1B; C2A, C2B; and C3A, C3B; and using Eq 1 calculate the resistance needed for the desired frequency. The capacitors should have close tolerances and a dielectric with a stable temperature characteristic, such as polystyrene or NPO ceramics (foil, film or high-dielectric ceramic will suffice for all but extreme temperature environments). Table 1 is a compilation of the values needed to cover 20 Hz to 20 kHz in three ranges when the capacitors are 0.1 µF, 0.01 µF and 1000 pF. Since R must have a 10:1 ratio, finding the appropriate shunt and series resistors for the potentiometer is a matter of calculating the resistance needed in shunt with the pot that produces 79.5 kΩ at maximum resistance and the resistance in series that limits the minimum resistance to 7.95 kΩ. If the potentiometer is a dual 100-kΩ unit, the shunt resistors are 389 kΩ and the series resistors are 7.95 kΩ. To allow for component tolerance and some frequency overlap, the shunt resisters are chosen to be 430 kΩ ±5% and the series resistors are 6.8 kΩ ±5%. Metal-film resistors are preferred, but carbon-composition resistors are tolerable in benign environments. The switching arrangement shown in Fig 2 uses three SPST switches, but of course, one 3-position rotary switch could be used.

Matching errors in R1 and R2 or C1 and C2 will cause the bridge gain to be less than 1/3 and require the amplifier gain to be somewhat greater than 3. The ganged pots used for R1 and R2 for a variable frequency oscillator need not be precision units; they only need to track. Single-turn carbon or ceramic elements are satisfactory, and absolute tolerance is not important if the individual units track well. Tuning capacitors should also be matched. Polypropylene

Table 1

Frequency	*switch position*	*C*	*R*
20 Hz	1	0.1 µF	79.5 kΩ
200 Hz	1	0.1 µF	7.95 kΩ
200 Hz	2	0.01 µF	79.5 kΩ
2 kHz	2	0.01 µF	7.95 kΩ
2 kHz	3	1000 pF	79.5 kΩ
20 kHz	3	1000 pF	7.95 kΩ

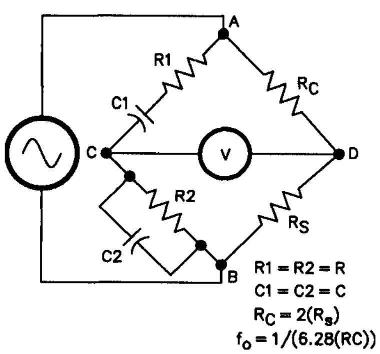

Fig 1—The Wein bridge is frequency sensitive and uses only resistors and capacitors.

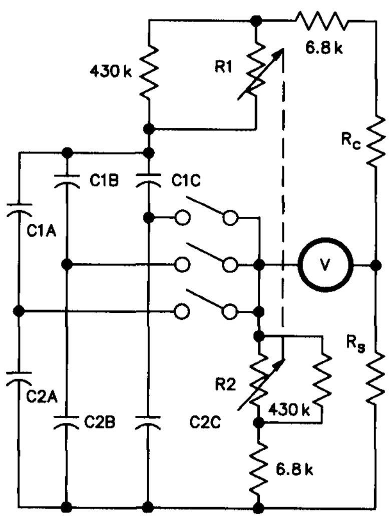

Fig 2—A simple switching arrangement that can control a three-decade tuning range.

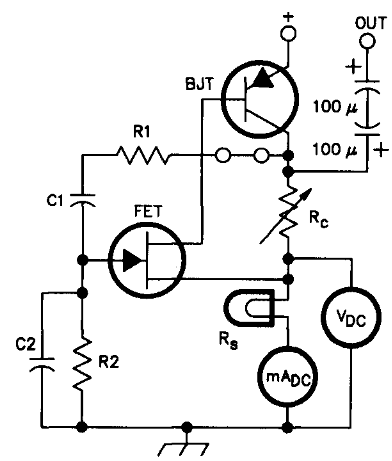

Fig 3—An oscillator can be designed empirically.

capacitors in 2% tolerances are an excellent choice, and 5% tolerances are satisfactory.

An oscillator based on the Wein bridge is shown in Fig 3. Stable oscillation requires a loop gain of exactly unity and a loop phase shift of zero degrees. That is, when the loop is opened at the input to the amplifier and a signal is applied to the input, the voltage at the output of the feedback network has exactly the same amplitude and phase as the input. If the loop gain is greater than unity, the voltage fed back will be greater than the input and oscillations will build up until the amplifier either saturates or cuts off. If the loop gain is less than unity, the voltage fed back will be less than the input and any excited oscillation will die out. A stable oscillation requires that the loop gain be held to exactly unity with changes in loading and supply-voltage variation. To achieve unity loop gain requires an amplifier with a voltage gain of 3 to compensate for the gain of 1/3 in the R-C legs of the bridge.

A hybrid JFET/PNP BJT amplifier comprised of an N-channel JFET (junction field effect transistor) and a PNP BJT (bipolar junction transistor) such as the one shown in Fig 3 has zero phase shift from JFET gate to BJT collector, and its input impedance is essentially infinite. The gain of the amplifier is determined by the JFET's g_{fs} (forward transconductance), the h_{FE} (current gain) of the BJT, and the ratio of R_C to R_S. When R_S is an incandescent lamp, the lamp's resistance increases as the rms current through the lamp increases. This decreases the ratio of R_C to R_S and reduces the amplifier's gain, tending to hold the output at a constant amplitude. Conversely, when the output decreases, R_S decreases, and the gain is increased, again tending to hold the output constant. Loading of the output, which shunts part the collector's ac signal, is also compensated. Current to the load decreases current to the lamp, which decreases its resistance and increases the gain, again tending to hold the output constant.

The response time of the lamp is in the order of hundreds of milliseconds, so its resistance is determined by the average rms value of several cycles of the oscillator frequency. At very low frequencies, below 10 Hz, the lamp's resistance can change with the instantaneous value of the current and cause distortion of the output.

The resistance of an incandescent lamp is neither constant nor a linear function of either the applied voltage or current. The lamp current, when operating at other than the rated voltage, can be expressed as:

$$\frac{I_{sp}}{I_{op}} = \left(\frac{V_{sp}}{V_{op}}\right)^{0.55} \qquad \text{Eq 3}$$

$$\frac{V_{sp}}{V_{op}} = \left(\frac{I_{sp}}{I_{op}}\right)^{1.82} \qquad \text{Eq 4}$$

where I_{sp} is the manufacturer's specified nominal current at the specified rms voltage, V_{sp}, and I_{op} is the current at the operating rms voltage, V_{op}.

Eq 4 can be rewritten to express the voltage across the lamp for various operating currents:

$$V_{op} = V_{sp} \bullet \left(\frac{I_{op}}{I_{sp}}\right)^{1.82} \qquad \text{Eq 5}$$

The resistance of the lamp at some arbitrary operating voltage can be obtained by manipulating Eqs 3 and 4 to produce:

$$R_{op} = \frac{V_{op}}{I_{op}} = \frac{V_{sp}}{I_{op}} \bullet \left(\frac{I_{op}}{I_{sp}}\right)^{1.82} \qquad \text{Eq 6}$$

Finding a suitable incandescent lamp for a desired amplifier operating point is a cut-and-try process, but a set of reiterative calculations can quickly zero in on an acceptable compromise. Table 2 is a compilation of several incandescent lamps and their calculated resistances when operated at 10 mA and 12 mA.

A 2N3906 is a good choice for the PNP BJT. The 2N3906 is specified to have a minimum h_{FE} of 100 at 10 mA of collector current. If the BJT's collector current is to be 10 mA_{av} (10-mA dc), the base current of the BJT, which is also the drain current of the JFET, must be 0.1 mA.

The following relationships of the parameters of a JFET are given in the book *Designing With Field Effect Transistors* by Evans:

$$\frac{I_D}{I_{DSS}} = \left(\frac{1 - V_{gs}}{V_{off}}\right)^2 \qquad \text{Eq 7}$$

where I_{DSS} is drain current with the gate shorted to the source, I_D is the drain current with a gate-to-source voltage of V_{gs}, and V_{off} is the gate-to-source voltage needed to reduce the drain current to zero.

Expressions for V_{off} and V_{sg} can be obtained by manipulating Eq 7 to produce:

$$V_{off} = \frac{V_{gs}}{\left[1 - \left(\frac{I_D}{I_{DSS}}\right)^{0.5}\right]} \qquad \text{Eq 8}$$

$$V_{sg} = V_{off} \bullet \left[1 - \left(\frac{I_D}{I_{DSS}}\right)^{0.5}\right] \qquad \text{Eq 9}$$

The data sheet for the 2N5457 N-channel JFET shows these typical values:

I_{DSS} = 3 mA

I_D = 0.1 mA at V_{gs} = 2.5 V (gate negative with respect to the source). Based on these typical values, V_{off} can be calculated with Eq 8 to yield:

$$V_{off} = \frac{2.5}{\left[1 - \left(\frac{0.1\text{ mA}}{3\text{ mA}}\right)^{0.5}\right]} = 3.06\text{V}$$

The V_{gs} needed to produce a drain current of 0.1 mA can be taken from the data sheet or calculated with Eq 9 to yield:

$$V_{gs} = 3.06\left[1 - \left(\frac{0.1\text{ mA}}{3\text{ mA}}\right)^{0.5}\right] = 2.5\text{V}$$

The values of I_{DSS} and V_{gs} for a particular N-channel JFET can be established with two simple tests that require minimal test equipment: a multimeter, a resistor in the range of

Table 2

Lamp No	V_{sp}	I_{sp}	R_{sp}	R_{op} *10 mA*	R_{op} *12 mA*
47	6.3	0.150	42	4.6	5.3
1813	14.4	0.100	144	21.8	25.3
1820	28	0.100	280	42.4	49.2
7367	10	0.04	250	80.2	93.1
7371	12	0.04	300	96.3	111.8
2316	18	0.04	450	144.4	167.7
1819	28	0.04	700	224.6	260.8
8099	18	0.02	900	509.7	592.0
1835	55	0.05	1100	294.0	341.3

several thousand ohms (2.4 kΩ is a good value) and a battery or voltage source between 6 V and 24 V.

1. Connect the resistor between the JFET source and the negative side of the battery.

2. Connect the gate to the negative side of the battery.

3. Connect the drain to the positive side of the battery.

4. Measure the voltage across the resistor, V_{gs}, and calculate I_D through the resistor, $I_D = \frac{V_{gs}}{R}$.

5. Short the resistor and measure the drain current, I_{DSS}.

The h_{FE} of the BJT can also be determined easily:

1. Connect the emitter to the positive side of the battery and connect a resistor from base to ground that will allow about 0.1 mA to flow:

$$R = \frac{V_{bat} - 0.6}{1 \bullet 10^{-4}}.$$

2. Measure collector current from the collector to the negative side of the battery. Calculate h_{FE} from $h_{FE} = \frac{I_C}{I_B}$.

With V_{gs}, I_D and I_{DSS} established, V_{off} and g_{fs} can be calculated. With V_{off} and I_{DSS} known, the V_{gs} needed for a particular I_D can be calculated with Eq 9. The current in R_S is the sum of the BJT collector current and the JFET drain current. As a "first cut," assume $I_D = 0.1$ mA and $V_{gs} = 2.5$ V for the JFET and $h_{FE} = 100$ for the BJT. The current in R_s, the lamp current, is 10.1 mA. The required R_S then is 248 Ω. From Table 2, the #1819 incandescent lamp is found to have the desired resistance at a current between 10 mA and 12 mA. A different JFET/BJT choice or different output current would probably dictate a different lamp choice.

The operating point of the amplifier can be computed by the simultaneous solution of the equations for V_{OP} and V_{sg}. The computation shows the operating point to be $I_D = 0.10474$ mA, and the lamp current to be 10.58 mA, which corresponds to a lamp resistance of $R_{op} = 235$ Ω. The dc operating point is $V_{sg} = 2.49$ V and $I_D = 0.1047$ mA. The current in the lamp, I_{op}, is then 10.58 mA, and the lamp resistance, R_{op}, is calculated to be 235 Ω.

The instantaneous current in the lamp is the sum of the dc current plus an ac component. The rms value of the dc and ac currents is not the simple sum of I_{dc} and I_{ac} but is the root of the mean square of the instantaneous value. The rms value can be found by sampling the current periodically, say every 90° of the ac component, squaring the value of the sampled value, finding the mean (the average) of the squares and then taking the square root of the mean. For example, the rms value of a dc current of 10.6 mA and a peak ac component of 3 mA, is obtained as follows:

Degrees	*Value*	*Squared*
0	10.6	112.36
90	13.6	184.96
180	10.6	112.36
270	7.6	57.76
	Mean square	467.44/4 = 116.86
	rms	10.81

The 3-mA ac component increases the rms value from 10.6 mA to 10.81 mA and the lamp resistance from 235.6 to 239 Ω. The voltage on the lamp is 2.50 V dc and 3.25 V pk (1.43 $V_{p\text{-}p}$).

The gain of the amplifier is dependent on the effective transconductance G_m, $G_m = g_{fs} \times h_{FE}$, of the JFET/BJT pair. Since forward transconductance is defined as the change in drain current for a change in gate-to-source voltage, the g_{fs} is effectively multiplied by the h_{FE} of the BJT. According to Evans the g_{fs} of the JFET is expressed as:

$$g_{fs} = \frac{2 I_D}{V_{off} - V_{gs}} \qquad \text{Eq 10}$$

Given: I_D=0.1047 mA, V_{off}=3.06, V_{gs}=2.48:

$$g_{fs} = \frac{2 \times 0.1 \times 10^{-3}}{3.06 - 2.48} = 3.61 \times 10^{-4}$$

and the effective transconductance G_m is:

$$G_m = g_{fs} \times h_{FE} \qquad \text{Eq 11}$$

when h_{FE}=100:

$$G_m = 3.39 \times 10^{-4} \times 100 = 3.61 \times 10^{-2}$$

The voltage gain from the gate to the source can be expressed as:

$$VG_{gs} = \frac{G_m R_s}{G_m R_s + 1} \qquad \text{Eq 12}$$

Therefore, with a lamp resistance of 239 Ω, the voltage gain from gate to source is:

$$VG_{gs} = \frac{3.61 \times 10^{-2} \times 239}{3.61 \times 10^{-2} \times 239 + 1} = 0.896$$

The voltage gain of the amplifier is:

$$VG = VG_{gs} \times VG_{cg} \times VG_{sc} \qquad \text{Eq 13}$$

where

VG_{gs} = voltage gain from gate to source, 0.896

VG_{cg} = voltage gain from collector to gate, 1/3

VG_{sc} = voltage gain from source to collector.

The voltage gain of the amplifier necessary for oscillation is unity, and *VG* of Eq 13 is unity. The voltage gain from source to collector, VG_{sc}, required for oscillations is calculated from Eq 13:

$$VG_{sc} = \frac{1}{0.333 \times 0.896} = 3.34$$

The voltage gain from source to collector is determined by the ratio of R_c and R_s. It can be shown that

$$\frac{V_C}{V_S} = 1 + \frac{R_C}{R_S} \qquad \text{Eq 14}$$

where V_C is the the voltage from collector to ground, and V_S is the voltage from source to ground.

Manipulating Eq 14 shows:

$$\frac{R_C}{R_S} = \frac{V_C}{V_S - 1} \qquad \text{Eq 15}$$

$$R_C = R_S \times \frac{V_C}{V_S - 1} \qquad \text{Eq 16}$$

The resistance of the #1819 lamp, R_S, has been established as 239 Ω, which requires an R_C of 559 Ω. R_C is the critical component in that it determines the value of R_S that will be forced to be achieved. In effect, R_C determines the peak current and the peak voltage across R_S.

The power supply voltage should be at least 1 volt greater than the peak collector voltage to avoid collector saturation of the BJT. Therefore, with 3.25 V_p across R_S and 7.6 V_p across R_C, a supply voltage greater than 10.85 V is required. If R_C is larger than 559 Ω, R_S will be forced to be greater than 239 Ω and the collector current will be greater than 3 mA peak. If R_C is less than 559 Ω, R_S will be less than 239 Ω and the collector current will be less than 3 mA.

The load should be capacitively coupled for dc isolation. Any nonpolarized capacitor with adequate capacitance will serve for coupling. Electrolytics can be connected in series with opposing polarity to act as a nonpolarized capacitor.

Ac current in the load will decrease the ac current in R_S and thus decrease R_S, which increases the gain of the amplifier and compensates for the current diverted to the load. The minimum load should be greater than 1 kΩ so that the peak current demands on the BJT are not excessive. High BJT currents mean high JFET drain currents that can exceed the capability of the JFET, force the drain current to zero and cut off the positive peaks of the output voltage. The coupling capacitor should be greater than 100 µF to pass 20 Hz with minimum loss and minimum reactive loading on the amplifier.

When measuring the ac voltage, the ac voltmeter should not respond to a dc voltage. While most ac voltmeters are capacitively coupled and do not respond to dc, not all are. A way to check is to set the meter to the ac mode and connect it to a dc voltage. If the meter indicates a voltage, it will not give a true indication of the ac component of an ac voltage biased with a dc voltage.

Conclusion

The Wein-bridge oscillator makes a good low-distortion audio source and can even be extended into the high ultrasonic range—to even 1 or 2 MHz. The circuit is very tolerant and forgiving because of the high negative feedback in the JFET source. With the design equations given, an oscillator can be designed that produces an extremely stable low-distortion signal.

About the Author

Presently semi-retired and doing consulting work on RF systems, Parker Cope, W2GOM, received a BSEE in 1950 and worked for Bendix Radio for 16 years before forming his own company, Scan Inc. Over the years he's worked for several companies including Numicon-Mosstype, Mohawk Developement Systems and Delco. Parker has taught radio and electronics for RETS in Baltimore, Maryland and Ivy Tech in Fort Wayne, Indiana. He currently holds 23 patents and disclosures, and he has had articles published in magazines such as EDN, 73 Magazine *and* Popular Electronics. *He has held an Advanced class license since 1950 (ex W3SGV).*

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The Mechanical Filter in HF Receiver Design

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By William E. Sabin, WØIYH

The high-quality HF communications receiver for AM, SSB, CW and narrowband data modes presents many interesting design problems in terms of sensitivity, strong signal immunity, frequency control/readout and selectivity. A major goal is to achieve a high ratio of performance to cost. This article discusses the circuit design aspects of a new generation of Collins mechanical filters in the IF section as a way to help realize these goals. The present discussion focuses on the analog receiver signal path.

The example filters to be discussed here are the Collins Low Cost Series (526-8634/5/6-010), made by the Filter Products Division of Rockwell International, Costa Mesa, CA. These filters are centered at 455.0 kHz and come in 3 dB/60 dB bandwidths of 0.5/2.0, 2.5/5.2 and 5.5/11 kHz. They have a maximum insertion loss of 6 dB. They use the newly perfected torsional mode design, which leads to performance/cost and performance/size ratios which are unprecedented in the long history of the IF mechanical filter.[1] Fig 1 conveys their impressive dimensions (1.25×0.5 ×0.25 inches). A wide range of other miniature torsional mode filters are also in the product line. The Collins Radio Company (now part of Rockwell International) started manufacturing mechanical filters in 1952.

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Receiver Topology

A brief review of the analog receiver system architecture will help to put the mechanical filter into perspective as a receiver component.

Fig 2(A) is a block diagram of an *archaic* HF general coverage receiver which uses a 455-kHz IF frequency. In order to avoid spurious responses which are from signals 910 kHz away (the image response) and others, notably the one only 227.5 kHz away (the fourth order "IF/2" response), multiple tuned circuits at signal frequency are needed. Even-order responses like the IF/2 may be somewhat reduced if *balanced* mixer circuits of some kind are used. But still, the gang-tuned and gang-switched preselector and local oscillator assembly is much too expensive and too inadequate for today's performance and cost requirements. Instead, the "upconversion" approach shown in Fig 2(B) is an improvement.

[1]Notes appear on page 20.

The implication and the cost burden of upconversion is that the frequency stability requirements for the very high frequency local oscillators (LO) mandate a synthesized, low phase-noise first local oscillator which is now digitally switched and tuned, and an equally stable second mixer fixed, or vernier tunable, LO. The state-of-the-art in synthesizer design and cost has advanced so dramatically that upconversion is now one accepted approach, even for low-cost receivers.

The tightly controlled and stable center frequency and passband of the mechanical filters make them especially desirable where synthesized, accurate local oscillators are used.

The same image and IF/2 responses mentioned above must now be rejected by the first IF filter. Its stopband requirements are therefore very stringent; 90-dB rejection, minimum, is needed in today's world of really big HF signals, coming from large, high-gain receiving antennas. An additional 100-MHz IF filter can be used to achieve the required rejection. More commonly, a more desirable IF filter is used by including a third, lower IF in the 10-MHz region. The spurious responses of the first mixer are controlled by a 1.6-30 MHz high-pass/low-pass filter or possibly by PIN diode-switched half-octave filters. The first IF filter helps to minimize the spurious responses of the second mixer. Fig 2(C) is the method usually found in practice today in high quality, narrowband HF receivers.

A low-cost alternative for the approach in Fig 2(B), occasionally used, would be an image-reducing second mixer. This can reliably add an additional 20 dB or perhaps 30 dB of image and IF/2 rejection, which may not be enough, at the relatively low cost of some additional mixer and LO circuitry.

And of course the various LO and IF frequencies in the multiple conversion receiver must be optimal in terms of harmonic intermodulation (IM) products (harmonics of the LOs combining with harmonics of the RF/IF to produce spurs within the IF passbands).[2,3,4] For this reason, the first IF is preferably greater than three times the highest signal frequency. But if a sufficiently high-level first mixer is employed, a ratio of greater than two-to-one may also be an acceptable approach (it does a better job of rejecting harmonic IM products). Also, internal "tweets" (various LO frequencies leaking into places where they don't belong) must also be minimized, therefore internal shielding and lead filtering are critical aspects of the design. An interesting example is when the 110-MHz LO, Fig 2(C), leaks into the first mixer. A vulnerability to a 10-MHz antenna signal occurs (IF feedthrough). A very important consideration is the distribution of gain and noise figure along the signal path. The high frequency IF filters preceding the mechanical filters are usually substantially wider (but see later discussion), and since AGC is usually produced only by signals which pass *through* the mechanical filter, the early stages are vulnerable to strong signals which are inside the wider band but outside the mechanical filter band. Therefore the early stages must have low gain, low noise figure and high intermodulation performance. These conflicting goals lead to some difficult tradeoffs. In other words, the desire to put the "knothole" close to the antenna is not realized in the upconversion receiver which is intended for *narrowband* modes.

A consequence of low front-end gain is that the gain *after* the mechanical filters must now be *increased*, and this can cause high levels of wideband noise to be generated, which degrades receiver sensitivity.[5] In a narrowband CW/FSK receiver, in particular, a second narrow filter downstream is often used to establish the required narrow noise bandwidth. This in turn can cause AGC loop design problems because of the time delays that the cascaded filters introduce. The second filter may be kept outside the AGC loop. Alternatively, an audio bandpass filter in conjunction with an image-reducing product detector is an acceptable solution, except that wideband IF noise may affect the performance of the AGC circuitry.[5]

A compromise approach is shown in Fig 2(D). The wide AM filter, with its lower group delay, is always in place and an IF amplifier drives the switched, narrower filters at a higher signal level (say 20 dB or so). The AM filter provides at least some 6-kHz wide "roofing" protection for the circuitry preceding the other filters.

Another approach, often used, is to use the design in Fig 2(C) and let the 10-MHz filter be the roofing filter (a crystal filter) with a bandwidth of, say 10 kHz. Its shape factor is not critical since the mechanical filters do the "hard work." This is an excellent method, but involves a relatively low-cost crystal filter (4 poles).

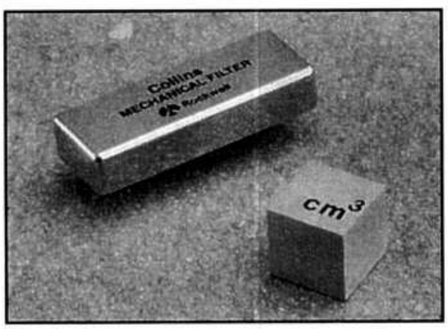


Fig 1—Collins mechanical filter with size comparison.

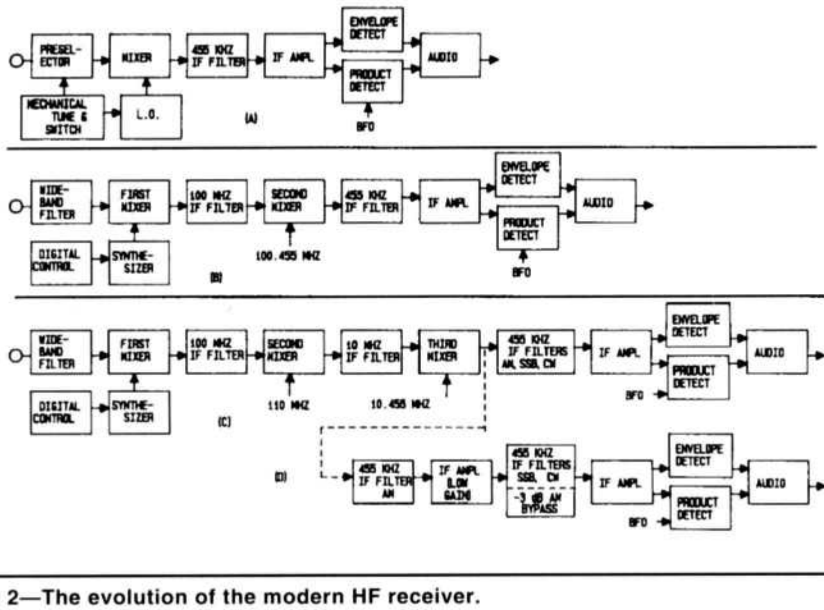


Fig 2—The evolution of the modern HF receiver.

A further elegant escalation is to use a more elaborate 10-MHz crystal filter and then manipulate the various local oscillators in such a way that the intersection of the crystal-filter passband and the mechanical-filter passband produces a variable bandwidth and a variable center frequency. As these are changed, the net tune-frequency remains constant. Pairs of AM, SSB or CW filters with matched frequency responses are most effective. This method creates a valuable "operator's aid" feature.

Impedance Matching the Mechanical Filter

The termination requirements for the mechanical filters are shown in Fig 3. The source Z_S and load Z_L should be 2000 Ω, within 5%, and as purely resistive as possible across the 6-dB bandwidth, and should not become highly reactive in the transition bands (see later discussion). The matching networks establish the correct value of load impedance, R_{LOAD}, at the output terminal of the driver stage (to establish the desired gain, stability and maximum output excursion for that stage) and the correct value of source impedance, R_{OUTPUT}, for the output stage (to establish the desired input excursion, noise figure and stability for that stage). They also determine the ratio V_{OUT}/V_{IN} which is needed for correct gain distribution. The physical resistors, R_{NET}, are usually required in order to help meet *all* of the above requirements simultaneously (see following text for a further word of clarification of this point). Some signal power is lost in them, but at this point in the signal path some resistive loss is tolerable. Some of this resistive loading resides within the QX product of the matching network inductors. Also, in the interest of accurate filter termination, the resistive loading supplements the loading by the active devices, which often show considerable variation over temperature and production tolerances.

Variations in the filter input and output impedances $Z_{F\ IN}$ and $Z_{F\ OUT}$ also affect Z_{LOAD} and Z_{OUTPUT} (Fig 3). Figs 4 and 5 show typical AM filter impedance magnitude and phase plots. In particular, the out-of-band overloading tendency at the output terminal of the driving stage could be degraded by a large increase in Z_{LOAD} outside the filter pass band. The matching networks should be designed with this potential problem in mind. Equally important is that the filter will work properly if it "sees" the correct Z_S and Z_L at each end.

The L Network

The method shown in Fig 6 is a simple network which provides the necessary impedance transformation from some high value down to 2000 Ω without "coloring" the frequency response of the mechanical filter. The idea is to let the mechanical filter provide the selectivity. This method provides a very easy way to achieve resonance and impedance matching simultaneously. The behavior of this network can best be visualized with the Smith Chart diagram of Fig 6, in which the center point is normalized at 2000 Ω. Starting at some desired high value of R at 455 kHz, the shunt coil has an inductive susceptance which traverses the path AB. It arrives at the 2000-Ω constant resistance circle at

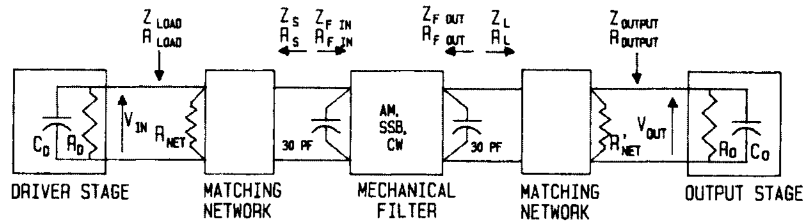


Fig 3—Matching mechanical filter-to-driver and output stages.

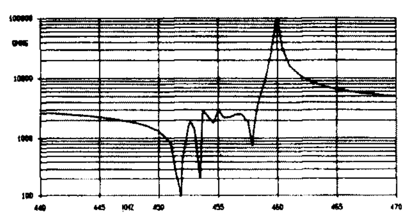


Fig 4—Filter input-Z magnitude vs frequency.

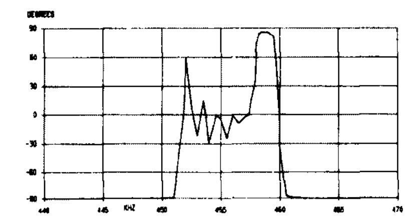


Fig 5—Filter input-Z angle vs frequency.

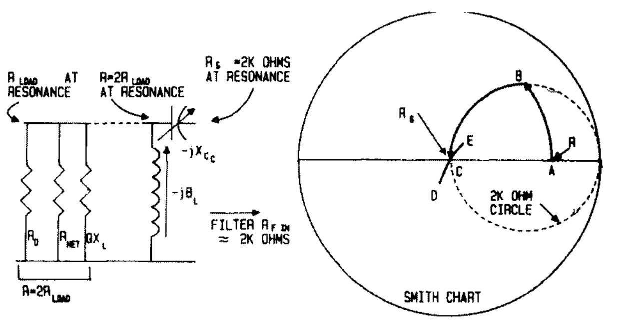


Fig 6—Matching network (L-Network) without stray capacitance.

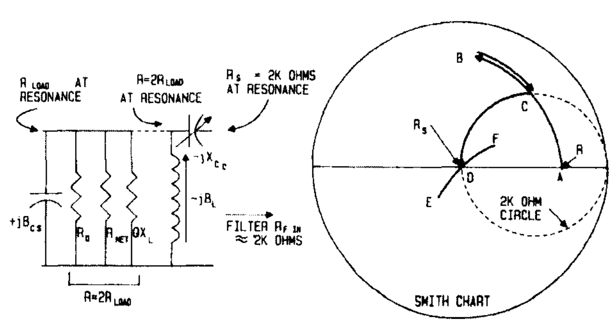


Fig 7—Matching network (L-Network) with stray capacitance.

point B. A series capacitive reactance then traverses the path BC, where C is 2000 Ω. As the frequency varies from 450 kHz to 460 kHz the value of Z_S (or Z_L) varies along the path DCE.

A complication occurs because of the stray capacitance, C_S, shown in Fig 7. This consists of the capacitances of the coil, the wiring and the active device involved. The Smith Chart of Fig 7 shows that now the inductive susceptance must be significantly greater, path AB. The shunt capacitive susceptance of C_S lies along the return path BC, as shown, and point C is on the 2000-Ω circle, just as before. In other words, the "effective" inductive susceptance is the path AC. Series capacitive reactance then traverses CD to the 2000-Ω point. But when we look at the spread (EDF) of Z_S (or Z_L) from 450 kHz to 460 kHz we see that it is significantly greater than before. The stray C_S sharpens the selectivity of the network.

Too much LC selectivity, especially if it is repeated at input and ouput, produces excessive rolloff at the passband edges. It will also cause impedance mismatch, causing excessive ripples at the passband edges. This is based on my experience with both actual circuits and accurate *SPICE* simulations. The "passband" involved might be that of a single AM filter or a set of USB/LSB, or even four-channel multiplex, SSB filters which are switch selected as discussed in a later section. The L network is an excellent answer to this problem.

For the best results over the 450- to 460-kHz band, especially for the wider filters, the coil and the sum of its self-capacitance and other stray C_S should be resonant as far as possible above 460 kHz. Coil catalogs usually give the value of inductance at some low frequency, where capacitive effects are small, and also the self-resonant frequency from which C_{COIL} and L_{TRUE} can then be estimated.

The coil and capacitor C_C must be high-quality components which are stable over temperature and time. C_C should be an NPO trimmer in parallel with a stable fixed capacitor. Other parameters such as stray C_S and active device I/O impedances should be as constant as reasonably possible. These stabilities become more necessary as R in Fig 6 or 7 becomes larger because then, for some small fractional change of L_{EFF} or C_C, the mismatch and the passband response degrade more rapidly.

Eqs 1 and 2 give values of L_{EFF} and C_C (coupling capacitor) for a given value of R, where R (as shown in Figs 6 and 7) is equal to twice the desired value of R_{LOAD} (or R_{OUTPUT}) indicated in Fig 3. The variation in Z_S (or Z_L) over frequency can then best be determined either from a Smith Chart program or by *SPICE* simulation.[6] A further word of explanation: R is the total shunt resistance across the high impedance terminals, *not* including the contribution of the mechanical filter. If it is twice R_{LOAD} or R_{OUTPUT} then the filter is properly terminated and, simultaneously, the load impedance of the driver or the source impedance of the output stage, in the passband, is the desired value. In other words, the filter and its matching network provide the other R which, in parallel with the R mentioned above, gives the desired R_{LOAD} (or R_{OUTPUT}) within the filter pass band.

$$L_{EFF} = \frac{1.564\times10^{-5}\times R}{\sqrt{R-2000}} \qquad \text{Eq 1}$$

$$C_C = \frac{1.224\times10^{-13}}{L_{EFF}} + \frac{L_{EFF}}{R^2} \qquad \text{Eq 2}$$

$$L_{TRUE} = L_{EFF}\left[\frac{\sqrt{1+4KL_{EFF}^2}-1}{2KL_{EFF}^2}\right]; \qquad \text{Eq 3}$$

$$K = 8.173\times10^{12}\, C_S^2$$

where in Eq 3, L_{TRUE} is the catalog low frequency value of the inductor.

Fig 8 shows the L network as it is used to transform from a low impedance line R_{LINE} up to 2000 Ω. In this case the $Q'X_L'$ product of the coil is part of the mechanical filter's source or load resistance (R_S or R_L) and the L network transforms R_{LINE} up to a value R' which, in parallel with $Q'X_L'$, is the desired 2000 Ω. A word of clarification: The series combination R_{LINE} and C'_C in Fig 8 is equivalent to a *parallel* combination which consists of the resistive component R' and a capacitive component X_C' which is "tuned out" by L'_{EFF}. Eqs 4, 5 and 6 can be solved recursively for L'_{EFF} and C'_C, starting with a value of R'=2000 Ω. This process converges rapidly because $Q'X_L'$ is usually much

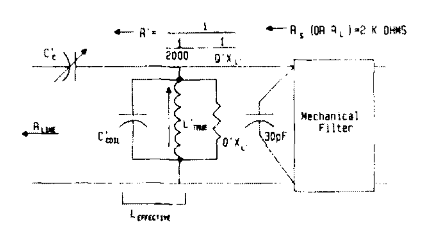


Fig 8—Matching mechanical filter to lower impedance R_{LINE}.

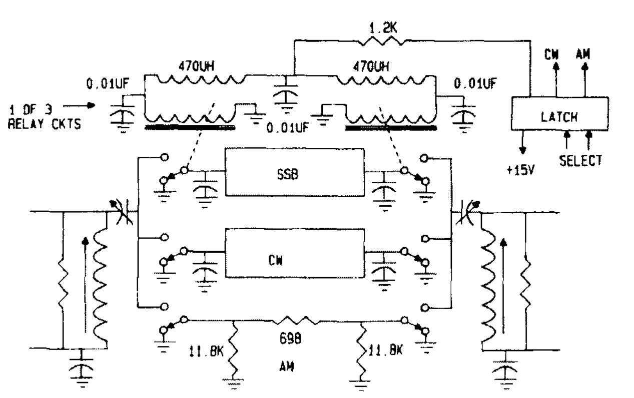


Fig 9—Relay switching of mechanical filters.

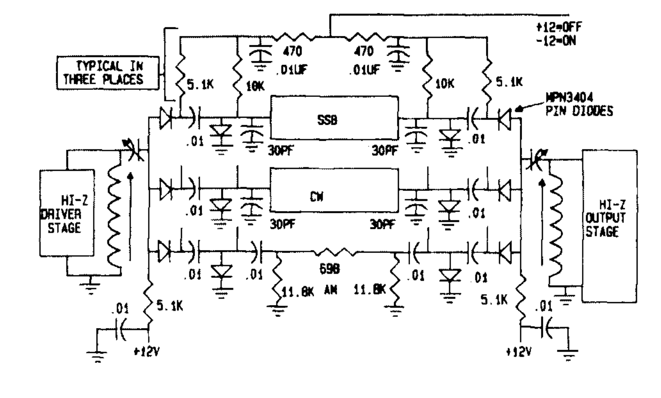


Fig 10—Diode switching of mechanical filters.

greater than 2000 Ω. Be sure, also, that the value of R_{LINE} does not vary.

$$C'_C = \frac{1}{2.859\times10^6 \times \sqrt{R_{LINE}(R - R_{LINE})}} \qquad \text{Eq 4}$$

$$L'_{EFF} = \frac{1+8.173\times10^{12}\times C'^2_C R^2_{LINE}}{8.173\times10^{12}\times C'_C} \qquad \text{Eq 5}$$

$$R' = \frac{1}{\frac{1}{2000} - \frac{1}{Q'X'_L}} \qquad \text{Eq 6}$$

$$L'_{TRUE} = L'_{EFF}\left[\frac{\sqrt{1+4KL'^2_{EFF}} - 1}{2KL'^2_{EFF}}\right] \qquad \text{Eq 7}$$

$$K = 8.173\times10^{12}\, C'^2_{COIL} \qquad \text{Eq 8}$$

Other Design Factors

A receiver which has perhaps 100 dB or more of gain compression due to AGC has stringent stopband attenuation requirements. In order not to spoil the stopband rejection of strong adjacent channel signals, my experience has been that the inductors must be well shielded, both electrostatically and magnetically. Circuit layout, shielding and ground management are critical. Procedure: temporarily disconnect the mechanical filter input and output leads and fix all the problems. A later section on filter switching contains further discussion of leakage. I have measured better than 100 dB of ultimate attenuation for these mechanical filters, therefore the stray paths around the filters must be at least that good. I have also noticed that small amounts of stray coupling can produce some anomalous and undesirable frequency-response distortions both inside and outside the passband. These leakage problems are a lot more manageable at the 455-kHz frequency than at the much higher IF frequencies that are often used.

The inductor core material should be correct for 455 kHz. For high impedance levels, ferrite with a μ_r of 125, using type 61, Q1 or 4C4 material provides small coils. For lower impedance levels, powdered iron assemblies with a μ_r of 20 (Carbonyl C) or 35 (Carbonyl HP) are preferred. Since the coil is heavily loaded (low operating Q) the coil Q need not be greater than 50 or so as long as its QX product is included in the impedance transformer calculations.

When designing matching circuits, it is necessary to be careful about intermodulation (IM) in the core material (the mechanical filters are excellent in this respect).[1] It is natural to want to use inductors whose dimensions are appropriate for these small filters, but troubles can easily occur. During the receiver design process, determine the maximum two-tone signal level that the mechanical filters and the inductors are expected to encounter and for which in-band IM performance is guaranteed. Using a two-tone signal of the appropriate level, check to see that the IM generated by the cores is at least 20 dB better than the requirement. Changes in core material (powdered iron is preferred) and size and perhaps a reduction in gain ahead of the filters may be needed.

Another good reason for using the L network is that it makes the inductance value large and therefore the volts per turn small, compared to other approaches which use smaller values of inductance, thereby improving IM performance. The inductor may, however, be slightly larger physically. C_{COIL} also tends to increase (try to minimize).

A special case is when the filter input network is at the output of the last mixer. This core is vulnerable to wideband IM distortion which can ruin the front end intermodulation performance of the receiver. My experience with solid-state receiver circuitry is that small (but not too small), fully shielded powdered iron core coils at this location cause very little trouble if the front-end amplification is minimized.

Quite often we desire to drive the mechanical filter from a diode double-balanced mixer which "likes to see" a pure resistance output load (50 Ω) over a wide frequency range, in order to maximize the mixer's dynamic range. One excellent approach is to insert a low-gain, low-noise buffer amplifier whose resistive input impedance is determined by lossless feedback and

Transformer Coupling for Mechanical Filters

Fig A shows how wideband transformers can be used to connect a driver stage and an output stage to a mechanical filter. The main goals are to get the desired value of gain for the driver stage, the desired value of source (generator) impedance for the output stage and to terminate each end of the filter *exactly* as recommended: 2000 Ω in parallel with 30 pF. The input side is shown, and the methods to be discussed are also applied to the output side.

The main idea is that all of the shunt resistances, capacitances and inductances to the left of the ideal transformer, including those of the driver stage, a fixed resistor R1 (to set the gain of the driver stage) and also the real-world transformer itself, when modified by the square of the turns ratio *N*, are combined with C1 at the input of the filter to provide 2000 Ω in parallel with 30 pF across the passband and transition bands.

The most difficult item is the transformer design. For a step-down transformer, the 30-pF requirement can be difficult to obtain if *N* is large, so moderate values of *N* should be used. This in turn implies a rather low value of R1 and a high g_m for the driver stage, to get the required gain. R1 is needed to prevent a large increase in gain, and therefore possible instability, outside the filter passband and also to help terminate the filter. If type-Q1 ferrite cup-cores are used, it is easier to experiment with the transformer design.

A very desirable approach is to make accurate IF impedance measurements, using an RF impedance bridge, looking to the left from the filter input terminals. The transformer design, R1 and C1 are iterated until the correct impedance is achieved. Under some conditions it may even be possible to *eliminate* the transformer, which would be especially nice.

These procedures are then repeated at the filter output.—*William Sabin, WØIYH*

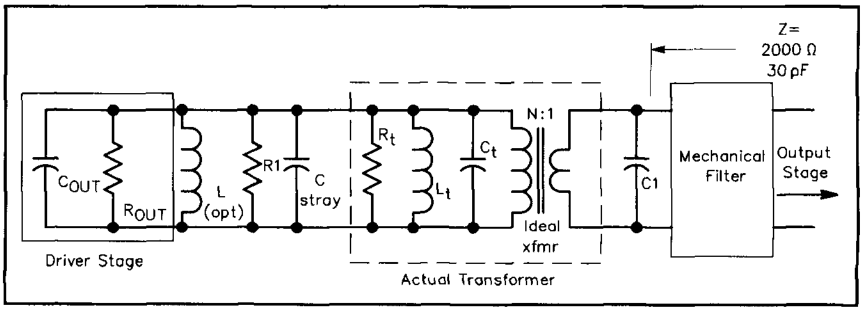


Fig A

which is invariant with respect to the amplifier's output load variations (ie, unilateralized).[4] Some kind of diplexing (LP/HP or BP/BR) between mixer and amplifier, to improve wideband resistive loading of the mixer, is highly recommended. A simple (read "crude") compromise is a 3-dB pad. Also, mixers which are called TIM (termination insensitive mixers) are commercially available and should be considered. Quite often, some reduction in mixer performance can be traded for cost reductions. Maximum intermodulation performance is not always *the* prime consideration.

Balanced active FET mixers, as used in medium performance receivers, tend to be more tolerant of output load impedance variations when driving a mechanical filter because a buffering action similar to that described above is inherent in their design.

Finally, a very desirable approach to circuit design with a mechanical filter, as Note 1 points out, is to design its driver (or load) amplifier stage to provide a 2000-Ω output (or input) impedance, thus minimizing the number of tuned LC networks. This may be easy to do, using recently perfected wideband opamps from many vendors as IF amplifiers.[7] Some other examples are the NE602 mixer and the NE605 mixer/FM discriminator devices, which have input/output resistances in the 1000- to 2000-Ω range.

It is necessary, though, to restrict somehow the noise bandwidth of the later amplifier stages, as mentioned before, so some tuned circuits will be needed. The IF properties of any IC, such as noise figure and two-tone intermodulation distortion, at maximum expected signal level and especially if used before the filter, should be checked out. Check also for crossover distortion, which can degrade the IM ratio at lower signal levels. If a gain-controllable amplifier is used, check the IM over the expected range of gain control. The gain-control function should have, ideally, a constant dB-per-volt slope characteristic so that the AGC loop will be stable.

Switching Circuitry

Two methods of switching mechanical filters will be considered: relays and diodes. The use of rotary gang switches has been pretty well obsoleted in recent times because of size restraints, packaging problems, the use of automated assembly and the desire to have software control of the radio. Relay switching is an excellent and simple method, but involves electromechanical hardware which may be objectionable in some cases. Diode switching is more complicated and may be subject to intermodulation in the diodes but may have greater reliability in difficult environments. It is also an excellent method.

Fig 9 shows a relay switching method. In AM mode (see Fig 2(D)) a 3-dB, 2000-Ω pad is switched in. The relays are ultraminiature SPDT types, Potter & Brumfield type T81 or similar (for example, Radio Shack 275-241), with separate relays at input and output. The filter I/O terminals are grounded when the filter is "deselected," and better than 100 dB of isolation is obtainable if the layout is carefully done. One cause of serious response degradation, especially in the stopband, is due to capacitance from the center arm to the relay coil forming a sneak path around a "selected" filter. In one case, this capacitance was 6 pF. A net value of just 0.1 pF from input to output causes very serious problems. The decoupling circuitry for the relay coils helps to prevent problems of this kind. The relay method is very straightforward and low-loss; the relays are cheap (about $3 each).

The diode switching method is shown in Fig 10. The diodes are Motorola MPN3404 PIN band-switching types with 1.5-pF reverse biased (12 V) and 1.5 Ω or so forward biased (2.5 mA). Similar types of diodes are available in surface-mount packages. Note the absence of RF chokes in the signal path which can spoil the 100-dB isolation by spurious inductive coupling effects.

In this circuit we take a slightly different approach to the L network in order to minimize the signal loss due to the biasing resistors. This loss could become quite large (in one particular case I had about 4.8 dB per switch) if we use the approach in Figs 6 and 7. In Fig 10 the combined value of the diode biasing resistors and the QX product of the inductor (after it has been transformed down) is the 2000-Ω pure resistance loading which the filter demands. In order to make this work efficiently, the driver stage should be a current source (ideally) and the output stage should have a very high input impedance, both of which are easy to get at 455 kHz. The values for the L networks are then chosen to get the desired load impedance for the driver stage and the source impedance for the output stage. The values given in Fig 10 can be considered "starters" and can be tweaked as required (especially the 10-kΩ resistors). *SPICE* simulations are very helpful if accurate Q values are used.

In my prototype setup, using the AM filter, two-tone third-order inband IM products (at 453.5/456.5 kHz) are better than 70 dB below each tone (454.5/455.5 kHz) at an input level of 0.0 dBm per tone. The total isolation for a pair of turned off switches is better than 100 dB, if the layout is properly done (100 dB is a lot of dBs in such a very small volume but, as mentioned before, it is very desirable).

The commercially available GaAs SPDT switches, designed for 50-Ω circuits, are less desirable because they would require a pair of tuned matching transformers for each mechanical filter.

A possible source of signal leakthrough concerns the grounding of the input/output pins of a deselected (switched-out) *wideband* filter or an attenuator. If the points at which these pins are grounded are at a different signal potential than the case of the filter, a small signal can be coupled into the filter and can induce a wideband output which may be down only 80 dB or so. This can degrade the ultimate attenuation of a switched-in *narrowband* filter. I have seen this kind of problem.

There is another leakage problem which I have found troublesome. Because of the large amount of amplification at the 455-kHz IF frequency, especially at low signal levels when the AGC is not operating, or just barely operating, it is possible for the high-level IF output to find its way back into the input of either the mechanical filter or the 455-kHz amplifiers preceding the filter. The symptom is that for low input signal levels the ripple in the passband increases, due to the feedback signal and the input signal adding or subtracting at various frequencies in the passband. As the input signal level increases, it overrides the feedback leakage, which tends to remain constant, and this rippling effect is reduced somewhat, but not adequately. The lead filtering methods shown in Figs 9 and 10 help to eliminate this effect, but other sneak paths, especially on PC boards, can cause the problem. Shielding, lead filtering and circuit layout improvements are usually indicated.

IF Circuit Alignment

The tuneup of the mechanical filter networks and all the other IF circuitry involves simultaneous adjustments for resonance and correct passband response (low ripple and flat response). The use of a sweep generator and logarithmic amplitude display, for example a spectrum analyzer-tracking generator pair or scalar network analyzer with 455 kHz capability, should be considered indispensable test equipment.

The Signetics NE604 log video detector chip ($5) makes an excellent lab-built 70-dB input device (total parts cost $50) for a low-frequency ($300) oscilloscope which can be used in conjunction with a low-cost ($150) function generator with sweep capability to produce a very low-cost alignment tool. I have built and used an instrument using this approach; it works quite well and does not tie up a very expensive ($10000 minimum) piece of test gear at a test station. A high-impedance input probe can be attached at various points for signal tracing and troubleshooting operations.

Conclusions

The small size, low cost and high-quality performance of the torsional mode mechanical filters contribute to the design of a superior HF receiver in a small package. Layout of the filter circuitry in a small volume is a critical item. The filters must be accurately terminated and impedance matched to the driver and output stages. The receiver block diagram, as discussed briefly in connection with Fig 2, is some acceptable compromise between cost and performance, with low intermodulation and high narrowband sensitivity as conflicting but resolvable goals.

The numerous contributions and critical review of this article by Bob Johnson,[8] Bill Domino and Peter Ysais, at Rockwell Filter Products are greatly appreciated. The encouragement of Lee Cornet is also appreciated. Their many years of experience with mechanical filter design, usage and marketing are prominent throughout this article.

Notes

[1] Domino, W. J. and Johnson, R. A., "Miniature Precision Bandpass Filters Solve IF Design Problems," *RF Design*, October, 1991.

[2] Bain, Richard, "Mixer Spurious Plotting Program," *RF Design*, May, 1989, errata, July, 1989, page 13.

[3] Sabin, W. E., "A Spurious Response Program for Wideband Mixer Circuits," *RF Design*, October, 1989.

[4] Sabin, W. E. and Schoenike, E. O., Editors, *Single-Sideband Systems and Circuits*, Second Edition, Chapter 4 (Receiver Design), McGraw-Hill, NY, 1995.

[5] Sabin, W. E., "Measuring SSB/CW Receiver Sensitivity," *QST*, October, 1992. See also *QST* Technical Correspondence, March, 1993.

[6] Hayward Electronics, 7700 SW Danielle Ave, Beaverton, OR, "Microsmith," available from ARRL.

[7] Breed, Gary, "High Speed IC Applications Circuits", *RF Design*, Feb, 1992.

[8] Johnson, Robert A., *Mechanical Filters in Electronics*, John Wiley & Sons, NY, 1983. The definitive work on every aspect of mechanical filters.

□□

PHEMT Preamp for 6 cm

Reprinted from Dubus Technik IV, *Preamps and Receivers.*

By Rainer Bertelsmeier, DJ9BV

Abstract

A preamp equipped with a PHEMT provides top-notch performance in noise figure and gain as well as unconditional stability for the 6-cm amateur band. The noise figure is 0.65 dB at a gain of 13 dB. It utilizes the NEC NE32684 X-band PHEMT and provides a facility for an optional second stage on board.

Circuit Description

This LNA utilizes a specially designed quarter-wave stripline coupler in the input. This innovative technique provides about 0.1 dB less loss than a high-Q ATC100A, hence 0.1 dB less noise figure than comparable LNAs (see Fig 2) and last, but not least, it's very inexpensive. Using the NEC32684 PHEMT, a noise figure of 0.65 dB can be measured at 5.760 GHz at a gain of 13 dB.

Fig 1 shows the circuit diagram and Fig 4 the 34 × 47-mm PC board. It's printed on 0.508-mm thick RT-Duroid 5870. The quarter-wave coupler at the input has a length of 9.3 mm and a slit width of 100 µm. The substrate thickness makes the coupler low loss and provides the low-inductance plated-through holes necessary for stability.

A separate bias circuit as described in Note 1 provides an independent adjustment facility for drain current and voltage as well as regulation.

Construction

Construction is performed using a microstripline technique. The active bias circuit, which provides constant voltage and current, is on a separate PC board.

Refer to the parts layout in Fig 3 for construction.

1. Mount the SMD parts onto the bias board and check the functioning of the circuit.
2. Prepare the PC board to fit into an aluminium box.
3. Prepare holes in the box for the SMA connectors. Note: input and output connectors are asymmetrical; use the PC board to mark the connector positions.
4. Drill holes for through contacts (0.9-mm diameter) in the PC board and connect through with 0.88-mm CuAg (silver-plated copper) at the indicated positions.
5. Solder all resistors onto the PC board.
6. Solder all capacitors onto the PC board.

Glücksburger Str 20
D-22769 Hamburg
Germany

7. Solder the NE326 PHEMT onto the PC board. Use only an insulated soldering tool; ground the PC board, your body and the power supply of the soldering tool. Never touch the PHEMT at the gate—only at the sources or the drain—when applying it to the PC board—and solder fast (less than 5 seconds).

8. Mount the PC board into the box with M2 screws.

9. Mount the SMA connectors, and solder the inner line to the stripline.

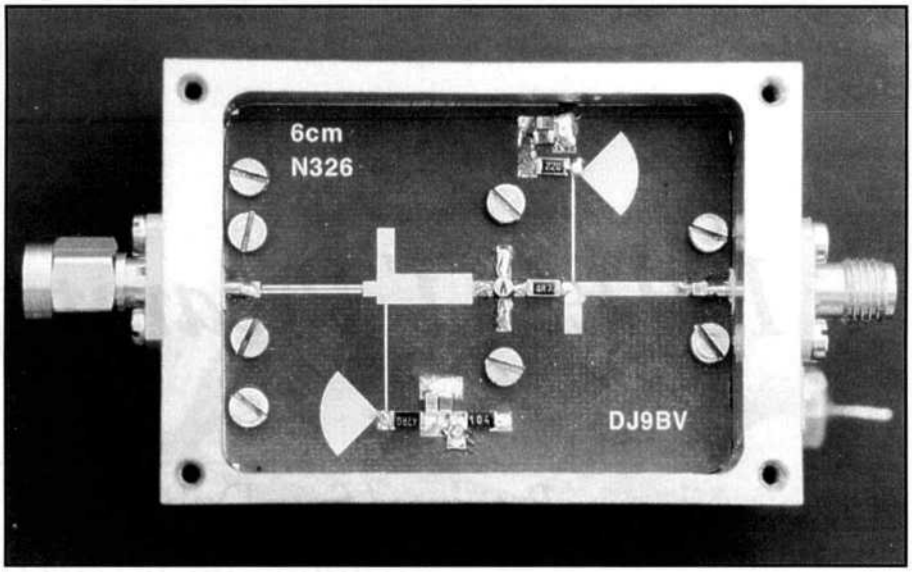
10. Mount the bias board into the box.

11. Connect D1 between the feed-through capacitor and the bias board. Connect the gate terminal and drain terminal of the RF-PC board to the bias board.

12. Connect 12-V B+, and adjust P1 for 12-mA drain current (measure 260 mV across R4 on the RF board). The drain voltage should be about +2 V.

13. Glue conducting foam to the inside of the top cover and mount the top plate.

14. Your small wonder is now finished.



The LNAH-5.7-N326 6-cm preamp.

Measurement Results

Noise Figure and Gain

Measurements were taken with an HP8510 network analyer and an HP8970B/HP346A noise figure analyzer, transferred to a PC and plotted.

Fig 7 shows the measurement results for gain and noise figure. A typical noise figure of 0.65 dB at a gain of 13 dB can be measured at 5.76 GHz. The preamp is rather broadband; a low noise figure is provided from 5.4 to 6 GHz.

Stability

A broadband sweep from 0.2 to 20

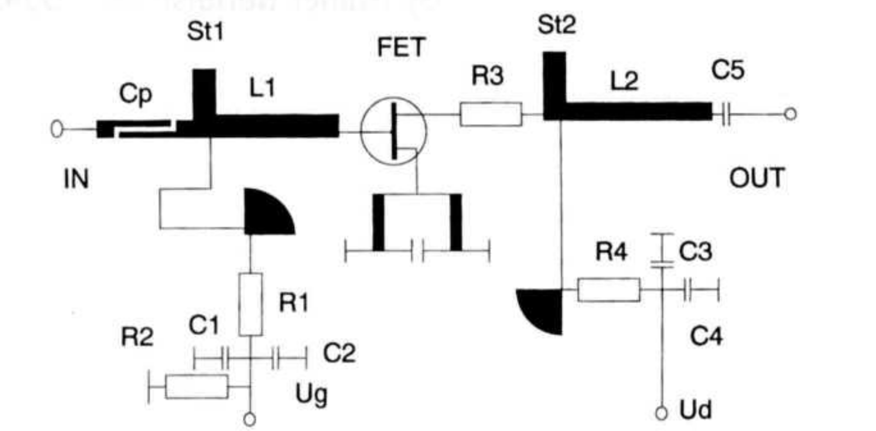


Fig 1—Circuit of the LNAH-5.7-N326 preamp.

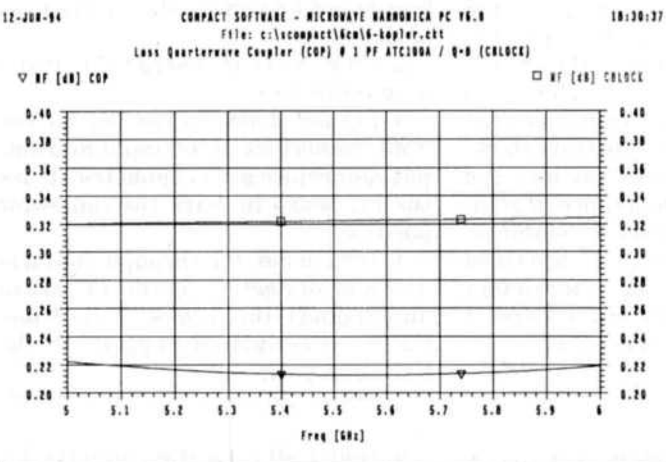


Fig 2—Loss of input coupler versus ATC100A/1 pF.

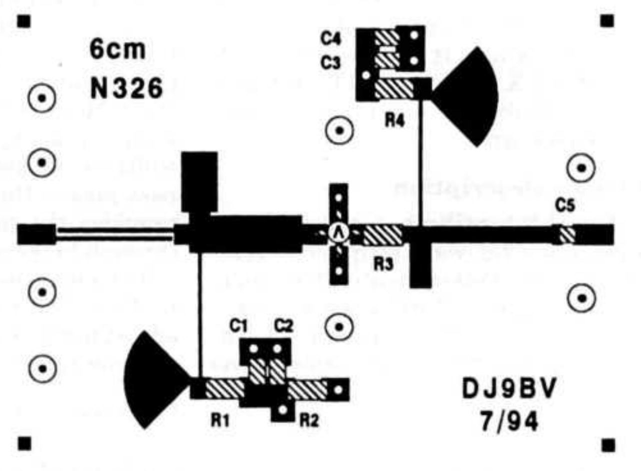


Fig 3—Parts layout of the LNAH-5.7-N326.

GHz shows a stability factor K of not less than 1.1, and the B1 measurement was always greater than zero. These two properties indicate unconditional stability.

The preamp is a no-tune design, unconditionally stable and has a very low noise figure.

Acknowledgments

I'd like to thank Dr. U. L. Rohde, KA2WEU/DJ2RL, from Compact Software, Inc, and its German distributor, Klaus Eichel, DL6SES/KF2OO, from TSS, for supplying the *Microwave Harmonica* software, which proved to be an excellent tool for the design. Thanks, too, to Rainer Jäger, DC3XY, for providing kits for these preamps, and last but not least, Dieter Briggmann, KC6GC, for the professional measurement of S-parameters

Parts List for the Bias Board

Part Number	*Type*	*Value*	*Manufacturer*	*Package*
C1, C6	SMD-C	0.1 µF	Sie	1206
C2, C4	SMD-Elko	1 µF/16 V	Sie	1206
C3, C7, C8	SMD-Elko	10 µF/16 V	Sie	1210
C5	SMD-C	0.01 µF	Sie	0805
T1	Si-PNP	BC807	Sie	
IC1	Regulator	LM317	Na	LM317LZ
IC2	Inverter	LTC1044SN8	LT	SO8
D1	Si-Diode	1N4007	Mo	
C2, C3, C4	Si-Diode	1N4148	Mo	
P1	SMD-Pot	1 kΩ	Vitrohm	4310
P2	SMD-Pot	100Ω	Vitrohm	4310
R1	SMD-R	10Ω	Sie	1206
R2, R3	SMD-R	220Ω	Sie	1206
R4	SMD-R	680Ω	Sie	1206
R5	SMD-R	33Ω	Sie	1206
R6	SMD-R	12 kΩ	Sie	1206
R7	SMD-R	4.7 kΩ	Sie	1206
R8	SMD-R	10 kΩ	Sie	1206
R9	SMD-R	22 kΩ	Sie	1206

Reuse of the Bias Board

The bias board (see Note 1) is reusable for any grounded-source FET or HEMT circuit. The voltage range at the Ud terminal is set by R4. R4 serves as a safety measure to limit the maximum output voltage if P1 suffers a random open condition during adjustment. The following table displays the voltage range versus value of R4.

Any resistors in the drain path have to be considered in respect to the drain current needed to determine the appropriate value of R4.

Ud versus R4

R4 (Ω)	U at IC1/Out	Ud (V)
560	2.15 to 3.4	1.5 to 2.8
680	2.2 to 3.75	1.55 to 3.1
820	2.25 to 4.05	1.6 to 3.4
1000	2.25 to 4.4	1.6 to 3.85

Measurements of the LNAH-5.7-N326 Preamp

Device:	NE32684
Noise Figure:	0.65 dB typical at 5760 MHz
Gain:	13 dB typical at 5760 MHz
Input RL:	10 dB
Output RL:	10 dB
Bandwidth:	NF < 0.8 dB, 5.3 to 6 GHz
Stability K:	> 1.1 from 0.2 to 20 GHz

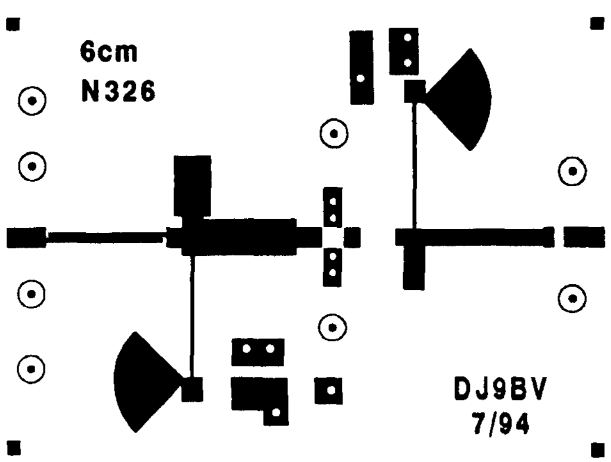


Fig 4—LNAH-5.7-N326 preamp PC board.

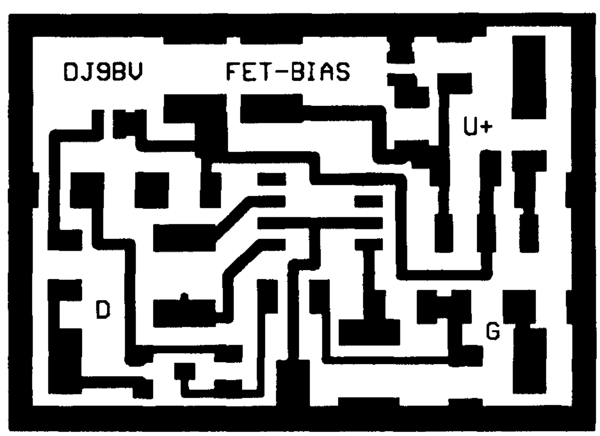


Fig 5—Bias circuit PC board.

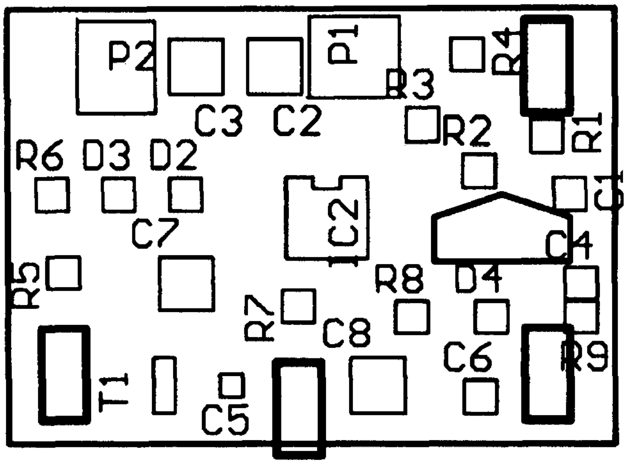


Fig 6—Bias circuit parts layout.

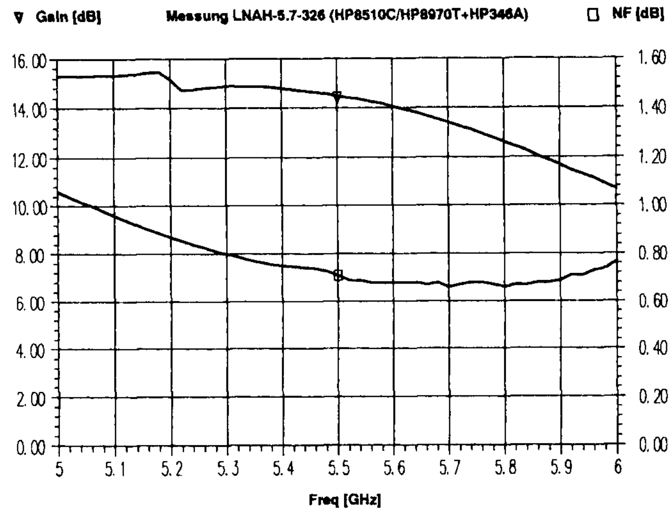


Fig 7—Measured noise figure and gain of the LNAH-5.7-N326 preamp.

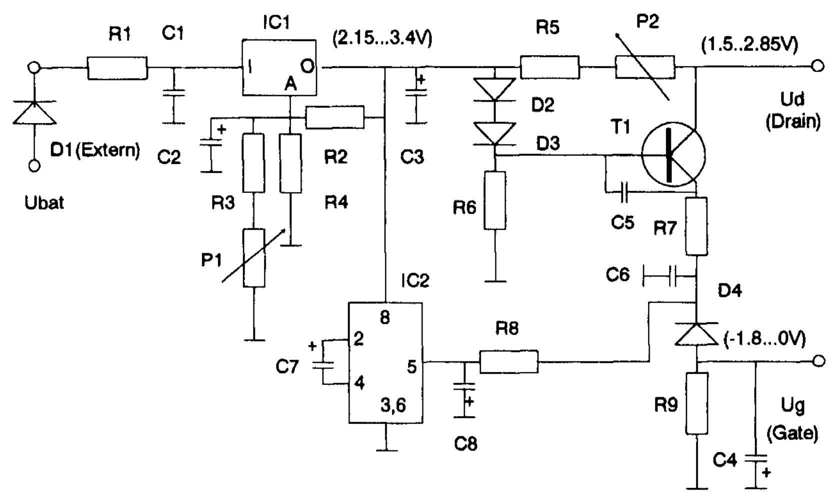


Fig 8—Bias circuit for grounded-source FETs.

and noise figures. Without their help this work would not have been possible.

Notes

Parts, kits and PC boards are available from Rainer Jäger, DC3XY, Breslauer Str.4, D-25479 Ellerau, Germany. Tel (+49) 41-06-73430.

Ready-made and calibrated units are available from Frank Schreyer, DD1XF, Maimoorweg 32, D-22179 Hamburg, Germany. Tel (+49) 40-642-8253.

[1]Bertelsmeier, R, DJ9BV, "PHEMT LNA for 13 cm," *QEX*, Oct 1995, pp 7-11.

Parts List for the LNAH-5.7-NB324 Preamp

Part Number	*Type*	*Value*	*Manufacturer*	*Package*
C1, C3	SMD-C	100 pF	Sie	
C2, C4	SMD-C	1000 pF	Sie	0805
C5	Chip-C 50 mil	1 pF (500 CHA 1R0 JG)	Tekelec	CHA
R1	SMD-R	47	Sie	1206
R2	SMD-R	100 k	Sie	1206
R3	SMD-R	4R7	Sie	1206
R4	SMD-R	22	Sie	1206
FET	GaAs Fet	NE32684	NEC	
Bu1, Bu2	Coaxial	SMA	Radiall/Suhner	
PC board	RT Duriod 5870	LNAH-5.7-N326 35×47×0.508 mm	DC3XY	
G	Aluminum box	35×74×30 mm		

RF

By Zack Lau, KH6CP/1

A VHF+ Preamp Design Technology Update

Among the topics discussed at the 1995 Microwave Update in Arlington Texas were preamplifier stability and filtering. Tommy Henderson, WD5AGO, discussed stability improvements to his 1296-MHz preamps. Chip Angle, N6CA, pointed out that a low-noise preamp is useless if your receiver is overloaded by strong out-of-band signals, and he presented a set of filter designs for 900 to 2400 MHz to solve this problem. The Microwave Update 1995 *Proceedings* are available from the ARRL as publication number 5366.

Preamp technology below 450 MHz really hasn't changed—the best devices are still the MGF1801/2116 on 144/222 MHz and the MGF1302 on 432 MHz. I suppose you might try a half-wave cavity input filter instead of the traditional quarter-wave designs, but even on 432 MHz such a preamp is rather large and bulky.[1]

According to Kent Britain's latest *DUBUS* article, power GaAs FETs achieve the best noise figures because they are easier to match with low loss.[2] The impedances of lower-noise devices are just too difficult to match without incurring extra losses—losses that raise the total noise figure. On the other hand, the greater impedance transformation needed for these devices requires circuits with a higher loaded Q, which means that the selectivity of the circuit is better when it's matched for best noise figure. The skirts of a single resonator aren't particularly steep, so you may want to add additional resonators to make a high-quality band-pass filter.

Since background noise on 2 meters is so high, even for EME work, you might wonder why some people use a \$50 device such as the MGF 1801 to get a 0.2-dB noise figure when a \$5 device will give a 0.3-dB noise figure—with better input selectivity. Well, a number of us were fortunate enough to buy these high-priced items at \$5 each when there was a surplus source. Unfortunately, I don't know of a current surplus source! While you don't *need* a noise figure this low at 2 meters, it's quite useful for calibrating noise-figure meters, since the laws of physics do set an obvious lower limit on the noise figure. Of course, it is a bit of work to make low-noise preamps for each band just for calibration purposes, so some people have taken another approach. They bring a wideband preamp, such as one made from a GaAs

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[1]Notes appear on page 28.

MMIC, to the noise figure contest, then use its measured noise figure to calibrate their set-up at home.

At 2304 MHz and above there have been significant changes. For casual applications it's tough to compete with the simplicity of a single GaAs MMIC like the MGA 86576. When optimized for narrowband use, you can usually get a noise figure from this device of under 2 dB, at least on the 13-, 9- and 6-cm amateur bands. The only difficulty is the need for excellent grounding—don't expect it to be stable even if you are using lots of plated through holes on 1/16-inch boards. (1/16-inch glass-epoxy G-10 or FR-4 [fire resistant] circuit board is probably the most common board used for inexpensive RF work.) If you must use such a thick board, mount the device on the ground plane and bring the input and output leads up through the board. Even better, use a good, thin board. I've had no problems grounding device leads through 15-mil Rogers Duroid.

When dealing with glass-epoxy board, the thickness is only nominal; it doesn't make much difference to the typical digital chip how thick the board is. On the other hand, one of the things you pay for with an expensive microwave substrate is some attempt at keeping the thickness of the board to a close tolerance.

In applications where high gain is acceptable, a GaAs MMIC makes an excellent second stage.[3] However, I wouldn't count on such a two-stage preamp winning a noise-figure contest unless you provide a stable receive converter. Since converters typically have 20 dB of RF amplification for low noise, adding another 45 dB results in 65 dB of gain, a bit much to have on a single frequency. Not only must the converter have excellent shielding, but it should also be unconditionally stable. Otherwise, the system may oscillate—even if your preamp is unconditionally stable by itself.

For the ultimate in noise figure of 2304 MHz room-temperature preamps, the Hewlett Packard ATF 36077 and NEC 32684A devices both do quite well. Perhaps because of competition, the newer devices are not only lower noise but also lower cost! It may also be time to investigate the NEC 32584C, which is NEC's latest low-noise PHEMT.

I think I've proved that source biasing isn't a disadvantage in terms of obtaining the lowest noise figure—my source-biased designs have been competitive in noise figure contests. As Charlie, G3WDG, pointed out in his Microwave Update talk, a poorly designed negative-bias circuit can degrade the noise figure of a preamp. He suggested precisely adjusting the quarter wave decoupling line used in the bias circuit to prevent it from injecting noise. This is just what I did for my 10-GHz preamp—I added the bias lines and adjusted them for minimal effect on the RF performance of the circuit.[4]

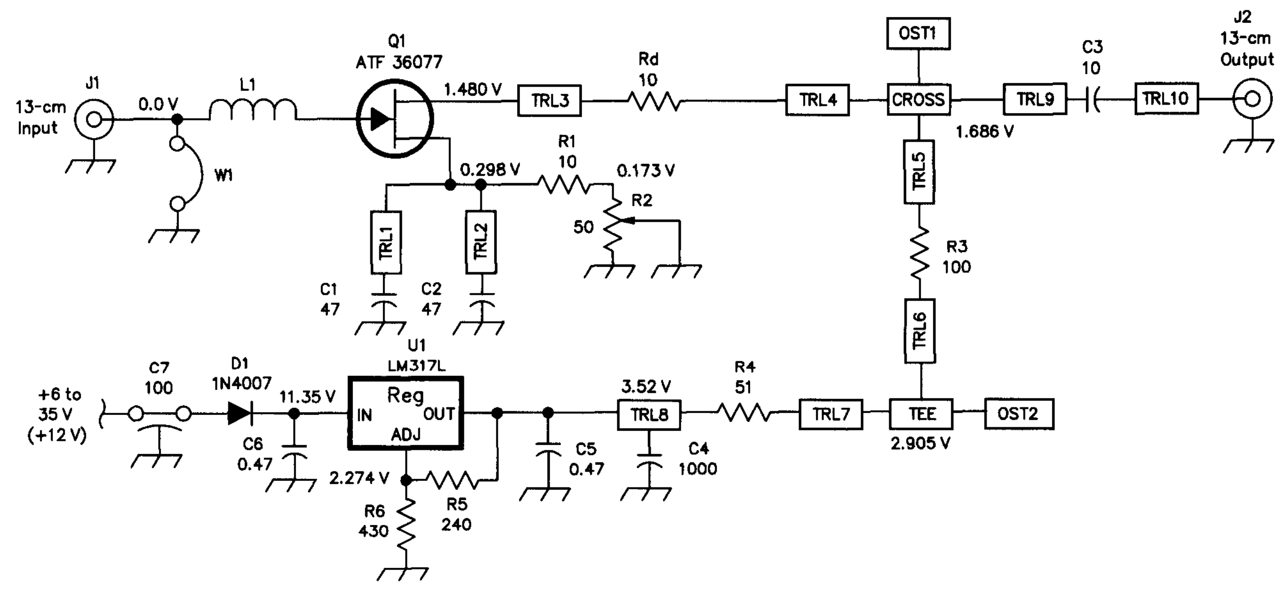

Fig 1—Schematic of the 2.3-GHz preamp, with measured dc voltages shown at key points.

C1, C2—47-pF ATC 100A chip capacitors. Substitution not recommended.
C3—10-pF chip capacitor.
C4—1000-pF chip capacitor.
C7—100-pF feedthrough capacitor. Value not critical. Anything from 100 pF through 0.1 μF should work fine.
D1—1N4007 reverse polarity protection diode. Other diodes like the 1N4001 will work.
J1—SMA connector with captivated center contact. Pin sticks out 50 mils to attach W1 and L1.
J2—SMA panel connector.
L1—Inductor made from 0.60 inches of no. 32 silver-plated copper wire. 2 turns on #50 drill bit used as a temporary form. No. 20 stranded Teflon wire is a good source for the no. 32 wire.
RD1—100-Ω chip resistor. Both 0805 and 1206 case versions will work, although the layout is for the 1206 case.
U1—LM317L adjustable three-terminal regulator.
W1—Wire loop made from no. 32 silver-plated copper wire. The overall wire length is 0.5 inches, with 0.075 inches used for connections at each end. One end goes to the center pin of J1, and the other end goes to a ground point 235 mils away.

A misconception about competitive preamp design is that you need to build a number of preamps to get a "good one." I typically need to build just two samples, assuming I haven't made a mistake in the design and construction process. Building more than two rarely results in a significantly better preamp. Maybe this just reflects my particular construction skills—I do mount SMA connectors by drilling and tapping four holes to accept 2-56 screws. This requires that the holes be accurately located since you don't get to ream out these holes! And it allows me to use inexpensive stainless-steel connectors instead of gold-plated connectors that can be soldered.

The December 1992 *QEX* 10-GHz preamplifier I designed seems to be quite reproducible, at least if a 1.0-dB noise figure is acceptable. I've seen a number of these, built by a variety of people, that work just fine without tuning. However, to get a 0.7- or 0.8-dB noise figure with this design requires tuning, which is an art that also requires an expensive noise figure meter. And it's also important to use the proper noise source, such as an HP 346A. The 346B won't yield useful numbers at the low noise figures of interest to amateurs due to its change in source impedance as it's pulsed on and off. It ought to be possible to use the 346B with a 10-dB attenuator, but I've seen people have trouble doing this. It may be necessary to use a precision attenuator with a good return loss (or SWR).

You might consider using the new NEC 32584C or the Mitsubishi MGF 4917D for new 10-GHz designs. Toshi, JE1AAH, brought a 10-GHz preamp to Microwave Update that uses the new Mitsubishi PHEMT, and it did just as well as my tuned NE32684A design.

While certainly a lot better looking, those expensive milled aluminum cases might actually degrade preamp performance slightly compared to the brass enclosures I use. The difficulty with aluminum cases is in connecting the ground plane of the circuit board to the input and output connectors. The brass enclosure makes it easy to obtain soldered connections of relatively low loss. With the milled cases, the best option for getting good connections seems to be gluing the circuit board to the case with conductive epoxy.

A number of people have reported good results with the two-stage MGF 1302 amplifier I published in May and June 1993 *QST*. While it doesn't have the lowest noise figure, it seems to be easily duplicated and works well without tweaking. As a bonus, configuring it for receive or transmit use (noise versus power) is just a matter of changing the biasing. It makes a good candidate as a second stage 10-GHz preamp to follow a low-noise PHEMT.

At 1296 MHz and 903 MHz, the lowest noise figures are obtained with cavity preamplifiers and PHEMT transistors. However, the stability is often marginal. I think this is only a problem if the preamp oscillates in your system. A more serious problem is interference from commercial radio services. I'm actually using a no-tune filter ahead of my 1296-MHz transverter, but this configuration probably isn't reproducible. When you combine low loss with a narrow filter, it is quite easy to get severe notches in the passband unless the filter is precisely tuned. This isn't a problem with the normal no-tune designs, since the filters have a fair amount of loss. For most people, a tuned low-loss filter made from copper sheet and tubing is a much better choice.[5]

On the higher bands, a filter well worth considering is a waveguide high-pass filter. Not only is this low loss, but it offers excellent rejection of low-frequency signals. You might consider feeding a 3456-MHz dish with a small horn, as opposed to the more popular triband feed, to get this effect.[6] While the triband feed has some rejection of out-of-band signals, it may not be enough with the tremendously powerful or nearby transmitters sometimes encountered.

A 2.3-GHz Preamp Project

Figs 1 through 3 show a preamp design for 2.3 GHz that uses an HP

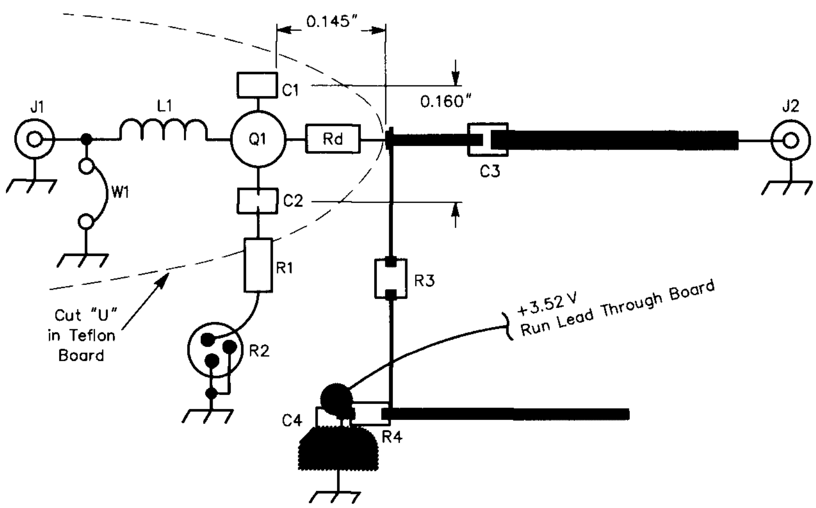

Fig 2—Parts placement diagram for the 2304 MHz preamplifier.

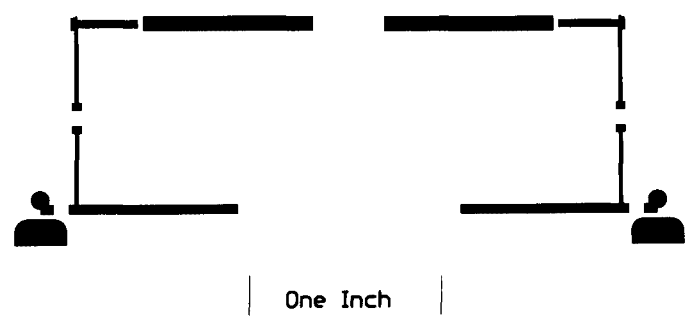

Fig 3—Etching pattern for the 13-cm preamplifier output network. The board is 1/32-inch Teflon board with a dielectric constant of 2.55.

ATF 36077 PHEMT. While the performance of the preamplifier is only marginally better than previous designs, it is presented for the economy-minded amateur who wants a high performance preamp using inexpensive parts. The price of the new HP part is less than half the cost of the NEC 32684A. The noise figure of this design is about 0.3 to 0.4 dB, with a gain of 16 to 18 dB.

Some people have encountered problems obtaining the 100-pF chip ATC capacitors I've used in my 2.3/2.4-GHz preamps. You may have to get them directly from American Technical Ceramics.[7] While the 47-pF chip capacitors are cheaper, computer modeling indicates you ought to stick with the 100-pF capacitors, at least when source biasing the NEC 32684A. The 47-pF ATC 100A chip capacitors seem to work better with the HP ATF 36077 PHEMT. Placement of these capacitors is rather critical—the spacing between them determines the amount of source inductance the transistor sees.

According to my computer model, Rd needs to be 51 Ω for unconditional stability. This sacrifices about 0.1 dB in noise figure compared to making it 10 Ω. The 10-Ω value seems adequate in practice, according to my measurements. Terminating one port of the preamplifier with a sliding short and looking at the other port with a spectrum analyzer indicates no sign of oscillation. The transmission line for the short is a 4-inch length of 3/32-inch tubing over a brass ground plane. Spacing is about 25 mils. This can be a hazardous test. Not only can you damage your preamplifier, but it is possible to damage the test equipment hooked up to the preamp since it is quite possible for some preamps to put out 100 mW or more.

This design is quite similar to the 2304-MHz design I published for the NEC 32684A in the November 1993 *QEX*. The most significant difference is in the quarter-wave bias line to the drain. I added a 100-Ω resistor in the center of it to enhance stability. This is the disadvantage of source biasing: the need to make changes to a design for stability when switching devices.

The board is etched on 1/32-inch Teflon circuit board with a dielectric constant of 2.55. A "U" is cut into the board to accommodate the PHEMT and input network. A piece of unetched circuit board fills in this gap, providing a nice ground plane for the input and source biasing circuits. Just as in some 2-meter preamp designs, you solder chip capacitors to a ground plane, and the capacitors then support the transistor in the air. It helps to solder the capacitors in before attaching the 25-mil-thick brass walls that enclose the preamp. I found a wall height of 0.75 inches to work well. Hold off on soldering the other chip components in until the walls are attached; these parts can be damaged if you bend the flexible Teflon board. The brass walls stiffen the board and make the preamplifier more rugged.

For best noise figure, it's necessary to adjust the inductor and wire loop using a noise figure meter. I found it necessary to spread out the inductor a bit. Tuning improved the noise figure of the second preamp I built by 0.08 dB.

Notes

[1] Lau, Z., KH6CP/1, "RF: A 1296 MHz Cavity Preamplifier," *QEX*, January 1995, pp 26-29.

[2] Britain, Kent, WA5VJB, "Power-GaAsFET Preamps," *DUBUS*, 3/1995, pp 11-16.

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