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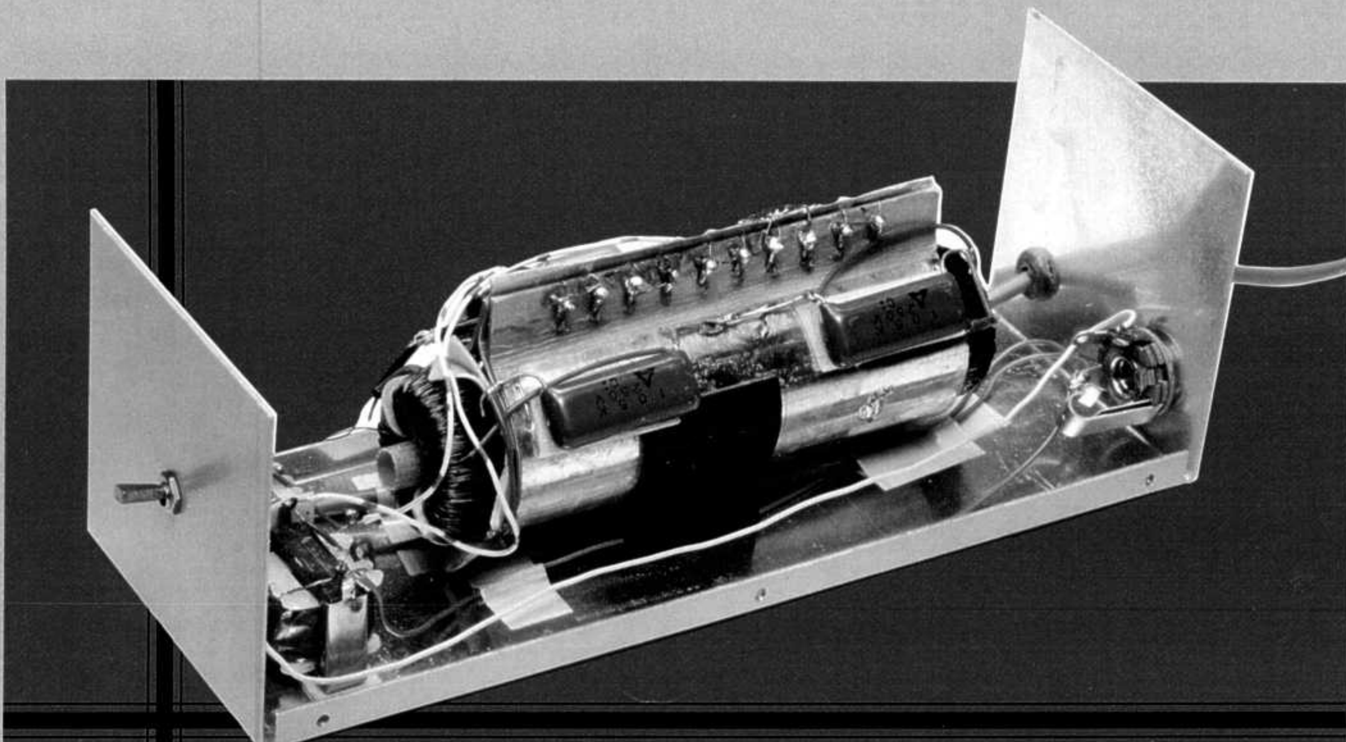

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COLUMNS

ABOUT THE COVER

Ed Wetherhold, W3NQN, is well-known for his passive audio-filter designs. The cover photo shows a filter he designed for use on SSB.

Purposes of *QEX*:

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

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

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Empirically Speaking. . .

Spectrum Management

Who keeps, gets or loses which pieces of the radio-frequency spectrum is increasingly becoming blood sport. The stakeholders in this game are legion: the International Telecommunication Union, industrialized and developing countries, Congress, the National Telecommunications and Information Administration, The Federal Communications Commission, broadcasters, citizens concerned about the education of children, minorities, females, land mobile operators, radio amateurs — and that's not even the whole tip of the iceberg.

On October 14, the Columbia Business School hosted a spectrum management seminar in New York City. This meeting was important enough to entice CCIR Director Richard C. Kirby, WØHLCT, to make a special trip from Geneva to attend. He named these specific issues for the early 1990s:

- Mobile services want more space in the 1-3, 4.2-4.4 and 5-5.25 GHz ranges
- High-frequency broadcasters are looking for more frequencies, particularly below 9 MHz
- Geostationary satellite broadcasters are eyeing 500-3,000 MHz for sound broadcasting and 12.7-23 GHz for high-definition TV.

You'll remember TV Answer as one of the groups who would like a piece of the amateur 220-MHz band.[1] Their lawyer, Henry Goldberg, was there speaking up for entrepreneurs with new gadgets that need spectrum. He said that there are five kinds of people filing with the FCC:

- Those who would gain spectrum
- Those who would lose spectrum
- Those who might experience more interference
- Those competing for the same spectrum
- Those with competing products (not necessarily RF)

His recipe?

- Don't ask permission from the FCC to get into business
- If you can't bypass the FCC, go to them with spectrum
- If not, slug it out with every other kind of interest

Practically every speaker decried the lack of sufficient data about spectrum utilization, propagation and just about everything else. That translates to "If you can't dazzle them with balderdash, complain about not having enough facts."

There was also considerable discussion about spectrum auctioning, even ownership of spectrum just like land. Don't panic yet—they're mainly talking about building fences around certain bands, not ham bands. Nevertheless, selling or leasing spectrum rights is discussed more and more these days in polite company, without the anonymity of a shopping bag over one's head.

This brings us to *All I Really Need to Know I Learned in Kindergarten*, by Unitarian minister Robert Fulghum. There's a book by that title (ISBN 0-394-57102-9), but it started out as a list of kindergarten lessons and was widely photocopied, read on the air and exchanged on BBSs. It also appears to have an uncanny applicability to spectrum management:

Share everything.
Play fair.
Don't hit people.
Put things back where you found them.
Clean up your own mess.
Don't take things that are not yours.
Say you're sorry when you hurt somebody.
Wash your hands before you eat.
Flush.
Live a balanced life.
Watch out for traffic, hold hands and stick together.

Now, doesn't that just about say it? *—W4RI*

[1]Sumner, "TV Answer: What was the Question?," p 9, *QST*, Mar 89.

CW and SSB Audio Filters Using 88-mH Inductors

By Ed Wetherhold, W3NQN
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The advances in audio filtering using active RC and switched-capacitor filters (SCFs),[1,2] may give you the impression that the simple, passive, LC audio filter is obsolete and of no real use. In spite of its dated technology, however, the passive LC audio filter is still preferred by many radio amateurs and professional filter designers. E. Christian in *LC-Filters: Design, Testing and Manufacturing*[3] writes: "If the design of filter networks would have started with active RC filters, the invention of coils would have been considered a major breakthrough!"

In this article, I'll show you how LC-filter theory is applied to the design and construction of inexpensive, high-performance, CW and SSB audio filters. I've included information that will demonstrate how passive audio filters are—in some ways—superior to active audio filters.

Passive LC audio filters have several advantages: They don't need a power supply, are less susceptible to signal overload, are inexpensive and easy to build, and are less likely to fail and are easier to troubleshoot and repair. The filter's disadvantages of greater bulk, fixed bandwidth, fixed center frequency and a slight insertion loss are generally not that important to the radio amateur. Once a desired bandwidth and center frequency are selected, there is little need to change them. Insertion loss is easily compensated for by increasing the receiver audio gain. Matching the input and output impedances of the LC audio filter to the audio output of a receiver and headphones is solved easily by using a filter design that allows the use of inexpensive impedance-matching transformers.

Surplus 88-mH inductor stacks are especially well-suited for building passive CW and SSB audio filters.[4] These high-Q inductors come in a cardboard-covered, five-inductor cylindrical stack, complete with terminals. Additional coils can be mounted at the ends of the stack when the filter design requires more than five inductors. The physical shape of the 88-mH inductor stack, and the position of the stack's terminals, make the five-resonator CW and SSB audio-filter construction very easy.

CW Filter Construction and Design

I'll discuss the design of narrow- and broadband CW audio filters. The narrowband filter uses one 88-mH inductor stack with two modified 88-mH inductors fastened to each end. Fig 1 shows how the inductors and capacitors are assembled to make this CW filter. The inductor stack mounts to a base with a 1⅜-inch plastic mounting clip. The broadband CW audio filter is physically similar to the narrowband filter, the difference between the two being that the broadband filter uses one additional inductor instead of two. Electrically, the broadband filter is about about 1.5 times wider than the narrowband filter.

Designs 1 and 4 of Table 1 summarize the measured performance parameters of the narrowband and broadband designs that were constructed using the design data presented in Table 2 (see designs 1-5). The schematic diagram shown in Fig 2 applies to both the narrowband and broadband designs. Figs 3 and 4 are pictorial diagrams that show you how to wire the narrowband and broadband CW filters using the existing terminals on the inductor stack.

Table 2 lists the several CW-filter designs available. Pick a design having the desired center frequency (Fc) and bandwidth. A 750-Hz Fc is usually preferred because the sidetone in many transceivers is at this frequency. The narrowband CW-filter designs are best for DXing or QRP reception where optimum selectivity is required. The broadband CW filter designs are more suited to CW net operation where the wider bandwidth permits reception of off-frequency stations. In either case, the measured shape factor (given in Table 1) indicates both the narrowband and broadband designs have good skirt selectivity. The measured relative attenuation of designs 1 (narrowband) and 4 (broadband) is shown in Fig 5. Notice that both CW-filter designs have a relatively flat passband, ensuring that the filter will not ring. If you want a CW-filter design with different Fc and bandwidth, calculate a design for your own special requirements using the procedure explained in Appendix A.

R1 (see Fig 2) serves to maintain a relatively constant audio level when the filter is switched in or out of the circuit.

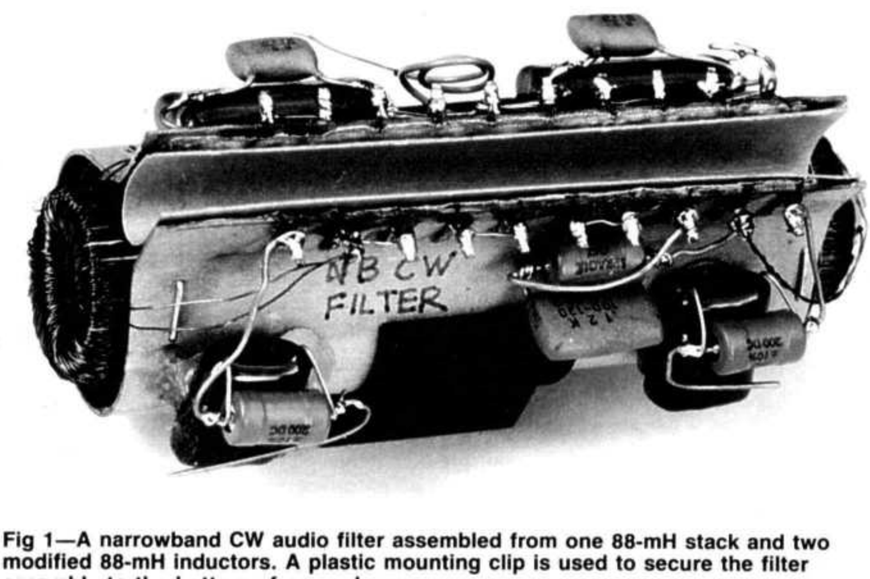


Fig 1—A narrowband CW audio filter assembled from one 88-mH stack and two modified 88-mH inductors. A plastic mounting clip is used to secure the filter assembly to the bottom of an enclosure.

[1]Notes appear on page 7.

Table 1
Summary of Measured Performance Parameters of Five-Resonator Filters.

No.	*Filter Type*	*Center Freq (Hz)*	*3-dB BW (Hz)*	*3-dB F-Lo (Hz)*	*3-dB F-Hi (Hz)*	*30-dB BW (Hz)*	*30-dB F-Lo (Hz)*	*30-dB F-Hi (Hz)*	*Insertion Loss (dB)*	*Shape Factor*
1	CW, Narrow-band	746	254	630	884	544	522	1066	2.2	2.14
4	CW, Broad-band	592	357	440	797	649	348	997	1.5	1.82
6	SSB	1056	2628	372	3000	4519	232	4751	0.2	1.72

Notes

1. All filter responses were measured using a source and load impedance within 10% of the Rt design values given in Table 2.
2. The center frequency is calculated using the measured 3-dB frequencies and the equation:

$$F\text{-center} = \sqrt{F3Lo \times F3Hi}$$

where
F3Lo is the 3-dB F-Lo frequency and F3Hi is the 3-dB F-Hi frequency.

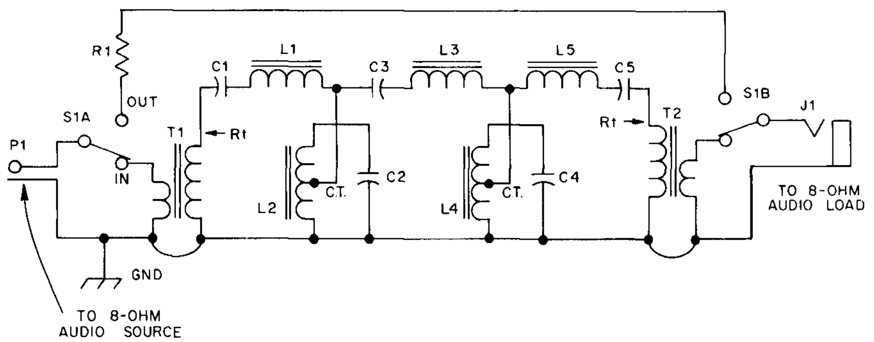
3. All attenuation levels are measured relative to a 0 dB level at the frequency of minimum insertion loss.
4. The shape factor, an indication of the filter selectivity, is calculated by dividing the 30-dB bandwidth by the 3-dB bandwidth.

Table 2
Component Values and Calculated Design Parameters of Five-Resonator Filters

No.	*Filter Type*	*Center Freq (Hz)*	*C1, C5 (μF)*	*C2, C4 (μF)*	*C3 (μF)*	*L1, L5 (mH)*	*L2, L4 (mH)*	*L3 (mH)*	*Rt Design (ohms)*	*3-dB BW (Hz)*	*RC (%)*
1		750	0.512	1.036	0.1706		43.5		230.0	275	
2	CW, Narrow-Band	651	0.680	1.20	0.2267		49.9	264	213.7	256	0.0441
3		490	1.20	1.20	0.400		88.0		213.7	256	
4	CW, Broad-band	592	0.820	0.820	0.410	88	88.0	176	208.4	389	
5		651	0.680	0.820	0.340		73.0	176	208.4	389	6.3
6	SSB	1031	0.271	1.084	0.542		22.0	44.0	224	2705	

Notes

1. Rt is the design impedance of the filter. The actual source and load impedances seen by the filter should be within 10% of the design value.
2. RC is the reflection coefficient of the Chebyshev low-pass filter prototype upon which the bandpass filter is based.
3. The 3-dB BW is the bandwidth used in the band-pass filter design calculations explained in Appendixes A and B.
4. L2 and L4 are made by removing turn-pairs from a bifilar-wound, 88-mH inductor. See Appendix C for an explanation of how to determine the number of turn-pairs to remove and the lead connections required.
5. In designs 1, 2 and 3, three 88-mH inductors are connected in series to make L3. In designs 4 and 5, two 88-mH inductors are connected in series for L3. In design 6, the windings of two inductors are connected parallel-aiding to make 22-mH inductors. The two 22-mH inductors are then connected in series to make the 44-mH value required at L3.
6. Mylar® or metallized Mylar capacitors, selected to be within 1% of the design values, are recommended for optimum filter performance and easy mounting on the inductor stack.



T1, T2—8:200-ohm, 0.4-W transformer (Mouser 42TU200). High-Z and low-Z winding resistances are 12 and 0.5 ohms, respectively.
S1—DPDT toggle switch.
R1—0 to 12 ohms, ½ W, 10% (see text).

Fig 2—Diagram of the CW audio filter for designs 1-5 presented in Table 2. Note: The filter sees a termination impedance consisting of the 200-ohm secondary impedance (transformed from 8 ohms), the 24-ohm transformer winding resistances referred to the high impedance winding, and the resistance of L1 (or L5), for a total impedance of about 230 ohms. Therefore, the filter is designed for any convenient impedance within 10% of 230 ohms to assure a flat pass-band response.

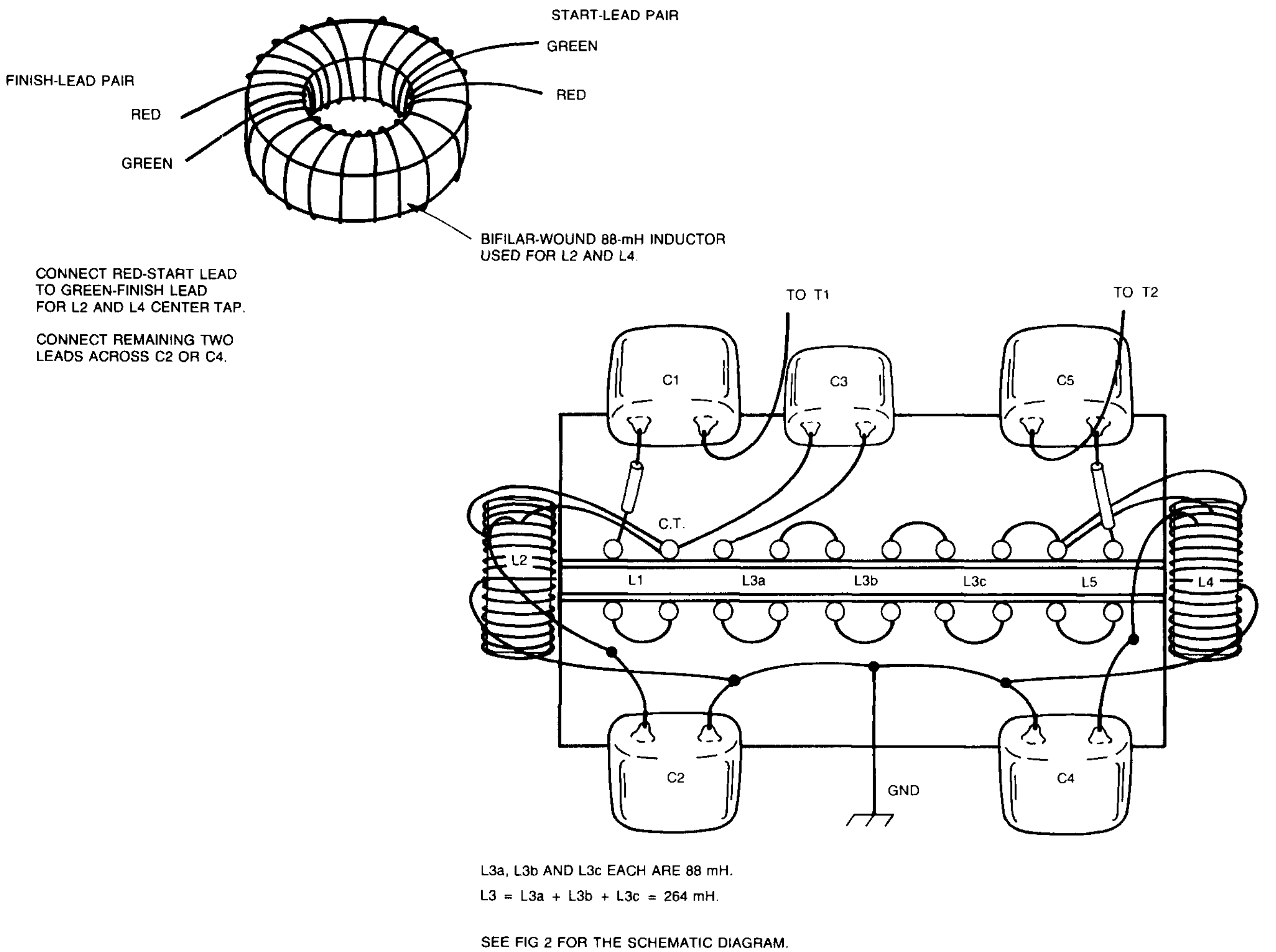


Fig 3—Here's how to connect the leads of the modified bifilar-wound, 88-mH inductor to make L2 and L4, and how to wire the 88-mH inductor stack for the narrowband CW audio filter. At A, the lead-connection instructions for L2 and L4 are shown. Wiring of the 88-mH inductor stack and the added inductors, L2 and L4, is shown at B. See designs 1-3 in Table 2 for the component values. L2 and L4 are fastened to each end of the inductor stack with silicone sealer.

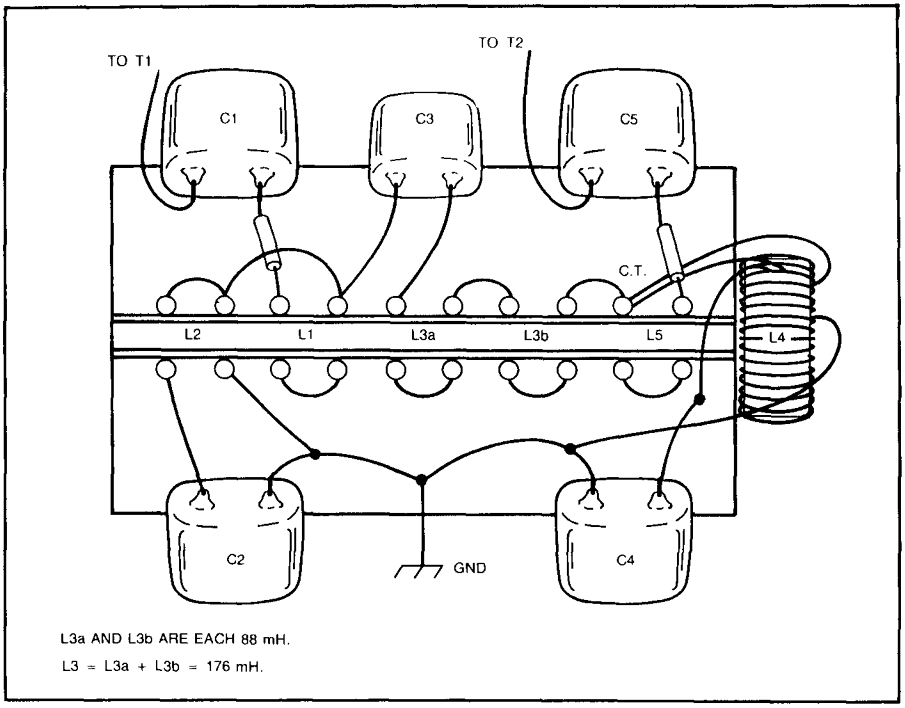


Fig 4—Here's how to wire the 88-mH inductor stack for the broadband CW audio filter. See designs 4 and 5 in Table 2 for the component values. See Fig 3A for the details on the lead connections for L4. L4 is fastened to the stack with silicone sealer.

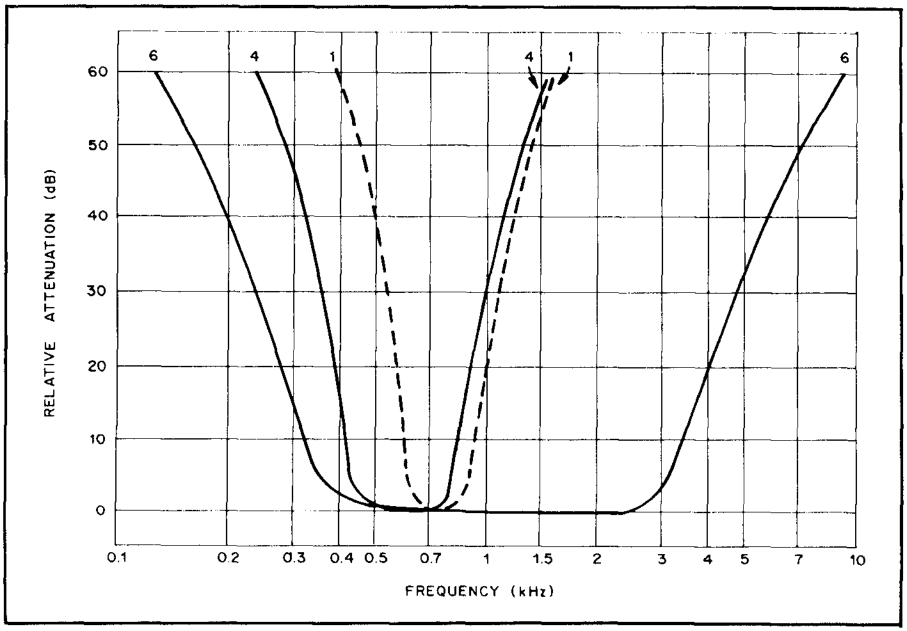


Fig 5—The measured relative attenuation responses of the CW and SSB audio filters. The curves are labeled with the same numbers used in Table 2. Table 1 summarizes the measured performance data.

Its value is determined by trial and error, and it will vary from 0 to 12 ohms or more depending on the filter design and the audio system.

SSB Audio-Filter Design

Only one design is needed for the SSB filter; that's shown in Fig 6. Fig 7 shows the wiring of the assembly. Table 2 (design no. 6) lists the design parameters and component values for the SSB filter. The measured response of the filter is summarized in Table 1 and shown in Fig 5. If a slightly different bandwidth or center frequency is desired, see Appendix B for an explanation of the design procedure. Like the CW filters, the nominal impedance level of the SSB filter is also 200 ohms. This makes it possible to switch between separate CW and SSB audio filters using a single pair of 8:200-ohm transformers.

Modifying the Inductors

The narrowband and broadband CW audio filters may require inductance values for L2 and L4 that are less than 88 mH. In this case, the required values are easily obtained by modifying a special inductor that is bifilar wound and has polyurethane-film-insulated wires. The more commonly available surplus inductor has two separate windings on opposite halves of the core, and its wires have insulation that must be scraped clean for soldering. By comparison, the bifilar-wound inductors are much easier to modify because turns are removed *in pairs* instead of individually, thus reducing the effort by half. Also, scraping the insulation from the toroid leads is unnecessary. The polyurethane-insulated leads are directly solderable: When a soldering-iron tip with a temperature of 750 to 800° is applied to the leads, the insulation vaporizes, and the exposed wire can be soldered. Determining how many turns to remove from the 88-mH toroid to get the different inductance values required (such as 43.5, 49.9 or 73.0 mH) for L2 and L4 is important.

The original inductance of the bifilar-wound inductors varies from about 87 to 89 mH. For best results, the inductor's exact value should be determined before the modification calculations are started. If you don't have an inductance-measuring bridge, you can still make accurate inductance measurements by resonating the inductor with a capacitor whose value has been measured accurately (to better than 0.5%), and then determining the resonant frequency of the combination with an audio-frequency signal source and a frequency counter. The inductance is then calculated from the measured capacitance and frequency.

Appendix C gives details on how to measure the inductance of the toroid and

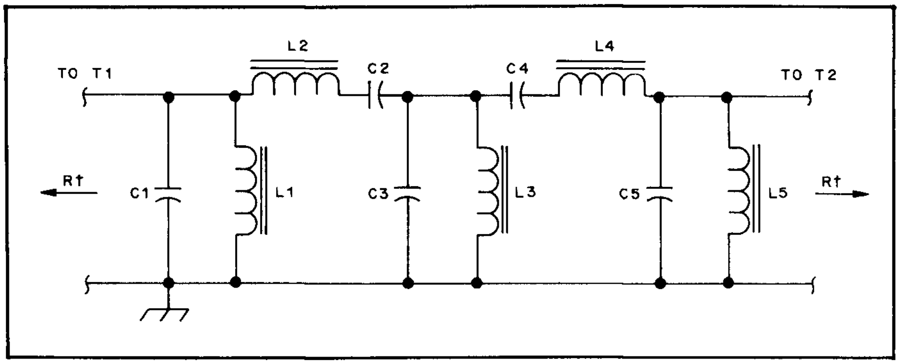


Fig 6—Schematic of the SSB audio filter. See design 6 in Table 2 for the component values and design parameters. The measured performance parameters are shown in Table 1. Fig 7 shows the wiring of the SSB audio filter.

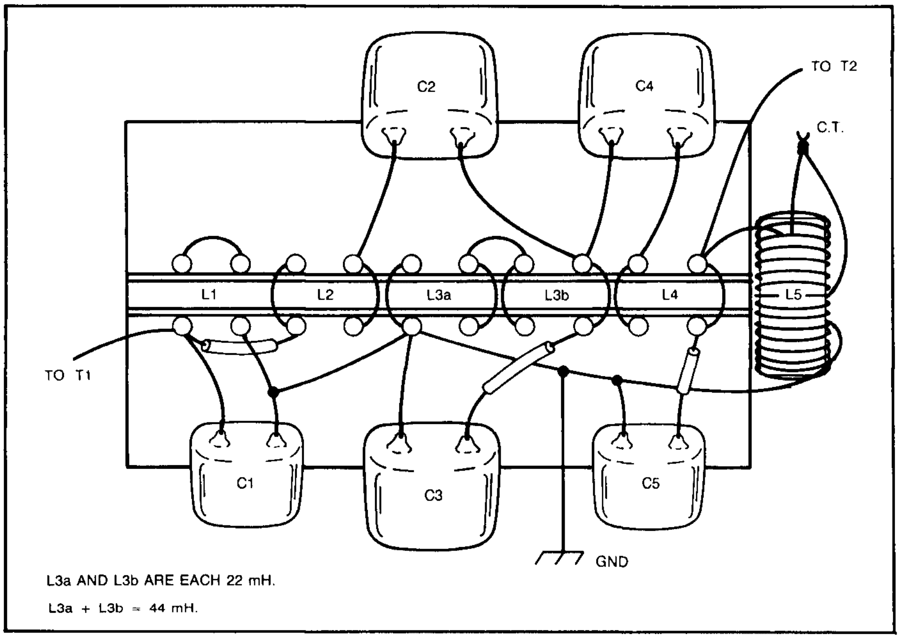


Fig 7—This is how to wire the 88-mH inductor stack for the SSB audio filter. The center-tap connection of L5 is not used in this application. L5 is fastened to the stack with silicone sealer.

how to modify it. If capacitance and frequency meters are unavailable, assume the inductance of the bifilar-wound inductor is 88 mH. To obtain a desired inductance (Ld), calculate the turns to remove (Td) using the equation

$$Td = 714 - (76.1105 \times \sqrt{Ld}) \qquad \text{(Eq 1)}$$

where

Ld is the desired inductance in mH and
Td is the number of turns to remove.

For example, to make a 43.5-mH inductor from a bifilar-wound 88-mH inductor, remove 212 turns or 106 turn-pairs. The number of turns removed is about 30% of the original turns.

Five-Resonator Filter Background

Although the design details of the five-resonator CW and SSB audio filters have been widely published,[5-13] the previously limited supply of inductor stacks prevented wide distribution of inductors for amateur use. With the present cooperation of the C & P Telephone Co of Maryland, it's now possible for any ham to construct these filters. Also, the earlier CW audio filter designs required eleven inductors and 8:1000-ohm transformers. By comparison, the latest design requires only seven inductors (at most), and uses the same impedance level as the SSB audio filter.

Summary

The low cost, easy assembly and high performance of these passive LC audio filters make them attractive to beginning and experienced radio amateurs. The design and construction details given here ensure easy assembly of the filters. If you want additional audio selectivity to enhance your communication effectiveness, this is your chance to get it.

Notes

[1] R. Schellenbach and F. Noble, "Switched-Capacitor Filters—An Emerging Technology for Amateur Radio Use," *QST*, Mar 1984, pp 19-25. See also R. Schellenbach and F. Noble, "Digital Switched-Capacitor Filters—A Practical Construction Project, *QST*, Jul 1984, pp 11-15.

[2] R. Arndt and J. Fikes, "SuperSCAF and Son—A Pair of Switched-Capacitor Audio Filters," *QST*, Apr 1986, pp 13-19. See also Feedback, *QST*, Oct 1986, p 51.

[3] E. Christian, *LC-Filters: Design, Testing and Manufacturing, Preface*, p ix (New York: John Wiley & Sons, 1983).

[4] I have made arrangements with the C & P Telephone Company of Maryland to make these inductor stacks available to me at no charge for distribution to radio amateurs. (The cooperation of the C & P Telephone Company in making their inductors available is greatly appreciated.) US and Canadian hams should send me a stamped, self-addressed 9 × 4½-inch envelope for details on how to obtain these inductors and other parts (mounting clip, matched capacitor set, 8:200-ohm transformers, etc) required in the filter construction. ARRL affiliated radio clubs will receive preference if I receive a letter from the club secretary specifying the number of amateurs who want to construct these CW or SSB audio filters.

European club members who want to obtain these inductors should write to Dave Aizlewood, G4WZV, 36 King Street, Winterton, South Humberside, DN15 9TP. A kit of parts including a matched capacitor set, inductors and transformers for a 750-Hz center frequency costs £12.00 plus £1.50 postage. Make your check payable to "G-QRP Club." The ARRL and *QEX* in no way warrant these offers.

[5] E. Wetherhold, "Modern Design of a CW Filter Using 88- and 44-mH Surplus Inductors," *QST*, Dec 1980, pp 14-19.

[6] E. Wetherhold, "High-Performance CW Filter," *ham radio*, Apr 1981, pp 18-25.

[7] E. Wetherhold, "A High-Performance CW Filter (for QRP reception)," *SPRAT, The Journal of the G-QRP Club*, Issue No. 32, Autumn 1982, pp 3-5. (The article describes an 88-mH, two-stack design with one added 44-mH inductor.)

[8] E. Wetherhold, "A CW Filter For the G3RJV 'Superex' Receiver," *Short Wave Magazine*, Aug 1983, pp 307-311.

[9] E. Wetherhold, "How to Build A CW Filter For The Novice Operator", Parts 1 and 2, *CQ*, Feb and Mar 1985, pp 70-73 and 72-76, respectively.

[10] E. Wetherhold, "A CW Filter For the Radio Amateur Newcomer", *Radio Communication*, Jan 1985, pp 26-31.

[11] E. Wetherhold, "Easy-to-Build One-Stack CW filter Has High Performance and Low Cost," *SPRAT, The Journal of the G-QRP Club*, Issue No. 54, Spring 1988, pp 20-21.

[12] E. Wetherhold, "A Simple, High-Performance CW filter," and "A Passive Audio Filter for SSB," p 28-3, M. Wilson, ed., *The 1988 ARRL Handbook* (Newington: ARRL, 1987), pp 28-1 to 28-3.

[13] E. Wetherhold "(One-stack) CW filter Combines High Performance and Low Cost," *Radio Handbook*, 23rd edition, W. Orr, Ed., (Indianapolis: Howard W. Sams and Co, 1987), p 13-4.

APPENDIX A

Derivation and Calculations of CW Filter Element Values and Parameters

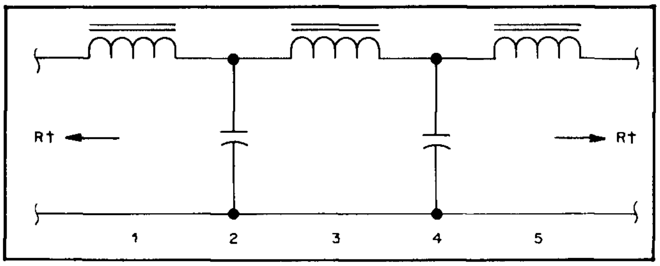


Fig A1—The L-in/out configuration is used for the narrowband and broadband CW filter designs. Components are numbered left to right and are listed in Table A1 with their normalized values.

Table A1

Reflection Coeff (%)	*G1, G5 (H)*	*G2, G4 (F)*	*G3 (H)*	*G3/G1 Ratio*
0.0441	0.1054	0.2625	0.3162	3.000
6.3	0.1642	0.2657	0.3284	2.000

The Chebyshev component values (G1-G5) are normalized for 1-ohm terminations and for a 3-dB cutoff frequency of 1 Hz.

The derivation procedure follows:

1) Assume the filter termination impedance (Rt) will be the sum of an 8-to-200-ohm transformed impedance, the transformer winding resistances transferred to the high-Z winding and the resistance of L1 or L5, or a total of about 230 ohms. Also, assume the design Rt will be 4 × 230 = 920 ohms, and inductors L1 and L5 will be 4 × 88 = 352 mH.

2) For a narrowband CW audio filter design, use the normalized values for RC = 0.0441%, where L1 = 352 mH, and L3 = 3 × L1. For a broadband design, use the values for RC = 6.3%. The calculations for a narrowband design follow.

3) Calculate the 3-dB bandwidth (B3) based on

Rt = 920 ohms,
L1 = 0.352 H and
G1 = 0.1054 H:
B3 = (G1 × Rt) / L1
= (0.1054 × 920) / 0.352
= 275.5 Hz.

Because of inductor losses, the actual 3-dB bandwidth will be about 8% narrower than the calculated B3.

4) Find the values of C2 and C4 based on

B3 = 275.5 Hz:
C2 = G2 / (Rt × B3)
= 0.2625 / (920 × 275.5)
= 1.0357 μF.

Summarizing the low-pass filter design values:

L1 and L5 = 0.352 H,
L3 = 1.056
H, C2, 4 = 1.0357 μF,
Rt = 920 ohms and
B3 = 275.5 Hz.

This low-pass filter design will be transformed into a band-pass filter having a center frequency (Fc) selected by the designer. Fig A2 shows the band-pass filter schematic diagram.

5) Assume that Fc is to be 750 Hz. Calculate L2 based on C2 = 1.0357 μF where L2, C2 and Fc are expressed in millihenrys, microfarads and kilohertz, respectively:

L2 = 25.33 / (C2 × Fc^2)
= 25.33 / (1.0357 × 0.75^2)
= 43.48 mH.

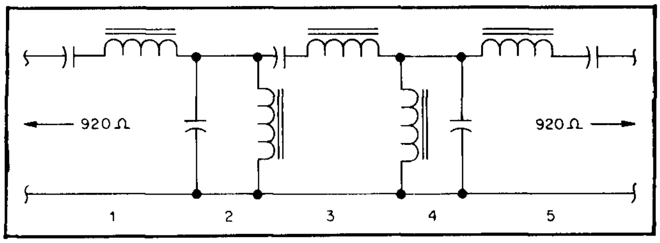


Fig A2 —Band-pass filter schematic diagram.

L2 and L4 are made by removing about 30% of the turns from an 88-mH inductor. (See Appendix C for details.) Because the values of L2 and L4 can be varied, you can pick any Fc between 500 and 1000 Hz.

6) The relative bandwidth percentage

(BW%) = 100 × B3 / Fc
= 275.5 / 7.5
= 36.7%.

The minimum Q required to obtain a close approximation of the calculated response is

Qmin = 20 × Fc / B3
= 54.

Because inductor Q at 750 Hz is only about 40, the difference between the calculated and actual responses will be somewhat different. Nevertheless, an inductor Q of 40 will be adequate for this application.

7) C1, C5 and C3 are calculated based on

L1 and L5 = 352 mH,
L3 = 1056 mH and
Fc = 0.750 kHz:
C1 = 25.33 / [L1 × (Fc^2)]
= 0.127929 μF, and
C3 = C1 / 3
= 0.12793 / 3
= 0.042643 μF.

Summarizing all values calculated so far:

C1, C5 = 0.12793 μF	L3 = 1056 mH
C2, C4 = 1.0357 μF	B3 = 275.5 Hz
C3 = 0.042643 μF	Fc = 750 Hz
L1,5 = 352 mH	Rt = 920 ohms
L2,4 = 43.48 mH	

All LC products = 45.03 × 10^{-9} to give Fc = 750 Hz.

8) The three series branches are moved to the L2 and L4 center taps. The series-branch reactances become one-quarter of their former values. The values of L1, L3 and L5 are quartered and the values of C1, C3 and C5 are quadrupled. C2, C4 and L2, L4 remain unchanged. The values of L1, L3 and L5 now can be realized with one or more 88-mH inductors. Rt becomes 230 ohms. Fig A3 shows the final diagram and component values.

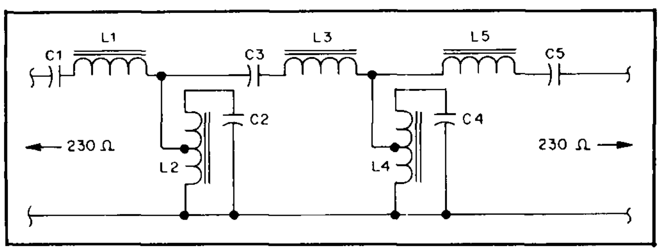
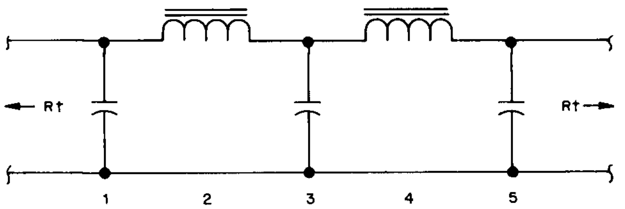


Fig A3—Schematic diagram and component values of the CW filter design for Fc = 750 Hz and Rt = 230 ohms.

C1, C5 = 0.512 μF,
C2, C4 = 1.036 μF,
C3 = 0.1706 μF,
L1, L5 = 88 mH,
L2,4 = 43.5 mH,

APPENDIX B

Derivation and Calculations of SSB Audio-Filter Element Values and Parameters



The components are numbered 1 to 5, left to right.

G1 and G5 = 0.1642 F,
G3 = 0.3284 F,
G2 and G4 = 0.2657 H,
G3 / G1 = 2.
Rt = 1 ohm.

The reflection coefficient for a 5th-order Chebyshev design = 6.3%.

Fig B1—The filter schematic is shown with the normalized component values (G1-G5) of the low-pass filter prototype used in designing the SSB band-pass filter. The component values are normalized for 1-ohm terminations and for a 3-dB cutoff frequency of 1 Hz. A reflection coefficient of 6.3% was selected so that G3 is twice G1.

The steps in the derivation procedure follow:

1) L2 and L4 must be 22 mH to allow use of a standard inductor value. Rt is to be 224 ohms so an 8:200-ohm transformer (Mouser 42TU200) can be used. The exact Rt value is based on an 8-ohm source transformed to 200 ohms and the transformer winding resistances referred to the high-impedance winding (200 + 12 + 12 = 224 ohms).

Find the 3-dB bandwidth (B3) of the desired band-pass filter based on the low-pass filter prototype in Fig B1:

B3 = (Rt × G2) / L2
= (224 × 0.2657) / 0.022
= 2705.3 Hz.

2) Calculate the values of C1 and C3 for a 3-dB bandwidth of 2705.3 Hz:

C1 and C5 = G1 / (Rt × B3)
= 0.1642 / (224 × 2705.3)
= 0.27096 μF.
C3 = 2 × C1
= 2 × 0.27096 μF
= 0.5419 μF.

3) Transform the low-pass filter prototype in Fig B1 into a band-pass design by resonating all components to Fc. See Fig B2 for the band-pass filter schematic diagram.

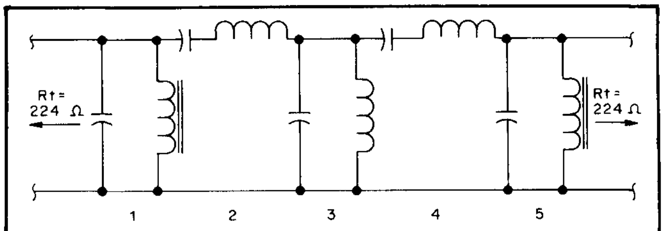


Fig B2—Band-pass filter schematic diagram.

4) Calculate the passband center frequency based on the values of C1 and L1. C1 has already been determined. L1 must be 88 mH to allow use of a standard inductor value that gives an Fc near the center of the desired SSB band-pass filter. Since the desired 3-dB lower and upper frequencies are about 300 and 3000 Hz, Fc should be within 10% of

$\sqrt{300 \times 3000}$ = 950 Hz.

Calculate the actual value of Fc by using the values of C1 and L1:

Fc = 159.155 / $\sqrt{\text{L1} \times \text{C1}}$

where Fc is expressed in hertz, L1 is in henrys and C1 is in microfarads.

L1 = 0.088 H and
C1 = 0.27096 μF.
Fc = 159.155 / 0.1544166
= 1030.686
= 1030.7 Hz.

5) Calculate the value of L3:

L3 = L1 / 2
= 0.088 / 2
= 44 mH.

6) Based on the 3-dB bandwidth (B3) and Fc, calculate the lower and upper 3-dB frequencies, F3L and F3U:

F3L = $\sqrt{\text{Fc}^2 + \text{X}^2}$ − X

where X = B3 / 2 = 1352.65 Hz

F3L = $\sqrt{1030.7^2 + 1352.65^2}$ − 1352.65
= 347.94 Hz.
F3U = F3L + B3
= 347.94 + 2705.3
= 3053.2 Hz. As a check,
Fc = $\sqrt{\text{F3L} \times \text{F3U}}$
= $\sqrt{347.94 \times 3053.2}$
= 1030.7 Hz.

Reviewing the calculated design parameters, you can see that the Fc, B3, F3L and F3U values for the SSB audio filter application are satisfactory, and the design is acceptable. If not, the Rt value of 224 ohms may be varied within 10% to obtain slightly different parameter values.

The relative bandwidth percentage is BW% = 100 × B3 / Fc = 262.5%. The minimum required inductor Q to obtain a close approximation of the ideal response is Qmin = 20 × Fc / B3 = 7.6.

The results of these calculations indicate that there should be no difficulty in using the 88/22-mH inductors to obtain a band-pass response that closely approximates the ideal response because the inductor Q at 1000 Hz is about 50. Only when a relative bandwidth of less than about 20% is required will the actual bandpass response be significantly less than ideal because of limited inductor Q.

7) Calculate the value of C2 and C4 based on L2 = 22 mH and the Fc value where C2, L2 and Fc are expressed in microfarads, millihenrys and kilohertz, respectively:

C2 and C4 = 25.33 / (L2 × Fc^2)
= 25.33 / (22 × 1.0307^2)
= 25.33 / 23.3715
= 1.0838 μF.

Note: All circuits are tuned to Fc = 1030.7, and all tuned circuits have an LC product of 23.84 × 10^{-9}.

By using other Rt values that can be matched with standard transformers (such as 8:500 or 8:1000 ohms), other distinctly different bandwidths may be obtained.

APPENDIX C

How to calculate the number of turns to remove from a bifilar-wound 88-mH inductor to obtain any desired inductance.

1) A bifilar-wound inductor is identified by its red- and green-colored insulated wires. The polyurethane-insulated wires are solderable at 750 to 800°F, and the leads do not need to be scraped to remove the insulation. Caution: The fumes generated during soldering are irritating to the lungs and eyes, so keep your face away from the fumes, and solder only in a well-ventilated area.

2) Measure the original inductance, Lo, with the two windings connected in series-aiding. To do this, connect the red start lead to the green finish lead, and connect the other two leads to an inductance bridge. An alternate method of finding the inductance is to resonate the inductor with a known capacitance and calculate the inductance based on the capacitance and resonant frequency values. For example, connect the inductor in parallel with a 0.27-μF capacitor (its value previously measured to an accuracy of better than 0.5%) and lightly couple (using 470-pF capacitors) an audio generator and an ac voltmeter to the tuned circuit. Vary the generator frequency until a voltage peak is indicated by the voltmeter. Measure the resonant fre-

quency (it'll be approximately 1033 Hz) with a frequency counter and calculate the inductance using the equation:

$$Lo = 25.33 / (F^2 \times C),$$

where Lo, F and C are expressed in millihenrys, kilohertz and microfarads, respectively.

3) Remove 50 turn-pairs (total turns removed = 100) and again connect the windings in series-aiding. Measure the modified inductance, Lm.

4) Calculate $To = 100 \times R / (R-1)$,

where

$$R = \sqrt{Lo / Lm}$$

To = original number of turns on inductor core, Lo = original inductance in the series-aiding connection, and Lm = modified inductance after removing 100 turns (50 turn-pairs). For example, if

Lo = 89.10 mH and
Lm = 67.09 mH, then
R = 1.152417796 and
To = 756 turns.

5) Calculate:

$$S = (To - 100) / \sqrt{Lm}$$

where Lm is the modified inductance after removing 100 turns (50 turn-pairs) from the inductor.

For example, if an inductor has

Lo = 89.10 mH,
To = 756 and
Lm = 67.09 mH for 100 turns removed, then

$$S = (756 - 100) / \sqrt{67.09}$$
$$= 80.0894.$$

6) Use the following general equation (applicable to all bifilar-wound inductors with Lo = 89.10 mH) to find the number of turns to remove to obtain a specific inductance:

$$Td = To - (S \times \sqrt{Ld}),$$

where Td is the number of turns to remove from an unmodified 89.10-mH inductor, To is the number of original turns on the inductor core, Ld is the desired inductance in mH, and S is the value calculated in (5).

For example, for the values given in (5) and if the desired inductance (Ld) is 43.48 mH, then:

$$Td = 756 - (80.0894 \times \sqrt{43.48})$$
$$= 756 - 528$$

= 228 turns, or 114 turn-pairs to be removed from the original inductor. Because 100 turns have already been removed, an additional 128 turns (228-100 = 128) or 64 turn-pairs must be removed to get 43.48 mH.

Bits

FCC Upholds RF Lighting Standards

The FCC has denied Linear Corporation's request to reconsider its decision (General Docket 83-806) concerning existing technical standards for RF lighting devices. (RF lighting devices convert 60-Hz ac electrical power into RF energy. This RF energy stimulates a gas inside a glass bulb, giving off light. RF lighting devices normally operate below 30 MHz.)

In its petition, Linear Corporation contended that the FCC's decision to retain the existing standards for RF lighting devices fail to promote the use of radio services in accordance with the Communications Act of 1934. Linear Corporation disagreed with the Commission's decision with respect to issues such as interference to Part 15 devices, end-of-life phenomenon (the increase of emission levels from an RF lighting device when the device nears the end of its useful life), single-user interference and labeling requirements. Linear Corporation also alleged that rule changes were made in the classification of RF lighting devices, and in the field strength requirements for industrial, scientific and medical (ISM) equipment, without proper public notice.

The FCC concluded that equipment used under the provisions of Part 15 should continue to be required to accept harmful interference from any other devices, including RF lighting devices. The Commission specified three main reasons for rejecting reconsideration; (1) that the original proceeding was initiated to study the interference potential of RF lighting devices, (2) that operating conditions were beyond the scope of the original proceeding and (3) that changes in the basic policy of Part 15 should be treated in a separate proceeding.

In addition, the Commission concluded that existing emission limits for RF lighting devices are adequate to protect authorized services. The existing limits for emissions from RF lighting devices are the same as those currently applied to computing or digital equipment. The two emissions limits are non-consumer (industrial, commercial or business applications) and consumer (predominately residential applications). The consumer limits are more restrictive than the non-consumer limits to account for the higher probability of interfering with broadcast receivers and other equipment in a residential environment. The Commission stated that these emission limits have been effective in preventing harmful interference.

In general, the Commission found that Linear Corporation did not present any information in its petition which had not been previously considered. The Commission also found that Linear Corporation's petition did contain any persuasive arguments or evidence to warrant a change in policy regarding the status of Part 15 devices, end-of-life phenomenon, single user interference, and labeling requirements. Finally, the Commission indicated that no changes were made in the classification of RF lighting devices or in the permissible field strength levels for ISM equipment. Instead, the Commission only clarified what is meant by consumer and non-consumer applications.—*Tom Francis, NM1Q*

Kilovac Relays and Product Guide

Kilovac® has introduced four new models to its line of relays for hot-switching applications. Power-switching applications, such as the capacitive discharge of a heart difibrillator, electrostatic discharge simulator (ESD), safety interlock switches and pulsed lasers, are notorious for causing contact erosion. The Kilovac relays offer a solution to inherent contact degradation resulting from these types of hot-switching applications. These four additions to Kilovac's family of gas-filled relays meet a broad range of application requirements, with voltage ratings from 3.5 kV to 100 kV and up to 1.2 kA peak.

The new Kilovac relays are filled with an electronegative gas that offers excellent arc-quenching characteristics. As a result of the gas environment, the relays have an extended life and offer improved overall performance. Kilovac claims these relays are the smallest and lightest relays with these ratings (see the accompanying table) available in industry today.

New Series Part No.	*Form*	*Operating Voltage (kV dc)*	*Continuous Current (A dc)*	*Single-Quantity Price*
HC-5	SPDT	3.5	8	$108
KC-15	SPDT	15.0	12	$212
KC-16	SPDT	15.0	12	$202
K6OC	SPDT	30.0*	10	$265

*Requires encapsulation by customer; dimensionally equal to the KC-16.

A detailed description of all Kilovac's relays is available on request. Contact Kilovac at PO Box 4422, Santa Barbara, CA 93140, tel 805-684-4560, and ask for Kilovac's High Voltage Relay Product guide.—*Paul K. Pagel, N1FB*

Path Selection—Part 1

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Mountaintop microwave DXing, ATV work and repeater linking are among pursuits which require evaluation of the communication paths. Path analysis involves both terrain analysis and working out a path budget—in decibels, not dollars. Fortunately, neither of these related tasks is difficult. In fact, they can be done from the comfort of your shack. A little planning with maps and a calculator or computer can save a long drive or hike up a mountain, or it might reveal a longer workable path than the one you had in mind.

UHF and microwaves are usually thought of as being suitable only for line-of-sight communications. But "line of sight" has meaning in radio communications that few noninitiates understand. Path analysis involves the understanding of how radio signals travel over the earth, earth geodesy, and interpreting terrain effects. This article provides a review of these techniques. In part 1, I'll discuss techniques for determining the radio horizon for a particular set of conditions. Next month, I'll cover obtaining and using topographic maps to select actual communications paths.

Radio Refraction

Radio signals traveling through free space essentially move in a straight line. Terrestrial communications, however, take place through the earth's atmosphere, which makes signals almost *never* travel in a straight line. Unlike free space, the earth's atmosphere is neither a vacuum nor homogenous. At different altitudes and locations along a given path, the temperature, moisture content and atmospheric pressure vary, significantly impacting signal paths. Usually, these variations make air less dense at higher altitudes than near the ground. Density usually decreases with altitude by a fairly predictable gradient, but not even that always holds true. Not only does the gradient sometimes change in slope, but temperature and humidity inversions, and other anomalous conditions, can cause temporary—sometimes bizarre—changes in the direction and uniformity of the gradient.

Radio signals travel a bit slower in a dense medium that in a thinner one. Therefore, normal atmospheric conditions cause signals to refract (bend) towards the earth's surface. This tendency means that the earth is effectively larger (or flatter), as far as RF is concerned, than it really is. The amount of this flattening effect is designated by the symbol K, for the earth's *effective* radius. If the earth had no atmosphere and there was thus no signal refraction, K would be 1 (K = 1 is also known as K for "true earth").

In temperate, inland climates, normal atmospheric conditions have a gradient that causes K to be about 4⁄3 (1.33). This is a commonly used rule of thumb for radio refraction. But, like all rules of thumb, the actual value is usually something else. The atmosphere in dry, mountainous areas usually has a gradient which yields a K value closer to 1.25. In warm, humid areas like the Gulf coast, the gradient makes K closer to 1.6. The atmospheric gradients change seasonally and diurnally (with the time of day). Table 1 shows typical summer and winter daytime (1500 UTC) K values at various locations.

If you know the temperature, atmospheric pressure, relative humidity and elevation at a given location, you can estimate K for that location. The gradient that determines refractivity correlates well in most of the world with the surface radio-refractivity index.

Table 1—*Typical K Values*

Location	*K*
Gulf and California coast, summer	1.62
Dixie and west coast, summer	1.54
Inland plains and northeast, summer	1.47
Gulf and California coast, winter	1.40
Northern Dixie and west coast, winter	1.34
Inland plains and northeast, winter	1.30
Mountains, summer and winter	1.26

The following series of equations can be used to make close estimates of K. First, you must make an estimate of the radio-refractivity index at sea level from the following equation:

$$N_0 = \left(\frac{77.6}{T}\right) \times \left(\frac{4810\ U\ e_s}{T}\right) + P \qquad \text{(Eq 1)}$$

where
N_0 = sea-level radio refractivity index
T = temperature in Kelvins
P = atmospheric pressure in millibars
e_s = saturation vapor pressure in millibars
U = relative humidity (expressed as a decimal)

The following conversion factors can be used to arrive at the units used in Eq 1:

$$\text{Kelvins} = 273.15 + {}^\circ\text{C} \qquad \text{(Eq 2)}$$

$$^\circ\text{C} = ({}^\circ\text{F} - 32)/1.8 \qquad \text{(Eq 3)}$$

$$P_{mb} = 33.86\ p_{in.\ Hg} \qquad \text{(Eq 4)}$$

Vapor pressure is the portion of the total atmospheric pressure that is attributable to water vapor in the air. Moist air exerts greater pressure than dry air. Saturation vapor pressure, e_s, is the pressure added when the air holds as much water vapor as possible at a given temperature. The value of e_s in millibars may be estimated from temperature using

$$e_s = \exp\,(1.805 + 0.0738T - [2.89 \times 10^{-4}]\ T^2) \qquad \text{(Eq 5A)}$$

where
T = temperature in degrees Celsius
exp (x) = e^x (e = 2.7183, the base of natural logarithms)

or

$$e_s = \exp\,(0.401 + 0.0467T - [8.93 \times 10^{-5}]\ T^2) \qquad \text{(Eq 5B)}$$

where
T = temperature in degrees Fahrenheit.

For locations not at sea level, it is now necessary to make an adjustment for the surface elevation. Higher elevations have lower refractivity indexes than lower elevations. Refractivity index can be estimated from

$$N_s = N_0 \exp(-0.1057 \times h) \qquad \text{(Eq 6A)}$$

where

N_s = surface radio refractivity index
N_0 = radio refractivity index at sea level
h = surface elevation in kilometers

or

$$N_s = N_0 \exp(-0.03222 \times h) \qquad \text{(Eq 6B)}$$

where

h = surface elevation with respect to sea level (SL) in thousands of feet

The earth's effective radius, K, can now be estimated from the following:

$$K = \frac{1}{(1 - 0.04665 \exp[5.577 \times 10^{-3} N_s])} \qquad \text{(Eq 7)}$$

Eqs 5 through 7 are regression, or curve-fitting, equations. These equations are descriptive of the relationship between variables, but do not define that relationship. Regression equations are used when the definitive mathematical relationship is too complex for practical work.

Example: Pullman, Washington, is 2600 feet above mean sea level. On a warm summer day, with a temperature of 82 °F, atmospheric pressure of 30.1 inches of mercury (Hg) and a relative humidity of 45%, what is K?

To solve this, first convert the parameters to those used by Eq 1. From Eqs 2 and 3, the temperature is found to be 300.9 K. From Eq 4, the atmospheric pressure in metric units is found to be 1019.2 mb. The saturation vapor pressure is estimated with Eq 5B and found to be 37.7 mb. From these parameters, the sea level radio refractivity index, N_0, is found to be 332.7. At 2.6 thousand feet ASL, Eq 6B gives a surface refractivity, N_s, of 306.0 at Pullman. Finally, Eq 7 gives us an estimate of the corresponding K: 1.346. This is very close to the rule of thumb for average, temperate climates—K = 4/3 (1.333).

To demonstrate how this rule of thumb can vary, and to further test the calculations, what happens to K if all parameters stay the same except the relative humidity, which is reduced to 65%? To 25%? The correct answers are 1.432 and 1.281, respectively—quite significant variances.

Earth Curvature

Once the effective earth's radius is known, it is possible to calculate the curvature of the earth at any point along the path from the following equations:

$$H = \frac{d_1 d_2}{12.746\ K} \qquad \text{(Eq 8A)}$$

where

H = earth curvature in meters, and
d_1 and d_2 = distances in kilometers of the point from each end of the path

or

$$H = \frac{d_1 d_2}{1.5\ K} \qquad \text{(Eq 8B)}$$

where

H is in feet
d_1 and d_2 are in miles.

These heights must be added to each point along the path and to the terrain elevation to compensate for earth curvature.

Fig 1 shows a 47.36-mile path that was plotted as if the earth was flat (K = ∞). (Although 4/3 curved graph paper is available, it is difficult to find, and using it restricts you to one assumed value of earth curvature.) The centerline between the two antennas in Fig 1 is plotted as the top curved line, allowing for earth curvature. Mathematically, this is the same drawing as the earth curvature above the terrain, or as drawing the terrain on 4/3 curved graph paper.

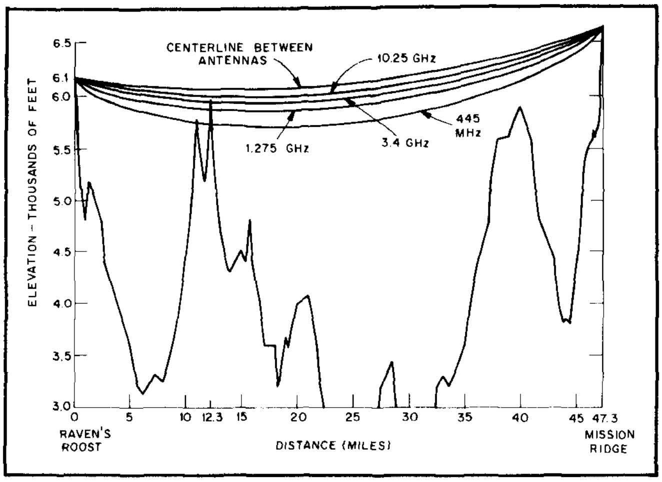

Fig 1—Cross-sectional view of the terrain between Raven's Roost and Mission Ridge. Antenna height at both ends of the path is 15 feet. The top curved line represents the centerline between the antennas, showing earth curvature. The curves plotted are the first Fresnel-zone radii for (from the second curve from the top) 10.25, 3.4 and 1.275 GHz, and 445 MHz. See text.

Assuming a path length of 75 kilometers and the atmospheric conditions from the refractivity example above, the earth curvature to add at midpoint on the path (with relative humidity at 45% and a K of 1.346) is 82 meters (from Eq 8A). But, at 25% humidity, K drops to 1.281, increasing the curvature allowance to 86 meters. At 65% humidity, K increases to 1.432, decreasing the curvature allowance to 77 meters.

Fresnel Zones

Fresnel (correctly pronounced "frayNEL," but more commonly pronounced "freNEL") zones are ellipsoid areas surrounding the direct path between the two antennas. The lower four curves in Fig 1 are first Fresnel zone radii plotted for four UHF and microwave amateur bands: 10.25, 3.4 and 1.275 GHz, and 445 MHz. Fresnel zones are important because they can be used to predict obstruction loss, and because knowing where they are can help you to minimize the adverse effects of reflections on permanent paths.

If you draw lines from any point on the boundary of an odd-numbered Fresnel zone to each antenna, the length of those two lines will be an odd multiple of a half wavelength longer than the direct path. In the case of the first Fresnel zone, one of particular interest, the distance is ½ λ longer. If you draw lines from any point on the boundary of an even-numbered Fresnel zone to each antenna, the length of those two lines will be an even multiple of a half wavelength longer than the direct path. In the case of the second Fresnel zone, the distance is one wavelength longer.

The loss curve first described by Bullington,[1] a Bell Labs scientist, is shown in Fig 2. The curve shows that if a signal clears 0.6 of the first Fresnel zone radius (designated $0.6F_1$) over an obstruction, received signal strength will not be attenuated by the obstruction. For clearances less than $0.6F_1$, the amount of obstruction loss depends upon the nature of the obstruction. So-called knife-edge obstructions have the least amount of loss, while smooth, spherical obstructions have the greatest amount of attenuation.

[1]Notes appear on page 13.

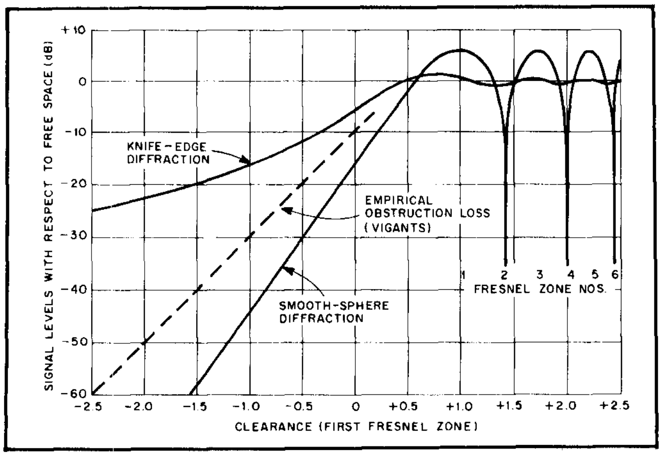

Fig 2—Path loss resulting from obstructions, as a function of Fresnel-zone signal clearance. The upper curve represents signal loss over knife-edge surfaces. The least attenuation results from such obstructions. Vigants' empirical results are shown by the broken line. The lower curve represents losses resulting from rounded surfaces. These obstructions cause complete signal cancellation at even-numbered Fresnel-zone clearances. See text.

The dashed line in Fig 2 is based on observations made by Vigants,[2] another Bell Labs scientist. The dashed line tracks actual abnormal atmospheric conditions in northeastern Florida. The Bullington curve shows that, especially for smooth, spherical obstructions, clearances of an even-numbered Fresnel zone can result in near total cancellation of the received signal because of cancellation from the indirect signal. This happens even if ample line-of-sight clearance is available. This is especially important for permanent paths over water or other reflective surfaces, but has relatively minor consequences for amateur work.

At any point along the path, any Fresnel zone radius can be calculated from

$$F_n = 17.314 \sqrt{\frac{d_1 \, d_2 \, n}{fD}} \qquad \text{(Eq 9A)}$$

where
F_n = Fresnel zone radius in meters
d_1 and d_2 = the distances described in Eq 8
n = Fresnel zone number
f = frequency in GHz
D = path length in kilometers

An alternative to Eq 9A is

$$F_n = 72.064 \sqrt{\frac{d_1 \, d_2 \, n}{fD}} \qquad \text{(Eq 9B)}$$

where
F_n is in feet
d_1, d_2 and D are in miles.

Because $0.6F_1$ is of most interest, Eq 9 can be simplified somewhat by eliminating n from the equation and making the constants in Eq 9A and 9B 10.39 and 43.24, respectively.

In Fig 1, notice the effect of the obstruction 12.3 miles from the Raven's Roost site. This 6000-foot ridge is as close to a true knife edge as you are likely to encounter on a practical path. Although the radio signal has line of sight between each antenna at K = 1.25 (normal for the path), the first Fresnel zone radii are more and more obstructed as the frequency gets lower.

At 10.25 GHz, the clearance of the signal is $0.95F_1$, meaning 95% of the first Fresnel zone is cleared. From Fig 2, it is evident that because this is greater than $0.6F_1$, no obstruction loss will be noted. At 3.4 GHz, the clearance is $0.55F_1$, so only a small amount of obstruction loss will be noted. However, at 445 MHz, the clearance is only $0.2F_1$, so obstruction loss on the order of 3 or 4 dB will occur. Had that obstruction been a smooth sphere instead of a knife edge, the loss could have been as much as 13 or 14 dB.

To calculate the clearances just cited, it is first necessary to determine the "flat earth" height of the radio signal as it passes over the obstruction. The earth seems flat to a radio wave when $K = \infty$. A simple ratio gives us that height:

$$h_x = h_1 - \frac{d_1(h_1 - h_2)}{D} \qquad \text{(Eq 10)}$$

where
d_1, d_2 and D are the same as in Eq 9,
h_1 = height of the antenna at the d_1 end
h_2 = height of the antenna at the d_2 end
d, h, and D must be in the same units

For the path in Fig 1, h_1 = 6638 feet, h_2 = 6175 feet, d_1 = 12.3 miles, D = 47.36 miles, and $d_2 = D - d_1$ = 35.06 miles. Using these values in Eq 10 gives h_x = 6295 feet. This is the clearance that would occur if the earth was flat ($K = \infty$). We are more interested in the clearance at a normal K (in this case about 1.25), so we must use Eq 8B. Doing so, we find we must add a 230-foot earth-curvature allowance to the 6000-foot peak, which is the closest potential obstruction.

Our clearance, then, at K = 1.25 is 6295 − 6230 feet or 65 feet. At 1275 MHz, what is the first Fresnel zone radius and how much of it is obstructed by the peak when K = 1.25? Applying Eq 9, we find that F_1 = 193 feet at this frequency, 12.3 miles from Raven's Roost. Therefore we find that 65/193 or 34% of the first Fresnel zone ($0.34F_1$) is cleared.

Next month, I'll continue the discussion with notes on using topographic maps and other information to select optimum paths.

Notes

[1]K. Bullington, "Radio Propagation Fundamentals," *Bell System Technical Journal*, May 1957, pp 593-626.
[2]A. Vigants, "Microwave Radio Obstruction Fading," *Bell System Technical Journal*, Jul-Aug 1981, pp 785-801.

Bits

RF Technology Expo 89

The 1989 RF Technology Expo will be held at the Santa Clara Convention Center, Santa Clara, California, from February 14-16, 1989. Over 120 exhibitors will be present. The program includes:

- Measurement techniques
- Receiver topics
- VCO and synthesizer design
- Passive-circuit design
- MMIC technology
- RF power topics
- UHF and microwave components
- Manufacturing and reliability engineering
- Design techniques

For more information, write RF Technology Expo 89, 6300 S Syracuse Wy, Suite 650, Englewood, CO, 80111.—*Rus Healy, NJ2L*

Correspondence

NTIS Publications

Every year, the US Government publishes thousands of technical reports on almost every subject imaginable. Obviously, technological subjects head the list, with radio and electronics included in that broad area. To distribute these reports, the government has set up an organization called the National Technical Information Service (NTIS). NTIS is a branch of the Department of Commerce. NTIS is self-funding, which means that it charges its customers to cover its operating expenses, so that it does not require a government subsidy to operate (surprise!). NTIS distributes reports in two forms: paper copies (8½ × 11-inch photocopies) and microform (commonly called microfiche). For a variety of reasons, the least-expensive method is microfiche. But, don't let the fact that you don't own a microfiche reader discourage you. Most public libraries have at least one reader, and will usually let you use it.

NTIS also distributes copies of reports by some foreign sources. This is helpful because some subjects that have only limited interest in the US are heavily studied in other countries. Recently, I read three NTIS publications on antennas that might be of interest to you.

(When ordering an NTIS publication, please include the NTIS order numbers. NTIS has a handling charge of $3 per order for prepaid orders. The mailing address is: National Technical Information Service, 5285 Port Royal Rd, Springfield, VA 22161.)

A Practical Guide To The Design of Rhombic Aerials, by J. N. Taylor, Royal Aircraft Establishment (UK), April 1985, 39 pp. NTIS order number N87-14566. Paper cost, $12.95; microform, $6.95.

For many years, those with an interest in the design of rhombic antennas were restricted to searching for design information in old books and technical journals. This information void has now been filled by a technical paper from England. This guide uses charts derived from computer projections to reduce the burden of calculations. In addition, the guide covers optimization for design variations such as: reduced length, reduced height, reduced height and length, reduced height and increased length, maximum gain and desired elevation (take-off) angle. The author also considers various design tradeoffs and problems, such as the effects of real ground.

The only problems with this guide are related to the computer analysis. No listing of the programs used to derive the charts is given (even though all the formulas needed are in the report). Also, the radiation patterns were prepared with a printer with very limited graphics capabilities, resulting in poor contrast and choppy appearance, which makes the patterns difficult to read on the copies of this report provided by NTIS.

Despite these shortcomings, those of you with the real estate for a large antenna and who need wideband antenna performance will be able to design your own rhombic without need for information not contained in this book.

Beverage Antennas For HF Communications, Direction Finding and Over-The-Horizon Radar, by J. Litva and B. J. Rook, Communications Research Center (Canada), 1976, 188 pp. NTIS order number DAD013787/5G1. Paper cost, $19.95; microform, $6.95.

This very complete research report covers a wide variety of questions regarding Beverage antennas. The report not only considers theoretical matters, but gives a considerable body of data of actual experiments carried out over a five-year period. Of special interest is the fact that to obtain the experimental data, radiation patterns were measured for full-size HF Beverages using aircraft and tethered balloons. This makes possible the comparison of calculated to real results.

The major portion of the report deals with arrays of Beverages to obtain particular characteristics. For example, the use of a linear array of Beverages to increase efficiency is discussed in detail. Considered are the spacing of multi-Beverage arrays for increased efficiency and their use as transmitting antennas. Also, the linear array of Beverages is compared to some other antennas for signal-to-noise performance. Rosette arrays for direction finding are covered as well.

The report is rounded out with a Fortran program listing that allows calculation of critical performance characteristics of a given Beverage antenna system. For those of you wanting to expand your existing Beverage antennas or trying to decide what antenna to put up for that low-band station, this report has a wealth of information.

On Increasing The Radiation Efficiency of Beverage-Type Antennas, by E. K. Miller, R. J. Lytle, D. L. Lager and E. F. Lane, Lawrence Livermore Laboratories (USA), 1977, 22 pp. NTIS order number UCRL52300. Paper cost, $9.95; microform, $6.95.

This paper presents an extensive discussion of methods of modeling the real performance of Beverage and the methods of increasing antenna efficiency. The conflicts in the various mathematical modeling techniques are discussed, with the authors pointing out those that come closest to describing real performance. Also presented is the increase in efficiency of two-wire Beverages at various wire spacings. The authors give a great deal of attention to the critical matter of current attenuation over the length of the Beverage antenna.

Even though heavily concerned with theoretical matters, this paper serves up a few important practical points that I have not seen in other publications. *—Dom Mallozzi, N1DM, 26 Carey Ave, Apt 8, Watertown, MA 02172*

Bits

Courses on Mechanical Behavior and Service Life of Plastics

L. J. Broutman and Associates and the Society of Plastics Engineers (SPE) are cosponsoring three short courses for plastics designers, engineers, chemists and R&D staff, and managers in plastics design, construction and quality control. The courses are designed to demonstrate some of the more important aspects of the mechanical behavior of plastics. The first course is a two-day session (February 13 and 14, 1989) entitled "Mechanical Behavior of Plastics." The second course will be held February 15 and 16, 1989, and is called "Service Life Predictions for Plastic Parts." The third course is "Impact Behavior and Testing of Polymers and Composites." All the courses will be held at the Hyatt Regency O'Hare Hotel in Chicago, Illinois. For more information, contact Michael R. Roop, Director of Educational Services, L. J. Broutman & Associates, 3424 S State St, Chicago, IL 60616, tel 312-842-4100.*—Rus Healy, NJ2L*

>50
Focus on technology above 50 MHz

By Bill Olson, W3HQT
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Reflector Antennas

In the last column, I discussed electromagnetic waves and antenna polarization as an introduction to a series on antenna fundamentals. The subject of antennas is very complex and thousands of volumes have been written on the subject. (Makes a two-page article in *QEX* seem a bit inadequate!) Many of the comments I've received from readers indicate that approaching some of these complex subjects in plain language and "keeping it simple" have been useful. In the same vein, I'll keep this column simple—with apologies to those of you who keep Kraus' *Antennas* on your nightstands![1] This column is for the reader who asks "I bought this four-foot dish at a flea market for four bucks—is it any good for anything?" I'll discuss reflector antennas—mainly parabolic reflectors—and I will try to answer a few of the most-often-asked questions.

Parabolic dishes are not the only type of reflector antenna, but parabolics are certainly the most prevalent. Basically, any radiator that increases gain by the use a reflecting surface of some shape can be thought of as a reflector antenna. Some of the more common types are the corner reflector, the backfire and short-backfire antennas, the spherical reflector and various parabolic designs (such as the parabola and the cylindrical parabola). Secondary reflector antennas such as the Cassegrain and Gregorian reflector systems are also forms of reflector antennas. I guess even a dipole over a plane reflector fits the above definition, though this configuration is most often used as a feed antenna for a dish or as a reference gain antenna. Remember that a reflector antenna has *two* parts: (1) the reflector and (2) the feed antenna. Their functions are quite distinct—the feed illuminates the reflector, and the reflector focuses the energy. A parabolic reflector without a feed is just a birdbath, *not* an antenna! In addition, the feed antenna must be matched to the reflector, as we shall see.

Most reflector antennas, with the exceptions of the parabolic and corner reflectors, are used very little in amateur work. The corner reflector is a dipole radiator "standing in the corner," so to speak, of a bi-planar reflector (Fig 1). Advantages of the corner reflector are relatively high gain with simple, noncritical construction. The corner reflector can be either vertically or horizontally polarized, and the sides can be made from solid sheet or screen. The geometry of the corner reflector provides about 10 dB gain over a dipole, and *no more*, no matter how big we make the sides of the reflector. The parabolic dish, on the other hand, has gain that is limited only by the reflector's size.

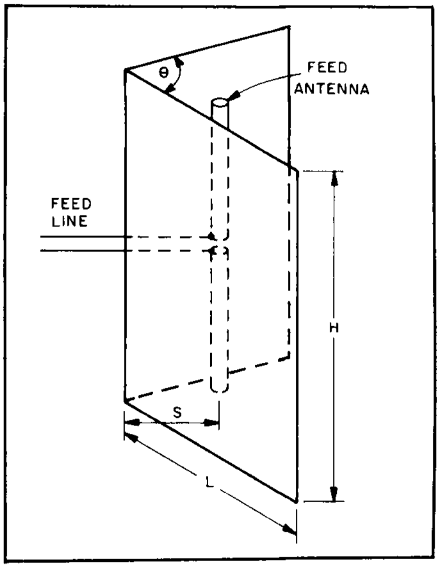

Fig 1—A corner-reflector antenna. $L \geq 1\ \lambda$, $H \geq \frac{5}{8}\ \lambda$, $S = 0.5\ \lambda$, and $\Theta = 60°$. Gain will be between 8 and 10 dBd when a ½-λ dipole is used to illuminate the reflector.

The Parabolic-Dish Antenna

Just about everyone on Earth is familiar with parabolic-dish antennas. If you live in the city, you see dishes on tops of buildings, towers, and so on. If you live in the country, you can see dishes in almost every yard. (Even if you don't live on Earth, there're plenty of these things out there too!)

A perfect parabolic reflector would convert a spherical wave emanating from a point source at the focus of the dish into a plane wave. That is, if a source radiating in all directions is placed at the focal point of a parabolic dish, all the energy that hits the dish should be reflected—all in the same direction, and in phase. A dish's gain depends upon how large the dish is, how much of the feed energy hits the dish, and other factors.

Dish Geometry

A parabolic-dish antenna is a paraboloid, which is a three-dimensional solid generated by rotating a parabola about a line joining its origin and focus (see Fig 2). The mathematical equation for a parabola is

$$Y^2 = 4fx \qquad \text{(Eq 1)}$$

where

f = focal length
x = depth of the dish at any distance from the origin (Y).

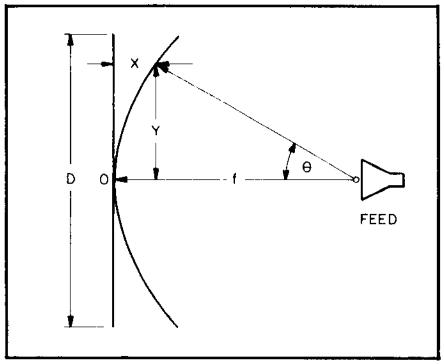

Fig 2—Cross-sectional view of a parabolic reflector. See text.

For any given dish of diameter D and depth, c, the focal length can be found by substituting ½D for y and c for x. This yields the equation: $(\tfrac{1}{2}D)^2 = 4fc$. Solving for f,

$$f = \frac{D^2}{16C} \qquad \text{(Eq 2)}$$

Hence, if a reflector is known to be a paraboloid, the focal length can be found by plugging the dish depth and the diameter into Eq 2. The f/D ratio can also be determined after you know f. The f/D ratio of a reflector is important, because it affects the design of the feed antenna, as we shall see. Once a feed is optimized for a reflector with a particular f/D ratio,

[1]Notes appear on page 16.

however, it will be right for any dish with the same f/D ratio, regardless of dish diameter. Also remember that the focal point of a dish is a *fixed point*, determined *only* by the geometry of the dish.

Dish Gain

One of the primary parameters of concern regarding any antenna is, of course, gain. We all know that the bigger the antenna, the more gain (assuming, of course, an optimally designed antenna). With dish antennas, this statement holds true, and gain is directly proportional to the projected area of the dish (proportional to the square of the diameter). The equation for gain is:

$$G = \eta \frac{4\pi A}{\lambda^2} \quad \text{(Eq 3)}$$

where

A = area
λ = wavelength
η = feed efficiency of the antenna (always less than one, and typically 0.45 to 0.55 for amateur antennas).

Dish efficiency is always less than 100%, because of difficulty in fully illuminating the reflector, resistive losses, blockage by the feed antenna, surface inaccuracies, and so on.

The bigger the dish, the greater the gain. As the reflector gets larger, the feed gets smaller, proportionally, and more closely approximates a point source. The energy reflected from the dish tends more and more to be in phase, planar, and with less dispersion. What is *not* so obvious is how different dish geometries and feed configurations affect antenna efficiency and gain.

Dish efficiency is affected by many factors, but the most important has to do with optimum illumination of the reflector. An ideal feed would be a point source radiating energy uniformly in a cone-shaped pattern—fully illuminating the dish, but nothing else. *All* the energy from the feed would then be reflected as a plane wave from the dish. In real life, because no feed antenna has a sharp cutoff, there is some spillover. That is, some of the energy from the main lobe (as well as some sidelobe energy) from the feed antenna misses the reflector and is lost. If the feed antenna has too wide a beamwidth, there is excessive spillover, and if the beamwidth is too narrow, a sub-optimal amount of the dish is illuminated. In both cases, gain is less than optimum. The optimum feed antenna has a clean pattern with similar E- and H-plane patterns, and the power density at the edge of the dish is 10 to 12 dB below the density at the center. (Historically, engineers have used 10 dB as a rule of thumb for reduced sidelobes. We sometimes see numbers up to 15 or even 20 dB.) See Fig 3. For edge-illumination power densities greater than −10 dB (−8 dB, for instance) gain is decreased and sidelobe energy goes up. For power densities less than −10 dB, gain also decreases, but the pattern is cleaner. For example: for an edge-illumination density of −20 dB, the first sidelobe is at least 40 dB weaker than the main lobe of the dish, while for an edge-illumination density of −6 dB, the first sidelobe is less than 20 dB weaker than the main lobe.

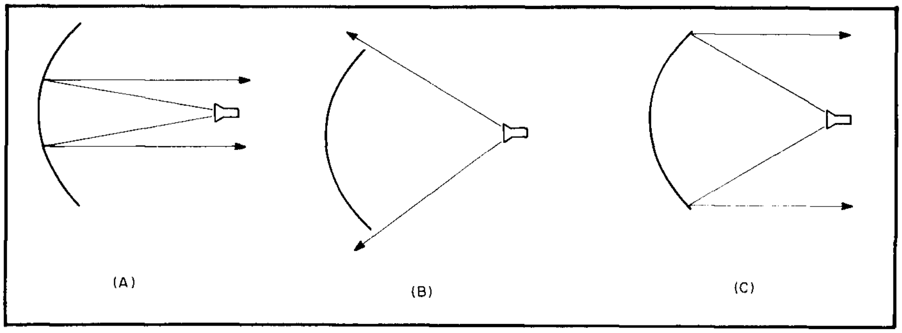

Fig 3—Illumination of a parabolic reflector is critical to antenna performance. An under-illuminated dish is shown at A (feed-antenna beamwidth too narrow), and excessive feed-antenna-energy spillover is shown at B (feed antenna beamwidth too wide). Proper dish illumination is shown at C.

How does f/D ratio figure into the efficiency and edge-illumination picture? Dishes with high f/D ratios are relatively flat, and the feed is relatively far from the reflector. The feed antenna must have a fairly narrow beamwidth for optimum edge illumination. Dishes with low f/D ratios (sometimes known as spaghetti bowls) are more difficult to feed. For example, a dish with an f/D ratio of 0.25 (known as a focal-plane dish, because the focus is in the same plane as the rim of the dish) has a subtended angle of 180°. Optimum illumination for such a dish would occur with a feed antenna that had a −10-dB beamwidth of 180°—a difficult feed to design.[2] Feed antennas on dishes are easy to support. Because of the short distance from the dish to the feed, high f/D dishes require more extensive feed-support structures. Dishes with low f/D ratios are quieter on receive than other dishes, because of the low spillover and cleaner pattern. So, once again, there is no simple answer as to which antenna is the best. Trade-offs, trade-offs . . .

Conclusions

Where do we use a dish and where do we use a Yagi? Table 1 shows the gains of different-diameter dishes at various amateur frequencies, assuming 50% illumination efficiency. As you can see, you probably wouldn't want to use a four-foot dish at 432 MHz, because a short Yagi has the same gain, and a *lot* less wind area. On the other hand, the same dish would have a gain of nearly 40 dB at 10 GHz! The advantages of dishes are high gain at the higher frequencies and multiband capability with a single reflector. Disadvantages are high wind area and weight for larger dishes. For instance, you probably wouldn't want to put a 16-foot dish on top of your tower for 432-MHz tropo when four long Yagis would do the same job!

Next month, I'll discuss some practical feeds for illuminating various dish types, and I'll offer some tips for obtaining or building your own dish.

[1]J. D. Kraus, *Antennas*, 2nd edition (New York: McGraw-Hill Book Co, Inc, 1988).
[2]Actually, the required beamwidth at the −10-dB points is more than 180°. This is a result of additional wave attenuation in space for the signals traveling to the edges of the dish—the edges are farther from the feed than the center. This additional path loss is called *space attenuation*, and its effect is most pronounced with low-f/D-ratio dishes.

Table 1—Gain of Parabolic Reflectors, Assuming 50% Feed Efficiency

	Gain (dBi)					
Diameter (feet)	*432 MHz*	*1296 MHz*	*2304 MHz*	*3456 MHz*	*5760 MHz*	*10.368 GHz*
1	—	9	14	18	22	27
2	6	15	20	24	28	33
4	12	21	26	30	34	39
8	18	27	32	36	40	45
16	24	33	38	42	46	51
32	30	39	44	48	52	56

Components

By Mark Forbes, KC9C
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Rocklin, CA 95677

It's Christmastime already! I hope all of you get all the ham radio gear and projects that you told Santa about. If you haven't made up your mind about what you want, here are some last-minute suggestions.

Inductor Sources

Inductors are among the more difficult items for circuit tinkerers to find. Radio Shack® carries inductors of some popular values, but we're often stuck for other values. I've pulled together information that should give you a place to begin your search for that odd-value inductor or transformer.

J. W. Miller Inductors

J. W. Miller, a name virtually synonymous with inductors, has a new inductor catalog available. The catalog covers Miller's entire inductor line, and is a must for most RF experimenters. The catalog covers everything from tiny molded chokes through large line-filter chokes capable of carrying 20 A.

To get a copy of the catalog, write to J. W. Miller, 19070 Reyes Ave, Rancho Dominguez, CA 90221. Request their *Inductors For Electronics* catalog.

Coilcraft Surface-Mount Inductors

Coilcraft has a wide selection of surface-mount inductors and transformers. They offer two kits for designers that could be quite handy to hams. The first kit is a set of fixed, surface-mount inductors. Included in the kit are six each of 64 different inductance values in the range 0.004 μH to 1 mH. This is known as Kit C100 and sells for $125.

The second Coilcraft kit for designers is a set of adjustable inductors in the range 0.1 μH to 10 μH. Kit C101 contains 45 inductors comprising 11 different values, and sells for $50.

If you like to build RF circuits, or are planning to experiment with RF design, either of these kits would be a good way to ensure that you have the right inductor on hand, and to allow you to empirically try other values and observe their effects. You can order either kit from Coilcraft, 1102 Silver Lk Rd, Cary, IL 60013, tel 312-639-6400. You might want to take a look at their catalog before you place an order. The catalog contains several useful application notes in addition to the technical data on the components.

IBM® PS/2® Model 30 286

"Why is a personal computer being talked about in a Components column?" you may ask. Just as the perception of microprocessors has evolved from "the center of the universe" to a building block component, PCs themselves have become components in many systems. Packet-radio systems are an example of this: With the availability of commercial TNCs, very few people actually assemble ICs, resistors, and so forth as components in a packet station. Instead, the components of a packet radio system are the antenna, transceiver, TNC, and computer.

The PS/2 family has been somewhat maligned in the press, and I, too, haven't been overly impressed with the line's price-v-performance aspects. The introduction of the IBM PS/2 Model 30 286, however, changes that.

The PS/2 Model 30 286 marries the AT bus to the OS/2™ operating system. Thus, it's an excellent vehicle for running programs today, and can be upgraded to OS/2 when its user requires this. (Although the 30 286 can run OS/2, it is not a Micro Channel Architecture machine.)

At $1995, the IBM Model 30 286 is also economically priced. If you are outfitting your station with a computer for the first time, or are contemplating a system upgrade, the IBM PS/2 Model 30 286 is definitely worth considering.

Inresco Circuit Saver

This rather innovative product is promoted by Inresco as "destined to replace fuses and circuit breakers on your PC board." Circuit Savers protect circuitry by opening in response to sense excessive current. When the overcurrent condition ends, they automatically reset, eliminating manual replacement or resetting. The trip time of a Circuit Saver is in the 100-μs range; total isolation is achieved in about 300 μs.

Note that these devices are intended for replacement of fuses and circuit breakers, but not for voltage-surge protectors. Circuit Savers sense current rather than voltage, and are much too slow to be used for overvoltage protection.

At ratings from 50 mA to 1 A, Circuit Savers sell for around $5 each in small quantities. That's certainly comparable to a circuit breaker—and the Circuit Saver need not be reset manually. Each Circuit Saver is good for about 100 trips. For more on this product, write Inresco, 645 Ocean Rd, Pt Pleasant, NJ 08742; tel 201-892-5881.

NEC Microwave Prescaler Family

NEC RF and Microwave Semiconductor Products has announced three prescalers in the microwave range. The UPB584B/G and the UPB585B/G both operate over a range of 500 MHz to 2.5 GHz. Conveniently, they operate from a 5-V power supply. The B configuration is an 8-lead ceramic flat package, and the G configuration is an 8-pin miniDIP.

The UP584 is a divide-by-2 prescaler, and the 585 is a divide-by-4 device. Both parts can be used in frequency synthesisizers. Because these parts can operate at 1.2 GHz, they could serve as the basis for a nice project to acquaint Novices with their 1.270- to 1.295-GHz band.

Also included in the family is the UPB587B/G. This device is a low-power, divide-by-2, 4, or 8 prescaler. Typical supply voltage is 3.

Additional information can be obtained from NEC's exclusive agent: California Eastern Laboratories, 3260 Jay St, Santa Clara, CA 95054.

Polycore RF Devices

Polycore RF Devices has just released a new catalog covering their line of power FETs and MMICs. Polycore power FETs cover frequencies from HF through 2 GHz and can produce output powers up to 300 W. Most of the devices are proprietary designs, but a few are second-sources for Motorola and Philips products.

Polycore's F3000-series Superpower FETs for 50-V operation have just been released. The output power of these devices is as high as 300 W—a level at which their 50-V rating makes power supply requirements easier to satisfy. Minimum gain of F3000-series FETs is 13 to 16 dB; the line covers 1 to 300 MHz.

You can obtain Polycore's short-form catalog by writing Polycore RF Devices, 1107 Tourmaline Dr, Newbury Park, CA 91320.

Micro Crystal Division

Crystals are available in a variety of cuts from Micro Crystal; the line includes surface-mount and leaded-package styles. The surface-mount crystals, which may be very useful for UHF and microwave projects, available at fundamental frequencies from 8 to 35 MHz. At less than $10 each, they are an interesting alternative to standard through-hole-mounted crystals.

Micro Crystal Division is located at 35 E 21st St, New York, NY 10010.

Reader Response

I'm always interested in hearing from readers. If you have a particular area of interest that you'd like to hear more about, let me know. I'd also like everyone to drop me a QSL or postcard for a mini-survey: Just jot down what types of project you like to build and what your most recent project was. Mail your card to me at my byline address. 73!