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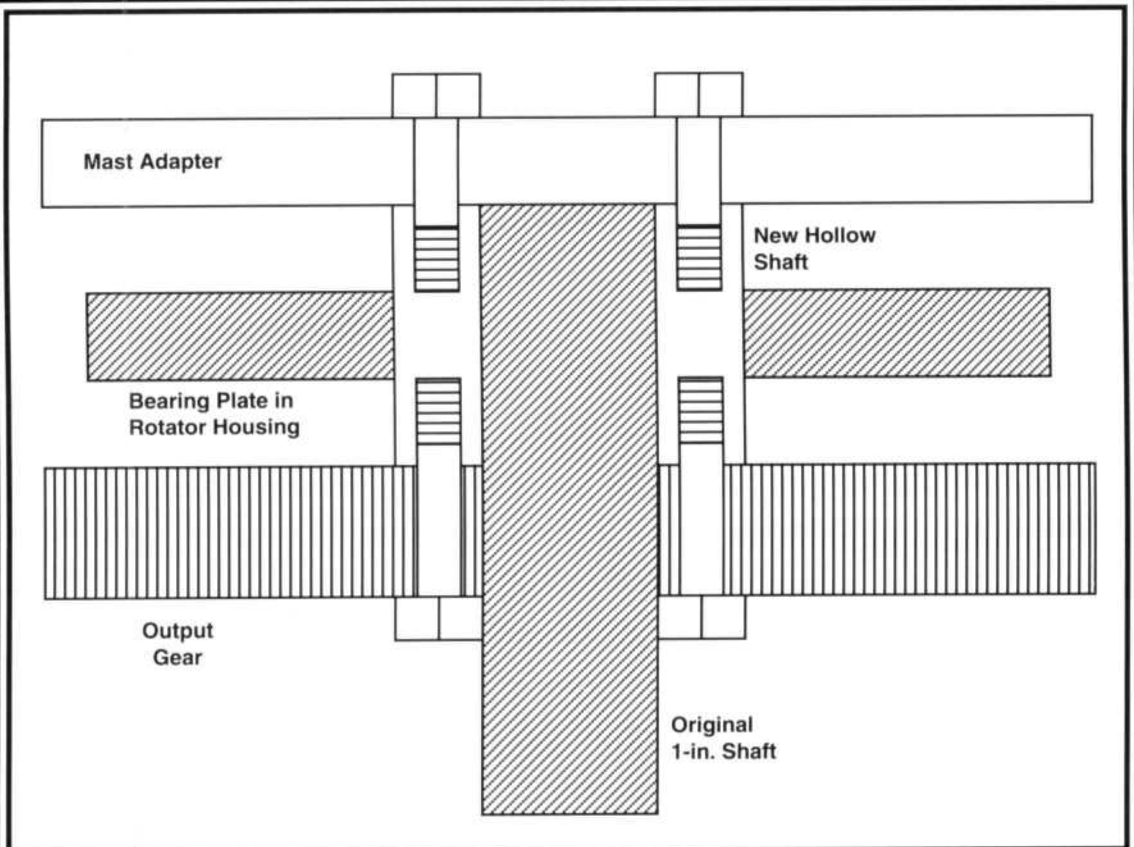
ARRL Experimenter's Exchange

June 1995



Making Better Rotators

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QEX

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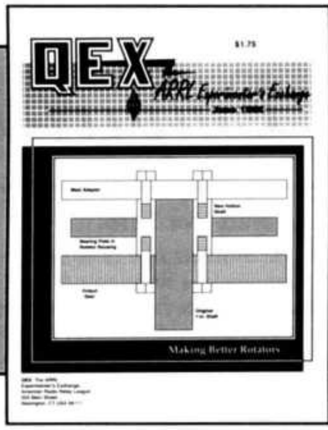

About the Cover

W6QHS has some ideas for improving existing rotators under heavy loads.

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Features

Columns

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THE AMERICAN RADIO RELAY LEAGUE

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

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

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

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

Empirically Speaking

Getcha Red Hot Info!

Several months ago (February 1995) in this space, we suggested that amateur experimenters should arrange to access the Internet in order to keep up with what's going on in the world of amateur experimentation. This month, we want to tell you about some particular examples that show the power and the pervasiveness of the Internet—the World Wide Web in particular—as a medium for this kind of information exchange.

In the June 1994 issue, we reprinted a paper presented by Phil Karn, KA9Q, at the 1994 TAPR Annual Meeting, in which Phil showed how forward error correction (FEC) could be used to improve the performance of digital communication links. Now, Phil has established a Web page that demonstrates these principles by the use of audio files and explanatory text. This page (http://www. qualcomm .compeople/pkarn/fecdemo/index.html) contains several 64-kbit/s μ-law PCM audio files (.au file format). You'll need compatible audio player software (and a sound card, of course); Netscape V1.1 for Windows has a built-in player that will do the job. This page demonstrates not only Phil's concepts, but also how the Web can be used to foster leading-edge amateur technology.

Another example is reported in this month's "Digital Communications" column, in which Steve Bible, N7HPR, gives an overview of spread spectrum as it applies to Amateur Radio. Steve has established several Web pages—with links to yet other pages—that provide information about amateur spread spectrum. The advantage of the Web for this kind of overview presentation is that as new information becomes available the Web pages can be immediately updated. It need no longer take days, weeks or months to get information out to the folks working in a particular field.

Web pages can also address those pesky frequently asked questions. The GRAPES 56-kbit/s modem designed by WA4DSY is a perennial favorite. As newcomers to packet develop a thirst for higher speeds, they often want to know about this high-speed project. They can learn about it on the Web (http://www. mindspring.com/~bobm/ grapes/grapes.html).

The Web isn't the only mechanism on the Internet for disseminating information. The Web is ideal for "publishing" text, graphics and binary data, but it's not optimal for holding discussions. This kind of activity is better served by Usenet newsgroups and specialized mail reflectors. In the latter category is ARRLCAD, an e-mail reflector (mailing list) through which you can now share your questions, answers, ideas and views about *ARRL Radio Designer* and other amateur-radio-related circuit and antenna design and simulation tools with other users. To subscribe to ARRLCAD, send an email message to:

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with text that reads:

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in which *FirstName* and *LastName* mean exactly that. The reflector software will confirm your subscription with a informational welcome message.

No doubt we'll see more and more of this sort of thing. 'Net-based information services are growing faster than anyone can keep track, but if you see something on the Internet that you think other *QEX* readers would find of interest, drop us a line (by email, natch!) and we'll report it here.

This Month in *QEX*

Antenna rotators used by amateurs have a disturbing tendency to fail when large arrays are used under high-wind conditions. Why? As D. B. Leeson, W6QHS, has found, it's often because of limitations in the "Torque Capacity of Keyed Rotator Shafts."

The venerable R-C phase-shift network used to produce quadrature audio signals for SSB generation has been described in *QEX*, and elsewhere, before. But Tetsuo Yoshida, JA1KO, takes a new tack in "Polyphase Network Calculation using a Vector Analysis Method," which leads to an improved design approach.

Crystal ladder filters are the design of choice for home builders, mostly because the crystal requirements are less stringent than those of lattice designs. But making SSB-bandwidth filters has been problematic with existing design techniques. Wes Hayward, W7ZOI, attacks this problem with "Refinements in Crystal Ladder Filter Design." The result is an approach that allows easy design and construction of wide-bandwidth filters.

Finally, Harold Price, NK6K, gives his "Digital Communications" column over to Steve Bible, N7HPR, who provides a good overview of spread spectrum and its application to Amateur Radio, along with pointers to more detailed information.—*KE3Z, email: jbloom@arrl.org (Internet)*

Torque Capacity of Keyed Rotator Shafts

Just how much array can that rotator handle? It depends in part on the shaft's torque handling ability.

By D. B. Leeson, W6QHS

Antenna rotator failures are an all too common disappointment, especially when large arrays are rotated in windy conditions. Even heavy-duty rotators such as the Hy-Gain HDR300 and Orion 2300/2800, which advertise a braking torque of 7500 in-lb or greater, seem prone to suffer failure of the keyed output shaft. Efforts to use multiple keys with setscrews, as well as applying Loctite Stud and Bearing Lock to the output gear and mast adapter plate, have resulted only in minor delays in the onset of excessive play in the rotators. The problem is reduced but not eliminated by use of antenna wind balance and resilient shock isolators[1].

A letter from the author of an earlier article on big rotators, Victor Mozarowski, VE3AIA, alerted me to the possibility that a small-diameter keyed shaft may not be able to handle the torque specified by the rotator manufacturer.[2] Since the design of a keyed shaft is a well-documented engineering area, I confirmed this by some simple calculations.

My conclusion is that rotators rated at or above 5000 in-lb of torque must have output shafts in the range of 1.5- to 2-inch diameter, with appropriately sized keys, depending on the choice of steel alloy and tempering. I believe that a 1-inch keyed shaft cannot reliably provide the advertised braking torque capacity (7500 to 17000 in-lb) with oscillating antenna loads.

Shaft Torsion Limits

The standard reference for mechanical questions is *Machinery's Handbook*, which has sections devoted to shaft design and key design.[3,4] Shaft torque limits are also covered in mechanical engineering texts.[5]

The formula for the torque capacity of a cylindrical shaft, ignoring stress concentrations and strength reduction due to keyways, is:[6]

$$T_{max} = f_s J / R,$$

where f_s is the maximum allowable torsional shearing stress in lb/in^2, J is the polar moment of inertia of the shaft

[1]Notes appear on page 8.

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cross section, and $R = D/2$ is the radius of the shaft. For solid shafts:

$J = \pi R^4/2 = \pi D^4/32$, where D is the diameter of the shaft, while for hollow shafts:

$J = \pi(D^4 - d^4)/32 = \pi D^4[1 - (d/D)^4]/32$

where D is the outer diameter and d the inner diameter.

For a solid shaft:

$J/R = \pi D^4/32 \div 2/D = \pi D^3/16$

while for a hollow shaft:

$J/R = \pi D^3[1 - (d/D)^4]/16$

The factor $[1 - (d/D)^4]$ is denoted $1/B^3$ so that:

$J/R = (\pi/16)(D/B)^3$

where B is defined as a factor depending on the ratio of inner to outer diameter ($B = 1$, $d = 0$ for solid shaft), given by the formula:

$$B = \sqrt[3]{1/[1-(d/D)^4]}$$

The factor B is also tabulated in *Machinery's Handbook*.

Thus, for the case of a cylindrical shaft the maximum operating torque is given by:

$$T_{max} = \frac{\pi fs(D/B)^3}{16}$$

or if we know T_{max} and want to determine the required D:

$$D = B\sqrt[3]{\frac{16T_{max}}{\pi f_s}}$$

The factor $16/\pi = 5.1$.

The value of f_s, the maximum allowable torsional shearing stress, ranges from 8000 lb/in^2 for "commercial" steel up to much higher values, as much as 40000 lb/in^2, for specially hardened steel alloys. The recommended value of f_s in pure torsion is 30% of the elastic limit for the material in tension but not more than 18% of its ultimate tensile strength.

Keyways and Shock Loading

In my experience, rotator output shafts commonly fail primarily by deformation of the material around the key, with the result that excessive play develops. This ultimately leads to failure of the key or keyway.

The effects of keyways, shock loading and stress risers are treated by additional factors. For a shaft with a standard keyway (width one-quarter and depth one-eighth of the shaft diameter), the *Handbook* recommendation is that shafts with keyways be designed for 75% of the working stress recommended for a solid shaft of the same diameter and material.

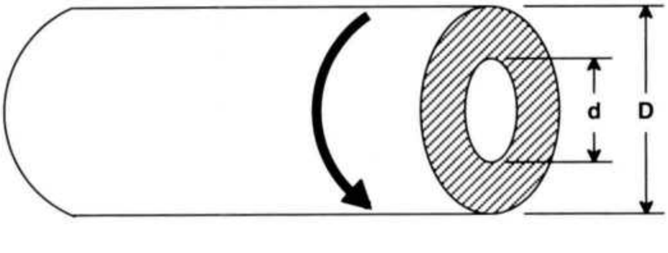

Fig 1—Radii of a shaft.

Combined shock and fatigue factors for rotating shafts range from 1 for a steady load applied gradually up to a range of 1.5 to 3 for loading suddenly applied with heavy shocks.[7]

Stress concentrations arising from steps in shaft diameter are accounted for by a torsional stress-concentration factor that ranges from unity for gradual changes in diameter to as much as a factor of 2 or 3 for abrupt steps in shaft diameter.[8] If the shaft is subject to bending as well as torsion, the problem is further complicated, but this case does not apply to common rotator configurations. All these factors reduce the torsion capacity of a shaft.

The formula for the maximum allowable torque for a keyed shaft with shock loading is:

$$T_{max} = \frac{K_k f_s}{5.1K_t}(D/B)^3$$

where K_k is the 0.75 factor to account for the keyway, D, B and f_s are defined above, and K_t is the shock loading factor. This can be rearranged to calculate the shaft diameter required for a torsion capacity T_{max}:

$$D = B\sqrt[3]{\frac{5.1K_t T_{max}}{K_k f_s}}$$

For a keyed shaft, the variables are the shaft diameter and the material. Table 1 shows that for ordinary steel a 1-inch keyed shaft is limited to 471 in-lb and that a 2.18-inch diameter and a ½-inch key are required to provide 5000 in-lb capacity.

For special steel, f_s=0.3 times the elastic limit in tension, but not more than 0.18 times the ultimate tensile

Table 1—Keyed Shaft Torque Limits (Without Shock Isolator, K_t = 2.5)

	"common" steel		*1025*	*17-4*	*4130*
D, inch	1.00	2.18	1.84	1.33	1.28
Key size	¼	½	½	5/16	5/16
B	1	1	1	1	1
K_t	2.5	2.5	2.5	2.5	2.5
f_s	8000	8000	13500	36000	40000
75% of f_s	6000	6000	10125	27000	30000
T_{max}, in-lb	471	5000	5000	5000	5000

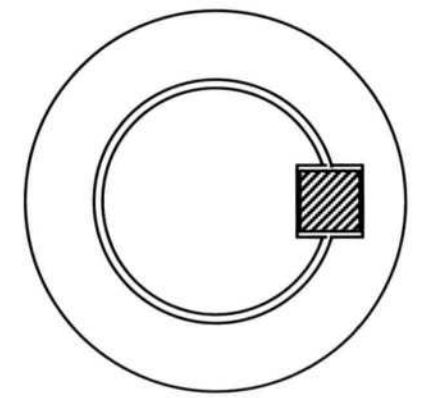

Fig 2—End view of a keyed shaft.

strength, usually given in thousands of lb/in^2 (ksi). As an example, for 1025 steel f_s = 13.5 ksi, the smaller of 0.3 x 45 = 13.5 ksi and 0.18 x 90 = 16.2 ksi.

If hardened 1025 steel is used, the diameter can be 1.84 inch; if hardened 4130 axle steel is used, a 1.28-inch diameter shaft and a 5/16-inch key will provide a 5000 in-lb capacity. An excellent shaft material is 17-4 precipitation hardening (PH) stainless, which when hardened at 900° (H 900) provides 200 ksi ultimate strength and 185 ksi yield strength, and hence a calculated f_s = 36 ksi (36000 lb/in^2). Under high shock loading a keyed 17-4 H900 shaft, as used in the new Hy-Gain HDR300A,while definitely a big step in the right direction, must be 1.33-inch diameter for 5000 in-lb when designed to the above guide lines. With a high-strength shaft, the gear, key and mast adapter may become problems.

If we consider the case of a rotator fitted with a effective torsion shock isolator, we can reduce the shock factor, perhaps to 1.5 (load suddenly applied, minor shocks only). Table 2 shows the results for this case.

With a shock loading factor of K_t=1.5, we can carry 5000 in-lb with a 1.55-inch 1025 shaft using a 3/8-inch key. Much higher tensile steel, such as hardened 4130 chrome-moly, allows the use of a 1.08-inch shaft with an effective shock isolator (lower K_t) for 5000 in-lb. But this requires a 1.250-inch diameter to accept even 7500 in-lb of braking torque. A hardened 17-4 stainless shaft with isolator requires a calculated diameter of 1.12 inch for 5000 in-lb of torque.

This all suggests that a 5000 in-lb rotator (whether rotating or braking torque) application is likely to be beyond the capability of a keyed 1-inch shaft made of anything other than the best hardened axle steel, and then only with an effective shock isolator. A 2-inch steel shaft seems a better choice to provide reliable operation at 5000 in-lb of torque and above.

Capability of 2-Inch Masts

A related question is the torque capability of the 2-inch masting commonly used in antenna installations, as shown in Table 3.

Table 2—Keyed Shaft Torque Limits (With Shock Isolator, K_t = 1.5)

	"common" steel		*1025*	*17-4*	*4130*
D, inch	1.00	1.00	1.55	1.12	1.08
Key size	1/4	1/4	3/8	1/4	1/4
B	1	1	1	1	1
K_t	1.5	1.5	1.5	1.5	1.5
f_s	8000	13500	13500	36000	40000
75% of f_s	6000	10125	10125	27000	30000
T_{max}, in-lb	785	1320	5000	5000	5000

Table 3—2-Inch Mast Torque Limits

Mast torque	*"common" steel*	*1025*
D, inch	2	2
Wall, inch	3/16	1/4
d, inch	1.624	1.5
B	1.21	1.13
K_t	2.5	2.5
f_s	8000	13500
T_{max}, in-lb	2854	5813

The typical mast can handle the 5000 in-lb we would hope it could, but it cannot handle much more under high shock load without exceeding its strength limits. However, I am is not aware of any mast failing in torsion in typical installations. A 4130 chrome-moly mast can carry substantially more torque and bending moment, as can a 1025 3-inch mast.

Strength of Bolted Flanges

A bolted flange can be a stronger alternative to a keyed shaft. As compared to the case of a shaft in torsion, all of the bolt material resists the tendency of the flanges to shear off the bolts, so the shear strength is substantially higher in this case. Torque isolator assemblies are similar to bolted flanges and are limited by the shear capacity of the bolts themselves. The allowable torque for a flange with bolts in pure shear is given by the formula:

$$T_{max} = \frac{NAf_sR}{K_t}$$

where N is the number of bolts, A is the cross section area of the bolt, f_s is the shear strength of the material, as above, and R is the radius to the center of the bolts. Shock loading can be accounted for by a factor K_t, as above. The additional torque capacity due to the friction between the flange plates is generally not claimed in conservative designs.

If the bolt is also subject to bending, as in the case of holding a resilient isolating donut, the maximum allowable shear stress p_t under combined shear and bending is less than f_s (typically 75%), and the total force is the vector sum of the shear and bending forces.

If the force F on the bolt acts at a distance h from the flange, and the radius of action is R, the total moment on the bolt is:

$$M = \sqrt{(FR)^2 + (Fh)^2} = F\sqrt{R^2 + h^2}$$

so the formula for T_{max} becomes:

$$T_{max} = \frac{NAp_tR}{K_t\sqrt{1 + (h/R)^2}}$$

Table 4 shows the torque capacity of a torque isolator flange using 3/8-inch bolts with 1.5-inch radius holding an isolator disk at a mean height of 0.75 inch, or with spacers, 1.5 inch above the flange, for various steels.

This shows that typical hardened bolts should be able to provide 5000 (but not the advertised 17000) in-lb so long as the bending moment is not excessive. I use a torque isolator design with bolt-head clearance holes in the end

Table 4—Torque Isolator Flange Torque Limits

	"common" steel	*Grade 5 + spacer*	*Grade 5*
Radius, inch	1.5	1.5	1.5
Height (mean)	0.75	1.5	0.75
Bolt diameter	0.375	0.375	0.375
Area	0.08*	0.08*	0.08*
f_s	8000	75000	75000
p_t	6000	56250	56250
Force/bolt	464	4350	4350
N bolts	4	4	4
K_t	2.5	2.5	2.5
T_{max}, in-lb	996	7380	9330

* Area at thread root diameter

plates to eliminate spacers under the donut so that the bolts are as short as possible, and I use Grade 8 bolts held in place with Loctite thread locking compound. To avoid fatigue failure of the bolt threads, it is also important to mount the bolts in counterbored holes (see Fig 3) so the shear force is carried by the full shank diameter rather than the threads of the bolt.

Strength of Bolted Rotator Attachment

We can estimate the ultimate torsion strength of a rotator mounted on a tower plate. For the standard Ham-M bolt pattern, the radius is 2.12 inch and the bolt diameter is 5/16 inch. Because the bolts are in shear across their threaded area, the bolt capability must be derated to the area at the thread root diameter, which is 0.0522 in^2 rather than the full-diameter area of 0.0767 in^2. The bolt shear strength is 3900 lb, so without any derating for fatigue or shock the torsion capability of four bolts at 2.12-inch radius is approximately 33000 in-lb. If the rotator is mounted on a standard Rohn 45 plate (measured thickness is 0.11 inch), the bearing strength of each hole is 3450 lb, so the total bearing strength for four holes is approximately 29000 in-lb. Derating these numbers by a shock and fatigue factor of 2 to 3 would represent a more realistic estimate.

Typical aircraft and automotive practice places the shear forces applied to a bolt or stud on the unthreaded portion of the bolt shank, not on the threaded portion. The threaded portion has a substantially reduced diameter, the root diameter of the threads, as well as stress risers due to the thread shape and manufacture. Fatigue life is greatly extended for bolts with rolled, rather than cut, threads; these are found on AN and Grade 8 bolts and studs. The drawing of Fig 3 shows how to pilot the holes on the bolt shank rather than the threads.

Unfortunately, antenna rotator practice is to place the shear stress directly on the bolt threads. A better design would use a bolt with an unthreaded shank carried through the plate and 1/4 inch into the non-ferrous rotator housing, so that the shear force could be held by the unthreaded bolt shank in both materials. This would require the threads to be undercut in the rotator housing, but would be much stronger in shear.

A typical 5/16-inch hardened steel bolt can withstand a shear force of 5750 lb at full diameter or 3900 lb at the thread root diameter. Table 5 summarizes the above calculations, including the effect of shock loading. The bearing forces of the 0.11-inch steel plate and 1/4-inch aluminum housing are given in Table 6. From this we can conclude that the bolts holding a rotator to a steel plate as used in Rohn towers will in fact provide substantial safety margin at 5000 in-lb, and that the mounting plate is thick enough to provide the desired bearing force. An aluminum rotator housing would seem to provide substantial margin, but in practice the bolts tend to work loose in the tapped holes because of the effect of oscillating loads and the lower bearing strength of a casting, even with the larger bolt pattern used in the HDR300. I've experienced cases in which the bolts became loose and the rotator threads required repair with stainless thread inserts (Helicoil is a commonly-found brand name for this type of insert), which ought to be part of the original design of a heavy-duty rotator.

Table 5—Torque Limits Due to Rotator Mounting Bolts

	Thread root dia	*Shank dia*
Diameter, inch	0.3125	0.3125
Area	0.0522*	0.0767
f_s	75000	75000
Force/bolt	3900	5750
Radius, inch	2.12	2.12
N bolts	4	4
K_t	2.5	2.5
T_{max}, in-lb	13288	19524

* Area at thread root diameter

Table 6—Torque Limits Due to Bearing Force

	Steel	*Aluminum*
t, inch	0.11	0.25
f_b, lb/in^2	100000	56000
D, inch	0.3125	0.3125
Bearing force/bolt	3440	4375
Radius, inch	2.12	2.12
N bolts	4	4
K_t	2.5	2.5
T_{max}, in-lb	11660	14840

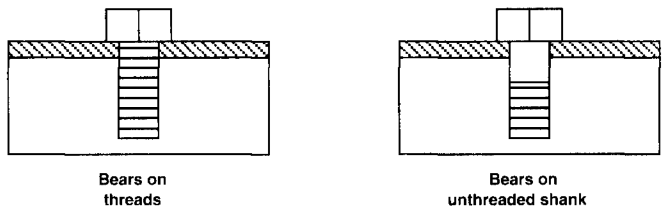

Fig 3—The best use of a bolt would result in shear forces being applied to unthreaded parts of the bolt, as at right.

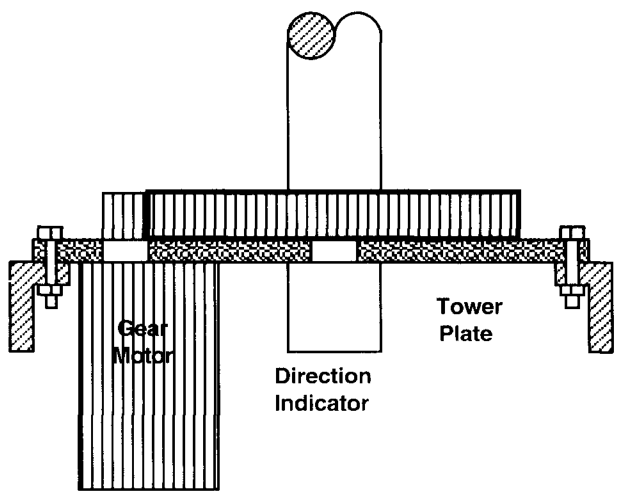

Fig 4—A more reliable rotator design would connect the output gear directly to the mast or the mast shock isolator.

Conclusions and Recommendations

Rotators using 1-inch keyed output shafts seem prone to failure under oscillating loads, and calculations suggest that even the best hardened-steel shafts will not reliably provide the braking torques specified by rotator manufacturers. Bolted or pinned flanges (like an automotive flywheel mounting) might help.

It has been suggested that using a longer key, or two keys in precision keyways, with setscrews holding them in place, improves the reliability of the shaft joints. My own experience is that this is at best a marginal improvement that still does not meet the manufacturers' torque specifications over an extended time. Manufacturers have recognized the problem and have introduced shafts made of stronger alloys, but larger-diameter output shafts seem indicated.

While I understand the economies of scale involved, I'm still amazed that a common washing machine, with its big motor, gearbox and control circuitry, can be bought for less than even an average antenna rotator. A more reliable rotator design would not use an output shaft but would connect the output gear directly to the mast (or the mast shock isolator), as is found in the classic and reliable prop-pitch motor design. A promising design could use an external output gear (perhaps built like a ring rotator, but with external teeth and a diameter small enough to fit inside a typical Rohn 25 or 45 tower) driven by a standard gear motor. Fig 4 shows such an arrangement.

This approach has the additional feature of integrating the cost of the tower rotator plate into the cost of the rotator itself, at the same time permitting the selection of an adequate plate thickness. A set of angle brackets would be required to attach to the tower legs, but the large radius would minimize the mechanical difficulty. The use of automotive engine mounts for some shock isolation at the plate mounting points has been suggested.

Appendix A — Strength of Bolts and Bearing Holes

Approximate shear and tensile strength of UNC AN bolts (similar to SAE Grade 5):[10]

Diameter, inch	*Shear, lb**	*Tensile, lb*
3/16	2070	1800
1/4	3680	3360
5/16	5750	5660
3/8	8280	8470
7/16	11200	11680
1/2	12760	15730

* for full diameter; if shear is across thread root diameter, it must be derated

Approximate bearing strength of steel sheets on bolts and pins:

Thickness, inch	*1/4 dia.*	*5/16 dia*	*3/8 dia*
0.049	1225 lb		
0.065	1625	2030	
0.083	2075	2594	3112
0.095	2375	2970	3560
0.120	3000	3750	4500
0.188	4680	5860	7030
0.250	6250	7800	9375

Thickness required for bearing strength approximately equal to bolt shear strength:

Diameter, inch	*Thickness, steel*	*Thickness, aluminum*
3/16	0.120 inch	0.188
1/4	0.188	0.250
5/16	0.188	0.3125
3/8	0.250	0.375
7/16	0.250	0.4375
1/2	0.250	0.500

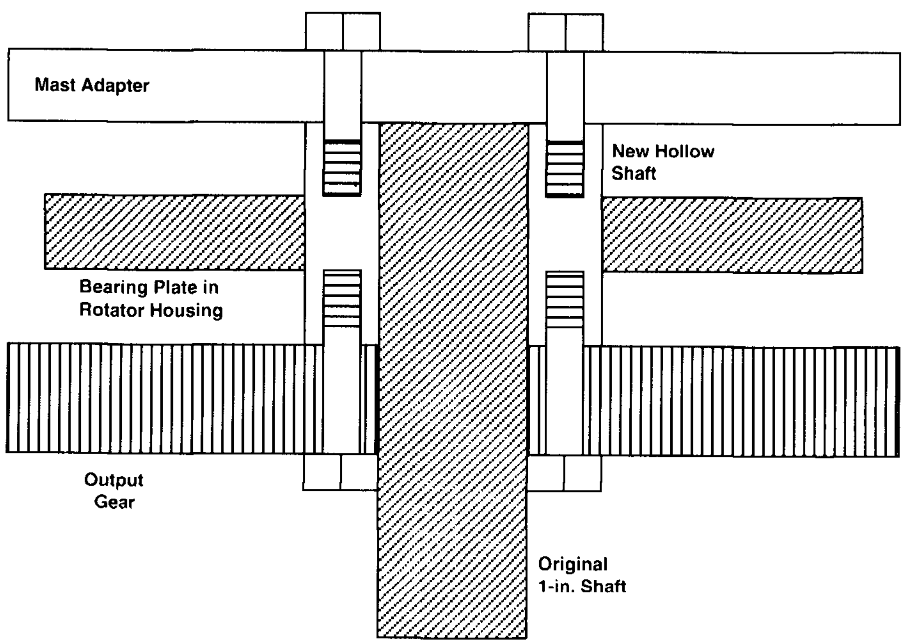

Fig 5—Improving existing rotator designs would involve use of 2-inch output shafts and proper use of bolts.

A stronger rotator could be also patterned after the internal gear arrangement of the TailTwister or Yaesu rotators (but with a motor and gears large enough to provide the required torque; even 0.1 horsepower at 1 rpm yields 6300 in-lb).

In my own station I have separated the control box and the motor power supply, locating the transformer under relay control at the base of each tower. This permits controlling several rotators from a single indicator, which can be of almost any type, and makes it easy to provide dc rotator power with a few additional components.[9] Favorable experience here with the TIC 1032 Ring Rotator, using two dc gearmotors, reflects two advantages of dc motors: (a) the motor provides the same torque as an ac motor but is not limited to 1800 or 3600 rpm, so with proper gearing more output torque is available, and (b) a dc motor acts as a brake if the terminals are shorted when not in use.

Existing rotator designs can be improved with 2-inch output shafts, perhaps with bolted or pinned connections instead of keyways. A 2-inch OD shaft with 1-inch ID could be fitted over an existing shaft and bolted/pinned to the output gear, which would provide nearly as much torsion capability as a solid 2-inch shaft. (See Fig 5.) The upper case half would require a larger bearing, but the stress at the mast plate joint would also be reduced. A suitable mast adapter plate can be made using a taper bushing (Morse Q-bushing) with or without a keyway. The use of an automotive shaft seal should also be considered, to keep water out of the gear box.

In any event, the disappointing performance of heavy-duty rotators with 1-inch output shafts seems confirmed by the engineering calculations here, and it is hoped that manufacturers will respond to the problem and offer an affordable solution.

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Polyphase Network Calculation using a Vector Analysis Method

Vector analysis leads to better design of 90° phase-shift networks.

By Tetsuo Yoshida, JA1KO

Introduction

It may seem difficult to understand the physical meaning of the polyphase networks that produce a 90° phase difference over a broad audio-frequency band. Many articles have been written on this subject since the polyphase network was first introduced to obtain a SSB signal.[1,2,3] However, no easily understood explanation has been published to date.

I intend to offer a simplified explanation of the method used to calculate polyphase networks using vector analysis based on basic trigonometry.

The amplitude response of a polyphase network is normally attenuated at mid-band frequencies when capacitors and resistors are selected in the traditional way. I propose a new system of selecting the capacitors and resistors that creates polyphase networks that have a flat amplitude response across the entire audio-frequency band.

The Vector Diagram of the Polyphase Network

A typical polyphase network is shown in Fig 1. When a push-pull audio signal is applied to the input terminals (1 to 4), four output signals having 90° phase differences appear at the output terminals at the right. (Capacitors and resistors in the same column have the same values.)

As shown in Fig 1, the same audio signal, E_i, is applied to input terminals 1 and 2, and an out-of-phase audio signal, $-E_i$, is applied to input terminals 3 and 4. The vectors at input terminals e_{01}, e_{02}, e_{03} and e_{04} are in a line, as shown in the vector diagram of Fig 2.

Now, let us consider the state of the output vectors in the first column, e_{11}, e_{12}, e_{13} and e_{14}, when nothing is connected to the output terminals of the first column. As e_{01} and e_{02} are the same and equal to E_i, it is evident that e_{11} is also same as E_i at all frequencies. Similarly, e_{13} is equal to $-E_i$. The vector e_{12} lies on the circle of Fig 2, which has a diameter of $2E_i$. At zero frequency, e_{12} coincides with e_{02} (or E_i), as the impedance of C1 is infinite. At a very high frequency, e_{12} coincides with e_{03} (or $-E_i$), as the impedance of C1 is negligibly small.

Since at this point nothing is connected to the column output terminals, the same current that flows through R1 flows through each associated C1.Vector e_r, the voltage appearing across R1, leads the voltage appearing at C1, vector e_c, by 90°. The amplitude of each vector is calculated

[1]Notes appear on page 15.

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as follows:

$$e_r = R_1 \cdot i$$

$$e_c = \frac{i}{j\omega C_1} \qquad \text{Eq 1}$$

where i is the current flowing through C_1 and R_1. The phase difference θ between e_{11} and e_{12} is:

$$\theta = 2\tan^{-1}\left(\frac{e_r}{e_c}\right) = 2\tan^{-1}(\omega C_1 \cdot R_1) = 2\tan^{-1}\left(\frac{\omega}{\omega_1}\right) \qquad \text{Eq 2}$$

where

$$\omega_1 = \frac{1}{C_1 \cdot R_1} \qquad \text{Eq 3}$$

At a frequency of $0.5774/(2\pi C_1 \times R_1)$, for example, the angle between e_{11} and e_{12} is 60°, and at a frequency of $1/(2\pi C_1 \times R_1)$, it is 90°. Vector e_{14} also lies on the same circle but on the opposite side.

Now, let's connect the second column (C2 and R2). To make things simple, the impedance of the second column is set high compared to that of the first column. If the impedance of the second column is assumed to be high, vectors e_{11}, e_{12}, e_{13} and e_{14} in Fig 3 will be same as those of Fig 2.

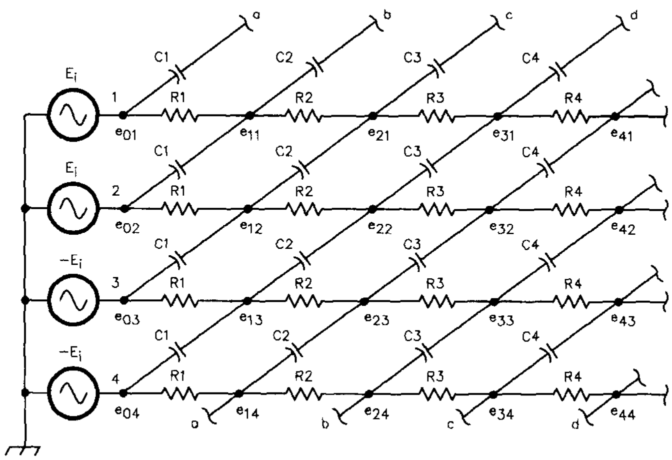

Fig 1—A typical polyphase network.

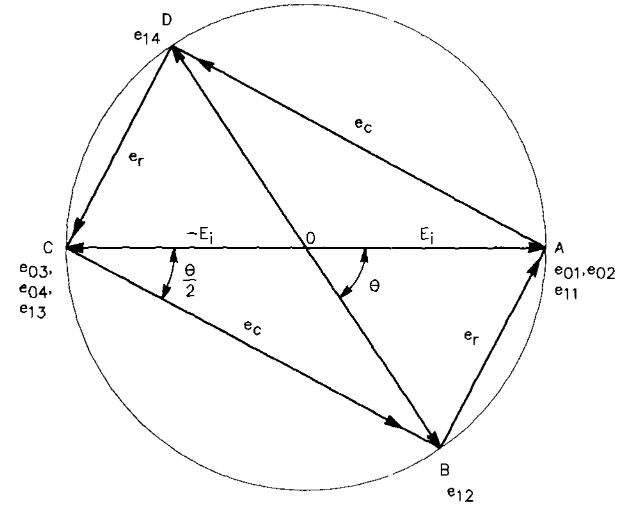

Fig 2—The vectors at the input and the output terminals of the first column.

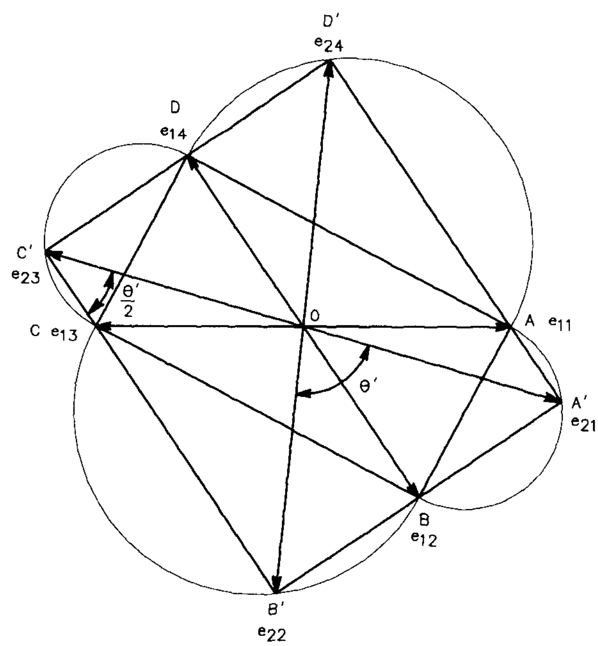

Fig 3—The vectors at the input and output terminals of the second column.

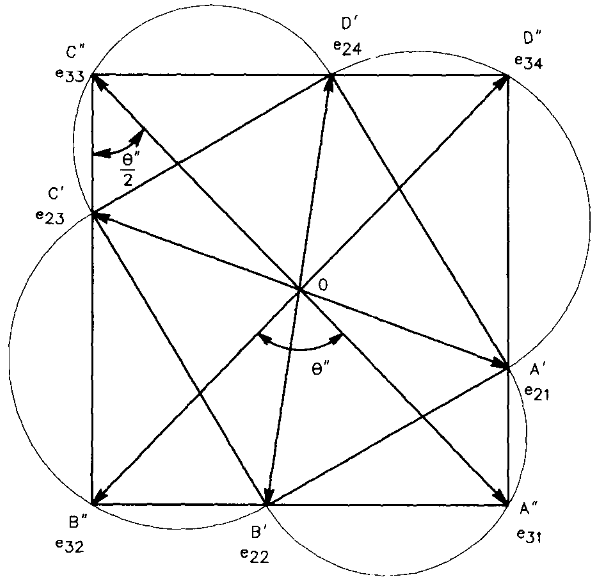

Fig 4—The vectors at the input and output terminals of the third column.

Vector e_{21} lies on the circle having a diameter AB, and vector e_{22} lies on the circle having a diameter BC. In Fig 3, it is easily shown that the angle between vectors e_{21} and e_{22} (that is the angle A′OB′) is twice the angle A′C′B′. Therefore, the angle between e_{21} and e_{22}, θ′ is:

$$\theta' = 2\tan^{-1}\left(\frac{A'B'}{B'C'}\right) \qquad \text{Eq 4}$$

In the case of R1=R2 and C1=C2, at a frequency of $0.5774/(2\pi C_1 \times R_1)$, the angle θ′ is 81.7868°, which is closer to 90° than the angle θ=60° at the output terminal of the first column.

Applying the same assumption of higher impedance, a third column can then be added. In Fig 4, the angle between vectors e_{31} and e_{32} is twice the angle A″ C″ B″ and can be obtained by:

$$\theta'' = 2\tan^{-1}\left(\frac{A''B''}{B''C''}\right) \qquad \text{Eq 5}$$

Similarly, the angle at a frequency of $0.5574/(2\pi C_1 \times R_1)$ is 87.7958° which is closer to 90° than that of the second column output.

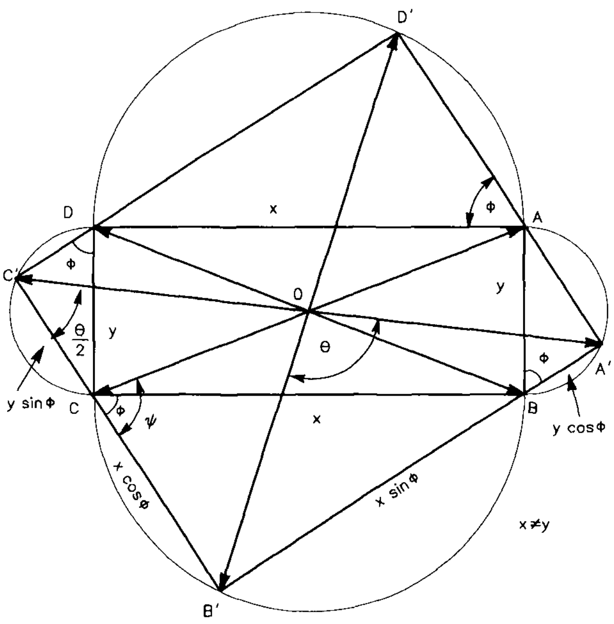

Fig 5

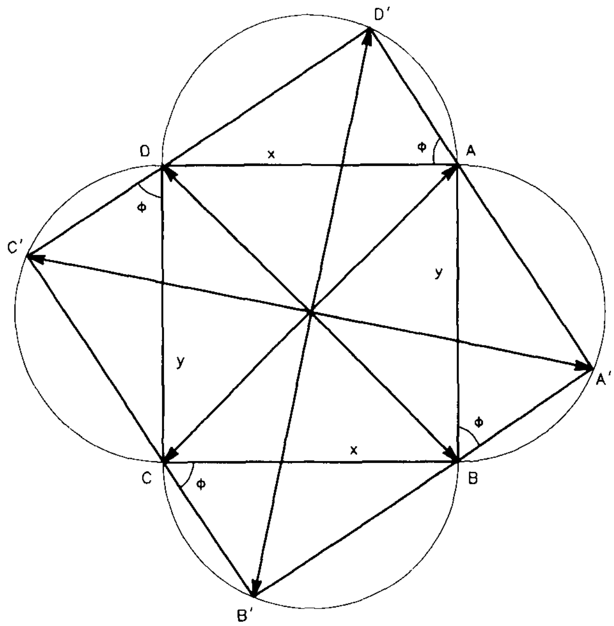

Fig 6

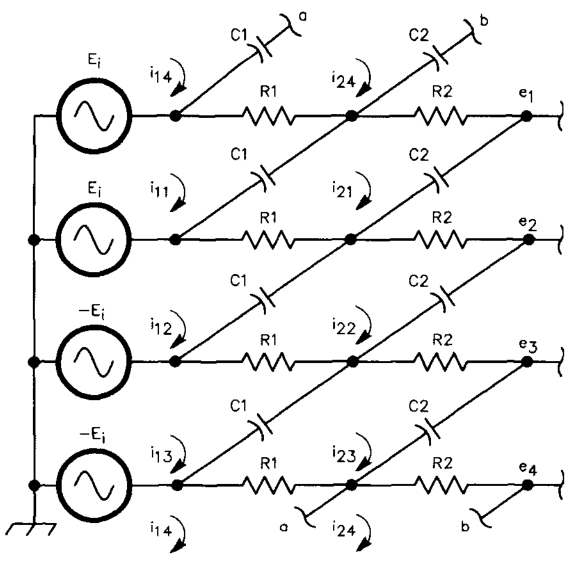

Fig 7—The circuit used to evaluate the impedance assumption.

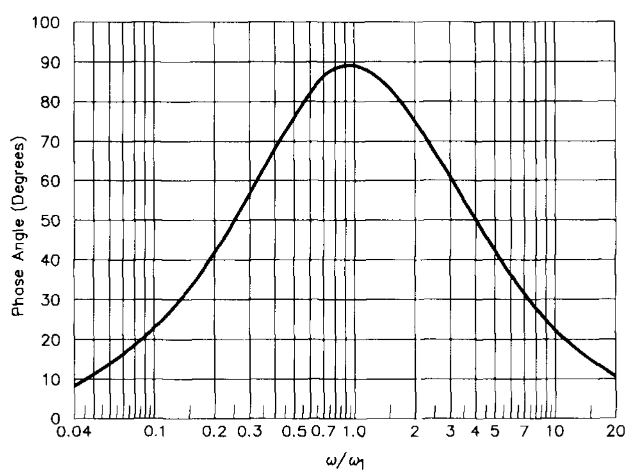

Fig 8—The phase difference between the vector e_1/E_i and e_2/E_i of Fig 7 ($\omega_1 = \omega_2$, any n).

As the audio signal passes through the columns, the phase angle between the output audio vectors gets closer and closer to 90°. Fig 5 explains this process. In this figure, A, B, C and D show the input vectors of a column and A′, B′, C′ and D′ show the output vectors of the same column. Applying the same assumption indicated above—the impedance of the next column is high enough compared to that of the preceding column:

$$A'B' = y\cos\phi + x\sin\phi$$
$$B'C' = x\cos\phi + y\sin\phi \qquad \text{Eq 6}$$

The ratio of the arms of the output rectangle is:

$$\frac{A'B'}{B'C'} = \frac{y\cos\phi + x\sin\phi}{x\cos\phi + y\sin\phi} \qquad \text{Eq 7}$$

As the ratio of the arms of the input rectangle is y/x, every time an audio signal passes through a column, the ratio changes as follows:

$$\frac{\dfrac{A'B'}{B'C'}}{\dfrac{AB}{BC}} = \frac{1 + \dfrac{x}{y}\tan\phi}{1 + \dfrac{y}{x}\tan\phi} \qquad \text{Eq 8}$$

If y/x is smaller than 1, the above ratio is larger than 1, and if y/x is more than 1, the above ratio would be less than 1. In all cases, the rectangles composed of four vectors become closer, forming a square as the vectors pass through columns.

When a column is added after the vectors achieve a 90° angle, the output vectors would still have the same 90° angle; the angles between the vectors are not affected. Fig 6 shows that if the input vectors become square, the output vectors are square. In this case, the result of the equation (Eq 8) becomes 1.

This is a very important result. In the polyphase network, if the vectors of an audio signal form a 90° angle, this 90° angle remains unchanged even if more columns are added afterwards, and this is the reason why the polyphase network is able to form a 90° phase shift over a very wide range of audio frequencies.

Evaluation of the Assumption

The above vector analysis is based on the assumption that the impedance of each column is high compared to that of the preceding column. If this assumption isn't true, what effect does the impedance of a column have on the vectors produced by the preceding columns?

The effect can be evaluated using the circuit shown in Fig 7, which is a two-column polyphase network. Applying Kirchhoff's law:

$$\begin{aligned}
&R_1(i_{11} - i_{24}) + \frac{1}{j\omega C_1}(i_{11} - i_{21}) = 0\\
&R_1(i_{12} - i_{21}) + \frac{1}{j\omega C_1}(i_{12} - i_{22}) = 2E_i\\
&R_1(i_{13} - i_{22}) + \frac{1}{j\omega C_1}(i_{13} - i_{23}) = 0\\
&R_1(i_{14} - i_{23}) + \frac{1}{j\omega C_1}(i_{14} - i_{24}) = -2E_i\\
&R_1(i_{21} - i_{12}) + \frac{1}{j\omega C_1}(i_{21} - i_{11}) + \left(R_2 + \frac{1}{j\omega C_2}\right)i_{21} = 0\\
&R_1(i_{22} - i_{13}) + \frac{1}{j\omega C_1}(i_{22} - i_{12}) + \left(R_2 + \frac{1}{j\omega C_2}\right)i_{22} = 0\\
&R_1(i_{23} - i_{14}) + \frac{1}{j\omega C_1}(i_{23} - i_{13}) + \left(R_2 + \frac{1}{j\omega C_2}\right)i_{23} = 0\\
&R_1(i_{24} - i_{11}) + \frac{1}{j\omega C_1}(i_{24} - i_{14}) + \left(R_2 + \frac{1}{j\omega C_2}\right)i_{24} = 0
\end{aligned} \qquad \text{Eq 9}$$

substituting:

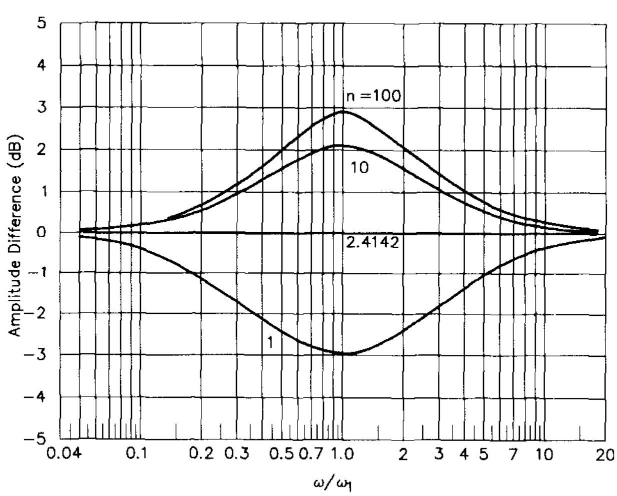

Fig 9—The difference in amplitude between the input and output terminals of Fig 7 ($\omega_1 = \omega_2$).

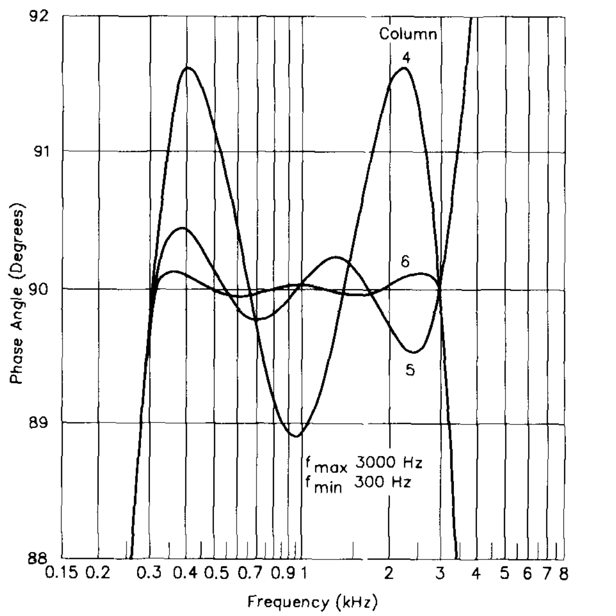

Fig 10—Phase difference of 4-, 5-, and 6-column polyphase networks.

$$\omega_1 = \frac{1}{C_1 \cdot R_1}, \omega_2 = \frac{1}{C_2 \cdot R_2}, n = \frac{R_2}{R_1} \qquad \text{Eq 10}$$

and showing the output vectors as e_1, e_2, e_3 and e_4, Eq 9 is solved as:

$$\frac{e_1}{E_i} = \frac{1 + \frac{\omega^2}{\omega_1 \cdot \omega_2} + j\left(\frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}{1 - \frac{\omega^2}{\omega_1 \cdot \omega_2} + j\left(\frac{2\omega}{n \cdot \omega_2} + \frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}$$

$$\frac{e_2}{E_i} = \frac{1 + \frac{\omega^2}{\omega_1 \cdot \omega_2} - j\left(\frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}{1 - \frac{\omega^2}{\omega_1 \cdot \omega_2} + j\left(\frac{2\omega}{n \cdot \omega_2} + \frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}$$

$$\frac{e_3}{E_i} = -\frac{1 + \frac{\omega^2}{\omega_1 \cdot \omega_2} + j\left(\frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}{1 - \frac{\omega^2}{\omega_1 \cdot \omega_2} + j\left(\frac{2\omega}{n \cdot \omega_2} + \frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)} \qquad \text{Eq 11}$$

$$\frac{e_4}{E_i} = -\frac{1 + \frac{\omega^2}{\omega_1 \cdot \omega_2} - j\left(\frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}{1 - \frac{\omega^2}{\omega_1 \cdot \omega_2} + j\left(\frac{2\omega}{n \cdot \omega_2} + \frac{\omega}{\omega_1} + \frac{\omega}{\omega_2}\right)}$$

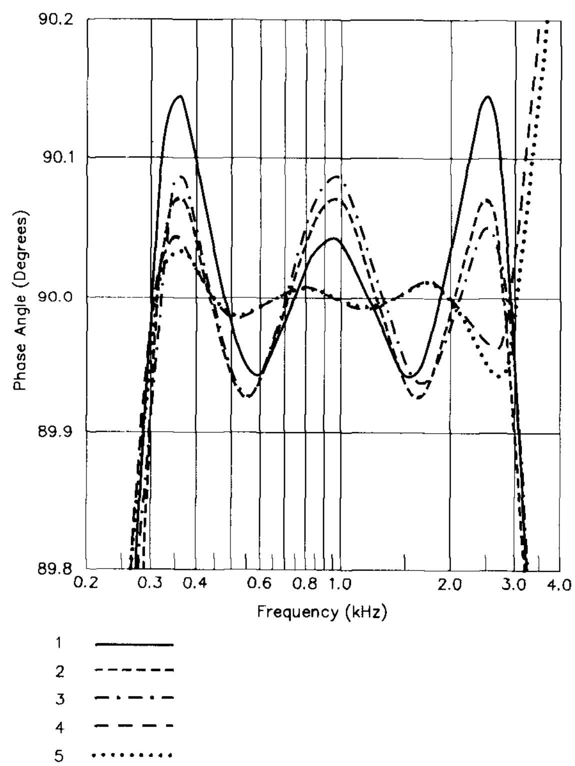

Fig 11—A comparison of standard networks with Chebychev networks.The networks are described in Table 1.

To make it simpler, suppose $\omega_1=\omega_2$. The phase difference for this case between vectors e_1/E_i and e_2/E_i is shown in Fig 8. The amplitudes of the four vectors of Eq 8 are all identical and are shown in Fig 9. Fig 9 shows that the amplitudes of the output vectors are increased as n, the ratio of resistance of the next column to that of the preceding column, is increased. If n is 1, as has been the case so far, the output vector is decreased by 3 dB compared to the input vector at the center frequency. If $\omega_1=\omega_2$, the output vector will have the same amplitude as that of the input vector when n is 2.414, or 1 plus $\sqrt{2}$. When n is increased more, the output vector becomes larger than the input vector—as much as 3 dB. It is interesting that although the amplitude is affected by n, the phase is *not* affected by n at all, as shown in Fig 8. This confirms that the assumption introduced here is generally acceptable.

I compared Fig 8 against values calculated using a vector method with the same assumption and found it to be true. Therefore, it can be concluded that, as far as the angles of the output vectors of polyphase networks are concerned, the vector analysis method based on basic trigonometry is applicable.

This analysis leads to another important fact about the ratio of the amplitudes of the output vectors to those of the input vectors. All the vector amplitudes of Eq 11 have the same value. Setting it to unity, the following equation is obtained:

$$n = \frac{m + 1 + \sqrt{m^2 + 6m + 1}}{2m} \qquad \text{Eq 12}$$

where

$$m = \frac{\omega_2}{\omega_1} \qquad \text{Eq 13}$$

This means that the resistance of each column must be n times of that of the preceding column, and that value changes according to the ratio of the time constants of the the two columns.

Calculation of the Angle between Output Vectors

Referring to Fig 5 and Eq 6, the angle between output vectors and the phase-shift value between input and output vectors are obtained by a series of calculations using the following equations:

$$f(p) = \frac{1}{2\pi C_p \cdot R_p}$$

$$S(p) = \tan^{-1}\left\{\frac{f}{f(p)}\right\} = \phi$$

$$X(p) = X(p-1) \cdot \cos\{S(p)\} + Y(p-1) \cdot \sin\{S(p)\}$$

$$Y(p) = Y(p-1) \cdot \cos\{S(p)\} + X(p-1) \cdot \sin\{S(p)\}$$

$$H(p) = 2\tan^{-1}\left\{\frac{Y(p)}{X(p)}\right\} = \theta$$

$$K(p) = \tan^{-1}\left\{\frac{Y(p)}{X(p)}\right\} = \frac{\theta}{2} \qquad \text{Eq 14}$$

$$L(p) = \tan^{-1}\left[\frac{Y(p)}{X(p-1) \cdot \cos\{S(p)\} - Y(p-1) \cdot \sin\{S(p)\}}\right] = \Psi$$

$$M(p) = L(p) - K(p)$$

where

p is the number of columns, from 1 to c, and p–1 means the number of the preceding column.

C_p and R_p are the capacitance and resistance of the pth column.

$H(p)$ is the angle between the pth column output vectors of the 1st row and 2nd row.

$M(p)$ is the phase shift between the pth and (p-1)th columns of the first row.

$X(0)$ and I(0) are the input audio vector components, and $X(0)=1$ and $Y(0)=0$ can be used as initial conditions.

Examples of calculated values for 4-, 5- and 6-column polyphase networks are shown in Fig 10. In this calculation, the node frequencies of each column, which are equal to $1/(2\pi C \times R)$, are determined by the following equations:

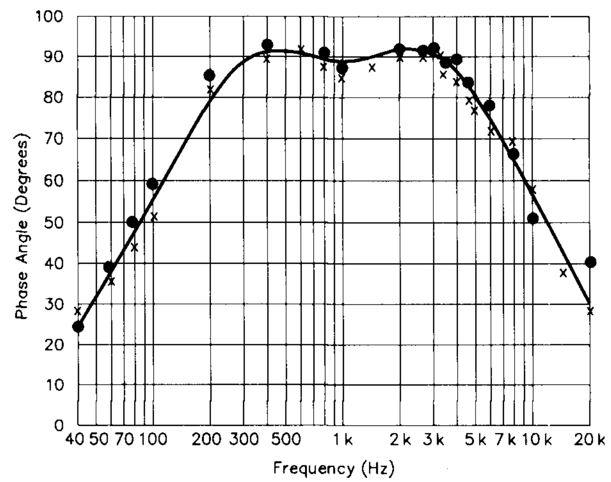

Fig 12—The phase angle of two example 4-column networks.
• Resistance value obtained using Eq 12.
x Constant resistance value.

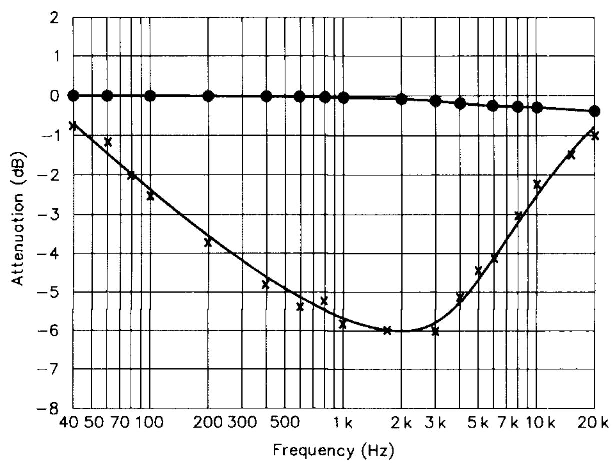

Fig 13—The attenuation of the 4-column networks.
• Resistance value obtained using Eq 12.
x Constant resistance value.

$$f_p = f_{min} \cdot \left(\frac{f_{max}}{f_{min}} \right)^{\left(\frac{p-1}{c-1} \right)} \quad \text{Eq 15}$$

where

f_{min} is the lowest frequency of the audio pass band and is f_1.

f_{max} is the highest frequency of the audio pass band and is f_c.

The resistance used in each column is increased over that of the preceding column by n (Eq 12).The column capacitance is obtained by:

$$C_p = \frac{1}{2\pi f_p \cdot R_p} \quad \text{Eq 16}$$

In this case, the node frequencies makes a geometrical progression. A beautiful Chebychev curve may be obtained if the node frequencies are selected carefully. Fig 11 shows a comparison of a 6-column network designed using the equations mentioned above (curve 1) with a Chebychev 6-column network by W9CF (curve 2) (see Note 2). The node

Table 1

Data of the networks (f in Hz, C in μF, R in kΩ)

Mark	*1*	*2*	*3*	*4*	*5*
f_1	300.0	314.2	312.1	300.0	312.1
f_2	475.5	435.5	442.1	440.3	442.1
f_3	753.6	720.3	723.4	646.3	658.2
f_4	1194.3	1249.5	1326	948.7	970.5
f_5	1892.9	2066.8	2122	1392.5	1463
f_6	3000.0	2864.5	2842	2043.9	2010
f_7	_	_	_	3000.0	3121
R_1	10.000	10.000	10	10.00	10
R_2	19.539	10.000	10	20.188	20
R_3	38.176	10.000	10	40.754	39
R_4	74.592	10.000	10	82.273	82
R_5	125.744	10.000	10	166.09	160
R_6	284.767	10.000	10	335.30	330
R_7	_	_	_	676.89	680
C_1	0.05305	0.05065	0.051	0.05305	0.051
C_2	0.01713	0.03655	0.036	0.01791	0.018
C_3	0.005532	0.02210	0.022	0.006042	0.0062
C_4	0.001787	0.01274	0.012	0.002039	0.0020
C_5	0.0005769	0.007701	0.0075	0.0006881	0.00068
C_6	0.0001863	0.005556	0.0056	0.000232	0.00024
C_7	_	_	_	0.00007838	0.000075

Table 2

f_p in Hz, C_p in μF, R_p in kΩ

Number of Column	*First Network* f_p	C_p	R_p	*Second Network* f_p	C_p	R_p
1	312.1	0.0068	75	294.7	0.018	30
2	643	0.0033	75	664	0.0047	52
3	1415	0.0015	75	1457	0.0012	91
4	3121	0.00068	75	3215	0.00033	150

frequencies and component values are shown in Table 1 mark 1 and mark 2.

The angle deviations from 90° in the audio passband are 0.07 degrees better in the case of the 6-column Chebychev than for the geometrical progression circuit described here, for a ratio between f_{max} and f_{min} of 10. It is more practical, however, to choose node frequencies determined by using capacitors and resistors of standard values that are readily available. An example is shown in Table 1 mark 3. Curve 3 in Fig 11 shows a case where standard value capacitors and resistors are used in a Chebychev network. When the number of columns is increased by one —to 7 columns instead of 6—and the exact-value C_s and R_s are used (mark 4 Table 1), the improvement in deviation is 0.1° (curve 4 of Fig 11). When standard-value C_s and R_s are used in a 7-column network (mark 5 Table 1), the angle deviations from 90° are slightly better (curve 5 of Fig 11) than the value of the ideal Chebychev 6-column network (curve 2 of Fig 11).

This means that it may be more economical to increase the number of columns than to try to procure uncommon values of resistors and capacitors.

An Actual Example and Conclusions

A pair of polyphase networks of four columns each were prepared and measured. The first network was constructed using the current basic design in which the values of all resistances are equal. The second network was designed using the technique proposed here, in which the values of resistance are increased according to Eq 12. The component data are shown in Table 2. The measured circuit responses are shown in Figs 12 and 13.

Fig 12 shows the angles between the output vectors of each network. The measured values here show no significant phase difference between the two networks. On the other hand, Fig 13 shows the attenuation of the networks, and it is quite clear that the network using increasing resistance values displays better frequency response.

In conclusion, I've explained why polyphase networks can produce a wide-band 90° phase difference and have shown a method that improves the amplitude response of such networks. Although the analysis is based on certain assumptions, the measurements taken from the actual sample network confirm that this theory can be valuable for practical network design.

Notes

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[2]Gschwindt, A. (HA5WH), "Some reflections on the four-way phasing method," *Radio Communication*, 1976, January, pp 28-33.

[3]Schmidt, Kevin (W9CF), "Phase-Shift Network Analysis and Optimization," *QEX*, 1994, April, pp 17-23. □□

Refinements in Crystal Ladder Filter Design

Improved design techniques can result in much better wide-bandwidth filters.

By Wes Hayward, W7ZOI

Hams have been building their own crystal filters since the earliest days of single sideband. Early motivations were economic; commercially built filters were either too expensive or unavailable. Quartz crystals offered a method for achieving the selectivity needed in SSB transmitters and receivers. The trend continues, especially among QRP enthusiasts.

Recent work by Carver and by Makhinson has taken the process further.[1,2] Both examined the construction and design of very high-performance crystal filters. Their goal was to build filters offering performance that was not commercially available. This work produced some spectacular filter performance. Both based their work on one of several versions of *X.EXE*, a computer program for crystal filter design that I wrote several years ago.[3]

Recent communications (including those with Makhinson) mention limitations in the design method used in *X.EXE*. Wide filters were difficult, if not impossible. Examination of the underlying filter theory revealed a simple *circuit modification* to extend the design methods. This circuit extension produces filters with wider bandwidths at a greater variety of center frequencies, including filters using overtone modes. The new freedom allows us to build filters with improved time-domain performance, the primary goal in the filter designs of Carver.

Before examining circuit modifications, the original mathematical methods will be presented—allowing the limitations to be evaluated—beginning with a review of L-C filters using crystal-like circuits. The problems encountered when we substitute crystals into this framework are then presented, leading to the methods used in *X.EXE*. Finally, the circuit modifications are presented that extend the capabilities.

This article is practical to the extent that it produces filters that are inexpensive and relatively easy to build. The filters really work well. The methods, however, are mathematical. Hence, the article is more analytical than is usual in amateur literature.

[1]Notes appear on page 21.

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L-C Filter Background

The bandwidth of a band-pass filter defines a filter Q. This parameter, $Q_f = F_{center} / Bandwidth$, must be less than the unloaded Q of the resonators. A filter is then completely determined if the following conditions are met:

1) The singly loaded end section Q is

established in accordance with the polynomial of choice (Chebyshev, Bessel, etc).

2) The couplings between resonators are set to fit the polynomial.

A practical, although uncommon, example would use L-C *series* tuned circuits with large L and small C. This topology is presented in Fig 1.[4]

Unloaded resonator Q is easily measured. Once known, a *normalized* Q can be calculated for a filter as the ratio, $Q_0 = Q_u / Q_f$. Tables in Zverev list the values of Q_0 that are needed for a given polynomial filter to be realizable.[5] If the resonators are not good enough, it will be impossible to realize some filter types.

Consider an example, a 4th-order Butterworth filter, which has normalized parameters $q_1 = q_4 = 0.765$, $k_{12} = 0.841$, $k_{23} = 0.541$, and $k_{34} = 0.841$. Assume that the inductors are 10 µH while the series capacitor is 101.3 pF. Assume also that both L and C are fixed; we can't adjust them. (This restriction is the same one we encounter with crystals.) The L-C combination is series resonant at 5 MHz. Assume further that $Q_u = 250$, a typical value for toroid inductors. Examination of Zverev's tables shows that $Q_0 > 3.7$ will allow the construction of a 4-resonator Butterworth filter. The unloaded resonator bandwidth is 20 kHz, so 4th-order Butterworth band-pass filters with a bandwidth of 74 kHz or more are realizable. We will design this filter for $B = 200$ kHz, so $Q_f = 25$ at 5 MHz.

The end section Q is denormalized according to the equations summarized on page 92 of Note 3:

$$Q_{END} = \left(\frac{1}{q \cdot Q_f} - \frac{1}{Q_u} \right)^{-1} \qquad \text{Eq 1}$$

where q (lower case) is the normalized end Q. The value for q was 0.765 for both ends of the Butterworth filter, so denormalized $Q_{END} = 20.71$. The inductor reactance at the filter center frequency is $\omega L = 314.2\ \Omega$, so the end sections will have a singly loaded Q of 20.71 if the end termination is 15.2 Ω. The coupling coefficients are denormalized with regard to Q_f:

$$K_{12} = \frac{k_{12}}{Q_f} \qquad \text{Eq 2}$$

where k_{12} (lower case) is the normalized coupling coefficient between the 1st and 2nd resonator. The coupling capacitor is then:

$$C_{12} = \frac{C_0}{K_{12}} \qquad \text{Eq 3}$$

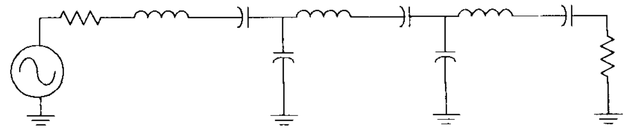


Fig 1—L-C filter topology using series tuned circuits.

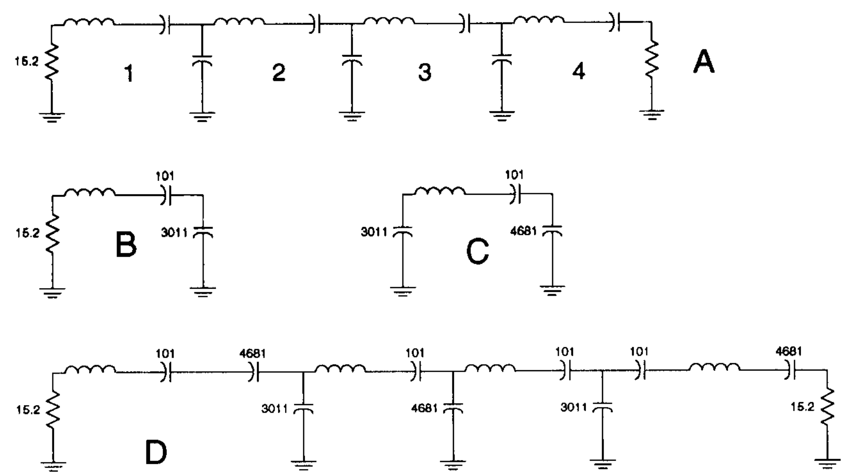


Fig 2—Evolution of a 4-resonator band-pass L-C filter including tuning. See text.

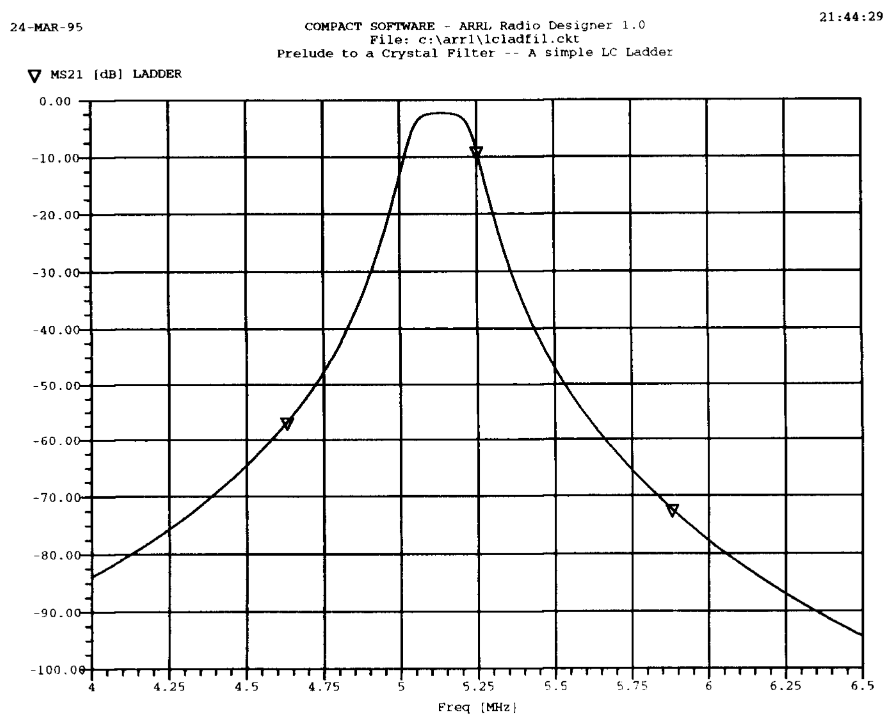


Fig 3—L-C filter response. The filter is that of Fig 2D.

where C_o is the nodal capacitance for the resonator, 101.3 pF for the example. With $k_{12} = 0.841$, the coupling capacitor is $C_{12} = 3011$ pF, a large but realizable value. C_{34} has the same value. A similar process generates the middle coupling capacitor, $C_{23} = 4681$ pF.

The almost-complete filter is shown in Fig 2A. We must still tune the filter. The simple resonators all started on the same 5-MHz frequency, but this synchronous tuning was upset when we inserted coupling capacitors. Mesh 1, shown in Fig 2B, is resonant at 5.084 MHz while mesh 2 (Fig 2C) tunes to 5.137 MHz. Owing to symmetry, mesh 3 is identical to mesh 2 while mesh 4 is the same as mesh 1. A filter becomes properly tuned when we insert extra series C in meshes 1 and 4, forcing all four meshes to be resonant at the same frequency, 5.137 MHz. The final circuit is shown in Fig 2D. The calculated response of this filter is presented in Fig 3.

The Quartz Crystal

The equivalent circuit for a crystal is given in Fig 4. The motional parameters, L_m and C_m, form a series tuned circuit while ESR, the *equivalent series resistance*, models crystal losses. This is, so far, no different from the series tuned circuits considered in the previous L-C example. However, the quartz crystal is complicated by an additional element, C_o, the parallel crystal capacitance. This component is the result of the basic physical crystal structure, a slab of quartz with metalization on both sides. This is a simple parallel-plate capacitor that is unrelated to the piezoelectric properties of the material. C_o of Fig 4 includes any stray capacitance that might be in the crystal package. The value for C_o is usually related to the motional capacitance by an approximation, $C_o = 220C_m$.[6]

The component values are much different than what we might realize with off-the-shelf L and C. For example, a 10-MHz crystal that I've used for numerous filter experiments has $L_m = 0.02$ H, $C_m = 12.67$ fF, and $C_o = 3$ pF. (1 fF = 0.001 pF = 1 femtofarad.)

As an initial approximation, the filter designer might ignore C_o. The crystal is then nothing more than a series resonator, and the design of filters is *exactly* like the L-C example. Unfortunately, this works only for the narrowest of filters. As the filter becomes wider, the discrepancy between design and measured results is extreme. (It is not necessary to build the filters to see these effects. The difficulties are quite evident in computer simulation with C_o included in the crystal model.) The coupling capacitors calculated when C_o is ignored are too large and may not be in the right ratio. The result is a bandwidth that is too narrow and a distorted shape. A smaller value of motional C should be used when calculating coupling elements. But what are appropriate motional elements? An answer was found in a *reactance slope approximation*, the method that formed the basis for *X.EXE*.

Consider a simple lossless series tuned circuit, Fig 5. The complex impedance is given as:

$$Z = j \cdot X = j\left(\omega \cdot L - \frac{1}{\omega \cdot C}\right) \qquad \text{Eq 4}$$

where $\omega = 2\pi f$. At resonance:

$$\omega_o \cdot C = \frac{1}{\omega_o \cdot L} \qquad \text{Eq 5}$$

which allows us to eliminate C:

$$j \cdot X(\omega) = j\left(\omega \cdot L - \frac{\omega_o^2 \cdot L}{\omega}\right) \qquad \text{Eq 6}$$

Differentiating this expression with respect to angular frequency, ω:

$$\frac{\partial X}{\partial \omega} = L + \frac{\omega_o^2 \cdot L}{\omega^2} \qquad \text{Eq 7}$$

This expression can be solved for inductance as a function of reactance slope with frequency, providing a new definition for effective inductance:

$$L = \frac{\dfrac{\partial X}{\partial \omega}}{1 + \dfrac{\omega_o^2}{\omega^2}} \qquad \text{Eq 8}$$

Consider now the equivalent circuit for the crystal shown in Fig 4. Ignoring ESR, the series tuned circuit portion has the impedance:

$$Z_{SER} = j\left(\omega \cdot L - \frac{\omega_o^2 \cdot L}{\omega}\right) \qquad \text{Eq 9}$$

This is converted to an admittance, $Y = 1/Z$, and the admittance of the parallel capacitance, C_o, is added, yielding:

$$Y = j\left[\frac{\omega \cdot C_o \cdot L \cdot (\omega^2 - \omega_o^2) - \omega}{L \cdot (\omega^2 - \omega_o^2)}\right] \qquad \text{Eq 10}$$

Converting back to impedance form, the reactance becomes

$$X = \frac{L_m \cdot (\omega_o^2 - \omega^2)}{\omega \cdot C_o \cdot L_m \cdot (\omega^2 - \omega_o^2) - \omega} \qquad \text{Eq 11}$$

This reactance is differentiated with respect to angular frequency, ω, with the result inserted into Eq 8 to produce the desired result, an expression for effective resonator inductance of a crystal in a filter:

$$L_{EFF} = L_m \cdot \left[\frac{2 \cdot A}{(A-1)^2 \cdot \left(1 + \dfrac{\omega_o^2}{\omega^2}\right)} - \frac{1}{A-1}\right] \qquad \text{Eq 12}$$

where

$$A = L_m \cdot C_o \cdot (\omega^2 - \omega_o^2) \qquad \text{Eq 13}$$

We define $X(\delta f)$ as the ratio of the effective inductance to the motional inductance, with δf being an offset frequency in Hertz above the crystal resonance. Then, Fig 6 shows $X(\delta f)$ plotted as a function of offset for a typical 5-MHz crystal with $L_m = 0.098$ H. Two parallel capacitance values are used, resulting in the two curves. The lower curve is for $C_o = 2$ pF while the upper one is for $C_o = 5$ pF. The effective inductance ratio is plotted for offsets up to 2500 Hz. The effect is dramatic, especially when C_o is large.

This effect, an increase in effective inductance resulting from added parallel capacitance, is a familiar one. It happens when we parallel an inductor in a low-pass filter to generate fre-

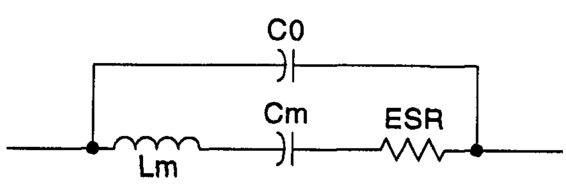


Fig 4—Quartz crystal model.

Fig 5—Ideal series tuned circuit.

quencies of high attenuation in the stopband (elliptic filters). It also happens when we build a trap for use in an antenna. In both cases, the addition of parallel C causes the inductor to behave like one with a larger value.

Crystal Filter Design

Now that we have isolated an effect—increased effective inductance due to holder capacitance—that will complicate filter performance, we can try to compensate for it. From Fig 6, if the filter is narrow, with a bandwidth of only a few hundred Hertz, the inductance ratio is very close to 1 over the entire band. We can then build effective filters by allowing the inductance to equal the motional value. Wider-bandwidth filters are built by evaluating the effective L over the passband. The L_{eff} value calculated at the top of the passband seems to work well for SSB filters. That value, and the resulting motional capacitances, are used for filter design. This is done in *X.EXE*.

The results shown in Fig 6 describe a particular 5-MHz crystal, although the effect is a general one. Going to a higher crystal frequency partially alleviates the problem, while lower frequencies complicate the outcome. This is the effect that led Makhinson (Note 2) to observe that effective lower-sideband ladder filters with an SSB bandwidth are best built for center frequencies between 6 and 12 MHz.

The approach used in *X.EXE* seems to produce useful and predictable filters when $L_{eff} < 2L_m$. The rest of the design is no different than the L-C method outlined above. The reader should review the results obtained by Carver and Makhinson to see what is possible with these methods.

Motional inductance and capacitance can be measured with a variety of methods. The scheme shown in the sidebar is the one I presently use with most crystals.

Additional Refinements

The methods presented are not new. Rather, they represent but one mathematical formulation for the design. This method grew from correspondence with Dr. Dave Gordon-Smith, G3UUR. Dave had used similar analytical techniques to derive a unique set of normalized filter tables for crystal filter design with arbitrary crystals with any value for C0.[7] While not published, Dave has freely distributed his work (and tables) to many amateur experimenters over the years. His analysis predated and contributed to the "L_m effective" approach that I've presented above (Eq 12). Dave also originated the crystal measurement scheme presented in the sidebar.

Wider Bandwidth Crystal Filters

The design methods outlined so far do not work well (if at all) when wide-bandwidth filters are to be built. Even when the methods appear to be working with regard to filter shape and bandwidth, the resulting circuits may have excessive group delay near one end of the passband. This is another consequence of C_o, the nemesis of our problem. There is a simple solution to the problem: add circuitry that will cancel the effects of C_o.

The modified design procedure will be illustrated with an example. The circuit desired is one with a bandwidth of 3 kHz, suitable for AM reception. The center frequency is 3.58 MHz, the frequency of readily available color-

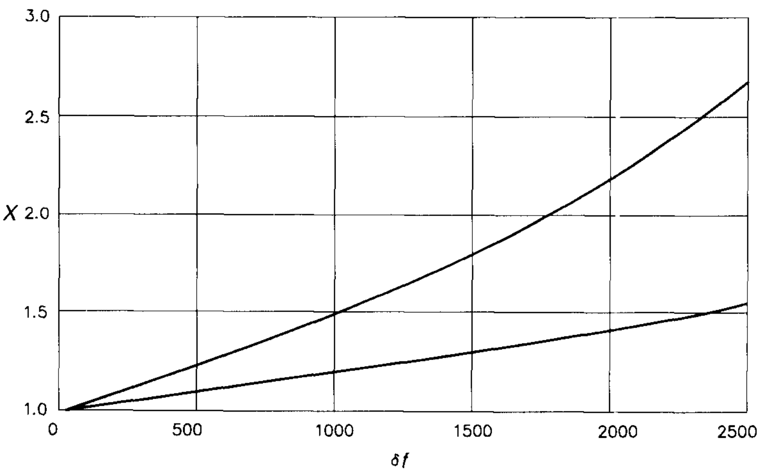


Fig 6—*X*, defined as L_{eff}/L_m, is plotted for frequency offset, δ*f*, above the crystal series-resonance frequency in Hertz. These 5-MHz crystals had parallel C of 2 pF (lower curve) and 5 pF.

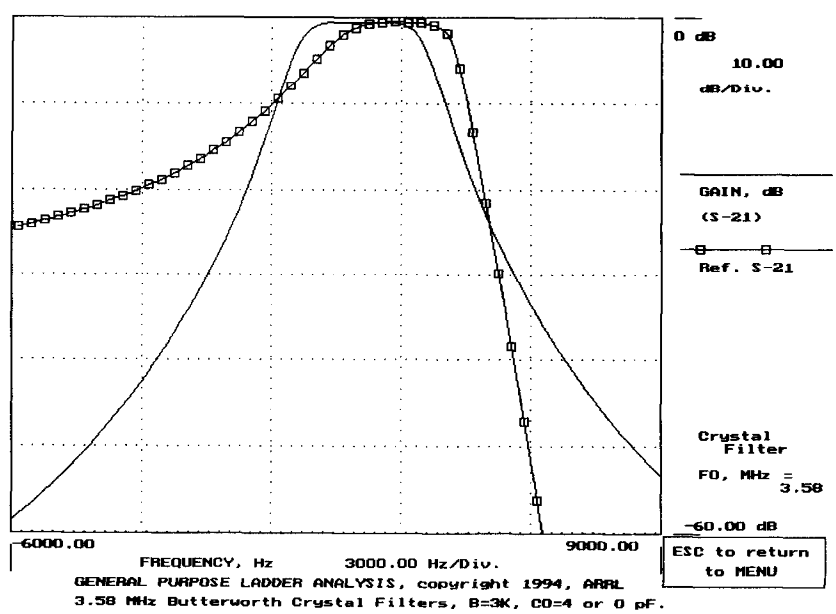


Fig 7

burst crystals. Circuit design begins with an attempt at using the standard methods. The available crystals have a motional L of 0.117 H and C_0 of 4 pF. This data, along with the *k* and *q* data for a Butterworth polynomial were entered into *X.EXE*. The resulting filter was then analyzed; the results are presented in the *reference plot* (small squares) of Fig 7. The filter is a poor one at best, showing severe asymmetry. Further, the bandwidth is slightly less than the desired 3 kHz. This attempt at synthesis is clearly unsuccessful; this is an application requiring the modified circuit.

The scheme that is used to cancel C_0, the parallel crystal C, is to add inductance in parallel with the crystal, creating a parallel resonance with C_0. The inductance is made a bit smaller than needed and a small trimmer capacitance is added, allowing for easy adjustment. The resulting filter is shown in Fig 8. The filter was designed with *X.EXE*, but with a value of 0 pF entered for C_0.

The added inductors are wound on ferrite toroids. This was required for this low frequency filter; the needed 150-µH inductance was not practical with iron-powder toroids. The solid curve in Fig 7 shows the calculated result for this filter. The shape is clean and very symmetrical, something of a rarity with lower-sideband-ladder crystal circuits.

This filter was built and tested with the results shown in Fig 9. The only subtlety encountered was with adjustment of the trimmer capacitors. I eventually shorted all four crystals with pieces of wire. Then the crystals were individually unshorted, the related capacitor was adjusted for a deep null about 20 kHz away from the passband, and the crystal was again short circuited. Only then were the shorts removed. No further adjustment was needed. This method should work well with filters with many more crystals.

It is not surprising that adjustment should be difficult. Fig 10 shows the calculated response for this filter over a wider frequency range of 3.4 to 3.8 MHz. The nulls (calculated!) next to the desired passband are over 150 dB below the desired response! The "wings" that reappear at wider separations from the desired responses can be a problem. The crystal filter will have to be "protected" with a suitable L-C band-pass circuit; one or two resonators will probably do the job.

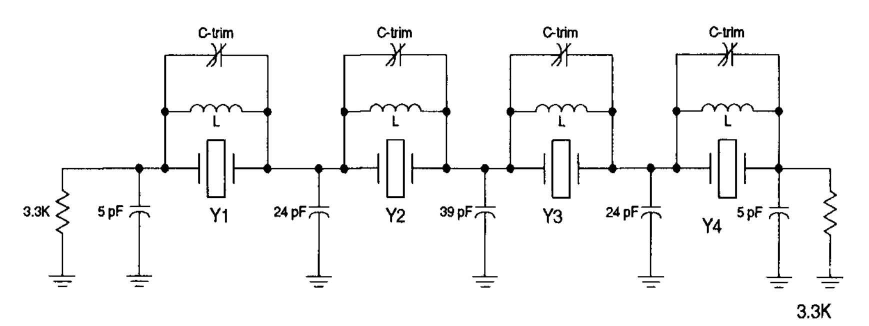


Fig. 8—A low-frequency AM filter with BW = 3.5 kHz.
Y1,2,3,4—3.58-MHz surplus color-burst crystals. L_m = 0.117 H, C_0 = 4 pF
L—151 µH, 48 turns #30 on FT-50-61 Ferrite toroid (Amidon).
C-trim—3-12 pF ceramic trimmer.

The G3UUR Method for Measuring Quartz Crystal Motional Parameters

This simple circuit may be used to measure the motional parameters of fundamental-mode quartz crystals. A crystal to be evaluated is placed in the circuit at Y1 and oscillation is confirmed. The frequency is measured. Then the switch is thrown and the frequency is measured again. Typical values are C_p = 470 pF and C_s = 33 pF. C_m will have the same units as C_s:

If

$$C_s \ll C_p$$

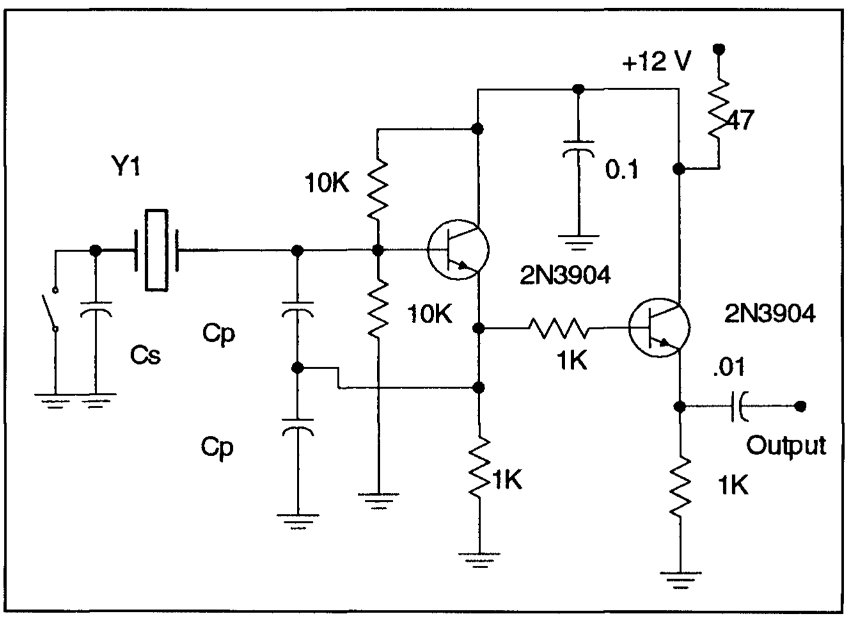
then

$$C_m \approx 2 \cdot C_s \cdot \frac{\Delta F}{F}$$

and

$$L_m = \frac{1}{\omega^2 \cdot C_m}$$

where $\omega = 2\pi F$ with F in Hertz. ΔF is the frequency difference observed when the switch is activated. Example: use capacitors mentioned above, 10-MHz crystal; $F = 1\times10^7$, ΔF = 1609 Hz, to yield L_m = 0.0239 H and C_m = 10.6 fF. (1000 fF = 1 pF.)



Filters with Overtone Crystals

The example filter used readily available low-frequency crystals. The instrumentation used for measurement is that available to most experimentally active amateurs. Similar filters have been built at VHF with overtone crystals. Such filters are not difficult if small inductors are added to each crystal, along with small trimmers. The inductors can now use iron-powder cores. The difficulty with overtone crystals results from the physical nature of the crystals: while motional inductance remains essentially constant, independent of overtone number, the motional capacitance decreases in proportion to the square of the overtone number. Hence, a third-overtone crystal will have one ninth of the motional C that is seen at the fundamental frequency. The coupling capacitors will be reduced by a similar ratio, complicating filter realization. Swept instrumentation will probably be required for the construction of really good overtone filters.

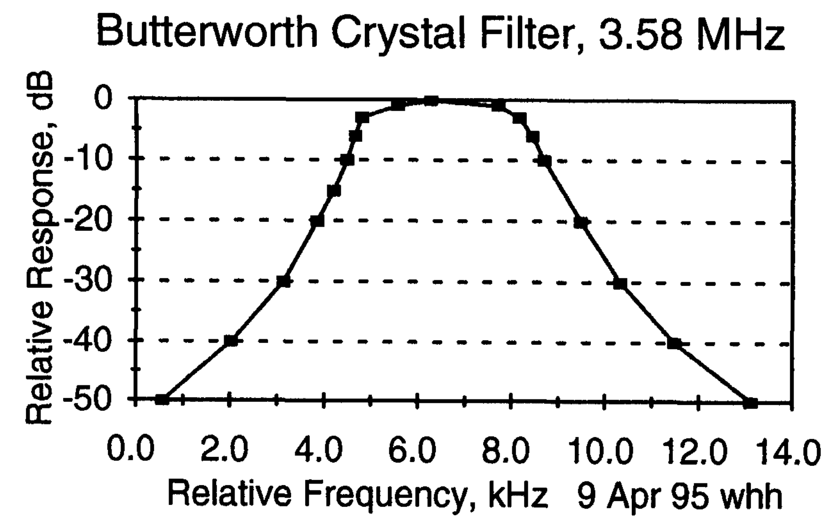


Fig 9—Measured response of an AM-bandwidth crystal filter at 3.58 MHz.

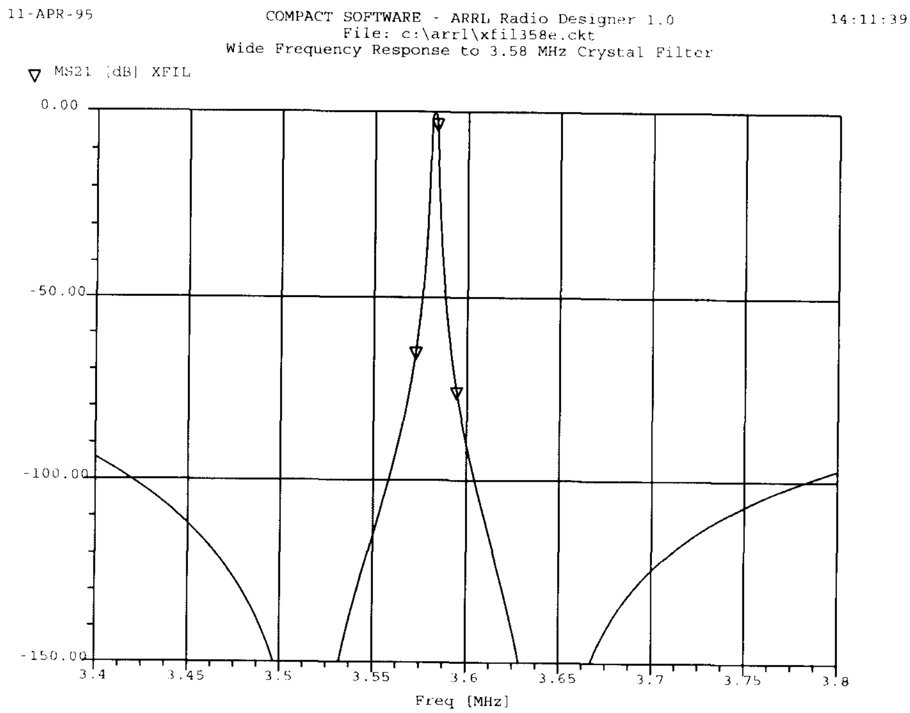


Fig 10

Conclusions

The filters that we have examined are very practical. Of greater significance are the general methods that allow the experimenter to move toward high-order filters with improved selectivity and more constant group delay.

Acknowledgments

Any long-term project like this one is the result of several sessions of midnight pondering and numerous discussions with others with similar interests. In this vein, I want to acknowledge discussions with Larry Lockwood, W7JBY, and Dr. Dave Gordon-Smith, G3UUR. It was Larry who first suggested that I examine reactance slope methods and Dr. Gordon-Smith who confirmed the approach through communications of his earlier works. More recent communications with Bill Carver, K6OLG/7, Jacob Makhinson, N6NWP, and Ulrich Rohde, KA2WEU, have provided the emphasis for the later work.

Notes

[1]Carver, Bill, "High Performance Crystal Filter Design," *Communications Quarterly*, Winter 1993.

[2]Makhinson, Jacob, "Designing and Building High-Performance Crystal Ladder Filters," *QEX*, January 1995.

[3]Hayward, Wes, *Introduction to Radio Frequency Design*, ARRL 1994. See the crystal filter design programs on the disk distributed with the book including *X.EXE* and *MESHTUNE.EXE*.

[4]See Note 3, page 92.

[5]Zverev, *Handbook of Filter Synthesis*, John Wiley and Sons, Inc, 1967.

[6]Bottom, V. E., *Introduction to Quartz Crystal Unit Design*, Van Nostrand Reinhold Co, 1982.

[7]Gordon-Smith, Dave, G3UUR, private correspondence, June 26, 1982.

Digital Communications

By Harold E. Price, NK6K

Spread Spectrum: It's Not Just for Breakfast Any More!

Don't blame me, the title is the work of this month's guest columnist, Steve Bible, N7HPR (srbible@cs.nps.navy.mil). While cruising the net recently, I noticed a sudden bump in the number of times spread spectrum (SS) techniques were mentioned in the amateur digital areas. While *QEX* has discussed SS in the past, we haven't touched on it in this forum. Steve was a frequent cogent contributor, so I asked him to give us some background. Steve enlisted in the Navy in 1977 and became a Data Systems Technician, a repairman of shipboard computer systems. In 1985 he was accepted into the Navy's Enlisted Commissioning Program and attended the University of Utah where he studied computer science. Upon graduation in 1988, he was commissioned an Ensign and entered Nuclear Power School. His subsequent assignment was onboard the USS Georgia, a Trident submarine stationed in Bangor, Washington. Today, Steve is a Lieutenant and is completing a master's degree in computer science at the Naval Postgraduate School in Monterey, California. His areas of interest are digital communication, amateur satellites, VHF/UHF contesting and QRP. His research area closely follows his interest in Amateur Radio. His thesis topic is "Multihop Packet Radio Routing Protocol Using Dynamic Power Control." Steve is also the AMSAT Area Coordinator for the Monterey Bay area. Here's Steve, I'll have some additional comments at the end.

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Steve Spreads It On (Okay, that one was Harold.)

The column title says it all. What was once a communication mode shrouded in secrecy has entered the consumer market in the form of wireless Ethernet links, cordless telephones, Global Position Service (GPS), Personal Communications System (PCS) and digital cellular telephony (CDMA). And what are radio amateurs doing with spread spectrum today? Perhaps very little since AMRAD performed early experiments in amateur spread spectrum in the 1980s and formed the early regulatory rules that govern amateur SS today. In this column I would like to reintroduce the topic of amateur spread-spectrum communication and discuss what it is and how we can experiment with spread spectrum today. Hopefully, this column will prod you into thinking again about spread-spectrum communication when you see that there are several low-cost building blocks available on the market today. Interspersed throughout the column, I'll throw in the Part 97 rules and regulations that deal directly with amateur spread spectrum.

Historical Background

In 1980, the FCC expressed a desire to extend spread-spectrum communication outside of the military-only realm and allow radio amateurs to experiment with spread-spectrum communication. The FCC, in following Title 47, Section 303 of the Code of Federal Regulations (CFR) shall...

...(g) Study new rules for radio, provide for experimental uses of frequencies, and generally encourage the larger and more effective use of radio in the public interest...

What this meant was that a new mode of communication was opening up for experimentation and exploration by amateurs.

In 1980, AMRAD took the lead and forged the beginnings of amateur spread-spectrum experimentation. The results of their experimentation were documented in the *AMRAD Newsletter*, *QEX*, *QST*, and compiled into a single book entitled *The ARRL Spread Spectrum Sourcebook*. This is a good book and is recommended for anyone learning about spread-spectrum communication. Though it is becoming a bit dated by today's standards, what with advances in technology since the late 1980s, it is nonetheless a good guide that also provides a historical perspective into the merging of SS into Amateur Radio. In the sidebar I've included a selected bibliography so you can find other sources of information, ranging from the practical to the theoretical.

What is Spread Spectrum?

A spread-spectrum system is one in which the transmitted signal is spread over a wide frequency band, much wider, in fact, than the minimum bandwidth required to transmit the information being sent.[1] Spread-spectrum communication cannot be said to be an efficient means of utilizing bandwidth. However, it does come into its own when combined with existing narrowband systems occupying the frequency. The spread-spectrum signal, being "spread" over a large bandwidth can coexist with narrowband signals, only adding a slight increase to the noise floor that the narrowband receivers see. As for the spread-spectrum receiver, it does not see the narrowband signals since it is listening to a much wider bandwidth at a prescribed code sequence which I'll explain later.

First, let's introduce five spread-spectrum techniques:

Direct Sequence Systems—Direct sequence is one of the most widely known and used spread-spectrum systems, and it is relatively simple to implement. A narrowband carrier is modulated by a code sequence. The carrier phase of the transmitted signal is abruptly changed in accordance with this code sequence. The code sequence is generated by a pseudorandom generator that has a fixed length. After a given number of bits, the code repeats itself exactly. The speed of the code sequence is called the chipping rate, measured in chips per second (cps). For direct sequence, the amount of spreading is dependent upon the ratio of chips to bits of information (chips per bit). At the receiver, the information is recovered by multiplying the signal with a locally generated replica of the code sequence. See Fig 1.

Frequency Hopping Systems—In frequency hopping systems, the carrier frequency of the transmitter abruptly changes (or hops) in accordance with a pseudorandom code sequence. The order of frequencies selected by the transmitter is dictated by the code sequence. The receiver tracks these changes and produces a constant IF signal. See Fig 2.

Time Hopping Systems—In a time-hopping system, the period and duty cycle of a pulsed RF carrier are varied in a pseudorandom manner under the control of a coded sequence. See Fig 3. Time hopping is often used effectively with frequency hopping to form a hybrid time-division, multiple-access (TDMA) spread spectrum system.

Pulsed FM (Chirp) Systems—A pulsed FM system is a spread-spectrum system in which a RF carrier is modulated with a fixed period and fixed duty-cycle sequence. At the beginning of each transmitted pulse, the carrier frequency is frequency modulated, causing additional spreading of the carrier. The pattern of the frequency modulation will depend upon

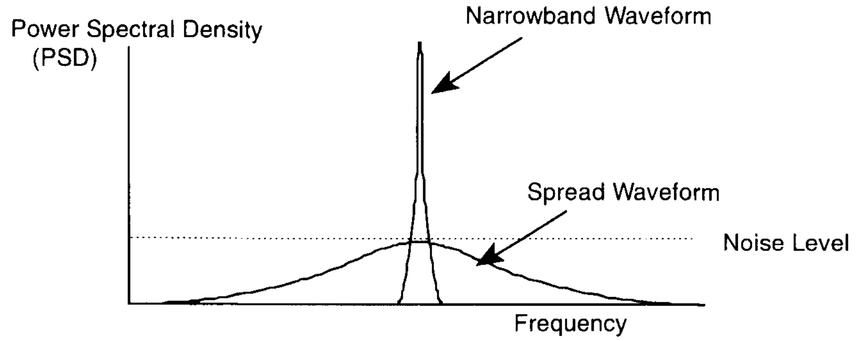

Fig 1—Comparison of a narrowband signal with a direct-sequence spread spectrum signal. The narrowband signal is suppressed when transmitting spread spectrum.

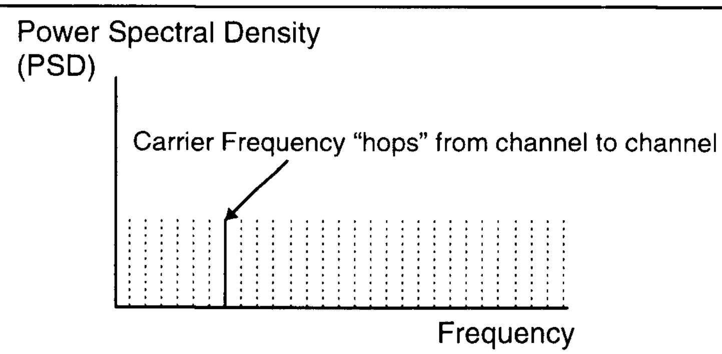

Fig 2—An example of a frequency-hopping spread-spectrum signal.

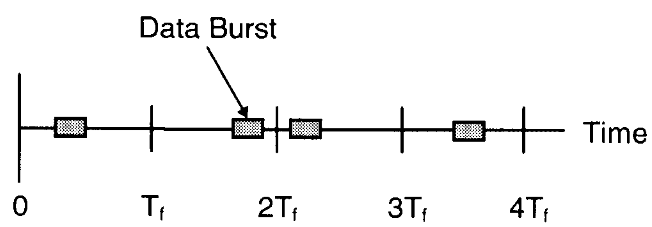

Fig 3—Time hopping spread spectrum. Each burst consists of *k* bits of data, and the exact time each burst is transmitted is determined by a PN sequence.

[1]Notes appear on page 27.

the spreading function that is chosen. In some systems the spreading function is a linear FM chirp sweep, sweeping either up or down in frequency.

Hybrid Systems—Hybrid systems use a combination of spread-spectrum methods in order to benefit from the properties of each system. Two common combinations are direct sequence and frequency hopping. The point of combining the two methods is to capitalize on characteristics that are not available from a single method.

Why Spread Spectrum?

This could easily degenerate into a simple listing of advantages and disadvantages, but spread spectrum has many unique properties that cannot be found in any other modulation technique. As amateurs, we should exploit these properties and search for useful applications. Think of spread spectrum as another useful tool in our repertoire of modulation methods. For completeness, I will list some advantages and disadvantages that you will see for typical spread-spectrum systems. Bear in mind that these come about because of the nature of spread spectrum, not because they are direct attributes.

Advantages:

- Resists intentional and unintentional interference,
- has the ability to eliminate or alleviate the effect of multipath interference,
- can share the same frequency band (overlay) with other users, and
- provides privacy due to the pseudorandom code sequence (code division multiplexing).

Disadvantages:

- Bandwidth inefficient, and
- implementation is somewhat more complex.

Other Properties

There are several unique properties that arise as a result of the pseudorandom code sequence and the wide signal bandwidth that results from spreading. Two of these properties are selective addressing and code-division multiplexing. By assigning a given code to a single receiver or a group of receivers, they may be addressed individually or by group separately from other receivers that are assigned a different code. Interference between groups of receivers can be minimized by choosing codes that have low cross-correlation properties. In this manner, more than one signal can be transmitted at the same time on the same frequency. Selective addressing and code-division multiple access (CDMA) are implemented via these codings.

A second set of properties is low probability of intercept (LPI) and antijamming. When the intelligence of the signal is spread out over several megahertz of spectrum, the resulting power spectrum is also spread out. This results in the transmitted power being spread out over a wide frequency bandwidth and makes detection in the normal sense (without the code), very difficult. Since LPI is not a typical application for amateurs, it would be best to rename this property "reduction of interference." Thus spread spectrum can survive in an adverse environment and coexists with other services in the band. The antijamming property results from the wide bandwidth used to transmit the signal. Recall Shannon's information-rate theorem:

$C = W \log(1 + S/N)$
C = capacity in bits per second
W = bandwidth
S = signal power
N = noise power

where the capacity of a channel is proportional to its bandwidth and the signal-to-noise ratio on the channel. By expanding the bandwidth to several megahertz—even several hundred megahertz—there is more than enough bandwidth to carry the required data rate and have even more to spare to counter the effects of noise. This antijamming quality is usually expressed as "processing gain."

So for the amateur, the properties of code-division multiplexing, coexistence in an adverse environment and processing gain are all excellent reasons to experiment with and find useful applications for spread spectrum in Amateur Radio. Coupled with these reasons, amateurs can also enjoy increased data rates in digital data (packet radio) that cannot be done with conventional amateur or commercial radios due to physical (ie, bandpass filters) and rules restrictions. For example, narrowband systems in the 70-cm band are limited to a maximum data rate of 56 kbit/s and a bandwidth of 100 kHz, there are no such restrictions in the 33-cm band and up.

Perhaps one of the most important reasons to use spread spectrum is its ability to discriminate against multipath interference. A RAKE receiver implementation for direct sequence SS allows individual signal paths to be separately detected and then coherently combined with other paths.[2] This not only tends to prevent fading but also provides a path-diversity effect, resulting in very rugged links in terrestrial mobile communications.[3]

Building Blocks

Spread-spectrum signals are demodulated in two steps:

1) the spectrum-spreading (direct sequence, frequency hopping) modulation is removed, and

2) the signal is demodulated.

The process of despreading a signal is called correlation. The spread-spectrum signal is despread when the proper synchronization of the spreading code between the transmitter and receiver is achieved. Synchronization is the most difficult aspect of the receiver. More time, research, effort and money has gone into the development and improvment of synchronization techniques than into any other area of spread spectrum. The problem of synchronization is further broken down into two parts: initial acquisition and tracking.

There are several methods to solve the synchronization problem. Many of these methods require a great number of discrete components to implement. But perhaps the biggest breakthrough has been from the use of digital signal processing (DSP) and application-specific integrated circuits (ASIC). DSP has provided high-speed mathematical functions that can slice up the spread spectrum signal into many small parts to analyze, synchronize and decorrelate it. ASIC chips drive down the cost by using VLSI technology and creating generic building blocks that can be used in any type of application the designer wishes. With the fast-growing Part 15 and personal communications system (PCS) spread-spectrum market, many ASIC manufacturers have been designing and selling ASIC chips that take care of the most difficult problem in spread spectrum—despreading and synchronization. With a few extra components, the amateur can have a fully functioning spread-spectrum receiver.

One manufacturer of a spread spectrum demodulator ASIC is UNISYS (Unisys Communications Systems Division, DSP Components, Dept 9065, M/S F1F12, 640 North 2200 West, Salt Lake City, UT 84116-2988;

tel: 801-594-4440; fax: 801-594-4127). Their PA-100 performs the functions of despreading and demodulation, carrier-recovery loop (frequency or phase), pseudo noise (PN) code detection, PN code-tracking loop, data synchronization and automatic gain control. It is programmable and offers a wide range of choices of data rates, modulation types, processing gains, PN codes, loop bandwidths, and tracking and acquisition procedures. It is capable of chipping rates up to 32 Mcps and data rates up to 64 Mbit/s. The PA-100 is controlled via a simple 8-bit interface. The chip is a 208-pin plastic Metrix Quad Flat Package (MQFP). The cost of the chip is $167.00 in single quantity and $67.00 in lots of 1000.

Where Does Part 15 Fit into All This?

Many of the spread-spectrum devices on the market today are listed as Part 15 devices. This refers to a device operating under the provisions of Title 47, Section 15.247 of the Code of Federal Regulations (CFR). There are three frequency bands allocated to this service:

- 902 to 928 MHz (26-MHz bandwidth)
- 2400 to 2483.5 MHz (83.5-MHz bandwidth)
- 5725 to 5850 MHz (125-MHz bandwidth)

Operation under this provision of this section is limited to frequency hopping and direct-sequence spread spectrum. No other spreading techniques are permitted. Section 15.247 defines the technical standards under which these systems must operate. For example, the maximum peak output power of the transmitter shall not exceed 1 W. If transmitting antennas of directional gain greater than 6 dBi are used, the power shall be reduced by the amount in dB that the directional gain of the antenna exceeds 6 dBi. This equates to a maximum transmitter EIRP of +6dBW (1 W into a 6-dBi gain antenna).

Part 15 equipment operates on a secondary basis. Users must accept interference from other transmitters operating in the same band and may not cause interference to the primary users in the band. Primary users are government systems such as airborne radiolocation systems that emit a high EIRP, and Industrial, Scientific and Medical (ISM) users. Thus the Part 15 device manufacturer must design a system that will not cause interference with and be able to tolerate, the noisy primary users of the band. And this is where spread-spectrum systems excel because of their low-noise transmissions and ability to operate in an adverse environment.

Amateurs should realize that under the present Part 97 rules and regulations governing amateur spread-spectrum, taking a Part 15 spread-spectrum device and adding an amplifier to it would break the rules. Even though it would be transmitting within the amateur spectrum, it more than likely would not be using one of the specified spreading codes assigned to amateur operation (refer to Section 97.311(d)—SS emission types). However, this should not deter the amateur from using Part 15 devices in their experimentation or use in the amateur service. The device should be monitored to ensure that it remains under the Part 15 regulations so no Part 97 regulations apply. Amateur traffic can flow though Part 15 devices, and they do not require a call sign since they do not require a license. However, the amateur should realize that when the traffic enters the amateur bands, for example, through a gateway, then Part 97 rules begin to apply.

Further Part 97 Rules and Regulations

Any amateur contemplating experimentation with spread spectrum in the amateur bands (excluding Part 15 devices) should become familiar with the present Part 97 rules and regulations governing it. Here are some excerpts that bear emphasizing:

Section 97.119 Station identification

(a)(5) By a CW or phone emission during SS emission transmission on a narrow bandwidth frequency segment. Alternatively, by the changing of one or more parameters of the emission so that a conventional CW or phone emission receiver can be used to determine the station call sign.

Section 97.305 Authorized emission types

Spread spectrum is permitted on the following bands (over the entire band unless otherwise indicated):

UHF: 70 cm (420-450 MHz), 33 cm (902-928 MHz), 23 cm (1240-1300 MHz), 13 cm (2300-2310 and 2390-2450 MHz*).

SHF: 9 cm (3.3-3.5 GHz), 5 cm (5.650-5.925 GHz), 3 cm (10.00-10.50 GHz), 1.2 cm (24.00-24.25 GHz).

EHF: 6 mm (47.0-47.2 GHz), 4 mm (75.5-81.0 GHz), 2.5 mm (119.98-120.02 GHz), 2 mm (142-149 GHz), 1 mm (241-250 GHz), above 300 GHz.

Operation on all of the above bands is on a secondary basis. No amateur station transmitting in these bands shall cause harmful interference to, nor is protected from the operations of the primary service.

(***Note**: Recent rule making has allocated 2390-2400 MHz and 2402-2400 MHz to the amateur community on a primary basis.)

Section 97.311 SS emission types

[**Note**: Sections (a) through (d) set the technical standards for spread-spectrum emissions.]

(e) The station records must document all SS emission transmissions and must be retained for a period of 1 year following the last entry. The station records must include sufficient information to enable the FCC, using the information contained therein, to demodulate all transmissions. The station records must contain at least the following:

(1) A technical description of the transmitted signal;

(2) Pertinent parameters describing the transmitted signal including the frequency or frequencies of operation and, where applicable, the chip rate, the code rate, the spreading function, the transmission protocol(s) including the method of achieving synchronization, and the modulation type;

(3) A general description of the type of information being conveyed, (voice, text, memory dump, facsimile, television, etc.);

(4) The method and, if applicable, the frequency or frequencies used for station identification; and

(5) The date of beginning and the date of ending use of each type of transmitted signal.

(f) When deemed necessary by an EIC to assure compliance with this part, a station licensee must:

(1) Cease SS emission transmissions;

(2) Restrict SS emission transmissions to the extent instructed; and

(3) Maintain a record, convertible to the original information (voice, text, image, etc.) of all spread spectrum communications transmitted.

(g) The transmitter power must not exceed 100 W.

Rules Reform

Needless to say, by today's standards, practices and improvements in technology, the above Part 97 rules and regulations on amateur spread spectrum are extremely restrictive, especially in the case of the few fixed spreading codes dictated by Section 97.311(d)(1). The ARRL is reviewing suggestions from the ARRL Future Systems Committee for changes to these rules and regulations to allow less restriction and freer experimentation.

Getting Around the Rules—Legally

In the meantime, there is a Special Temporary Authority (STA) to allow amateur spread-spectrum experimentation. Under this STA, Section 97.305(c) is waived to the extent that particular amateur stations are authorized to transmit spread-spectrum emissions on frequencies in the 6-m (50 - 54 MHz), 2-m (144 - 148 MHz), and 1.25-m (222 - 225 MHz) bands. Section 97.311(c) is waived for these stations to the extent that the prohibition against hybrid spread-spectrum emissions is lifted; and Section 97.311(d) is waived for these stations to use other spreading codes.

To participate in this STA it is requested that you have a bona fide purpose of experimenting and advancing the art of amateur spread spectrum. Contact Robert Buaas, K6KGS, 20271 Bancroft Circle, Huntington Beach, CA 92646. Please include your name, address, call sign, the expiration date of your license and the details of your experiment. Do include an abstract of the project and a proposed set of goals you are trying to obtain. The information that you collect through your experimentation will be helpful in the advancement of amateur spread spectrum, but it will also be useful for justification of rules changes before the FCC.

Areas to Expand and Research

Typical SS applications such as wireless Ethernet use point-to-point communication. They link two subnets over distances of several miles with external Yagi antennas and less than 1 W of power. Amateurs would rather use the traditional CSMA/CA technique they are familiar with in today's packet radio. However, with the requirement of correlating the spreading code it would require a network node that has multiple receivers to listen in on the channel and detect when an outlying node is trying to communicate with it. Here's where amateur experimentation can advance the art of spread spectrum: by creating a CDMA spread-spectrum packet-radio network. By using the techniques employed by GPS, relatively short codes can be use to minimize receiver acquisition time. These codes would also need to have good cross-correlation properties to minimize multiple-access interference between nodes.

Power control is required to control the reuse of the frequency beyond that allowed by code-division multiplexing. It also behooves us to explore good power control to limit interference and to reduce the power consumption and drain on batteries.

Routing of packets through a network is typically a software issue, but with the ability to do code-division multiplexing, how do we route packets from one subnet to another when they do not use the same code sequence?

Driving cost down has always been a top goal of any designer, and even more so since each amateur is experimenting with their own money. Amateurs tend to be a frugal lot and will find any means available to build a system that costs as little as possible. This spawns innovative and creative methods to achieve this means. Then these means tend to be passed back to the commercial sector to benefit everybody.

CDMA is not the exclusive province of direct-sequence systems; CDMA can also be used with frequency hopping. TDMA is not the exclusive province of narrowband systems; TDMA can also be used with direct sequence or frequency hopping.

This Isn't New

In the 1982 AMRAD letter (reprinted on page 4-11 of the *ARRL Spread Spectrum Handbook*), Hal Feinstein, WB3KDU, wrote:

Spread spectrum has found its way into packet radio. Spread spectrum allows each node to have a unique code which acts as a hard address. Another node in the system can send data to that node by encoding that data with the spread spectrum address for the receiving node. Traffic for other nodes does not interfere because it would have a different code. Among the reasons cited for employing spread spectrum for packet switching are privacy, selected addressing, multipath protection and band sharing. But it is interesting to note that a load is taken off the contention collision approach because now a single frequency is not in contention among the nodes wishing to transmit. The load is divided among the nodes addresses, and each that is interested in sending data to a target node competes for that node only.

This is the CDMA part of SS. This is one of those areas the FCC really wants hams to experiment with. I think the paper has a lot of insight—and it was written over 13 years ago.

PANSAT—A Spread Spectrum Satellite

The Space Systems Academic Group (SSAG) at the Naval Postgraduate School (NPS) in Monterey, California is actively designing and building an amateur satellite named PANSAT (see Fig 4). PANSAT is the acronym for Petite Amateur Navy Satellite. PANSAT is to become a packet digital store-and-forward satellite very similar in capabilities to the existing PACSATs in orbit today. The tentative launch date of PANSAT is late 1996 or early 1997, as a Get Away Special (GAS) payload from the Space Shuttle.

One big difference between today's PACSATs and PANSAT is that PANSAT will use direct-sequence spread spectrum for the communication uplink and downlink.

PANSAT is being designed from the ground up as an amateur satellite. The only military mission of PANSAT is as a training vehicle for the education of military officers in the Space Systems Curricula by the design, fabrication, testing and operation of a low-cost, low-Earth-orbit (LEO), digital communication satellite. One of the engineering objectives of PANSAT includes the evaluation of the performance of spread-spectrum packet-radio communication using the amateur community as the user base.

In order to facilitate the evaluation of spread-spectrum performance, the SSAG is designing a low-cost spread-spectrum modem and RF package to be presented to the amateur community in kit form. The goal is to have the design of the spread-spectrum radio/modem available before the launch of PANSAT to allow amateurs to build them and become operational via terrestrial means. This presents an exciting exchange of technology that provides the amateur a low-cost unit to build and experiment with. As the designs and developments progress they will be presented in the amateur press.

Future and Summary

Now is the time to begin experimenting with spread spectrum communication on a wider scale. Technology has advanced to the point where amateurs can afford to build systems. The build-

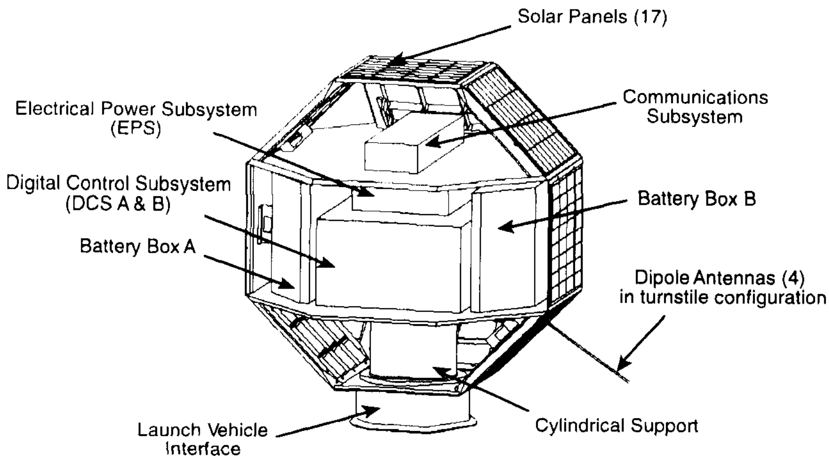

Fig 4—A cut away view of PANSAT, a direct-sequence spread spectrum satellite being designed and built at the Naval Postgraduate School in Monterey, California.

ing blocks are available now in the form of application specific integrated circuits. The recent flood of consumer devices that employ spread spectrum has also driven the price down. In many cases the amateur can either use these devices under their present type acceptance or modify them for amateur operations. However, the amateur should remain aware of the rules and regulations governing the particular device, whether it falls under Part 15 or Part 97 of the FCC Rules and Regulations, and remain within their guidelines. Any amateurs who wish to expand beyond the present Part 97 rules in bona fide experimentation are encouraged to join in the Special Temporary Authority.

Spread-spectrum systems exhibit unique qualities that cannot be obtained from conventional narrowband systems. There are many research avenues exploring these unique qualities. Amateurs in their inherent pioneering nature can and will find new and novel applications for spread-spectrum communication that the commercial sector may not even think of. And due to the frugal propensity of the amateur, they will certainly find the least expensive way to implement it, thus driving down the cost.

Amateurs should realize that there is plenty of room to explore spread-spectrum techniques. All that remains now is to pick up a few good books on the subject and warm up the soldering iron. And as you progress upon this road less traveled, make sure you take notes along the way. Then share your discoveries with your fellow amateur to help all of us expand the horizon with this exciting mode of communication called spread spectrum. It is no longer shrouded in secrecy and it's not just for breakfast anymore!

Web Crawling

Here are two web pages of interest. I've started a general amateur radio SS page, http://www.sp.nps.navy.mil/ss. See also the PANSAT page at http://www.sp.nps.navy.mil/pansat/pansat.html

TANSTAAFL from NK6K

Harold here with a final word. Way back in 1983, we pitched packet as a way of sharing a narrowband channel in the time domain. Compared to the then-current technology, ASR 33 or glass TTY and keyboard, with real-time, hand-typed text, we could place several users on a channel where only one fit before. Now, of course, there are too many users, too many bytes and too few channels (at least for the RF modems we have). SS, with its noninterference properties, also promises free channel sharing. Is it a free lunch? What is the limit on channel reuse, and on sharing between narrowband and SS users? The question seems to be akin to, "How many angels can dance on the head of a pin?" As with most real systems, modeling can only take you so far, especially with the number of variables that would be involved in an amateur SS system. Here is a case where hams can again add to the practical, rather than theoretical, knowledge pool. We have to take advantage of the spectrum we have, the STA we have, and actually put some hardware on the air and see what happens. Let's actually experiment, not ignore SS because it might interfere with other modes. Let's find out if it does, and see what can be tweaked to avoid it. We have the relative luxury of being able to try new things on the air, without proving we can generate revenue from them before spectrum is allocated. Get on with it.

Selected Bibliograhy

Books

Extensive research-oriented analysis:

M.K. Simon, J. Omura, R. Scholtz, and K. Levitt, *Spread Spectrum Communications Vol I, II, III.* Rockville, MD. Computer Science Press, 1985.

Intermediate level:

J.K. Holmes, *Coherent Spread Spectrum Systems*, New York, NY. Wiley Interscience, 1982.

D.J. Torrieri, *Principles of Secure Communication Systems*, Boston. Artech House, 1985.

Introductory to intermediate levels:

G.R. Cooper and C.D. McGillem, *Modern Communications and Spread Spectrum*, New York, McGraw-Hill, 1986.

R.E. Ziemer and R.L. Peterson, *Digital Communications and Spread Spectrum Systems*, New York, Macmillan, 1985.

R.E. Ziemer and R.L. Peterson, *Introduction to Digital Communications*, New York, Macmillan, 1985.

Practical

R.C. Dixon, *Spread Spectrum Systems*, John-Wiley & Sons, 1984.

Journals

There have been several special issues of IEEE publications that are devoted to spread spectrum systems. *IEEE Transactions on Communications*: August 1977 and May 1982. *IEEE Journal of Selected Areas in Communications*: May 1990, June 1990, and May 1992.

Notes

[1]Dixon, R.C., *Spread Spectrum Systems*, John-Wiley & Sons, 1984, p 7.

[2]RAKE is not an acronym. It is called RAKE because the filter arrangement of the receiver is like a garden rake.

[3]Gilhousen, K., Qualcomm Inc, USENET newsgroup discussion.

Upcoming Technical Conferences

The Central States VHF Society Conference

July 27-30, 1995, Sheraton Colorado Springs Hotel, 2886 South Circle Drive, Colorado Springs, CO 80906, 1-800-325-3535 or (719) 576-5900.

Contact: Lauren Libby, KXØO, President at (719) 593-9861, email: 75151.2442@compuserve.com or Hal Bergeson, WØMXY, Vice President and Program Chairman, 809 East Vermijo Avenue, Colorado Springs, CO 80903, (719) 471-0238, email: bergeson@ppcc.colorado. edu.

Events: Thursday night, side trip to "Flying W Ranch" for western music and beef barbeque. Friday and Saturday, technical and operating talks for all, beginners to the experienced ham. There will also be a special "Young People's" VHF program for those under 21. Saturday night's banquet features a presentation by Arnie Coro, CO2KK, on VHF opration in Cuba and a taste of Cuban culture. Sunday morning, SMIRK will sponsor a breakfast get-together.

Special Programs: Several "wives and kids" activities are being planned, including a "British High Tea" and a tour of the Glen Eyrie castle, a historical landmark in Colorado Springs. Shopping, galleries, local landmarks, unique shops and eating places are also part of this program.

Special Note: Colorado Springs is a summer tourist town. The town and surrounding area is a great place for a summer vacation. However, that presents special problems when it comes to booking room reservations. The Central States VHF Conference has a large block of rooms reserved for the Conference. Please take advantage of that as you will find it difficult to find rooms otherwise. Get your reservations in early; when our block of rooms is gone, you may have trouble finding another room.

Eastern VHF/UHF Society Conference

August 25-27, 1995, Quality Inn & Conference Center, 51 Hartford Turnpike, Vernon, CT 06066.

Contact: San Hilinski, KA1ZE, Chairman, Pilgrim Drive, Tolland, CT 06084, tel: (W) (203) 649-3258, (H) (203) 872-6197; Ron Klimas, WZ1V (address below) or Rae Bristol, K1LXD (address below).

Call for papers: Deadline for camera-ready papers **is July 21, 1995**. Papers should be sent to Ron Klimas, WZ1V, Conference Secretary & Proceedings Editor, 458 Allentown Rd, Bristol, CT 06010; tel: (W) (203) 768-4758 (H) (203) 589-0528 (BBS) (203) 768-4758 weeknites/weekends;Internet: klimas@uhavax.hartford.edu.

Events: Friday, check-in and hospi-tally room activities; Saturday, registration, formal talks and bandsessions; Saturday evening, banquet; Sunday, VHF-SHF Swap Meet and antenna measuring.

Registration: Registration at the door will be $25. Preconference registration, before August 20, is $20. Sunday-only registration is $5. Registration fees should be sent to: Rae Bristol, K1LXD, 328 Mark Drive, Coventry, CT 06238, tel: (203) 742-8650

Reservations: The Quality Inn is offering special rates of $51.50 per night, single or double. Call Lori Torizer at 1-800-235-4667. Be sure to mention the Eastern VHF/UHF Society to receive the special rate.

Other activities: A shopping center, $2 movie theater and amusement area are on site.

1995 ARRL Digital Communications Conference

September 8-10, 1995, La Quinta Conference Center, Arlington, TX—just minutes from Dallas/Fort Worth Airport. Co-hosted by Tucson Amateur Packet Radio (TAPR) and the Texas Packet Radio Society.

For more information contact the TAPR office at 8987-309 E. Tanque Verde Road #337, Tucson, AZ 85749-9399, tel: (817) 383-0000; fax: (817) 566-2544; Internet: tapr@ tapr.org

Call for papers: Deadline for receipt of camera-ready papers is **July 21, 1995**. Contact Maty Weinberg at ARRL HQ (tel: (203) 666-1541; fax: (203) 665-7531; Internet: lweinberg @arrl.org) for infomation on submitting papers.

The 1995 AMSAT Annual Meeting and Space Symposium

October 6-8, 1995, in Orlando Florida. For more information contact: Bob Walker, 6601 SW 16th Street, Plantation, FL 33317, (305) 792-7015, email: n4cu@amsat.org.

Call for papers: Those interested in submitting papers are asked to send in a summary by July 1, 1995. The deadline for camera-ready copy is August 12. Inquiries should be sent to Bob Walker at the above address.

Microwave Update 95

October 26-28, La Quinta Inn, Arlington, TX.

For more information contact: Al Ward, WB5LUA, 2306 Forest Grove Estates Road, Allen, TX 75002 or Kent Britain, WA5VJB, 1626 Vineyard, Grand Prairie, TX 75052-1405.