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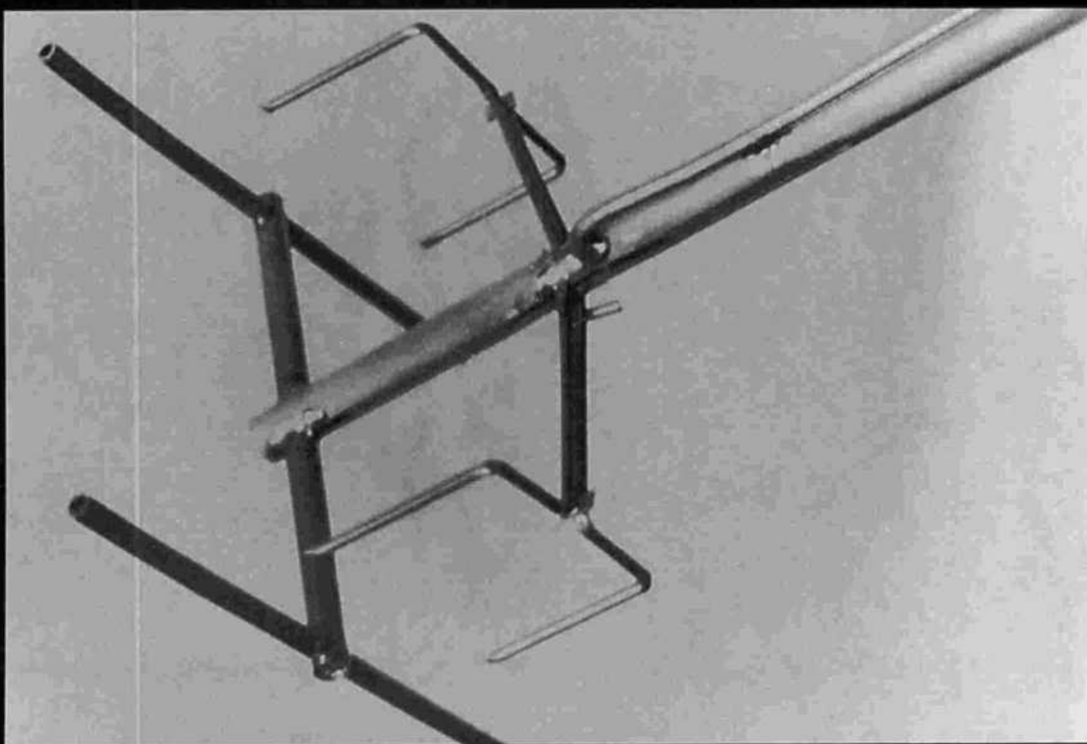
ARRL Experimenter's Exchange

February 1996



A Feed for Deep Parabolic Dishes

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QEX

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David Sumner, K1ZZ
Publisher

Jon Bloom, KE3Z
Editor

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Joseph Costa
Technical Illustrator

Joe Shea
Production Assistant

Advertising Information Contact:

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

Circulation Department

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

Offices

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

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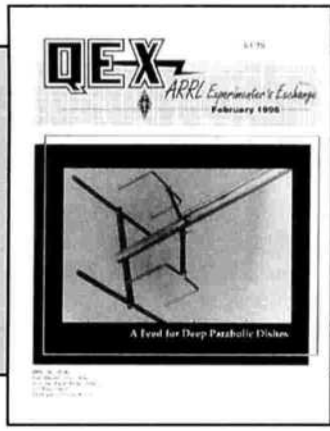

About the Cover

Horn feeds are easy, but they don't always work well on deep dishes. W7PUA's dipole-reflector feed does, though.

ISSUE NO. 168

Features

Columns

February 1996 QEX Advertising Index

THE AMERICAN RADIO RELAY LEAGUE

Purpose of *QEX*:

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

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and correspondence for publication in *QEX* should be marked: Editor, *QEX*.

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*.

Empirically Speaking

Electronic *QEX*: Who Cares?

Almost nobody, apparently. In last December's *QEX*, we asked whether readers were ready for *QEX* in electronic form. We expected a flood of email. Boy, were we in for a rude awakening!

We received, via email and postal mail, a total of two dozen responses from readers. *Two dozen!* Obviously, no matter what those 24 people had to say there wasn't going to be a mandate to produce *QEX* in an electronic form.

In fact, a number of those who wrote did so to express concern that the printed *QEX* was going to go away. That was never part of our thinking, and we apologize if we gave the wrong impression. All we were proposing was the *addition* of an electronic version. So those concerns weren't really valid. Now, it would be one thing if those messages had been received along with hundreds of responses of: "Yes, give me an electronic version!" They weren't.

We received a few responses telling us why the reader would *never* want an electronic *QEX*. The arguments were cogent, but since, again, we weren't about to ram an electronic *QEX* down anyone's throat, these views didn't really tell us whether we should spend the time and effort needed to do an electronic version.

What did tell us what we wanted to know was the fact that we received only about a dozen positive responses, ranging from a fervent desire for *QEX* via the Internet to a lukewarm, "I guess so—why not?"

Well, we can't afford to spend what it would take to make *QEX* available in electronic form for those dozen people so, sorry folks, you'll just have to make do with print.

We'll admit to some disappointment. We were looking forward to the challenge (not much of one, really) of making *QEX* fit for the 'Net. It seems to be the coming thing—or maybe all the current Internet hype is just that. Is the Internet going to be the "CB radio" of the 1990s, with a boom followed by a bust? We rather doubt it, but it's clear now that the medium is at least part of the message, and the medium of choice for *QEX* readers is still the printed page.

We're still thinking of doing *something* electronic, such as "electronicizing" the occasional *QEX* article to use as a teaser on the ARRL Web site. (Of course, we do often offer electronic files that support our articles—*QEX* is not quite wholly in the print medium.) But for now, any large-scale electronic *QEX* project just isn't in the cards.

For those readers who are happy with the print medium, we hope you'll continue to enjoy *QEX* in that form for a long time to come. We'll be here, in the form you want. And if you change your mind about an electronic *QEX*, please let us know.

This Month in *QEX*

Want to feed a deep dish at 1296 MHz? Having trouble finding a feed that makes best use of the dish area? You're not alone! Horns become difficult when the dish is deep, but try one of the "Dipole-Reflector Parabolic Dish Feeds for *f/D* of 0.24 to 0.4," by Bob Larkin, W7PUA. It just may solve your feed need.

"A Homebrew 2-m Repeater" can cost very little. Even better is one that requires no radio modifications, and that's what John Thomas Crago, KB8DAN, and the West Virginia Institute of Technology ARC have devised. It uses simple, reproducible circuitry and isn't limited to use on 2 m.

A clean VFO isn't too hard to build, but a *stable* VFO, that's something else. You could spend days selecting temperature-compensating components to reduce the drift, but Klaas Spaargaren, PAØKSB, shows an electronic method for "Frequency Stabilization of L-C Oscillators" that can tame that drifting VFO.

If you're experimenting using DSP, you'll want to add "Correlation of Sampled Signals" to your DSP toolbox. Correlation is a powerful component of many DSP algorithms, and Jon Bloom, KE3Z, shows how it works—and how to work it.

Get ready for those 1996 conferences! "Upcoming Technical Conferences" gives you the details about the Central States VHF Society Conference and the TAPR/ARRL Digital Communications Conference.—*KE3Z, email: jbloom@arrl.org.*

Dipole-Reflector Parabolic Dish Feeds for f/D of 0.24 to 0.4

Illuminating a deep dish can be a problem, but here's a feed that does the job.

By Bob Larkin, W7PUA

Deep parabolic reflectors with focal-length-to-diameter ratio (*f/D*) less than 0.4 have been widely manufactured due to their ability to have low sidelobes. Typically, though, the price of improved sidelobe response has been a loss in forward gain. I have a 48-inch dish with a 0.375 *f/D* that is rugged and has a surface smooth enough for use up through 10 GHz. For the microwave frequencies it is practical to get adequate gain performance using a horn type of feed along with a flange to reduce the feed gain.[1,2] At lower frequencies, such as 1296 MHz, the size of the horn has become quite large and aperature blockage becomes a problem. Likewise, the physical size of the horns at lower frequencies make them substantial wind catchers and awkward for portable operation.

Feeds using a dipole with reflector (a two element Yagi) have been used for many years.[3] They suffer from poor front-to-back ratio and poor balance between the patterns in the azimuth and elevation planes. Still, they are small and easily built, and I have usually ended up using this configuration. When I wanted to improve my feed for 1296 MHz, the dipole-reflector feed seemed like a good starting place. This article describes improvements that can be made to the dipole-reflector feed to make it a good deep-dish feed for linear polarization.

The basic approach is to use wire analysis with *EZNEC* to improve the feed pattern.[4] Then a computer program was written to analyze the efficiency with which the feed is able to illuminate the dish surface. This allows you to determine the range of *f/D* ratios for which a particular feed is applicable. This latter technique is not limited to the dipole-reflector feeds but can be used with any feed for which a calculated or measured pattern is available. After the design was complete, measurements were made, first of the feed and then of the entire parabolic dish antenna, to confirm the predicted performance.

Evolution of the Double-Handlebar Dipole-Reflector Feed

Dipole Feeds

Probably the simplest and certainly the oldest feed is the half-wave dipole.

[1]Notes appear on page 9.

2982 N. W. Acacia Place
Corvallis, OR 97330
email: boblark@proaxis.com

Heinrich Hertz used such a feed in 1888 to provide a one-way QSO on about 432 MHz! Two problems exist for the dipole feed. It has no front-to-back ratio and therefore loses at least half of its efficiency in the energy that cannot illuminate the dish. Additionally, the directivities in the E- and H-planes are not the same. There is no directivity in the H-plane (this is the elevation plane perpendicular to a horizontal dipole) whereas the 10-dB beamwidth is about 135° degrees in the E-plane (the azimuth plane for a horizontal dipole).

Dipole with Reflector Feed

The simple addition of a reflector increases the front-to-back ratio of the dipole feed to about 10 dB. The remaining reverse radiation lowers the efficiency of the dish antenna by about 10% and raises the sidelobes considerably. Fig 1 shows the patterns in the two planes. The E-plane 10-dB beamwidth is 125°, which is too narrow for low-*f/D* dishes, but the H-plane beamwidth is much too wide at 240°. In spite of these shortcomings this feed is very simple to construct, lightweight, and thus widely used.

Handlebar Dipole with Reflector Feed

The dipole with reflector was the starting point for this study. Some experimentation with *EZNEC* showed that it was possible to broaden the E-plane pattern and to greatly increase the front-to-back ratio by shortening the length of the dipole to 20 to 40% of a half wave. This decreases the directivity in the E-plane because there is less in-phase radiating length that is concentrating energy toward the dish center. This same effect helps front-to-back ratio since the problem with the dipole and reflector is a lack of current in the reflector element, which gives poor cancellation of the backward wave. By decreasing the perpendicular radiation of the dipole it is possible for better cancellation to occur, with a 20-dB front-to-back ratio being achievable.

The short dipole presents a difficult impedance matching problem. This is resolved by adding rods on the ends of the short dipole that are folded back towards the reflector as shown in Fig 2A. This "handlebar" configuration brings the impedance at the center back to the range of an ordinary dipole, but since the two added rods are close together and have opposite currents they are not major sources of radiation. The resulting pattern (Fig 2B) shows that the E-plane pattern now has a wider beamwidth, and the front-to-back ratio is up to about 23 dB, while the excessive H-plane beamwidth remains.

This handlebar configuration can be a simple modification to existing dipole-reflector feeds. It is well worth doing and produces only minor changes in the antenna impedance.

Double-Handlebar Dipole with Reflector Feed

At this point the handlebar feed is looking fine in the E-plane for feeding the deep, low-*f/D* dish. The remaining problem of excessive beamwidth in the H-plane is easily solved by feeding a pair of the handlebar dipoles in phase. The spacing between the feeds can be adjusted to vary the H-plane beam-width over a wide range. Beamwidths at the 10-dB points from about 125° to 240° can be chosen, making this feed suitable for *f/D* from 0.24 through 0.45.

This double-feed configuration has obvious family resemblances to the EIA standard-gain antenna feed.[5] However, the wavelength-square reflector of the EIA antenna gives it higher gain and greater aperture blockage. As will be

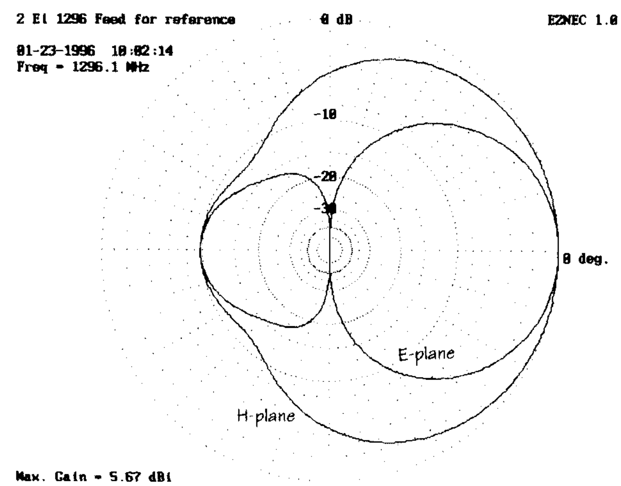

Fig 1—Calculated E- and H-plane patterns for a half-wave dipole with a 4.5-inch long reflector spaced at 2.2 inches.

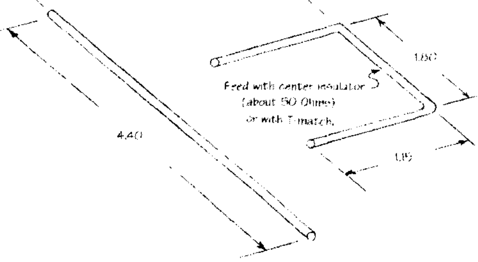

Fig 2A—Handlebar dipole with reflector geometry.

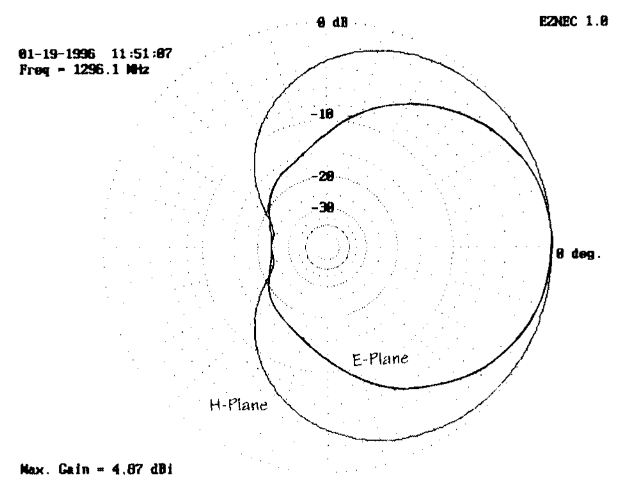

Fig 2B—Calculated E- and H-plane patterns for the antenna shown in Fig 2A.

seen, the EIA feed is not suitable for *f/D* less than 0.4.

A Feed for *f/D*=0.375

Several parameters need to be optimized to achieve the desired beamwidths of the feed pattern. *EZNEC* was used for this, and the following parameters were varied:

- the length of the short dipole section parallel to the reflector,
- the distance between the short dipole and the reflector,
- the length of the rods at the ends of the short dipole, and
- the spacing between the two dipole-reflector antennas.

The first three parameters vary the E-plane beamwidth and the feed impedance. The last parameter varies the H-plane beamwidth, with a small effect on feed impedance.

Using this procedure, I picked the configuration of Fig 3 called Type A as being most suitable for my 0.375 dish. Fig 4 shows the resulting feed patterns. The E-plane pattern is –8 dB at the dish edge (67°), and the H-plane pattern is –10 dB. Fig 5 shows that the efficiency of feeding various dish depths peaks between 0.375 and 0.4. *f/D* ratios between 0.32 and 0.48 all show gain efficiencies above 0.7, quite acceptable from a gain point of view. However, it is undesireable to use this feed with *f/D* greater than the 0.4 peak because sidelobes of the dish pattern will start to rise rapidly. It is better to create a new design with a narrower beam.

Matching the Feed Characteristics to the Dish *f/D*

A recent article by Paul Wade, N1BWT, provides an excellent background against which to evaluate new feed systems.[2] He shows that the feed must be matched to the particular dish *f/D*. If the feed pattern is too wide, the spillover efficiency is poor, reducing the gain and hurting the receiving performance by causing poor overall sidelobes. On the other hand, if the feed pattern is too narrow, the dish will have poor illumination at the edges and as a result have low gain. Both of these effects are shown in efficiency curves such as Fig 5. (These curves are generated from the *EZNEC* calculated patterns, but they can come from the measured feed patterns if those are available.)

Efficiency calculations can be used to replace the usual rules-of-thumb, such as having the feed pattern down 10 dB at the dish edge, although nothing in the results done so far have suggested the rules-of-thumb are not close to correct. The biggest advantage is that it allows one to inspect the sensitivity of not operating at the best point, or for understanding the tradeoffs between best gain and good sidelobe performance. One should also note that there are no corrections needed for the additional path loss to the edge of the dish. This is all included implicitly in the efficiency calculation.

The efficiency calculations were made using equations derived by several workers during the 1940s and given in a paper by C. C. Cutler.[6] Cutler's paper is excellent background reading for anyone working with dish

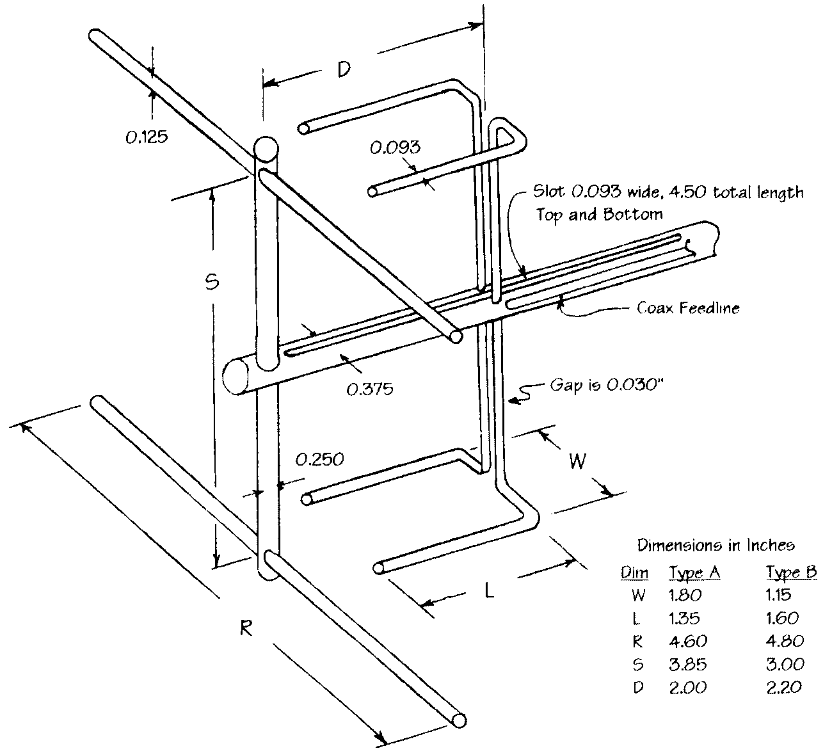

Fig 3—Double-handlebar feed.

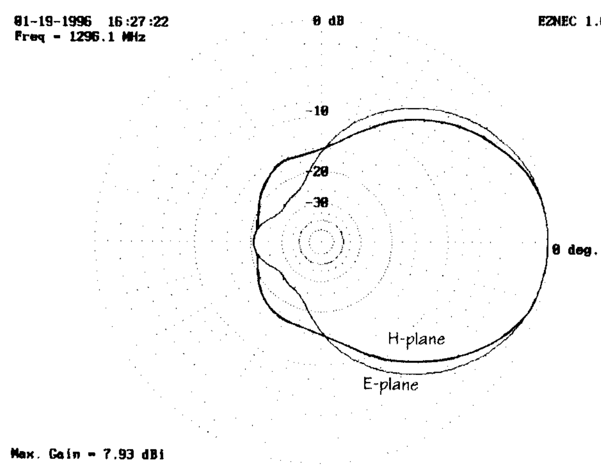

Fig 4—Calculated E- and H-plane patterns for the Type-A double-handlebar feed, suitable for *f/D* from 0.32 to 0.4.

feeds. It includes some mathematical treatment of the subject but a full understanding of the math is not required to read the paper. DISHFD1.BAS (see "Calculation of Parabolic Reflector Aperture Efficiency") is a BASIC program for calculating dish feed efficiency from a file with the feed pattern.

You may have noted that the efficiencies shown in the curves tend to be greater than 0.7 (70%). As pointed out by N1BWT, most experimenters report that 0.5 to 0.55 efficiency is difficult to achieve, and larger values are rare. The causes, which are discussed in Cutler's paper are:

- diffraction effects at the dish edge creating higher sidelobes,
- polarization changes in the reflected wave due to the dish geometry,
- obstruction of the reflected wave by the feed, and
- phase errors due to both the feed and the dish surface.

Beyond these effects are ohmic losses in the feed and, to some extent, the reflector, as well as impedance matching losses between the transmitter or receiver and the feed. Obviously, we are not going to achieve the efficiency that is predicted by looking only at spillover and illumination uniformity.

Another general observation about the efficiency curves is that they are typically quite broad. For instance, the peak of Fig 5 is 0.73, so a half dB of gain reduction would be 0.89 × 0.73 = 0.65. This level is obtainable for *f/D* ranging from 0.28 to more than 0.5. As noted before, an *f/D* greater than 0.4 or so should not be used since the sidelobe performance would be detrimental for reception. This leaves a fair range of *f/D*, from 0.28 to 0.4, as being ideal for this feed. Interestingly, these are the dish depths that are generally felt to be difficult to feed. Maybe the low-*f/D* dish has more value than generally believed, not only because we are able to develop better feeds for these deep dishes, but also because the under-illumination of a dish is not very detrimental to gain performance.

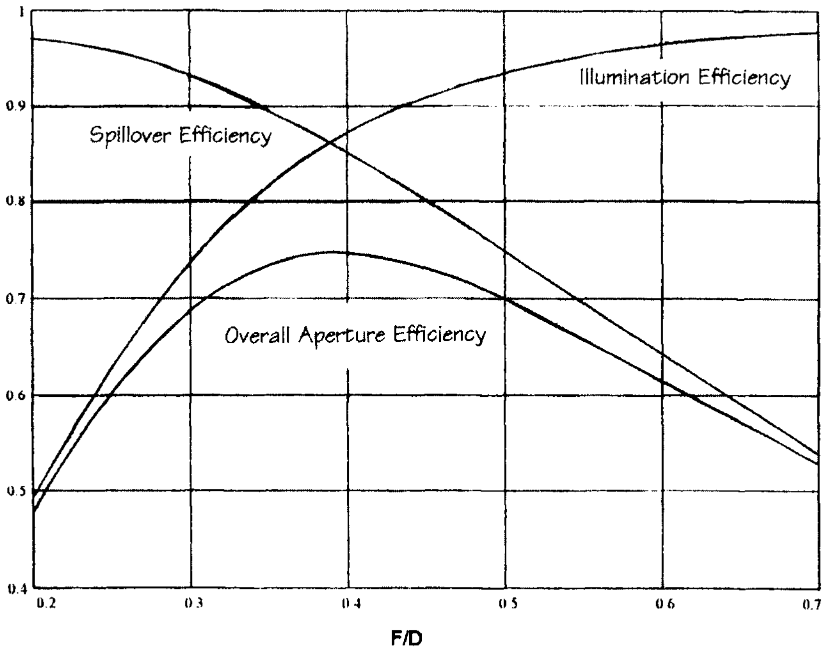

Fig 5—Aperture efficiency of the Type-A double-handlebar feed showing illumination and spillover efficiencies.

A Feed for Dishes with *f/D*=0.24 to 0.3

The success with the Type-A feed for *f/D*=0.375 suggested exploration of a feed for deeper dishes. This resulted in the Type-B feed of Fig 3. The dipole portion that parallels the reflector was shortened further relative to the Type-A feed, and the dipole-reflector pairs were moved closer. Fig 6 shows the resulting beam patterns, and Fig 7 shows that the efficiency now peaks at *f/D*=0.3. The efficiency is is above 0.7 from *f/D*=0.24 to 0.36 but again, sidelobe considerations dictate that *f/D* above 0.3 should use a higher-gain feed such as the Type A.

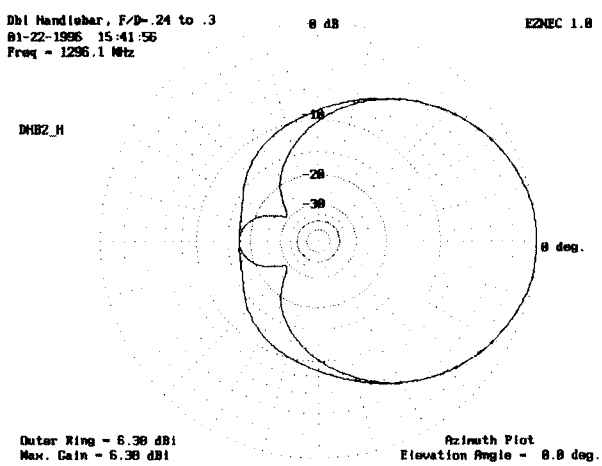

Fig 6—Calculated E- and H-plane patterns for the Type-B double-handlebar feed, suitable for *f/D* from 0.24 to 0.3.

Construction of the Feeds

Fig 3 and the photograph, Fig 8, show the construction of the double-handlebar feed. The reflectors are supported conventionally at the end of a boom. The driven elements are made from 3/32-inch copper tubing available in hobby stores. The support rods for the driven elements are the same piece of tubing that forms the transmission line. Tabs, about 0.15 × 0.2-inch of 0.031-inch thick teflon-glass printed circuit board, were soldered at the feedpoints for additional support for the transmission line and to set the spacing between the support rods at 0.031-inch. The support rods are slightly bent near the boom to make the spacing reasonably uniform. The boom is split longitudinally for a half wave, and the transmission line between the two driven elements is soldered to the boom at the center of the slot. This provides a solid mechanical support along with an isolated RF connection. The 50-Ω feed line is brought up one side of the boom, and the center conductor goes through a hole in the side of the boom to connect to the opposite side.

The slot in the boom needs to be a quarter wavelength long to electrically isolate the feed. However, the short dipoles are closer than this to the reflectors. The difference is made up by bending the support rods at a slight angle towards the reflectors until the dimensions are as shown in Fig 3. You should try to end up with the reflector directly behind the dipoles, but the dimensions are not critical.

In order to use readily available materials, the $^{3}/_{8}$-inch boom was made from two pieces of brass hobby tubing with $^{3}/_{8}$- and $^{11}/_{32}$-inch outside diameters. The wall thickness is designed to allow these pieces to telescope together, forming a stronger piece of tubing with a $^{1}/_{32}$-inch wall. The tubing is soldered together at the edges and also once or twice as the longitudinal slot is being cut. Soldering the support rods at the center of the slot will ensure that the inner and outer tubing of the boom are electrically connected.

A piece of 0.141-inch 50-Ω transmission line is soldered to the outside of the boom up to the feedpoint, where a $^{5}/_{32}$-inch hole has been drilled in the side of the boom. A second $^{5}/_{32}$-inch hole is drilled across from the first to allow easy soldering of the 50-Ω line across the boom at the midpoint of the slot. The other end of the 50-Ω line is then terminated in an RF connector.

Note that it is also possible to feed the antenna through the center of the boom, but this requires more metal crafting than the arrangement shown.[6] The arrangement of soldering the feed line to the outside of the boom works well.

When any feed is adjusted for best

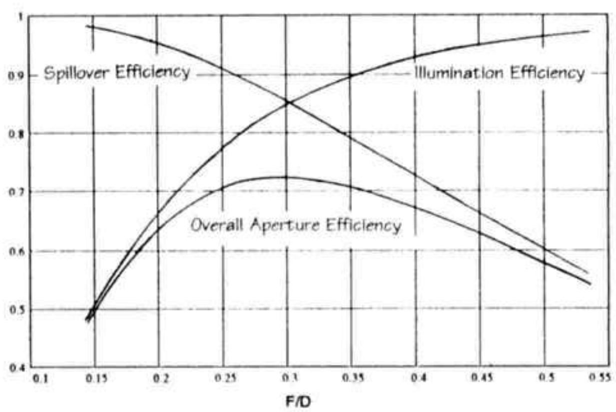

Fig 7—Aperture efficiency of the Type-B double-handlebar feed.

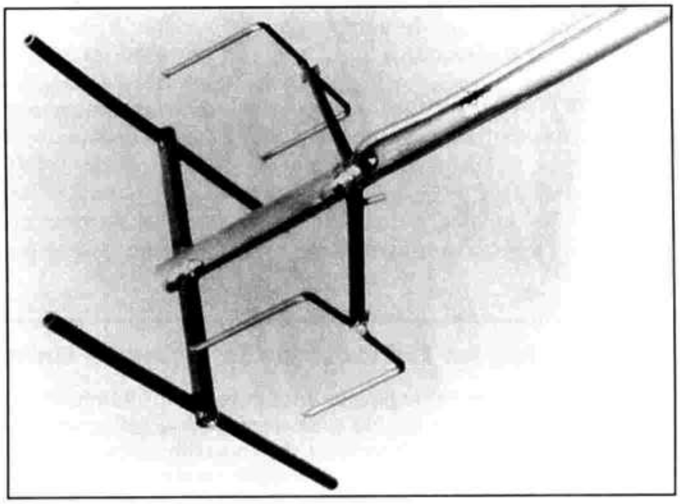

Fig 8—A Type-A double-handlebar feed.

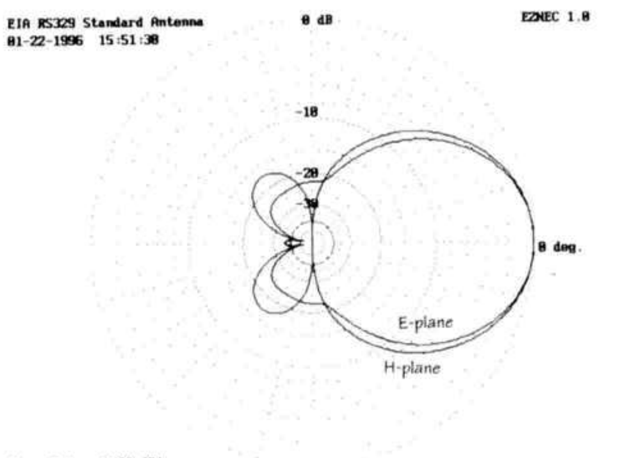

Fig 9—Calculated patterns for the EIA standard-gain antenna.

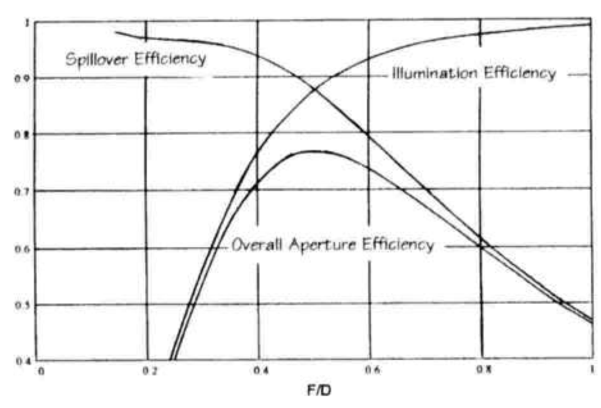

Fig 10—Aperture efficiency of the EIA antenna when used as a parabolic dish feed.

match it should be installed on the dish at the final focal position. This is due to the reflection from the dish that comes back and is caught by the *blockage* aperture of the feed with resulting changes in match. The phase center of the double-handlebar feed is roughly halfway between the short dipoles and the reflectors.

A match of about 2.0 VSWR is achieved with the arrangement as described above. This was brought down to 1.3 by two capacitive stubs formed by 0.5-inch lengths of 50-Ω transmission line connected across the feed point. These stubs are soldered to the outside of the boom on the opposite side from the feed line, with one above the boom and one below to maintain symmetry between the feed halves. It may also be possible to improve the match by changing the length of the driven elements. This is acceptable as long as both driven elements are adjusted the same amount.

Measurements

Once the feed was matched, it was time to see if the measured performance agreed with the calculated performance. After all, no one has ever had a QSO with a computer-generated antenna pattern. The first step was to measure the feed by itself. This is practical since the aperture size of the feed is less than a wavelength across, and the measuring range need only be about two wavelengths, or 18 inches at 1296 MHz.[5,7] Elevating this short range 5 or 6 feet in the air minimizes reflection problems. The feed lines are arranged behind the antennas, and several ferrite beads are placed on the feed lines near the antennas to minimize conducted radiation problems. Measurements down 20 dB are reasonably free of reflection errors, and measurement of absolute gain to better than a dB is possible.

Using this short range, it was possible to confirm the calculated patterns. The high backlobe with the dipole and reflector feed was easily seen, as was the supression achieved with the new feeds.

The feeds were now ready for measurements using the dish. Measurements were made on a *slant* range with the sense antenna at 25-feet height and a distance of 44 feet. The dish center was about 5 feet above the ground. This geometry kept the radiation from the antenna substantially off the ground, minimizing problems from reflections. The sense antenna was a dipole supported about 8 feet away from a metal tower. There were reflections from the tower, but by sweeping the frequency from 1200 to 1400 MHz it was possible to measure the reflection level as –16.5 dB. The error from the reflection was removed from the measurements by measuring the maximum and minimum values of the transmission loss between the antennas (see "Correcting for Reflections in Antenna Gain Measurements"). When the feeds were changed on the dish, it was observed that the transmission loss changed essentially the same amount at all frequencies, indicating that the level changes with frequency were due to the tower reflections. An HP8714B network analyzer was used for the measurements, which allowed easy calibration of the connecting cables.

Using this set-up and the 48-inch f/D=0.375 dish, the following gains were measured along with the calculated aperture efficiencies:

Dipole with reflector feed
20.2 dBi 0.39 efficiency
Double-Handlebar Type A
21.3 dBi 0.50 efficiency
Double-Handlebar Type B
20.9 dBi 0.45 efficiency

Correcting for Reflections in Antenna Gain Measurements

When a single reflection is encountered in measuring antenna gain, it is possible to correct for the reflection if some parameter can be varied to change the phase of the reflection. The antenna range used to measure the gain of the parabolic reflector with feed had a reflection from the support tower used to hold the test dipole. By sweeping the frequency over a 100-MHz band at 1296 MHz, it was possible to change the phase of the reflection enough to produce two peaks and one null.

If the magnitude of the reflection voltage relative to the direct wave is called r, then we are measuring 20 log(1+r) dB at the peaks and 20 log(1-r) dB in the nulls. If the reflection is very weak, we can take the average of the peak and null, in dB. But as the reflectons get stronger there is a problem in that the peaks can only get 6 dB stronger than the desired signal, whereas the nulls can become as deep as one can measure. The averages in dB will produce signal levels that are too low. To give the correct answer, one must convert the dB values back to voltage magnitudes and *then* take the average. The curve in Fig A does this, with the result expressed back in dB.

For example, if we measure –37 dB (loss) in our antenna range at the peaks and –42 dB at the null, the difference is 6 dB. Entering the graph at 6 dB we see that the correction factor from the peak is 2.5 dB. Thus the correct loss is –37 – 2.5 = –39.5 dB. This is 0.5 dB higher then dB averaging would have indicated.

This technique is limited by the variation in the gain of the antennas with frequency and is suitable only for broadband antennas. Also, there is a 6 dB increase in antenna-range loss between dipoles when the frequency is doubled. For highest accuracy, this effect should be calculated for the peaks and nulls using 20 log(f_c / f) where f is the frequency of a peak or null and f_c is the center measurement frequency. A 5% variation in frequency gives 0.45 dB too high a value of peak or null when low in frequency and 0.42 dB too low a value when high in frequency.

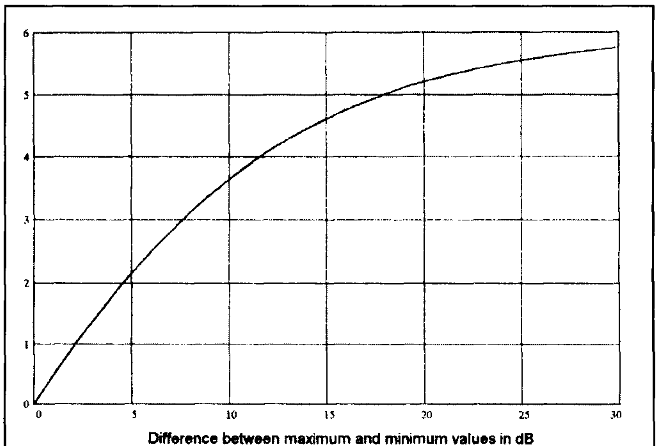

Fig A—Correction factor in dB for measurements with single reflections. Subtract this correction value from the maximum value measured.

The relative numbers are consistent with the calculated feed patterns and the resulting illumination efficiencies. The accuracy of the absolute gains is probably 1 dB or better.

Sun noise measurements were also made.[8,9] These tests include effects due to sidelobes that *see* the Earth and so are representative of the performance that would be obtained receiving satellite signals or other nonterrestrial sources. These measurements consistently showed the Type-A feed outperforming the dipole with reflector by even more than the gain measurements show. This indicates that the improved sidelobe performance with the new feed is more beneficial for reception than transmission, as expected.

I did not have an $f/D = 0.24$ to 0.3 dish to test, so the final testing of the Type-B feed intended for these deeper dishes will have to wait for a later date.

The "EIA" Feed

The EIA standard-gain antenna is sometimes used as a feed.[5] It's larger than the double-handlebar feeds described here and has somewhat higher gain. As such, it is more suitable for dishes with f/D from 0.4 to 0.5. Fig 9 shows the *EZNEC* calculated patterns for this antenna, and Fig 10 shows the resulting aperture efficiency versus f/D. The EIA antenna has a wavelength square reflector blocking the aperture; that can be an appreciable problem for lower frequencies where this can be a significant portion of the dish area.

Other Issues

Phase Centers

As the feed gets physically larger, the tendency is for the phase center to move as the angle of view is varied. Calculations done on these feeds indicate that a spacial spread of up to a half wavelength does not have phase changes of greater than about 15°, which is acceptable. Measurement of these feeds with a vector network analyzer confirmed these calculations.

Diffraction

Associated with the usual design procedure for reflector antennas is the *optical* assumption. This says that all behavior will be in accordance with the area and anglular orientation of the reflector surface, just like we see an optical mirror behave. Unfortunately, at 1296 MHz my 48-inch dish is only about 5 wavelengths across. This means that the discontinuity at the edge of the dish is affecting the entire operation of the antenna. Our predicted "best feed" does not take this into account, with the result that we will not achieve either the gain or sidelobe performance that we expect. *EZNEC* may be able to compute the gain and patterns for the entire feed plus reflector system. This would give some insight into these edge effects.

Focal-Point Adjustment

It is tempting to adjust the focus of the antenna by varying the feed location along the center axis. This usually produces a well-defined maximum in the response. Care must be taken in using this approach, however, since the impedance match to the feed also changes with feed location. This means that part of what is being measured is the loss in power transfer to the feed rather than just the change in reflector focus. To make this approach valid, it is necessary that the feed be rematched at every feed setting, or that the power transfer be calculated and taken into account.

A simpler way to deal with focusing is to carefully measure the phase center of the feed, then permanently place this point at the dish focal point. The match to the feed can then be given a final adjustment with the feed set in position. The drawback is that one must have a system for measuring RF phase to determine the phase center. This requires a *vector* measuring system such as a vector voltmeter or a vector network analyzer.

Other Frequencies

The dimensions are not too critical, and these feeds would work well at 2304 or even 3456 MHz. Starting around 3456 MHz, though, it is probably best to use a circular horn feed.[2] The physical size of these horns ceases to be an issue, and the simplicity of the horn becomes an advantage. Scaling of these feeds down to 903 or 432 MHz should work well, and the small size of the feeds makes their use worthwhile.

Circular Polarization

These feeds were intended for linear polarization. However, it should work well to nest a pair of double-handlebar dipoles at right angles with a half-wave-diameter circular reflector. Feeding these 90° out of phase would produce circular polarization. The slotted-boom feed would not be useable, of course, but a pair of transmission lines could be brought out to the feed point on the boom along with support insulators. I would be interested to hear if someone tries this.

Conclusion

The double-handlebar feed for parabolic reflectors can be used for f/D = 0.24 to 0.4 with excellent results. Two designs can cover this range of reflector depths. The feeds were analyzed for aperture efficiency, and low f/D reflectors were found to have little, if any, gain penalty using these feeds. The feeds are easily constructed, low in wind resistance and aperture blockage, and well suited for frequencies up to 3.5 GHz.

One footnote to this project is that the Type-A feed was installed on the 4-foot dish and using 30 W, the first contact was with W7ID at 350 miles. This was home-station-to-home-station over a mountainous path under dead-band conditions.

About the Author

Bob Larkin, W7PUA, has been active since he was first licensed in 1951 as WN7PUA. Moving to New Jersey, he was licensed as W2CLL for 12 years. Bob is now a consulting engineer for communication companies and holds an MSEE from New York University. His interests are primarily VHF/UHF and microwaves, and he is active on all bands from 50 to 10368 MHz, with a particular interest in pushing the limits of weak-signal work.

Notes

1 Malowanchuk, B. W., "Selection of an Optimum Dish Feed," *Proceedings of Central States Conference*, 1979.

2 Wade, P., N1BWT, "Practical Microwave Antennas, Part 2-Parabolic Dish Antennas," *QEX*, Oct 1994, pp 13-21. This is an excellent summary of feed design and current amateur design trends.

3 Silver, S., *Microwave Antenna Theory and Design*, McGraw-Hill, 1949. This is the antenna book from the Radiation Laboratory series, available in most university or technical libraries. Dover Publications did a soft-cover reprint in 1965. Dipole-with-reflector feeds are well covered in chapter 8. Chapter 10 has good information on horn design, and chapter 12 covers the feeding of parabolic dishes.

4 *EZNEC* Program, available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.

5 Straw, R. D., N6BV, Editor, *The ARRL Antenna Book*, 17th edition, ARRL. Chapter 27 covers antenna measurements and the EIA standard gain antenna.

6 Cutler, C. C., "Parabolic-Antenna Design for Microwaves," *Proceedings IRE*, Vol 35, Nov 1947, pp 1284-1294. Also reprinted in *Reflector Antennas*, Love, A. W., Editor, IEEE 1978.

[7]Turrin, D., "Antenna Performance Measurements," *QST*, Nov 1974.

[8]Hoch, G., "Determining the Sensitivity of Receive Systems with the Aid of Solar Noise," *VHF Communications*, 1980, Number 2.

[9]Curry, Jr., W.H., "Antenna-Performance Measurements using Celestial Sources," *Ham Radio*, May 1979.

[10]The program DISHFD.BAS can be obtained in the file DISHFD1.ZIP downloadable from the ARRL BBS at 860-594-0306, or via the Internet at **http://www.arrl.org/files/qex** or **ftp://ftp.arrl.org/pub/qex**.

Calculation of Parabolic Reflector Aperture Efficiency

Notes 3 and 6 show the method for calculating aperture efficiency when the feed pattern is known. Loss of gain due to phase errors is not included, and the feed pattern is assumed to be axially symmetric. The equations involve the calculation of integrals, these being the same as areas bounded by specific curves. The following BASIC program takes a feed pattern specified in dB at 10° points, starting from the center axis, and from this calculates the aperture efficiency. Also calculated are the contributions from the two associated quantities, illumination efficiency and spillover efficiency. The feed pattern is from a text file having the format described in the program.

In order to be as accurate as possible with data points only every 10°, a smooth curve is placed through sets of four voltage magnitude data points. This smooth curve, expressed as a cubic polynomial, is then used to calculate the pattern every degree between the central two 10° points. This process is repeated every 10° until all 180° points have been found in 1° steps. The integrals are then calculated from these detailed data points with the angle subtended by the reflector from the feed as a parameter. This angle is also related to the dish *f*/*D*, and both quantities are printed out.

This program, along with several data files, is available for downloading.[10]

```
REM DISHFD1.BAS ver 1.0
REM
REM QBasic program to calculate the efficiency of a parabola as a
REM function of the feed amplitude characteristics and the f/D of
REM the dish.     Bob Larkin, W7PUA    20 Dec 95

DIM FEED#(20), IFEED#(180), ETC#(17), ETAS#(17), ETAI#(17), C#(4, 5)
DIM IN1#(180), IN2#(180)
PI# = 3.14159265359#

REM Files should have 19 entries on separate lines. The values should be
REM in dB attenuation relative to the peak gain of the feed. Thus, there
REM should be a 0.0 entry somewhere and no entries should be negative.

INPUT "Input file name, including path and extension, for feed data"; NAMEF$
OPEN NAMEF$ FOR INPUT AS #1
FOR I% = 1 TO 19
  INPUT #1, FEED#(I%)
  FEED#(I%) = 10# ^ (-FEED#(I%) / 20#)  'Convert to Voltage magnitude
NEXT I%
FEED#(0) = FEED#(2)  'Provides symmetry at the axis for interpolation
FEED#(20) = FEED#(18) 'This does it at 180 degrees as well

REM Next interpolate 18 times by fitting a cubic through adjacent 4 points
REM and interpolating the polynomial at each degree.
FOR I% = 1 TO 18  'Over all 10 degree sectors
  REM For numerical accuracy, we will always consider angle as (-10, 20)
  FOR J% = 1 TO 4  'Over 4 data points, set up equations
    C#(J%, 1) = 1
    FOR K% = 2 TO 4 '3 coefficients per equation
      C#(J%, K%) = (10# * (J% - 2)) ^ (K% - 1)
    NEXT K%
    C#(J%, 5) = FEED#(I% + J% - 2)
  NEXT J%

  REM Now solve 4 equations for 4 coefficients
  FOR J% = 1 TO 4
    FOR L% = J% TO 4
      IF ABS(C#(L%, J%)) > 1E-10 THEN GOTO NONZ
    NEXT L%
NONZ:  FOR K% = 1 TO 5
      CT# = C#(J%, K%)
      C#(J%, K%) = C#(L%, K%)
      C#(L%, K%) = CT#
    NEXT K%
```

```
    CT# = 1# / C#(J%, J%)
    FOR K% = 1 TO 5
      C#(J%, K%) = CT# * C#(J%, K%)
    NEXT K%
    FOR L% = 1 TO 4
      IF L% = J% THEN GOTO LUP
      CT# = -C#(L%, J%)
      FOR K% = 1 TO 5
        C#(L%, K%) = C#(L%, K%) + CT# * C#(J%, K%)
      NEXT K%
LUP:  NEXT L%
  NEXT J%
  REM Now interpolate to each degree point
  FOR J% = 0 TO 10
    AJ# = J%  'Evaluate the polynomial
    V# = C#(1, 5) + C#(2, 5) * AJ# + C#(3, 5) * AJ# ^ 2
    V# = V# + C#(4, 5) * AJ# ^ 3
    IFEED#(10 * (I% - 1) + J%) = V#  'Save in deg by deg vector
  NEXT J%
NEXT I%

REM Now integrate the interpolated curves to find the antenna efficiencies
REM Collect integrands, U*tan(theta/2) and U^2 * sin(theta) in IN1(), IN2().
REM Note that integration period 1 is 0 to 1 degrees, centered on 0.5 deg.
DENOM# = 0#
FOR I% = 1 TO 180
  U# = .5# * (IFEED#(I% - 1) + IFEED#(I%))
  THETA# = PI# * (I% - .5) / 180#
  HTHETA# = .5# * THETA#
  IN1#(I%) = U# * TAN(HTHETA#)
  IN2#(I%) = U# * U# * SIN(THETA#)
  DENOM# = DENOM# + IN2#(I%)
NEXT I%
DENOM# = PI# * DENOM# / 180#  'Scale to be an area
PRINT
PRINT "I. eff = Illumination efficiency, S. eff = Spillover efficiency"
FOR TH1% = 5 TO 90 STEP 5
  NUM# = 0#
  FOR I% = 1 TO TH1%     'TH% is half angle subtended by dish
    NUM# = NUM# + IN1#(I%)
  NEXT I%
  NUM# = PI# * NUM# / 180#
  DENI# = 0#
  FOR I% = 1 TO TH1%
    DENI# = DENI# + IN2#(I%)
  NEXT I%
  DENI# = PI# * DENI# / 180#
  K# = 2# / ((TAN(PI# * TH1% / 360#)) ^ 2)
  FOD# = 1# / (4 * TAN(PI# * TH1% / 360#))
  ETA# = K# * NUM# * NUM# / DENOM#
  EI# = K# * NUM# * NUM# / DENI#
  ES# = ETA# / EI#
  PRINT USING "Half Angle=## deg, f/D=#.### "; TH1%; FOD#;
  PRINT USING "Feed=###.##dB "; 8.68589 * LOG(IFEED#(TH1%));
  PRINT USING "Eff=#.### I. eff=#.### S. eff=#.###"; ETA#; EI#; ES#
NEXT TH1%
INPUT "Enter to continue"; ZZ$
STOP
```

A Homebrew 2-m Repeater

This design from the West Virginia Institute of Technology Amateur Radio Club uses off-the-shelf radios—without modification.

By John Thomas Crago, KB8DAN

Introduction

Here's a project that homebrew enthusiasts can appreciate. What do you get when you take a handful of bright Electrical Engineering and Electrical Engineering Technology students with Amateur Radio licenses, anxious to get their feet wet in electronics, and put them in a college ham radio club? Well, when the extent of the club's activity depends primarily on the incomes of college students, you end up with a lot of low-cost, innovative ideas for group projects and homebrew equipment.

In the fall semester of 1994, the West Virginia Institute of Technology Amateur Radio Club was deciding on our first group project of the term. It had been observed in previous years that the mountainous terrain in which WV Tech is tightly snuggled made reaching outside repeaters on 2-m a difficult task, to say the least. Also, since the majority of the club members were housed on campus and the dorms have very strict rules about outside antennas, 2-m operation was restricted to mostly low-power, "rubber ducky" equipment. It had even been noted several times that with the campus arranged amidst large mountains, hand-held communications via simplex were often unreliable from one end of the town of Montgomery, West Virginia, to the other. The club station had the capability of reaching distant repeaters but is located a mile away from most of the members.

1453 Winslowe Dr, Apt 201
Palatine, IL 60067

The obvious solution was to install a strategically located repeater station on campus that could aid our signals farther down the long valley that guides the Kanawha River to Charleston. Of course, the purchase of a new, commercial repeater station was out of the question. (College students are seldom noted for their excess pocket change!) It was at this point that the club decided to design and construct our own simple 2-m repeater station.

Considerations

In the planning process, we considered what, exactly, we were capable of. Floor space was a concern. The Amateur Radio club shares a room with a class laboratory, so the amount of available space is limited.

We also had a few things to keep in mind as far as the equipment was concerned. The radio equipment was partly owned by the club and partly borrowed from fellow hams. Since the radios were not entirely ours to do with as we pleased, modifications to any equipment were out of the question. The transceivers themselves were nothing out of the ordinary: simple 2-m transceivers available to any amateur. Application then became a matter of how to take two every-day transceivers and arrange them into a functional repeater station. This called for an interface that could somehow link the two transceivers so that one would automatically retransmit whatever the other received.

Of course there were also legal matters to keep in mind. The station must be capable of identifying itself at regular intervals when in use. Other user-friendly matters such as hang-time and time-out functions found in most repeaters today would also be handy assets.

Theory of Operation

Fig 1 shows the features of the repeater design. Explanations of the individual repeater functions follow.

Repeater Activation

The repeater transmitter must know when the receiver is detecting an incoming signal so it can key-up and relay the signal being received by the receiver. Remembering that any internal modifications to the borrowed equipment are not allowed, the problem was how to inform the repeater transmitter that the receiver was active. Our particular receiver was equipped, as are many 2-m transceivers, with a *receiver active* light. The light comes on when a signal comes through the receiver. The transmitter could easily monitor this by using the circuit in Fig 2.

Though Fig 2 is basically a light detector, it is also a simple analog-to-digital (A/D) converter. Using a photocell, the voltage division between the resistor, R, and the photocell can be varied greatly. Ohm's law says that the voltage across a resistor is equal to the current flowing through that resistor multiplied by the value of the resistor in ohms. A photocell's resistance increases in dark conditions and decreases a great deal when submitted to a light source.

In other words, by placing the photocell over the top of the receiver's active light and covering it to prevent any outside light from entering, the photocell will keep constant track of when the receiver is on or off. The voltage drop across the photocell under dark conditions will result in a large portion of the +5 V entering into the CMOS 4049 inverter/buffer, producing a low out of the 4049. When the receiver is active and the light is on, the voltage drop across the photocell is relatively low and the inverter will output a high to turn on the repeater controller circuitry. The CMOS 4049 chip was chosen for this particular application due to its low-current consumption.

Controller

When a signal is present at the receiver, the 4049 provides a +5-V input to the controller circuitry. When you

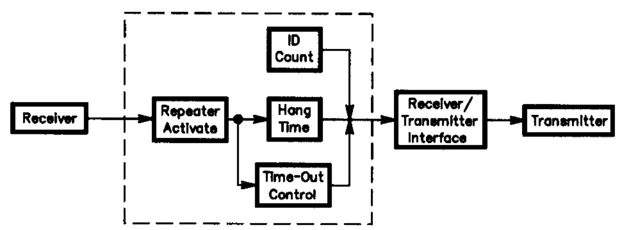

Fig 1—Block diagram of repeater operation.

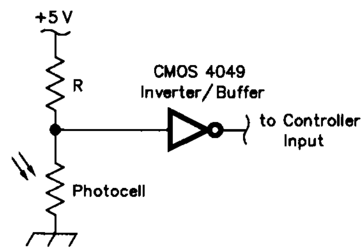

Fig 2—Receiver-on light detector.

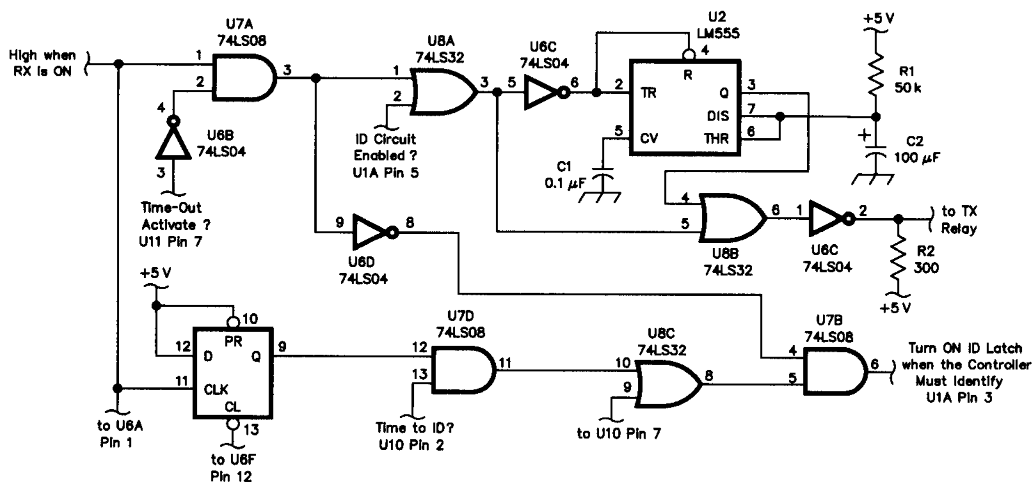

Fig 3—Repeater activation circuit.

are designing the voltage divider and determining the value of the resistor to be used with your specific photocell, keep in mind that a TTL high level is actually around 3.8 V to 5 V. Voltages over 5 V are not recommended for very long periods of time.

The high input does a few things, as can be seen from the schematic in Fig 3. It turns AND gate U7A on, allowing the U6C 7404 inverter to cause a current to flow through the coil of the TX relay, closing the PTT switch. The high at the input also sets D flip-flop U1B. U1B stays on until the repeater eventually identifies. U1B ensures that as long as there is a signal present in the repeater, the controller will ID every ten minutes (or however long the ID timer is set for).

The high input also passes through inverter U6A (Fig 4) and holds the clear pin of the time-out counter, U11, low, which makes U11 begin counting. U11 will count until it reaches a predetermined time-out period. For this controller, the time-out is set for around one and a half to two minutes. The use of variable resistors and/or capacitors at the 555 timer IC, U3, will allow fine tuning of the timing of the counters. If the input into the controller stays high for more than the time-out period, pin 7 of U11 will go high,

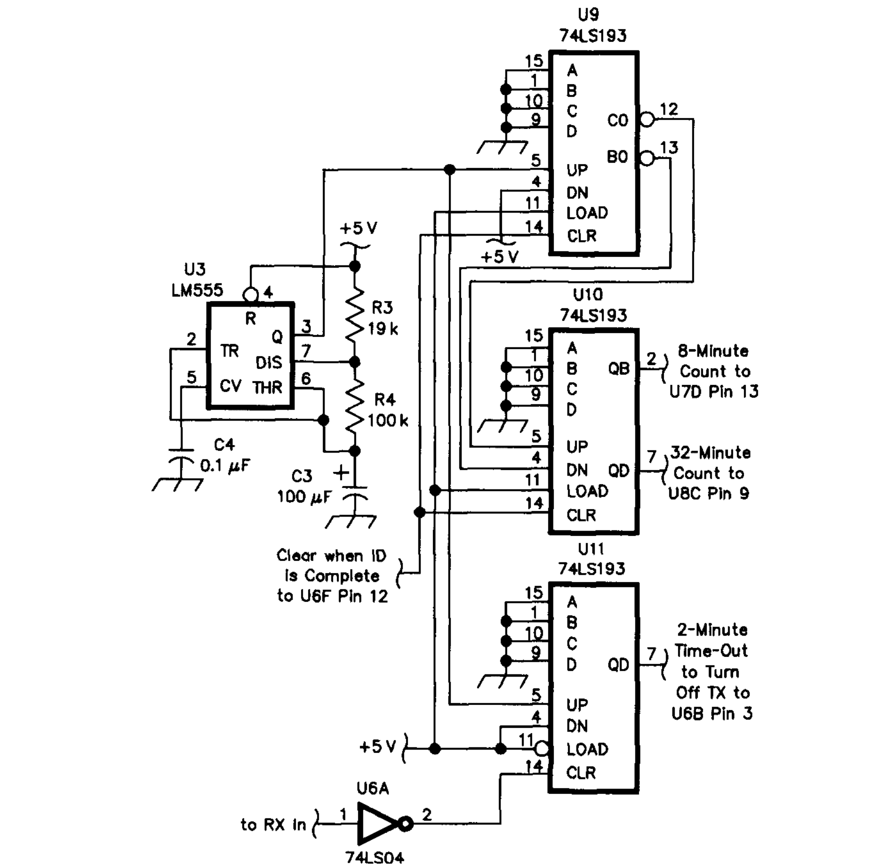

Fig 4—Repeater timing circuit.

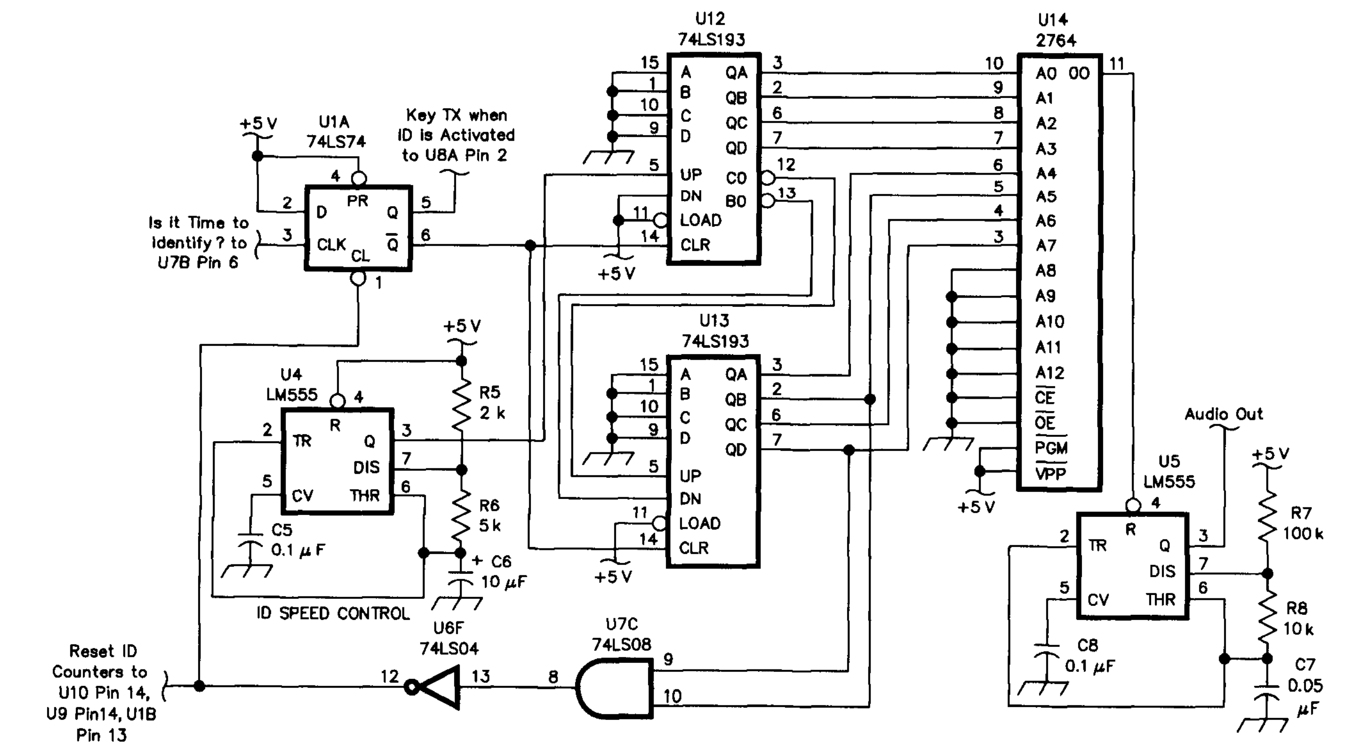

Fig 5—Morse code ID circuit.

pass through inverter U6B and cause a false condition at the input of U7A. This turns the output of U7A off, and the transmitter will also be turned off. However, thanks to OR gate U8A, the ID circuitry will still be allowed to operate through the transmitter. In this way, if the repeater has not identified for eight minutes and a carrier comes on the repeater for a period of two minutes without a break, the repeater will time out, even if the carrier is still present, and allow the controller to ID. The repeater will remain off the air until the carrier is eventually dropped and counter U11 is cleared. This is the operation found in most repeaters and is very useful in providing strict observance of FCC rules.

With the repeater capable of IDing in the presence of a carrier, we must make sure that its ID is on time. The repeater must ID every ten minutes when in use. Note that counters U9 and U10 (Fig 4) are chained together in such a way that the counting of U9 will be allowed to continue over to U10, providing for a counter capability twice that of U11 since all three counters are connected to the same timing oscillator, U3. Pin 2 of counter U10 turns on at approximately eight-minute intervals. This pin is connected to the input of AND gate U7D. The output of U7D also depends on flip-flop U1B. If there is a received signal present, or if there was one present at any time since the last repeater ID, the flip-flop is set and the output of U1B will enable pin 12 of U7D. Since counter U10 says it is time to ID, and U1B says there is a carrier present, pin U11 of U7D goes high and goes on to pin 5 of U7B.

Here another decision is made. If there is a signal present at the exact time the repeater wants to ID, pin 8 of inverter U6D is low and the information is not allowed to pass to pin 6 of U7B. This prevents an unwanted repeater ID on top of a transmitting signal. However, as soon as the carrier drops, at a time no longer than two minutes (8-minute ID plus 2-minute time-out equals the 10-minute FCC ID requirement), the output, pin 6, of U6D goes high and sets flip-flop U1A. U1A now goes high at pin 5 and low at pin 6. Pin 5 turns on OR gate U8A that turns on the transmitter in the same way the incoming carrier would. Pin 6 of U1A is low and removes the clear from the counter pair, U12 and U13. This allows timer U4 to cycle counters U12 and U13 to access each address of the Morse ID EPROM, U14 (Fig 5).

The ID circuitry and tone oscillator use a principle that has appeared in the *ARRL Handbook*. This design uses the same idea with different chips. The counters activate successive addresses of U14, reading the binary message programmed into the EPROM. The method of programming the EPROM for a particular message is covered on page 14-9 in the 1994 edition of the *ARRL Handbook*. The output of the U5 tone oscillator is coupled to the mike input of the transmitter via a small coupling capacitor.

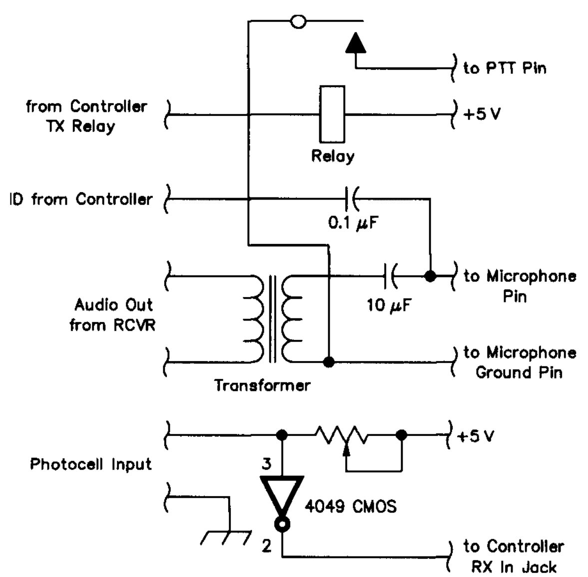

Fig 6—Receiver/transmitter interface.

As soon as counters U12 and U13 finish the ID process, their outputs force AND gate U7C high, clearing flip-flops U1A and U1B. U1B will immediately go high when another signal is received, and U1A will turn back on when the controller must ID again. Note that counter U10 provides another opportunity to enable the repeater to ID at an interval much greater than that of ID pin 2 of U10. If pin 2 turns on after 8 minutes, pin 3 turns on after four minutes, pin 2 after eight minutes, pin 6 after 16 minutes, and pin 7 after 32 minutes. This implies that the repeater will ID every 32 minutes even if there has been no repeater activity. Though not a requirement by law, this may be a desirable feature if a repeater control station wishes to ensure the repeater is operating normally at regular time intervals without needing to inspect it via remote control. This option may be omitted from Fig 4 with no adverse effect to the circuit. This can be done easily by taking the output of U7D directly to the input, pin 5, of U7B and removing U8C entirely.

Another function of the controller is the hang time. The 555 labeled U2 (Fig 3) is configured as a one-shot that will take a high and extend it out for a given amount of time even after the input has dropped. When the incoming signal goes high, U2 receives a low from the output of inverter U6C. When the signal drops from the input of the repeater, U2 will lengthen the high by an amount of time given by the calculations later in this article. The controller used by our repeater has a hang time of 5 seconds between transmissions. If no other carrier is detected within these 5 seconds, the repeater is turned off until U2 is reset by the presence of another incoming carrier.

To someone unfamiliar with logic circuitry, the controller circuitry may seem confusing at first. Stepping through the circuit operation a few times will help you follow the flow of events in the controller.

Receiver / Transmitter Interface

The special feature of this repeater design is that it can be used with most any receiver and transmitter found at a hamfest or in ham shacks and closets. Our repeater design actually required no receiver or transmitter modifications—the covers never even had to be removed.

The audio output of the receiver may be fed to the microphone pin of the transmitter mike input, transferring the signal from the receiver to the transmitter for retransmission of the signal. Another design decision must be made at this point. Many common microphone input impedances range from a few hundred to several hundred ohms. The output of the receiver will most likely be an 8-Ω impedance. The most desirable method of impedance matching is by using an audio transformer. This transformer is readily found at many electronics and circuit retailers—ours was found at the local Radio Shack. Without the use of an impedance transformer between the two radios, the signal out of the transmitter would sound like a very worn out audio tape, if it's audible at all!

Another item that should be placed in the receiver/transmitter interface is the transmitter relay. The transmitter microphone jack consists of several pins. Consultation with the transmitter's user's guide will most likely provide information as to which pin does what when connected to the microphone. The pin marked PTT, push-to-talk, and the microphone ground are the pins used for the keying action of the transmitter. When these two pins are shorted together, it simulates the action of pushing the microphone button and the transmitter is turned on.

Fig 7 shows the proper connection for the small Radio Shack relay used in our design. As soon as a carrier is detected at the controller input, U6C goes low, very close to ground voltage level. This passes current through the relay, closing the connection, which in turn shorts the pins of the microphone input, keying the transmitter. Fig 8 shows the pin-out for our particular transmitter.

Design Equations and Calculations

For the 15-Second 555 Timer

The timer used with the time-out and ID counters is a 15-second timer. The counters used are binary counters, which implies that each state indicates a 15-second change in time: When pin 1 is high, 15 seconds has elapsed; When pin 2 is high, 30 seconds has elapsed; and when pins 1 and 2 are both high, 45 seconds have gone by. When each successive pin turns on, all of the previous pins before that one are off and the time elapsed is 2^M times the time set by the counter timer, where M is the number of pins below the one currently high. For example, since we had two 4-bit counters chained together, the second pin on the second counter represents a time of:

$2^5 \times 15$ seconds
$= 32 \times 15$ seconds
$= 8$ minutes

The calculations for the 15-second 555 timer follow. Values for R3 and R4 were calculated:

$f = (1.46)/[(R3+R4) \times C3]$, where
$C3 = 100\ \mu F$
$f = 1/T$, where T is the period of one timer cycle
$= 1/15$ seconds

Therefore,

$(R3+2R4) \times C3 = 21.90$
$R3+2R4 = 219000$
$R3 = 19\ k\Omega$ and $R4 = 100\ k\Omega$

Of course, these values could have varied as long as the algebraic equation above was satisfied.

ID Speed Control

FCC rules require that any Morse code repeater ID must be sent at no greater than 20 words per minute. Our particular message consisted of 153 bits that we wanted to send in a period of around twelve seconds. The 153 bits were derived by the number of memory spaces, both 1s and 0s, programmed into the EPROM from the time the message began until it finished. With this in mind:

$12/153 = 0.078$ sec/bit
$f = 1/0.078 = 12.7$ Hz

Say an average five-character American call, using the EPROM programming method shown on page 14-9 of the 1994 *ARRL Handbook*, will take up approximately 65 bits; we get:

65 bits$\times$0.078 sec/bit=5.07 sec
1 minute=60 sec
60 sec/5.07 sec=11.83.

This puts the speed of the CW identifier well below 20 wpm if we count the call sign as one word.

The timing circuit component calculations are:

$f = (1.46)/[(R5+2R6) \times C6]$, where
$C6 = 10\ \mu F$
$(R5+2R6) \times C6 = 0.1176$
$R5+2R6 = 11758.39$

R5 and R6 use readily available values at 2 kΩ and 5 kΩ, respectively.

Hang-Time One-Shot

The hang-time length is by no means vital, but some users may find it annoying to have a long period of empty carrier after a call, or to have the repeater cut out abruptly even after a small break in transmissions. Our hang time is 5 seconds and was determined by calculating component values for U2:

$T = 1.1 \times R1 \times C$

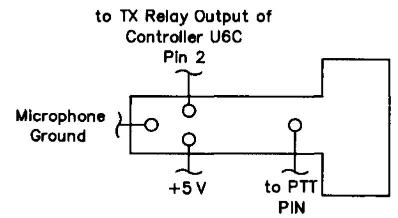

Fig 7—Relay connections.

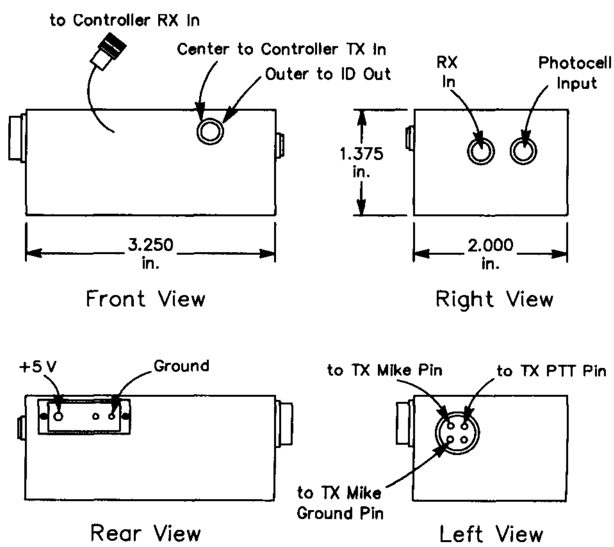

Fig 8—Exact layout and dimensions of the repeater interface.

where T is the time of the one-shot. C was again 100 µF. The value of R1 was approximated at 50 kΩ. The value here, again, is not vital. A half second either way of the desired hang time will hardly be noticeable to the average repeater user.

Photocell Receiver-On Detector

The photocell is simply a resistor whose resistance depends upon the light absorbed by the face of the photocell. Most photocells have differing resistance ranges, so the following values may be modified for any particular photocell.

Our photocell had a resistance of 270 kΩ in the dark and a much smaller 800 Ω in the light. The diagram of the photocell circuit shows how the voltage divider circuit is assembled. The 4049 buffer transforms the voltage level between the resistor and photocell to TTL level. The calculations show how the 200-kΩ value of R was derived. The voltage divider is expressed by Kirchhoff's voltage law so that:

RX light on (signal present)

[(5 V)(800 Ω)]/[(200 kΩ+800 Ω]= 0.02 V, which is much less than 0.8 V. The CMOS inverter/buffer makes this low level a high and turns the transmitter on.

RX light off (no signal present)

[(5V)(27 0kΩ)]/[200kΩ+270kΩ]= 2.87 V, which is much greater than 1.6 V. The CMOS inverter/buffer makes this high level a low and turns the transmitter off.

So when the receiver light is on and a signal is present, the photocell exhibits a low resistance value. The majority of the voltage drop then is across the other resistor, R. The voltage across the photocell, and thus the voltage entering the inverter/buffer, is small in comparison, well below the low-level voltage. The inverter's output then produces a high level that is sent to the input of the controller and to the TX relay, keying the transmitter. Note that a simple buffer, without the inverter, would work just as well if the photocell and resistor in Fig 2 were reversed.

Construction

With the design complete, it was time to put all of the circuitry in a package that could all be quickly assembled or disassembled to make it readily portable.

The assembly was done on a Radio Shack predrilled perf board. The individual chips were placed in wire-wrap sockets, and all connections were made by wire wrapping with 32-gauge wire. Wire wrapping so many components may seem tedious, but for a beginning builder this is one of the safest methods. The TTL ICs are very susceptible to heat, so an amateur homebrewer unaccustomed to the proper use of a soldering iron can easily destroy the internal components of a chip without leaving any outside evidence. This makes troubleshooting a nonfunctional controller quite annoying.

Metal is preferred for the enclosure. Keep in mind that all of the digital circuitry involved is easily corrupted by RF energy from nearby sources; the repeater cabinet will contain a couple of 2-m rigs as well as the controller. The controller and interface should be placed in metal enclosures with proper connectors to prevent unwanted RF interference. The entire repeater station was placed in a metal cabinet with the major pieces of equipment located on different levels. A metal shelf between each component gives additional isolation.

The station was assembled for an estimated cost of a few hundred dollars, including the receiver and transmitter. The actual controller and interface combined cost approximately thirty dollars. This cost estimate is flexible since most hams will have spare resistors and capacitors lying around the shack and may even have a few enclosures suitable for mounting the repeater equipment.

The homebrew repeater station described in this article is relatively simple to build, inexpensive to construct and maintain, and all of the elements required are readily available from most electronic dealers. The schematics may seem as if there are quite a few ICs used in the construction of the station, but the repeater can fit in a very small space. The exact layout and dimensions for the interface used by our club are shown in Fig 8.

The controller circuitry is housed in a 7×5×2-inch enclosure. A shielded microphone cable with appropriate connectors is used to relay the audio signal and PTT information from the interface to the transmitter input jack.

The entire station is shown in Fig 9. Two separate antennas are used for the transmitter and receiver. They are spaced about thirty feet apart horizontally and about sixty feet vertically. This vertical spacing delivers the amount of antenna isolation required

Parts List

Part	*Quantity*
2764 EPROM	1
74LS08 AND gate	1
LM555 timer	4
74LS193 4-bit counter	5
74LS04 inverter	2
74LS32 OR gate	1
74LS74 dual D flip-flop	1
4049 CMOS inverter	1
Audio matching transformer	1
5-V, 1-A relay	1
Photocell	1

Various capacitors and resistors.
Some parts distributors include:
Mouser Electronics, 1-800-346-6873
Digi-Key, 1-800-344-4539
Radio Shack, 1-800-THE-SHACK

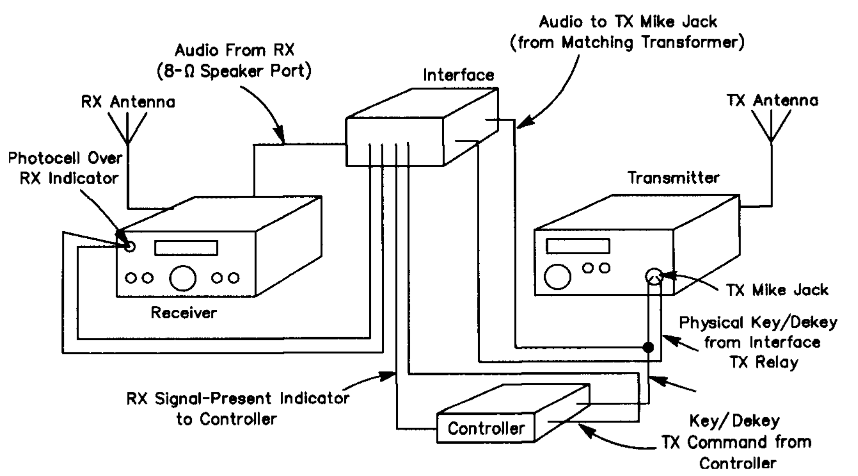

Fig 9—Entire repeater station.

without the use of a duplexer.

In Conclusion

Of course, repeater design is not limited to the methods chosen. A single microprocessor could eliminate the need for many of the TTL-level logic chips used in this design. It is, however, a difficult task for a beginning builder to program and implement a microprocessor device. The most difficult programming feature in this design was the entering of 1s and 0s into a limited memory field.

There are many other methods for detecting the presence of a signal at the receiver. If the receiver is free for cover-off modifications, the A/D converter could be operated from the squelch circuit inside the receiver. By detecting an interruption in the squelch field of the receiver, an incoming signal could be flagged immediately. Also, a phototransistor could be used in place of the photocell. The presence of incoming light would induce a current that could be used to close the contacts in the TX ON relay.

Also note that by using variable resistances for R1, R5 and R6, the duration of the hang-time and the speed of the CW identifier, respectively, may be controlled easily at the control operator's discretion.

Thanks to the Spring 1995 WVIT ARC for all of their help in designing, building and testing the repeater station: Tony Mauser, KB8GZW; Stacy Fisher, KB8YQK; Christopher Ferguson, KB8VRL; Lee Van Iderstine, N1SCQ; and Mark Wensyel, KB8TAC. Thanks also to Shawn Allen, KB8IYA, and Brian McClure, N8PQI. Special thanks to the club advisor, Dr. James Cercone, N8OZO.

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□□

Frequency Stabilization of L-C Oscillators

So you have a clean, pure L-C VFO, but it drifts. Here's a way of stabilizing its frequency.

By Klaas Spaargaren, PAØKSB

For many builders and users of simple receivers and transmitters, the long-term stability of oscillators forms a major problem. Usually, oscillator drift caused by temperature variations is reduced by the use of temperature sensitive components such as NTC capacitors. However, crystal stability is seldom achieved.

This article describes a digital method for frequency stabilization by which an HF L-C oscillator can achieve the stability of a crystal oscillator. This simple unit can be used for new construction or as an add-on to existing equipment.

Cort van der Lindenplantsoen 13
1181XP Amstelveen
Holland

Principles of Stabilization

Two simple principles exist for L-C oscillator frequency stabilization (I'll disregard complex synthesiser systems.) One method uses a digital frequency counter that measures a VFO frequency periodically. Any deviation between the measured values and a preset value results in automatic correction of the VFO frequency. The principle can be executed with simple means.[1]

This article describes the second principle, which uses a mixer to downconvert the VFO frequency to a low value for further treatment. It has significantly better performance than the technique mentioned above.

[1]Notes appear on page 23.

Basic Operating Principle

A special form of frequency-locked loop is used. Fig 1 shows the block diagram. The heart of the system is a digital mixer. The VFO frequency is mixed with a crystal oscillator to a low frequency, which is then compared to a reference frequency. Any deviation between the frequencies, caused by drift, generates a control voltage that corrects the VFO.

The digital mixer is a standard high-speed CMOS D-type flip-flop. Its operation as a digital mixer is not intuitively clear, so I will explain it in some detail.

In a D flip-flop the information on the data input (1 or 0) is transferred to the Q output on the low-to-high transition of the clock pulse. In this case, the clock pulse is a 10-Hz signal derived from a crystal oscillator.

The VFO signal is connected to the data input. In this way the D flip-flop acts as a digital sample-and-hold circuit. Since the output can change only at the rising edge of the clock pulse, the output frequency can never become higher than 5 Hz (half the clock frequency), but it is still determined by the two input frequencies. Let's see how.

If at each transition of the 10-Hz clock the VFO signal is high (or low)—exactly in phase with the clock—the output will remain high (or low). The output frequency is 0 Hz.

If the VFO frequency is tuned slightly upwards, say 1 Hz, an output frequency of 1 Hz will occur. If the VFO frequency increases further, the output increases equally up to its maximum value of 5 Hz. As the VFO frequency increases further, the mixer output goes back to 0 Hz and rises again as the VFO continues to be tuned higher.

If we keep the output of the digital mixer constant by automatically controlling the VFO, the VFO's frequency must be constant as well. We have made a frequency-locked loop.

At a VFO frequency that is 10 Hz higher, the digital mixer will generate exactly the same output frequency. This type of frequency-locked loop has many stable operating points, with 10-Hz spacing between the lock points.

In a stable loop, a fixed ratio N exists between the VFO frequency and the clock frequency. N is always a very large number. (See the calculations in the Appendix.)

Automatic frequency control of the

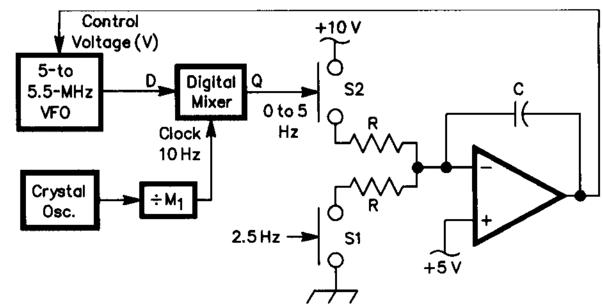


Fig 1—Block diagram of the basic frequency-stabilization system.

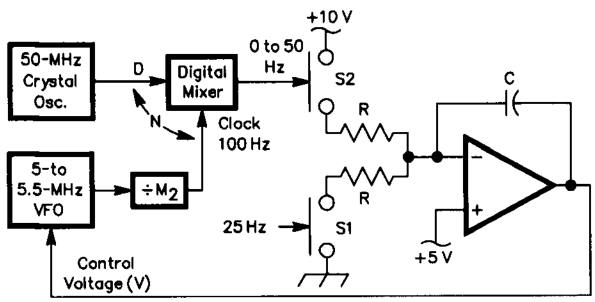


Fig 2—Block diagram of an improved system.

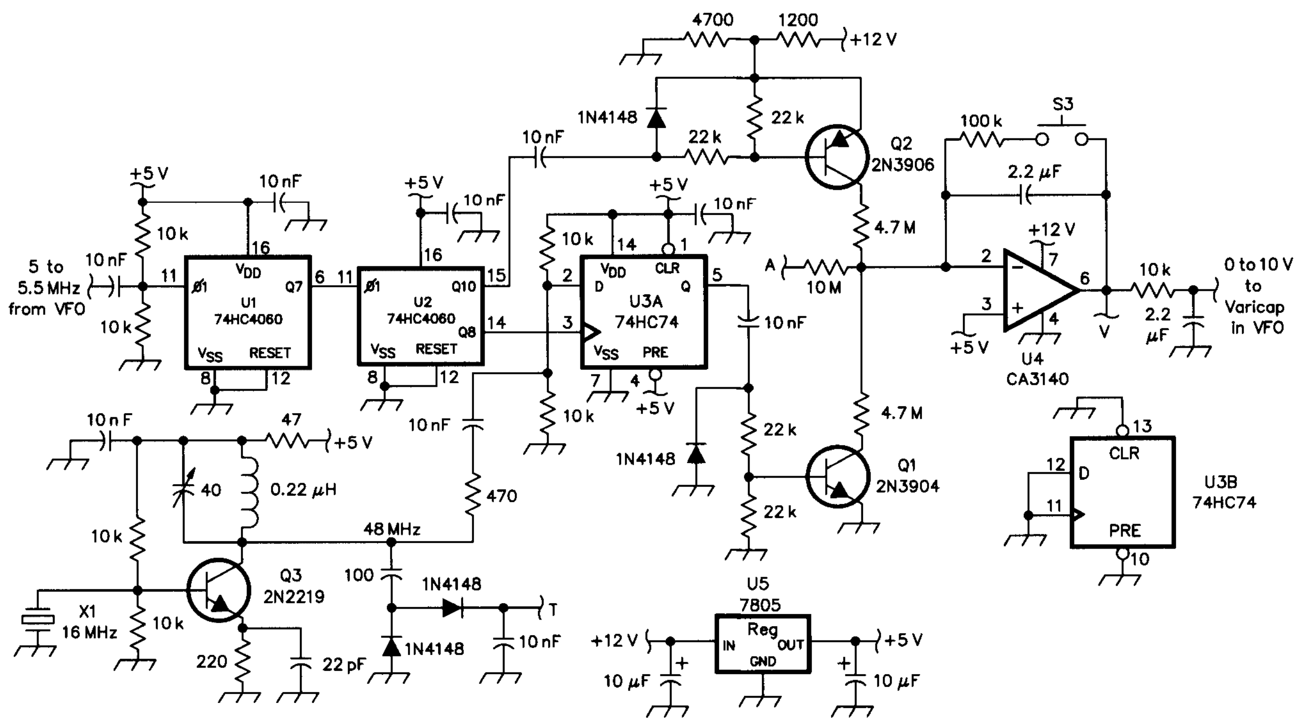


Fig 3-Schematic diagram of the frequency-stabilization circuit. The VFO circuit is not shown.

VFO is achieved in the following way:

The 0 to 5-Hz output frequency of the digital mixer is compared to another low frequency. (This frequency does not need to be crystal stable.) 2.5 Hz is a suitable value, as the output of the digital mixer will then stabilize in the middle of its range. This reference signal is conveniently derived from the same crystal oscillator as the 10-Hz frequency: ($f_{clock}/4$).

Both signals momentarily close switches S1 and S2, resulting in a slightly increased or decreased charge in the hold capacitor, C, of the integrator after each pulse.

The ouput voltage, V, of the integrator is connected to a voltage-variable capacitor in the VFO. So the VFO frequency also changes slightly up or down after each pulse until both low frequencies are the same. The frequency stays on such a stable lock point, and any slow frequency drift is corrected.

The long-term stability of the VFO is then determined only by the crystal oscillator from which the 10-Hz signal is derived.

Advanced Operating Principle

A considerable improvement is still possible. In the system described above, the correction pulses have a frequency of 2.5 Hz. If it were possible to increase that frequency and still maintain the 10-Hz lock points, a higher VFO drift rate could be corrected; or with the same drift rate, a faster correction with smaller individual steps would be possible.

This is what happens in the system shown in Fig 2. The main difference between Fig 1 and Fig 2 is the location of divider M. The crystal oscillator provides a 50-MHz signal to the digital mixer. Suppose that the VFO frequency of 5 MHz is divided by the fixed divider M_2, when M_2=50,000. This results in a 100-Hz signal to the clock input of the D flip-flop. Its maximum output frequency will now be 50 Hz, ten times as high as before. A little arithmetic will show that the spacing of the lock points is still 10 Hz.

N is the ratio between the frequencies of the crystal oscillator and the output of divider M_2. In the example N=500,000, M_2=50,000.

When the system is stable the following relation holds:

$$F_{vfo}/M_2 = F_{xtal}/N \qquad \text{(Eq 1)}$$

resulting in:

$$F_{vfo} = F_{xtal} \bullet M_2/N \qquad \text{(Eq 2)}$$

With these values the VFO frequency will be 5,000,000 Hz.

The next stable point is when N=N+1, so N=500,001. Again using Eq 2, this results in a VFO frequency of 4,999,990 Hz—giving a spacing of 10 Hz. (The Appendix shows that with this system the spacing varies with N and is only exactly 10 Hz with the numbers used above.)

Operation in Practice

During normal manual tuning of the VFO, the action of the controller is not noticed, as it is slow. After manual tuning, the frequency creeps to the nearest 10-Hz lock point, which is never further away than 5 Hz up or down. (A musician with absolute pitch might notice such a small change in a receiver CW tone, but I can't.)

Of course the action of the controller must be small to avoid overshoot of the frequency. Each corrective pulse to the integrator may change the VFO frequency a very little, less than 1 Hz. You don't want the VFO to "hunt."

Acceptable Distance between Lock Points

In the system described, the spacing between lock points is 10 Hz. I found that a spacing of up to 40 Hz is still acceptable. This means that there is quite some freedom in the design, allowing other VFO or crystal oscillator frequencies to be used.

In the practical system described below, the spacing between lock points varies between 15.9 Hz and 19.2 Hz when the VFO is tuned from 5 to 5.5 MHz.

Detailed Circuit

The practical circuit is shown in Fig 3. Two cascaded binary dividers, U1 and U2, divide the VFO frequency by 32768 in 15 cascaded stages (7 in U1 and 8 in U2). This forms the clock signal for the 74HC74 D flip-flop, U3A. (Only one of the two flip-flops in the IC is used.)

The crystal oscillator operates at 48 MHz. I used a 16-MHz crystal in third-overtone mode in a design taken from *Solid State Design for the Radio Amateur*.[2] I added a diode detector with which the peak-to-peak amplitude of the 48-MHz voltage can be measured.

Switches S1 and S2 of Figs 1 and 2 are formed by transistors Q1 and Q2. These are normally off. The on time is determined by the differentiating R-C networks in their bases and is less than 1 ms per pulse. Via the 4.7-MΩ resistors, the output of the integrator changes slightly after each pulse.

The output voltage of the integrator has a range of 0 to 10 V. After turn-on it starts approximately in the middle of its range. The small ripple on the output signal of the integrator caused by the correction pulses is smoothed by an R-C filter before it is connected to the varicap in the VFO.

In case the output should become saturated (0 or 10 V), it can be brought to midrange again by momentarily pressing S3. I advise mounting S3 on the front panel of the equipment in which the stabilizer is built. Press it now and then when you start to retune the radio.

No details are given of the VFO and associated buffer circuits, but the tuning range of the varicap in the VFO must be larger than the expected drift. A convenient sensitivity is 1 kHz/V. The phase of the action of the varicap is not important. (The frequency-locked loop will stabilize in the range of 0 Hz to maximum or in the range of maximum to 0 Hz of the output of the digital mixer.)

The amplitude of the VFO signal to the first digital circuit must be approximately 4 V p-p. The input impedance of the digital circuits is high.

The unit needs a supply voltage of 12 V, draws a current of some 20 mA and has its own 5-V regulator, U5, to supply the crystal oscillator and digital circuits with 5 V.

Construction

I used the "ugly construction" technique as described in *The 1995 ARRL Handbook*. (I prefer the name *amateur-type surface-mounted technology!*)

I placed the ICs upside down on the copper cladding of a piece of printed-circuit board. The supply connections of all ICs are decoupled with 10-nF ceramic capacitors that, together with the ground supply connections, keep the ICs in place. I used thin insulated wire to make all connections and soldered directly to the IC pins.

For the coil in the overtone oscillator circuit, I used a small RF choke of 0.22 µH. A 1/2-inch, 6-turn air coil wound on a 1/4-inch form worked fine also.

The 2.2-µF capacitor in the integrator circuit must be a type with little leakage—polycarbonate or polystyrene. The resistors connected to the input of U4 should be mounted close to the IC.

I suggest mounting the circuit in a small metal box to avoid any influence of RF signals from transmitters and to avoid interference to reception caused by the switching signals of the CMOS ICs.

Adjustment and Testing

The only adjustment is the 40-pF trimmer capacitor in the 48-MHz overtone oscillator. Oscillation can be confirmed by measuring the dc voltage at test point T, which must be about 4 V. The amplitude at this point can be adjusted with the trimmer; the precise frequency is not important.

When all the wiring is done the stabilizer should work. If not, the individual circuits can be tested in the following way. With a VFO frequency of 5 MHz the output of U1, pin 6 must be 39 kHz, and the outputs of U2, pins 15 and 14 must be 152 Hz and 38 Hz, respectively. Besides confirming proper operation of the ICs with an oscilloscope, a multimeter can be used as well. If U1 and U2 operate properly, their outputs will be square waves and a dc voltmeter will indicate 2.5 V on the outputs. If they do not work, their outputs will be 0 or 5 V.

The integrator and the action of the varicap in the VFO can be tested as well. Measure the dc output voltage of the integrator on test point V. The impedance at that point is low, so any multimeter can be used.

Disconnect the 4.7-MΩ resistors from Q1 and Q2 and press S3. The output voltage should be 5 V. Connect point A to ground for a few seconds. The output voltage should increase at a rate of 1 V every 4.4 seconds. When point A is open, the output should remain stable at the last value. Connecting point A to +12 V should give the opposite action (but slightly faster).

When the output voltage is changed from 5 to 6 V, the VFO frequency should change about 1 kHz. In normal

Appendix

Fig A shows a simplified block diagram of the frequency stabilization system. The simplification concerns the output of the frequency comparator, which is assumed to be 0 Hz here and which is a low frequency in the actual system. The difference is not significant for these calculations. The divider *N* of Fig A is not really present in the actual circuit; it symbolizes the ratio of the two frequencies at the inputs of the digital mixer.

The following relation holds for the frequency-locked loop:

$$\frac{f_v}{M} = \frac{f_x}{N} \quad \text{(Eq 1)} \Rightarrow f_v = \frac{f_x M}{N} \quad \text{(Eq 2)}$$

The next lock point exists when *N*=*N*+1. The difference *D* between lock points then becomes:

$$D = \frac{f_x M}{N} - \frac{f_x M}{N+1} = f_x M\left(\frac{1}{N} - \frac{1}{N+1}\right)$$

The term between brackets can be written as:

$$\frac{N+1-N}{N(N+1)} = \frac{1}{N(N+1)}$$

As *N* is always large compared with the 1 in the denominator, only a negligible error results if the term is simplfied to:

$$\frac{1}{N^2} \Rightarrow D = \frac{f_x M}{N^2} \quad \text{(Eq 3)}$$

From Eq 1 it also follows that:

$$N = \frac{f_x M}{f_v} \quad \text{(Eq 4)}$$

Substituting Eq 4 into Eq 3 results in:

$$D = \frac{f_v^2}{M f_x} \quad \text{(Eq 5)}$$

We can now fill in the values used in the practical system (Fig 3) with:

$f_v = 5.0\,\text{MHz}, \quad M = 2^{15}, \quad f_x = 48\,\text{MHz}$

These values result in the the difference *D* between lock points as:

$$D = \frac{\left(5 \times 10^6\right)^2}{48 \times 10^6 \times 2^{15}} = 15.98\,\text{Hz} \quad f_{clock} = \frac{5 \times 10^6}{2^{15}} = 152.58\,\text{Hz}$$

For a VFO frequency of 5.5 MHz we get:

$$D = \frac{\left(5.5 \times 10^6\right)^2}{48 \times 10^6 \times 2^{15}} = 19.23\,\text{Hz} \quad f_{clock} = \frac{5.5 \times 10^6}{2^{15}} = 167.84\,\text{Hz}$$

The frequencies of the correction pulses are $f_{clock}/4$: 38.14 Hz at 5 MHz and 41.96 Hz at 5.5 MHz.

Design Method for Other Frequencies

Given the frequencies of the VFO and of the overtone oscillator in MHz and the desired distance between lock points *D* in Hz, calculate *M* by:

$$M = \frac{f_v^2}{f_x D} \times 10^6$$

Select the nearest power of 2 and check if *D* is below the desired value according to:

$$D = \frac{f_v^2}{M f_x} \times 10^6$$

In case *D* deviates too much from the desired value, take the next higher value for *M* and repeat the calculation for *D*.

Example

A VFO operates between 37 and 38 MHz and the overtone oscillator operates at 80 MHz. The required distance between lock points, *D*, is 20 Hz. Calculate *M*:

$$M = \frac{38^2}{80 \times 20} \times 10^6 = 902.500$$

Select *M*=1,048,576 (20 cascaded binary dividers). Calculate *D* at 37 and at 38 MHz, resulting in 16.3 Hz and 17.2 Hz, respectively. As *D* is below the desired value of 20 Hz over the whole range of the VFO, the selected value of *M* is the correct one.

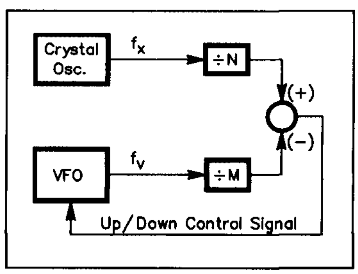


Fig A—A simplified block diagram of a frequency-locked loop.

operation, point A is left open; it is used for test purposes only.

Testing of the frequency discriminator can take place in the following way. Reconnect the 4.7-MΩ resistor to Q2. The integrator output should decrease at a rate of about 1 V every 100 seconds. The 4.7-MΩ resistor connected to Q1 has the opposite effect. The VFO frequency should become stable with both resistors in place and point A open.

(This 1 V per 100 s corresponds with a change in VFO frequency of 1 kHz per 100 s. This is the maximum drift rate that can be corrected. In 100 s, 3800 correction pulses have been generated, so the change in VFO frequency is 0.26 Hz per correction pulse.)

The values given above are typical values. Because of component tolerances, results of ±50% from the values given are possible and are fully acceptable.

Results and Further Experimentation

I have used the circuit in Fig 3 on a number of L-C oscillators, including one operating at 38 MHz. In all cases I was able to obtain virtually drift-free operation.

In this design I used 74HC integrated circuits. I found that 74HCT types also work well without any circuit modification.

According to the data sheets, the maximun toggle frequency of a 74HC74 is 76 MHz (59 MHz for a HCT version), and the maximum input frequency for a 74HC(T)4060 is 88 MHz. My ICs still operated at up to 95 MHz. This means that the circuit can be used with considerably higher VFO and crystal frequencies than mentioned so far. To verify this I have used the system on a 38-MHz L-C oscillator in combination with an 80-MHz overtone oscillator (of different design than that in in Fig 3). In this case M=1.048.576 (20 cascaded binary stages using the output of the first IC from pin 1 instead of pin 6. Be aware that in a 4060 IC not all divider ouputs are available.) At 38 MHz I got reliable operation with a spacing between the lock points of 17.2 Hz. After a cold start the loop stabilized within seconds and stayed on the same lock point for hours.

The quality of the L-C oscillator itself must already be very good: a T9 tone without appreciable hum, jitter, microphony, hand effects, etc, with long-term drift as the only deficiency.

There are two drawbacks to this stablization technique. One is that it is not easily possible to work split frequency or use an RIT function when the stabilizer is in operation. After a number of jumps from one frequency to the other, the VFO may lose its original lock point. Secondly, with a fixed value of M, the distance between lock points varies with the square of the VFO frequency, making the system less suitable for a multiband VFO. In such a case, M should be selectable, which increases complexity.

Conclusion

With this simple, low-cost circuit the long-term stability of HF L-C oscillators can achieve the stability of a crystal oscillator, while maintaining fine tuning steps.

Notes

[1]Spaargaren, Klaas, PA0KSB, *Ham Radio*, Dec 1977. (This is still published in the latest edition of the British *RSGB Radio Communication Handbook*.)

[2]*Solid State Design for the Radio Amateur*, published by ARRL.

Correlation of Sampled Signals

Correlation is a powerful signal-processing technique. Here's how to add it to your DSP toolbox.

By Jon Bloom, KE3Z

Digital signal processing (DSP) relies on numerous techniques, many of which stem from counterparts that have been used for years in analog signal processing. Among these are correlation techniques. These techniques are not widely understood among DSP experimenters, probably because their use in analog systems is not that common.

The correlation of two signals is essentially a measure of how alike the two signals are. You can probably imagine how useful such a measure can be. But computing a correlation value gives us some unique ways of using signals—ways that may not be quite so obvious. In this article, I'll introduce the concept of correlation, show how correlation values can be computed and provide an example of the use of correlation.

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

Correlation Concepts

Suppose we have two sequences of sample values, which we'll call sequence x and sequence y. Each of these sequences has N samples of a signal, with both signals having been sampled at the same sampling rate. Both sequences contain samples that have positive values, when the signal being sampled was above 0 V, and negative values, when the signal was below 0 V.

We can compute the *crosscorrelation* of these two sequences thus:

$$r_{xy} = x[0]y[0] + x[1]y[1] + ... + x[N-1]y[N-1] \quad \text{Eq 1}$$

where $x[0]$ is the first sample of sequence x, $x[1]$ is the second sample of that sequence and so on. r_{xy} is the crosscorrelation value. In more compact form, Eq 1 becomes:

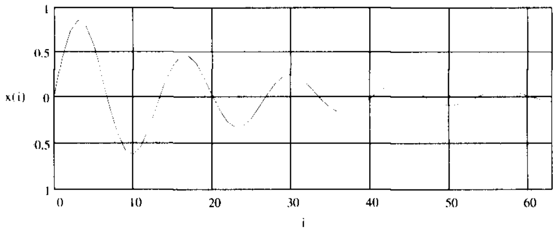
$$r_{xy} = \sum_{i=0}^{N-1} x[i]y[i] \quad \text{Eq 2}$$

Let's consider several possibilities. Suppose the two signals are exactly the same. In this case, each of the $x[i]$ values is the same as the corresponding $y[i]$ value. Since the two values must either both be positive or both be negative, the product of the two will always be positive, and the resulting sum will be a large positive number. Now suppose that the two signals are exactly the same but 180° out of phase. In this case, each $y[i]$ term will be the negative of the corresponding $x[i]$ term. Their product will always be negative, and the sum will be some

This first section just creates an arbitrary test signal of 64 samples, numbered 0 to 63.

$lsamp \quad 63$

$x_0(n) \quad \exp(\ 3\cdot n)\cdot\sin(30\cdot n)\cdot\Phi(n) \qquad i \quad 0..lsamp \qquad x(i) \quad x_0\left(\frac{i}{lsamp\ \ 1}\right)$



Now we compute the autocorrelation function of the signal—the correlation of the signal with itself, shifted in time. The horizontal axis is the number of samples of shift; the vertical axis is the autocorrelation of the signal.

$$r_{xx}(j) \quad \sum_{i=0}^{lsamp} x(i)\cdot x(i\ \ j) \qquad j \quad 0..lsamp$$

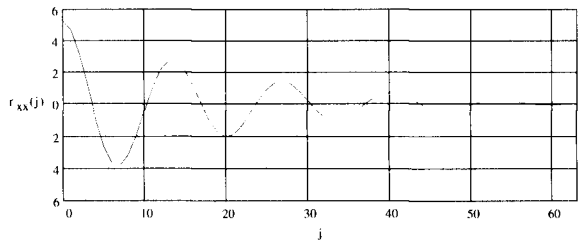
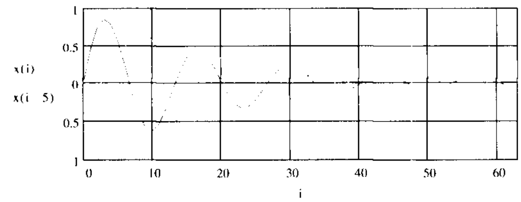
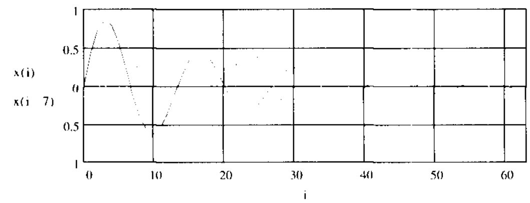


Fig 1

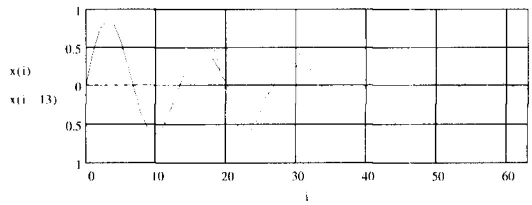
Here are some examples that show the unshifted signal, x(j), and the shifted signal, x(i-j), for various values of j.



for $r_{xx}(5)$
(j=5)



for $r_{xx}(7)$
(j=7)

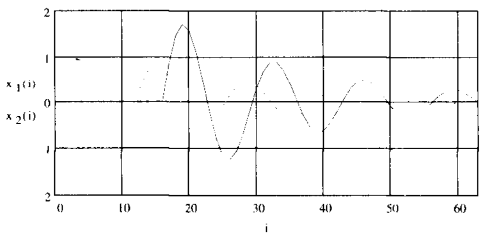


for $r_{xx}(13)$
(j=13)

Fig 2

Now we are going to use two signals. Each of the signals is a modified version of our original test signal. The x_1 signal is our original signal, shifted in time and scaled in amplitude. The x_2 signal is our original signal, scaled in time, but by a different amount than the x_1 signal.

$x_1(i) \quad 2\cdot x_0\left(\frac{i\ \ 16}{lsamp + 1}\right) \qquad x_2(i) \quad x_0\left(\frac{i\ \cdot\ 12}{lsamp + 1}\right)$



Here's the crosscorrelation of our two new signals, x_1 and x_2:

$$r_{x1x2}(j) \quad \sum_{i=0}^{lsamp} x_1(i)\cdot x_2(i\ \ j)$$

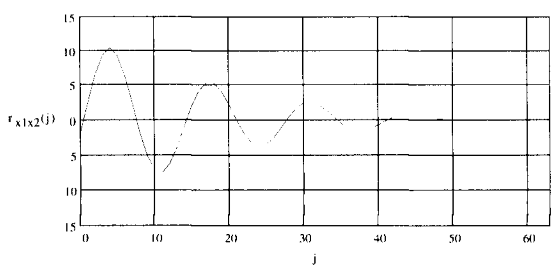
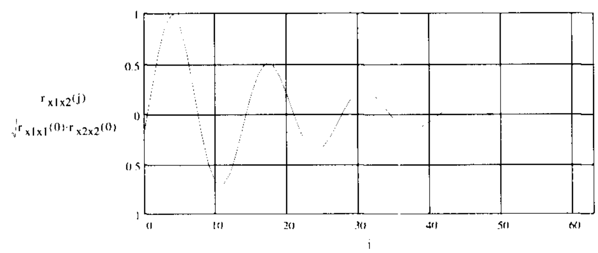


Fig 3

Now we compute the normalized crosscorrelation. To do this, we must first compute the autocorrelation of x_1 at j=0 and the autocorrelation of x_2 at j=0.

$$r_{x1x1}(j) \quad \sum_{i=0}^{lsamp} x_1(i)\cdot x_1(i\ \ j) \qquad r_{x2x2}(j) \quad \sum_{i=0}^{lsamp} x_2(i)\cdot x_2(i\ \ j)$$

$r_{x1x1}(0) = 20.862 \qquad r_{x2x2}(0) = 5.242$



Inspection of the normalized autocorrelation graph reveals that the maximum of the crosscorrelation function occurs when j=4, where the normalized crosscorrelation is:

$$\frac{r_{x1x2}(4)}{\sqrt{r_{x1x1}(0)\cdot r_{x2x2}(0)}} = 0.998$$

This tells us that: a) the difference in time (or the phase shift) between the two signals is 4 sample periods, and b) the signals are almost exactly alike since the value of the normalized crosscorrelation for this 4-sample shift is nearly 1.

Fig 4

large negative value. Finally, suppose the two signals are different. Some of the $x[i]y[i]$ terms will be positive, some will be negative, and the resulting sum will lie somewhere between the large positive value and the large negative value. This is the basic idea behind correlation. We are simply calculating a measure of the "sameness" of the two sequences.

One possible way the third case—comparing two different signals—could arise in practice is if the two signals were in fact the same but were out of phase by some value other than 180°. We can directly apply Eq 2 to measure how alike two sequences are only if the two signals are in phase. But it's likely that we don't know whether they are in phase. If we are bothering to calculate the cross-correlation, we may not even yet know if they are similar signals, much less anything about the relative phase!

This leads us to the need to calculate a crosscorrelation *function*. This means simply that we are going to calculate the crosscorrelation between the two signals, shift one of the signals by one sample, recalculate the cross-correlation, shift the signal again, recalculate, and continue until we have exhausted the samples. Essentially, we are shifting the y signal to the right each time and computing how well it "lines up" with the x signal. Mathematically, the crosscorrelation function will be:

$$r_{xy}(j) = \sum_{i=0}^{N-1} x[i]y[i-j] \qquad \text{Eq 3}$$

Here, j is the shift count, and we can calculate $r_{xy}(j)$ for any value of j between 0 and N–1. (Notice that when j is 0, Eq 3 simplifies to Eq 2.) When we know the two signals are the same, where $y[i]=x[i]$, we call the computed correlation the *autocorrelation*. Normally, we refer to the autocorrelation function of the sequence x as $r_{xx}(j)$.

It's a bit easier to see how correlation works by looking at autocorrelation. Fig 1 shows a *Mathcad* worksheet with an example of autocorrelation. The top graph shows a plot of the sequence values for a signal I made up to use in the example. (Note that, while the sample values only exist at integer sample numbers, the plot connects the values together to form a continuous line. It's just more "visual" to show the plot this way.)

The graph at the bottom of Fig 1 shows the computed autocorrelation function, $r_{xx}(j)$. As expected, $r_{xx}(j)$ is at its maximum positive value when j=0. But notice the periodic nature of the autocorrelation. Why this happens in this case is explained by Fig 2, which shows the two signals that are being multiplied and summed, $x[i]$ and $x[i-j]$, for various values of j. When j is 5, the two signals are out of phase by some amount, so $r_{xx}(5)$ is a value between the maximum positive value that occurs at j=0 and the most negative value. That most negative value occurs at about j=7 or so, and the second graph of Fig 2 shows that the $x[i]$ and $x[i-j]$ signals are just about completely out of phase when j=7. The third graph of Fig 2 shows that at j=13, the signals are about back in phase, and from Fig 1 we see that $r_{xx}(j)$ reaches a maximum again at this value of j. But the two signals aren't as similar as they were at j=0, so the maximum at j=13 is lower in value than the one at 0.

Normalization

Fig 3 extends our example. Here, our example signal has been used to form two signals of different amplitudes, as shown in the top graph, and the two signals are out of phase by four sample periods. When we compute the cross-correlation function of the two signals, as shown in the bottom graph, the maximum occurs at j=4. This shows that we can find the time (phase) difference between the two signals by computing the crosscorrelation function. But remember that our original purpose was to measure the "sameness" of the two signals. We now know that they are different in time by four sample periods, but what do we know about how alike their shapes are? Not much, really. The maximum of the crosscorrelation function has a value of just over 10, from the graph. This value depends very much on the amplitude of the signals, but usually we will want to ignore amplitude differences. It's the *shapes* of the signals we want to compare.

We can get around this problem by *normalizing* the amplitudes. That is, if we compute the amplitude of a signal, then use that information to create a scaling factor by which all of that signal's samples are multiplied, we'll adjust the amplitude of the signal to some standard value. That way, we'll always get the same result from our crosscorrelation function no matter what the amplitudes of the two signals are.

But what "amplitude" are we talking about? Peak amplitude? No, that doesn't work too well. Suppose that one of the two signals of Fig 3 had a brief noise spike right at its maximum amplitude point. That would significantly change the peak amplitude without really much changing the overall shape of the waveform. And we can't just take the average of the samples, because if the signal is ac-coupled the average will be 0. What we need is the *energy* in the signal.

This being a discrete-time (sampled) signal, we calculate the energy of a signal, x, so:

$$E = \sqrt{\sum_{i=0}^{N-1} x[i]x[i]} \qquad \text{Eq 4}$$

If we now divide each of the x sample values by the result of Eq 4, we will have normalized the signal. Similarly, we compute the energy in the y signal and divide its sample values by the result. But notice that the part of Eq 4 underneath the radical sign is just $r_{xx}(0)$. Now we can come up with an equation, from Eq 3 and Eq 4, that computes the *normalized crosscorrelation*:

$$r_{xy}(j) = \frac{1}{\sqrt{r_{xx}(0)r_{yy}(0)}} \sum_{i=0}^{N-1} x[i]y[i-j] \qquad \text{Eq 5}$$

And Fig 4 shows the computation of the normalized crosscorrelation of our signals of Fig 3. Now the maximum crosscorrelation occurs with a value very close to 1. (It's not exactly one because in the process of shifting the signals over, in Fig 3, I lopped off the ends of the signal, and I lopped off more of the x signal than I did of the y signal; they're not quite the same signal any more.) The result of Eq 5 will always have a maximum amplitude of 1 if the signals are identical, –1 if the signals are identical but one is inverted from the other, and something between 1 and –1 if the signals are different. And the value of j at which the maximum occurs tells us the time difference between the two signals.

Fixes

There is one remaining problem. In Eq 3, we are shifting the y signal to the right and calculating how well it "lines up" with the x signal. But what if the y signal was *lagging* the x signal by a small amount? We'd never see the point where the two signals line up because we never shift the y signal left. One simple solution to this problem would be to also plug negative values of j into Eq 3, letting j range from –(N–1) to N–1. But that means we'd have to do twice as many calculations.

Notice that Eq 3 results in use of negative indices of the y sequence, when $i-j$ becomes less than 0. So far, we've treated the value of any such sample as zero. But Fig 5 shows a different approach. Here, we copy the values of the y signal samples into the negative indices of the y sequence. This effectively *wraps* the y signal around. Now when we use Eq 3, the maximum of the crosscorrelation will occur near the right side of the graph if the y signal lagged the x signal, as shown by the bottom graph of Fig 5. What we are doing can be thought of as treating the signals as periodic ones, where the N samples of each signal are preceded and followed by copies of the same N samples.

A Need for Speed

So, now we have a way of assigning a number to how "alike" two signals are and of measuring the time difference between them. This is neat stuff! But we paid a heavy price in processing time. Directly implementing Eq 3 requires a *lot* of computation! (Open these example worksheets in *Mathcad* and see how long they take to calculate!) It would sure be nice to find a quicker way to do these calculations.

The DSP-aware among you won't be surprised to hear that the solution to the speed problem involves the fast Fourier transform (FFT) algorithm. Showing why that's so would take more math than we need to involve ourselves with. I'll just get right to the result:

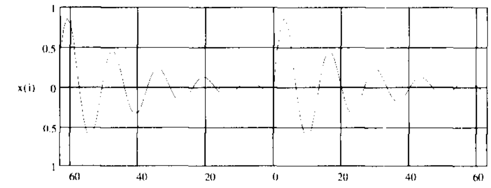
$$r_{xy}(j) = F^{-1}\left[X^*(j)Y(j)\right] \quad \text{Eq 6}$$

The same test signal is now created in a "wraparound" version, where x(-i)=x(N-i)

$nsamp := 64 \quad lsamp := nsamp - 1$

$x_0(n) := \exp\left(-3\cdot\frac{n}{nsamp}\right)\cdot\sin\left(30\cdot\frac{n}{nsamp}\right)\cdot\Phi(n) + \exp\left(-3\cdot\frac{nsamp+n}{nsamp}\right)\cdot\sin\left(30\cdot\frac{nsamp+n}{nsamp}\right)\cdot\Phi(-n)\cdot\Phi(n+nsamp)$

$i := -lsamp .. lsamp \quad x(i) := x_0(i)$



$x_1(i) := 2\cdot x_0(i-12) \quad x_2(i) := x_0(i-16)$ Shifted, scaled signals (like Fig 3, but x_2 lags x_1)

$r_{x1x2}(j) := \sum_{i=0}^{lsamp} x_1(i)\cdot x_2(i-j) \quad j := 0 .. lsamp$

$r_{x1x1}(j) := \sum_{i=0}^{lsamp} x_1(i)\cdot x_1(i-j) \quad r_{x2x2}(j) := \sum_{i=0}^{lsamp} x_2(i)\cdot x_2(i-j)$

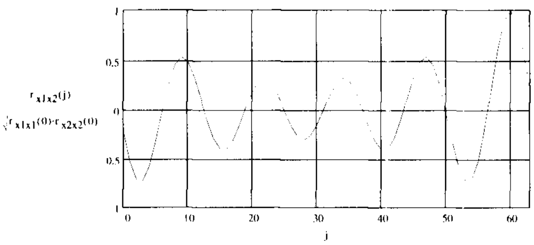


Fig 5

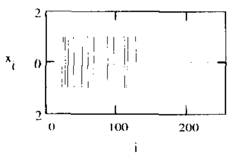
Read a 128-sample sync vector (32 bits x 4 samples/bit) into a 256-element vector (array) and make a second 256-element vector that holds a shifted copy of the sync vector.

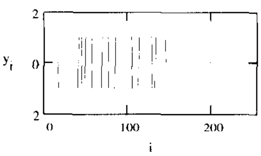
$nsamp := 256 \quad lsamp := nsamp - 1 \quad fsamp := 127$

$i := 0 .. lsamp$

$x_i := 0 \quad y_i := 0$

$x_i := READ(corr) \qquad k := 18 .. lsamp \quad y_k := x_{k-18}$





Compute the autocorrelations of the two vectors. This will be used to compute the normalized crosscorrelation.

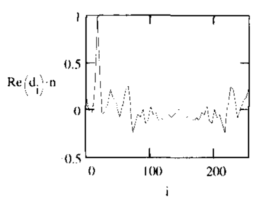
$r_{xx}(j) := \sum_{i=0}^{lsamp} x_i\cdot x_{i-j} \qquad r_{yy}(j) := \sum_{i=0}^{lsamp} y_i\cdot y_{i-j}$

Compute the crosscorrelation using FFTs.

$a := cfft(x) \quad b := cfft(y) \qquad n := \frac{1}{\sqrt{r_{xx}(0)\cdot r_{yy}(0)}}$ Normalization factor for the crosscorrelation

Note that we have to scale the result of the IFFT. This is because Matchcad scales by 1/sqrt(N).

$c_i := a_i\cdot\overline{b_i} \quad d := \sqrt{nsamp}\cdot icfft(c) \qquad e_i := \overline{a_i}\cdot b_i \quad f := \sqrt{nsamp}\cdot icfft(e)$



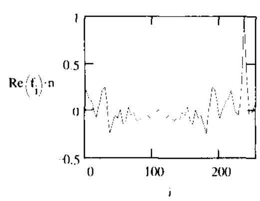
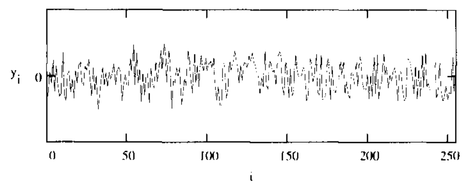


Fig 6

$noiseamp := 6$ Peak-to-peak noise amplitude

$y_i := rnd(noiseamp) - \frac{noiseamp}{2}$

$k := 18 .. lsamp \quad y_k := y_k + x_{k-18}$

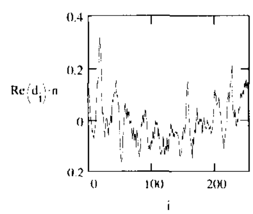


$r_{yy}(j) := \sum_{i=0}^{lsamp} y_i\cdot y_{i-j}$ Recompute the autocorrelation of this new signal.

Compute the crosscorrelation using FFTs.

$a := cfft(x) \quad b := cfft(y) \qquad n := \frac{1}{\sqrt{r_{xx}(0)\cdot r_{yy}(0)}}$ Normalization factor for the crosscorrelation

$c_i := a_i\cdot\overline{b_i} \quad d := \sqrt{nsamp}\cdot icfft(c) \qquad e_i := \overline{a_i}\cdot b_i \quad f := \sqrt{nsamp}\cdot icfft(e)$



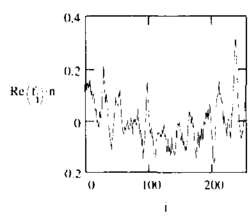


Fig 7

where $X^{*}(j)$ is the complex conjugate of the jth bin of the FFT of $x[n]$ and $Y(j)$ is the jth bin of the FFT of $y[n]$. F^{-1} is the inverse FFT function. In other words, to compute the crosscorrelation function, we first take the FFT of the x signal and compute the complex conjugate of each bin (by reversing the sign of the imaginary part) to get X^{*}. Then we take the FFT of the y signal to get Y. We then multiply (using complex arithmetic) each bin of X^{*} by the corresponding bin of Y. Finally, we take the inverse FFT of those products. The real part of the result will be $r_{xy}(j)$. (The imaginary part will always be 0.) The value of $r_{xy}(j)$ from Eq 6 is unnormalized, but we can normalize it the same way we did in Eq 5.

One of the nice features of this technique is that the FFT by nature treats the signal as a periodic one. The fix we applied to deal with the lead-lag problem is automatically included in the FFT calculations.

An Application Example

Some time ago, Phil Karn, KA9Q, described some concepts he was working on to develop a new link-layer protocol for amateur digital communication.[1] One of the elements of Phil's system was the use of a 64-bit *sync vector*, a string of bits that denotes the beginning of a frame.[2] To demonstrate how such a thing might be used, I've set up the example shown in Figs 6 and 7. To keep the graphs from getting too crowded, I use a 32-bit vector, and I set it up so that the signal was sampled at 4 times the bit rate (four samples per bit).

In Fig 6, two copies of the sync vector samples are used, with one (the y signal) lagging the other (the x signal) by 18 samples. At the bottom of Fig 6 are two graphs of the crosscorrelation function computed from these two signals, computed using FFTs. The left graph shows that the maximum of the crosscorrelation function occurs at j=18 and is equal to 1, as we would expect. The right graph was generated by changing which of the two intermediate FFTs was conjugated. This shows that you can select which of the two signals is the reference signal by simply selecting the FFT to conjugate.

Fig 7 is similar to Fig 6 except that I've added noise to the y signal—a lot of noise. The peak-to-peak amplitude of the noise is three times that of the signal. Now when we compute the crosscorrelation, we get the result shown at the bottom of Fig 7. Notice that we still have a maximum at j=18; we still can detect where the sync vector bits are in that noisy input signal. Now, though, the maximum crosscorrelation is significantly less than 1, because the clean sync vector and the noisy input signal aren't much alike any more.[3]

Conclusion

Correlation has many applications in DSP. I hope this article provides enough background to get you started exploring these applications.

The *Mathcad* worksheets shown here can be downloaded from the ARRL BBS (860-594-0306) or via the Internet from **http://www.arrl.org/files/qex** or **ftp://ftp.arrl.org/pub/qex** in file qexcorr.zip.

References

[1]Karn, P., KA9Q, "Toward New Link-Layer Protocols," *QEX*, June 1994, pp 3-10.

[2]Here we use the term *vector* in its linear algebra sense, meaning a one-dimensional matrix, or array, of values.

[3]In *Mathcad*, these graphs will change each time you calculate the document because the random noise changes.

Upcoming Technical Conferences

The Central States VHF Society Conference

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

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

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

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

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

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

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

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

1996 ARRL and TAPR Digital Communications Conference

The 15th ARRL and TAPR Digital Communications Conference will be held September 20-22, 1996, in Seattle, Washington (minutes from SeaTac airport).

It's that time again! Time to make your travel plans and put the finishing touches on your work for the upcoming 15th Annual ARRL and TAPR Digital Communications Conference.

The ARRL and TAPR Digital Communications Conference is an international forum for radio amateurs in digital communications, networking, and related technologies to meet, publish their work, and present new ideas and techniques for discussion. Presenters and attendees will have the opportunity to exchange ideas and learn about recent hardware and software advances, theories, experimental results, and practical applications. The Digital Communications Conference is not just for the digital expert, but for digitally orientated amateurs of all levels of experience.

The 1996 ARRL and TAPR Digital Communications Conference will be held on September 20-22, 1996, in Seattle, Washington. This year's conference location is just minutes away from the SeaTac (Seattle/Tacoma) Airport.

Not only is the Digital Communications Conference technically stimulating, it is a weekend of fun for all who have more than a casual interest in any of the ham digital communications modes. This includes BBS operators, networkers, DX-Cluster Sysops, software writers, modem designers and digital satellite communications enthusiasts. The ARRL and TAPR Digital Communications Conference is for all levels of digital operators—a *must* conference to attend to get active on a national level. Now, more than ever, Amateur Radio needs this great meeting of the minds, since it is important that we demonstrate a continued need for the frequency allocations we now have by pushing forward and documenting our achievements. The ARRL and TAPR Digital Communications Conference is one of the few ways to record our accomplishments and challenge each other to do more.

The conference is not just for the digital expert. This year's conference will again provide an entire morning with beginning and intermediate presentations on selected topics in digital communications. Some of the topics will include: APRS, satellite communications, TCP/IP, digital radio, spread spectrum and other introductory topics. Come to the conference and hear these topics presented by the experts! Don't miss this opportunity to listen and talk to others in this area.

In addition to the presentation of papers on Friday and Saturday, three workshops will be held during the conference. On Friday, Keith Sproul, WU2Z, will hold a workshop on APRS packet-location software. Keith is the Chair of the TAPR APRS Special Interest Group, developer of the Macintosh version—and more recent co-developer of the *Windows 95* version—of APRS and a leader in the area of APRS technology. This is a unique opportunity to gain insight into this fast growing new digital aspect of amateur operations that combines computers, packet radio and GPS (Global Positioning Satellites). On Sunday, Dewayne Hendricks, WA8DZP, will conduct a workshop focusing on how to utilize Part 15 wireless radios for ham applications. Dewayne is an expert in the area of commercial wireless systems; his company, WarpSpeed Imagineering, focuses on wireless Internet connectivity. This workshop presents an opportunity to learn how Personal Communications Technology (handheld and small-business wireless systems) can be used in the amateur service. A second Sunday workshop will focus on wireless networking using the WA4DSY 56K RF modem technology. This workshop will focus on the technology and accessories for creating and maintaining 56K networks using the WA4DSY modem and equipment compatible with it such as routers, digital driver cards, transverters and repeaters. Use of WA4DSY 56K equipment in the 219-220 band will also be discussed.

ARRL and TAPR especially welcome papers from full-time students to compete for the first annual student papers award. Two $500 travel awards will be given, one in each of the following categories: a) best technical/theory-oriented paper by a student, and b) best educational or community-oriented application paper by a student. The paper should relate directly to a wireless digital communication topic (see the guidelines at **http://www.tapr.org/** for more information). Papers coauthored by educators or telecommunications professionals are also eligible for this award, as long as a student is the first author. First-year awards have been funded through a grant by The ARRL Foundation, Inc. Deadline for receipt of finished student paper manuscript is June 11, 1996. Please note that this deadline is different from the general conference submission date. For full details and paper guidelines contact TAPR or check **http://www.tapr.org**.

Call for Papers—Anyone interested in digital communications is invited to submit a paper for publication in the Conference Proceedings. Presentation at the Conference is not required for publication. If you know of someone who is doing great things with digital communications, be sure to personally tell them about this! Papers are due by July 23, 1996, and should be submitted to Maty Weinberg, ARRL, 225 Main Street, Newington, CT 06111 or via the Internet to **lweinberg@arrl.org**. Information on paper submission guidelines are available on-line (**http://www.tapr.org**).

Local Co-Hosts—The 1996 ARRL and TAPR Digital Communications Conference is co-hosted by the Puget Sound Amateur Radio TCP/IP Group and Boeing Employees Amateur Radio Society (BEARS).

The Puget Sound Amateur Radio TCP/IP group is an informal group involved in an ongoing project to build and expand an amateur radio digital network in the greater Puget Sound area of the Pacific Northwest US. The Washington Experimenters TCP/IP Network (WETNET) uses TCP/IP as its primary transport protocol and currently has over 250 users. WETNET is linked to other amateur radio TCP/IP networks via the Internet. The Boeing Employees Amateur Radio Society (BEARS) is a general-interest amateur radio club for employees of the Boeing Company, headquartered in Seattle, Washington. The BEARS are an active amateur club, supporting radio classes, VHF/UHF repeaters and digital communications. BEARS has been instrumental in the construction of the Evergreen Intertie, an extensive network of interconnected repeaters in the Pacific Northwest.

What can you expect during the 1996 ARRL and TAPR Digital Communications Conference ?—A full day of papers and breakouts on Saturday for the beginner to the advanced amateur digital enthusiast.

Three workshops (see below): Friday (4 pm)—APRS, Conducted by Keith Sproul, WU2Z; Sunday (8 am)—How to Utilize Part 15 Radios for Ham Applications, Conducted by Dewayne Hendricks, WA8DZP; and Sunday (noon)—Wireless Networking using the WA4DSY 56K RF Modem.

Technology: The first annual Student paper session; a banquet with special guest speaker Lyle Johnson, WA7GXD (Lyle was one of the founders of TAPR and was instrumental in forming many of the current aspects of amateur digital communications. He is currently very active in building several digital aspects of the upcoming Phase 3D satellite); SIGs (Special Interest Groups) on Saturday following the banquet; informal get-togethers throughout the weekend; a meeting facility that is perfect for this type of meeting; vendor area and informal engineering discussions/demonstrations; an event at which the most important new developments in amateur digital communications are announced; digital "movers and shakers" from all over the world in attendance; and plenty of Washington state hospitality!

Conclusion—If you have attended a Digital Communications Conference in the past, just remember back to how much fun it was discussing the latest developments into the wee hours! If you have never been, then make your plans now to attend and find out how much fun the Digital Communications Conference can be. There are few activities where your participation can be so much fun and so important! What a great way to share and renew your enthusiasm for digital Amateur Radio! And enjoy getting together with colleagues from all over the world and bringing each other up to date on your latest work. All this, and more, for an unforgettable weekend of ham radio and digital communications. Make your travel and lodging arrangements now. We hope to see you at the ARRL and TAPR Digital Communications Conference on September 20-22!

Full information on the conference and hotel information can be obtained by contacting Tucson Amateur Packet Radio, 8987-309 E. Tanque Verde Road #337, Tucson, AZ 85749-9399. Phone: (817) 383-0000. Fax: (817) 566-2544. Internet: **tapr@tapr.org** Web: **http://www. tapr.org/**. [Note: If you need handouts or flyers for meetings contact TAPR about getting what you need!]

Hotel Information—Conference presentations, meetings and workshops will be held at the Quality Inn Seattle Airport, a complex co-located with the Radisson Hotel Seattle Airport. Rooms rates are $66/single-double and $76/triple. When making reservations with the hotel, be sure to indicate you are attending the ARRL and TAPR DCC conference. It is highly recommended that you book your room prior to arriving—a block of 75 rooms is reserved until September 6th, 1996. After the 75 rooms are booked, rooms will only be available in the Radisson hotel, but at a higher price. Be sure to book your rooms early! The hotel provides transportation to and from SeaTac Airport. You should contact the hotel to arrange airport transportation.

Quality Inn Seattle Airport (conference hotel), 17101 Pacific Highway South, Seattle, WA, 98188; tel: (206) 246-7000, fax (206) 246-1715.

Radisson Hotel Seattle Airport (alternate hotel), 17101 Pacific Highway South, Seattle, WA, 98188, tel: (206) 244-6000, fax (206) 246-6835.

Registration—Contact the TAPR office by phone, fax or e-mail (Internet: tapr@tapr.org) to preregister or for additional meeting information. MasterCard and VISA accepted. Preregistration (before Sept 1st) is $40.00; late registration or at the door is $45.00. (Conference registration includes a copy of the Conference Proceedings, sessions, meetings and lunch.) Saturday Evening Dinner (limited space) is an additional $19.00. (Dinner speaker: Lyle Johnson, WA7GXD.)

Workshops—*APRS Workshop:* Friday, 4 pm to 7 pm. Conducted by Keith Sproul, WU2Z. Registration $15.00; *How to Utilize Part 15 Radios for Ham Applications Workshop:* Sunday, 8:00 am to 11:00 am. Dewayne Hendricks, WA8DZP. Registration $15.00; *Wireless Networking using the WA4DSY 56K RF Modem Technology Workshop:* Sunday, noon to 3 pm. Registration $15.00. □□

Feedback

Two errors appeared in my article, "A Variable IF Selectivity Unit," November 1995 *QEX* (originally printed in September 1995 *Radio Communication*). In the Component List, under *Inductors*, RFC does not appear in the circuit and should be deleted, and under *Semiconductors*, TR1, TR2 should read MFE201. —*A. R. Thompson, GM3AHR* □□