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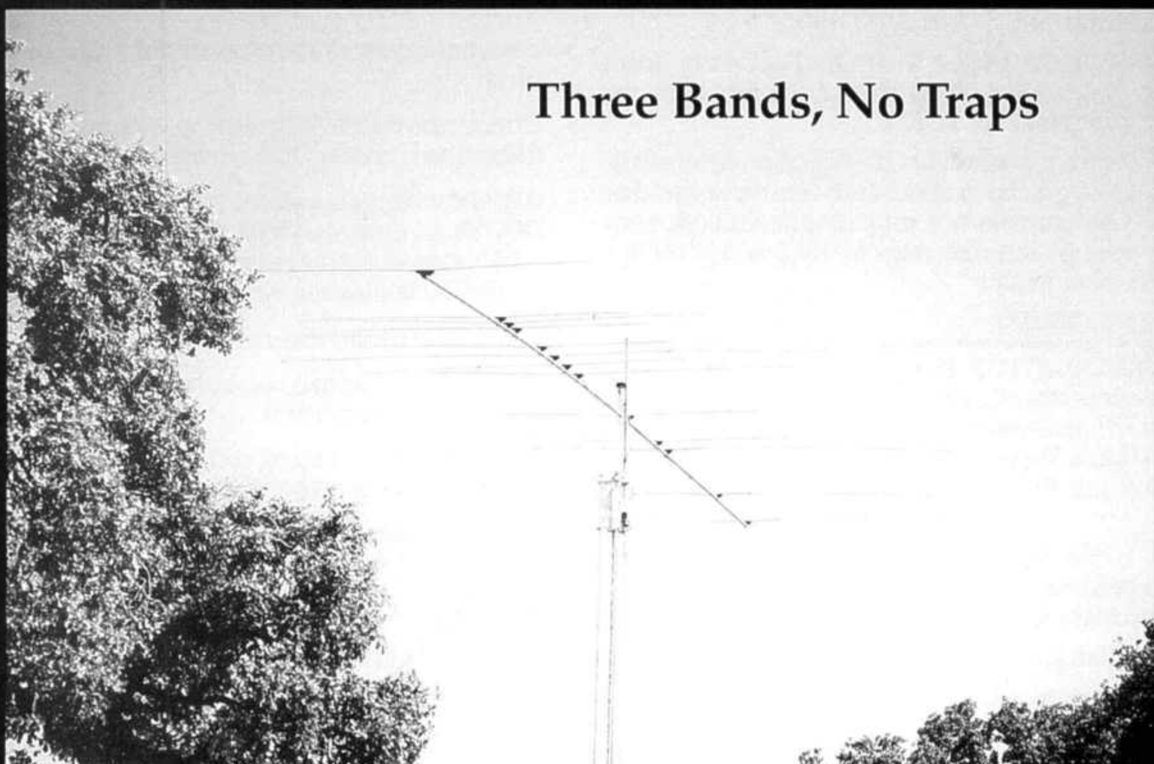

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About the Cover

Designing a tribander requires compromises in performance? Not the way K5RR has done it.

ISSUE NO. 176

Features

Columns

September 1996 QEX Advertising Index

Empirically Speaking

An HF Simulator for Amateurs

In the November 1994 "Empirically Speaking," we noted that the denizens of TAPR's HFSIG group were working on producing an HF simulator using DSP. It's past time that we reported to you that these efforts have met with some substantial success.

Building on work done by Alexander Kurpiers, DL8AAU, and Juergen Hasch, DG1SCR, Johan Forrer, KC7WW, has produced a simulator program that runs on the TAPR DSP-93 unit. Johan's software simulates standard CCIR test conditions and will distort a baseband signal in much the same way as if it had propagated via the ionosphere. The software is available at **ftp://ftp.tapr.org/tapr/SIG/hfsig/upload/dsp93sim.zip**.

Johan et al have done some outstanding work here, but there is still work to be done. For one thing, it would be useful to compare modem test results using the DSP-93 simulator to those obtained through a "professional" simulator, to validate the DSP-93 implementation. For another, it would be nice to port the software to other common DSP platforms.

Of course, the ionosphere is a complex medium, and the CCIR standard test conditions describe only a subset of the possible propagation effects, although a useful subset. The simulations performed in these DSP-93 programs probably represent about as complex a simulation as the hardware platform is capable of supporting. That suggests that a port to another DSP platform, one with a faster DSP chip, would allow simulation of additional concurrent propagation modes. Doing so would allow the simulation to step closer to the real ionosphere's effects. And the closer we can get to the real world, the better we can compare one HF modem to the next under all of the propagation conditions that might actually be encountered.

While we're on the subject, you may want to peruse the other files in the **/tapr/SIG/hfsig/upload** directory of **ftp.tapr.org**. Along with the HF simulator are a number of useful packages and snippets of code for HF digital systems, much of it contributed by KC7WW. Among the interesting items are torlib1.zip, the source code for Johan's PCTOR and PC-Pactor programs; ecc.zip, a collection of source to implement vaious error-correcting code schemes; and remez.zip, the well-known FIR filter design program. Give them a look.

This Month in *QEX*

Triband Yagi antennas for 20, 15 and 10 m have a long and venerable history. But most triband designs compromise antenna performance to achieve their three-band capability, either because they use traps to achieve multiband operation or because interactions between elements limit performance. It doesn't *have* to be that way. Using computer-based design tools, Richard C. Fenwick, K5RR, has designed "A High-Performance Triband Beam with No Traps" that gives monoband performance in a triband design.

These days, DSP audio signal processors (ASP) are common in ham shacks. Even though the operation of these ASPs is driven by software, almost all of the units are "store bought" units, as hams haven't, by and large, acquired the means of doing home-built DSP. That problem is solved with " A DSP-Based Audio Signal Processor," by Johan Forrer, KC7WW. His ASP software runs on a low-cost, commonly available Analog Devices DSP evaluation board and provides a good tool for learning how this processing is done—or for modifying the code to implement your own processing ideas.

In this month's "RF" column, Zack Lau, KH6CP/1, presents a 3456-MHz transverter design, an improvement on previous no-tune designs for this band.

It's conference season. "Upcoming Technical Conferences" tells you what's coming up, and "Conference Proceedings Available" tells you what you've missed that's worth catching up.—*KE3Z, email: jbloom@arrl.org.*

A High-Performance Triband Beam with No Traps

Computer-design tools show the way to a trapless tribander that performs well.

By Richard C. Fenwick, K5RR

When the traps in my new triband beam arced over the first time I operated when it was raining—during a contest, of course—I decided that I had to try to design my own triband beam. I had recently purchased Brian Beezley's brilliant *Yagi Optimizer* 6.5 (YO) and *NEC/Wires* 2.0 (NW) software, and the design described herein is the result of countless hours of analysis using these programs. In addition to the elimination of traps, I wanted high front-to-back and front-to-side ratios, which are not commonly found in commercial tribanders. I also wanted optimization in the phone bands. And finally, I wanted a gain of at least 7 dBd.

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Dallas, TX 75230

I doubt that anyone knows what the gain is for any commercial trapped tribander, as it is exceedingly difficult to calculate or measure.

Electrical Design

I found that a 4-element Yagi is a good choice for providing the desired performance. The resulting design effectively uses separate interlaced 4-element Yagis for each band. Interlaced Yagis typically suffer from considerable degradation in performance due to interaction unless special measures are taken. A 15-m beam will be affected by 20-m elements and a 10-m beam will be affected by both 20 and 15-m elements. Most of my efforts were directed at mitigating these effects.

Fig 1 shows the final electrical design. Using YO, I found that the effects of 20D2 on the 15-m beam could be compensated by moving 15D1 closer to 20D2 compared to where it would be if you scaled the dimensions of the 20-m beam. Optimum element lengths are unaffected. Using NW, I found the effect of 20D2 on the 10-m beam was minimal if it was placed between the 10-m reflector and driven element, and the effect of 15D2 was minimized if an open-sleeve dipole configuration was used for 10D1 and 15D2. The length and spacing of the 10-m sleeve elements were found to be extremely critical. The optimum half-length of 15D2 is about one inch longer than it would be in a 15-m beam by itself. The 10-m sleeve elements are equivalent to a 96.45-inch single half element in a 10-m beam by itself.

The antenna design was optimized for a height of 32 feet. Ordinarily, you'd want a much greater height, at least 70 feet, but I was building the antenna to be installed on an existing 32-foot tower at the home of my brother Bob, K6GX, located on a 550-foot hill overlooking San Francisco Bay. Calculated antenna patterns with the dimensions in Fig 1 are excellent at 70 feet at optimum frequencies, as shown in Fig 2. These calculations were done at 40 segments per half wave for actual ground with dielectric constant 13 and conductivity 5 mS/m, and 6061-T6 tubing. I doubt that any commercial triband beam will provide patterns remotely resembling these. I believe that you'd have to put monoband beams on separate towers to get front-to-back ratios of 30 dB on 10 and 15 m. The gain on 20 and 15 m is enhanced in the array environment by about 0.5 dB relative to optimized antennas by themselves, whereas in stacked arrays there is gain degradation on 10 and 15 m. (K4VX has calculated that a 6-element KLM 20-m beam degrades the gain of a 6-element KLM 15-m beam by 0.5 dB when spaced 10 feet apart, for example.)

An actual antenna would be built with tapered elements, of course. The W6QHS taper correction in YO was used to design the elements. Hairpin matching was used, also designed using YO.

Mechanical Design

This is a large antenna, using a 57-foot boom, so particular attention had to be paid to the boom construction. The 3-inch diameter boom from a KLM 20M6 was used, with 13 inches cut off the 20-m reflector end. Use of side bracing on the boom is highly recommended, but I wasn't able to use it owing to the mounting of the antenna on a tower using a Glen Martin Hazer mount. The boom was mounted on the mast about 27 feet from the 20-m reflector end, which appears to provide reasonably good weight and wind balance.

KLM element-to-boom brackets were used for all 13 elements. The element tubing came from five different antennas, mostly KLM. I doubt that anyone else constructing this antenna would use the element designs that I used, so I have not included the element design information in this article. I suggest that anyone attempting to build this antenna use the taper correction available in YO to finalize the element design.

Force 12 current-type baluns with 3-inch leads were used. Hairpins were constructed of 3/8-inch aluminum tubing on 1-inch centers, with 1/2-inch shorting straps above and below the tubes. Hairpin lengths were 20.5 inches for 20 m, 14.3 inches for 15 m and 13.5 inches for 10 m, measured from the center of the elements to the near edge of the shorting straps.

Measured Performance

I initially designed the antenna for matched frequencies of 14.210, 21.260 and 28.420 MHz. Upon erecting the antenna I found that the frequencies of lowest VSWR were 14.260, 21.300 and 28.725 MHz. Minimum VSWR was 1.2:1 or less on each band, which was a delightful surprise. This was measured at the base of the tower with a MFJ 259 SWR analyzer through 60 feet of RG-213 coax, more or less, and an Ameritron RCS4 coax switch. (Incidentally, I have found the MFJ 259 to be quite accurate and the RCS4 to be bulletproof.) I was satisfied with the minimum VSWRs, but not with the resonant frequencies, which I presume were not as intended owing to errors in the taper correction process. I then increased the lengths of all 20-m half-elements by 1/2 inch and all 10-m half-elements by 1 inch. The resulting VSWR curves are shown in Fig 3. I thought these were pretty good for an antenna that had gone up the tower only twice. In retrospect, I probably should have lengthened the 15-m elements as well, except for the driven element. The bandwidths are very close to that which was expected, and are also very similar to what is calculated on YO for equivalent single-band beams alone. A second VSWR minimum was observed on 10 m below 28 MHz.

I next measured the front-to-back ratios by transmitting a test signal into a second antenna located near ground level about 800 feet away and receiving on the triband beam. The MFJ 259 was used as a signal generator in these tests, a function which it performs well. Signal levels on the receiver S meter were noted as the triband beam was rotated. The S meter was then calibrated with a precision variable attenuator at each frequency and signal level at which a measure-

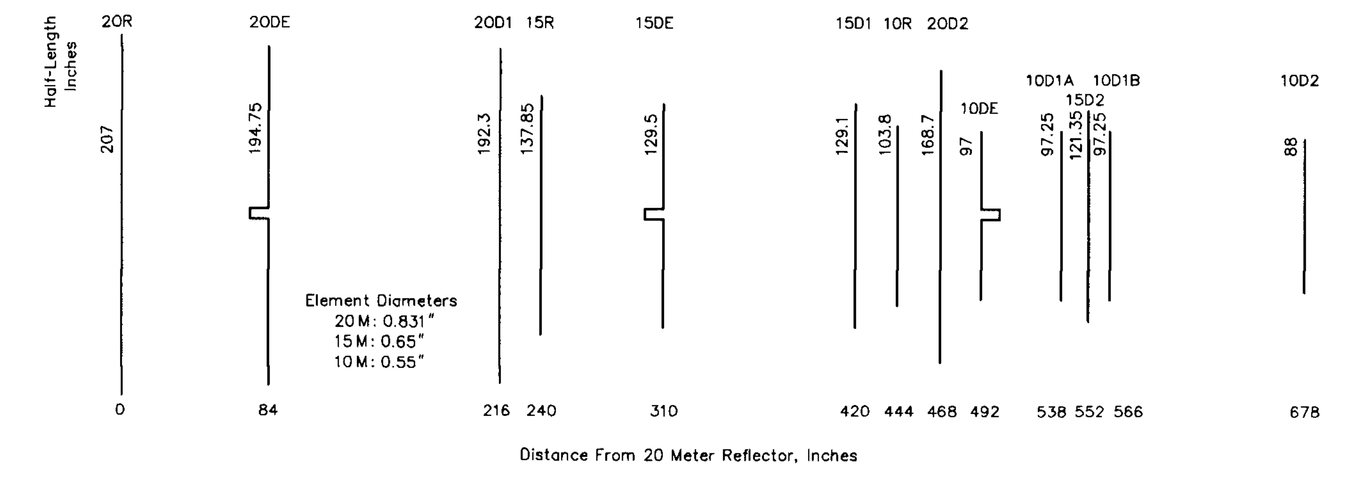

Fig 1—Dimensions of the high-performance triband beam.

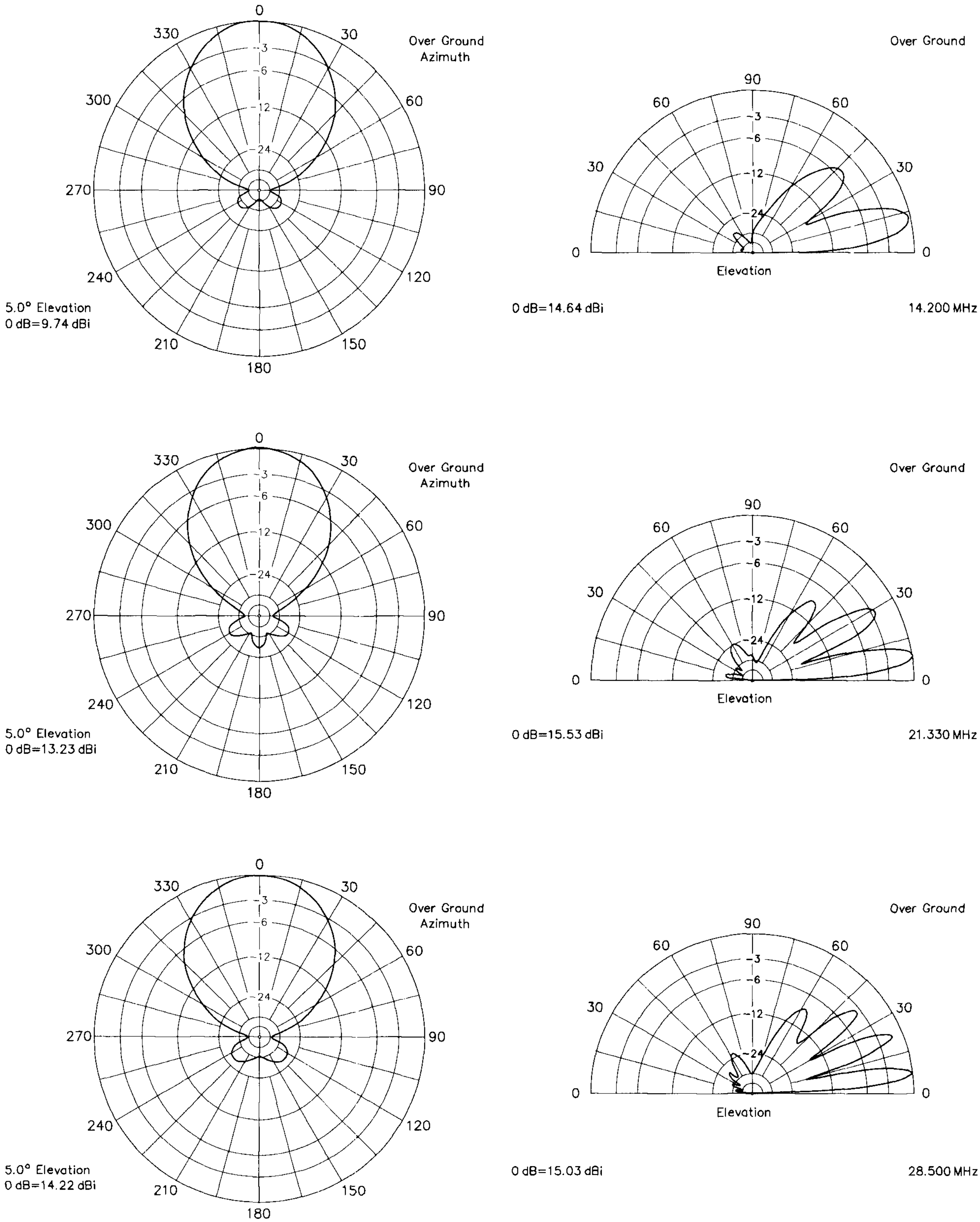

Fig 2—Calculated radiation patterns of the high-performance triband beam at a height of 70 feet.

ment was made. I believe the measurements to be accurate to within 1 dB. These measurements would apply to takeoff angles near the horizon. The results are shown in Fig 4, along with calculated values from NW, again using 40 segments per half wavelength and actual ground. The agreement is surprisingly good. Some error would be introduced into the 10-m front-to-back measurements because the 10-m beam was not being rotated about its center. The 20 and 15-m beams weren't either, but the magnitude of the problem would be much less.

Although the antenna was designed for the phone bands, it is certainly useable on the CW bands if an antenna tuner is employed. The front-to-back ratios are no worse than for 4-element beams alone when optimized for phone and probably better than most trapped triband beams.

Rear sidelobe levels were also noted as the beam was rotated. These measured about –25 dB on 15 m and –27 dB on 10 m, very close to the calculated values.

Over-the-air results have been very gratifying. At 14.2 MHz, the front-to-back ratio appears to be in the order of 40 dB and the front-to-side ratio in the order of 50 dB. I've never seen these kind of patterns in monoband beams, let alone tribanders, in 45 years of hamming.

Conclusion

The performance of this antenna met or exceeded all of my objectives, although a little more bandwidth would have been appreciated. I hope others will build it, and I will be happy to lend advice and assistance to those needing it. I am excited about the accuracy of *NEC / Wires*. This software and current desktop computers provide a powerful tool to the amateur antenna designer which not too many years ago could only have been dreamed of.

The beauty of all this is that you don't have to accept my claims for the performance of this tribander. You can analyze it on your own computer and verify the results.

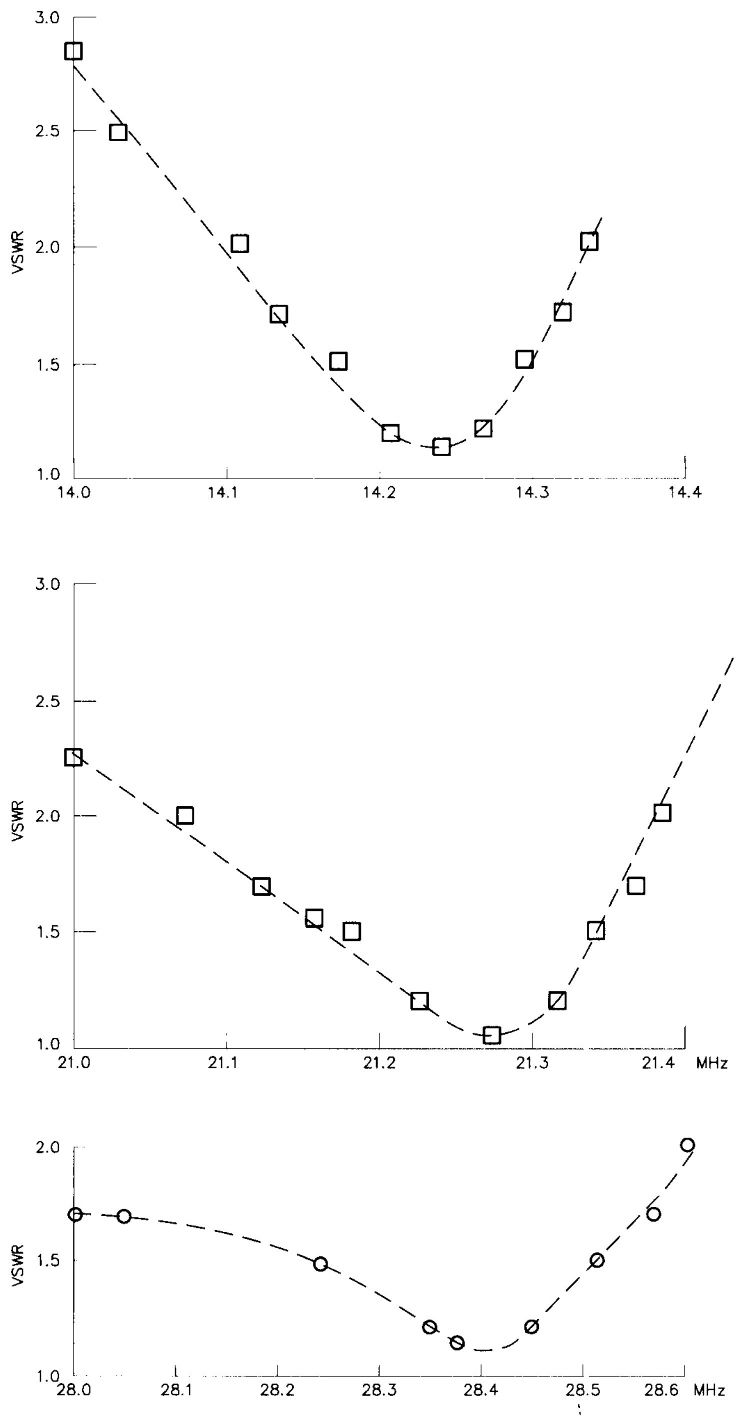

Fig 3—Measured VSWR.

About the Author

Richard Fenwick, K5RR, was first licensed in 1951. He holds a BSEE from Purdue University and MSEE and Engineering Degrees from Stanford University. In 1970, Richard co-founded Electrospace Systems, Inc, and is responsible for the Omega-T HV-3 and HV-5 antennas and the 2000C beam steering combiner. Now retired, he holds 12 US patents on antennas and has published numerous papers in QST *and professional journals.*

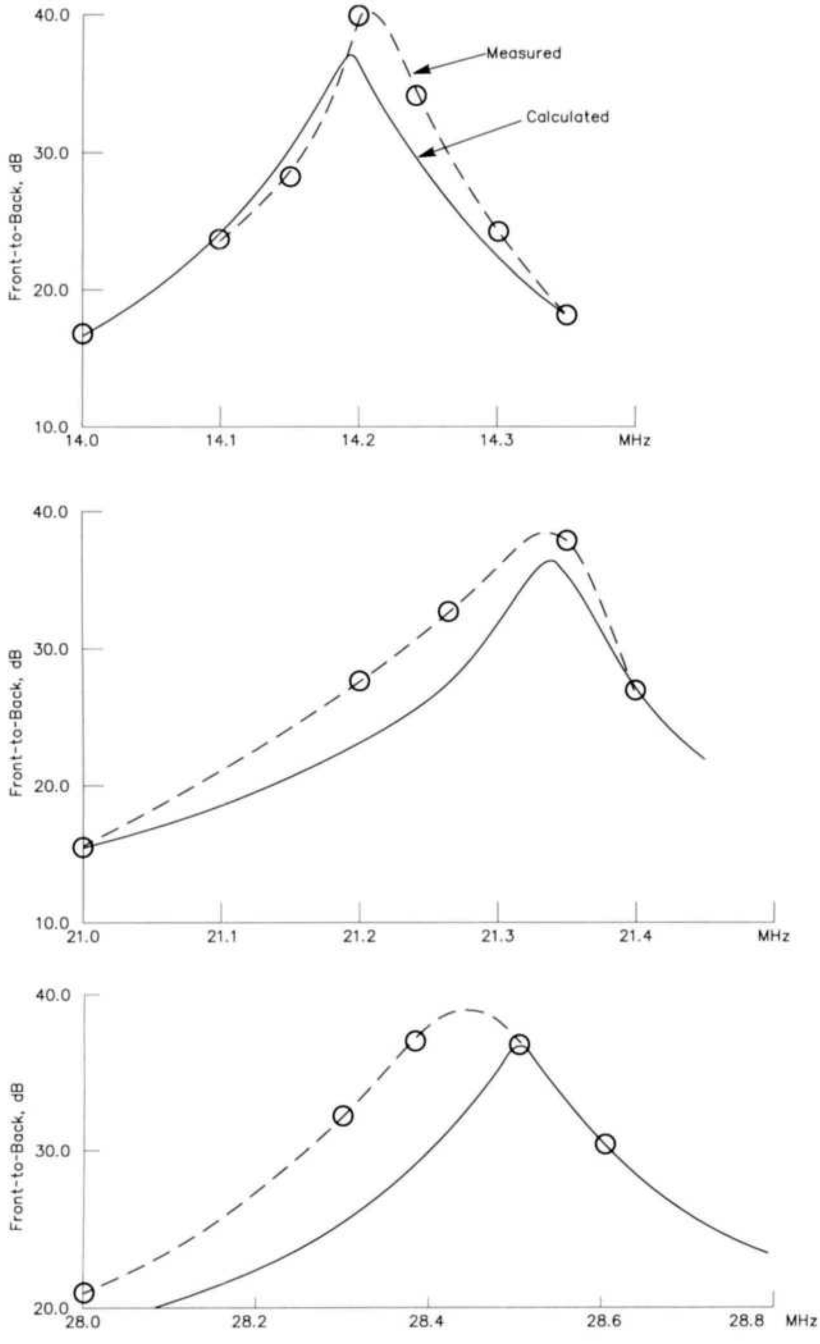

Fig 4—Measured and calculated front-to-back ratios.

A DSP-Based Audio Signal Processor

This project uses a low-cost DSP board and serves both as an introduction to DSP techniques and as a useful station accessory.

By Johan Forrer, KC7WW

This article presents the theory of operation and implementa tion details of a digital signal processor-based (DSP) audio signal processor (ASP). Such an ASP may be used with a communication receiver or incorporated as an integral part of a home-brew receiver project.

The ASP consists of several components: a digital beat-frequency oscillator (BFO), selectable band-pass filters for CW, SSB and other digital modes, a denoiser based on the least-mean-squares (LMS) technique and a Wiener-filter autonotcher (removing carriers or heterodynes). Several advanced concepts are applied in this project, such as multirate processing, adaptive filtering and frequency shifting. These all are of fundamental importantance to anyone wishing to learn more about the finer points of DSP. These principles may be considered the "tools of the trade" for working with DSP. This not only applies to audio signal processing, but also is becoming evident in contemporary digital radios.

The DSP platform used for this project is a low-cost evaluation module by Analog Devices called the *EZ-KIT Lite*.[1] However, any DSP platform with modest memory and processor speed may be used. The *EZ-KIT Lite* was considered ideal for this project because of its 16-bit audio interface, 33 million instruction per second (MIPS) ADSP-2181 DSP, 32k words of on-chip memory, included software development tools and low cost.

This article describes a number of components that make up an ASP, their functionality and how they are engineered and implemented. The objective is to expose the reader to the background that is essential for future involvement in DSP as there is no substitute or reward greater than trying it yourself.

Background

The W9GR DSP project that appeared in *QST* nearly four years ago encouraged many to build kits or to try their hand at DSP development.[2] Since then, several offerings of low-cost DSP evaluation modules (EVM) have brought powerful, yet affordable, DSP to the amateur experimenter. The *EZ-KIT Lite* used in this project is an example of contemporary EVMs that offer substantial amounts of on-chip memory combined with high clock rates—these modules are capable of doing serious DSP work.

A major hurdle for newcomers to DSP is the steep learning curve associated with DSP theory. In addition, implementation details for a typical

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[1]Notes appear on page 12.

DSP platform often seem a formidable prospect. The situation has improved a lot over the four years since the W9GR article was first written. Those interested in DSP now have access to reasonably good DSP filter design tools, simulation packages, and a wealth of literature and software examples, much of which is available on the Internet.

ASP Architecture

Module Descriptions

Fig 1 shows a top-down overview of the different modules of the ASP and their interconnectivity. It shows that there are several ways to interconnect modules depending on the type of processing required.

Input always passes through the digital BFO and may subsequently be routed "straight through" to the output combiner when no signal processing is desired. Otherwise, a specific filter may be placed in line, either with or without further processing. If filtering only is desired, the filter output is routed directly to the output combiner; the denoiser or autonotcher modules—or both—may be selected and placed in the signal path for further signal processing.

Digital BFO

Fine frequency resolution is a desirable feature for any receiving system. For example, changing received CW pitch to fall within a narrow-filter passband or adjusting an SSB signal for better clarity may be desired. It may also be an advantage to provide an alternative means to provide finer frequency resolution when working with a receiver that does not provide small frequency stepping. Such fine frequency resolution is essential for tuning RTTY, WEFAX or SSTV signals.

The digital BFO uses a Hilbert transformer to implement frequency shifting of signals received in the audio passband. The technique is an old one and is also used for other purposes such as the phasing method of SSB modulation/demodulation. Before the advent of DSP, such frequency shifting hardware was very difficult to build and tune. DSP achieves this nearly impossible feat with the Hilbert transform, which has the ability to do a perfect shift of a band of frequencies such that a constant group delay is exhibited thoughout the whole band.

Refer to references 10 to 12 for an in-depth treatment of the Hilbert transform, which is an essential part of the BFO. The operation of the BFO may further be formalized as follows.

Let the audio signal be represented by the time-domain signal $x(t)$. This is a real signal, ie, it has identical positive and negative frequency components. For the frequency-domain representation, let $F_p(X)$ represent the positive frequencies and $F_n(X)$ represent the negative frequencies. Thus:

$$F(x) = F_n(x) + F_p(x) \qquad \text{Eq 1}$$

The signal is subsequently band-limited by the codec's antialias filter and sampled at 18.9 ksps by the codec. The digitized signal is then passed through the Hilbert transformer as the first step in producing a special signal, called an analytic signal. The Hilbert transformer has the unique property that it delays all positive frequencies by +90° and all negative frequencies by −90°. Eq 2 shows this phase shifting (note the use $-j$ and $+j$ to indicate phase shifting) in the frequency domain.

$$\hat{F}(x) = -jF_p(x) + jF_n(x) \qquad \text{Eq 2}$$

To form the time-domain representation, $z(t)$, a complex signal, we combine $x(t)$, the real part, with the output of the Hilbert transformer, $\hat{x}(t) = H(x)$, the complex part, as shown in Eq 3. This is known as an analytic signal.

$$z(t) = x(t) + j\hat{x}(t) \qquad \text{Eq 3}$$

Eq 3 has interesting properties—it contains only positive frequencies. This is shown in Eqs 4, 5 and 6.

$$Z(t) = F(t) + j\hat{F}(t) \qquad \text{Eq 4}$$

$$Z(x) = F_p(x) + F_n(x) + j\left[-jF_p(x) + jF_n(x)\right] \qquad \text{Eq 5}$$

$$Z(x) = 2F_p(x) \qquad \text{Eq 6}$$

This property of Eq 6, that it only contains positive frequencies, is of great

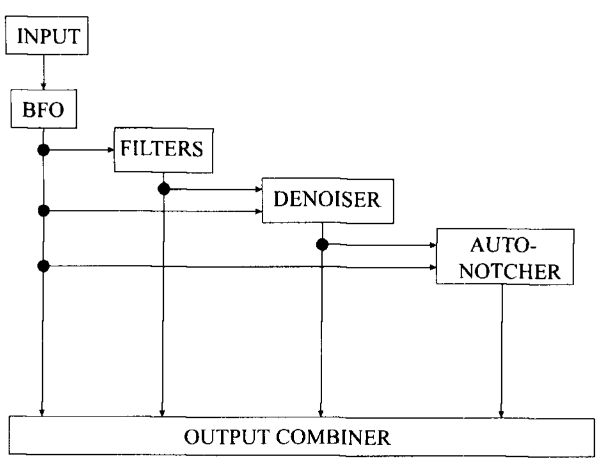

Fig 1—Outline of the audio signal processor (ASP) architecture. Each module has one output node and at least one input node. Module interconnections are user selectable and configured by software.

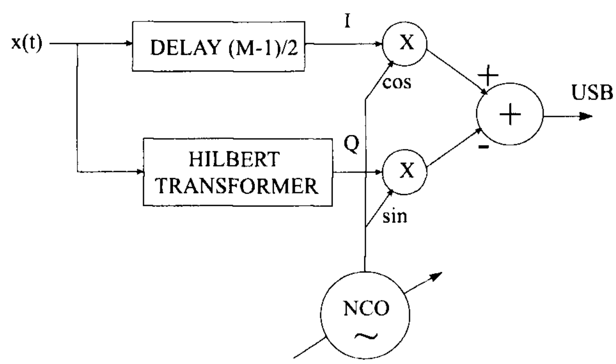

Fig 2—Hilbert transformer-based BFO. An analytic input signal is formed and mixed with a numerically controlled quadrature output oscillator (NCO). The real-valued output of the difference between the upper (in-phase or "I" branch) and lower (out-of-phase or quadrature "Q" branch) produce the upper sideband (USB). This corresponds to shifting the band of input frequencies up in frequency.

significance when working with phasing-type modulation schemes in communication systems. For example, selective cancellation of certain frequencies is possible without the need for additional filtering. Use of this property is made in the following section.

The next step is to do the actual frequency-shifting operation. What we are after is a real signal, $y(t)$, that has both positive and negative frequencies, with the positive frequencies shifted up in frequency—and the negative frequencies shifted down in frequency—by an amount determined by the BFO. This frequency shift is obtained by mixing (multiplying) a signal generated using a numerically controlled oscillator (NCO), which is our BFO, with the analytic input signal, $z(t)$, as shown in Eq 7.

$$y(t) = R_e\left[z(t)e^{j\omega t}\right] \qquad \text{Eq 7}$$

The NCO frequency is represented in complex notation as $e^{j\omega t}$. Note that we need to take the real part of the result, as after the frequency shifting is done we want to process only real numbers. Substituting and expanding Eq 7 results in Eq 8.

$$y(t) = R_e\left[(I + jQ)(\cos(\omega t) + j\sin(\omega t))\right] \qquad \text{Eq 8}$$

where I and Q refer to the in-phase and quadrature component signals of the analytic input signal. Eq 8 then reduces to our wanted signal as shown in Eq 9.

$$y(t) = I\cos(\omega t) - Q\sin(\omega t) \qquad \text{Eq 9}$$

A diagram of the processing is shown in Fig 2. Note that the negative sign produces the upper sideband (USB), or frequency up-shift. If a down-shift in frequency is desired, the sign should be a positive, in which case the lower sideband (LSB) results. The frequency response of the Hilbert transformer design used in the project is shown in Fig 6. See Notes 11 and 12 for details on determining the Hilbert transform coefficients.

Filters

The ASP includes the passband-shaping filters listed in Table 1. Digital filters can be tailored to meet specific criteria, such as the nature of their transition zones (filter skirts), stop-band rejection (adjacent-channel suppression) or pass- or stop-band ripple. Several interacting factors are involved: the sampling rate, steepness of the transition zones and the desired bandwidth. The designer can experiment with filter order as a means to achieve the desired effects, however, more often than not, some compromises must be made either in the final bandwith, transition zones and/or amount of stop-band rejection.

Filter requirements also vary. CW filters, for example, need steep skirts and good adjacent-channel rejection, typically in the order of –60 dB or better. Too narrow a filter is hard to use in practice because, if the signal is just slightly off frequency, the operator may have a hard time locating and placing it in the filter's passband. Two CW filters are provided, one wide and the other very narrow. Filters for SSB need a wider bandwidth, with somewhat relaxed adjacent-channel rejection requirements than for CW or RTTY. Stop-band rejection on the order of –40 to –50 dB is usually adequate. RTTY filters, like CW, also require good adjacent-channel rejection, the optimal bandwidth being determined by the baud rate and FSK shift. It's always a good idea to allow a little extra bandwidth for operator convenience—there usually is little degradation is performance by allowing a small amount of frequency tolerance.

Having some idea of filter requirements, practical implications need to be considered next. Finite impulse response filters (FIR) are used in this application. These were designed using the Parks-McClellan algorithm (Remez exchange, equiripple) using *Matlab*.[8] An excellent public-domain program is available as an alternative (Note 9), however, the reader may use one of any number of filter-design software packages. The magnitude responses for the six filters are shown in Figs 3 to 5 and comply with the specifications for bandwidth and stop-band rejection described in Table 1.

The codec is programmed to run at 18.9 ksps. At this rate, filter order becomes an important concern as it determines whether there will be sufficient clock ticks between samples to execute all of the DSP code. For example, a CW filter with stop-band rejection of –60 dB would require an extraordinarily large filter order to match the filter shown in Fig 3 and may be starving for clock cycles. However, the use of multirate processing reduces this computational load substantially. In the case of the CW filter, decimation by eight reduces the sample rate to 2.3625 ksps and only requires a filter order of 60 to meet the –60 dB stop-band rejection requirement.

Multirate processing requires additional overhead due to a low-pass filter at the decimating front end and a similar low-pass filter at the interpolating output stage. Note that CW operation normally does not require the noise processing functions because of the narrow bandwidth. Autonotching should also be bypassed for CW as the autonotcher would attempt to null out every CW signal in the passband!

For SSB and RTTY, a decimation

Table 1—Passband-shaping filters used in the audio signal processor. These are all finite-impulse response (FIR) filters. The associated frequency responses, passband bandwidth (BWp), stop-band bandwidth (BWs), and filter order are shown. The sampling frequency is 18.9 ksps. CW filters are centered at 800 Hz, SSB filters assume a voice-grade channel, and RTTY is for high-tone pair standard (2125/2295 Hz). The various stop-band bandwidth values were obtained by trial-and-error for the given sample rate and filter order.

Filter	*Freq.Response (Hz)*	*Bandwidth (Hz)*	*Sample Rate (kSPS)*	*Filter Order*
CW 100 Hz	750-850	BWp 100	2.3625	60
		BWs 400	2.3625	
CW 500 Hz	550-1050	BWp 500	2.3625	60
		BWs 700	2.3625	
SSB narrow	300-2700	BWp 1200	9.4500	120
		BWs 1400	9.4500	
SSB narrow	300-1500	BWp 2400	9.4500	120
		BWs 2600	9.4500	
RTTY 50 baud	2075-2345	BWp 270	9.4500	120
		BWs 465	9.4500	
RTTY 200 baud	2030-2390	BWp 360	9.4500	120
		BWs 560	9.4500	

factor of two is applied. These filters may optionally be used with noise reduction or autonotching. However, digital data signals may suffer waveform distortion effects and possibly other timing-related problems, so it usually is not a good idea to use such processing with data signals.

Denoiser

Denoising using DSP may be achieved by several means: least mean squares (LMS), autocorrelation and spectral subtraction.[3,5] The effectiveness of each of these depends on the nature of the noise and, to some extent, the nature of the signal. There also appear to be some psychological effects—operators often prefer one technique over another as a matter of taste. In certain instances it also appears that using a combination of these techniques may have advantages. This subject is beyond the scope of this article, and the reader is encouraged to study the references listed in Notes 2, 3, 4, 7 and 11 for the LMS technique and Notes 3, 5 and 6 regarding the spectral subtraction technique.

This project implements the LMS method as described by Hershberger.[2,4] The architecture for the LMS algorithm is shown in Fig 7. A delayed version of the input signal is passed through a tunable filter and then compared to the unprocessed input signal. The difference signal is then used to tune the variable filter in order to drive the difference signal to zero. The premise is that speech signals exhibit a substantial amount of coherence. That is, the delayed signal will have high correlation with the raw input. Noise, on the other hand, tends to have a random nature and will not show the same degree of correlation.

The variable filter thus becomes a time-varying filter. An FIR structure

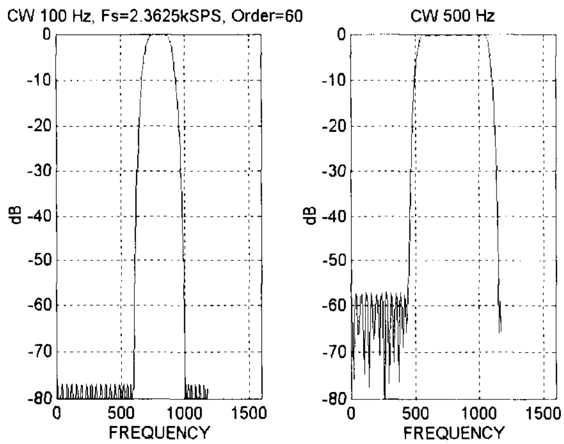

Fig 3—CW filter centered at 800 Hz. The left-hand figure has a 100-Hz bandwidth and the right-hand figure has a 500-Hz bandwidth. The sample rate is 2.3625 ksps and the filter order is 60.

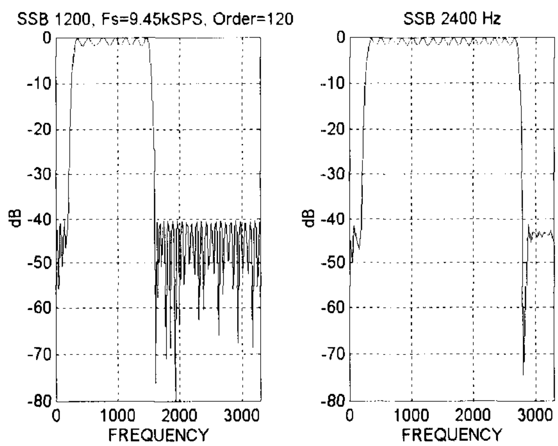

Fig 4—SSB filters. SSB filters assume a voice grade channel (300-3000 Hz). The left-hand figure is 1200 Hz wide and the right-hand figure is 2400 Hz wide. The sample rate is 9.45 ksps and the filter order is 120.

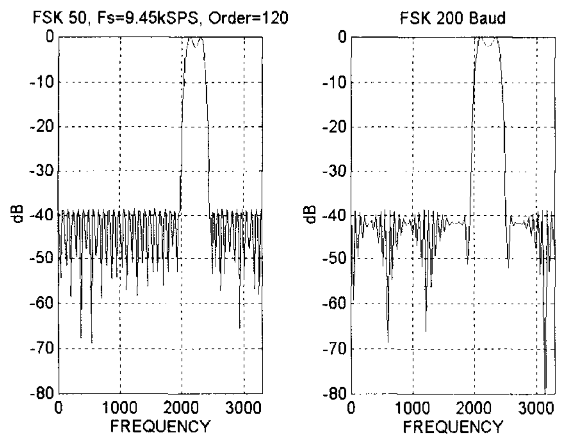

Fig 5—FSK filters. The FSK filters are the standard high-tone pair centered at 2210 Hz. The filter in the left-hand figure is designed for 50-baud operation and the filter in the right-hand figure is for 200-baud operation. The sample rate is 2.3625 ksps and the filter order is 60.

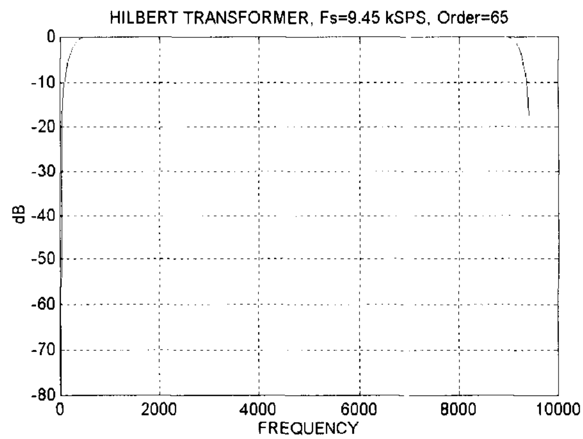

Fig 6—Magnitude response of the Hilbert transform of order 65. When used with a causal filter, it will have approximately unity gain, a group delay of (65-1)/2 samples, and approximately 90° phase shift.

is used as given in Eq 10.

$$y_k = \sum_{n=0}^{L} b_n(k)x_{k-n} \qquad \text{Eq 10}$$

The algorithm effectively tunes the set of coefficients, $b_n(k)$, in order to drive the difference signal, or error signal, ε, as small as possible. The LMS algorithm uses the method of steepest descent. In this case the set of coefficients may be considered a vector, B_k, at instant k, that needs to be updated for a minimum mean squared error (MMSE). The usual procedure for minimizing a function, in this case the error squared, ε^2, is followed. An estimate of the amount that the coefficient vector B_k needs to change is determined from the gradient, given in Eq 11.

$$\nabla_k = \frac{\partial E\left[\varepsilon^2\right]}{\partial B_k} \qquad \text{Eq 11}$$

Here, E refers to the "expected value," or mean.

Since we are dealing with a system where there potentially may be many local minima on the error surface, and subsequently little chance for achieving an absolute MMSE, the algorithm only makes small adjustments at a time to the coefficient vector, B_k, in order to steer it towards MMSE. This correction is shown in Eq 12.

$$B_{k+1} = B_k - \mu\nabla_k \qquad \text{Eq 12}$$

Here, the factor, μ, determines the rate of change. The gradient given in Eq 11 is difficult to compute for a dynamic system. However, it may be estimated from the instantaneous error as shown in Eq 13.

$$\hat{\nabla}_k = \frac{\partial \varepsilon_k^2}{\partial B_k} = 2\varepsilon_k \frac{\partial(x_k - y_k)}{\partial B_k} \qquad \text{Eq 13}$$

Where x_k is the reference (or input signal in our case), and y_k represents the output of the time-varying filter. Since x_k, the input signal, is independent of the output of the time-varying filter, it may be considered a constant and we may drop it from the partial derivative. If we then substitute Eq 10 for y_k, Eq 13 simplifies to:

$$\hat{\nabla}_k = -2\varepsilon_k X_k \qquad \text{Eq 14}$$

where X_k, is the L-element input vector (input delay line buffer). The estimated coefficient vector, B_{k+1}, in Eq 12 then becomes:

$$B_{k+1} = B_k + 2\mu\varepsilon_k X_k \qquad \text{Eq 15}$$

Eq 15 is the classic LMS formula. It is evident that the dynamics of the tuned filter depend on several factors; the rate at which the coefficients can be adapted, the sampling rate and the feedback loop factor, 2μ.

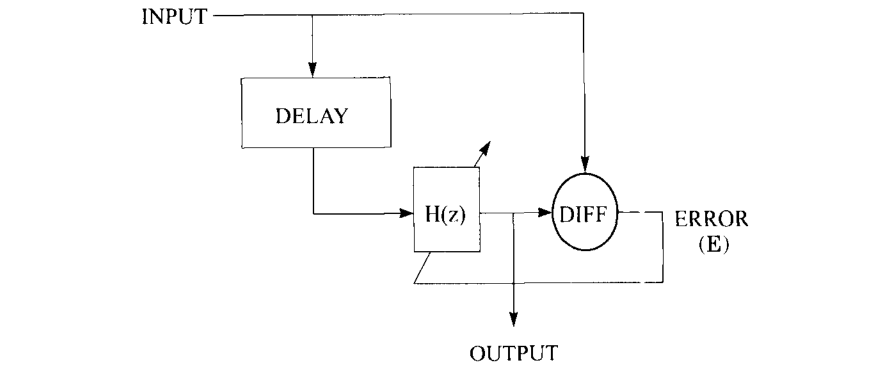

Fig 7—Generalized adaptive filter architecture for implementing the least-mean-square (LMS) denoiser.

Reyer and Hershberger further suggest the use of a "decay" factor to allow the filter to return to a quiet setting for situations where there is no input signal.[4] The final form for adapting the variable filter's coefficients is shown in Eq 16.

$$B_{k+1} = (1-d)B_k + 2\mu\varepsilon_k X_k \qquad \text{Eq 16}$$

Autonotcher

The autonotcher is based on a Wiener filter, which is basically identical to the LMS denoiser except that coherent signals are subtracted from the input.[2,4] This is achieved by simply using the same algorithm except for where the output is taken. This would correspond to the point in Fig 7 labeled ERROR. This makes code implementation really simple.

The working parameters such as the amount of delay in the signal path and the decay and convergence factors, and d and μ, respectively, are different for denoising than for auto-notching functions. These factors are maintained and adjusted separately.

A useful feature in this implementation allows for both the LMS denoiser and Wiener autonotcher to be placed in series. This is useful when monitoring SSB transmissions as it helps in removing heterodynes and reduces listening stress by removing background noise. Mileage, however, varies and much remains subject to personal preferences.

Software

Software for the ASP is available for downloading.[13] The software package consists of several modules—a control program that resides on the PC and DSP code for downloading to the *EZ-KIT*. Source code for both the PC control program, written in C, and the DSP, written in assembly language, is provided and may be used as a basis for further experimentation. I encourage you to explore and modify the code—remember that there is no greater reward than trying and succeeding in doing it yourself.

Acknowledgment

I wish to acknowledge the contributions that Adrian Nash, G4ZHZ, has made in providing ideas and suggestions for this project. The project started out on a DSP sound card and initially only provided the denoising function. Adrian's version provided the ideas for the BFO and filters. The version presented in this article combined all these functions and was ported to the *EZ-KIT*.

Notes

[1] *EZ-KIT Lite* is manufactured by Analog Devices, One Technology Way, PO Box 9106, Norwood, MA 02062-9196.

[2] Hershberger, D. "Low-Cost Digital Signal Processing for the Radio Amateur," *QST*, September 1992, pp 43-51.

[3] Hershberger, D. "DSP—An Intuitive Approach," *QST*, February 1996, pp 39-42.

[4] Reyer, S. E., and Hershberger, D., "Using the LMS Algorithms for QRM and QRN Reduction," *QEX*, September 1992, pp 3-8.

[5] Hall, D., "Spectral Subtraction Eliminates Noise from Speech in Real Time," *Personal Engineering*, May 1995, pp 51-54.

[6] Boll, S. F., "Suppression of Acoustic Noise in Speech Using Spectral Subtraction," *IEEE Transactions on Acoustics, Speech, and Signal Processing*, April 1979, Vol ASSP-27(2), pp 113-120.

[7] Widrow, B., Glover, J. R., McCool, J. M., Kaunitz, J., Williams, C. S., Hearn, R. B., Zeidler, J. R., Dong, E. and Goodlin, R. C., "Adaptive Noise Canceling: Priciples and Applications," *Proceedings of the IEEE*, December 1975, Vol 63 (12), pp 1692-1716.

[8]*Matlab*, The MathWorks, Inc, Cochituate Place, 24 Prime Park Way, Natick, MA 01760. **(http://www.mathworks.com)**

[9]Egil Kvaleberg, Hysebybakken 14A, N-0379 Oslo, Norway. (**http://www.oslonet.no/home/egilk**)

[10]Frerking, M. E., *Digital Signal Processing in Communication Systems*. 1993, Van Nostrand Reinhold. New York. (ISBN: 0442016166). (See various chapters on the Hilbert transform).

[11]Stearns, S. D. and David, R. A., *Signal Processing Algorithms in* Matlab. 1996, Prentice-Hall, Inc, NJ (ISBN: 0130451541), pp 339-343.

[12]Oppenheim, A. V. and Shafer, R. W., *Discrete-Time Signal Processing*, Chapter 10. 1989, Prentice-Hall, Inc, NJ (ISBN: 013216292).

[13]The software, in file qexasp.zip, is available from **http://www.arrl.org/qexfiles/**.

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By Zack Lau, KH6CP/1

3456-MHz Transverter

Jim Davey's 3456 no-tune transverter was an amazing piece of work—it was the first of the no-tunes and did a remarkable job of establishing the concept. Like many ground breaking designs, it had a number of minor problems. Here is my approach to fixing them.

The most serious problem is board radiation that unbalances the mixers and degrades spectral purity. This is caused by the combination of a low dielectric constant and a relatively thick circuit board. Unfortunately, high dielectric constant boards have about twice as much loss, a significant drawback since the transverter requires high-Q bandpass filters. As a result, I chose a much thinner, 15-mil board. This allows an aluminum cover to be placed over the circuit with negligible effects on circuit performance. You don't need absorptive rubber to shield this circuit. Microwave absorber material can be tough to find in small quantities.

Another drawback of the original design is the lack of voltage regulators. Performance was seriously degraded as the batteries ran down. By using three-terminal regulators, the circuits work just fine between 9 and 15 V. The LM2940T-8.0 is shown in Fig 1; it not only features a low drop-out voltage, but offers reverse polarity protection in case you hook up your batteries backwards! Beware, you can turn the regulator into an oscillator by substituting an inadequate value for C9. You need a sufficiently large amount of high-quality capacitance for stability. For details, consider downloading the data sheet from National's WWW page: **http://www.national.com/pf/LM/LM2940.html**.

Finally, the MMICs used were a bit marginal, operating near their upper frequency limit. This version uses newer MMICs with significantly enhanced performance at 3.5 GHz. The new MMICs have so much gain that it made sense to revise the circuit topology. Instead of dual mixers, I chose to use a single mixer and a splitter. It is common practice when using a single mixer to use the same band-pass filter for transmit and receive It is placed between the splitter and the mixer. Terminating the mixer with the splitter improves performance, while add-

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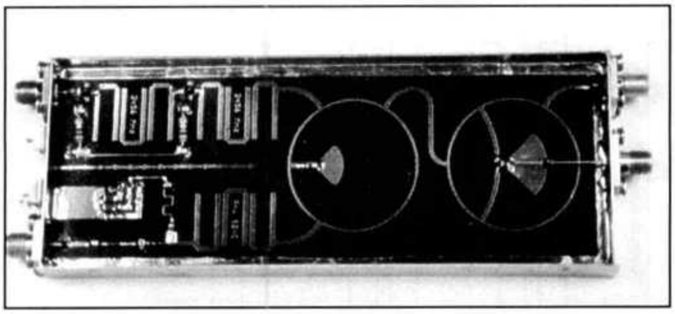
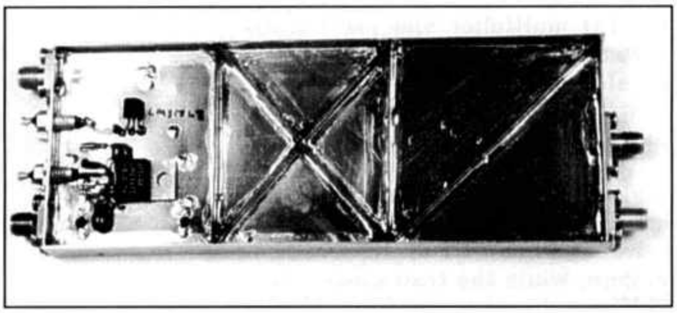
ing little extra complexity to the circuit. The situation would be different with pipe-cap or waveguide filters—then I'd have to spend nearly twice as much time fabricating the filters. In this situation, board space is saved because I don't need to place a small rectangle between two large circles.

There is also a subtle advantage for those of us using heat-transfer techniques for fabricating the boards. All the high-tolerance filters are concentrated in a relatively small area of the board. This makes it significantly easier to get an accurate reproduction. I noticed this when examining an LO board with low output (+4 dBm instead of +8 dBm). The output filter was stretched so much that the gaps between the resonators were a couple of mils wider than the design called for.



3456-MHz no-tune transverter, 14 dBm output, 1.7 dB NF.



3456-MHz no-tune transverter voltage regulators.

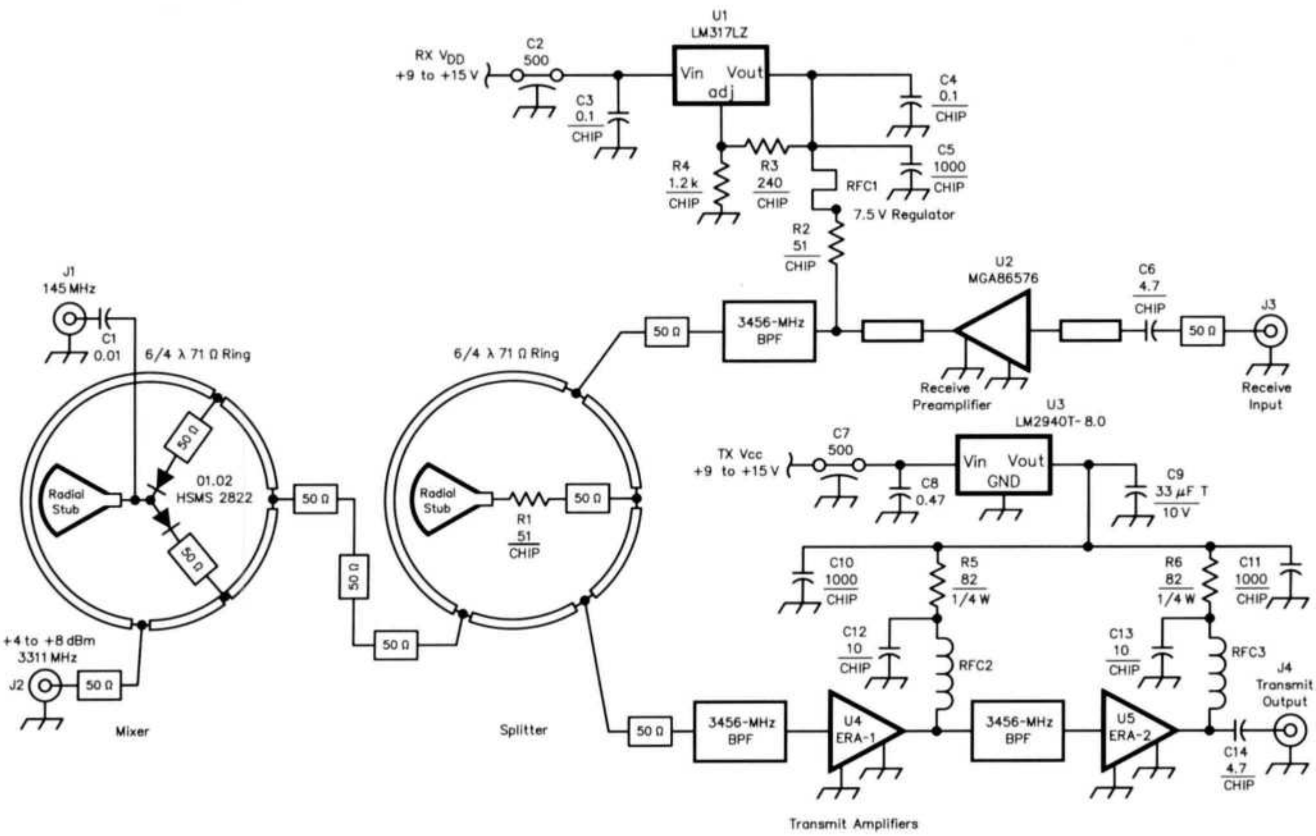


Fig 1—Schematic diagram of the 3456 transverter board.

C2, C7—Feedthrough capacitor; value not critical.
C6, C14—High-quality 4.7-pF chip capacitor like the ATC 100A. Not critical if you aren't worried about noise figure or power output.
C9—33-μF, 10-V tantalum. *The National Data Book* recommends a minimum of 22 μF.
D1, D2—HSMS 2822 packaged diode pair.
J1—Two-hole flange-mount SMA panel jack. Omni Spectra 2052-1652-02 works quite well.
J2, J3—Four-hole flange-mount SMA panel jack.
RFC1—Printed circuit board RF choke.
RFC2, RFC3—4 turns of no. 28 enameled wire close wound. 0.062-inch inside diameter.
U1—TO-92 case adjustable regulator.
U2—Hewlett Packard MGA 86576 GaAs MMIC.
U3—National LM2940T-8.0 low-drop-out regulator.
U4—Mini-Circuits ERA-1 HBT MMIC.
U5—Mini-Circuits ERA-2 HBT MMIC.

This effectively raised the center frequency of the filters. As Table 1 indicates, you can still get useable performance with 3 dBm of LO drive. To assist you in fabricating the filters, the dimensions are shown in Fig 2.

The single mixer significantly reduces the circuit board area required. Only 16 square inches are needed for the LO multiplier and main transverter board, about 2/3 of the original. It also simplifies transmit/receive switching—the mixer can be hooked up directly to a +3 dBm VHF transceiver, like the Rick Campbell mini R2/T2/LM2.[1] The effect of the splitter loss on performance is negligible—the receiver still has a 1.7-dB NF and 14 dB of gain, while the transmitter has a 1-dB compression point of +14 dBm with 12 dB of gain. The decrease in gain on receive may actually be an advantage by reducing it's susceptibility to mixer overload if a low-noise preamplifier is added. The output level is convenient for running surplus TWTAs or their solid-state replacements.

The new Mini-Circuits ERA-1 and ERA-2 have just the right amount of gain for the transmit amplifiers. On transmit, too much gain can be just as bad as too little gain. The more gain you have, the easier it is to make an oscillator. Feedback might help, but installing feedback networks deviates from the idea of a simple, reliable project with a minimum of parts. These MMICs use heterojunction-bipolar-transistor (HBT) technology. They do a good job of combining wide

[1]Notes appear on page 20.

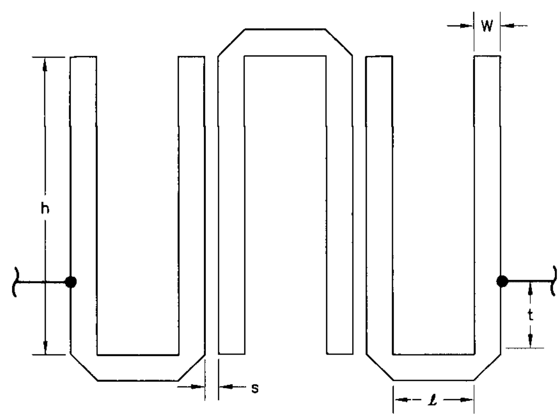


Fig 2—Dimensions of the 3312 and 3456-MHz band-pass filters on 15-mil 5880 Rogers Duroid.

	3456 MHz	3312 MHz
h coupled line height (mils)	486	508
s spacing between coupled lines (mils)	25*	25*
t tap height (mils)	28	34
w line width (mils)	50	50
ℓ uncoupled length (mils)	150	150

*Modeled spacing, see text.

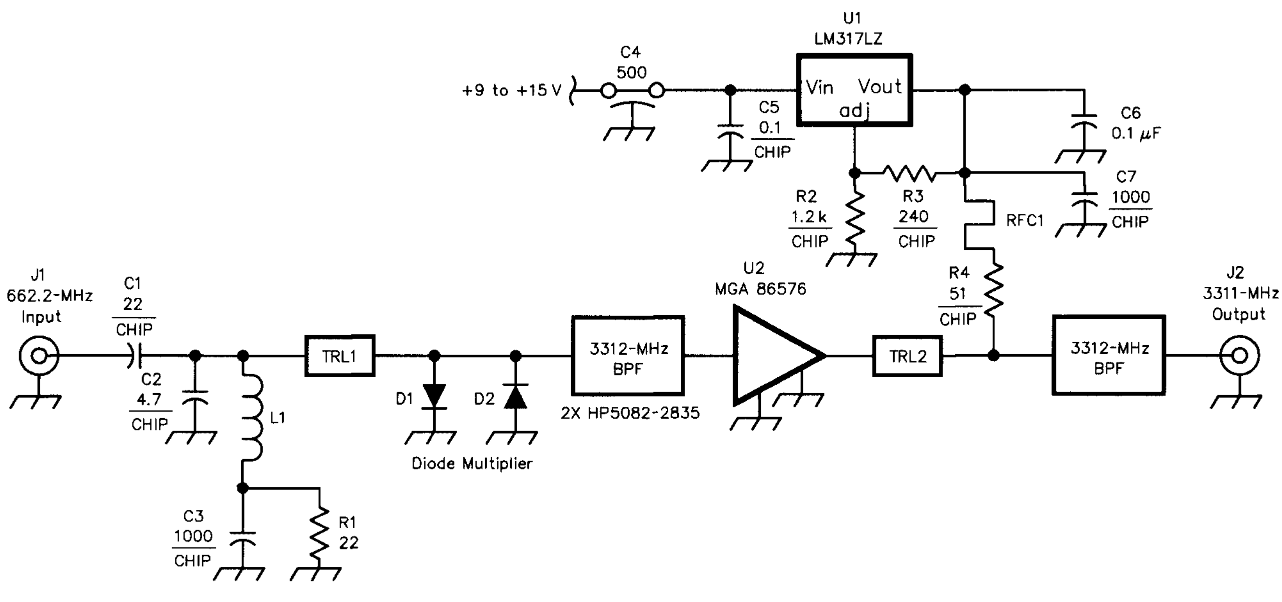


Fig 3—Schematic diagram of the 3311-MHz local oscillator multiplier.

C4—Feedthrough capacitor, value not critical.
D1, D2—Hewlett Packard 5082-2835 Schottky diodes.
L1—3 turns no. 28 enameled wire spaced 2 wire diameters. 0.089-inch inside diameter.
RFC1—printed circuit board inductor.
U1—LM317LZ adjustable regulator.
U2—Hewlett Packard MGA 86576 GaAs MMIC.

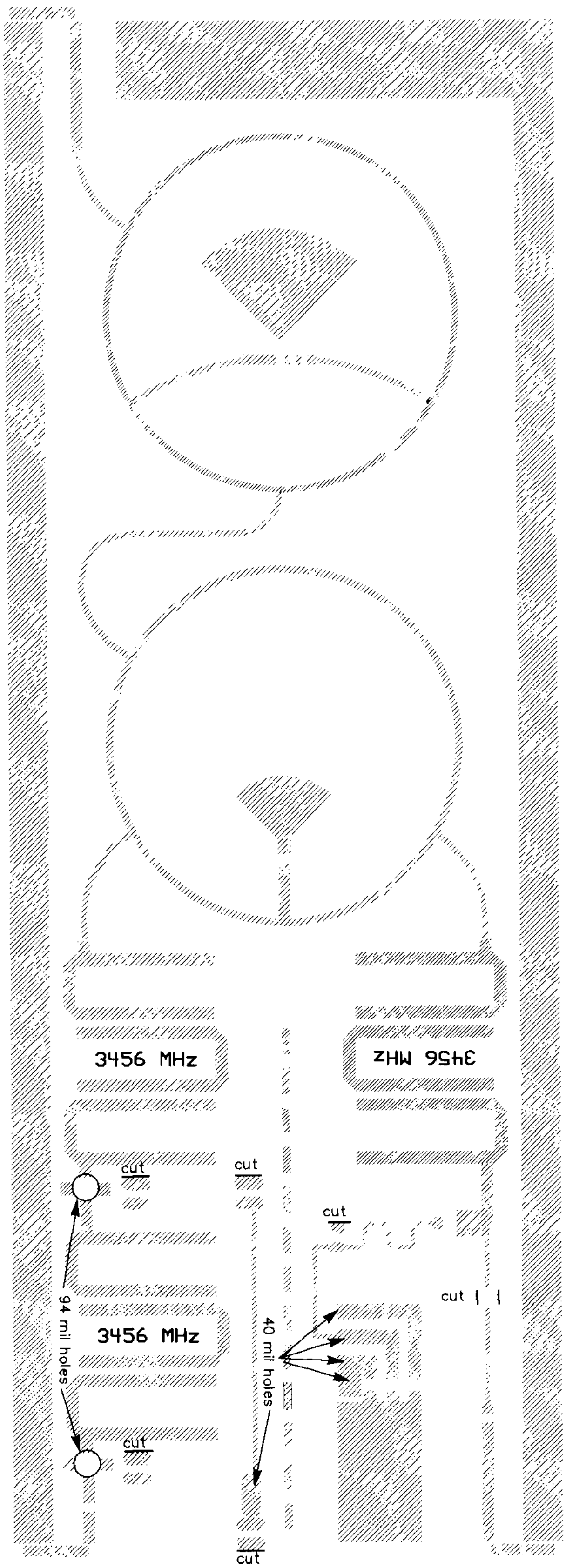


Fig 4—Cutting and drilling diagram for the transverter board.

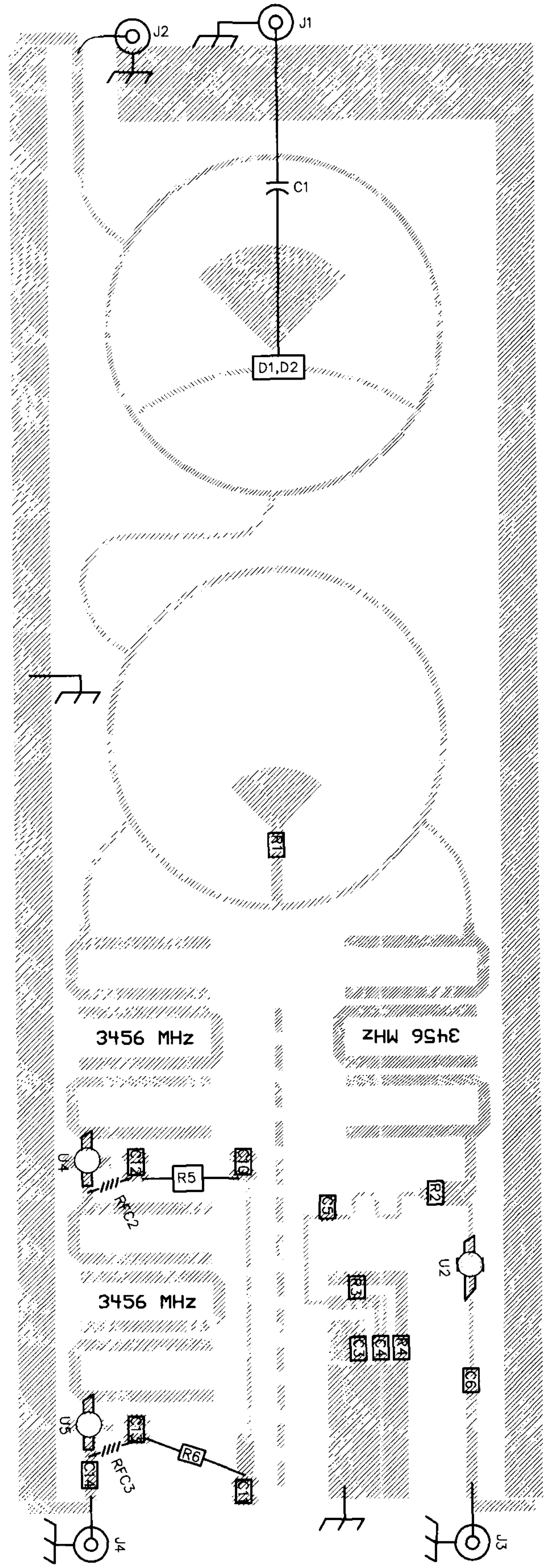


Fig 5—Trace side parts placement diagram for the transverter board.

bandwidth with a relatively low supply voltage. At 50 mA each, they draw a fair amount of current to generate 20 to 40 mW of RF. I think this a reasonable tradeoff, considering the complexity of the alternatives. A discrete FET design would be more complex, requiring a lot more design work.

I haven't experimented with the new ERA-4 or ERA-5 MMICs to see if more output power can be obtained without modifying the board. I'm still waiting for the ones I ordered in late May '96. I don't recommend using the high-gain ERA-3 MMIC between the bandpass filters—it is quite likely to be unstable unless drastic measures are taken. It may be necessary to shield the printed circuit board filters from each other. You might also experiment with narrowing the "waveguide" enclosing the circuitry. You may significantly reduce the chance of unwanted waveguide propagation by installing a shield on the optional grounding strip in the center of the board. This shield would lower the cut-off frequency by a factor of two. The low-loss nature of wave-guide is a significant disadvantage when attempting to build amplifiers—what better way to create an oscillator than to couple the input and output together with a low-loss transmission line?

Keep the parts close to the board to reduce their ability to launch signals into the waveguide. I laid out the board placing the RF chokes close to the edge of the wave-guide, as opposed to the center. Objects in the center of the waveguide couple into the waveguide better than those close to the edges. This is why you typically put detector diodes in the center of the waveguide when you want to maximize the signal to the diodes.

Keeping to the idea of simplicity, the receive preamplifier is a single Hewlett Packard MGA 86576 GaAs MMIC. It has about 24 dB of gain and a 1.6-dB noise figure. The NF is degraded an additional 0.1 dB by the 10 dB of converter losses. The LM317L could be replaced by a 78L07 or 78L06 regulator, but these are harder to find and require a larger input bypass capacitor for stability. With the three MMICs I used, the highest supply voltage resulted in the best gain and noise-figure performance. I didn't do any testing past the recommended 7-V device voltage. Unit 2 still had a 1.9-dB NF and 13.65 dB of conversion gain with a device voltage of 4.82 V, so the device voltage isn't terribly critical.

The mixer and splitter both use a 180° hybrid. A good reference on these

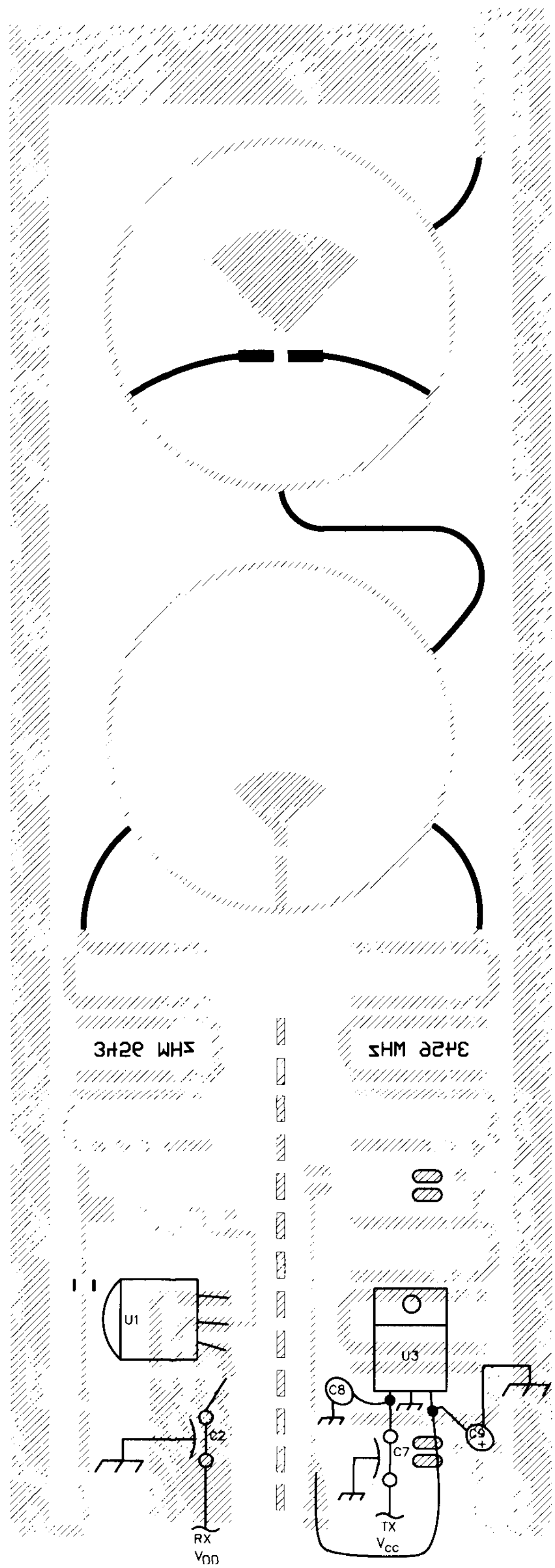


Fig 6—Ground-plane side parts placement diagram for the transverter board.

may be found in Chapter 6 of the *ARRL UHF/Microwave Experimenter's Manual*. The narrow bandwidth of the hybrid isn't a problem in this mixer application, due to the relatively low IF of 145 MHz. This is only 4% of the center frequency. The radial stub for the splitter is a little small, in an attempt to compensate for the stray inductance of the chip resistor.

The LO multiplier uses a diode multiplier, a pair of band-pass filters, and an MGA 86576 GaAs MMIC. FETs can be more efficient than diodes at frequency multiplication, but they tend to be more critical with regard to drive and tuning. The GaAs MMIC is a bit more expensive than a pair of the new ERA-3 MMICs, however, they produce just the right power level for driving a single mixer. An ERA-3 is more appropriate for driving a pair of diode mixers. Another advantage of the MGA 86576 is that it draws only 16 mA, compared to 35 mA for a single ERA-3. Actually, a pair of ERA-3s has too much gain, and running an ERA-1 and ERA-2 in cascade ups the current draw to 100 mA. This is six times as much as the GaAs MMIC draws.

The LO multiplier is designed to be used with a 662.2-MHz source. It can be easily modified to work with a more standard 552-MHz source by removing either one of the multiplier diodes, D1 or D2. As a 6× multiplier, +15 to +20 dBm of drive is needed. As a 5× multiplier, +13 to +20 dBm of drive can be used. This multiplier works much better with the correct number of diodes. If the stability of the LO amplifiers is marginal, it may be wise to add a 100-Ω resistor or resistive pad to the input of the multiplier. It isn't too difficult to envision cases where the constantly changing impedance of the diodes could cause problems.

I know that a lot of people are looking for practical ways to design the microstrip hairpin filters. The simple answer is there isn't any, at least for amateurs with little time or money. The programs that accurately simulate the discontinuities, such as the bends in the microstrip, still cost quite a bit of money. Trial and error, particularly with return loss measurements, is an effective way of designing complex filters, if you have the time. If you have a spectrum analyzer, I've found that upconverting a low-frequency signal generator with a mixer makes a good signal generator—if you add an isolator. The isolator significantly reduces the interaction between the filter and the mixer. If you aren't careful, some mixer/filter combinations can actually indicate that the filters have gain. Reflected signals can actually enhance the desired signal. As Jim Davey found out, return loss is a much more sensitive indicator of circuit performance than insertion loss.

Construction

Making these circuit boards is a stiff challenge—but some of us like challenges. Jim specified a tolerance of ±0.001 inches for the original boards. I doubt that I come that close with my etching techniques, but it's a target to shoot for. I was able to build three transverter boards, all with acceptable performance (Table 2). Mixing and LO spurs were at least 40 dB down. The second and third harmonics were only down 39 and 32 dB on one unit. This is no surprise, since there is no low-pass filtering of the output stage.

When comparing the dimensions of Fig 2 against the artwork, the careful examiner may notice that the spacing

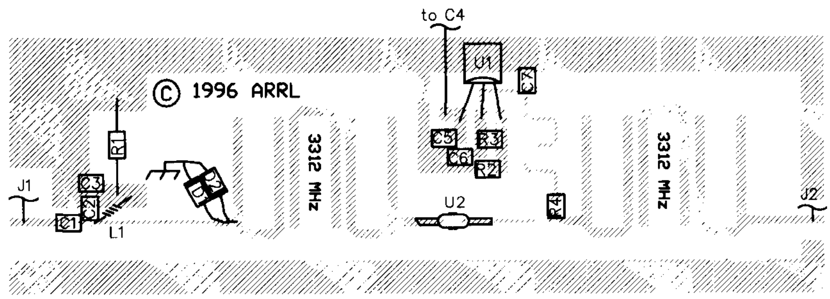


Fig 7—Parts placement diagram for the multiplier board.

Table 1—Effect of local oscillator power on transverter performance

Transmitter		*Receiver*	
LO Power (dBm)	*Output Power (dBm)*	*NF (dB)*	*Gain (dB)*
3.2	13.27	1.79	15.11
3.8	13.33	1.78	15.15
4.9	13.40	1.77	15.28
5.6	13.42	1.79	15.30
6.0	13.37	1.79	15.34
6.8	13.37	1.76	15.34
7.2	13.40	1.78	15.33
7.7	13.38	1.77	15.35
7.9	13.38	1.77	15.36

Table 2 —Test data for three converters

	Transmit		*Receive*	
Unit	*Power (dBm)*	*Gain (dB)*	*NF (dB)*	*Gain(dB)*
	(1 dB compression)		*(compressed)*	
1	14.40	12.9	1.78	15.34
2	14.10	12.6	1.75	14.95
3	14.83	12.1	1.64	12.94

Power was measured with an HP 8563E spectrum analyzer and confirmed with an HP 8481A/435B power meter. An HP 8970/346A was used to measure receive converter performance. Both an HP 8640B and a Marconi 2041 were used to generate the IF drive.

between coupled lines is actually 24 mils, as opposed to the 25 mils determined with the assistance of computer modeling. Since this was noted after several units were built and tested, I didn't bother revising the artwork. Due to the tight tolerances, I'd recommend you work with the Postscript files. If you can't download them off the 'Net from **http://www.arrl.org/qexfiles**, I can supply the file for noncommercial purposes if you enclose a 3.5-inch disk and addressed return envelope with postage.

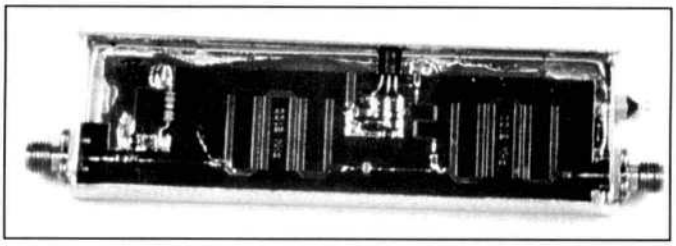
The boards are etched on Rogers RT/Duroid 5880 with a dielectric constant of 2.20, clad with 1 ounce rolled copper on two sides. The dielectric thickness is 0.015 inches.

When trimming the circuit board, don't forget to leave enough room for the SMA connectors. Normal square-flange connectors are 0.5 inches wide—thus the board needs to be at least 0.25 inches from the center of the 50-Ω microstrip traces to the edge. On the other hand, excessively widening the board increases the possibility of waveguide propagation, so you don't want to err too far in the other direction either. You can also use smaller SMA connectors—Digi-Key now advertises a line by Johnson Components (formerly EF Johnson Components). While a bit expensive, they supposedly work up to 26.5 GHz, as opposed to 18 GHz for standard connectors. They should be useful for 24-GHz work.

I used a hobby knife with a new no. 11 blade to cut the slits in the board for the grounding straps and MGA MMIC ground leads. The blade of the knife should just touch the outer edges of the pads marking where to cut the slits. I don't trim away any Teflon from the slits. Instead, I use a flat-bladed screwdriver to carefully close up the holes after the leads are passed through by reworking the remaining copper foil. The holes for the ERA MMICs are punched with a 94-mil hole punch and then touched up with the hobby knife. I've found that drilling large clean holes in thin Teflon board can be difficult. After drilling the 40-mil holes for the power supply parts, I countersink the ground plane side by hand with a 1/4-inch drill bit.

Some drill bits bite too deeply into the board, so you might practice your techniques on a few scraps of Teflon board. If you work slowly and realize there is a problem, you can "save" the board by carving away the excess copper with a sharp knife.

Even with a frame made of 25 × 500-mil brass sheet stock around the edges, I felt the board could use a bit of stiffening. To stiffen the board, I soldered some thin strips of unetched double sided circuit board to the ground-plane side. The assembly was stiffest when the boards were slid against the metal frame and soldered it. Forming an X or two seems to markedly stiffen the board. Brass could be used, but the circuit board is easier to solder and has less flexibility. A stiff board is important because most chip parts can't flex along with the board—too much flexing and something will break.



552 to 3312-MHz LO multiplier on 15-mil 5880 Duroid.

A two-hole flange mount works well for input connector J1. The thinner connector can be raised slightly compared to the other three connectors. This makes it easy to connect the input capacitor, C1. Alternately, higher walls could be used to enclose the transverter board.

While there are pads for grounding the ERA MMICs with copper foil, I found it easier just to bend and solder the leads. The leads are inserted into the holes and then bent against the body of the MMIC, flattening them against the ground plane. I used copper foil with Unit 2—there wasn't a significant difference in performance compared to the other two units.

I first test the voltage regulators to make sure they are putting out the proper voltages before installing the resistors that supply power to the MMICs. The actual voltage at the MMICs can vary a bit—the GaAs MMIC can draw between 9 and 22 mA, so the voltage drop across the 51-Ω chip resistor can be anywhere from 0.46 to 1.1 V. I'm not surprised at this range—manufacturing repeatable bias points has always been a weak point of FET technology compared to bipolar. The device voltage for the ERA MMICs is supposed to be between 3.2 and 4.4 V, nominally 3.8 V.

The shared mixer ought to make troubleshooting easier. If nothing works, there is a problem in the LO/mixer, or in both the transmit and receive amplifiers. While the conversion gain isn't excessive, it ought to be enough to hear the increase in noise when you turn on the receive amplifier.

Notes

[1]Campbell, Rick "A VHF SSB-CW Transceiver with VXO," *Proceedings of the 29th Conference of the Central States VHF Society*, ARRL, 1995, pp 94-106. Boards and kits are available from Kanga Products, Bill Kelsey, N8ET, 3521 Spring Lake Drive, Findlay, OH 45840, 419-432-4604. e-mail: kanga@bright.net or **http://qrp.cc.nd.edu/kanga/**.

Conference Proceedings Available

1996 ARRL and TAPR Digital Communications Conference

The 15th ARRL and TAPR Digital Communications Conference was held September 20-22, 1996, at the Quality Inn Seattle Airport in Seattle, Washington. Here is a summary of the papers presented. The Conference Proceedings are available from ARRL. ISBN: 0-87259-568-4; cost is $12, plus shipping; order number: 5684.

Learning DSP by Porting Programs to the TAPR/AMSAT DSP-93 Modem, John B. Bandy, WØUT

Linking BPQ Switches via Ethernet, Bill Barnes, N3JIX

javAPRS: Implementation of the APRS Protocols in Java, Steve Dimse, KO4HD

The Radio Amateur Digital System Artificial Intellegence Project, Garry W. Joerger, N5USG

Fast Flow Control in High-Speed Communication Networks, C.M. Kwan, R. Xu and L. Haynes

Nonlinear Channel Equalization Using Fuzzy CMAC Neural Network, C.M. Kwan, R. Xu, L. Haynes and J.D. Pryor

Optimization of Phase-Locked Loops With Guaranteed Stability, C.M. Kwan, H. Xu, C. Lin and L. Haynes

Baseband Group Delay Equalization of IF Filters for Data Communications, T.C. McDermott, N5EG

Easy to Follow Packet, James C. Nobis

Object-Oriented Modeling of a Satellite Tracking Software, M. Normandeau and M. Barbeau, VE2BPM

XNET: A Graphical Look at Packet Radio Networks, Richard Parry, W9IF

A 9600-Baud Modem for the LPT Port, Wolf-Henning Rech, N1EOW/DF9IC and Donald Rotolo, N2IRZ

Amateur Radio Digital Voice Communications, Paul L. Rinaldo, W4RI

WinAPRS: *Win*dows *A*utomatic *P*osition *R*eporting *S*ystem, Mark Sproul, KB2ICI, and Keith Sproul, WU2Z

Automatic Radio Direction Finding Using MacAPRS & WinAPRS, Keith Sproul, WU2Z

Circus of the Stars, Michelle Toon, KC5CGH

13-cm PSK Transceiver for 1.2 Mbit/s Packet Radio, Matjaz Vidmar, S53MV

23-cm PSK Packet-Radio RTX for 1.2 Mbit/s User Access, Matjaz Vidmar, S53MV

Packet and Internet, James Wagner, PhD, KA7EHK

Strategies for Improving Wide-Area

Networks, James Wagner, PhD, KA7EHK

The Word Storage Relay (WSR), Pat West, PE, W7EA

On-Air Measurements of HF Data Throughput Results and Reflections, Ken Wickwire, KB1JY

On-Air Measurements of MIL-STD-188-141A ALE Data Text Message Throughput Over Short Links, Ken Wickwire, KB1JY

CLOVER—The Technology Grows and Matures, Bill Henry, K9GWT

Construcing a Worldwide HF Data Network, Craig McCartney, WA8DRZ

Microwave Update 1996

Microwave Update 1996 was held October 4-6, 1996, at the Ramada Camelback Hotel in Phoenix, Arizona. Here is a summary of the papers presented. The Conference Proceedings are available from ARRL. ISBN: 0-87259-573-0; cost is $12, plus shipping; order number: 5730:

Improving System Integration for UHF and Microwave Communication, Kirk Bailey, N7CCB

A DSP Based Transceiver for UHF and Microwaves, Bob Larkin, W7PUA

NOISE: Measurement and Generation, Paul Wade, N1BWT

Optimizing TWT Power Output for Narrow Band CW/SSB Operation, James W. Vogler, WA7CJO

Ka-Band Beacon-Transponder for the AMSAT Phase-3D Satellite, Danny Orban, ON4AOD

High Performance 47-GHz Components, Toshihiko Takamizawa, JE1AAH

"No-Tune" Transverter for 10 GHz, Danny Orban, ON4AOD

The 1 dB Quest Revisited, John Swiniarski, WA1TFH

20 dB Gain Two-Stage 10 GHz Amplifier Using ERA-2 MMICs, Greg McIntire, AA5C

EME Communication at 1296 MHz and Higher Frequencies, Al Ward, WB5LUA

Simple Tracking Generators That You Can Build, John Petrich, W7HQJ

Just about as Cheap as you can get on 10 GHz, Kent Britain, WA5VJB

Feed Horns for Illumination of Parabolic Reflector Antennas for 1.3 and 2.3 GHz, Paul Chominski, SM0PYP/W6

The Cut Dish, John Anderson, WD4MUO/0

10 GHz Offset Feed Dish, Paul Wade, N1BWT

A Transportable Dish Mount, John Anderson, WD4MUO/0

UHF/Microwave Activity in Australia, Walter J. Howse, VK6KZ

The San Diego Microwave Group X-Band Repeater Update, Kerry Banke, N6IZW

2304 and 3456 No-Tune Transverter Updates, Steve Kostro, N2CEI

Phase Noise Part 1: What is it, and How it Affects Communications, Jeffrey Pawlan, WA6KBL

3456 Transverter Using TVRO Parts, Rick Beatty, NU7Z

Modification of TVRO LNBs for 10 GHz, Paul Wade, N1BWT, and Don Twombly, WB1FKF

Practical Microwave Antennas, Parts 1, 2 and 3, Paul Wade, N1BWT

1995 Microwave Update Noise Figure Results

EME Bibliography, Al Ward, WB5LUA

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Upcoming Technical Conferences

1996 AMSAT-NA Space Symposium and Annual Meeting

The 1996 AMSAT-NA Space Symposium and Annual Meeting will be held November 8-10, 1996, in Tucson, Arizona.

You know those times when you've been too busy or thought "ho-hum" about a ham event, so your buddy went, and you "missed it"!? This event is going to be good.

All the Southwest charm of Tucson—clear, broad, blue skies, at a season when our temperatures are the most comfortable. Friendly Southwest people and experiences.

Meet and share with hams from around the world.

Discuss and learn about the latest ham radio satellites under development.

Participate in Satellite Beginners' forums or immerse yourself in the intricate details of Amateur Space Technology.

Enjoy demonstrations of the latest satellite ground stations or take notes on how to put together the absolutely cheapest satellite ham station.

Bring your family; have them enjoy the tour to the KITT Peak Radio Telescope with you. Or see them in the evening after they have spent the day enjoying the tourist sites in the area.

Plan to *not* to miss it. For information call Heather Johnson, N7DZU, tel: 520-749-5106, email: n7dzu@azstarnet.com, or Larry Brown, NW7N, tel: 520-886-1957, email: nw7n@amsat.org.

Texas Packet Radio Society Fall Digital Symposium

The Texas Packet Radio Society Fall Digital Symposium will be held Saturday, December 7 from 8 AM to 5 PM, at the University of Texas campus, Sanchez Building of College of Education, Room 370. There is no cost to attend.

Topics covered will include (but not limited to): the status of the new hardware that Tucson Amateur Packet Radio is developing that will run the next decade's TexNet software and increase our network's capabilities in many ways; the recent threat to packet service providers; autonomy that arose at HamCom and isn't going away; what changes are needed in the Texas VHF and UHF bandplans to recognize and make room for new packet activities; is there any way to reduce the overcrowding that exists on the existing packet frequencies in the urban areas; what does it take to put up a new TexNet node; what is the status of the ROSE, NetROM and KA-node networks in our region; how to easily modify cheap old commercial gear for 9600 bps, and where to buy it; what is this Spread Spectrum stuff that I've heard about and how will it impact packet; APRS news; current TexNet firmware enhances the weather products available on the network and firmware now under development will tie in with APRS.

This symposium will have something for everyone. Hardly anyone goes home disappointed after participating.

Complete information and agenda may be found on the TPRS web site: **http://www.tprs.org** or by mail: Texas Packet Radio Society, PO Box 50238, Denton, TX 76206-0238.