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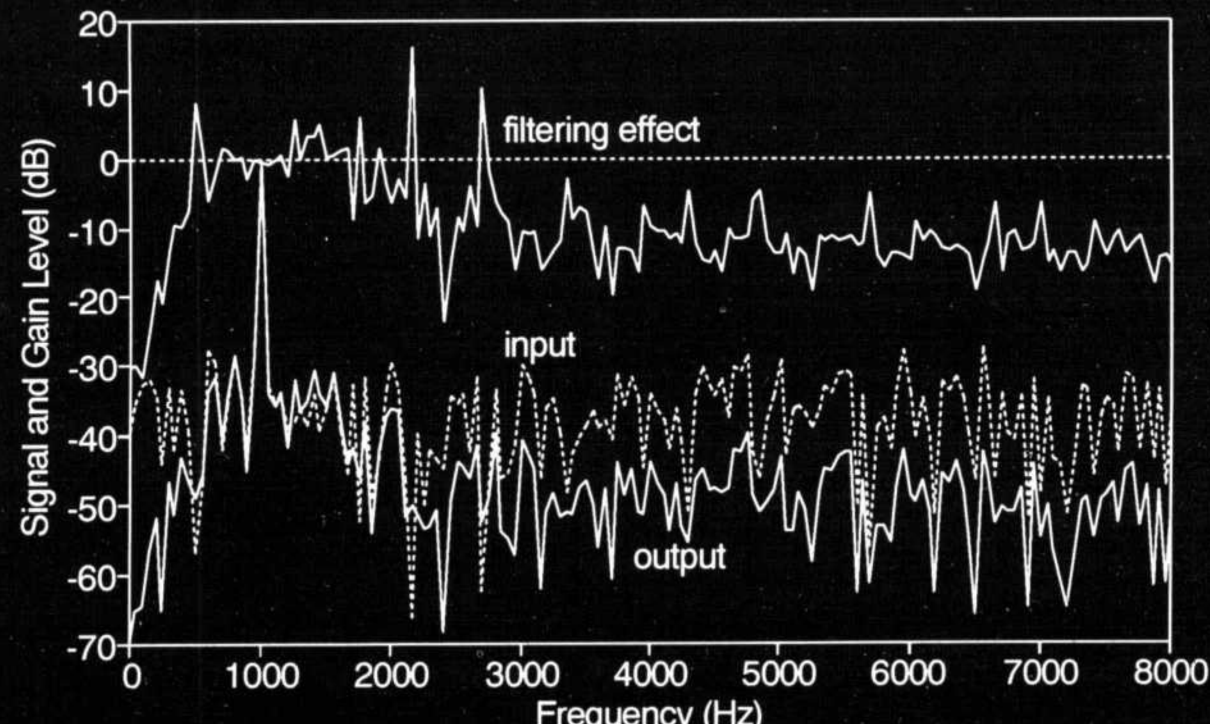
ARRL Experimenters' Exchange

SEPTEMBER 1992



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THE AMERICAN RADIO RELAY LEAGUE

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

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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...

Taking (Un)attendance

Much of packet radio as we know it today consists of a network of linked store-and-forward message systems—bulletin boards—scattered around the nation and the world. Where possible, these systems communicate via VHF packet links to forward messages. Where VHF links are not possible or aren't practical because of the delay of passing a message through multiple hops, often HF packet links are used instead. For the past five years, much of the US traffic forwarded on HF occurred under the auspices of a special temporary authority (STA) issued by the FCC to the ARRL. The administration of this STA by the League has allowed over 100 digital stations to operate without a control operator present for the purpose of forwarding.

The STA is soon coming to an end. The operations that are taking place under the STA are a clear benefit to the forwarding network since there is no other reliable real-time, long-distance data relay mechanism available to Amateur Radio. The League has proposed the establishment of rules for HF digital (RTTY and data) stations allowing unattended *response* to communications initiated by an attended station, but not unattended *initiation* of communications in the hope that this will allow the benefit of HF forwarding to continue while avoiding the potential problems of unrestricted unattended operation.

Not everyone is happy with this proposal. Some want no unattended digital operation; some want unlimited unattended operation; some want a mix. There has been a great hue and cry since the League's position was set forth at the July 1992 meeting of the Board of Directors. But when a survey in the January 1992 *QST* requested direction from the membership, it elicited only about 500 responses. Whether or not you responded to that survey, it is now time to stand up and be counted. Let your Director (see page 8 of *QST*) know what you think should be done about this issue. It's too important for you to leave it to others and hope for the best.

This Month in *QEX*

Digital signal processing is the "hot ticket" today in amateur radio—and other communications technologies. A fascinating look into DSP filtering techniques is presented in, "Using The LMS Algorithm for QRM And QRN Reduction," by Steve Reyer, WA9VNJ and Dave Hershberger, W9GR. This article is a companion piece to an article published this month in ARRL's flagship publication, *QST*. We hope to do more such companion pieces in the future, allowing authors the opportunity to expand on the technical niceties of their projects in *QEX* when *QST* page space limitations don't permit detailed coverage there.

Long-time amateur satellite operators may remember the 4×3×5 mode-J filter used with OSCAR 8. The advent of later mode-J birds, such as AMSAT's microsats, has reawakened interest in this design. This month, ARRL Technical Advisor and AMSAT stalwart Dick Jansson, WD4FAB, describes a new kit implementation of the filter in, "The 4×3×5 Filter—Revisited."

AM lives, and synchronous detection promises to help it live a fuller, richer life as ARRL Laboratory Engineer Mike Gruber, WA1SVF, reports in, "Synchronous Detection of AM Signals." Mike presents a complete theoretical discussion as well as practical circuits.—*KE3Z, email: jbloom@arrl.org (Internet)*

Using The LMS Algorithm For QRM and QRN Reduction

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In the September 1992 issue of QST, the article "Low Cost Digital Signal Processing for the Radio Amateur" describes an automatic notch filter and noise reduction filter using digital signal processing techniques.[1] This article presents the mathematical basis for the system.

Conventional notch filters built from op-amps, resistors, and capacitors are difficult to make variable. Depending on the circuit, compromises usually must be made with filter order, the number of ganged variable components, and notch bandwidth and depth as a function of tuning. Making an analog notch filter automatic has further trade-offs of acquisition time and Q versus sensitivity. Even if a switched capacitor filter is used, it is hard to make analog filters quickly and effectively track moving carriers. And an analog notch filter generally will only remove one interfering tone at a time.

In the area of noise reduction, although conventional analog noise blankers can successfully remove high-amplitude impulse noise, analog techniques have not been particularly effective at removing low level background QRN.

The solution? Digital signal processing (DSP)! These methods use a microcomputer to process the signals, allowing the desired flexibility at a reasonable cost. The proper choice of algorithm will create a fast, nearly complete cancellation of any periodic interference—multiple carriers, keyed CW, etc. It will also allow unusually effective reduction of background QRN.

The specific area of DSP applied here is called *adaptive filtering.* This involves computer-implemented filters that adapt themselves to incoming signals, enhancing the signal's desirable properties. Other applications of adaptive filters include active electronic mufflers for automobiles, ventilation system noise reduction, and cockpit sound reduction in helicopters. These powerful techniques are applied here in a ham radio environment—to cancel received audio interference.

Adaptive Filters

There are two basic methods for implementing adaptive filters. They depend on whether or not a reference signal for the interference is present. In the ventilation system example, a microphone could be placed "upstream" to obtain a sample, or reference, of the offending sound. This could be used to create an inverted cancellation signal "downstream," perhaps in a room where the noise is offensive. Production of the inverted signal then cancels the offensive sounds. The availability of such a reference is very helpful in designing the system.

A system with no reference is a much bigger challenge. A received SSB audio signal, corrupted by a carrier, has the voice and interference combined. While a human can describe which is which just by listening, making a computer algorithm determine the difference, especially when the interference is changing, requires a clever approach. The important factor is that the carrier is strongly correlated (periodic over a long term) and the voice is not. If the voice and carrier are delayed by a fraction of a second, the carrier will still "look" the same, but the voice won't. This factor allows an adaptive filter to discern the difference with no separate reference signal required.

The block diagram of Fig 1 shows how the system operates. The transmitted signal is subject to a periodic interference. This corrupted received signal (typically voice plus interfering carrier) then enters a subtraction block, where the signal through the lower path is subtracted from the corrupted signal. The lower path serves to isolate the interference, so it can be removed from the corrupted signal, leaving only the desired signal. The trick is to design the lower path so it properly passes only the interference.

If the interference is considered to be any periodic signal, the lower path will take the form of a bandpass filter, passing this signal. This filter is shown as an FIR (finite impulse response) digital filter. The FIR filter,

[1]Notes appear on page 8.

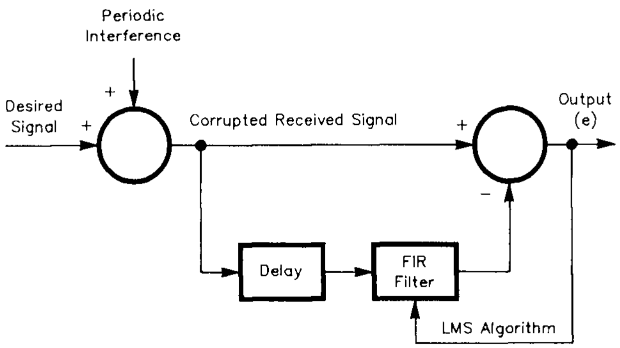


Fig 1—Block diagram of the interference reduction system. Note that no interference reference signal is used.

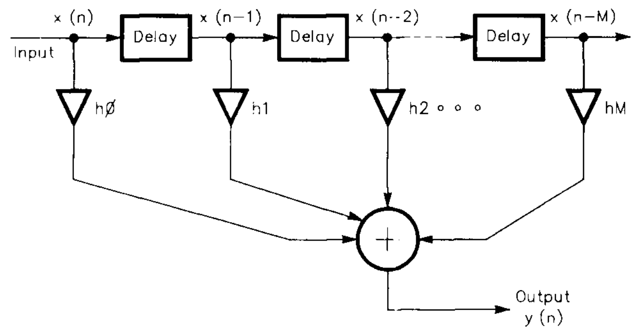


Fig 2—Signal-flow diagram of an FIR filter. x(n) represents the nth input sample and y(n) represents the nth output sample.

sometimes called a transversal filter, is a numerical technique operating on the samples of the incoming, corrupted, delayed signal. The FIR implementation is that of numerical convolution. Convolution, the mathematical process by which systems affect signals, can be implemented numerically in a DSP microcomputer. Thus the computer implementation becomes the filter itself. If the FIR filter can be made to change its characteristics, it can be part of an adaptive filter.

The lower path also includes the delay to decorrelate the voice signal. This delay comprises multiple samples, and its length is determined empirically.

The FIR Digital Filter

Fig 2 shows a finite impulse response digital filter. The samples enter the process from the left, and pass through the single-sample delays (or memory locations) as the sample clock ticks. At each tick of the sample clock, the delay memory values are shifted toward the right. Each of these samples are then weighted (multiplied) by the *h* values and the result is summed, providing the filter output. It is well known that such a process, with carefully chosen *h* values (called the unit pulse response), can implement lowpass, bandpass, highpass, and other common filter types. The sampling rate must be chosen to satisfy the sampling theorem.[2]

If the actual interference frequency were known, the unit pulse response values could be chosen and fixed. However, the interference can change, and so must the pulse response. Fortunately, these *h* values are simply numbers, not component values, and can be altered thousands of times per second. In fact, they can be changed after every sample, in order to adapt to the incoming interference.

The LMS Algorithm

The method for adapting the *h* values depends on how well the subtraction process is removing the interference. A common algorithm used for this purpose is called the least mean squares (LMS) algorithm.[3]

The LMS algorithm states that after each sample, *x*, the difference between the input and the developing cancellation signal should be found. It is called *e*, for "error." Now, each *h* value should be adjusted by adding an amount proportional to the product of *e* and the particular *x* which was used in association with that *h*. In other words,

$$h_{new} = h_{old} + 2uex$$

This operation is performed for each *h* value. The scale factor, 2*u*, controls the rate of convergence of the algorithm, and is adjusted empirically. The proof of the validity of the LMS algorithm is mathematically quite complex, but a certain intuitive feel can be gained from a simple example.

Suppose the only input is a pure sine wave. Ideally, the system should adapt and cancel it. Further assume the FIR coefficients (*h*) and the filter data values (*x*) are initially set to zero. The samples will arrive immediately at the subtracter with nothing to cancel them. For a while, they exit the system unattenuated. No change will occur in the *h* values, since there are no nonzero *x* values in the filter. Ultimately, samples will pass through the delay block in the lower path, and enter the FIR filter. Finally, there are nonzero *x* values which cause changes in *h*. The *x* values in the filter are samples of a sine wave. So the correction applied to each *h* looks like a scaled version of this sine wave and the set of *h* values starts to approximate this sine wave. The process continues until the error, *e*, becomes so small that no further changes occur. Fortunately, having the error become small is the same as cancelling the sine wave, since the error is the adaptive filter output.

This property where the filter coefficient set has the same appearance as the waveform to be cancelled is sometimes called a "matched filter." It is known that such a filter has bandpass properties. In some cases, especially in the presence of a solitary pure tone, as above, the coefficients will adapt until such time that the error is negligible, but the result may not have the appearance of a matched filter. Rather, it may approximate some other type of bandpass filter. With the addition of a speech signal, there is enough sample-to-sample variation that the filter is constantly attempting to adapt. Also, recall that the presence of the delay block decorrelates the speech, so the filter will not be successful at cancelling this desired signal.

Choosing the convergence factor, $2u$, is typically done empirically. If it is too small, the *h* values are changed very slowly. Worse, computation errors in dealing with such small numbers may prevent them from being changed at all. If $2u$ is very large, convergence can occur rapidly. Unfortunately, the system will respond radically to any slight error output, and may even adapt to the speech itself. A midpoint value is the best compromise. The specific method for choosing $2u$ for this system is presented later in this article.

What we have now is an automatic notch filter. By making some minor changes, we can change this into a noise rejection ("denoiser") filter. The LMS algorithm as described so far is used to reject correlated parts of the signal (pure tones) and to pass *less correlated* signal components (speech). The LMS algorithm may be used to perform the opposite function, of accentuating the *somewhat* correlated parts of the signal (speech) and rejecting the uncorrelated components. Here the objective would be to pass somewhat repetitive voice components and to reject completely uncorrelated signals such as hiss and thermal noise. To accomplish this, several things are done. First, the output of the FIR filter is taken as the system output. Second, the convergence parameter $2u$ is chosen differently. Third, the length of the delay line is adjusted (shortened) to recognize speech signals as correlated.

Correlation is a relative term. Pure sinusoidal tones (carriers) have the highest degree of correlation. Voice signals have somewhat correlated components (slow drawls and sustained vowel sounds such as "e-e-e-e-e"), and less correlated components such as sibilant and fricative "s," "ch," and "k" sounds. Examples of uncorrelated, non-repetitive sounds include white noise, thermal noise, static crashes, etc. By controlling the various LMS algorithm parameters, the system can be made to discriminate between highly correlated tones and somewhat correlated speech (autonotcher), or between somewhat correlated speech and highly uncorrelated hiss (denoiser).

In an automatic notch application, making $2u$ too large will cause the LMS algorithm to attack and remove the most correlated components of voice signals, adversely affecting intelligibility. Making $2u$ too small will result in slow carrier acquisition and a tendency to let all but the strongest carriers slip through.

Similarly, making the delay line too long will cause noticeable speech echoes and make it harder to track moving carriers. A delay line which is too short will blur the distinction between speech and unwanted tones.

In summary, traditional fixed analog filters discriminate between wanted and unwanted signals based on their *frequency*. The LMS algorithm, on the other hand, discriminates between wanted and unwanted signals as a function of their *degree of correlation*. The LMS algorithm may be operated to favor either the more correlated or more uncorrelated parts of the signal.

Although the LMS algorithm can provide some enhancement of voice signals by rejecting noise (uncorrelated) or tones (highly correlated), there are some things it cannot do. Rejection of off-frequency SSB interference or its intermodulation products is not an application of LMS, because these unwanted signals generally have the same degree of correlation as the desired voice signal.

Some hams have asked whether the LMS denoiser mode would be effective on FSK signals like RTTY and HF packet. The LMS denoiser works by forming bandpass filters around the most significant spectral lines in the signal. These bandpass filters in general will not be linear-phase filters, and consequently the data pulse shapes may become distorted. If a signal is HF packet, RTTY, or AMTOR, then we have *a priori* knowledge of the signal spectrum, and a *fixed* linear phase FIR filter could be designed and optimized for the specific signal type. This approach would be superior to the filtering performed by LMS.

LMS Algorithm Improvement via "Coefficient Decay"

We have developed a further modification to the LMS algorithm which we are calling "coefficient decay." This enhancement to the basic algorithm improves its performance in ham applications. Practical experimentation has revealed an unwanted side effect of the basic LMS algorithm. When the $2u$ convergence parameter is made aggressive (large), there is a "noise buildup" effect where hiss components are noticeably amplified. This phenomenon is due to the lack of a recovery mechanism in the "textbook" LMS algorithm. That is, when the input signal goes to zero, the *h* coefficients of the standard LMS algorithm will not change, and when there is a lot of noise in the signal, the coefficients will tend to wander aimlessly and may become quite large, increasing the unwanted noise part of the signal. The LMS algorithm may be enhanced by adding a subtle tendency of the *h* coefficients to return to zero by making the recursion formula:

$$h_{\text{new}} = (1-d)h_{\text{old}} + 2uex$$

where *d* (the decay parameter) is much smaller than one. A typical effective value for *d* is 0.00004, which makes 1-*d* about 0.99996. Using this value for *d*, in the absence of an input signal it takes about 17,000 iterations of the LMS coefficient adjustment recursion formula above for *h* to decay to one half of its original value. In other words, introduction of this coefficient decay mechanism gives the LMS algorithm a way to slowly recover and reset itself over a period of several seconds. A side effect of coefficient decay affecting the denoiser mode is a slight amount of volume expansion. Because expansion is itself another noise reduction method, this side effect is actually beneficial.

Simulation

Unlike analog filters, whose design is methodical and straightforward, design of an LMS filter involves a lot of subjective evaluation and experimentation. Listening tests using various kinds of real signals are essential.

Once parameter values for delay line length, convergence factor 2*u*, decay factor *d*, filter length *n*, and the sampling rate are all determined, it remains to verify that the LMS algorithm is operating as expected through simulation.

Results of these simulations are presented here. These simulations were performed by executing the actual TMS320C10 code developed for the hardware in the *QST* article mentioned earlier, using a developmental TMS320C10 simulator which runs on IBM PCs.

Fig 3 shows the LMS software in the autonotcher mode in response to a burst of signal at 1000 Hz followed immediately by a burst at 1500 Hz. As shown, the LMS algorithm adapts within a small number of cycles to the unwanted tone. As the tone changes frequency, the LMS algorithm must adapt again. The short intervals of sinewave cycles during the adaptation period sound like "clicks."

The system is still effective in the presence of a speech signal, simulated by random noise. Fig 4 shows an unwanted tone being eliminated while the speech-type signal is allowed to pass.

Fig 5 shows the frequency response of the notch produced by the LMS algorithm in response to the speech-plus-tone signal. Note that there are some additional notches produced of lesser depth than the desired notch. All notches are quite narrow, resulting in a minimal effect on the speech signal.

Fig 6 shows the impulse response of the FIR filter, which resembles a time domain "snapshot" of the undesired signal. This clearly illustrates the "matched filter" concept described earlier.

The next two figures show the operation of the LMS algorithm in the noise reduction mode. Fig 7 shows the three plots of the LMS algorithm operating on a tone plus wideband noise; the top trace shows the overall system frequency response. The lower two traces show the

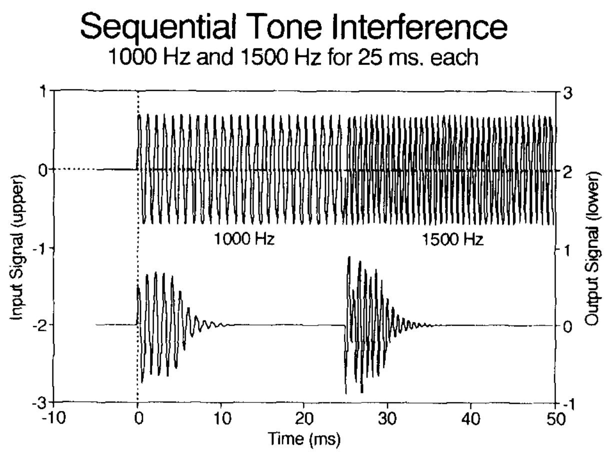


Fig 3

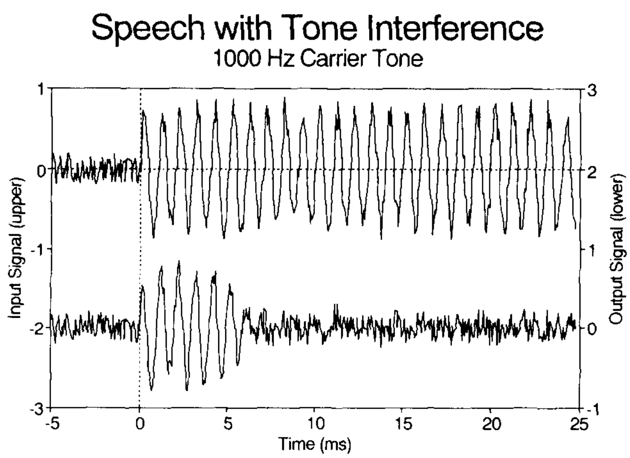


Fig 4

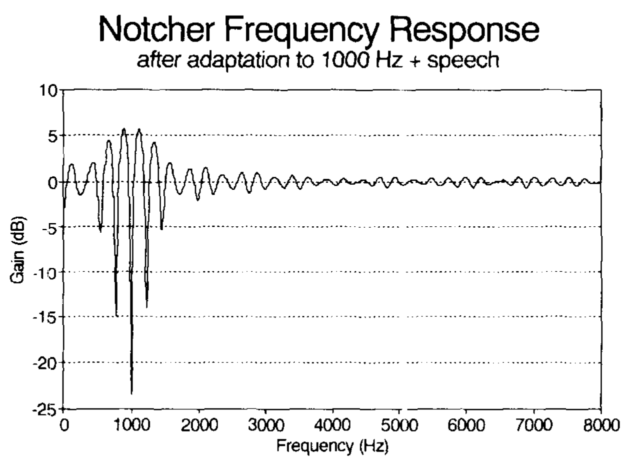


Fig 5

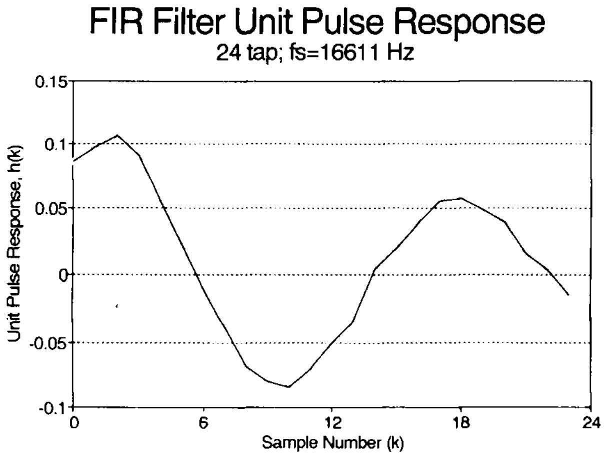


Fig 6

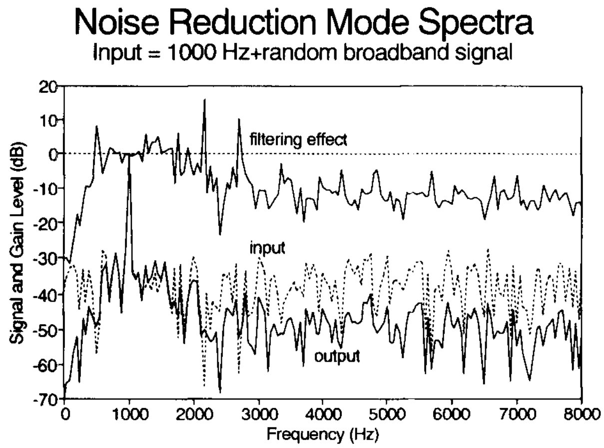


Fig 7

spectrum of the input signal superimposed on the spectrum of the output signal. The system frequency response is found by comparing the simulated input and output response levels. (Because of the 8-bit I/O precision, some of the spikes in the top trace may be disregarded as they represent the difference between two very small numbers with large rounding errors.) You can see that the 1000-Hz tone has been passed at full amplitude while the wideband noise away from the desired signal component has been reduced by as much as 10 to 15 dB. The received signal-to-noise ratio can therefore be improved by signal processing at the receiver instead of requiring the transmitting station to increase power.

Fig 8 shows the actual frequency response characteristic achieved by the system in the above simulation. It is determined by examining the FIR coefficients and computing the effective overall system frequency response after adaptation. The plot shows how a bandpass filter is automatically formed by filter adaptation around the 1000-Hz correlated signal component.

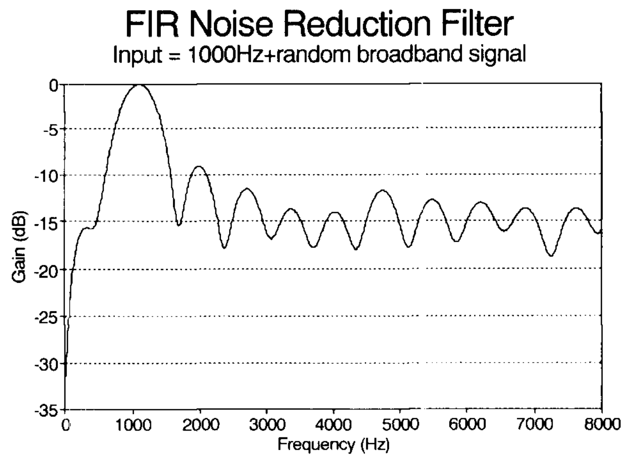


Fig 8

Practical Results

We have come to some observations through our use of LMS in our everyday amateur operations. The automatic notch mode operates very quickly to eliminate CW interference. Acquisition time is much faster than a commercial analog automatic notch filter we have experimented with. A swept analog automatic notch filter can only look at one narrow band of frequencies at a time. If a carrier comes on frequency, it will not be eliminated until the swept notch happens to encounter it (which may take a second or more). The LMS algorithm, on the other hand, is constantly evaluating the entire audio spectrum and there is no waiting for a sweep to encounter the interference. LMS autonotch acquisition time is measured in milliseconds.

Although the automatic notch mode of the LMS algorithm quickly gets rid of carriers, we find that HF listening is made much more pleasant by using the denoising mode most of the time. Only when the frequency is under siege by tuner-uppers and CW interference do we change over to the autonotcher mode.

Noisy FM signals are also cleaned up by the LMS denoising mode. We have found that a noisy FM repeater provides a most dramatic demonstration of the capabilities of LMS, perhaps due to the nature of FM where, unlike HF SSB, the background hiss-type noise is consistently present.

HF CW is also improved by LMS denoising, however at high speeds (30 WPM and up) the acquisition time tends to shorten the dits.

The denoising mode is very effective against background hiss. But when the QRN becomes very loud (such as strong impulse noise), then the LMS algorithm—as we have implemented it—loses its effectiveness. In this case, a conventional analog noise blanker would be more effective. The LMS denoiser algorithm will not extract a signal completely buried in the noise, but moderately noisy signals enjoy the greatest improvement.

We have found that the denoising mode reduces listening fatigue, and makes long-term HF monitoring much more pleasant.

We have also tried combining the denoising and autonotching functions with some success, but prefer to keep them separate. Using a low-cost DSP hardware platform with combined denoising and autonotching puts strain on DSP computational resources, driving down the sampling rate and filter lengths and reducing overall effectiveness. Even if more DSP horsepower were applied, combining the two functions results in performance compromises for the simple reason that the two modes are somewhat in conflict with each other: while the autonotcher mode is trying to remove correlated signals and pass uncorrelated signals, the denoiser mode is trying to do just the opposite!

Generally speaking, we leave the LMS denoising function in-line for most of our HF operations. When CW interference is present or expected, we switch over to the autonotcher mode.

Conclusion

The LMS algorithm has recently become economically practical for use in Amateur Radio applications. Modifications to the basic textbook algorithm improve its performance for amateur use. The autonotcher LMS mode is highly effective against CW interference, while the denoiser LMS mode reduces low-to-moderate-level hiss-type QRN. Price reductions on digital signal processors will allow further application of LMS and other algorithms to increase our enjoyment of Amateur Radio.

Dr. Steven E. Reyer, an advanced class amateur, was first licensed in 1967 at age 17. He is active on HF SSB and VHF packet, with his main Amateur Radio interest being in designing, experimenting and building. Steve holds a Ph.D. in Electrical Engineering (Digital Signal Processing) and is a Professor of Electrical Engineering at the Milwaukee School of Engineering. He also consults for industry in the areas of digital signal processing, communications and microprocessor systems.

Dave Hershberger was first licensed in 1965 at age 14 as WN9QCH, and holds an Extra Class license. Dave is active on HF CW and SSB and VHF FM and packet. Dave has a BA in mathematics from Goshen College and BS and MS degrees in electrical engineering from the University of Illinois. Dave is a senior staff engineer at the Grass Valley Group, a subsidiary of Tektronix which produces television broadcast equipment.

Notes

[1]D. Hershberger, "Low Cost Digital Signal Processing for the Radio Amateur," *QST*, Sep 1992.

[2]This theorem states that a signal must be sampled at a rate more than twice the highest frequency present in the signal. See B. DeCarle, "A Receiver Spectral Display Using DSP," *QST*, Jan 1992.

[3]B. Widrow et al, "Adaptive Noise Cancelling: Principles and Applications," *Proc. IEEE*, Vol 63, No. 2, Dec1975: 1692-1716

SYNCHRONOUS DETECTION OF AM SIGNALS

What Is It and How Does It Work?

By Mike Gruber, WA1SVF
ARRL Laboratory Engineer

Interest in synchronous detectors seems to be on the rise. This phenomenon is no doubt due in part to the commercial availability of synchronous detection in a popular, moderately priced short-wave receiver, the Sony ICF-2010. Simplified implementation utilizing current IC technology now makes synchronous detectors easier than ever for the experimenter to build and use.

Amplitude Modulation

A firm understanding of amplitude modulation is essential before being able to fully appreciate synchronous detection—especially sideband and carrier phase relationships. A brief summary therefore follows:

The typical AM signal is of a complex nonrepetitive nature, such as music or speech. For the sake of clarity and simplicity, however, the following discussion will consider only an AM signal modulated with a single sine-wave tone. The principles are the same.

An ordinary AM signal is generated by the nonlinear mixing of two signals—the RF carrier and the AF modulating audio, shown in Figs 1A and 1B, respectively. This mixing results in a wave form as shown in Fig 1C. For comparison, linear mixing is shown in Fig 1D.

This nonlinear mixing of the audio and radio frequencies of an AM signal produces two new frequency components—the sum and difference frequencies—commonly referred to as sidebands. The upper sideband (USB) frequency, equal to the RF plus the AF, the lower sideband (LSB) frequency, equal to the RF minus the AF, and the carrier frequency are all shown in the spectral display of Fig 2A.

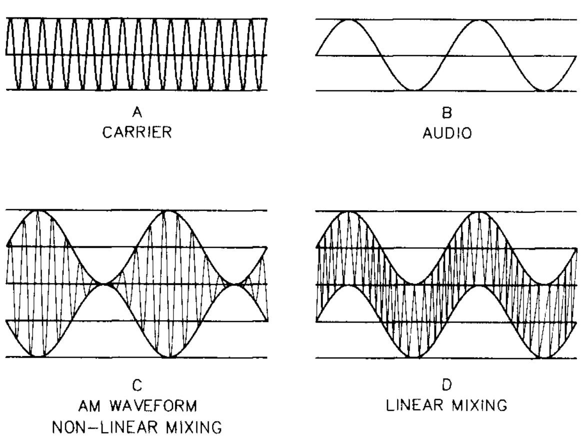


Fig 1

Each of these three components of an AM signal, as with any radio signal, can be represented as a rotating vector commonly referred to as a phasor.[1] The rotational speed of each phasor corresponds to the corresponding signal frequency. For example, the phasor that represents a 7.325-MHz signal (a popular BBC frequency here in North America) rotates 7.325 million times per second. Mathematical convention requires these phasors to rotate in a counterclockwise direction relative to a phasor of lower frequency. The phasor length corresponds to the signal amplitude.

Fig 2B shows three such phasors. They represent the carrier and upper (USB) and lower (LSB) sidebands of the AM signal in the previous figures. The USB, being higher in frequency, rotates more rapidly than the carrier, which rotates more rapidly than the LSB. In Fig 2C, the phase relationship between the three, at any arbitrary instant, is shown with the carrier used as a reference. Phasor rotation for USB and LSB, in this case, is shown as the difference between their actual rotation and the rotation of the carrier phasor. Fig 2D shows these same three phasors rotating about a single common point.

The envelope of the AM signal, or any radio signal for that matter, is at any given point in time determined by the vectorial sum of its phasors at that point in time. As an example, the AM signal depicted in Fig 3A shows one audio cycle of an AM signal broken up into 90 degree intervals. The corresponding resultant phasors for each of these intervals is shown inside the waveform. The

[1]A vector is used to denote magnitude and direction; a phasor, on the other hand, is used to denote magnitude and phase angle. The mathematics of both is the same.

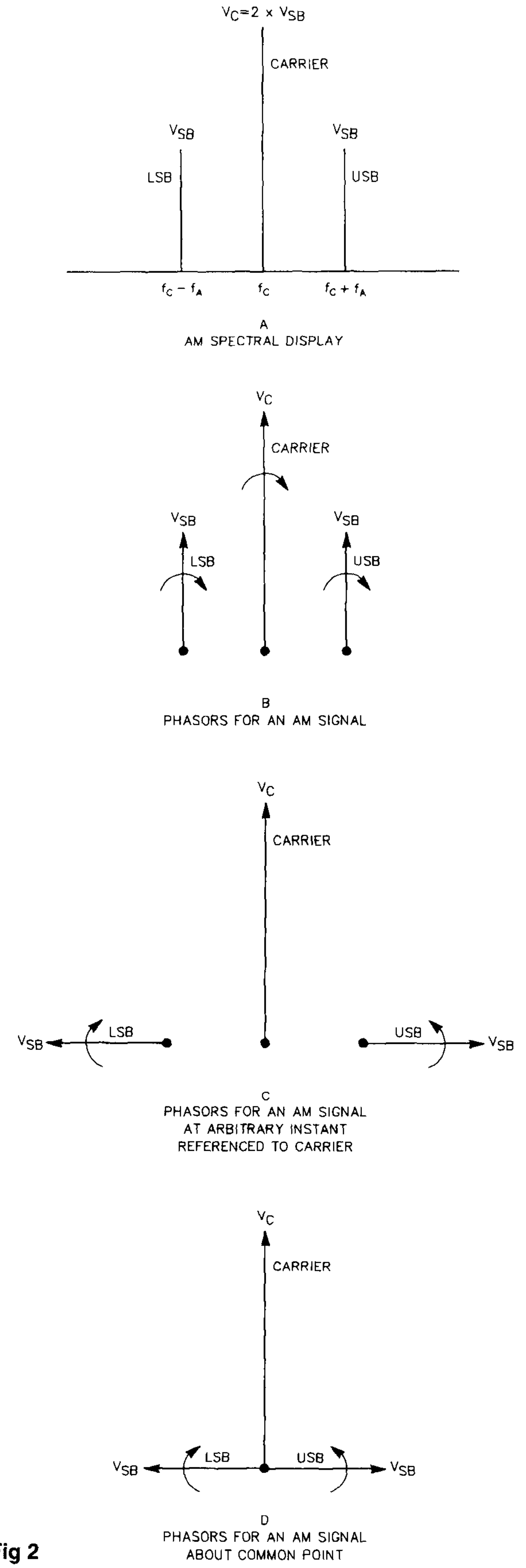


Fig 2

same phasors are also depicted about a common point in Fig 3B. The resultant is determined as follows:

1) 0 Degrees: The LSB and USB are 180 degrees apart from each other and, therefore, cancel each other out. The carrier level at this point is unaffected by the sideband components and is the same as if the carrier were unmodulated.
2) 90 Degrees: The USB, the LSB and the carrier are all in alignment and, therefore, reinforce each other. The RF envelope is the sum of both sidebands and the carrier.
3) 180 Degrees: The sidebands again cancel each other as at the 0 degree point. The carrier level is again unaffected by the sideband components. Note, however, that the sideband phasors have each advanced by 180 degrees.
4) 270 Degrees: Both sidebands are in alignment but are opposite the carrier. The net result is zero—the two sidebands cancel the carrier at this point.
5) 360 Degrees: This point completes one full audio cycle. The phasors have all returned back to the same positions they originally had at the zero degree point.

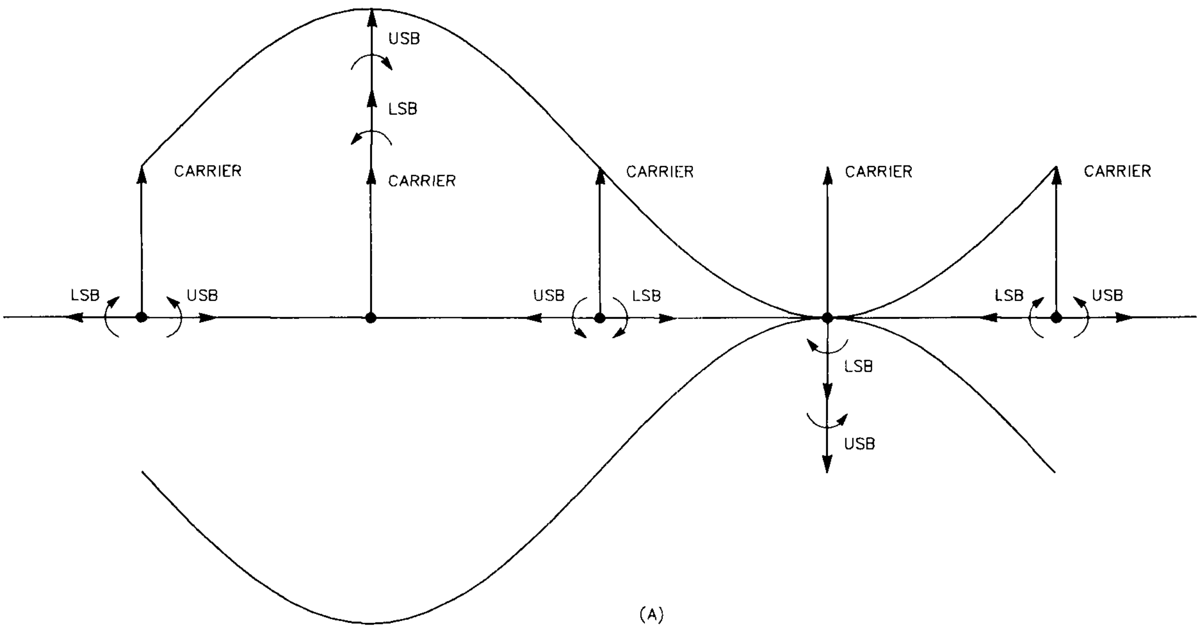
The reader may observe that only the upper half of the AM wave form envelope appears to be defined by the phasors detailed in Fig 3B. One must keep in mind that the carrier phasor is rotating at an RF rate—many times the audio rate of the sidebands relative to the carrier. This rotation of these phasors will cause the lower half of the AM wave form to be swept as time advances. (Picture in your mind's eye a strip of paper passing under a pen on the tip of the rotating resultant phasors. The AM wave form will be produced.)

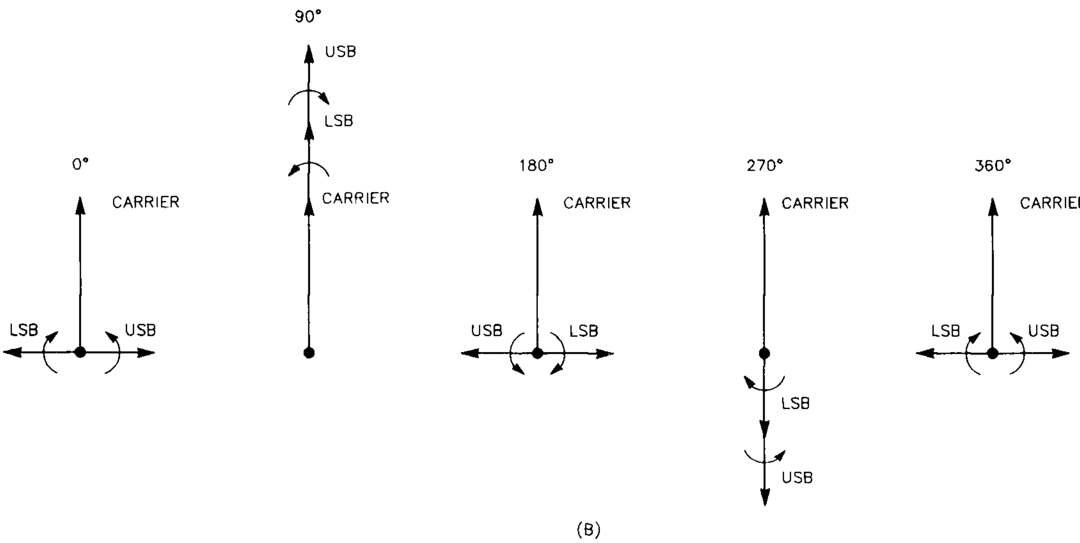
Envelope Detection

As indicated by the previous discussion, an ordinary AM signal consists of three parts, the upper sideband, the lower sideband and the carrier. An envelope detector, such as the common diode detector, is a nonlinear device. This nonlinearity introduces distortion when an AM signal is applied to it, and the desired difference frequency component is produced. In order for this detection process to function correctly, a full carrier is required. Should a selective fade reduce the carrier, severe "overmodulation" type distortion will result as shown in Fig 4; should noise be present on one of the sidebands, noise will also be present at the demodulated output. As we shall see, synchronous detection offers immunity from a selective carrier fade and provides the option to utilize phasing techniques to reject an unwanted sideband.

What is Synchronous Detection

The term "synchronous detection" is somewhat ambiguous. For purposes of this discussion, however, synchronous detection will refer only to a demodulation system having a locally generated carrier that is phase-locked to information derived from the transmitted





Figs 3A & 3B—Vectoral components of an AM signal

signal—usually the carrier. (The incoming AM signal is then mixed, or heterodyned, with the locally generated carrier.) This process essentially recovers the audio from the AM signal by translating to baseband.

The easiest, and probably the most common implementation of a synchronous detector, shown in Fig 5, is simply a phase-locked loop IC that is locked to the carrier of the signal. One notable exception, however, is the Costas loop. First described in 1956 by John P. Costas, K2EN, in a paper entitled "Synchronous Communications," this detection process obtains the necessary phase information for loop lock from the sidebands. This technique can therefore be used to demodulate double sideband AM with a full, reduced, or suppressed carrier. Despite the theoretical advantage of the Costas loop, its added complexity does not buy much when demodulating signals with sufficient carrier to capture and hold phase lock. Since the HF broadcasters will no doubt continue to transmit some form of carrier for some time to come, if for no other reason than phase locking, no fur-

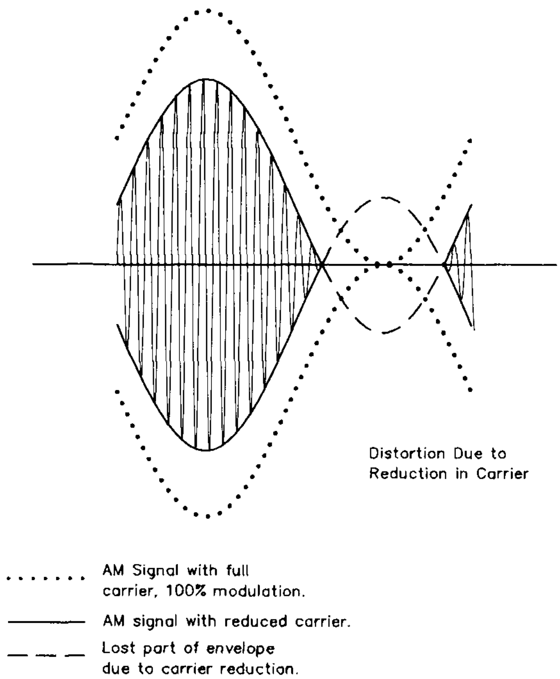


Fig 4

ther consideration will be given to the Costas loop in this discussion. The reader may, however, refer to the bibliography for further information on this topic.

Another similar detection scheme is called the "exalted carrier" method. A carrier sample is first taken from the AM signal and then amplified. It then typically undergoes a hard limiting process before being recombined with the original AM signal. Demodulation can then finally be accomplished by either envelope detection, or, as in true synchronous detection, by a translation process. This technique differs from true synchronous detection in that there is no locally generated carrier. Stability considerations and the simplicity of implementing true synchronous detection with modern IC chips, while offering no significant advantages, have now rendered exalted carrier detection somewhat obsolete.

Two terms often associated with synchronous detection are "synchrodyne" and "homodyne." Both of these terms are extremely ambiguous. Some of the more common meanings for them include: Synchrodyne—a receiver with synchronous detection, and, Homodyne—a receiver with exalted carrier detection. It is most important, however, if encountering these terms to understand their meanings as they apply to the literature at hand.

Why Synchronous Detection?

Synchronous detection offers several significant advantages over the common envelope detector. Unlike envelope detectors that require a minimum of 6.02 dB of carrier above the sidebands for proper demodulation of the signal, synchronous detectors need only sufficient carrier to capture and hold phase lock. Testing was conducted at the ARRL Laboratory with the sync detector of a Sony ICF-2010 to determine the practical level of carrier reduction possible with this circuit. A single-sideband test signal with variable carrier was used. Approximate results with speech and music audio signals are as follows:

	MAXIMUM LSB TO CARRIER	CARRIER BELOW PEP
CAPTURE LOCK:	Not Measured	−11 dB
HOLD LOCK:	9 dB	−16 dB

This characteristic of sync detectors provides increased immunity from selective fading of the carrier relative to the sidebands. This advantage can be significant when the reception of sky-wave signals is considered. In the above case, the carrier could fade by 15 dB before any noticeable degradation in the audio would become apparent. (The potential loss of phase lock dur-

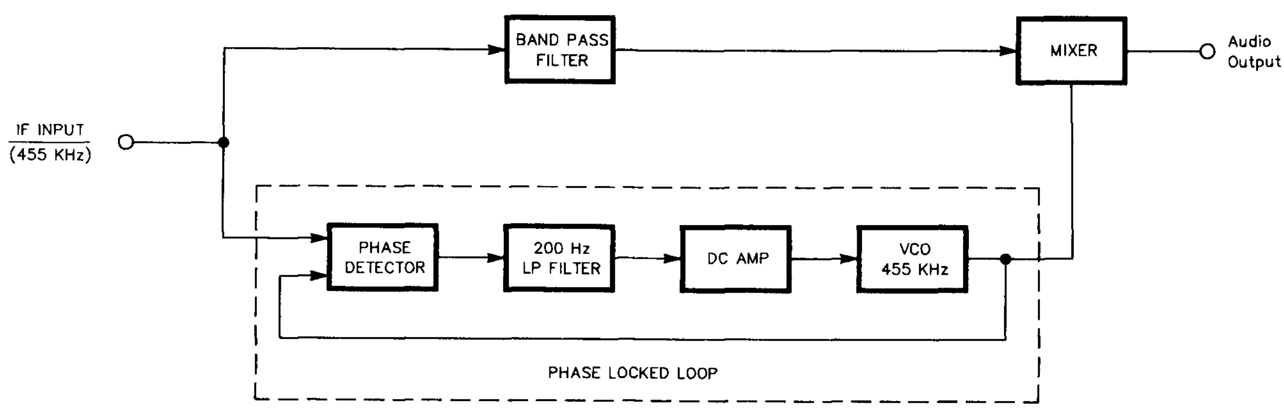


Fig 5—Synchronous detector block diagram

ing a severe fade, however, can result in a burst of noise instead of the less disagreeable reduction in audio typically associated with envelope detectors. The Sony design has provisions to overcome this problem.) Synchronous detectors are also compatible with AM signals having partially reduced carriers and/or one sideband.

Distortion as a result of the receiver being mistuned can in some cases also be eliminated. An envelope detector, being a nonlinear device, works correctly only when both sidebands are mirror images of each other and the carrier is of sufficient amplitude. If the receiver is tuned off to one side of the signal, the sidebands, and possibly the carrier, may be down the skirt of the IF filter. This asymmetry will cause distortion. A synchronous detector, however, if able to achieve proper phase lock on the carrier, is capable of producing an undistorted audio output from the same signal.

Another significant advantage of synchronous detectors is their compatibility with phasing techniques to select between the upper and lower sideband of an AM signal. The advantages include:

1) cancellation and reduction in noise and interference,
2) increased immunity from selective fading and phase shifting between sidebands, and
3) improved signal to noise ratio.

The primary characteristic unique to synchronous detectors that makes this possible is the ability to mix a synchronized carrier that is either in phase (I) or quadrature 90 degrees out of phase (Q) from the original carrier. Audio phasing techniques can then be used to cancel the undesired sideband while reinforcing the desired sideband. See Fig 6 for a block diagram and the associated phasor diagrams. The Sony ICF-2010 design is similar to this concept.

It is clear that synchronous detectors, especially in certain cases, can provide dramatic improvement in the reception of AM signals over the common envelope detector. Selective fading, overmodulation due to carrier fading, heterodyning, phase shift distortion and adjacent channel interference can be reduced or, in certain cases, be eliminated by use of this technology. Although these conditions are most common on the HF bands with crowded conditions and sky-wave propagation, the benefits of synchronous detection are also apparent on the MF AM broadcast band. Amateur AM activity, although enjoying the same benefits of broadcast AM reception, does require retuning if the stations drift or are more than a few hundred hertz apart from each other.

Synchronous Detection Considerations

Synchronous detection can be performed at both radio (RF) and intermediate (IF) frequencies—lending itself to both superheterodyne and direct conversion receivers. Frequency stability of both the transmitter and the receiver, however, becomes far more critical than with envelope detectors. A synthesized or crystal controlled tuning system is optimum and may be required to ensure stability and provide the predictable tuning resolution necessary to capture and hold lock. It is imperative to keep these factors in mind when determining the compatibility of a receiver for an outboard home-brew sync detector.

The frequency range of the phase-locked loop must be limited to reduce the possibility of capture and hold on anything other than the desired carrier. The Sony ICF-2010 often loses lock when tuned away from a station by more than three or four hundred hertz. It is important to note that unless phase lock has been achieved, the signal in most cases is completely unintelligible. An envelope detector, therefore, is best when tuning across the band or attempting to locate a particular station. Once a desired signal has been located, the synchronous detector may then be switched in. (The ICF-2010 incorporates electronic switching to accomplish this function automatically.)

Synchronous Detection Implementation

Probably the first question that comes to mind about synchronous detection is, "Can I build one with a single chip design." The answer to this question is a qualified yes. A simple phase-locked loop, or chip with a phase-locked loop, such as an AM stereo decoder or similar, will do the job. If, however, one desires to incorporate the selectable sideband feature into the project, an audio phase shift and audio summing network must be included in the design.

AM stereo chips for a 450-kHz IF typically operate with a VCO frequency of 3.60 MHz, or 8× the IF frequency. This makes it easy to obtain the quadrature phase necessary for a selectable sideband feature. The ICF-2010 receiver uses a Sony CX-857 AM stereo decoder chip. This IC is compatible with all AM stereo systems (including Kahn and C-Quam). This chip is particularly attractive for synchronous detection applications because it provides easy access to both the I and Q outputs. An additional 90-degree RF phase shift circuit is not required.

The audio phase shifting is no doubt the most difficult part of the circuit to produce. A wideband shift of 90 degrees total is required. The audio phase shift network in the Sony is provided on a separate board by four transistors, six capacitors and twenty-one resistors! This circuit provides all the necessary audio phase shifting and summing to produce the USB and LSB. These outputs are then fed to automatic switching circuitry in the ICF-2010 to provide for the upper or lower sideband depending upon whether the receiver frequency is above or below the carrier frequency.

Over the years, a number of synchronous detector construction projects have appeared in various amateur and electronic publications. They seem to range from vacuum tubes and discrete solid state components to

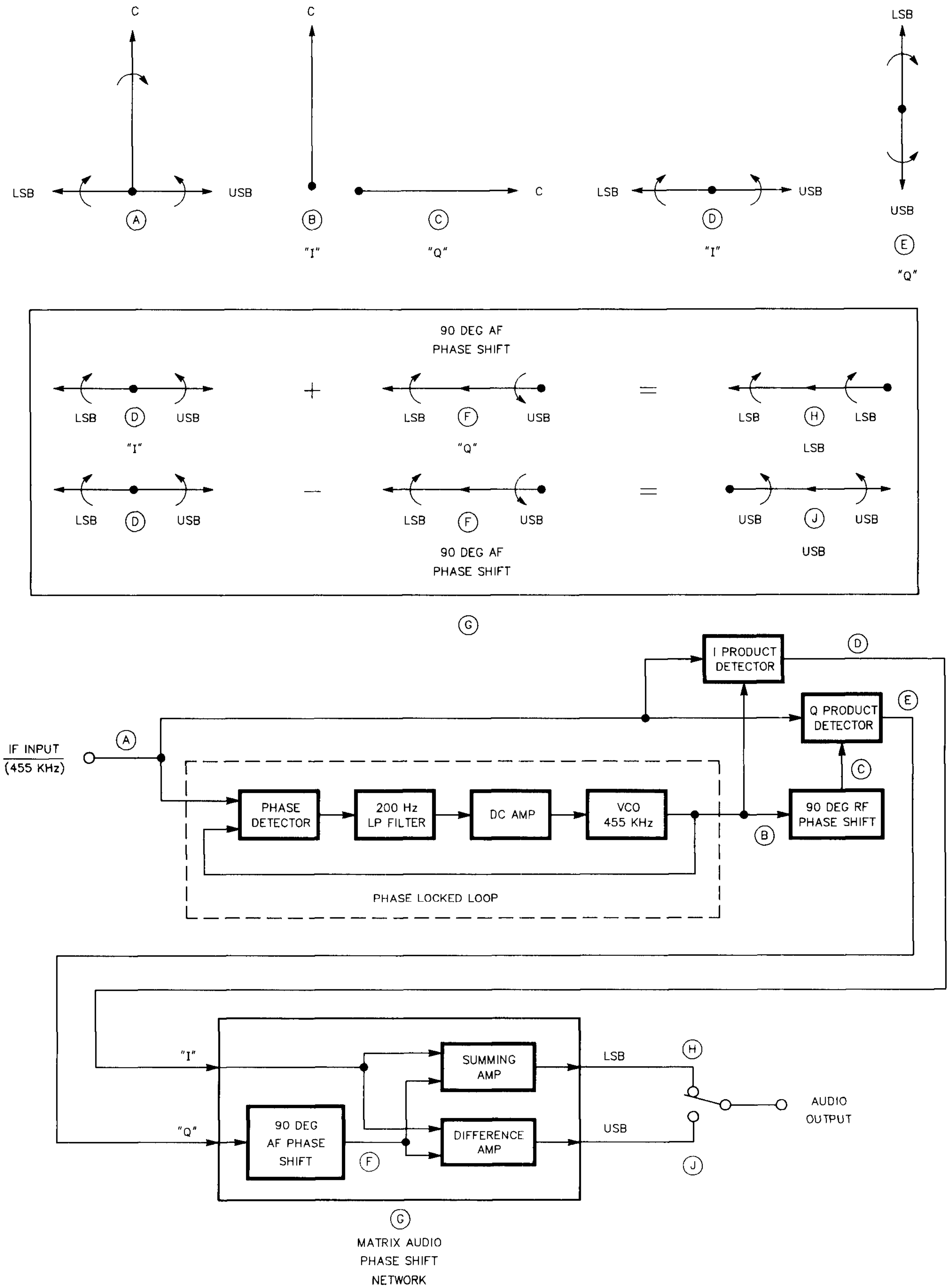


Fig 6—Synchronous detector with selectable sidebands block diagram.

HOW IT WORKS

The input to the sync detector is an AM signal as shown by the phasor diagram in Detail A. The I and Q locally generated carriers are shown by Details B and C, respectively; note the 90-degree phase shift in the Q channel. These carriers are then mixed with the original AM signal to produce the outputs shown by Details D and E. Although these signals are at audio frequencies, the USB and LSB components at this point are not in correct phase for proper audio recovery. A 90-degree audio phase shift is required and its affect is shown in Detail F. As shown by the vector analysis in the box of Details G, the desired sideband can be obtained at this point by either the sum or the difference of the I and Q channels. The desired sideband component from each channel is added in proper phase while the undesired component is simultaneously rejected by cancellation. This is shown in details H and J.

In the example of this figure, the USB is obtained by the difference between of the I and Q channels while the LSB is obtained by the sum of the I and Q channels. Other variations are possible. For example, only the sum is required to obtain either sideband provided an audio phase shift is in the I or Q channel. In this case, the USB is obtained by adding an audio phase-shifted I-channel to the Q channel. The LSB can likewise be obtained by adding an audio phase-shifted Q-channel to the I channel.

It is very important that a 90-degree phase difference exists between the I and Q channels. There is nothing "special" about single 90-degree shifts, however. For example, a plus 45- on one and a minus 45-degree shift on the other would do fine. You can use vector analysis, as shown here, to prove that to yourself.

obsolete ICs. Unfortunately, it appears that an article utilizing current IC technology and a selectable sideband feature, such as in the Sony ICF-2010, has yet to be published. A bibliography of these construction articles appears at the end of this discussion.

A synchronous detector with a selectable sideband feature has been constructed with spare Sony ICF-2010 parts by Steve Johnston, WD8DAS, and Paul Cianciolo, KB1RP. The parts used for their projects were the CX-857 AM stereo chip and the audio Phase Shift Network (PSN) module. Both Steve and Paul report excellent results with their detectors.

The circuit design that they used was essentially a combination of the circuit shown on page 8 of the CX-857 Data Sheet and the sync detector used in the ICF-2010. The schematic is shown in Fig 7. For those wishing to reproduce Paul and Steve's efforts, kit information and a complete source list for Sony parts and literature kit is included as an Appendix.

The construction of the detector is not particularly critical, but can be a bit tricky due to the surface-mount requirements of the IC. Paul used a piece of printed circuit board material mounted inside a Hammond Diecast 1590 B box, but almost anything similar should be satisfactory. Use construction techniques that are suitable for the frequencies involved and keep the coax input lead as short as is practical.

Proper selection of the candidate receiver for the sync detector is essential. The importance of stability cannot be overemphasized. The detector should have a free-running frequency equal to the receiver's IF and, have compatible impedances and signal levels.

The specifications for this project are as follows:

IF Center Frequency:	455-kHz typical, adjustable from 400 to 500 kHz
Input Impedance:	1 MΩ
Input Level:	500-mV typical, 100-mV minimum, 1-V maximum
Output Impedance:	10 kΩ
Output Level:	100-mV typical
Power Requirements:	12 V dc at 20-mA typical, but 4 to 14 V dc acceptable.

Paul and Steve both used a parallel tuned circuit for the VCO. (Paul used a 620-pF silver mica capacitor and a variable inductor centered approximately around 3.15 μH. The inductor could then be tuned to the desired free running frequency.) The ICF-2010, however, uses a

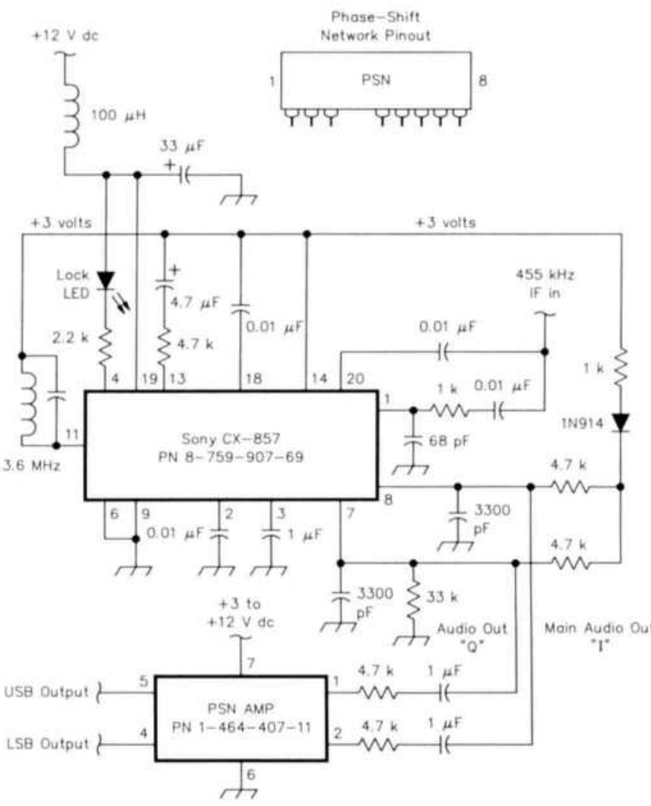


Fig 7—Synchronous detector.

I first became acquainted with selectable sideband synchronous detection when I purchased a new Sony ICF-2010 several years ago. The ability of this feature to reduce or eliminate certain types of distortion and interference was quite remarkable. It quickly became an indispensable asset to my shortwave listening enjoyment.

I soon began to desire an outboard sync detector for my other radios. To this end, albeit several years later, I built the unit described in this article. It has been used with great success in both my R-390A and FRG-7700 receivers. Selective fade distortion is virtually eliminated and the ability to select sidebands has proven to be a real asset. SWLs, AMers and broadcast band listeners are sure to enjoy this form of detection.—*Paul A. Cianciolo, KB1RP*

crystal for this purpose. The crystal version might be preferred in some cases, especially if receiver stability will allow for it due to the reduced locking range. The PLL is less likely to lock onto something other than the carrier of the desired signal.

The Future of Synchronous Detection

The introduction of single-sideband techniques to the HF broadcasting bands was established in the Resolutions and Recommendations of a World Administrative Radio Conference. Recommendation No. 517 (HFBC-87) specifies an initial 6 dB below PEP of carrier reduction. By the end of the year 2015, the final carrier reduction of 12 dB below PEP is to take effect and all DSB transmissions are to cease on these bands.

The initial 6 dB of carrier reduction will be only partially compatible with envelope detectors, and the final 12-dB reduction will require alternative detection methods. Synchronous detectors can be made fully compatible with such reduced carrier signals and will no doubt be the most popular detection technique for these new signals. It is expected, therefore, that the next generation of budget short-wave receivers will help proliferate at least some form of synchronous detection as full carrier DSB is being phased out. The resulting new chips and dedicated circuitry components should make for easy and inexpensive synchronous detection for mass produced radios as well as the home hobbyist.

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NOTE: The first nine and last two articles are primarily construction projects featuring complete construction details.

1. Dave Hershberger, W9GR, "Build a Synchronous Detector For AM Radio," *Popular Electronics*, April 1982, pp 61-71. Correction, July 1982, p 6. *(An additional error appeared in this article that was never published. The* envelope *position of switch S1B should be grounded. It is shown being unconnected.)*
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5. H. F. Priebe, Jr., K4UD, "Auto-Product Detection of Double-Sideband," *Ham Radio*, March 1980, pp 58-61.
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21. "AM Synchronous Demodulator" *Electronics & Wireless World*, September 1989, pp 858-860.
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Appendix

Parts

Sony CX-857 AM Stereo Chip
P/N 8-759-907-69 Price: $19.87
Sony Phase Shift Network
P/N 1-464-407-11 Price: $19.87

Literature

Sony ICF-2001D/2010 Service Manual
Sony CX-857 AM Stereo Decoder Data Sheets

Sources

Sony Service Company
Part Division
8281 NW 107th Terrace
PO Box 20407
Kansas City, MO 64195-0407
816 891-7550

Steinberg Electronics, Inc
2220 North Broad Street
Philadelphia, PA 19132
800 523-0894

Synchronous Detector Kits

Steve Johnston has indicated that he will be offering this project in kit form. The price is $139.00 for the kit and $199.00 for an assembled and tested unit. The cost for a single board is $15.00.You can contact him for further information at:

Steve Johnston
PO Box 3420
York, PA 17402-0420

The 4×3×5 Filter - Revisited

By Dick Jansson, WD4FAB
1130 Willowbrook Trail
Maitland, FL 32751-4837

Those amateurs who have been communicating through the OSCAR satellites long enough to remember OSCAR 8, will well remember experiencing some of the problems in operating their 70-cm receivers. This was caused by being desensed by their 2-m transmitters, when operating in Mode J. There are two basic phenomena at work in this particular ailment (to be discussed forthwith) and several cures that could be applied. With perseverance, most of us overcame these desensing problems to operate and enjoy AO-8.

With the advent of the more modern transceivers and selective GaAsFET preamplifiers, we have generally been able to forget the desense monster. AO-13 Mode J, and especially PACSAT Mode J communications, have again raised the issue of 70-cm receiver desensing. Let us examine the factors at work here, revisit some of the cures, and even look at a modern, commercially available version of one of these cures.

Receiver desensing is basically caused by the unwanted saturation of a receiver front end by RF energy from a nearby transmitter. This is most typically witnessed when the transmit frequency is lower than the receive frequency, particularly enhanced when the transmitter frequency is at ≈⅓ of that of the receiver. If the transmitter has unexpected and unwanted emissions at frequencies other than the desired one, it will commonly be at a third harmonic, and can affect the receiver operation with Mode-J weak signals. Conversely, if the receiver selectivity is such as to promote a modest sensitivity at the normal transmit frequency, even though the transmitter frequency is clean and well removed from that of the receiver, weak signal reception will be affected.

Curing these problems requires a degree of technical detective work, which is often viewed as an interesting challenge by many amateurs, even though the desense phenomena is annoying. The observer needs to look for clues regarding the culprit RF source/cause/path. First, a determination should be made of whether the desensing raises the noise floor of the receiver, even to showing an S meter reading, or is the problem one of the pure blocking of the signal. While not hard-and-fast rules, generally the transmitter is the culprit for the noise floor problem while the receiver blocking comes from poor selectivity. Obviously the use of a spectrum analyzer would be helpful. Getting a clear identification of the cause of the desensing is essential to applying the proper cure.

With the use of a linear amplifier, heterodyne generated SSB transmitter emissions generally promote very clean RF signals. Usually this allows excellent operation without emission concerns in the Mode J receive band. Really persistent emission problems can be corrected with the use of a cavity filter, much like one section of a repeater duplexer. One PACSAT operator recently corrected his Mode J desensing problems by merely giving his 2-m transmitter a really good tuning and eliminated a nasty spur. Having to rely on the cavity filter to reduce a transmitter spurious emission is considered to be a last resort.

Solving receiving desensing problems is mostly a case of providing enough isolation between the transmitter signal and the receive antenna. Sometimes the placement of the two antennas farther apart will provide the solution. The case of a "wide" receiver is not com-

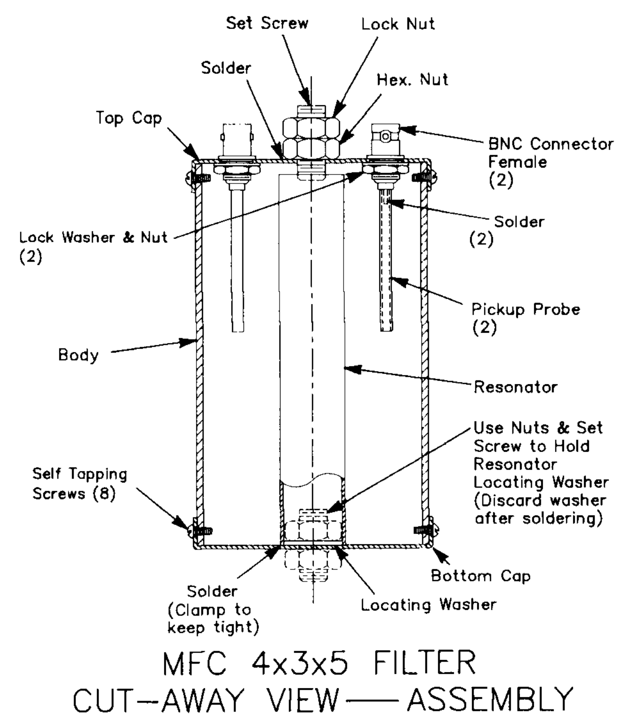


Fig 1

monly seen these days but the solution is one of improving the selectivity of the receiver.

In the days of operating through AO-8, the only practical low-noise preamplifiers that we had available used MRF-901 bipolar transistors, which provided notoriously broad-band reception. As a result, the 70-cm receivers were quite receptive to the 2-m transmitter signals. GaAsFET preamplifiers have quite good values of 3rd order intercept points, and reasonable tuned circuits in their inputs. Most often these conditions will allow sufficient receiver isolation.

PACSAT operation on Mode J, using compact, dual band antennas has again raised the issue of desensing, especially since this type of antenna has increased coupling between the two bands, compared to two separated antennas. Even the normally good selectivity of our modern preamplifiers and transceivers is not often sufficient for operations with the compact dual band antenna.

The 1992 edition of *The ARRL Handbook* shows the original capacitively coupled 4×3×5 MHz Filter for Mode J, on page 23-32. This filter was fashioned after a design by Joe Reisert, W1JR, and Jay Rusgrove, W1VD, and promoted by the ARRL for AO-8 Mode J operations. Many of us constructed this filter, and we spent many hours doing so. This was a real labor of love for some of us, and the results were well rewarded.

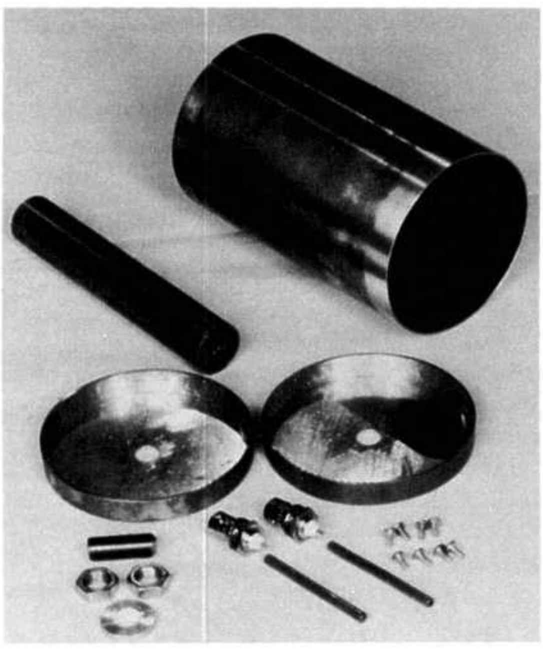
With the resurgence of PACSAT Mode J activities, there appeared a need for a more easily constructed simple filter. While excellent band-pass filters are available, with really good stop-band characteristics, the insertion loss of these filters makes their use in front of a good low-noise GaAsFET preamplifier rather prohibitive. Placing such a filter after the preamplifier will leave the preamplifier itself susceptible to input signal overload, although protecting the receiver. Single cavity, simple filters, such as the 4×3×5 unit have been measured to have insertion loss values less than 0.15 dB, meaning that their use in front of a 0.3-dB NF preamplifier would be quite acceptable. At the same time, such simple filters can provide upwards of 50-60 dB of rejection for the 2-m transmitter signals. My 1979 4×3×5 filter was silver plated and measured at less than 0.1-dB insertion loss!

To the rescue for all of these problems comes Microwave Filter Co, Inc (MFC).[1] This company has developed their Model 9397 filter specifically as a kit for the amateur satellite Mode J operator.[2] I was a little concerned about the kitting of this filter, remembering all of the hours that I spent building my own unit in 1979. The photo illustrations show you some pretty simple construction and neat results. Considering that I really took my time building this sample unit, and spent a lot of time taking the photographs of the work, I was surprised that I spent only 1+ hour on the assembly. The MFC 9397 version of this old friend of a filter employs some very clever design concepts that permit its rapid and certain assembly.

The MFC 9397 kit provides a drawing that is quite ample in illustrating the assembly required by the builder, Fig 1. Fig 2 shows the parts for the kit as it is received,

[1]Microwave Filter Co, Inc (MFC), 6743 Kine Street, East Syracuse, NY 13057, tel 1 800-448-1666.

[2]Model 9397 filter kit, $70 plus shipping and handling. Allow 30 days for delivery.



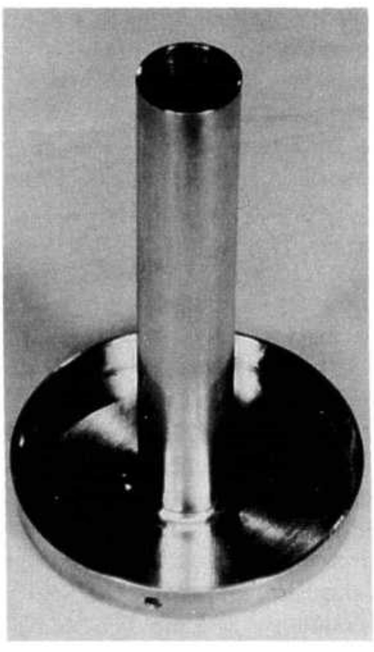
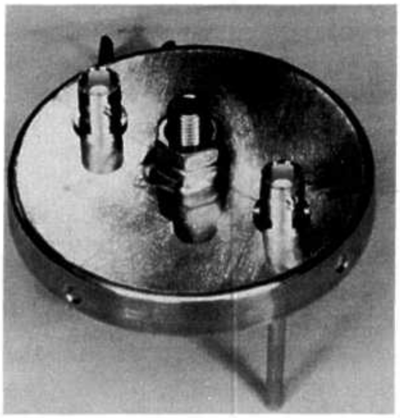
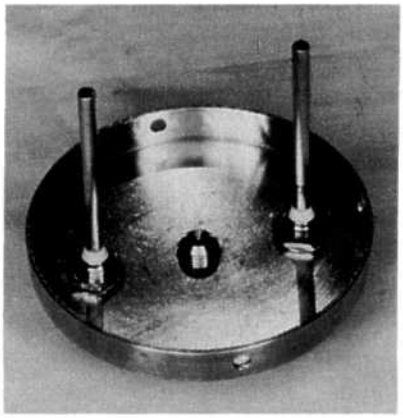


Fig 2–left, Fig 3–top, Fig 4–middle





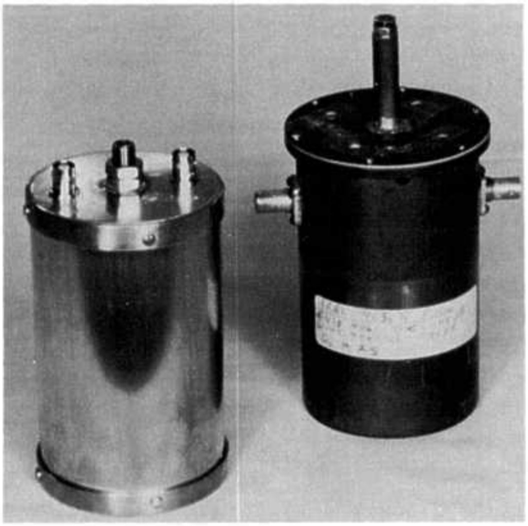


Fig 5–top left,
Fig 6–top right
Fig 7–left

including genuine copper tubing. Carefully follow the drawing instructions and locate and drill the four screw holes in each end cap. Similarly, drill the mating smaller holes in the large tube. Be sure to remove all burrs from the drilling operation, as these parts have a very close fit, and you do not need the aggravation of the parts not fitting together because of a drill burr.

The second step, Fig 3, is to assemble the large set screw, nuts and aluminum flat washer to the bottom cap. The center conductor fits over the washer rather tightly. Be sure to use some fine steel wool to scrub the adjacent surfaces of the center tube and the end cap. Also, make sure that the center tube is positioned perpendicularly to the end cap.

At this point the builder needs to solder the center tube and the end cap together. *Do not use lead-tin solder* for this operation. The recommended solder to use is that sold for plumbing solder and it contains a small amount of silver, with the remainder being tin. The electrical properties of the silver-tin solder are absolutely superior to the lead-tin alloys. Moreover, the silver-tin solder readily wets the joint and flows very nicely with the help of a little resin, noncorrosive flux, Fig 4. After soldering the two parts together, remove the setscrew and nuts. Careful clean the residual flux from the assembly.

Clean the solder area on the center of the top cap and using the large setscrew to concentrically locate one of the nuts, solder the cap and nut together. This will form the cavity tuning adjustment, along with a free nut for locking purposes, Fig 5.

Locate the two BNC connectors, nuts, and the small copper coupling tubes. The fussy builder may want to substitute the equivalent TNC connectors for the BNCs, but that will be a matter of personal preferences. Solder the coupling tubes to the connector center conductor and place the connectors on the top cap, tightening the nuts, Figs 5 & 6.

Mount the end caps on the large tube, locating them to the previously drilled screw holes. Attach the caps to the tube using the self tapping 4-40 screws provided. That step completes the filter assembly.

Fig 7 shows the assembled filter next to the 1979 filter. This filter will easily tune with the use of your 70cm receiver and any external signal source, even those from a satellite. Peak tune the filter and the job is done and ready for service.

VHF+ TECHNOLOGY

By Geoff Krauss, WA2GFP
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In spite of a number of calls and letters from readers wanting to know when my next column would appear, it is only after a six- or seven-month hiatus that I find myself with enough time to actually sit down in front of the word processor—the very end of June. I've missed all of the Spring VHF/UHF hamfests, conferences and contests, too. The "eater of time" has been the relatively sudden change in the state of world politics. After years of hoping for a lessening in the nuclear standoff situation, who would have expected that we would have such a sudden and complete reversal as to make the electronics business so abjectly subject to the amount of terminated and nonawarded contracts, and the resultant layoffs that we in the US have seen in the last half year or so? I think it's safe to say, from the gist of communications I've had over the last decade or so that I have been writing this column, that the majority of my readers derive at least a part of their income from, and are thus professionally involved in, the electronics industry. In fact, if one looks at the historical situation of the industry, most US electronics workers today have never known a time (ie, WWII and after) when a good proportion of total industry work was not the result of tax-dollar expenditures. In short, though, the desirable end of Cold War "hostilities" signaled by the break-up of the former USSR has led to much-reduced defense expenditures and job levels. Many of those whose jobs have survived must spend considerably more time on job-related tasks, and have less time for hobbies of any kind. My wish for those of my colleagues who find themselves caught up in this situation is that all have only the greatest success in the important task of economic survival, and that only enhanced hobby aspects follow thereafter!

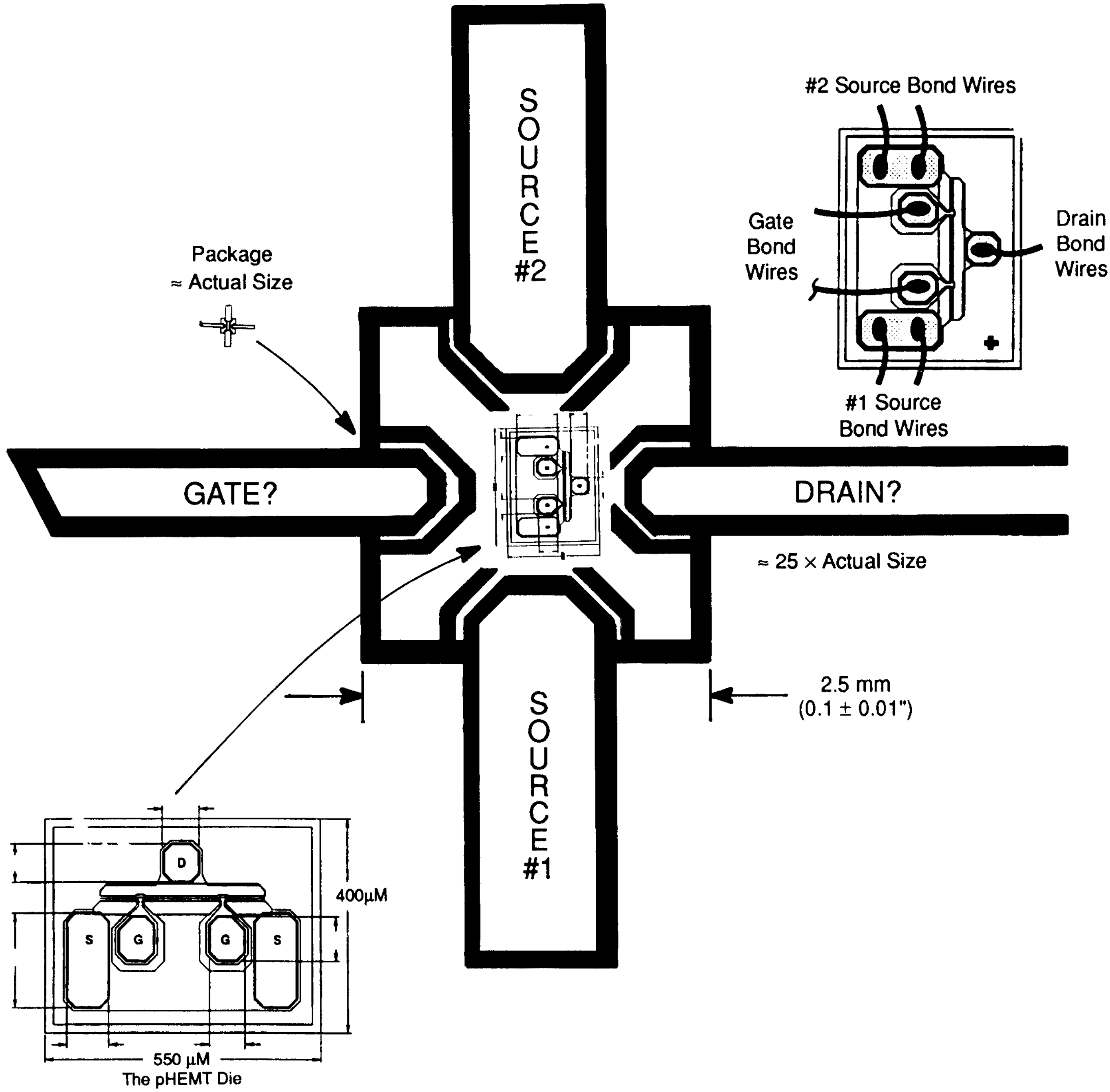
About the only Amateur Radio I've had time to think about has been one aspect of our VHF contests, as related to me by fellow VHF+ers who stopped me in the hall at work—the change in the contest rules to set up a separate Rover category that may be taken advantage of by a few contestants; they operate from a permanent site (home), or a semi-permanent location at which they have a big antenna, mains power, etc, to rack up a large score, then go out and do some true-rover operations, with auto, battery or generator power and fully transported equipment. Unfortunately, the present rules allow the two score portions to be lumped together and, since some of it was as a rover, the whole score is counted in the Rover category. The operator, or group, who goes out and only rovers cannot hope to attain the same magnitude of score, and is discouraged when the whole rover idea was to encourage people to go out in the field and put "rarer" grids on the air. See the May/June 1992 issue of the *NTMS Newsletter* for further discussion.

One reader who has done some very interesting work on the job is Michael Martin, AA4SG, who wrote a letter to me about some relatively low-cost (ie, $15-30) HEMT devices he came across in the course of a search for very-low noise figures at about 1 GHz. After pointing out that the Toshiba 2SK1325 is probably a better device than the heterojunction 2SK1619 GaAsFET I had mentioned some columns ago, Mike finds HEMTs are much better performers even though "most HEMTs are intended for applications well above 2 GHz and often have a maximum frequency of oscillation exceeding 20+ GHz." Mike notes that he has "had more than my share of 20+ GHz oscillations to get rid of, and believe me, the least of parasitics can cause problems." He lists some noise figures at 2 GHz (as manufacturers do not generally list NFs at lower frequencies) as:

Device	Manufacturer	2-GHz Noise Figure	at Vds/Is
ATF-35076	Avantek (Now)	0.13 dB	1.5 V dc/10 mA
ATF-35176	Avantek	0.14 dB	1.5 V dc/10 mA
ATF-35376	Avantek (H/P)	0.17 dB	1.5 V dc/10 mA
S8900	Toshiba	0.17 dB	2.0 V dc/10 mA
MGF-4310	Mitsubishi	0.30 dB	2.0 V dc/10 mA
NE32484A	NEC (California)	0.31 dB	2.0 V dc/10 mA
NE32184A	NEC (Elect Labs)	0.37 dB	2.0 V dc/10 mA
FHX35LG	Fujitsu	0.40 dB	3.0 V dc/10 mA

Now all of these devices may not be true HEMTs—NEC/CEL calls their NE32484A a "Pseudomorphic HJ FET." There is potential confusion here between HJ (heterojunction) FETs, which have been with us (as the sole device type called GaAsFET) for a number of years, the PHJ FET of the NEC nomenclature (also includes devices such as the NE42184A, NE76084 and NE76184A, according to one NEC/CEL ad I have here) and the PHemt (Pseudomorphic High-Electron-Mobility Transistor) as some of the devices by Avantek, GE, Microwave Technology (MwT) are denominated. By whatever name, though, any device with 12-GHz performance of less than 1-dB (typ 0.6 dB) NF at an associated gain Ga of 8-13 (typ 11) dB is *hot!*

Another problem, besides type naming, is packaging. Most readers realize that the last two numbers in each of the above-listed ATF and NE devices is the package designation (and also wonder why the other manufacturers don't follow this simple rule). The types are generally common to the industry and, universality, is a great help, however, some manufacturers do not package their devices (my employer, while being in the forefront of the HEMT pack, does not generally make its devices available for noncompany use and rarely packages the bare device die, so I cannot even use them as a source of UHF/SHF goodies!). A look at the accompanying drawings of device package and die outlines is instructive and I've used a half-page to show some figures scaled, as best I can on



The pHEMT Die

an office copier, to give appreciation of some of the problems. The packaged device is just about the size shown in the upper left—a microwave ceramic package of 0.1 inch on a side; the central "blowup" is the same packaged device, seen from the bottom, after about 25× magnification. This central view will show some of the lead-under-case details, which are overlaid in the center by a to-scale version of the device chip which is inside the package; a rotated and blown-up figure at the lower left shows the die/chip—note the side dimensions of 400 × 550 microns (millionth of a meter)! If the chip is put into the upper-left full-size drawing, it would appear as a small dot in the center of the package. The upper right view shows the same chip with its source/gate/drain bond-wire attachment locations. While you are looking, note that I have labelled the horizontally opposed pair of leads the gate? and the drain?, since some manufacturers (predominantly Japanese) diagonally cut the gate lead, and others (often US) cut the drain lead. *Always* refer to the data sheet for the device you are using to check, double check and triple check the orientation (you generally get one chance to connect in the device—removal and resoldering may be impossible).

There are few, if any, of us who have the equipment and/or skill needed to work with (move, attach, wire-bond, etc) the bare dice or chips. This may change in the future, but I would not bet a lot on that! History tells us that even the best hams of the 1920s did not have the tools and techniques to make their own vacuum tubes, and few individuals seventy years later can build or even repair/revacuum any form of tubes. I suspect that the same will hold for submillimeter-sized semiconductor chips. There is a chance that some VHF+ers who work for semiconductor fabricators may eventually be able to mount and wire some relatively small amount of chips for themselves and other SHFanatic, or even make up multichip microwave modules (eg, complete subsystems, with power supply, protection and the like on a single master substrate in a single package), but not very likely in this decade. That is why, even though the MwT-H4 chip (Microwave Technology, 4268 Solar Way, Fremont, CA 94538) is characterized between 1-18 GHz, it is very unlikely that you will soon get to use it at 3.0 V ds/12 mA=50% idss, to obtain data sheet NF/Ga values of:

Band (GHz)	902	1296	2.3	3456	5760	10368
NF (dB)	0.093	0.096	0.13	0.23	0.36	0.65
Ga (dB)	24	22	18	16.5	14	11.5

Because these noise figures are so low as to become almost meaningless, we can obtain a little better insight by converting to noise temperature, in degrees absolute, or K, by use of the formula: Tn(K)=290(F-1) where F is the noise FACTOR, or F=antilog(NF/10). So:

Freq	902	1296	2304	3456	5760	10368
Tn(K)	6.3 K	6.5 K	8.8 K	15.8 K	25.1 K	46.8 K

Other problems still abound, such as how to handle these devices safely (since ESD, or electrostatic discharge, can easily generate sufficient voltage spikes to punch through the extremely thin 0.1 to 0.5 micron gate oxides used), how to properly mount the devices, and most especially how to provide an input matching circuit with very, very, very low insertion loss (which loss directly adds to the stage NF and can often be in the 0.1-0.5 dB range itself) while still providing as much selectivity as possible, so that out-of-band signals do not overload or damage the supersensitive low-noise amplifier. It is in this latter area that VHF+ers have made some progress in the last decade, and have the best chance to make further advances in the near future.

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ARRL BOARD RECOMMENDS CREATION OF UNATTENDED DIGITAL STATIONS

Complying with the recommendations of the ARRL Committee on Amateur Radio Digital Communications, the ARRL Board of Directors adopted the following motion at their July meeting:

It was moved by Director Comstock, seconded by Director Heyn, that the General Counsel, with the assistance of the Executive VP and the staff, is authorized to prepare a draft Petition for FCC Rulemaking to permit the operation of a new category of amateur station, "unattended digital station," on RTTY/data frequencies below 30 MHz. Only amateur stations under the active control of a control operator would be permitted to communicate with unattended digital stations; unattended digital stations would not be permitted to engage in one-way communications; and appropriate safeguards would be required to prevent unattended digital stations from causing harmful interference to other amateur stations. The draft is to be circulated to the Executive Committee for review and final approval before filing. Further, the Digital Committee is requested to continue its study of the issue of unattended digital operation, with the objective of developing future recommendations for increased flexibility of operation of this class of station.

Upon adoption of the motion, Directors Turnbull, Burden, McConnell and Grauer requested to be recorded as voting against it.

Digital Committee Recommendations

The full text of the Digital Committee's recommendations follow.

Recommendations

I. Unattended fully-automatic operation of amateur digital stations should not be authorized below 30 MHz.

II. The FCC rules should be amended to allow unattended semi-automatic operation of digital stations on any frequency on which digital modes are authorized. Unattended semi-automatic stations may not initiate a contact, either with another station or via an undirected broadcast. An operator initiating a contact with an unattended station must first ascertain that no interference will be caused to existing communications and must monitor the progress of communications. If it becomes evident that the communications with an unattended semi-automatic station is interfering with other amateur communications, then the link with the semi-automatic station must be discontinued. An unattended semi-automatic station must be equipped with a time-out timer to insure that no signal is transmitted longer than five minutes in the event of the malfunction of control equipment or the loss of contact with the initiating station. Suggested wording for such an amendment is included in the appendix.

III. The FCC rules should be amended to allow the use of modem-dependent codes for the purpose of efficient data compression and error control on HF radio channels. The bandwidth of such signals should be restricted to 500 Hz, below 28 MHz and 2000 Hz between 28.0 and 28.3 MHz. The appendix suggests specific wording for the recommended rule change. A station using a modem-dependent code must still comply with 96.119 Station Identification.

IV. The League should publish a comprehensive tutorial-style operator's guide for HF digital operations clearly defining acceptable operating practices. Such a manual would delineate currently used informal subbands for the various modes and styles of operation and the good operating practices that are required for effective mutual cooperation and coexistence. This Committee will make specific recommendations for the content of this guide.

V. he League should publish technical standards or guidelines for the characteristics of signals generated by digital mode stations for the purpose of achieving the best possible use of the HF spectrum. *QST* should be used as a forum to educate the amateur community on the benefits and means of achieving acceptable signal quality and should review the technical characteristics of digital mode products with respect to published standards. This Committee will make specific recommendations for these technical standards.

Appendix A

The following is suggested wording for an addition to Part 97 authorizing unattended semi-automatic digital mode operation.

97.3 Definitions

( ) Unattended Digital Station - A station in the amateur service using an RTTY or data emission that is operated without a control operator present.

97.216 Unattended Digital Station

(a) Any amateur station licensed to a holder of a General, Advanced or Amateur Extra Class operation license may be an unattended digital station.

(b) An unattended digital station may operate on any frequency below 30 MHz that is authorized for RTTY or data emission for the class of operator license held.

(c) An unattended digital station may only use those RTTY or data emissions authorized by 97.305 and 97.307.

(d) No unattended digital station may initiate a contact with another station or may broadcast any undirected signal.

(e) The transmitter of an unattended digital station must be equipped with a time-out timer that will insure that no signal is transmitted for longer than five minutes in the event of the malfunction of control equipment or loss of contact with the initiating station.

(f) Any amateur operator initiating contact with an unattended digital station must first ascertain that no interference will be caused to existing communications, must be present for the duration of the contact, and must discontinue the contact if it becomes evident that communications with the unattended digital station is interfering with other amateur communications.

Appendix B

To encourage improvements in digital mode communications and especially to improved spectrum utilization on amateur HF bands, Part 97, 97.307 (f) (3) and 97.307 (f) (4), should read as follows:

(3) A RTTY or data emission using a specified code listed in 97.309 (a) of this Part may be transmitted. The symbol rate must not exceed 300 baud and, for frequency-shift keying, the frequency shift between mark and space must not exceed 300 Hz. A RTTY or data emission using an unspecified digital code under the limitations listed in 97.309 (b) of the Part also may be transmitted. If an unspecified digital code is transmitted the authorized bandwidth is 500 Hz.

(4) A RTTY or data emission using a specified code listed in 97.309 (a) of this Part may be transmitted. The symbol rate must not exceed 1200 baud and, for frequency-shift keying, the frequency shift between mark and space must not exceed 1 kHz. A RTTY or data emission using an unspecified digital code under the limitations listed in 97.309 (b) of the Part also may be transmitted. If an unspecified digital code is transmitted the authorized bandwidth is 2 kHz.

ADAPTIVE 9600-BIT/S MODEM ADDED TO DSP-12

The recently released version 1.99 firmware for the L. L. Grace's DSP-12 features a new adaptive 9600-bit/s modem that is intended to provide a dramatic improvement in 9600-bit/s performance, especially with UoSAT-22. The new release also incorporates some bug fixes and other enhancements including a new pipe mode.

The firmware may be downloaded from CompuServe's HamNet (Library 9). Its file name is GCE199.ZIP and it contains downloadable RAM and PROM image files and release notes.

SINGLE-PORT DSP MULTI-MODE CONTROLLER RELEASED

The DSP-1232 Multimode Data Controller is now available from AEA. A single port version of the DSP-2232, this controller features:

- Packet, AMTOR, Baudot, ASCII, Morse Code, NAVTEX, WEFAX.
- All satellite digital modes.
- K9NG/G3RUH 9600 bit/s; 2400 bit/s.
- Automatic identification of most types of digital signals.
- Software DSP modems (future upgrades installed on EPROM chips). Upload new modems into RAM from disk or telephone BBS.
- Software switchable radio port selection.
- Complete 18-kbyte personal mailbox accessible through packet and AMTOR.

This new controller eliminates the need for external modems for satellite work or high speed data, as all the modems exist in software.

The suggested list price is $799.

DISCLAIMER

The descriptions of hardware and software contained in this column are merely reports of what is being offered in the Amateur Radio marketplace. Your Gateway contributing editor, *QEX* and the ARRL in no way warrant the operability, functionality, suitability, or even the existence of what is being offered.

GATEWAY CONTRIBUTIONS

Submissions for publication in Gateway are welcome. You may submit material via the US mail to 75 Kreger Dr, Wolcott, CT 06716-2755, or electronically, via CompuServe to user ID 70645,247, or via Internet to horzepa@evax.gdc.com. Via telephone, your editor can be reached on evenings and weekends at 203-879-1348 and he can switch a modem on line to receive text at 300, 1200 or 2400 bit/s. (Personal messages may be sent to your Gateway editor via packet radio to WA1LOU@N4GAA or IP address 44.88.0.14.)

The deadline for each installment of Gateway is the first day of the month preceding the issue date of *QEX*.