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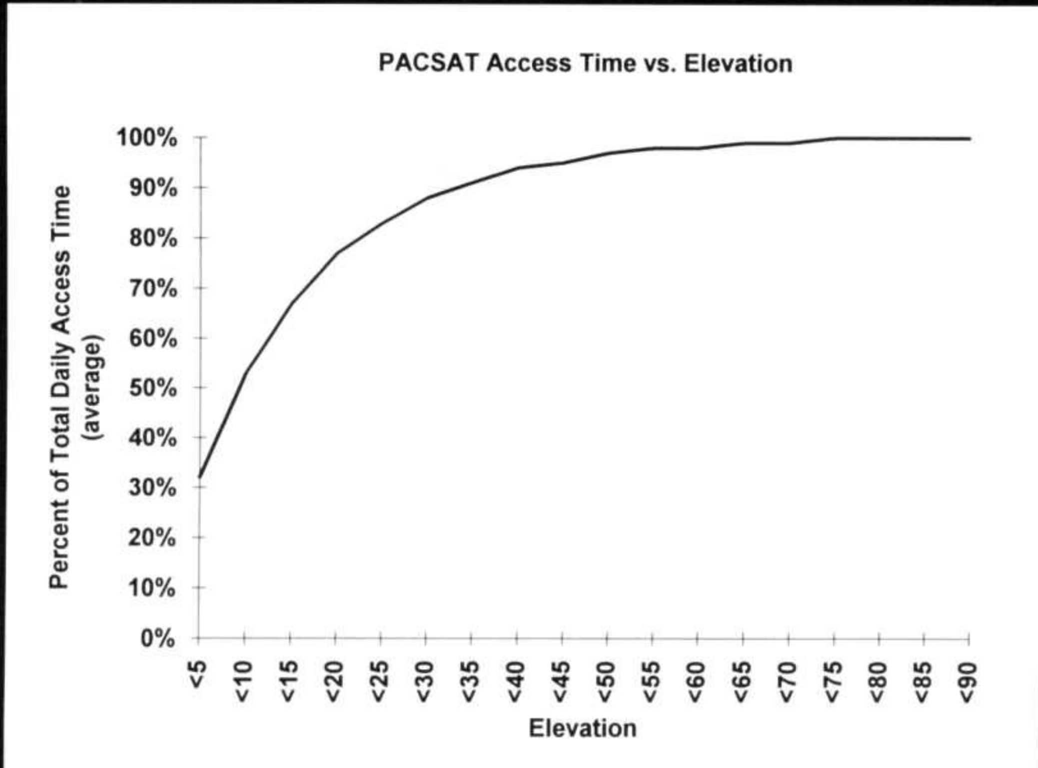
ARRL Experimenter's Exchange

February 1993



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About the Cover:

Think you need a kilobuck antenna system for low-earth-orbit satellites? Not so, as NK6K explains on page 14.

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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*.

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

Empirically Speaking...

EMI

No one active in Amateur Radio needs to be told that electromagnetic interference (EMI) is increasingly a problem to amateurs. Whether the interference is from the amateur's transmitter to consumer electronic devices or from those devices—and other sources—to the amateur's receiver, the residential amateur station is awash in EMI problems.

The ARRL has stepped up its efforts to deal with electromagnetic compatibility (EMC) problems over the past few years. The attack on these problems occurs on several fronts, including education of amateurs, consumers and industry, ARRL representation of amateurs in industry EMC organizations and on EMC committees, and a strengthening response by local RFI committees and the ARRL Field Organization—principally Technical Coordinators and Technical Specialists.

The ARRL book, *Radio Frequency Interference: How to Find it and Fix It* is one significant component of the overall mix of ARRL EMC activities. One of the things that book attempts to do is provide "cookbook" fixes to common RFI problems. Most of those fixes are the result of field experience; what works in one EMI situation is likely to work in another. It occurs to us that *QEX* readers might be able to contribute in this area. Cleaning up EMI is something many amateurs have had to do, and it's likely that the experimenter community within Amateur Radio does it best, primarily because that community is infused with the greatest level of understanding of the technical issues involved with EMI. So, we'd like to hear from you about your EMI successes. If you've solved a particularly knotty problem using original techniques, or even just by applying a combination of techniques, that information has value to the ARRL EMI program. Ed Hare, KA1CV, Laboratory Supervisor at ARRL Headquarters, would like to hear reports of EMI cases with as much technical detail as is available.

Another part of the overall EMC problem is that of measuring the susceptibility of devices to EMI. For example, there does not appear to be any particular standard way to test telephones for susceptibility to common-mode RF signals—by far the biggest culprit in cases of telephone interference. So it's difficult to say that telephone A is better than telephone B. And that's something we'd like to be able to do to encourage manufacturers of EMC deficient devices to improve their products.

Useful data would also be obtained from measurements of signals induced by amateur transmissions in house wiring, TV antenna and CATV wiring, and within consumer devices. Knowing the ranges of typical values would help determine what levels should be handled by consumer devices. Of course, some of this can be computed theoretically, but there's no substitute for empirical evidence to bolster the theory. And theoretical calculations have to rely on assumed field strengths in a "typical" amateur station—yet another area where empirical data is scanty—and needed!

Among the *QEX* readership, there are bound to be a few who have delved into these subjects. Won't you contact us? We need your help.

This Month in *QEX*

DSP filters are a mystery to many, but Frank Morrison, KB1FZ, does his part to dispel some of the mystery and provide a straightforward approach to designing one class of DSP filters in, "The Magic of Digital Filters."

Many projects require a readout of the voltage of current at some point in the circuit—even if the readout is labeled "S-units" or whatever. In a day when meters are becoming hard to find and expensive to boot, it behooves the project designer to explore other alternatives for visual display. Sam Ulbing, N4UAU, did so, and the result is "Making Meters With Chips."

In his "Digital Communications" column this month, Harold Price shows why PACSAT access doesn't require arrays of steerable antennas and discusses how TCP/IP may fit into the future of packet radio. He also takes a quick look at HAL's new Clover system. Zack Lau's "RF" column takes a fresh look at the newly controversial subject of the match presented by an amplifier to the load and discusses some of the issues involved in making cavity preamps.—*KE3Z, email: jbloom@arrl.org (Internet)*

The Magic of Digital Filtering

By Frank P. Morrison, KB1FZ
81 Old Garrison Rd
Sudbury, MA 01776

Most of you by now have at least heard, if not read about, the use of digital signal processing (DSP) in the newer transceivers and the wonderful things that it can do with respect to signal enhancement and the elimination of adjacent interfering signals, particularly in the CW mode. But how many of us have any understanding of what DSP really is and how it works? It may seem like magic; this discussion will try to look behind the legerdemain and answer some of the questions which you may have and give you the tools to design your own filters.

You may be familiar with various analog filters, such as Butterworth, Chebychev, and elliptical types, which are implemented using suitable combinations of inductance and capacitance. What is a digital filter, and how is it designed and implemented? The design of digital filters can be very esoteric, and there is a vast literature on the subject which can involve some very high-powered mathematics. Only one reference [1] is given below; it contains for those interested a very complete set of references to what has been published in the field. What we will try to do here is to present some of the fundamental concepts with as little math as possible. It will require *some* math, but bear with us.

Digital filters fall into two major categories, infinite impulse response (IIR) filters and finite impulse response (FIR) filters. The fundamental difference between the two types is that the output of an IIR filter is a function of both a number (N) of successive past input samples and a number (M) of successive past output samples, giving rise to an impulse response sequence of infinite length (which may have to be truncated at some point depending on the method of implementation), whereas the output of a FIR filter is a function only of a number (N) of successive past input samples, which limits the impulse response sequence length of N samples. Fine, but what is impulse response? It is the response of a filter to an impulse, or spike, of voltage at its input. In the digital world, such a spike becomes a single nonzero sample in a sequence of a large number of zero samples. The impulse response of a digital filter is a sequence of N samples in time, each separated by the sampling interval (the reciprocal of the sampling frequency). IIR filters are usually designed by first designing the equivalent analog filter, (Butterworth, Chebychev, or elliptical) and then mathematically transforming the analog design into a digital design with the same frequency response characteristics. These filters can have excellent amplitude properties, but the phase can be far from linear, particularly near cut-off frequencies. The design procedure is also mathematically complex. The design process for FIR filters, on the other hand, can be made relatively straightforward. FIR filters can be made to have a linear phase across the pass band, a desirable feature. We will therefore restrict our discussion to FIR filters.

The impulse response sequence of a FIR filter with linear phase has a fixed delay in the time domain. The

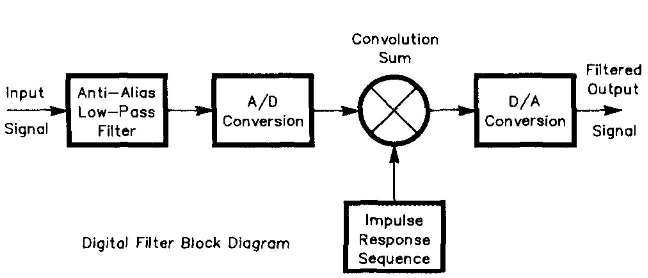


Fig 1

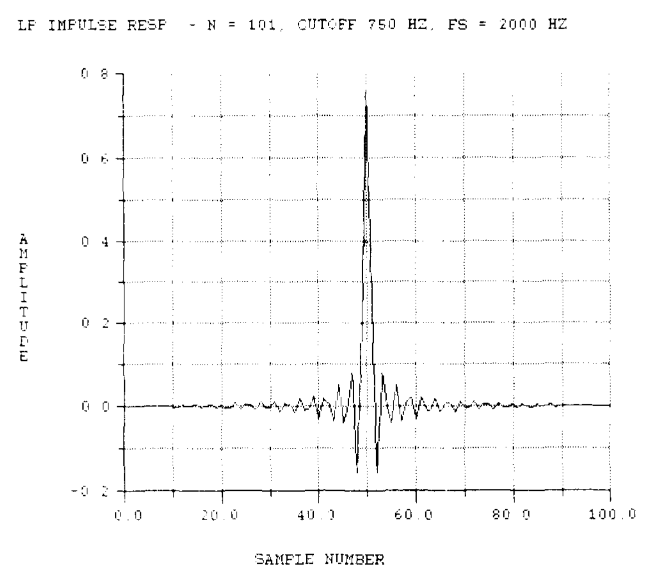


Fig 2

impulse response can be regarded as a series of N weights which, when successively multiplied in a certain way by N samples at a time of a digitized input signal, results in a filtered sequence from which the analog signal can be recovered at the output of a D/A converter. Fig 1 shows a block diagram of a digital filter implementation, and Fig 2 shows a typical impulse response sequence. If we specify a time delay of $(N-1)/2f_S$, where f_S is the sampling frequency in Hz, the output spike appears at the center of the response as seen in Fig 2. The multiplication process indicated in Fig 1 is actually a sum of N products. If $S_{out}(n)$ is the filter output signal sequence, $S_{in}(n)$ is the input signal sequence, and h(k), $0 \le k \le N-1$, is the impulse response (weighting) sequence, then:

$$S_{out}(n) = \sum_{k=0}^{N-1} S_{in}(n+k)h(N-k-1) \qquad \text{(Eq 1)}$$

The mathematical process defined by Eq 1 is called a convolution sum. Notice that the sign of k in the index of h in Eq 1 implies a reversal in time; this is not of concern here since the sequence h(k) is symmetrical. Thus once the filter has been designed, ie, the impulse response (weighting) sequence has been defined, and the signal to be filtered has been digitized at an appropriate rate (more on this later), one must merely implement Eq 1 to produce the desired filtered output sequence which is then passed to the D/A converter. The

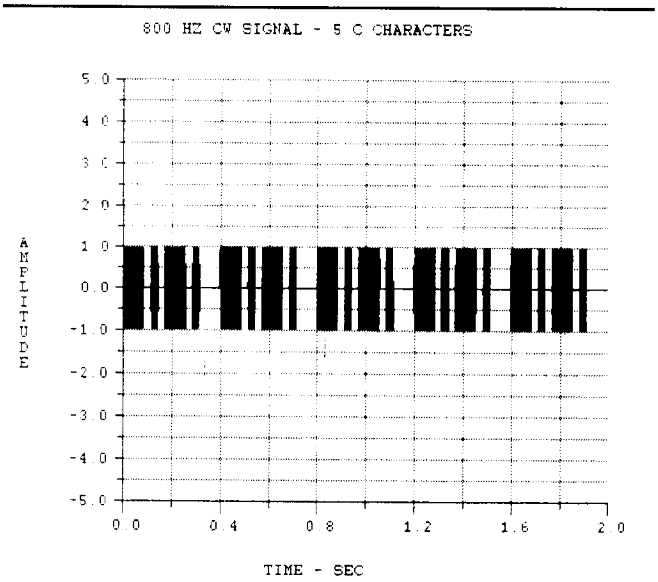


Fig 3

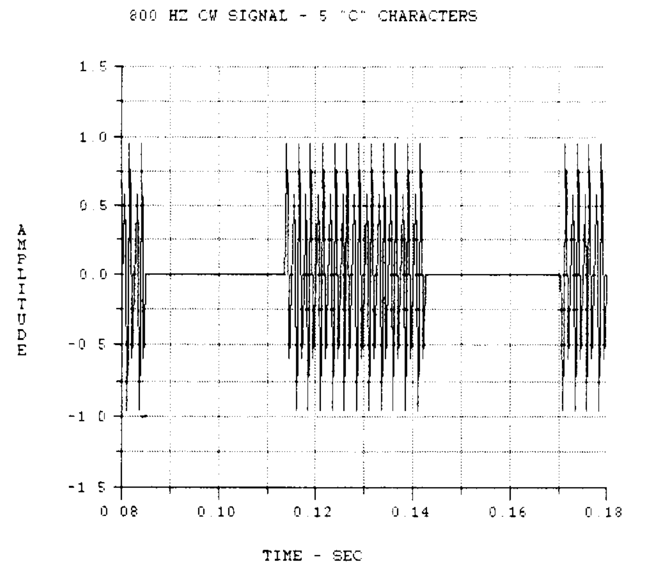


Fig 4

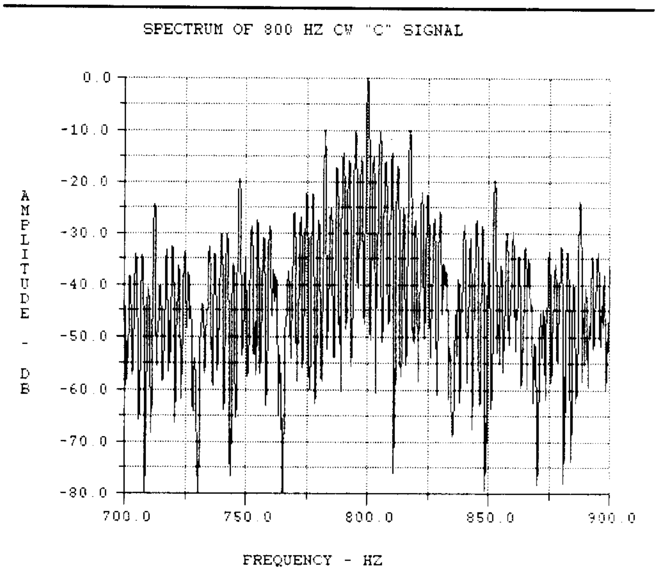


Fig 5

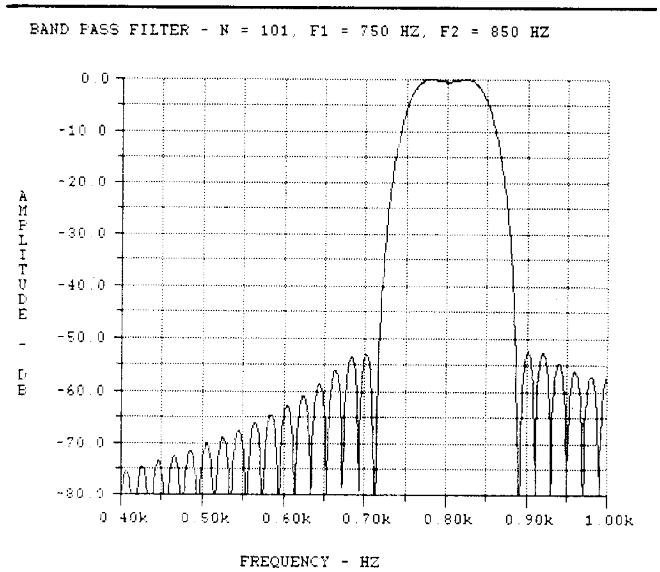


Fig 6

input signal, N samples at a time, is multiplied by the N weights and the N products are summed to produce one output sample.

What about the frequency response of the filter? In order to verify that a given impulse response sequence will produce the desired frequency pass and stop bands, the impulse response sequence must be passed through a Digital Fourier Transform (DFT), of which many of you have heard. You may have read of the Fast Fourier Transform (FFT); it is merely a specific algorithmic implementation of the DFT which makes its calculation very fast and suitable for real-time applications. The most common FFT is limited by design to input and output sequences whose lengths are a power of 2; the DFT will handle input and output sequences of arbitrary length. The DFT takes as input a sequence of samples in the time domain and transforms them into a sequence of samples in the frequency domain, which represent the frequency spectrum of the time signal. The DFT is given by:

$$F(k) = \sum_{n=0}^{N-1} S(n)e^{-j2\pi nk/N} \qquad \text{Eq 2}$$

where

$S(n)$ is the input time domain sequence, $0 \le n \le N-1$
$F(k)$ is the output frequency domain sequence, $0 \le k \le N-1$

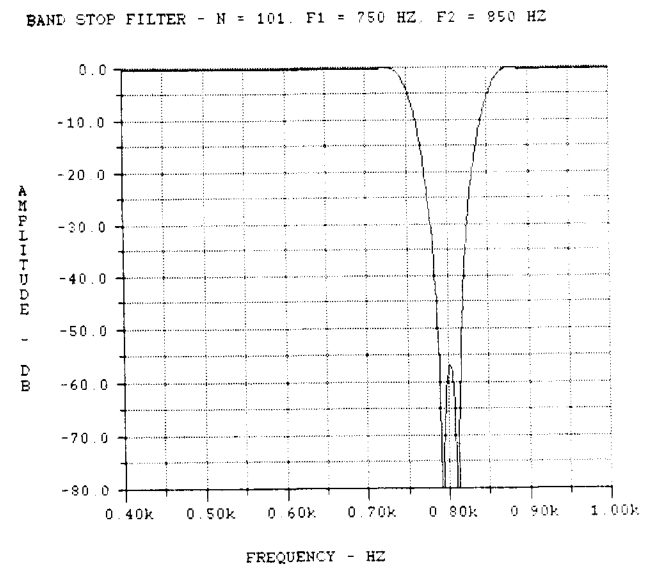


Fig 7

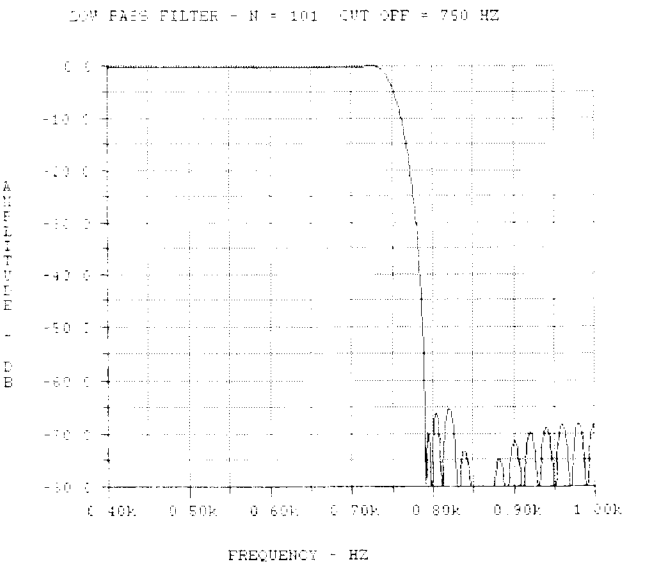


Fig 8

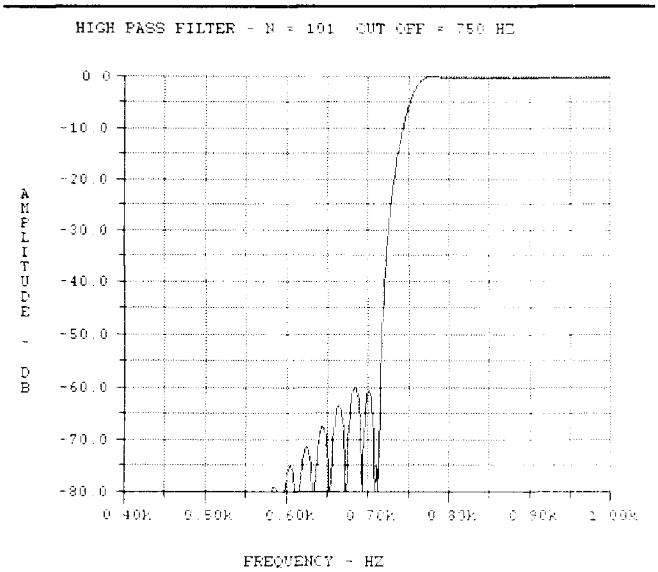


Fig 9

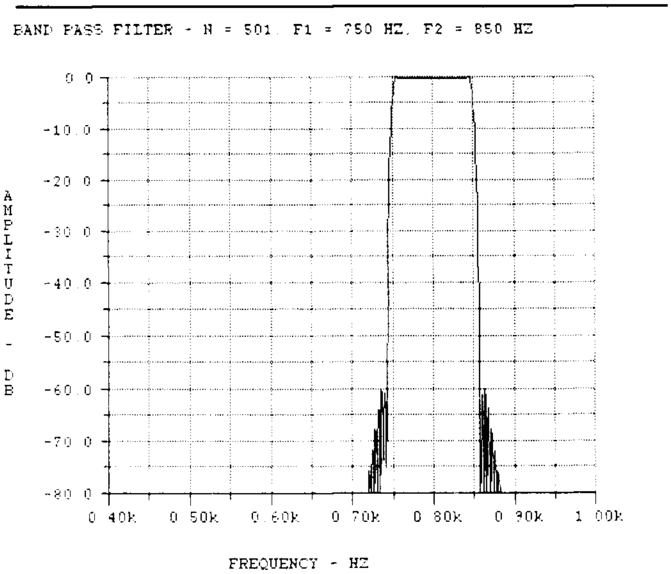


Fig 10

j = square root of −1 (imaginary)
e = base of Naperian logarithms = 2.71828
π = 3.14159

Don't be upset by the appearance of the imaginary quantity j; what it means is that the output will be a sequence of complex numbers of the form a + jb. Remember that a complex number can be thought of as representing a vector, which has both a length (amplitude) and direction (phase). The magnitude of the complex samples (square root of $a^2 + b^2$) represents the amplitude response of the filter, and the angle, given by the inverse tangent of the ratio of the imaginary part to the real part, is the filter output phase sequence (linear with frequency for the FIR filter).

If the sampling interval in the input time sequence is $\delta t = 1/f_S$ seconds, then the sampling interval in the output frequency sequence (frequency resolution) is $\delta f = 1/N\delta t$ Hz. A very important thing to remember about sampled data is the fact of life called aliasing; a sampled data stream cannot reproduce frequencies greater than 1/2 the sampling frequency. This is the reason for the anti-aliasing filter shown in Fig 1, which must be a low pass analog filter with a cut-off frequency (bandwidth) equal to or less than 1/2 the intended digital sampling frequency. Otherwise, any energy contained in the input signal at frequencies lying above 1/2 the sampling frequency will be folded down, or aliased, into the band 0-$f_S/2$, thus distorting the desired signal. This effect will be seen if the

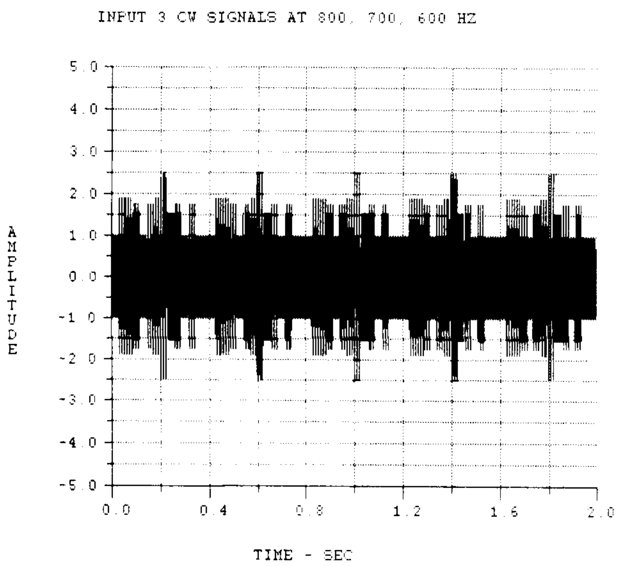


Fig 11

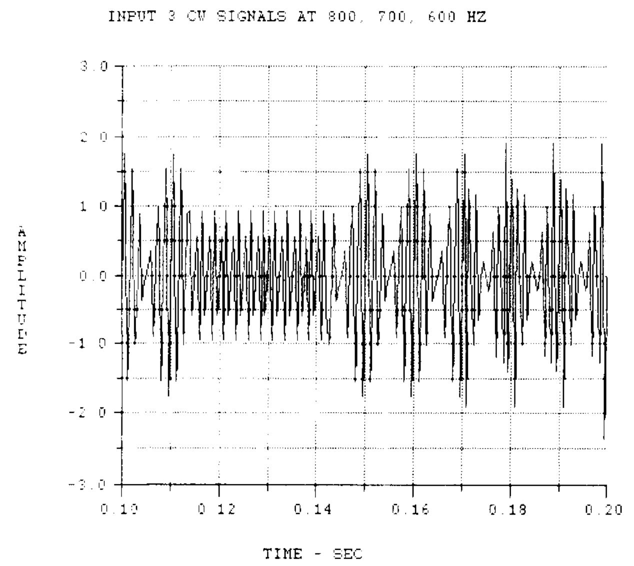


Fig 12

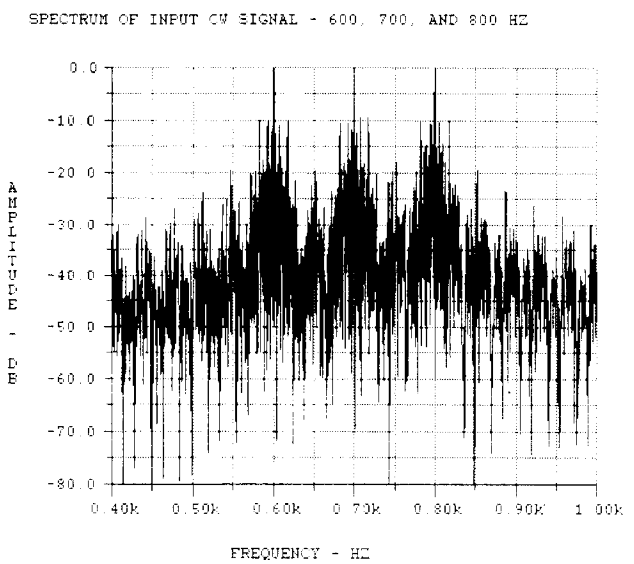


Fig 13

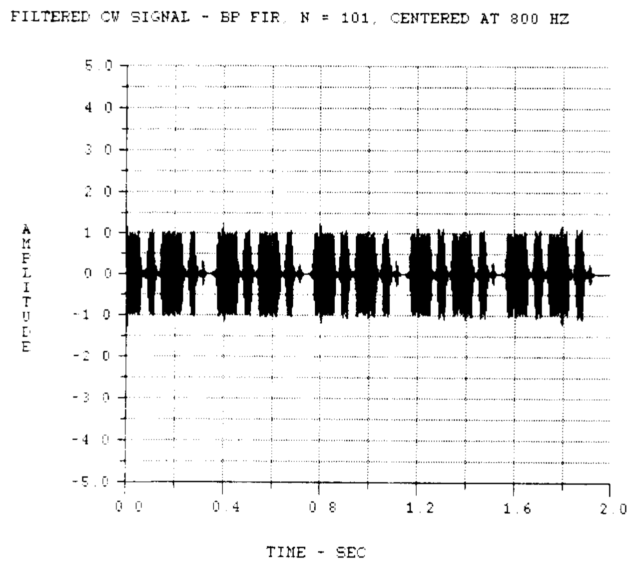


Fig 14

output of the DFT is plotted over the whole range $0 \le n \le N-1$; the response between $f_S/2$ and f_S will be a mirror image of the response from dc to $f_S/2$.

Designing A Filter

Before a filter can be designed, the spectral characteristics of the signal to be filtered must be understood. Let us assume that we wish to filter a CW signal in the audio frequency range of 400 to 1000 Hz. Fig 3 illustrates a 5-character CW signal (the character "C") at an audio frequency of 800 Hz. The equivalent word rate is 30 wpm. Fig 4 is a blow-up of Fig 3, to show some detail. Some license has been taken here in that the rise and fall times of the modulation envelope are instantaneous, ie, an ideal transmitter and receiver have been assumed along with no propagation effects. For present purposes, this is not considered a fatal defect. The spectrum of the signal is shown in Fig 5. We have often heard that the spectral width of a CW signal is about 100 Hz; this plot appears to bear this out, in that outside of this band (50 Hz either side of 800 Hz) the spectral components are down by 20 dB or more, except for the spikes at 747.5 and 852.5 Hz. This set of spikes, separated by 35 Hz, represent the length of a dot or space of 1/35 = 0.0286 second (note in Fig 3 that one character occupies 0.4 sec, and that there are the equivalent of 14 dots [spaces] in this time; 0.4/14 = 0.0286). It will therefore be assumed that for the purpose of designing either bandpass or band reject filters, a 100-Hz bandwidth will be adequate to reproduce the desired signal.

We will create one filter of each type, and show the frequency characteristic of each. The bandpass and bandstop filters will be 100-Hz wide, with cutoff frequencies of 750 and 850 Hz. The low- and high-pass filters will have a cutoff at 750 Hz. The impulse response length for each filter will be chosen to be 101 samples, which should give acceptable skirt rolloff. To avoid aliasing, and to provide some leeway between the upper cutoff and the folding frequency, a sampling rate of 2000 Hz is selected (this implies a low-pass analog anti-aliasing filter with a bandwidth of 1000 Hz in front of the A/D converter). To generate the desired impulse response sequences, the equations given in Appendix A are used. The resulting frequency responses, obtained by taking the DFT (Eq 2) of each impulse response sequence, are shown in Figs 6 through 9. To illustrate the effect of a longer impulse response sequence, Fig 10 shows the spectrum of a 501 sample band-pass impulse response. Note the considerable difference in the slopes of the filter skirts. If very sharp filters are needed, then the longer sequences can be used.

Filter Performance With a Composite CW Signal

For purposes of illustration, let's assume that we have three 30-wpm CW signals, at 800, 700, and 600 Hz. The 800-Hz signal is a series of the character "C", the 700-Hz signal a series of the character "Z", and the 600-Hz signal (transmitted by some lid) a series of the noncharacter dot-dot-dash-dash. The three signals are asynchronous in that the character boundaries are not aligned in time. Fig 11 shows how this signal appears in time; it doesn't look like one could copy any intelligence out of that mess. Fig 12 shows a blow-up in time of the signal. Fig 13 shows the spectrum of the composite signal; each signal shows essentially the same sideband structure seen in Fig 5. Let's put this signal through the band-pass filter of Fig 6, to see if the "C" signal can be recovered. The result is shown in Fig 14. Lo and behold, we have a copyable "C" signal, although it does appear a little ragged. Fig 15 shows a time blow-up; the rise and fall times of the envelope have been degraded, as might be expected, but it should be readable. Fig 16 shows the spectrum of the filtered "C" signal. Note the residual 600-Hz signal, down 70 dB. The 700-Hz signal is down 40 dB, but

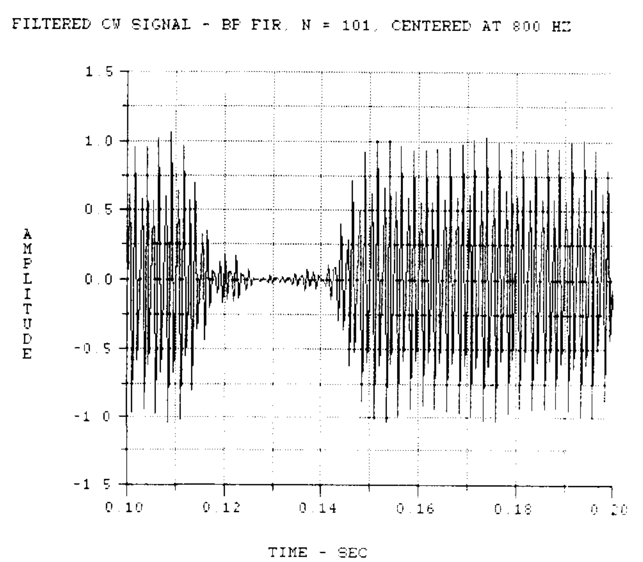


Fig 15

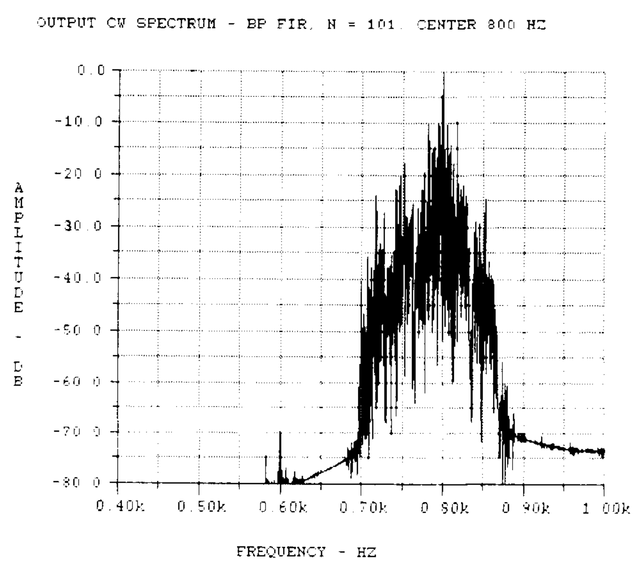


Fig 16

its upper sidebands are present and are contaminating the desired signal. This fact and the truncation of the sidebands of the "C" signal by the filtering process cause the degradation in pulse shape noted above.

The "C" signal at 800 Hz could also have been recovered using the high-pass filter in Fig 9. Actually, this filter would do a slightly better job because the "C" signal upper sidebands would not be attenuated thereby. The "Z" signal, lying between the other two, could be recovered by tuning the receiver until its frequency became 800 Hz and then using the band-pass filter. The lid's signal could be recovered by using the low-pass filter after appropriate retuning.

Summary

I have attempted to present some of the fundamentals of digital FIR filter theory and design, stripped of the mathematical rigor and formalism found in most of the literature. An example has shown that with these filters it is possible to separate closely adjacent CW signals with some success. The tools for generating FIR impulse response sequences have been given in the form of the equations in Appendix A. Again, to avoid too much mathematical detail, the derivation of these equations has not been included; if anyone is interested in that derivation, I would be glad to provide it. I hope that you have found this discussion interesting and informative.

Reference

1: *Digital Signal Processing*, by Alan V. Oppenheim and Ronald W. Shafer, Prentice-Hall 1975, Chapter 5, paras 5.4 through 5.61.

Appendix A

This appendix presents the equations from which can be calculated the impulse response (weight) sequences for FIR filters of the low-pass, band-pass, band-rejection and high-pass types.

The following definitions apply:

h(n)	–	sequence of impulse response (weight) samples
N	–	number of samples (weights) in the impulse response sequence (N is normally an odd number)
n	–	sample index, $0 \le n \le N-1$
β_1	–	first transition band amplitude = 0.5886
β_2	–	second transition band amplitude = 0.1065
f_1	–	lower cut-off frequency, Hz
f_2	–	upper cut-off frequency, Hz
f_s	–	data sampling frequency, Hz

Low-pass impulse response (weighting) sequence

$$h(n)=\frac{\sin\left[\frac{\pi}{N}(2M-1)\left(n-\frac{(N-1)}{2}\right)\right]}{\sin\left[\frac{\pi}{N}\left(n-\frac{(N-1)}{2}\right)\right]}$$

$$+2\beta_1\cos\left[\frac{2\pi}{N}M\left(n-\frac{(N-1)}{2}\right)\right]+2\beta_2\cos\left[\frac{2\pi}{N}(M+1)\left(n-\frac{(N-1)}{2}\right)\right]$$

where M=integer value of $(N-1)f_2/f_s+1$

Band-pass impulse response (weighting) sequence

$$h(n)=2\cos\left[\frac{\pi}{N}(L+M)\left(n-\frac{(N-1)}{2}\right)\right]\frac{\sin\left[\frac{\pi}{N}(L-M-1)\left(n-\frac{(N-1)}{2}\right)\right]}{\sin\left[\frac{\pi}{N}\left(n-\frac{(N-1)}{2}\right)\right]}$$

$$+2\beta_1\left\{\cos\left[\frac{2\pi}{N}L\left(n-\frac{(N-1)}{2}\right)\right]+\cos\left[\frac{2\pi}{N}M\left(n-\frac{(N-1)}{2}\right)\right]\right\}$$

$$+2\beta_2\left\{\cos\left[\frac{2\pi}{N}(L+1)\left(n-\frac{(N-1)}{2}\right)\right]+\cos\left[\frac{2\pi}{N}(M-1)\left(n-\frac{(N-1)}{2}\right)\right]\right\}$$

where M=integer part of $(N-1)f_1/f_s+1$
L=integer part of $(N-1)f_2/f_s+1$

Band-rejection impulse response (weighting) sequence

$$h(n)=\frac{\sin\left[\frac{\pi}{N}(2M-1)\left(n-\frac{(N-1)}{2}\right)\right]}{\sin\left[\frac{\pi}{N}\left(n-\frac{(N-1)}{2}\right)\right]}$$

$$+2\cos\left[\frac{\pi}{2}\left(\frac{(N-1)}{2}+L+1\right)\left(n-\frac{(N-1)}{2}\right)\right]\frac{\sin\left[\frac{\pi}{N}\left(\frac{(N-1)}{2}-L\right)\left(n-\frac{(N-1)}{2}\right)\right]}{\sin\left[\frac{\pi}{N}\left(n-\frac{(N-1)}{2}\right)\right]}$$

$$+2\beta_1\left\{\cos\left[\frac{2\pi}{N}L\left(n-\frac{(N-1)}{2}\right)\right]+\cos\left[\frac{2\pi}{N}M\left(n-\frac{(N-1)}{2}\right)\right]\right\}$$

$$+2\beta_2\left\{\cos\left[\frac{2\pi}{N}(L-1)\left(n-\frac{(N-1)}{2}\right)\right]+\cos\left[\frac{2\pi}{N}(M+1)\left(n-\frac{(N-1)}{2}\right)\right]\right\}$$

where M=integer part of $(N-1)f_1/f_s+1$
L=integer part of $(N-1)f_2/f_s+1$

High-pass impulse response (weighting) sequence

$$h(n)=2\cos\left[\frac{\pi}{N}\left(\frac{(N-1)}{2}+M+1\right)\left(n-\frac{(N-1)}{2}\right)\right]\frac{\sin\left[\frac{\pi}{N}\left(\frac{(N-1)}{2}-M\right)\left(n-\frac{(N-1)}{2}\right)\right]}{\sin\left[\frac{\pi}{N}\left(n-\frac{(N-1)}{2}\right)\right]}$$

$$+2\beta_1\cos\left[\frac{2\pi}{N}M\left(n-\frac{(N-1)}{2}\right)\right]+2\beta_2\cos\left[\frac{2\pi}{N}(M-1)\left(n-\frac{(N-1)}{2}\right)\right]$$

where M=integer part of $(N-1)f_2/f_s+1$

Note: In all the above equations, the value of sin(Ax)/sin(Bx) is A/B for x=0.

Making Meters With Chips

By Sam Ulbing, N4UAU
5305 NW 57th Lane
Gainesville, FL 32606

Have you ever built a project that needed a meter? Did you spend a lot of time and money to meet that need? Maybe you should consider using ICs to replace meters—it's cheaper and easier. When you see what these chips can do, I think you'll agree with me: Building a project with an analog meter is like building a QRP rig with vacuum tubes!

I'll discuss two chips that are ideal for making meters. Each chip is an analog-to-digital (A/D) converter designed for direct visual display. As you'll see, the chips are different from one another. But what they have in common is that they are:

- low cost—much less than analog meters
- highly accurate
- easily adaptable to measure different voltages or currents
- easy to read
- able to include alarms or special indicators
- good looking!

The ICL7106 All-Purpose Voltmeter

The ICL7106 is a voltmeter chip that drives a 3½ digit display, meaning it will display values up to ±1999, providing a resolution of better than 0.1%. With this kind of available resolution, the accuracy of the readout depends on the accuracy of the external resistors you use to build the meter. Fig 1 shows the basic components of the IC.

Fig 2 shows the circuit of an ICL7106 voltmeter. The chip requires +5 and –5 volts. In this circuit, the +5 Vdc is provided by a 78L05 regulator IC, allowing a supply voltage of from 7 to 20 volts. –5 Vdc is developed using an ICL7660 voltage converter. The ICL7106 A/D sampling rate is set by R1 and C1 to approximately 48 kHz. This rate results in about three readings per second and rejects 60-Hz noise. C2 is a reference capacitor which, in conjunction with C4, C5 and R2, allows the ICL7106 to calculate the input voltage. The reference voltage is set by R3 and R4. Ideally, C5 should be a polypropylene capacitor because of the low dielectric absorption of such capacitors. I've used ceramic and mylar capacitors here, though, with good results.

The meter in Fig 2 reads ±20 Vdc, but you can select the range of your meter circuit by choosing the appropriate value for R5. The chip accepts a maximum input voltage of 200 mV, so select R5 to make the maximum input voltage correspond to 200 mV. Since the readout at the maximum input voltage will be 1999 (with a decimal point displayed at the appropriate spot), you probably will want to set the range to one of the voltages in Table 1, which shows the values of R5 needed.

The digital section of the circuit in Fig 2 connects the chip to an LCD display module. The LCD display is activated when an ac voltage, provided by the ICL7106, is applied to a given segment of the display, relative to the display backplane. The ICL7106 applies a continuous square-wave signal to the backplane and drives each of the display segments either in phase with this backplane signal (segment is off) or out of phase with the backplane

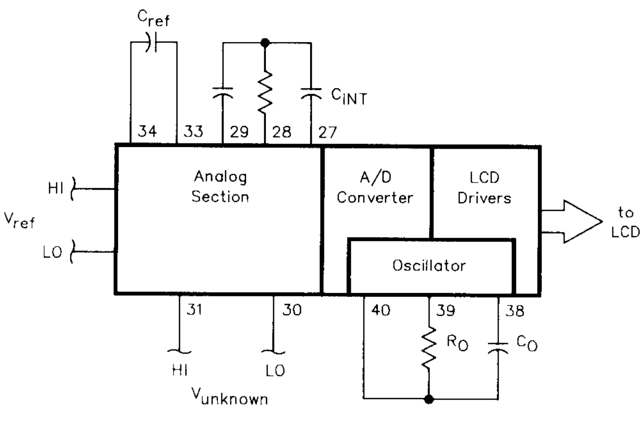


Fig 1—Block diagram of the ICL7106 voltmeter IC.

Table 1—Values of R5 in the ICL7106 circuit

	R5 values	
Full-scale V	*Exact*	*1% value*
200.0 mV	none	none
2.000 V	111.1 kΩ	110 kΩ
20.00 V	10.10 kΩ	10.0 kΩ
200.0 V	1.001 kΩ	1.00 kΩ

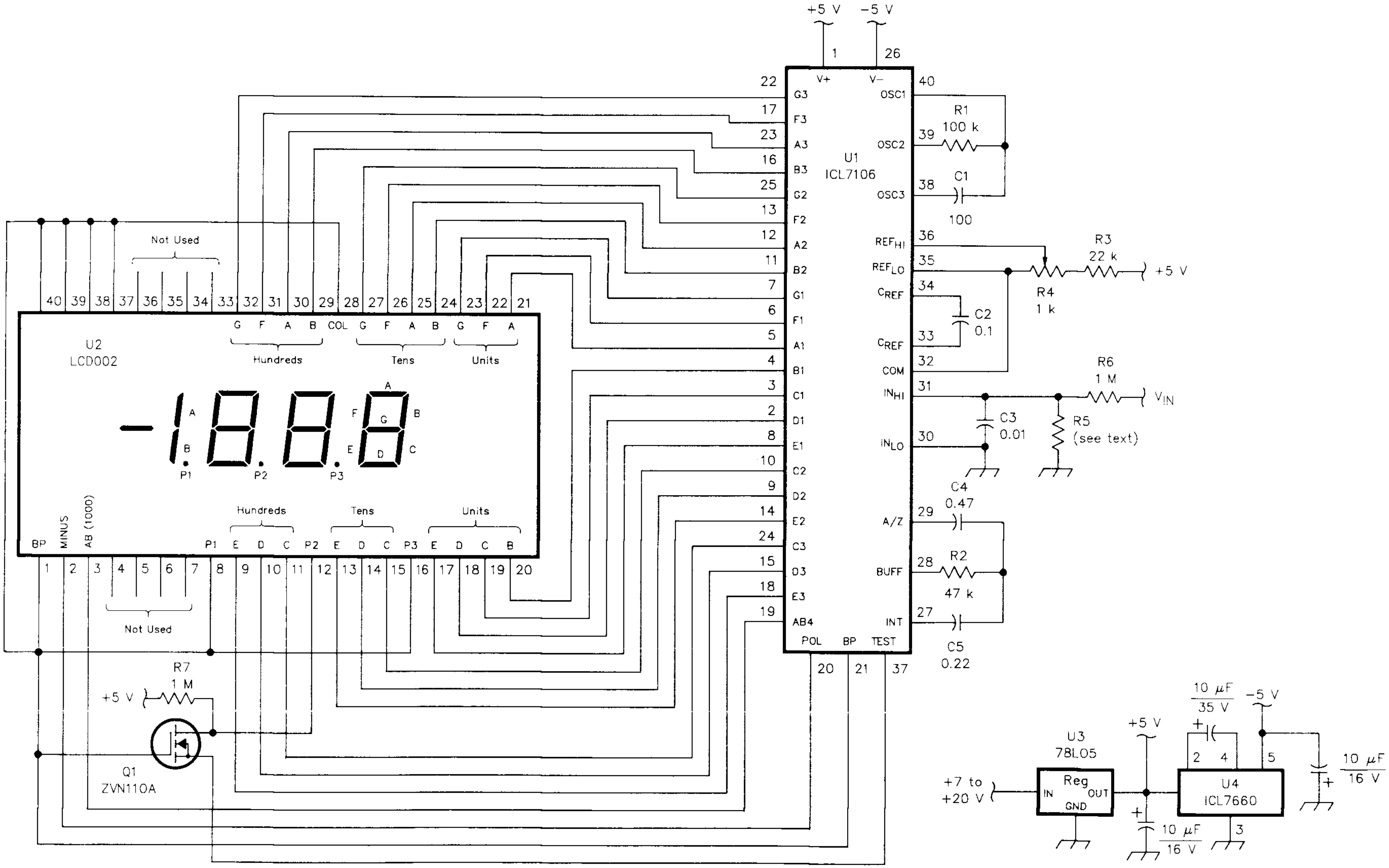
Using the nearest standard 1% resistor value introduces a small error. An exact resistor or trimmer pot can be used if necessary.

Fig 2—Schematic diagram of a general-purpose voltmeter using the ICL7106. The range of the voltmeter is set by selecting the value of R5 (see text).

C5—0.22 µF capacitor, polypropylene, ceramic or mylar (see text).
Q1—ZVN110A or equivalent N-channel MOSFET.

U1—ICL7106 LCD A/D converter IC.
U2—LCD002 LCD display module
U3—78L05 5-volt voltage regulator IC.
U4—ICL7660 voltage converter IC.

All parts are available from Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel: 800 344-4539.



(segment is on).

For a given range, you may want to illuminate one of the decimal points on the display module shown in Fig 2. The unused decimal points are wired to the backplane to ensure they stay off, while the needed decimal point is wired to the drain of Q1. Fig 2 shows this arrangement for a 20-volt display. If using an LCD display module other than the one referenced in Fig 2, check the pin connections—they may differ.

That's it. You now have an accurate digital voltmeter for your project. To convert it to an ammeter, remove R5 and connect V_{in} to a shunt resistance that has one end connected to ground. The shunt resistance should be of a value that produces 200 mV across the shunt at full-scale current. Table 2 shows typical shunt resistances.

Table 2—Shunt resistances for the ICL7106 ammeter

Full-scale I	*Shunt resistance*
200.0 μA	1 kΩ
2.000 mA	100 Ω
20.00 mA	10 Ω
200.0 mA	1 Ω
2.000 A	0.1 Ω

The LM3914

While the ICL7106 provides a numeric display this often isn't the most effective visual aid. Sometimes, the "moving pointer" nature of an analog meter provides the desired visual indication. A digital display that provides this kind of animation, as well as actual measurement capability, is produced by the LM3914. Along with this difference, LM3914-based displays are less expensive, smaller and simpler than ICL7106-based displays.

The LM3914, an 18-pin chip from National Semiconductor, drives 10 LEDs in a bar-graph display. Two reference voltages, the high reference and the low reference, set the range of the voltage measured by the chip.

Fig 3 shows the basic LM3914 voltmeter circuit. The supply voltage applied to pin 3 of the LM3914 can be between 3 and 20 Vdc, although it should be at least 1.5 volts higher than the voltage to be measured. In this basic circuit the low reference (pin 4) is at 0 volts, while the high reference (pin 6) is set by the reference voltage supplied by the chip on pin 7.

While an external reference voltage can be used, use of the internal reference source is usually easier. A 1.25-volt reference is supplied between pins 7 and 8. Placing a resistance (1 kΩ in our circuits) across these terminals causes a current to flow through the 1-kΩ resistance: I=1.25/1000=1.25 mA. This current also must flow through R2, developing a voltage across that resistor that is proportional to the resistance: V=I×R2. This voltage is added to the 1.25 V across R1 to form the high reference voltage, setting the upper limit of the measurement range.

Actually, an exact measurement requires a bit of fine tuning of the R2 value. This is because some of the current from pin 7 flows to pin 6 and through the internal voltage divider, causing the current through R2 to be slightly less than theoretical. The small bias current from pin 8, along with chip-to-chip variations, are also factors. I do a rough calculation of R2 and then use a small pot to trim the reference voltage to the exact value.

In this circuit, as the measured voltage increases more LEDs light up. This provides a bar-graph style display. You also may elect to have a "moving dot" display. Which display you get is controlled by the voltage at pin 9. If this pin is left floating, you get the moving-dot display. If the pin is connected to the supply voltage, the bar-graph display results.

An extension of the basic meter-display function of the LM3914 is shown in Fig 4. This circuit provides an

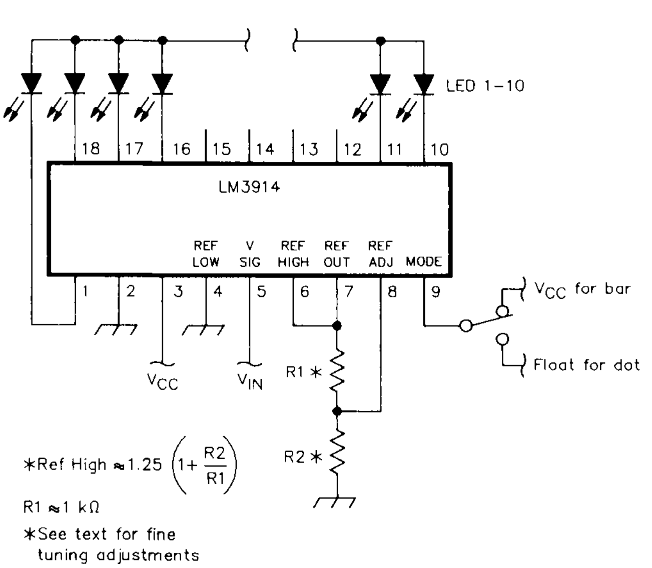


Fig 3—The basic LM3914 voltmeter circuit.

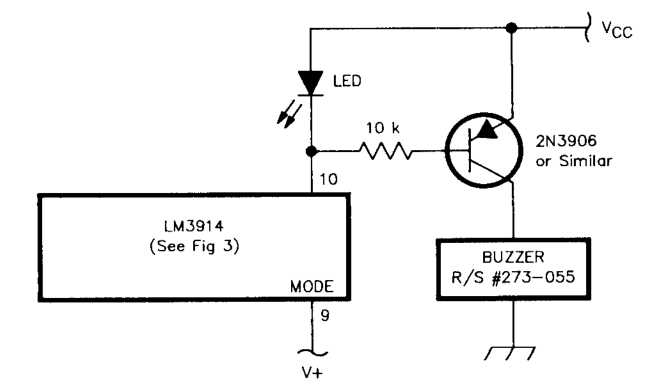


Fig 4—Use of the LM3914 as an over-voltage alarm.

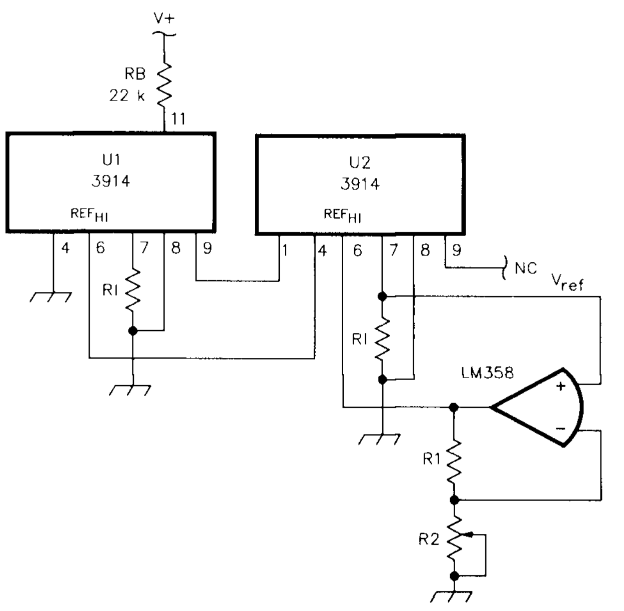


Fig 5—Cascading of LM3914s to achieve a higher-resolution display. Use as a moving-dot display is shown. For use as a bar-graph display, remove RB, disconnect U1 pin 9 from U2 pin 1, and connect both U1-9 and U2-9 to Vcc. Select RI to adjust LED brightness (see text). Reference voltages are set as follows:

$$REF_{HI-U2}=V_{ref}(1+R1/R2)$$

$$REF_{HI-U1}=REF_{HI-U2}/2$$

For example, for a 10-volt meter, R1=100 kΩ, R2=14.3 kΩ.

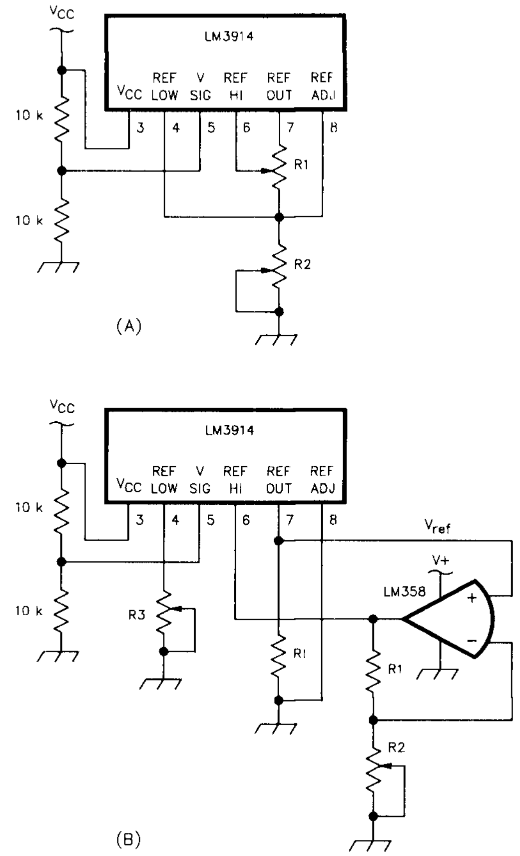


Fig 6—Two circuits to measure voltage on an expanded scale. Use the circuit at A when the difference between the high and low reference voltages is less than 1.25 volts. Adjust R1 to set the range, the difference between the voltages on pis 6 and 4, and use R2 to set the absolute voltage on pin 4. The circuit at B is used when the reference difference is more than 1.25 volts. The equations for the resistances are:

$$V_{REFHI}=V_{ref}(1+R1/R2)$$

$$V_{REFLO}=V_{REFHI}[R3/(R3+10000)]$$

Select RI to set LED brightness (see text).

over-voltage alarm by sensing when the uppermost LED comes on. The alarm circuit shown drives a piezoelectric buzzer. I put this circuit on a battery in my shack once, and when the alarm went off my unsuspecting wife ran from the house screaming, "fire!" Try getting that kind of reaction from an analog meter!

Cascading LM3914s for Higher Resolution

The resolution of a 10-segment, 10-volt display is only about 1 volt. While this is useful in many circumstances, greater resolution is often needed. It can be obtained by cascading LM3914 chips to provide a display using more LEDs. Displays of up to 100 LEDs can be obtained in this manner. Fig 5 shows a 10-volt meter using two LM3914s, for 20 LEDs. With twice the resolution of the 10-LED meter, it still can be built for only about $6.

Several changes are needed to the basic circuit of Fig 3 in order to cascade the ICs. Each chip needs its own low and high reference voltages. My approach to providing these voltages is shown in Fig 5. The on-board reference voltage is fed to an op amp, and the op-amp gain is set to give the desired high reference voltage. The other reference voltages are developed by a voltage divider. Since the high and low reference pins of the LM3914 are at opposite ends of a resistive voltage divider, successive reference voltages can be obtained by simply cascading the references: Connect the low reference of a chip to the high reference of the next lower chip in the cascade.

These changes are effective for the bar-graph display mode. For the moving-dot mode, a few other changes are needed: Pin 11 of each chip except the highest one should be connected through a 22-kΩ resistor, and each chip's pin 9 should be tied to pin 1 of the next highest chip. These changes force the highest LED of a chip to turn off when the lowest LED of the next higher chip lights.

An R1 value of 1 kΩ isn't totally arbitrary. The current drawn from pin 7 adjusts the current through the LEDs, controlling their brightness. This is why no current-limiting resistors are used with the LEDs. The current through each LED is about 10 times the current drawn from pin 7.

Expanded-Scale LM3914 Meters

So far, all of the LM3914 circuits have measured from 0 to some higher voltage. In applications where only a narrow voltage range is of interest, the LM3914 really shines, since it becomes a much higher-resolution device. I've used this design for several years to keep track of my shack's 12-volt battery voltage. By setting the low reference to 11.8 volts and the high reference to 13.8 volts, the meter will monitor the voltage with 0.2-volt resolution.

Fig 6 shows two variations on the expanded-scale meter. Only a few changes from the basic circuit are needed. In these examples, the meter is measuring its own supply voltage. To keep the measured voltage well below the supply voltage, a resistive divider is used. In Fig 8A, the difference between the low and high reference voltages is less than 1.25 volts, which can be obtained directly from the reference circuit. While this technique can be used for larger reference-voltage differences by tapping down on R2, interactions with the internal voltage divider begin to make calibration difficult. In Fig 8B, an op amp is used to avoid this problem.

The resolution of the expanded-scale meter can be as good as 1%. Not bad for a $3 meter!

Digital Communications

By Harold Price, NK6K
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The first column of a new year is usually a time for reflection on previous columns. This makes it easy for the columnist to enjoy the holidays as he or she paraphrases the past. Alas, a rehash of the only other column in this series would be an obvious ploy. The only way to allow for time with friends and family is to present the following potpourri of Things You Should Know in 1993. First, some things that are so blindingly obvious, you've forgotten them.

Improving Network Throughput

Even advanced technical types may be able to use this hint. I noticed that the problem is still prevalent while listening to the Pittsburgh area packet frequencies by ear: many stations have their *transmit keyup delay* parameter set too high. Proper attention to this parameter can give a surprising increase in throughput. One local BBS has its delay set 400 ms too long. This adds to the turnaround time, and is especially painful when uploading to the BBS, when it sends only short acknowledgment frames. During file transfers with a window size of one frame, an extra 400 ms can add 20% or more to the transfer time at 1200 baud; it doubles the transfer time at 9600 baud.

The easiest way to select the proper value is to use trial and error. Each transmitter will be different, and you need to allow for differences in the time it takes for the receiving station's squelch (if any) to open. BBS stations will want to set their delay a little longer to allow for slow users, but 400 ms is too long. If you're using the default, get out your manual and read up on it. Many TNCs use different units for the time represented by the parameter. NET/NOS users also need to pay attention to this parameter. One user's guide gives an example of the TXDELAY parameter set to 500 ms. This is too long in almost all cases.

Inexpensive Elevation Rotor for PACSATS

Here's an idea for a very cheap ($0.00) elevation rotator for a PACSAT satellite station. You don't really need one for occasional use. If you have an antenna with a 20-degree beam width in the elevation plane (most smaller Yagis will have more), you can hear 70% or more of an average pass from most locations. Fig 1 shows the daily access times at NK6K for AO-16 or LO-19, UO-22 access tunes are similar. The graph shows that most of the time is spent at low elevations, 88% at less than 30 degrees. All low-earth-orbit satellites will exhibit similar characteristics. You can borrow a 1200-baud PSK or 9600-baud FSK modem and get a taste of the action on these spacecraft without a major investment in tracking hardware. Since the software handles all the data work, you can steer the azimuth with one hand and the downlink with the other. You don't need a third hand for elevation.

There will be at least three amateur PACSAT-class spacecraft launched in 1993. KITSAT-B will be a near copy of KO-23 (KITSAT-A) and should be sending pictures as good as the one on the cover of the October 1992 *QEX*. ITAMSAT and UNAMSAT, from Italy and Mexico, are AMSAT-NA style microsats. With these launches, amateurs will have 70 Mbytes of data storage in orbit, which is more than most of us had on the ground when PACSATs were first proposed in 1981!

TCP in '93

This may be the year that you give TCP/IP a try. TCP/IP is closer to "real networking" than an unembellished AX.25 link. TCP/IP, or more properly, the Internet Protocol Suite, is a generic name for many different protocols that are used for computer-to-computer communication. If you are in Amateur Radio to give yourself a hands-on technical education, you owe it to yourself to bring TCP/IP up on your system. Unfortunately, this is not as easy as you might think.

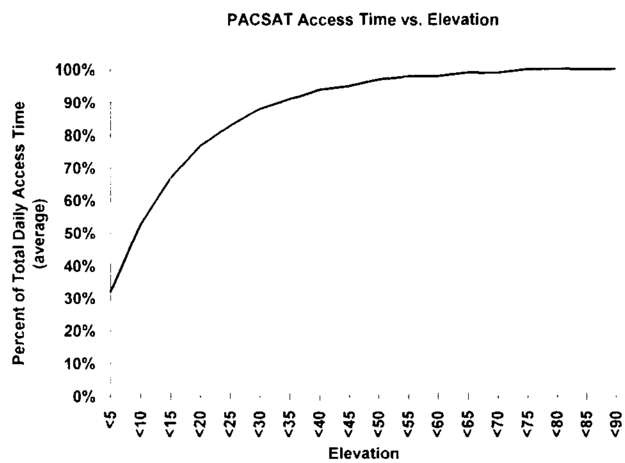

Fig 1—*Most of a PACSAT's time is spent at low elevations.*

NOS (or its predecessor NET) is the software that almost all Amateur Radio TCP/IP packages are based on. There are several tutorials, beginner's guides, help files, and FAQs (Frequently Asked Questions lists) out there to help you get started with NOS. A common thread in all of them is this concept, quoted from Mike Gallaher's FAQ: "The best place to get NOS is from someone you know who is already using it." More generally, the best way to get on the air is to have someone help you. This in itself is somewhat of a problem, since getting at someone who is on the air usually requires you to already be on the air. There are several off-the-air sources for amateur TCP/IP information. The ones I've verified recently are:

- CompuServe's HAMNET Forum, data library 9. Contains some information files and older releases. The leading edge types seem to avoid the *hoi polloi* on HAMNET. Someone needs to cross-fertilize. Network development is no fun without users to find errors, and HAMNET is fertile ground. As part of the research for this column, I placed a copy of the Mike Gallaher FAQ on DL9.
- WB6YMH BBS (310) 541-2503
- N8EMR BBS (614) 895-2553
- ChowdaNet (401) 331-0907

The best place to get TCP/IP information is via FTP off the Internet. Newcomers to TCP/IP aren't likely to have access, however. If you do, try: ucsd.edu in hamradio/packet/tcpip. Almost everything of interest will be there somewhere. Information on what is where is sparse, however. At this writing, the FAQ, a good place to start, is in: hamradio/packet/tcpip/incoming/FAQ-1292.

The best place to get a pointer to the pointers is from someone who is already on the air. There are several IP address administrators; you probably have one in your area. Bob Hoffman, the coordinator for my area, tells me that some coordinators are listed in the *ARRL Operating Manual.* While these guys may not have time to help you out themselves, they should know someone who does. Implementations are available for DOS, OS/2 PM, Windows/NT, Macintosh and several UNIX platforms. Work is ongoing, so get involved.

TCP/IP and Innovation

The use of the internet protocols in amateur packet radio is a good thing in general. Along with giving people hands-on experience that they might not be able to get elsewhere, and the mom and apple pie aspects of standard, interoperable protocols, there is a down side to the movement toward the internet suite. I attended the 1992 ARRL Computer Networking Conference in New Jersey, where there were the usual number of people describing problems and their approach to solving them. The question/comment period after each paper showed a trend that has been getting worse recently. Rather than a discussion of the problem or solution, someone in the back would yell "That's covered by RFC xxxx," and that would be that. (RFCs are what the internet world calls its standards documents.) The problem is that, while the RFC does address the general problem, the solution sometimes assumes bigger, faster systems and data rates. While a solution involving flooding on a set of interlinked 10-Mbit token rings might be viable in some arenas, it is not likely to be optimum for a 9600-bps, or even 1200-bps, RF link. There is a tendency to view the very-low-bit rate networks as degenerate cases, or simply uninteresting problems. An existing standard should not be rejected out of hand, and minor tweaks to it may suffice. But we still need to allow for solutions that are tailored to our unique situation. Between the limitations imposed by physics, Part 97, and the wallet, we amateurs don't always fit into the general case.

CLOVER

I'm excited about CLOVER because it was designed to take the real world into account[1]. CLOVER is a combination of hardware and software that implements a data transfer system optimized for use on HF. It uses modern techniques, unencumbered by holdovers from the days when "hardware" meant motors and gears. HAL's PCI-4000 PC adapter card implements CLOVER with an 8-MHz MC68EC000 CPU and a 20-MHz DSP56001 DSP engine.[2] This is serious processing power, and it is put to use to overcome the various limitations of HF, including the ionosphere as well as the audio stages of the common HF rig. CLOVER uses adaptive equalization, variable data rates, and Reed-Solomon coding to ensure some throughput in the worst of conditions, and faster data rates when conditions allow.

HAL has just started shipping the units and the first users are getting on the air. I've seen the user documentation; it is certainly comprehensive. At least two HF forwarding BBS implementors are adding CLOVER support, and HAL is interested in helping others to port software. With all that hardware on the card, plus software development to amortize, the cost of the PCI-4000 is a bit above the DSP-based TNCs, so I don't expect that everyone will have one. Still, if the card is as good in practice as it is on paper, I'd expect to see a substantial portion of HF forwarding traffic being sent by this mechanism by the end of 1993. Some of the RTTY/AMTOR big guns are already trying this new mode.

I hope to be able to report on the initial user experiences with CLOVER in my next column. As always, you are encouraged to talk back, either to the editor, or directly to my email address.

[1]Raymond C. Petit, "CLOVER-II: A Technical Overview," Proceedings of the 10th ARRL Computer Networking Conference, ARRL 1991, pp 125-129.
[2]HAL Communications Corp, Box 365, Urbana, IL 61801.

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Newington, CT 06111
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Conjugate Matching of Nonlinear Amplifiers

The November 1991 issue of *QST* had a very interesting and somewhat controversial article by Warren Bruene, W5OLY, the inventor of a popular SWR measuring circuit. The article, "RF Power Amplifiers and the Conjugate Match," presented a set of measurements that indicated that conjugate matching was not present in the amplifiers Bruene measured. Not surprisingly, this generated so much controversy that an additional article in the May 1992 *QST* was published to further explain what was happening. Let's look at conjugate matching in an amplifier so simple that we can know exactly what is happening.

Consider Fig 1, a simple class-C amplifier. V_{in} is large enough to produce a square-wave output. When Q1 is on, the drain-to-source resistance, R_{ds}, is nearly zero. When Q1 is off, R_{ds} is effectively infinite. Given that Q1 is on half the time and off half the time, what is the output impedance of the FET? (If it makes it any easier for you, you can make $R_{ds(on)}$ 0.1 Ω and make $R_{ds(off)}$ 10 kΩ.) Does the output impedance equal R_L? How can it, since R_L hasn't even been picked yet? You could look at page 24 of *Solid State Design*, which says that $R_L=V_{CC}^2/2P_O$, where P_O is the output power, but this equation is wrong for our sample circuit, since it assumes a sine wave output rather than the unfiltered square wave we actually have. You have to be careful when pulling equations out of books—you really should make sure that the underlying assumptions still hold true. Modifying the equation for our case gives: $R_L=V_{CC}^2/P_O$.

Note that this equation implies that the source resistance is zero. Otherwise, there is a maximum output power, which occurs when the load and source resistances are equal. But solving the equation for output power gives $P_O=V_{CC}^2/R_L$, which implies that the output power will increase as long as R_L is decreased. I would like to be able to publish an accurate derivation of the source impedance, but I have yet to discover one.

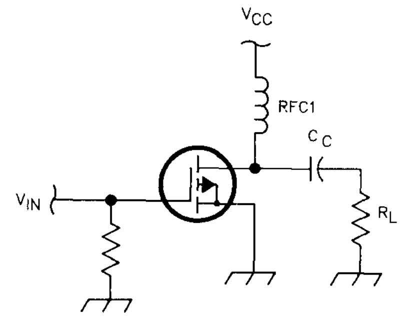

Fig 1

Now, what happens when you hook up an SWR meter, transmission line, and load to this transmitter, as shown in Fig 2? According to transmission line theory, as long as $Z_O=R_L$, any pulse sent down the transmission line will just disappear into the resistor, R_L. Practically, we know this isn't true. All you have to do is come up with a large enough pulse to destroy R_L, and you'll get a reflection! But this does explain why the SWR meter reads 1:1 even though the load and source resistances are not equal. There is no reflected power measured because there is no reflected wave to measure—all the power traveling down the coax goes into the antenna. It

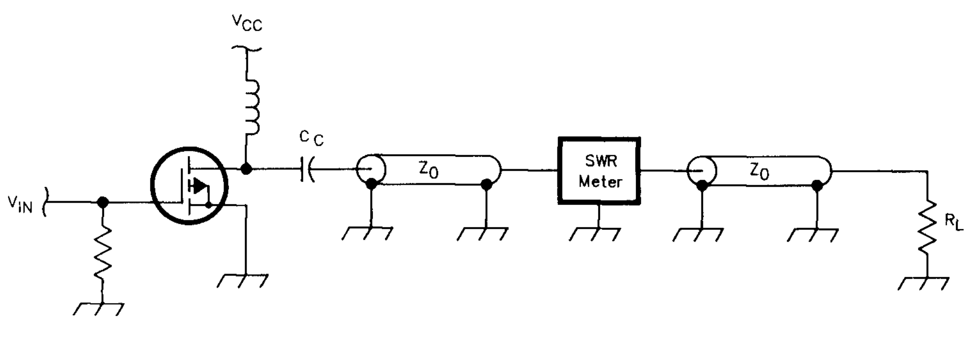

Fig 2

isn't always true that an SWR meter indicates that the load and source are conjugately matched.

Of course, this undoubtedly goes against truths heard long enough to be self-evident.

Coherent CW Experimenters Need Help

Speaking of simple transmitters, one group of people having second thoughts is the Coherent CW, or CCW, experimenter community. Why second thoughts? Because the need to build a rig "from scratch" has inhibited the exploitation of CCW. Not everyone interested in CCW necessarily wants to build their entire station, preferring to spend more time on actual experimental issues. It may make more sense for these people to modify existing transceivers, such as the Kenwood TS930 or TS450. In particular, these transceivers use a single reference oscillator, making it easy to substitute a highly accurate reference—needed for CCW—in place of the stock oscillator. Having had little success getting information from manufacturers, the two CCW experimenters listed below are interested in knowing what other commercial synthesized transceivers use a single reference, as opposed to multiple reference oscillators. Please help them if you can:

Cliff Buttschardt W6HDO
950 Pacific St
Morro Bay CA 93442

Peter Lumb G3IRM
2 Briarwood Ave
Bury St. Edmunds
Suffolk IP33 3QF
ENGLAND UK

Designing Cavity Preamplifiers

So, you want the ultimate in VHF/UHF low noise amplifiers? It's tough to compete with a properly designed cavity amplifier— except in terms of ease of construction and cost. With proper design, one can combine high selectivity with low noise, an ideal combination for preamplifiers.

You might ask, "why can't I just combine a high selectivity filter with a low noise preamplifier, simplifying the design into two easier designs?" You could, but you generally incur additional loss every time you have to make an impedance transformation. In designing the system this way, you might expect to make a transformation from the filter down to 50 ohms and another transformation from 50 ohms to whatever impedance the active device of the preamplifier wants to see. Not a problem if you can afford a dB or two of loss, but what if you need a receiver with a 21-Kelvin noise temperature (0.3 dB NF) for low-power EME work?

Since many of our readers do like math, I'll take the liberty of highlighting some useful equations. For a single tuned circuit, the insertion loss is given by: $20\log(1-Q_l/Q_u)$ dB where Q_l is the loaded Q and Q_u is the unloaded Q.

This equation tells us to make the ratio of the loaded Q to the unloaded Q as small as possible. Otherwise, the loss of the circuit may be too high. (Incidentally, many people get Q_l and Q_u confused. This has resulted in arguments whether high Q or low Q is best. You can get low losses with a high Q_u, relative to Q_l, or a low Q_l, relative to Q_u.)

Generally, $Q_l^2 > R_a/R_b - 1$, with $R_a > R_b$, where R_a and R_b are the two resistances being matched. You might want to derive the exact equations for your network, but this equation is close enough for our discussion. With real networks used for low noise preamplifiers, there is a minimum Q that depends on how different from the source impedance you allow the filter input impedance to be. This might mean that an intrinsically lower noise device, such as a PHEMT, might actually be worse than an ordinary GaAs FET at a given frequency because the required Q_l is higher. It has been argued that one could just use an ideal transformer—perhaps a low-loss transmission line transformer, to match the impedances. But the dimensions for a 50-to-5000 ohm, 432 MHz (for example) transmission line transformer are a bit unreasonable!

There are also limits on how high Q_u can get. While it is true that the loss of a cavity decreases as its volume—and Q_u—increases, there are other considerations. Perhaps the most important is the possibility of multiple modes—different field distributions in the cavity. I think this sets a limit on the size of a cavity—the different field distributions will result in different coupling factors, which translates into a truly unreliable preamp.

Then there is the issue of silver solder. I have looked into this and found that there is little improvement in the resistivity of tin/lead solder when small amounts of silver are added. Perhaps high-silver-content solders (40+ percent) may be a significant improvement, but I don't have any data yet.

Coupling Into a Cavity

This is perhaps the biggest question concerning cavities. There are a variety of methods: loop coupling, probe coupling, etc. Unfortunately, many of the designs are empirically derived—determined by cut and try. The modern design method is to let a computer solve it via number crunching, but I've not seen an inexpensive program that will handle this for amateurs. For now, empirical designs are probably still the rule.

Getting RF Out of the Device

The preceding discussion has been about efficiently getting RF into the device. Present common wisdom is that you don't worry about getting RF out of the device efficiently. While an efficient output network *may* result in the best amplifier, it usually results in a lousy oscillator instead. Even at 1.3 GHz, modern FETs have more than enough gain for most applications, allowing you to waste some of it. A 47-Ω drain resistor and a coupling capacitor seem to work very well with the Avantek

ATF10135, giving 15 to 17 dB of gain. That's less than the device is capable of giving, but what you gain for this loss (excuse the pun) is ease of tuning. Typically, you need at least two adjustments for the input circuit—coupling and approximate resonance. If you had two output adjustments as well, you would probably wear out the adjustments before you found the optimum settings!

Now you know "Zack's secret to a good preamp": don't tune the preamp to death. Gold-plated parts are nice, but there really isn't much gold there, so don't wear it out. Tune the preamp just a few times, as further tuning is unlikely to improve performance. It also helps to know how tuning screws shift as you tighten the lock nuts. With typical screws, further clockwise motion forces the screw into the cavity. But tightening the lock nut pulls the screw back out of the cavity, so preset the tuning screw to account for this. (Of course, it takes experience to learn just *how much* to preset the tuning.) I find that copper tubing makes poor threaded holes—it may be necessary to solder a nut in place rather than relying on the tapped hole, though the tapped hole may be electrically superior.

Biasing the Device

So far, I've always source biased my cavity preamps. The rationale for this is that grounding the gate of the FET allows for minimal RF losses in the input circuitry. If gate biasing, where does one optimally add a blocking capacitor and RF choke to bring in negative bias? Most implementations I've seen put the RF choke at a high impedance point—not a good idea if you want minimal loss. Of course, the question really depends on the FET used: a high gain PHEMPT may not be stable with source biasing. Then again, that instability might just be due to the designer not working hard enough.

What Device is Best?

Ideally, I'd choose a device that could be made unconditionally stable with reasonable drain loading and source feedback. In other words, it wouldn't oscillate no matter how you adjusted the cavity, while still giving you good noise figure and gain. But the latest devices seem to be getting worse, not better, in this regard. The Mitsubishi power devices such as the MGF1801 and MGF2116 seem to do the best on 144/222 MHz. At 432 MHz, the inexpensive MGF 1302 seems to do as well as anything else, possibly because the advantage to be gained from a better device is small. At 903 and 1296 MHz, the choices seem to be the MGF 1302, Avantek ATF10135, and various PHEMTs. As I mentioned before, the loss-versus-impedance-matching issue may prevent PHEMTs from outperforming ordinary GaAs FETs at 0.9 and 1.2 GHz, but there hasn't yet been enough published on designs using the different devices to draw a conclusion.

Finally, cavity preamps might be worthwhile on the higher bands, now that we have devices that want to see such high impedances for optimum noise figure. It will be interesting to see the progress made in this area in the next few years.

Bits

Award Nominations Open Now

Nominations are now being sought for the 1993 awards to be presented at the ARRL Atlantic Division Convention. The Convention is held in association with the Rochester, New York, Hamfest, May 21-23, 1993. The awards are commemorated by handsome plaques to be presented at the hamfest banquet.

"Amateur of the Year" nominees should be outstanding all-around amateurs from the Atlantic Division with a strong record of service to the amateur community. An award for lifetime service to Amateur Radio, the "Grand Ole Ham," is open to Atlantic Division OMs and YLs who have been licensed at least 30 years or are at least 50 years of age. The Atlantic Division "Technical Achievement" award may be presented to an individual or to a group.

Complete information on the awards and nomination prodedures is available from Richard Goslee, K2VCZ, 24 Elaine Dr, Rochester, NY 14623. The deadline for nominations is April 1, 1993.

TAPR 1993 Annual Meeting

The Annual Meeting of Tucson Amateur Packet Radio (TAPR) will be held on the weekend of March 6th and 7th, 1993 at the Best Western Inn at the Airport, 7060 S. Tucson Blvd, Tucson, Arizona, adjacent to Tucson International Airport.

In addition to the usual presentations of the latest and greatest developments in packet radio, TAPR will host a workshop on digital signal processing (DSP) for radio amateurs. This workshop, to be conducted by Jon Bloom, KE3Z, ARRL Senior Engineer, is designed to teach the fundamentals of digital signal processing. The target students are radio amateurs who understand the basics of complex (modulated) signals and are familiar with computer programming. The course is intended to bridge the gap between these two subjects, with the result that the student will be able to begin programming DSP applications almost immediately. The student's math background should include algebra and trigonometry.

The topics covered by the presentation include:

- Discrete-time systems
- Sampling theory
- Digital filters
- Signal generation and detection
- Discrete and Fast Fourier Transforms, with applications
- Applications of DSP to Amateur Radio
- Development tools
- Review of available literature

We are expecting to have the traditional "pizza bash" and other informal activities on Friday night, March 5, with the meeting all day Saturday, March 6 and the DSP workshop on Sunday, March 7. Registration for the Saturday meeting is $15, including a buffet luncheon. There will also be a $5 registration fee for the DSP workshop on Sunday to cover the cost of printed workshop materials. A steak dinner on Saturday night will be available for $13.95. TAPR will have a hospitality suite where you can gather informally, join TAPR or purchase kits and software. The meetings will run from 9:00 AM to 5:00 PM both days.

A block of rooms has been reserved at the special rate of $53 per night, single or double occupancy, including full American breakfast and happy hour reception. For reservations, call the Inn at the Airport at 1-800-772-3847, or in Arizona at 602-746-0271, fax 602-889-7391 (mention TAPR to get this rate).

If you are planning to attend or have a project you would like to present at the meeting, please call or write the TAPR office and let us know. We also would like to know if you are planning to attend the DSP workshop, so sufficient materials will be available. To have your paper included in the printed proceedings of the meeting, camera-ready copy should be submitted to TAPR no later than February 19, 1993. For further information, contact the TAPR office.

Tucson Amateur Packet Radio
PO Box 12925
Tucson, AZ 85732-2925

tel: 602-749-9479 (10 AM to 3 PM, Tue-Fri)
fax: 602-749-5636

1993 West Coast VHF/UHF Conference

Steve Noll, WA6EJO, has announced preliminary plans for the 1993 West Coast VHF/UHF Conference. This year's event will be held May 21-23 at the Ventura Holiday Inn in Ventura, California. Registration begins at noon on Friday, and the hospitality room will be open. Technical talks and expanded vendor exhibits will be held all day Saturday, along with noise-figure measurements, a banquet (with speaker) and awards ceremony. The conference concludes Sunday with a breakfast (with speaker), an indoor swap meet and antenna range measurements. Plans also call for a codeless Technician class during the weekend. The ARRL will again publish a conference proceedings book.

General admission will be $15. Reservations are required for the Saturday night banquet ($25) and the Sunday breakfast ($10). Proceedings will be offered to conference attendees at the special price of $10. Rooms are $58 (contact the Holiday Inn at 1-800-842-0800). The Technician class is $125 (contact Loraine McCarthy, N6CIO, at 714-979-2633).

For registration forms and more information, contact Steve Noll, WA6EJO, 1288 Winford Ave, Ventura, CA 93004-2504.

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