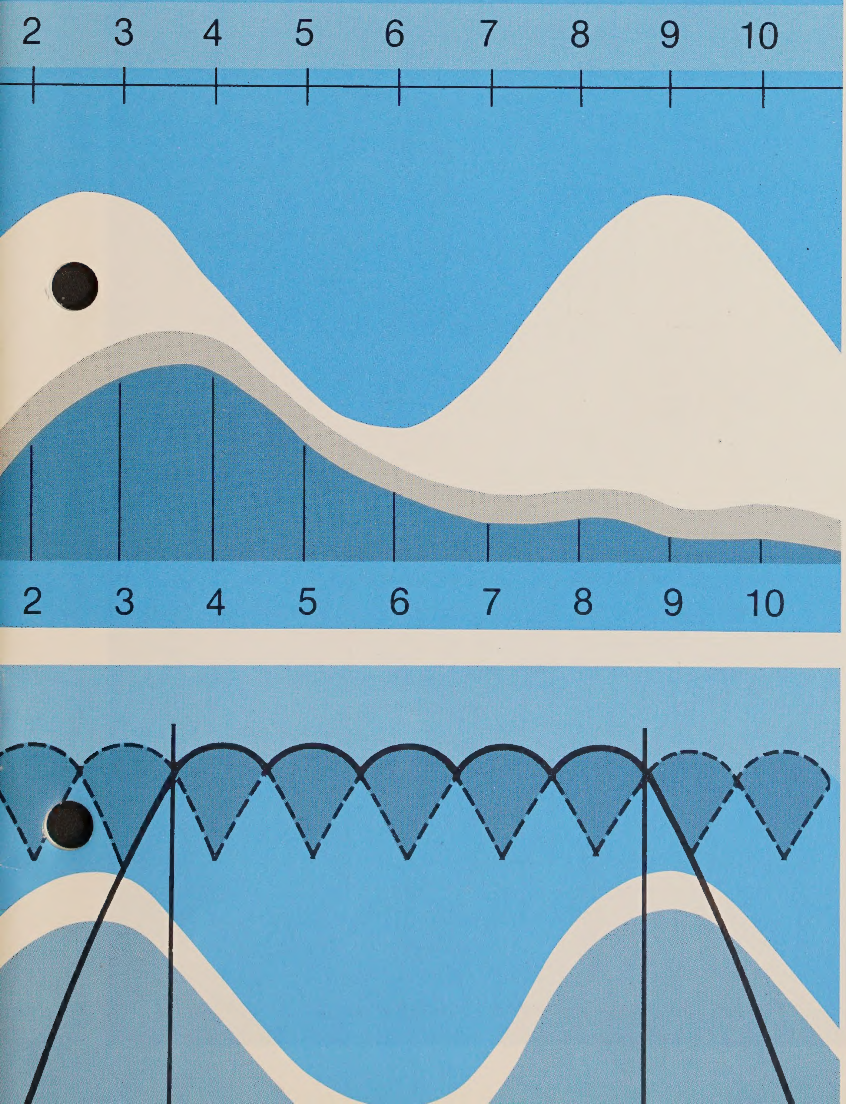
VOL. 9 NO. 3 MAY/JUNE 1982

Digital Signal Processing for Multichannel Receiving Systems



WATKINS-JOHNSON COMPANY
Tech-notes

Multichannel receivers exhibiting close amplitude and/or phase matching have seen considerable use in the fields of direction finding, antenna beam forming, and adaptive arrays. Examples of these systems are shown in Figure 1. In all of these applications the objective of the multichannel receivers is to provide accurate phase and, in some cases, amplitude information of the signal from each element of an antenna array. A typical multichannel receiver is shown in the Figure 2 block diagram. A key feature in multichannel receivers is sharing of common local oscillators between channels. Since the incoming signals are multiplied by the same local oscillators, their phase relationship remains preserved at the IF output. This assumes that all components and RF cables are matched. In many cases, matching is

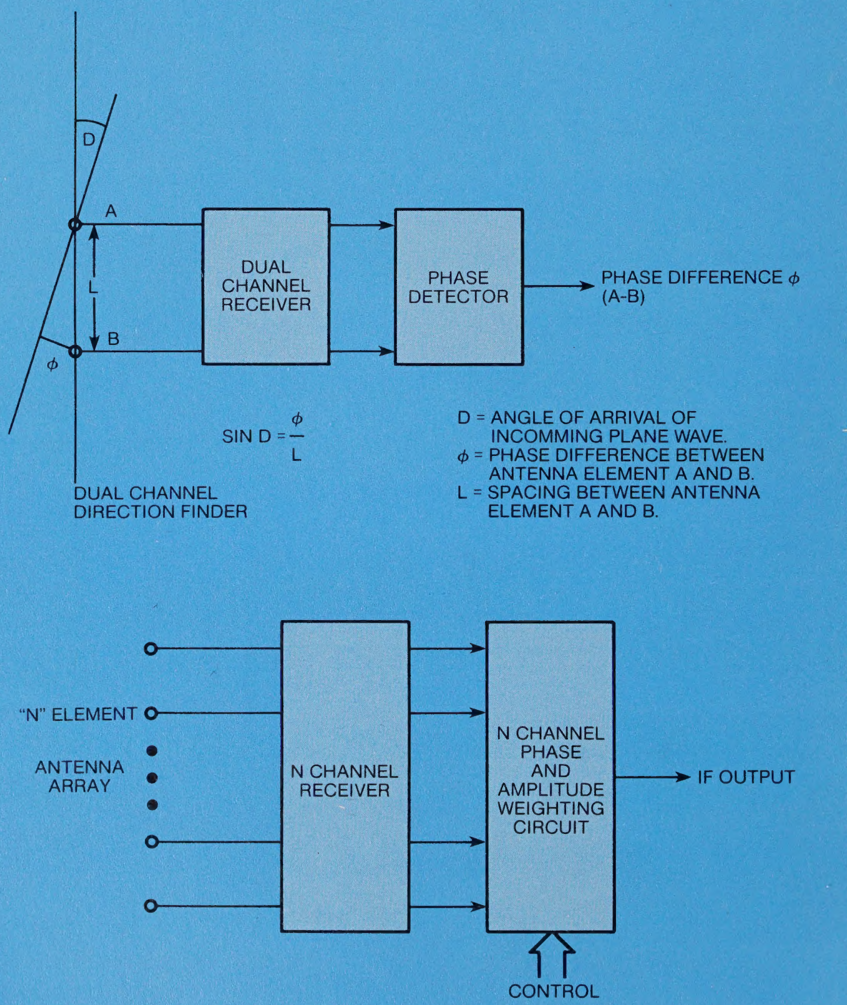


Figure 1. Beam forming system or adaptive array.

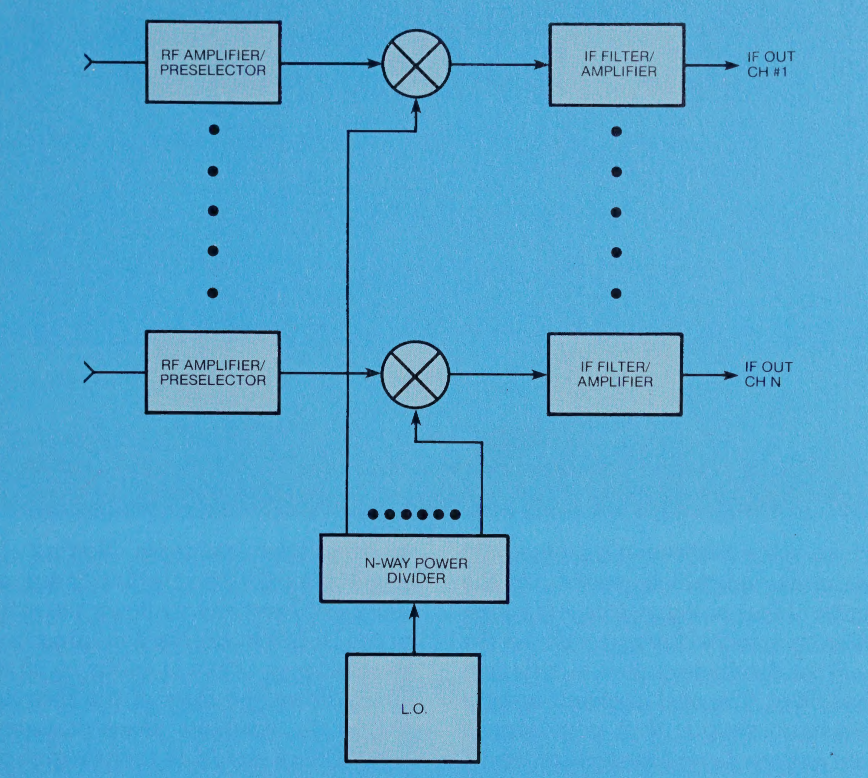


Figure 2. Block diagram of a typical multichannel receiver.

considered too expensive, especially in wide tuning-range receivers, so minimal effort is expended in achieving match, but various calibration schemes are employed to remove the static error between channels.

Figure 3 shows a common method used for calibration. A reference signal at the tuned frequency is fed into all channels; a measurement is then made, and the system is either "zeroed out" with variable phase shifters and attenuators, or the measured values are stored and subtracted out. Another scheme sometimes used with two-channel receivers is shown in Figure 4. In this method, no reference is required. The incoming signal is measured, the transfer switch is thrown, then the incoming signal is measured again. The reading is subtracted, cancelling out the error. The only penalty is that two readings of the incoming signal are required. This approach can be used on more than two channels, but rapidly becomes time-consuming as the number of channels increases. All of the above schemes can calibrate out phase errors due to linear time delay mismatch, such as unequal cable lengths.

The most difficult problem occurs as the result of group delay mistracking between the band-limiting filters in the IF. In most receivers, these filters are by far the narrowest and, therefore, contribute the largest delay. Mistracking across the passband can cause serious

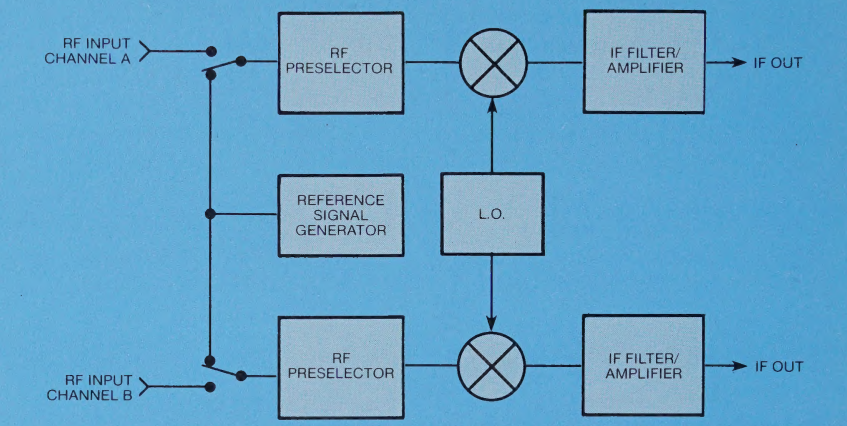


Figure 3. Dual-channel receiver with reference signal generator for phase and amplitude calibration.

errors, since calibration removes only the phase mismatch existing in the center of the passband. Phase tracking between matched filters to $\pm 1°$ over the 3-dB bandwidth is realizable, although expensive. The real problem is with wideband signals that have significant energy down the skirts of the IF filters. Figure 5 shows a typical 2-channel IF strip with both amplitude and phase detection capability. Most accurate phase detectors require signals that are normalized in amplitude. This is usually accomplished by hard limiting after bandpass filtering. A result of hard limiting is that if a signal has significant energy down the skirt of the IF filters, the hard limiting can cause energy which is as far as 10 to 20 dB down the skirt of the filter to weight the resultant phase measurement. Phase matching in the region of 10 to 20 dB down in the skirts of the filters is very difficult, especially if the filters have the steep skirts required for good adjacent-channel rejection. This problem is most serious and imposes a limitation which is difficult, if not

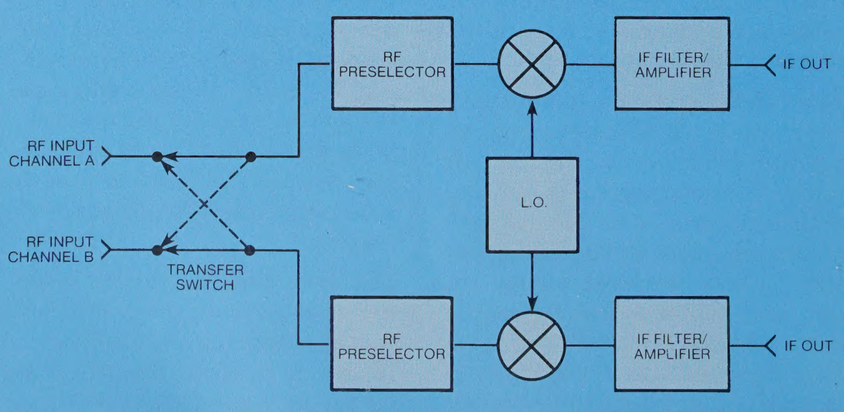


Figure 4. Dual-channel receiver with transfer switch for phase and amplitude calibration.

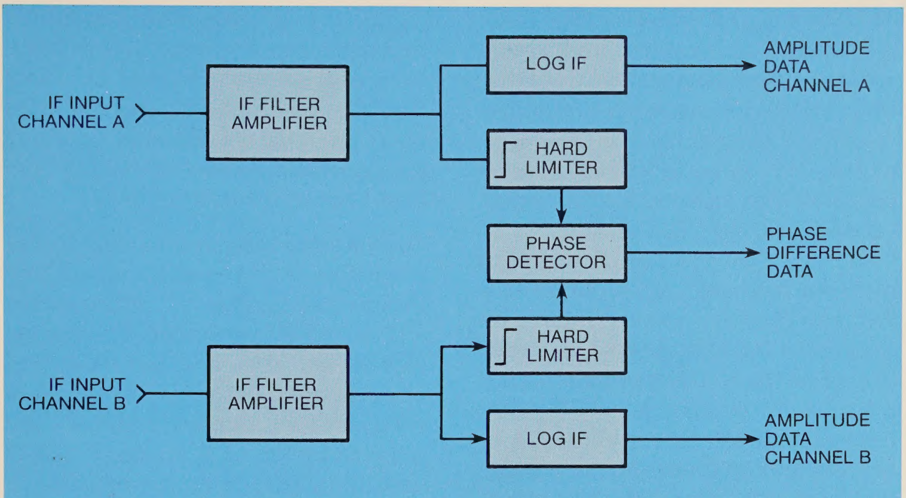


Figure 5. Typical dual-channel IF capable of providing both amplitude and phase data.

impossible, to correct. Fortunately, there is a reasonable solution using digital signal processing. Digital filtering permits realization of filters having precise amplitude and phase match, even far down the "skirts." Refer to Figure 6 for a block diagram of a typical multichannel receiver utilizing a digital processor. The main difference between this system and a conventional system is that the hard limiters and phase detectors are replaced by A/D converters. Since speed is a primary limitation in A/D converters, the IF is reduced to as low a frequency as possible. In digital signal processing systems, the dynamic range of the system is dependent upon the resolution of the A/D converter, where there is a trade-off between speed and resolution. Note that analog filtering must still be used, but the matching requirement for these filters is not severe, since there are methods to correct both group delay and amplitude errors with the digital processor.

To gain some insight into the mechanics of a multichannel receiver

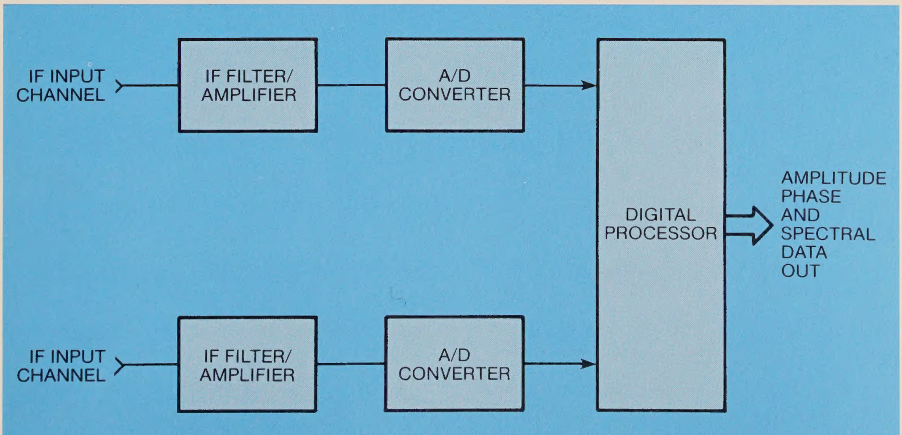


Figure 6. Typical multichannel receiver IF with digital processor.

utilizing a digital processor, it is useful to generate an example and assign some numbers. It should be noted that there are many approaches to implementing digital signal processing techniques; the example shown here is patterned after an experimental system developed at W-J CEI Division. Figure 7 shows a dual-channel IF strip with an input frequency of 21.4 MHz. It is assumed that the widest IF bandwidth required is 100 kHz. Since it is desired to operate the A/D converter at as low a frequency as possible, an IF center frequency of 100 kHz is chosen. This permits analog filters having a bandwidth of approximately 20 percent wider than the desired IF bandwidth, or 120 kHz, to be employed. When the IF at 21.4 MHz is mixed with an LO at 21.5 MHz, an IF of 100 kHz with a 3-dB bandwidth of 40 to 110 kHz is generated. Digitizing the IF can now be accomplished by sampling the IF and performing an A/D conversion at a 400-kHz rate. This information can then be processed. There are numerous digital signal processing algorithms that can be implemented in both hardware and software. One of the most flexible for multichannel systems where group delay and passband amplitude corrections must be made is the Fast Fourier Transform (FFT).

To gain some insight as to what occurs when the FFT is used for signal processing, refer to Figure 8, which shows a 16-point FFT performed. The sample rate is 400 kHz, and the waveform being sampled in time is a 50-kHz sine wave. Sixteen amplitude samples are required for the FFT, taking 37.5 μs of time. In this example, only real data in the time domain is sampled and is responsible for the aliasing effect noted.

Some FFT algorithms can accept complex amplitude data input. This requires an I and Q output from each channel. If complex amplitude data is used, then the aliasing frequency is equal to the sample rate, or 400 kHz. This does not violate the Nyquist criterion, since each sample would con-

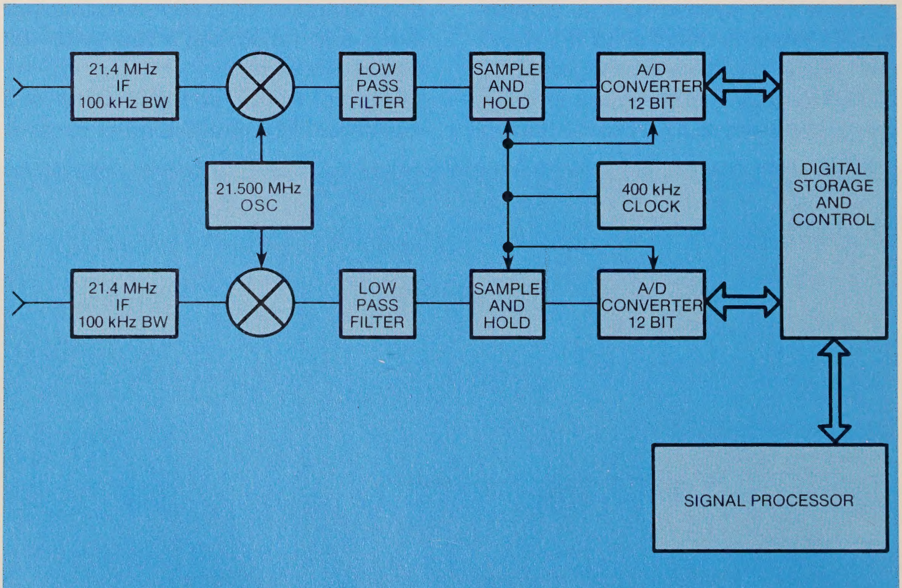


Figure 7. Dual-channel receiver utilizing digital signal processor.

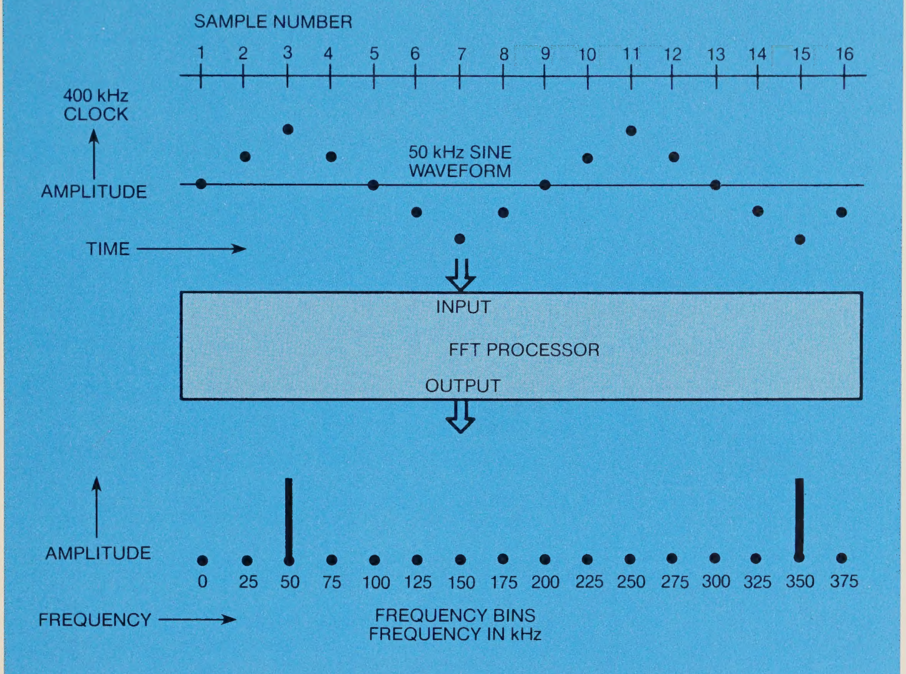


Figure 8. Example of 16-point FFT.

sist of two actual samples, one real and one quadrature.

The effect of this processing is as if there were N, in this case 16, contiguous filters with center frequencies spanning the range of 0 to 375 kHz. The bandwidth of each filter, or bin, is equal to,

$$\frac{\text{Sample Rate}}{N}$$

or, in this example, 25 kHz. The actual numerical output representing each frequency bin is a complex number whose amplitude represents that of the input component, and whose phase represents that of the input component referenced to the phase of the sample clock. If two or more channels are being sampled at the same time by the same sample clock, then the phase difference between channels is obtained simply by complex division, since in complex division the angles are subtracted.

In the example given, everything is ideal; that is, the input waveform is a sine wave whose frequency exactly corresponds to the center of a frequency bin (50 kHz, which falls in the center of bin number 3). If the example were not ideal, the result would be quite different, as is shown in Figure 9, where the frequency is 62.5 kHz, corresponding to a frequency halfway between bin 3 and bin 4. Note that all frequency bins are responding. This is caused by side lobes resulting from the fact that only a finite number, 16 in this case, of amplitude samples was taken. The spectrum that results is approximately the same as that of a rectangular RF burst that is 37.5 μs long at a frequency of 62.5 kHz. This effect will

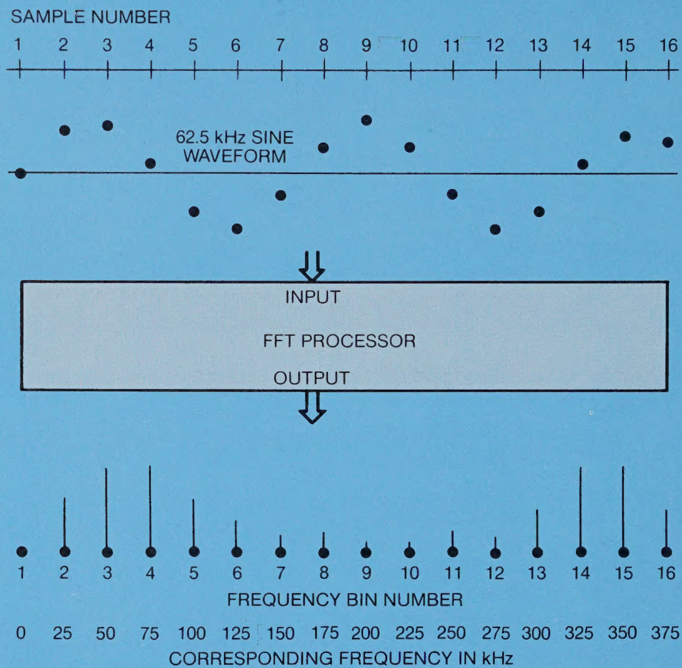


Figure 9. Example of FFT showing side lobe response.

occur with any frequency component that does not lie exactly in the center of a frequency bin. Fortunately, this effect can be greatly minimized by weighting the amplitude of the samples in time before performing the FFT. The function which describes the weights is called a *window function.* For an excellent description of window functions and their performance, refer to Proceedings of the IEEE, Vol. 66, No. 1, January 1978, "On the Use of Windows for Harmonic Analysis with the Discrete Fourier Transform," by Fredric V. Harris.

The effect of applying a window function and then performing an FFT is shown in Figure 10. Notice the considerably improved performance. The penalty for using a window function is the broadening of the apparent bandwidth of each frequency bin, somewhat similar to increasing the overlap on a set of analog contiguous filters. Nevertheless, it is a small price to pay for the increased performance. If a window function is chosen which can suppress sidelobes by approximately 70 dB, the increase in the width of each frequency bin is typically 1.68:1. In comparing this digital analysis technique to analog techniques, it is interesting to compare the "shape factor" of the resulting frequency bin. In terms of bandwidth, in this example where the unweighted frequency bin is 25 kHz wide, the 3-dB bandwidth of the weighted bin is 42 kHz, the 6-dB bandwidth is 45.25 kHz, and the 60-dB bandwidth is approximately 175 kHz, or a 3-dB to 60-dB shape factor of 4.17.

In staying with the example shown, we are only interested in frequencies from approximately 50 kHz to 150 kHz. Therefore, frequency bins 1 through 3 and bins 7 through 16 are set to zero. A

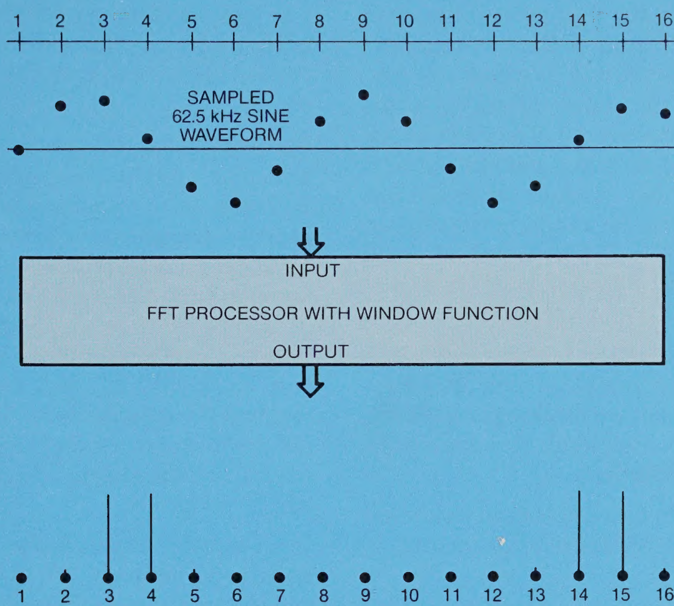


Figure 10. Example of FFT showing effect of window function.

plot of the resultant amplitude response is shown in Figure 11. The actual 3-dB bandwidth is 92 kHz. It can be seen that to eliminate interfering signals, frequency bins in the passband can be set to zero, thereby generating notches. Other signal enhancement schemes can be employed, such as setting the bandwidth to closely match the signal, or processing only those frequency bins exceeding a preset threshold. To correct group delay and amplitude errors in the system, one has only to input a wideband signal, such as noise, of

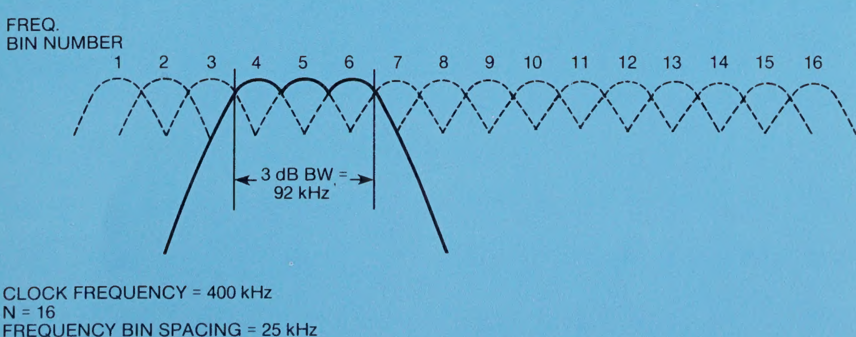


Figure 11. Composite response of frequency bins number 4 through 6.

equal phase and amplitude into all channels, measure the differences in corresponding frequency bins, and generate correction. This corrects any group delay as well as amplitude mismatch in the entire system up to the calibrating source. If N is made larger, thereby making the bandwidth of each frequency bin smaller, fine-grain correction can be made, resulting in realizable amplitude match of small fractions of a dB and phase match of small fractions of a degree.

Digital processing permits realization of a multichannel receiver with capability which would be extremely difficult, if not impossible, to achieve with analog components.

Grouping frequency bins can yield effective filters with shape factors approaching 1:1.

In the example shown, if N were 512 with a sample clock of 400 kHz, the frequency bin spacing would be 781.25 Hz.

$$S_B = \frac{400 \text{ kHz}}{512} = 781.25 \text{ Hz}$$

Figure 12 shows the composite response of bin number 65 through 193 for this case. Assuming the same window function as employed in the previous example, the 3-dB bandwidth is 100.8 kHz and the 60-dB bandwidth is 105.5 kHz, resulting in a shape factor of 1.047:1.

Once signals are transformed from the time domain to the frequency domain and reduced to numerical values, subsequent processing is greatly simplified. Phase and amplitude differences between channels can be measured for direction finding applications, or weights both in amplitude and phase can be applied to each channel as required for antenna beam forming or adaptive arrays. Actual signal recognition, enhancement, and processing can be performed.

If an analog output in the frequency domain is required, an inverse FFT can be performed on the processed time domain data. An example of this is shown in Figure 13.

By relaxing analog and RF hardware requirements, digital signal processing is proving to be a cost effective approach to achieving performance that otherwise is unattainable at almost any cost.

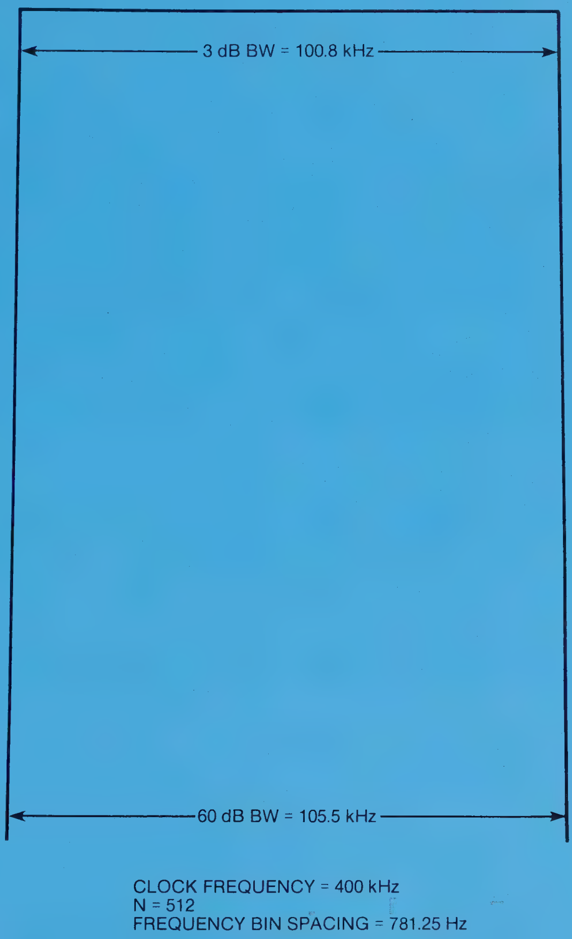


Figure 12. Composite response of frequency bins number 65 thru 193 when n = 512.

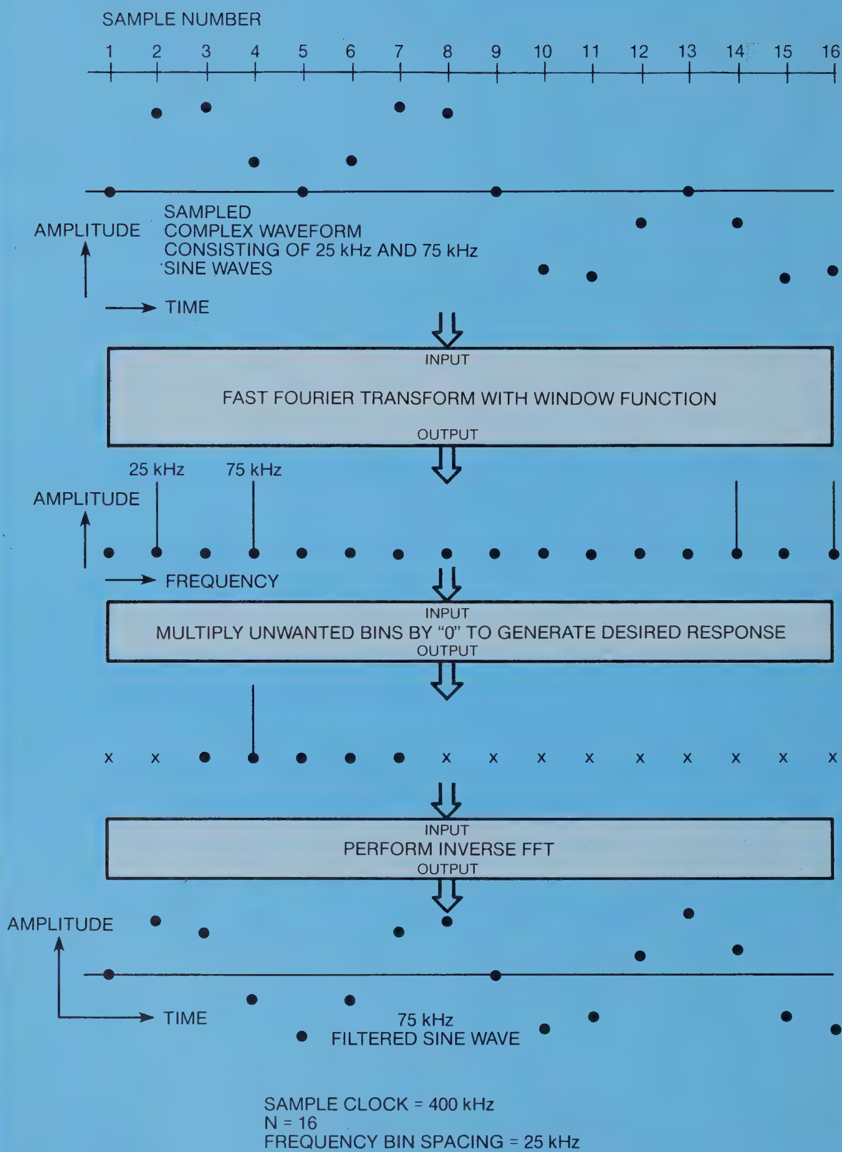


Figure 13. Example of a digital processor being utilized as a filter.

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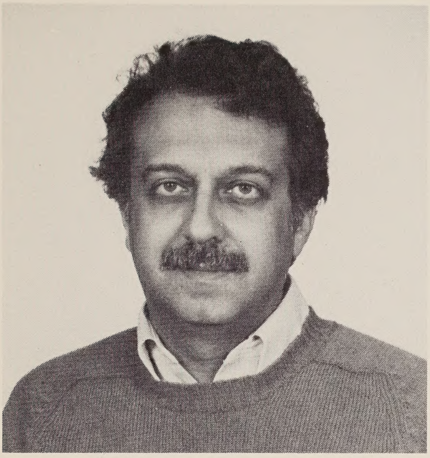

Benedict J. Nardi

Mr. Nardi is responsible for design and development of new products and research into new product areas.

He was responsible for the development of a radio direction finder and associated antenna systems, a high performance portable receiver featuring synthesized L.O.'s, various special purpose compact antennas and miniature receivers, improvements in the WJ-8940 EMI/RFI receiving system resulting in the "B" version, the prototype WJ-8622 compact VHF/UHF receiver, the WJ-9023 digitally controlled receiving system and the TH-Series solid-state microwave tuning heads. He is also responsible for the design of several AM demodulators and compressive RF amplifiers used in the QRC-259 program.

Prior to joining the CEI Division, he was responsible for design and development of digitally tuned microwave frequency synthesizers and various microwave hardware, including broadband frequency multipliers, mixers, upconverters and filters.

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Printed in U.S.A.