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The use of an A/D converter is considered for digital sampling of modulated waveforms obtained from an antenna. Demodulation is assumed to occur following the digital sampling process, either by using a high speed digital processor or by using a more conventional receiver following reconstruction of the analog waveform by a D/A converter. Since many communication carriers are not conditioned for jam-resistance by coding and spectrum-spreading techniques, a large instantaneous linear dynamic range is necessary to permit the detection and modulation of a small desired signal in the presence of a strong inband interfering signal. Consequently, our interest lies in A/D converters having a large number of quantization bits — a condition that is necessary for large linear dynamic range. The effects of A/D device parameters such as quantization resolution (number of bits), sampling rate, and aperture uncertainty are discussed.

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19. ABSTRACT (Continued)

A 12-bit A/D converter having an aperture uncertainty of 25 psec and a maximum sampling rate of 5 MHz was used to experimentally determine the magnitude of the spurious-free dynamic range that could be obtained with present off-the-shelf technology. A larger number of quantization bits would have resulted in unacceptably slow speed and an unacceptably low maximum frequency limitation. A smaller number of bits would have yielded a smaller spurious-free dynamic range. Since we did not have a digital processing facility to analyze the A/D directly, a D/A converter, deglitcher (sample-and -hold), and a high quality spectrum analyzer were used to evaluate the A/D performance. Consequently, the results reported are characteristic of the composite A/D-D/A-deglitcher system rather than the A/D alone. These results, however, are directly applicable to a first generation towed communication buoy system which would probably use the composite A/D-D/A-deglitcher architecture.

For narrow-band applications in which second-order mixing products between input frequencies fall out of band, a spurious-free dynamic range of 66.6 dB was observed. This measurement was obtained with two-tone excitation at 1.54 and 1.57 MHz using a sampling rate of 1.236 MHz and an analysis window of 100 kHz centered on the test tones. For most data observed, the maximum spurious signal was formed from a two-frequency, third-order interaction. If three-tone excitation had been used, third-order intermodulation products of the form $F_1+F_2-F_3$ would reduce the spurious-free dynamic range by an additional 6 dB.

For broadband applications, the inclusion of second-order mixing products reduces the spurious-free dynamic range to about 55 dB. This measurement was obtained with three-tone excitation at 20, 45, and 100 kHz using a sampling rate of 5 MHz and an analysis window extending from 10 to 200 kHz.

The Use of a High-Speed 12-Bit A/D Converter for Digital Sampling of Communication Signal Carriers

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December 31, 1984

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CONTENTS

iii

THE USE OF A HIGH-SPEED 12-BIT A/D CONVERTER FOR DIGITAL SAMPLING OF COMMUNICATION SIGNAL CARRIERS

INTRODUCTION

The objective of this report is to evaluate the utility of using a high-speed, high-resolution A/D converter to digitize radio frequency (RF) and intermediate frequency (IF) waveforms that are obtained from a receiving antenna. This approach to digital conversion of analog signals is of considerable interest for towed communication buoy applications because the large number of communication/navigation bands that are serviced and the diversity of modulation schemes employed prevent the use of complete receivers at the buoy for each channel of interest. Space limitation, cost, and limited lifetime expectancy associated with the buoy platform generally dictate a split receiver architecture in which the RF and IF sections are located in the buoy and the succeeding stages of the receiver are located at the platform at the opposite end of the tow cable umbilical link. With the split-receiver architecture, digital sampling at the buoy involves working directly with the RF or IF carriers that have not been demodulated. Consequently, digital sampling at this stage demands that demodulation be accomplished at the far end of the link by appropriate processing of the sampled RF/IF waveform.

Demodulation of the digitally sampled RF/IF waveform can be accomplished using a fast digital processor [1,2,3,4]. Software algorithms have been developed [2] to perform the heterodyning, demodulation, filtering, and automatic gain control functions normally accomplished with analog electronic hardware. Carriers with different forms of modulation are accommodated by calling upon the appropriate demodulating algorithm. This all-digital approach to signal processing is desirable because it eliminates the need for subsequent D/A conversion and analog frequency translation which would be necessary if standard analog receivers were to be used. The additional processes of D/A conversion and frequency translation will contribute some degradation to signal quality in the form of distortion (differential nonlinearity of D/A) and phase jitter (frequency translation).

The motivation for digitizing the received signal waveforms at the buoy is to reduce the loss in signal quality when the communication carriers are transferred across the tow cable transmission line(s). Traditional methods of analog signal transfer over electrical transmission lines within the tow cable have suffered from the effects of limited bandwidth (less than 1 MHz-km bandwidth-distance product), poor crosstalk rejection (30 dB) between pairs of transmission lines, ground loops, and noise pickup from external sources - - particularly at VLF frequencies.

The combination of digital modulation and fiber-optic transmission lines within the tow cable offers a solution to the noted problems of signal transfer. The optical fibers provide a large bandwidth-distance product (several hundred MHz-km) and freedom from external noise pickup, crosstalk, and groundloop effects. The use of digital modulation circumvents the problem of limited linear dynamic range associated with optical sources such as lasers and LEDs.

For the length of tow cable typically used in a communications buoy application, the optical loss will be sufficiently small that a data rate of several hundred megabits/second is possible with negligible error rate. Consequently, the only degradation associated with carrier transfer over the tow cable link will be that asssociated with the digital sampling process. The limitations of the sampling and quantization processes are governed by a variety of factors as explained below.

<u>The Sampling Theorem [5,6,7]</u>

The sampling theorem, as applied to low-pass signals, states that if the waveform being sampled has spectral content from DC to a maximum frequency F_m, the sampling rate must be at least $2F_m$ (the Nyquist rate) to prevent aliasing (the introduction of spurious low frequency components). In many situations it is impractical to abruptly limit the high frequency end of the waveform spectrum by employing an ideal (physically unattainable) rectangular low-pass antialiasing filter. Antialiasing filters, which provide acceptable phase shift and ripple characteristics within the pass-band, will always have a finite high-frequency roll-off rate. In this case the sampling rate must be higher than twice the 3 dB cut-off frequency to prevent contamination from aliasing products originating from the power contained in the skirt of the filtered spectrum. For a uniform response in the passband region and monotonically decreasing response in the filter roll-off region, the assumption of an overall linear system implies that all discrete aliasing products will individually be at least N dB down from maximum inband signal level if the sample rate is at least twice the frequency corresponding to the $-N$ dB point of the high frequency cut-off of the waveform spectrum.

When the waveform to be sampled is band-pass limited, it is not necessary to sample at a rate of twice the highest frequency of interest in the band. Aliasing can be avoided by sampling at a rate of at least twice the bandwidth of interest provided the appropriate relationship exists between the sampling frequency and the center frequency of the band. The relationship required between these quantities is developed in the following paragraphs.

If a CW signal of frequency F is sampled at a rate F_s, it can be shown [5(p.509), 6(p.297)] that the spectrum of the sampled waveform will contain spectral lines at frequencies F, $|F-F_s|$, $|F-2F_s|$, $|F-3F_s|$,....., $|F+F_s|$, $|F+2F_s|$, $|F+3F_s|$,....... In order to determine the effect that the aliasing products between F and harmonics of F_s (listed above) have on a sampled waveform of finite bandwidth, it is instructive to vary the frequency F and observe the changes in the spectral positions of the

2

intermodulation products. In Fig. 1 the relative positions of the spectral components are illustrated for two cases (a,b) where F is less than the Nyquist frequency ($F_s/2$) and for one case (c) where $F_s/2 < F < F_s$. In Fig. 1(a) F is located at $0.3F_s$ and the closest aliasing product is $F_s - F$ located at $0.7F_s$. In Fig. 1(b) the position of F has been changed to $0.4F_s$ and $F_s - F$ shifts to $0.6F_s$. If F is set equal to $F_s/2$, then $F_s - F$ is also equal to $F_s/2$ resulting in an overlap of fundamental and aliasing products. In Fig. 1(c) F is positioned above $F_s/2$ and, consequently, $F_s - F$ drops below $F_s/2$. Figs. 2(a) and 2(b) illustrate the relationship of fundamental and aliasing products for $F_s < F < 3F_s/2$. It is clear from the illustrations that as long as the maximum waveform frequency is less than the Nyquist frequency, there will be no contamination of the waveform spectrum. If $F > F_s/2$, then the sampled waveform spectrum will contain an aliasing component unless the passband (assumed to be symmetric) of the waveform is limited according to the relationship

$$BW = 2 \text{ X minimum value of } |F - NF_s/2| \tag{1}$$

where N=1,2,3,......... The maximum bandwidth (BW) of the sampled waveform is $F_s/2$ and this condition is achieved only when the center frequency of the symmetric pass band is set at an odd multiple of $F_s/4$. Therefore, to achieve the full Nyquist bandwidth when sampling symmetric-spectrum waveforms such as an IF carrier, the center frequency F_c must be related to F_s as

$$F_c = (2N-1)F_s/4 \tag{2}$$

In radio (communication) receiver applications, this relationship can be satisfied by designing the last heterodyning (mixing) stage to produce a final intermediate frequency equal to an odd multiple of $F_s/4$ or by appropriately changing the strobing frequency (F_s) of the A/D converter. Although there is no upper limit imposed on the choice of F_c by the sampling theorem, an upper limit is set by uncertainties or jitter in the sampling aperture.

Aperture Uncertainty.

Under ideal conditions, the sampling of a waveform is accomplished instantaneously at each sampling time. Mathematically this is equivalent to multiplying the continuous waveform by a periodic unit amplitude rectangular pulse that has a vanishingly small temporal width. Under these conditions (and ignoring quantization round-off inaccuracies) the samples are an accurate representation of the continuous waveform at the sampling instants. In real situations, the acquisition of the sample takes a finite period of time which may be viewed as a sampling aperture. The existence of a finite sampling aperture is significant only when the product of the time rate of change (slew) of the waveform and the aperture width approaches the magnitude of the least significant bit of the A/D converter. According to Kester [8], the sampling error that occurs for a

constant and repeatable aperture width is equivalent (at least in a first-order analysis) to an effective fixed sampling-time delay in conjunction with a zero-aperture sampling switch. Since compensating propagation delays can be incorporated into the logic circuits, major performance limitations are not caused by constant sampling apertures, but rather by variations or jitter in the aperture caused by phase modulations of the strobe signal from extraneous sources such as noise, power-line frequency pickup, and digital ground current effects. True, uncorrectable aperture error is generally described by the aperture-jitter parameter and is specified by most A/D and track-and-hold device manufacturers.

The maximum frequency that can be sampled with error less than one half of a least significant bit (LSB) can be calculated from knowledge of the aperture jitter "t" and the number of quantization bits in the A/D converter. For a sinusoidal signal of maximum amplitude A and frequency F_m, the maximum slew is $2\pi AF_m$. The maximum value of A is equivalent to 2^{N-1} bits for an N bit quantizer. Therefore,

$$2^N \pi F_m t \leq 1/2 \tag{3}$$

for the error to be less than LSB/2. Kester [8] recommends using a value of t measured at the 2-standard-deviation level in the statistical distribution of t values, rather than an rms value. Assuming the distribution to be Gaussian, we have $(t)_{2\sigma} = 2(t_a)$ where t_a is the rms value of t. Therefore, solving equation (3) for F_m we have

$$F_m \leq \left[2\pi 2^{N+1}t_a\right]^{-1} \tag{4}$$

For a 25 psec rms aperture uncertainty (jitter) and 8-bit quantization, an A/D converter can accurately digitize a 12.4 MHz sinewave. If the quantization is increased to 12 bits, the maximum frequency that can be accurately digitized decreases to 0.78 MHz.

Signal-To-Noise Ratio

For a system with a large number of quantization levels in which the random variable is uniformly distributed in the region between quantization levels, the rms noise arising from the quantization coarseness is given by [5(p.522)]

$$\text{rms quantization noise voltage} = \Delta/\sqrt{12} \tag{5}$$

where Δ is the physical measure of quantization step size (if the measure is in terms of bits, Δ is set to unity). For an analog bipolar input to an N-bit A/D, the maximum rms signal is

$$(V_{rms})_{max} = 2^{N-1} \Delta/\sqrt{2} \tag{6}$$

Therefore the signal-to-quantization noise ratio (SQNR) is equal to $(V_{rms})_{max}^2/(\text{mean-square of quantization noise voltage})$

4

$$SQNR = (3)2^{2N-1} \qquad (7)$$

or expressed in decibels

$$SQNR(dB) = 6.02N + 1.8 \qquad (8)$$

For a 12-bit quantizer, the SQNR is equal to 74 dB measured in the Nyquist bandwidth ($F_s/2$ Hz). If the sampled waveform is a narrowband signal, then theoretically the SQNR can be increased by the factor of 10 log ($F_s/(2 \times$ signal bandwidth)). However, our observations with a 12-bit A/D converter indicate that spurious signals (intermodulation products) in the form of $[MF_s \pm NF]$ will generally be present at amplitudes sufficiently high to seriously limit the utility of large gains in SQNR by bandwidth narrowing.

In addition to the quantization noise, the aperture uncertainty contributes a noise voltage of the form [8]

$$\text{rms aperture noise voltage} = \pi F 2^N t_a \ \Delta/\sqrt{2} \qquad (9)$$

Including the effects of both quantization noise and aperture uncertainty, the SNR is given by

$$SNR = \frac{3(2^{2N-1})}{1 + 6 \ (\pi F t_a 2^N)^2} \qquad (10)$$

There will be a 3 dB decrease in SNR relative to the SQNR when the second term in the denominator equals unity, which is equivalent to

$$F = \left[\sqrt{6} \ \pi \ t_a \ 2^N \right]^{-1} \qquad (11)$$

For a 12-bit quantizer and an aperture uncertainty of 25 psec, equation (11) predicts a frequency of 1.27 MHz.

Spurious Signals

Spurious signals arise from a variety of factors including differential nonlinearity of the transfer characteristic of the D/A converter (frequently caused by the inability to completely eliminate offsets at the merging points of the R-2R ladder networks), imperfect grounding techniques [9] or circuit layout which cause false or missing digital codes, and switching transients (glitches) in the D/A converter (which must be removed by a sample-and-hold device). Since most A/D converters use D/A devices as part of the conversion circuitry, the effect of differential nonlinearity also applies to A/D devices. Other contributors to spurious signals are feedthrough, integral nonlinearity, and sensitivity to power supply voltage levels. "Integral nonlinearity" is a term used by the engineering community to describe the nonlinear effects arising from the large scale curvature of the input/output transfer characteristic. The term "differential nonlinearity" is used to describe the nonlinear effects arising from discrete, small scale offsets in the transfer characteristic.

5

Most of these effects are described well in application notes provided by manufacturers such as Analog Devices, Burr-Brown, ILC Data Device Corp., Hewlett Packard, etc. and will not be dwelt upon here.

We emphasize, however, that the nature of the nonlinearity of an A/D or combined A/D - D/A system is different from that usually experienced with analog systems. The magnitudes of the intermodulation products do not necessarily diminish monotonically as the analog input to the A/D converter is decreased. High orders of intermodulation products (IM) between the input signals and the sampling frequency are always present. We have observed IM products at almost all orders through the fourteenth (i.e to 9 X (input frequency) - 5 X (sample frequency)) in the output spectrum of the deglitched D/A. The deglitcher is a sample-and-hold device used to acquire the D/A signal near the center of the D/A clock period (thereby eliminating transients or glitches that occur at the clock period transitions) and hold the sampled value until the midpoint of the subsequent D/A conversion period. The use of the deglitcher eliminates or reduces the magnitude of intermodulation products that would otherwise occur from the glitches in the D/A output at the sample period transitions.

A partial listing of intermodulation products observed using a 12-bit Analog Devices MOD-1205 A/D converter is given in Table 1. To analyze the output of the A/D converter, the parallel digital output was converted back to analog using an Analog Devices HDS-1250 D/A and an HTS-0025 Track-and-Hold Amplifier as a deglitcher for the D/A converter. The output of the deglitcher was analyzed using an HP8568A spectrum analyzer. With this procedure, the composite properties of the A/D - D/A system are interrogated rather than just those of the A/D alone. An A/D sampling rate of 1.2366 MHz was used to digitize a sinewave of 0.500 MHz frequency. A lowpass filter was not used for conditioning the reconstructed signal prior to spectral analysis because this would eliminate the higher frequency intermodulation products that we wish to observe and would be ineffective in removing intermodulation products below the signal frequency.

The spurious products were observed at frequencies corresponding to $(N_S F_S - NF)$ where F_S is the sampling frequency and F is the 0.5 MHz sinewave. The upper number in each matrix block of Table 1 is the level in dB of the spurious product relative to the 0.5 MHz sinewave; the lower number is the frequency in MHz.

The products in the first row correspond to aliasing components and are large in magnitude. The magnitude of the 0.5 MHz CW tone was -23 dB relative to full scale (+16 dBm) on the A/D converter; the observed noise floor at 0.5 MHz was -87.5 dB relative to full scale using a 1 kHz analysis bandwidth. It is clear that for a processing bandwidth of a few kHz, spurious signals limit performance more seriously than does quantization noise.

6

TABLE 1
Characterization of Intermodulation Products
From a 12-Bit A/D - D/A System

N \ Ns	1	2	3	4	5
1	-2.7* .7366	-11.3* 1.9732	(relative level, dBc) (frequency, MHz)		
2	-45.1 .2366	NO 1.4732			
3	-50.2 .2634	-63.1 .9732	NO 2.2098		
4	-53.7 .7634	-50.1 .4732	NO 1.7098		
5	NO 1.2634	OB .0268	NO 1.2098		
6	-62.6 1.7634	-58.3 .5268	-61.2 .7098	- 70.6 1.9464	
7	NO 2.2634	-63.4 1.0268	-51.5 .2098	OB 1.4464	
8	NO 2.7634	-64.1 1.5268	NO .2902	NO .9464	
9	3.2634	-69.4 2.0268	-61.6 .7902	-59.3 .4464	OB 1.683

Frequency $[IM (N,N)_s] = N_s F_s - NF$

$F = 0.50$ MHz
$F_s = 1.2366$ MHz
NO -- not observed
OB -- observed, but magnitude not recorded
*   -- aliasing product
dB_c -- dB relative to carrier

Frequency Response

The frequency response of the A/D conversion process is flat up to a frequency at which uncompensated aperture effects set in. However, when the dynamic performance of an A/D is observed by reconstructing the analog signal through a D/A and deglitcher, frequency response is strongly affected. The staircase output of the glitcher (rectangular pulses of period $1/F_s$) exhibits an amplitude response [9] of

$$A(F) = \sin(\pi F/F_s)/(\pi F/F_s) \tag{12}$$

when followed by a low-pass or band-pass filter.The power response of the deglitcher is shown in Fig. 3 as a plot of 20 log A versus F/F_s. Also shown are corresponding data measured for the A/D-D/A-deglitcher system at sampling frequencies of 1.236 and 5.000 MHz. The good agreement between measured data and the theoretical response of the deglitcher implies that the response of the A/D converter alone is flat over the range of frequencies studied and the observed response is attributable to the deglitcher.

Frequency Response of the SNR

When including a deglitcher in the signal reconstruction system, the previous considerations of frequency response indicate significantly reduced response for F/F_s greater than 0.45. However, since quantization noise will also be affected by the same mechanism, the SQNR may exhibit a flatter response than that of Fig. 3. Consequently, measurements of SQNR were made on our 12-bit A/D-D/A-deglitcher system at frequencies up to 1.47 times F_s as shown in Fig. 4. Although the response is again rapidly attenuated at F_s, it is less severely attenuated at frequencies approaching F_s. The data between F_s and $2F_s$ indicate that the reconstruction of bandlimited IF carriers with center frequencies greater than F_s is favorable provided the frequency selection guidelines are followed. The relationship between the desired and spurious signal levels for $F > F_s$ will be discussed in the Multiple-Tone Testing section.

A/D-D/A EQUIPMENT

Because of the application requiring A/D conversion of RF and/or IF carriers derived from an antenna, both conversion speed and quantization resolution (number of bits) are important parameters. In order to provide adequate instantaneous linear dynamic range to the front-end of the receiver, the A/D must have sufficient resolution to permit post-sampling recovery of a desired weak signal in the presence of a strong interfering signal that is offset in frequency but still within the IF passband. A large number of quantization bits (with corresponding linearity, freedom from spurious products, etc.) will provide a large instantaneous dynamic range.

8

Fast A/D conversion speed is required to accommodate several time-division-multiplexed (TDM) channels at the required bandwidths. A target goal for an initial application might be a 4-channel TDM service for HF, UHF, and VLF-LF (8-150 kHz; two identical channels). The effective conversion speed per TDM channel (1/4 the A/D conversion speed for a 4-channel TDM system) must provide a Nyquist bandwidth equal to the alias-free bandwidth of the channel considered.

For example, with F_s equal to 5 MHz the Nyquist bandwidth will be 625 kHz for a 4-channel TDM system. Considering the finite roll-off characteristics of practical bandpass and lowpass antialiasing filters, the usable 3 dB bandwidths will be significantly smaller than 625 kHz but still adequate for the channels considered. If we use the equation [10]

$$P/P_o = \left[1 + (Q \times BW/F_o)^2\right]^{-n} \tag{13}$$

to relate the power response (P/P_o) to bandwidth (BW) and Q of a single-tuned bandpass filter of "n" cascaded circuits, then for n = 4, Q = 13, and F_o (center frequency) = 1.5 MHz, the bandwidth is 638 kHz at the −60 dB points and 50 kHz at the −3 dB points -- suitable for most HF and UHF applications. For the VLF-LF band, a 7-element lowpass Chebyshev filter can be constructed to provide a response that is −1 dB at 194 kHz, −3 dB at 213 kHz, and −50 dB at 481 kHz. Consequently, for a 4 channel TDM system consisting of 2 VLF-LF channels and a UHF and HF channel, an A/D conversion speed equal to or greater than 5 MHz is desirable.

The aperture uncertainty of the track-and-hold (used ahead of the A/D to minimize the aperture jitter of the sampling process) must be small enough to support the highest IF frequency used. As previously discussed, the larger the number of bits in the A/D converter, the lower will be the frequency that can be sampled without aperture error. Obviously, with realistic technical constraints there must be a trade-off or a balance achieved in the demands on quantization resolution, conversion speed, and aperture uncertainty.

In the search for suitable A/D conversion hardware, products from Analog Devices, ILC Data Device Corp., Micro Networks Co., Phoenix Data Inc., and Datel Intersil Inc. were considered. Based on an overall comparison of component properties, with emphasis on number of quantization bits, conversion speed, aperture uncertainty, and AC linearity, the components available from Analog Devices seemed to be most appropriate and were procured for laboratory investigation.

9

The components included the following:

o MOD-1205 A/D Converter

    - 12-bit resolution

    - built-in track-and-hold with
      25 psec aperture uncertainty

    - conversion rate of DC to 5 MHz

    - AC Linearity:
      (DC-1 MHz):   spurious signals 70 dB//FS:
      (1 MHz-2.5 MHz): spurious signals 65 dB//FS:

    - transient response and overvoltage recovery time
      to 12-bit accuracy equal to 200 nsec.
    - power dissipation: 13 watts
    - size: approximately 222 cm^3
    - input impedance: 50 or 400 ohms

o HDS-1250 D/A converter

    - 12-bit resolution
    - linearity of $\pm$ 1/2 LSB
    - monotonicity guaranteed (-55 to 125ºC)
    - settling time: 50 nsec
    - output impedance: 200 ohms
    - temperature coefficients
      --linearity: 3 ppm/ºC
      --gain: 30 ppm/ºC
      --bipolar offset: 15 ppm/ºC
    - power dissipation: 0.9 watts
    - size: 3.6 cm^3

o HTS-0025 Track-And-Hold (Deglitcher)

    - acquisition time to 1% for
      1V output step: 20 nsec
    - settling time: 20 nsec
    - slew rate: 400 V/microsec
    - aperture uncertainty: 20 psec
    - droop rate: 0.2mV/microsec
    - linearity: .01%
    - power dissipation: 1.5 watts
    - size: 3.6 cm^3

    A slightly faster 12-bit A/D converter (CAV-1210) is also available from Analog Devices. This device converts at rates up to 10 MHz with 25 psec aperture uncertainty but has larger power consumption (18 watts) and size (287 cm^3) than the MOD-1205. For the initial laboratory investigation we chose to use the MOD-1205.

MULTIPLE-TONE TESTING

In order to determine the utility of the A/D system for digitizing an IF signal, two-tone IM tests were conducted to determine the maximum desired-to-spurious signal ratio that could be obtained within a 100 kHz spectral window centered on the test tones. In most cases the largest spurious signals were third-order IM products. A schematic of the apparatus used to generate clean test-tones is shown in Fig. 5. The 6 dB attenuators on the input and output ports of the power splitter were inserted to provide the isolation necessary to optimize the S/IM3 ratio. Two ANZAC AM-109 high performance amplifiers (10 dB gain, +50 dBm third-order intercept) were used to boost the level of the tones to the vicinity of the maximum drive level of the A/D converter (+16 dBm). Since the tones were separated in frequency by only 30 kHz, a coherent summing of the tones should be assumed, thereby limiting each tone to a level of 10 dBm or less. Since differential, rather than integral, nonlinearity usually limits performance more severely, maximum spurious signal generation may not occur at maximum drive level; however, large signal levels are required to stress the slew rate capabilities of the A/D. Typically, noise floor levels will rise at the higher drive levels.

Since we do not have a digital processing facility to directly evaluate the output of the A/D converter, it was necessary to convert the signals back to analog form using the HDS-1250 D/A and the HTS-0025 deglitcher in order to perform spectral analysis on the HP 8568A. Consequently, the results obtained are characteristic of the composite system rather than the MOD-1205 alone. Ideally, the D/A and deglitcher should have noise and linearity equivalent to at least 2 bits better than the A/D for dynamic testing to be representative of the A/D alone. Presently, this is difficult to achieve at the speeds of interest. Moreover, the test system we have used is representative of that which would be used in an early-development buoy system where digital signal processing would probably not be used.

The results of two-tone testing at an A/D sampling rate of 1.236 MHz are shown in Table 2. The sampling rate was reduced to approximately one fourth of the maximum sampling rate of 5 MHz to simulate conditions corresponding to a four-channel TDM system. The tones at 1.54 and 1.57 MHz are placed in a nearly optimal location (1.25 F_s) according to the guidelines previously discussed. Two-tone testing near 1.25 F_s (as apposed to 0.75 F_s or 0.25 F_s) is desirable because it corresponds to the highest IF frequency that could conceivably be used with this system (because of the onset of aperture jitter effects above 0.78 MHz).

At an input level of 5.8 dBm per tone (the maximum level achievable with the equipment configuation of Fig. 5), the spectrum of the composite input signal was free of noise and spurious signals to -77 dBc (dB relative to the carrier) using a 1 kHz analyzer resolution bandwidth and 3 Hz video filtering. The 28.2 dB loss between input and output levels of the A/D - D/A system corresponds closely to the 12 dB difference in full scale levels of the devices (16.2 dBm for A/D; 4.2 dBm for D/A) plus the 15 dB loss associated with the deglitcher frequency response at a frequency of 1.25 F_s (see Fig. 3). The maximum noise was observed at the highest drive level but spurious signal production was nearly constant over the upper 20 dB of the drive level range.

TABLE 2

Two-Tone Testing at 1.236 MHz Sampling Rate

Tone 1 at 1.54 MHz
Tone 2 at 1.57 MHz
Input clean to - 77dBc (1kHz resolution, 3 Hz video)
100 kHz spectral window centered on tones

INPUT TONE LEVEL (dBm)	OUTPUT TONE LEVEL (dBm)	MAX. SPURIOUS SIGNAL LEVEL (dBm)	OUTPUT NOISE LEVEL (dBm//1 kHz BW)
5.8	-22.4	-90.	-92.0
3.8	-24.4	-90.	-92.5
0.3	-27.9	-90.	-92.5
-5.3	-33.5	-90.	-94.0*
-9.3	-37.5	-90.	-94.0*
-14.2	-42.4	-89.	-94.0*
-19.3	-47.5	-92	-94.0*
-24.5	-52.7	-93	-94.0*
-29.2	-57.4	NO	-98.0
-34.0	-62.2	NO	-98.0
-39.5	-67.7	NO	-98.0

NO: Not observed above noise level

*: Noise floor of spectrum analyzer.
    Reference levels changed for last 3
    measurements to achieve lower analyzer
    noise.

Assuming that the maximum observed spurious product provides the major performance limitation, the system tested can be described by a maximum two-tone, spurious-free dynamic range of 66.6 dB (89-22.4 dB). Theoretically, each tone can be driven another 4.2 dB before exceeding the range of the A/D. This additional drive would increase the dynamic range to 70.8 dB if the spurious signal level remained at 68 dB, but this circumstance has not been demonstrated. Because of the hard-limiting effect of overdriving the A/D converter, it is best to avoid driving the system to the theoretical limit. If a three-tone excitation signal had been used in testing, intermodulation products of the form $F_1+F_2-F_3$, which are normally 6 dB larger than two-tone third-order products, would decrease the observed spurious-free dynamic range by 6 dB. A spectral display of the output of the A/D-D/A system at a drive level of 5.8 dBm is shown in Fig. 6.

Two-tone testing was also conducted at similar tone frequencies but at a sampling rate of 5 MHz to see if there was any significant improvement in performance by sampling in the region of $0.3F_s$. With the exception of changing the sampling rate, all other aspects of the measurement system remained the same. The results are listed in Table 3. The 13.2 dB system loss between input and output is again attributable to the 12 dB difference between full scale levels for the A/D and D/A plus the 1.4dB loss associated with the deglitcher frequency response at $F = 0.31F_s$. The highest system noise occurs at the highest drive level, but the largest spurious signal occurs at an approximately 15 dB lower level -- presumably at a level where the peak signal amplitude is crossing a site of differential nonlinearity. If we characterize the system by a constant spurious signal level of -75.5 dBm, a maximum spurious-free dynamic range of 68.1 dB exists. Thus, changing the region of signal occupancy from $1.25F_s$ to $0.3F_s$ improves performance by no more than 1.5 dB. Spectral displays of the A/D-D/A system output are shown in Figs. 7 and 8 for input drive levels of 5.8 and -9.4 dBm, respectively.

In the VLF-LF band (8-150 kHz), which is of particular interest for the towed communication buoy application, there is greater opportunity for spurious signal generation than exists for the previously investigated 100 kHz band centered at 1.5 MHz. If we consider test tones at F_1 = 20 kHz, F_2 = 45 kHz, and F_3 = 100 kHz, then second-order products such as $2F_1$, $2F_2$, F_1+F_2, F_1+F_3, F_2+F_3, F_3-F_1, F_2-F_1, F_3-F_2 will all fall into the band of interest-- unlike the previous situation. Therefore, conclusions based on the previous analysis cannot necessarily be expected to apply for this band.

The power spectrum of a 3-tone (20 kHz, 45 kHz, 100 kHz) composite waveform used to drive the A/D-D/A system is shown in Fig. 9. The spectrum is clean to -70 dBc. The noise floor of the spectrum analyzer is at the base of the rectangular notch near the right side of the spectrum. The response of the A/D-D/A system at a sampling rate of 1.236 MHz is shown in Fig. 10 using a smaller analysis bandwidth in order to better expose the spurious products. The major product is the $F_3 - F_1$ second-order component which limits the desired-to-spurious signal ratio to about 55 dB.

The existence of additional second- and third-order intermodulation products in the output of the A/D-D/A system is shown in Fig. 11 where the input level of the three tones was increased by 7.3 dB to -18 dBm/tone.

13

TABLE 3

Two-Tone Testing at 5.0 MHz Sampling Rate

Tone 1 at 1.544 MHz
Tone 2 at 1.565 MHz
Input clean to -77 dBc (1 kHz resolution, 3 Hz video)
100 kHz spectral window centered on tones

INPUT TONE LEVEL (dBm)	OUTPUT TONE LEVEL (dBm)	MAX. SPURIOUS SIGNAL LEVEL (dBm)	OUTPUT NOISE LEVEL (dBm//1 kHz BW)
5.8	-7.4	-79.0	-85.0
3.7	-9.5	-82.0	-87.5
0.5	-12.7	-83.0	-88.5
-4.0	-17.2	-76.0	-91.0
-9.4	-22.6	-75.5	-92.0*
-14.0	-27.2	-81.5	-92.0*
-19.2	-32.4	-80.0	-92.0*
-23.7	-36.9	-81.5	-92.0*
-28.9	-42.1	-81.5	-92.0*
-33.5	-46.7	-86.0	-92.0*
-38.7	-51.9	-79.0	-92.0*

*: Values limited by noise floor of spectrum analyzer

At this higher drive level the intermodulation products $F_1 + F_2$, $F_1 + F_3$, $F_2 + F_3$, $F_3 - F_1$, $F_3 - F_2 - F_1$, and $F_3 + F_1 - F_2$ are clearly visible, and the desired-to-spurious signal ratio is reduced to 51 dB by the $F_2 + F_3$ product. Apparently, digital sampling and analog reconstruction of signals in the 8 to 150 kHz band will be limited to desired-to-spurious signal ratios near 55 dB. Since the received signal levels in this band are rarely large, the 55 dB limitation may not be a handicap. In the next section we will examine noise power ratio measurements which are of considerable importance because of the preponderance of atmospheric noise in the VLF-LF band.

NOTCHED NOISE TESTS

In broadband applications, such as the VLF-LF band in which it is common to have a large noise background and a weak SNR, the effect of the broad band noise on the A/D-D/A equipment is of concern. For example, a weak signal that is sitting in a minimum in the noise spectrum can be totally obscured by an equipment induced nonlinear interaction among the broadband noise components. It is possible for noise products formed by the intermodulation process to fill in the spectral noise minimum and diminish the SNR existing prior to the digital sampling/reconstruction process.

The A/D-D/A system was tested for noise intermodulation characteristics by driving it with a broadband noise source having a notch near 100 kHz. The characteristics of the notched noise source are shown in Fig. 12. We would have preferred a response that was flatter in the non-notched region between 10 and 200 kHz and that rolled off more sharply beyond 200 kHz; however, faced with the lack of flat response power amplifiers in this frequency range and a limited time schedule, the noise source characteristic of Fig. 12 was the best that could be achieved. The notch at 100 kHz was produced by filtering after all stages of amplification using a laboratory constructed pi-filter having two series resonant and one parallel resonant circuit. Air-core inductors were wound for the notch filter with AWG #22 Teflon coated wire in order to avoid nonlinear performance which had previously been a problem using large ferrite cores. Attempts to insert a lower power notch filter prior to the final stage of amplification were unsuccessful because of nonlinearity in the amplifier at the power level required to drive the A/D.

Using graphical estimation, the total average noise power contained in the notched noise source was determined to be approximately 7.4 dBm. Since this magnitude represents the average of a random variable with zero mean, the probability that the random variable (voltage) will be within "k" standard deviations (σ) from zero is lower bounded using Chebyshev's inequality [6, p.99]

$$P(|\text{voltage}| \leq k\sigma) \geq 1 - k^{-2} \tag{14}$$

From (14) we conclude that there will be at least an 89% probability that the noise voltage will be within 3 standard deviations from zero. Consequently the A/D converter should be driven with broadband random noise

having an average power of 20 log 3 or (9.54 dB) less than the maximum allowable input level in order to prevent clipping the extremes of the noise voltage fluctuations. Based on a 50 ohm input impedance, the 7.4 dBm noise power corresponds to 0.524 volts rms. The 3σ value is then equal to 1.572 volts which is 2.29 dB below the 2.048 volt maximum drive level of the A/D.

The response of the A/D-D/A system to the notched noise input is shown in Fig. 13. The lower trace shows the power spectrum of the A/D-D/A system noise, and the noise level of the spectrum analyzer is indicated by the base of the rectangular drop-out toward the right side of the lower trace. The fill-up of the noise notch is caused by the proximity of the notch to the noise floor of the A/D-D/A system. An expanded spectral display in the region of the notch is shown in Fig. 14. Here, the level of the notch is approximately what would be anticipated from a summation of noise powers contributed by the system and the notch. Consequently, we conclude that in the interrogated region near 100 kHz, the presence of broadband noise over the region of approximately 20 to 200 kHz has no apparent detrimental effect to the performance of the A/D-D/A system provided the broadband noise voltage has a 3σ value less than the maximum voltage level of the A/D.

Notched noise tests were also conducted in the vicinity of 1 MHz frequency using a sampling rate of 5 MHz. This test was intended to simulate conditions existing in a very noisy IF band. The power spectrum of the notched noise source used to drive the A/D converter is shown in Fig. 15. Most of the noise power is contained within a 1 MHz bandwidth and the total broadband power is approximately 2.42 mW corresponding to a 3σ value of 1.044 volts on the 2.048 volt range of A/D (or -5.9 dB relative to 2.048 V). The depth of the notch is approximately 66 dB.

The power spectrum of the response of the A/D-D/A system to the notched noise source is shown in Fig. 16. Several differences between the input and output spectra are apparent. One difference is that the relative heights of the two maxima have been altered by about 2.5 dB in the output spectrum. Approximately 0.7 dB can be attributed to the frequency response of the deglitcher (see Fig. 3). The remaining 1.8 dB of the 2.5 dB discrepancy is presumably caused by aperture jitter effects which are likely to become noticeable beyond 0.78 MHz (the maxima occur at 0.83 and 1.33 MHz). A second difference between the input and output spectra is that the width of the notch at a given depth below the maxima is larger for the output power spectrum. Nonlinear interactions within a A/D-D/A system would be expected to fill-in the notch (particularly at its lower extreme) which should result in narrowing of the notch, not broadening. Consequently, this effect must be attributable to some other property or combination of properties such as deglitcher frequency response and aperture jitter.

The depth of the notch in the output spectrum is only about 64 dB, but this result is almost entirely attributable to noise contributions rather than nonlinear effects. We support this claim by the following reasoning. Based on a 66 dB notch depth in the source spectrum, the ideal condition

would be to have the output notch noise at -27.2 dBm (output maximum) - 66 dB = -93.2 dBm. The self noise of the A/D-D/A system with a dummy input load is -98.0 dBm at 1.0 MHz. Summing the contributions of self noise and impressed notch noise, we would expect the notch to occur at a level of -92.0 dBm. The observed value of -91.0 dBm is in close agreement.

The output power spectrum is shown in Fig. 17 for a 3 dB increase in source power. The depth of the notch has been reduced to 58 dB relative to the lower frequency maximum, and consideration of noise additions (as done previously) leaves a discrepancy of 6 dB, presumably caused by nonlinear mixing in the A/D-D/A system. The 3σ value of the source noise was approximately 1.47 volts which is 2.9 dB below the maximum input voltage of the A/D.

Finally, in Fig. 18 the output power spectrum is shown for an additional 2 dB increase in source power corresponding to a 3σ value of 1.86 volts (0.85 dB below A/D maximum). The notch depth has been reduced to 48 dB indicating a very large increase in intermodulation noise power.

We conclude from this series of tests that the A/D-D/A system is free of intermodulation noise contamination provided the broadband source noise voltage has a 3σ value (three times rms voltage) that is 5 to 6 dB less than the maximum voltage accepted by the A/D converter.

SUMMARY AND CONCLUSIONS

The use of an A/D converter has been considered for digital sampling of waveforms received from an antenna. Two modes of operation have been discussed. The first mode consists of direct A/D conversion of relatively low frequencies up to 150 kHz; the second mode consists of A/D conversion of a moderate bandwidth IF carrier (at a center frequency near 1.5 MHz) that has been derived from a much higher RF carrier. For the application of interest (towed communication buoy), we assume that the carriers received on the antenna are not demodulated. Instead, the A/D system will digitize the modulated carriers and demodulation will be carried out using the sampled data, either directly with a high-speed digital processor or indirectly by analog methods following carrier reconstruction with a D/A converter.

The effects of quantization resolution (number of bits), sampling speed, aperture jitter, frequency response (particularly that of the deglitcher), and spurious signal generation are considered. The important technical features are summarized as follows.

As a consequence of the sampling theorem, the maximum analog bandwidth that can be sampled digitally without introducing contaminating by aliasing products is $F_s/2$, where F_s is the sampling frequency. If the analog signal has a frequency higher than $F_s/2$, an additional constraint is necessary. If the bandlimited carrier is symmetric about center frequency F_c, the constraint requires that F_c be placed at an odd multiple of $F_s/4$ to prevent aliasing in bandwidth $F_s/2$.

The effect of variability (jitter) in the sampling aperture is to limit the maximum frequency that can be sampled if the quantization error is to be less than one least significant bit (LSB). For a constant aperture jitter condition, the maximum frequency that can be sampled without error is inversely proportional to the number of quantization levels. For the equipment used in our laboratory tests (12-bits A/D resolution, 25 psec aperture jitter), aperture error greater than 1/2 LSB is expected to occur at frequencies greater than 0.78 MHz.

A large dynamic range is required at the IF section of a communication receiver if small signals must be recovered in the presence of inband interference or jamming signals. Consequently, a large number of quantization bits are desired in the A/D converter. The relationship between signal-to-quantization noise ratio was given in (8). However, since A/D conversion speed decreases and the penalty of aperture jitter increases as the number of quantization bits are increased, a compromise must be accepted among the parameters of conversion speed, maximum frequency to be sampled, and dynamic range. The magnitude of the aperture jitter is basically dependent upon the state of the technology and, to some extent, the cost and size of the hardware the user can tolerate. The bandwidth required and the number of TDM channels establishes the required sampling rate. The A/D is then selected with as many quantization bits as will permit achievement of the required speed and aperture-limited maximum frequency.

The frequency response of the A/D conversion process is flat up to a frequency at which aperture jitter effects set in. However, if the analog signal is reconstructed using a D/A converter and deglitcher (sample-and-hold device), the frequency response of the deglitcher follows a $\sin (\pi F/F_s)/(\pi F/F_s)$ behavior when followed by low-pass or band-pass filtering. Consequently, the response vanishes at frequencies that are multiples of F_s and is significantly diminished at the relative maxima occurring at odd multiples of $F_s/2$. For example, at $F = 3F_s/2$ the response is down 13.2 dB. Fortunately, however, the deglitcher frequency response attenuates the quantization noise also, so that the signal-to-quantization noise level is not adversely affected at $F = 3F_s/2$ although it does vanish at multiples of F_s.

Spurious signal production provides the greatest performance limitation in the use of the A/D converter. For a single input signal of frequency F, spurious products of the form $|N_s F_s - NF|$ will be present, where N_s and N are integers. The spurious products corresponding to N = 1, N_s= (any integer) are the well-known, large amplitude, aliasing components that are avoided if the guidelines regarding bandwidth, sampling frequency, and choice of center frequency for passband (IF) waveforms are followed. Higher order terms (N > 1, N_s = any integer) are present at considerably lower amplitudes, but their effect cannot be avoided because of the large numbers of products generated. Using our A/D-D/A deglitcher system, we have observed spurious products up to fourteenth order ($5F_s-9F$). We emphasize that this result describes the performance of the composite A/D-D/A system, not the A/D alone. We had no facility to evaluate the A/D independently.

18

When several discrete frequencies are present at the input to the A/D converter, intermodulation products will be formed at frequencies corresponding to the sums and differences of various combinations of input frequencies. Observation of this phenomenon with the A/D-D/A system indicated that second- and third-order intermodulation products were considerably larger than fourth- and higher-order products.

The performance limitation imposed by spurious signal generation depends upon how the A/D system is to be used. The limitation will usually be less severe for narrow-band applications, such as sampling an IF pass-band, because large-amplitude, second-order intermodulation products usually fall away from the band of interest and will not contribute to distortion. Using our composite A/D-D/A-deglitcher system, we have observed 66.6 dB spurious-free dynamic range in a 100 kHz band centered near 1.55 MHz using a sampling frequency of 1.236 MHz. Since frequencies above 0.78 MHz are expected to be affected by aperture jitter, a few dB of the 66.6 dB dynamic range limitation may be caused by this effect. The observed spurious-free dynamic range would be 6 dB smaller for three-tone testing in which intermodulation products are formed by interactions among three input signals.

For wider band applications in which second-order intermodulation products cannot be avoided, the spurious-free dynamic range will be smaller. Measurements over a 10 to 200 kHz band using our A/D-D/A-deglitcher system with a three-tone input yielded only 55 dB spurious-free dynamic range at a sampling rate of 5 MHz.

Although the spurious-free dynamic ranges observed for the A/D-D/A system are not as large as are generally desired for the digital sampling of modulated communications waveforms, the A/D conversion process deserves serious consideration for the towed buoy application because the digital samples can be transferred over the tow cable to the final stages of the receiver without any further deterioration in signal quality. Because of the high bit-rate required (approximately 60 Mbit/sec for a four-channel TDM system), fiber-optic transmission lines are required. Fiber-optics are well suited to digital modulation applications of this nature, and fiber-optic tow cables have reached a satisfactory state of development.

The effect of broadband noise on the performance of the A/D-D/A system was found to be negligible provided the 3σ value of the broadband noise voltage waveform (three times the rms voltage) was at least 5 to 6 dB below the maximum level accepted by the A/D converter.
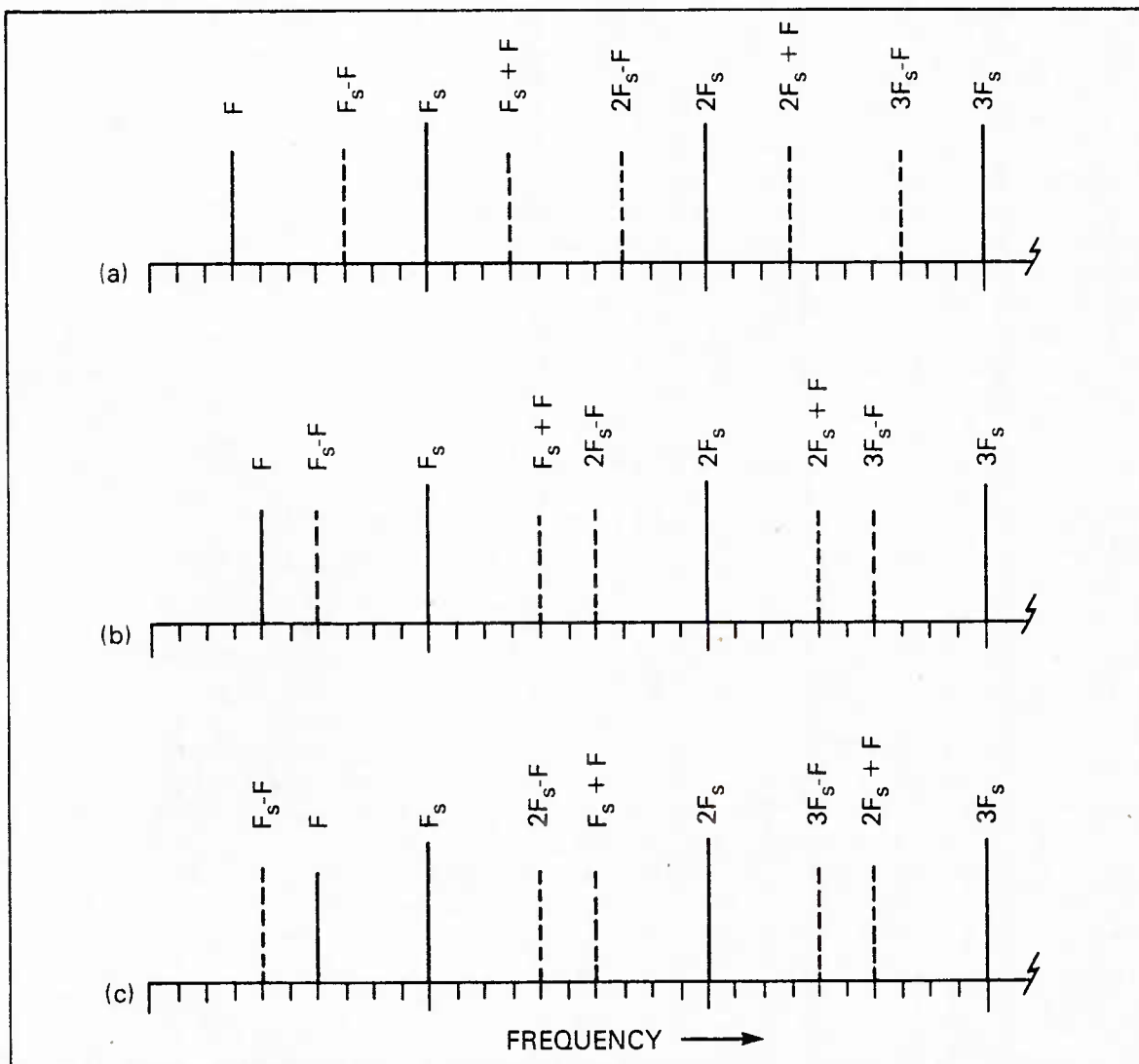
Figure 1

The frequency spectrum of a CW tone of frequency F when sampled periodically at rate F_s. The spectrum includes harmonics of F_s and aliasing components formed at frequencies of $|N_s F_s \pm F|$, where N_s assumes integer values. The illustrations show the repositioning of the aliasing products as F increases from below the Nyquist frequency (a,b) to above it (c).
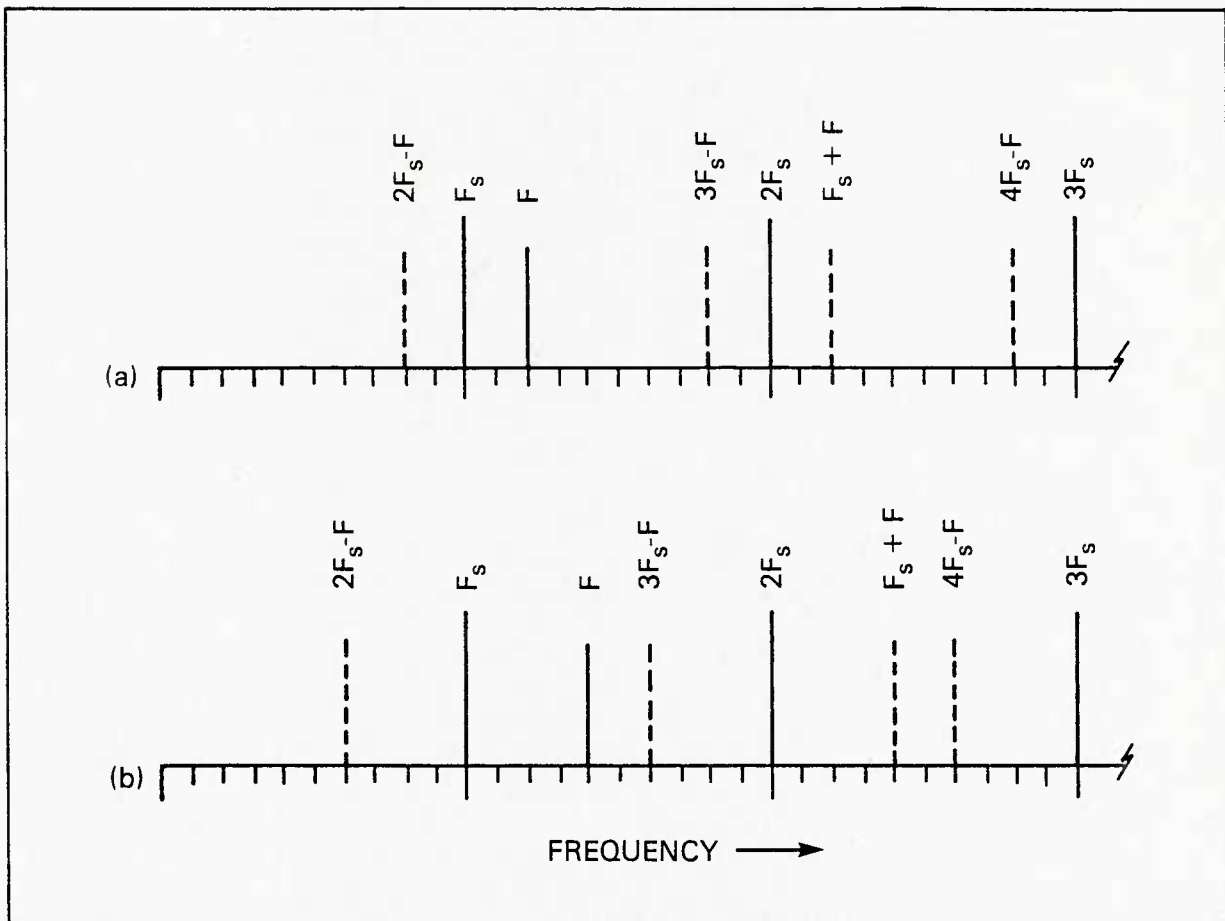
Figure 2

The frequency spectrum of a CW tone of frequency F when sampled periodically at rate F_S and when $F_S < F < 3F_S/2$. The illustrations demonstrate that to achieve an alias-free bandwidth of $F_S/2$, the center frequency of the band must be placed at an odd multiple of $F_S/4$.
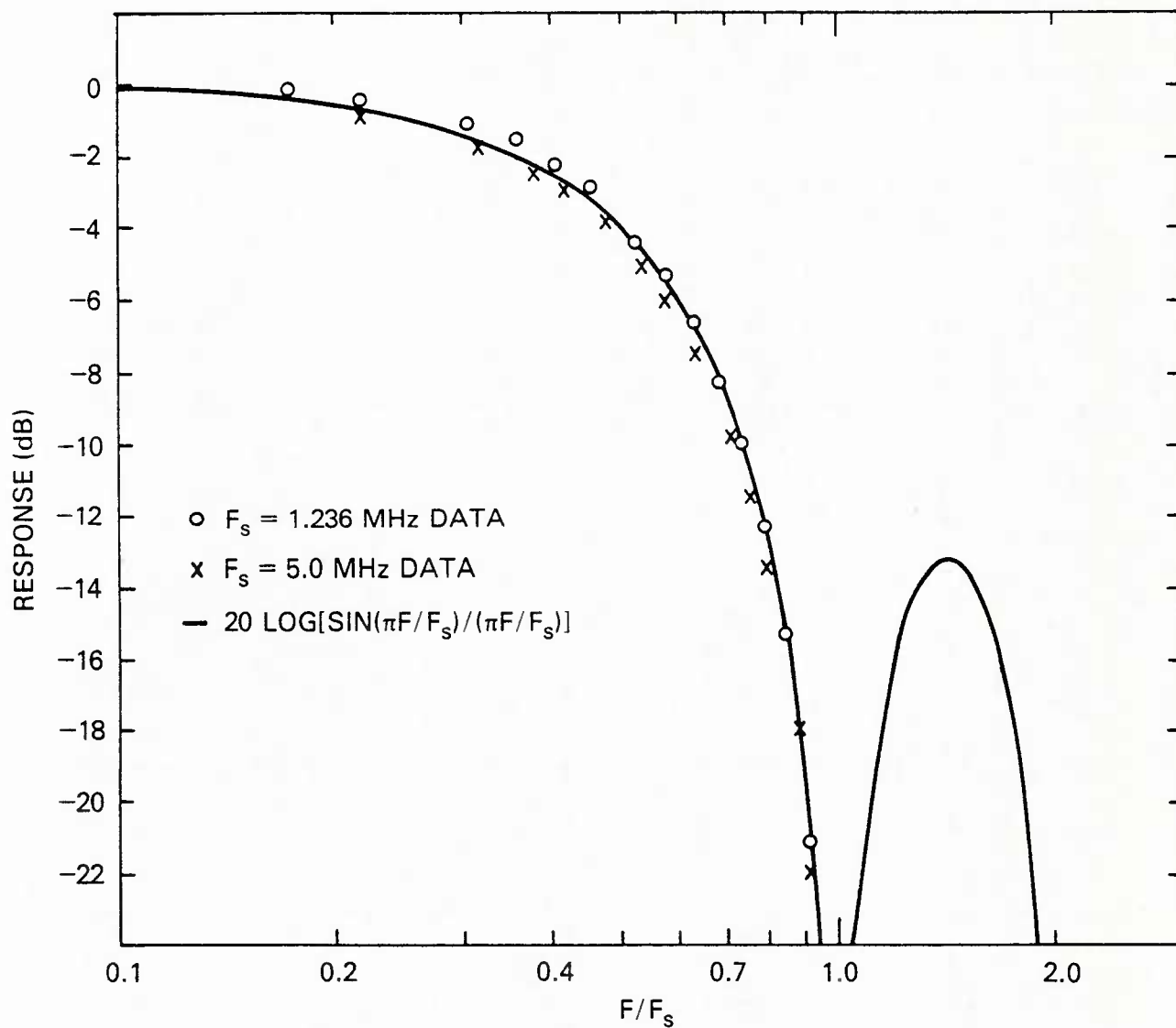
Figure 3

The frequency response of a sample-and-hold device used as a deglitcher
for a D/A converter. The solid line is the theoretical
$\sin(\pi F/F_s)/(\pi F/F_s)$ response; the data points are measurements
of the overall response of the A/D-D/A-deglitcher system.
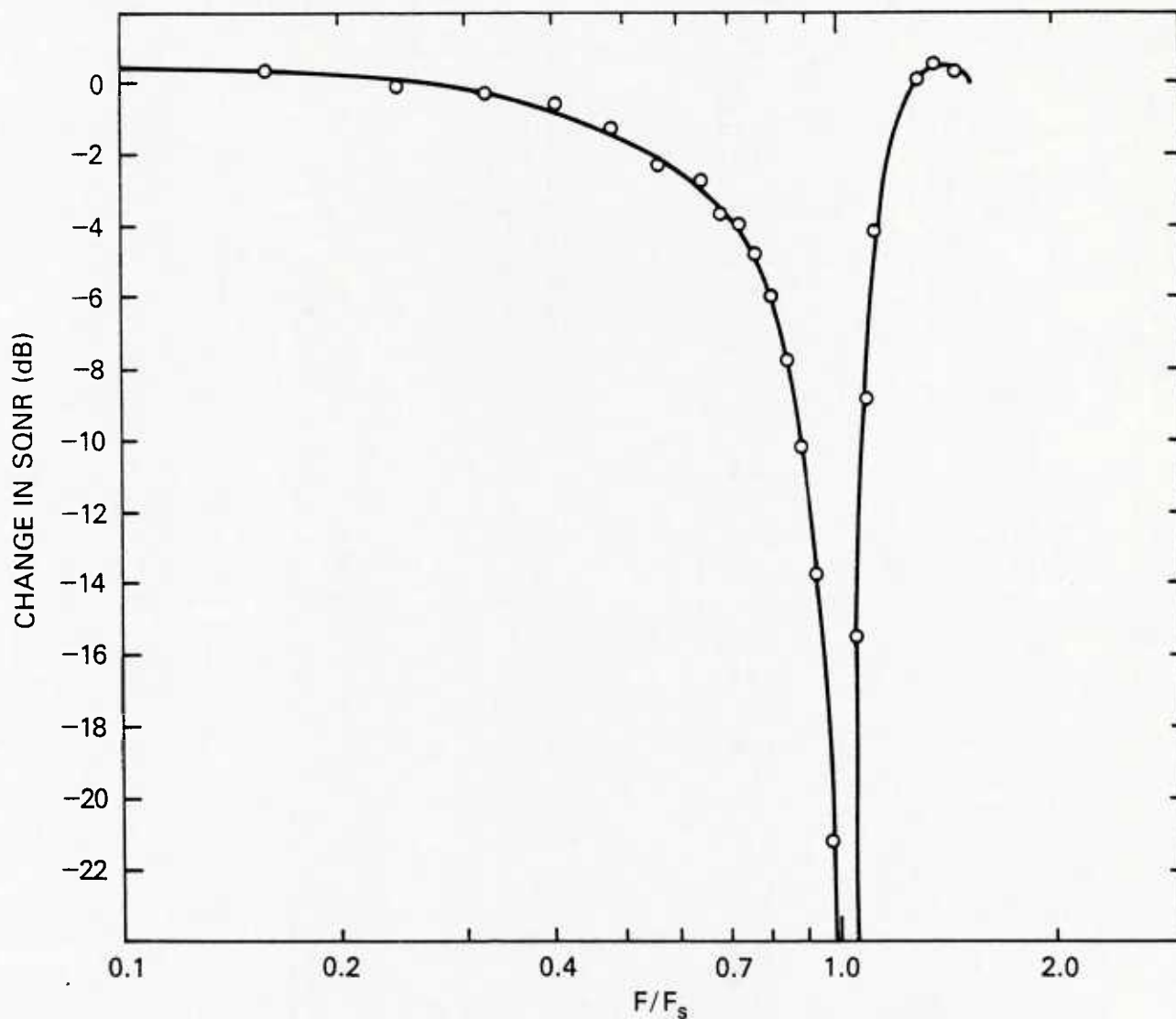
Figure 4

Frequency response of the signal-to-quantization noise ratio. Since the frequency response of the deglitcher affects quantization noise as well as signal, the signal-to-quantization noise ratio is degraded only in frequency regions near multiples of F_s.
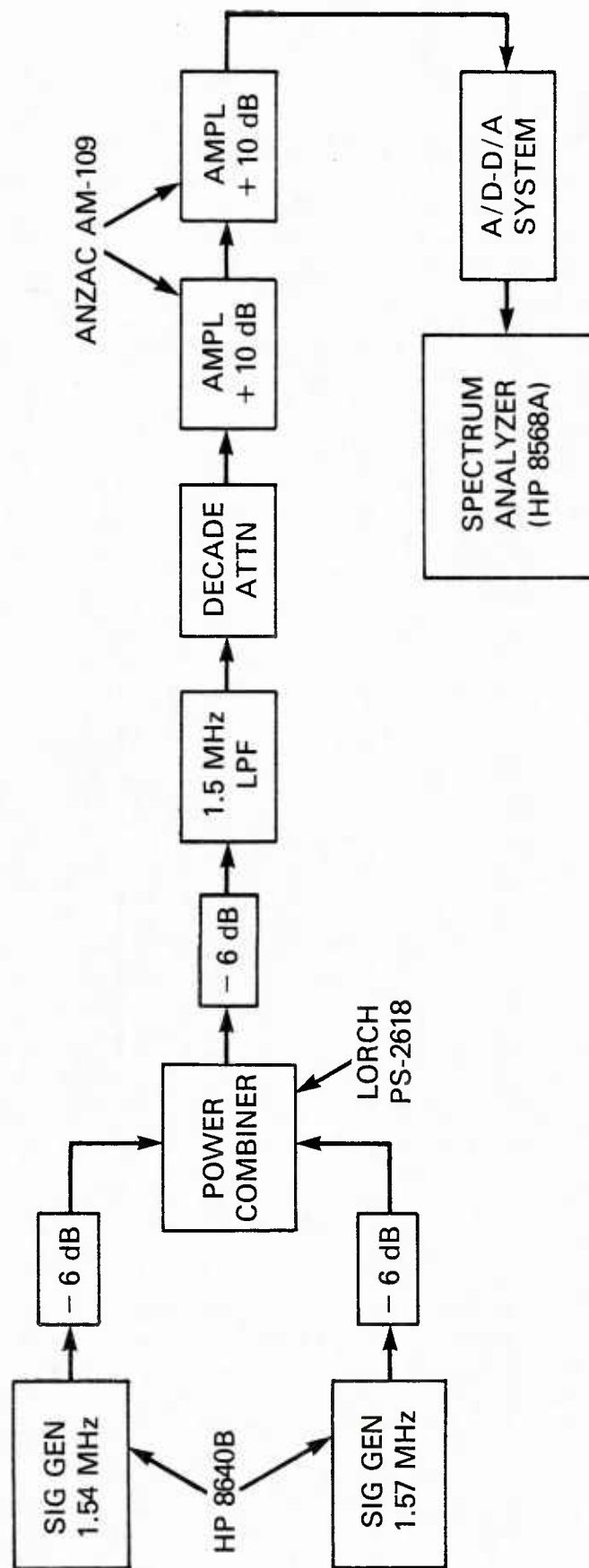
Figure 5

Apparatus used for two-tone testing of the A/D-D/A system near 1.5 MHz frequency.

24
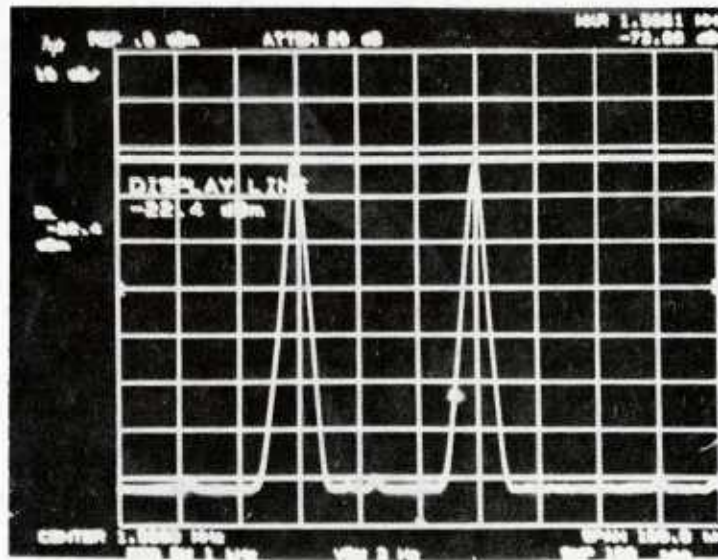
Figure 6

An exemplary spectral display of the output of the A/D-D/A system at a drive level of 5.8 dBm per tone and a sample rate of 1.236 MHz.
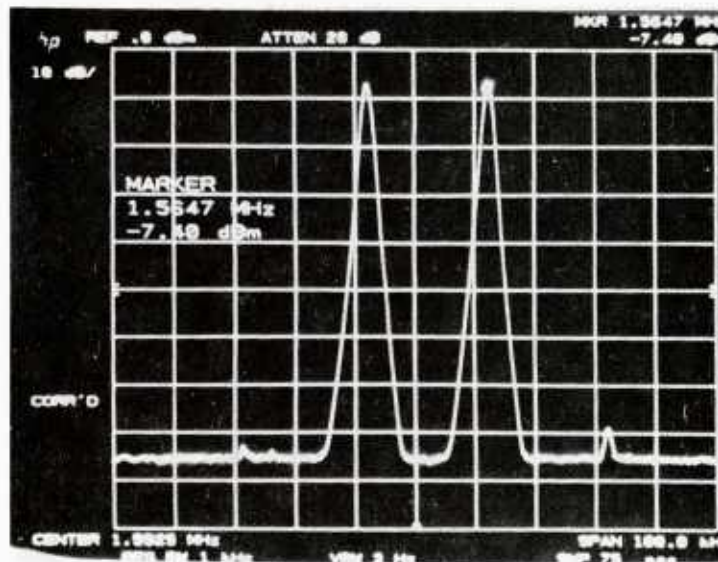
Figure 7

Spectral display of the output of the A/D-D/A system at a drive level of 5.8 dBm per tone and a sample rate of 5.0 MHz.
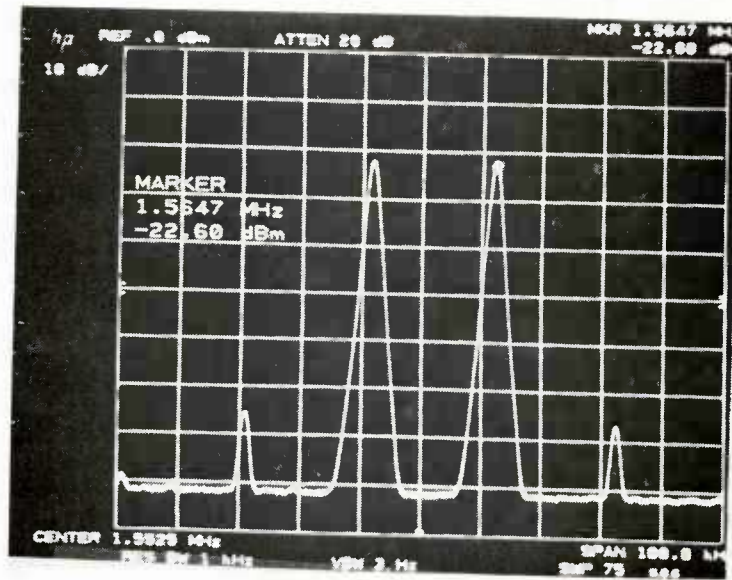
Figure 8

Spectral display of the output of the A/D-D/A system at a drive level of
−9.4 dBm per tone and a sample rate of 5.0 MHz. Note that the
distortion is worse than at the higher drive level of 5.8 dBm (Fig. 7),
indicating the differential nature of the nonlinearity.
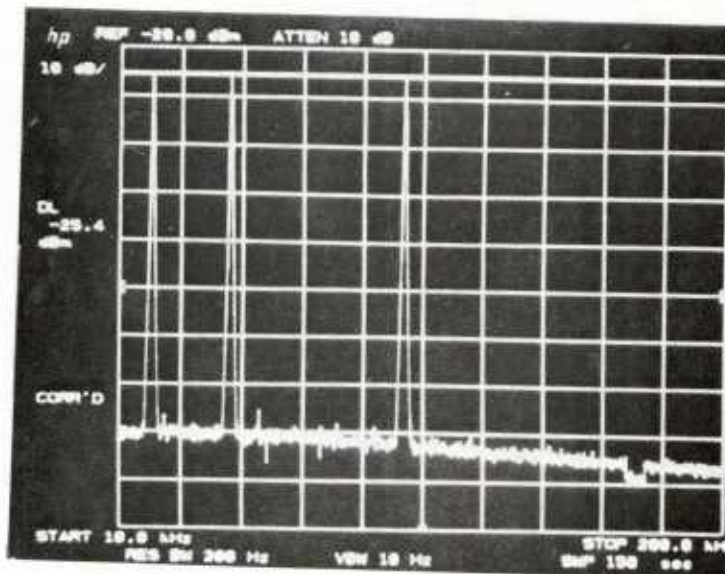
Figure 9

Power spectrum of a three-tone (20 kHz, 45 kHz, 100 kHz) composite
waveform used to drive the A/D-D/A system. The spectrum is clean to
−70 dBc.

26
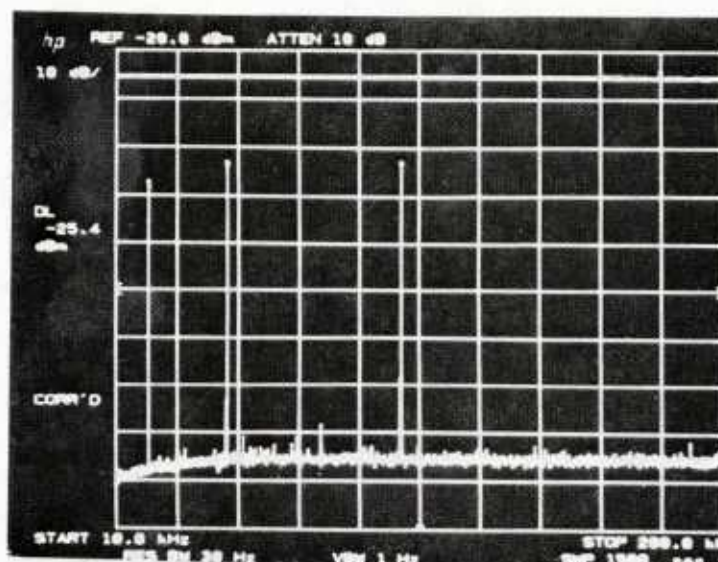
Figure 10

Response of the A/D-D/A system to the source waveform of Fig. 9 at a sampling rate of 1.236 MHz.
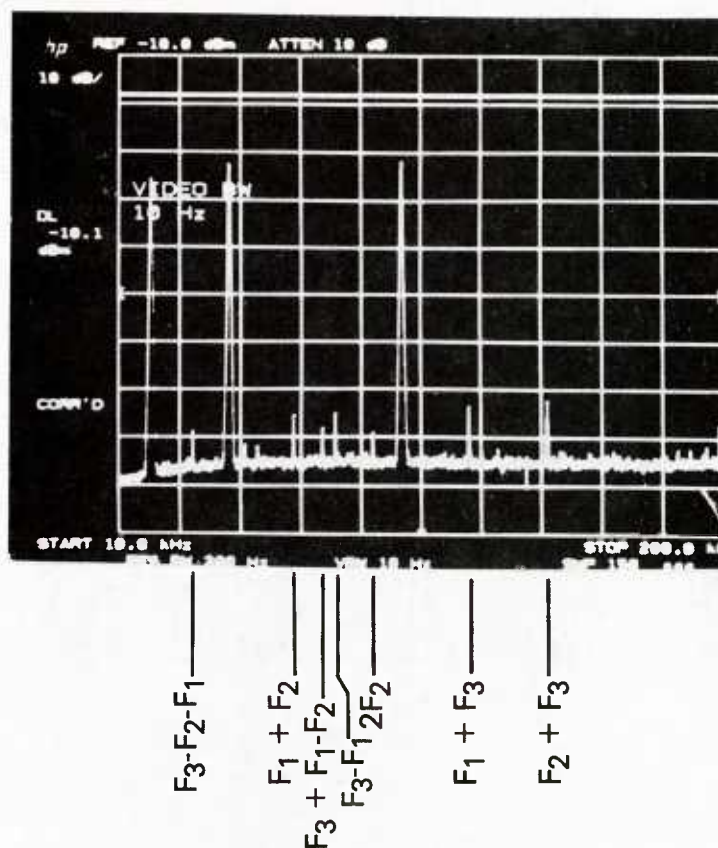
Figure 11

Response of the A/D-D/A system to the source waveform of Fig. 9 at 7.3 dB higher drive level. Second- and third-order intermodulation products are easily identified.
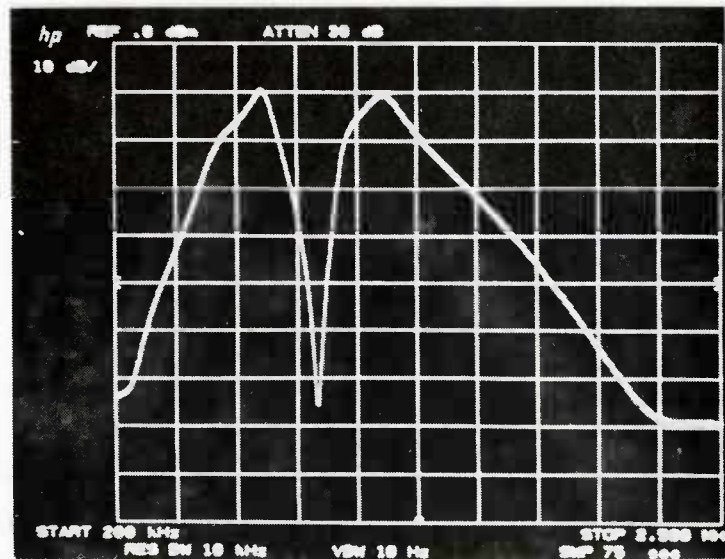
27
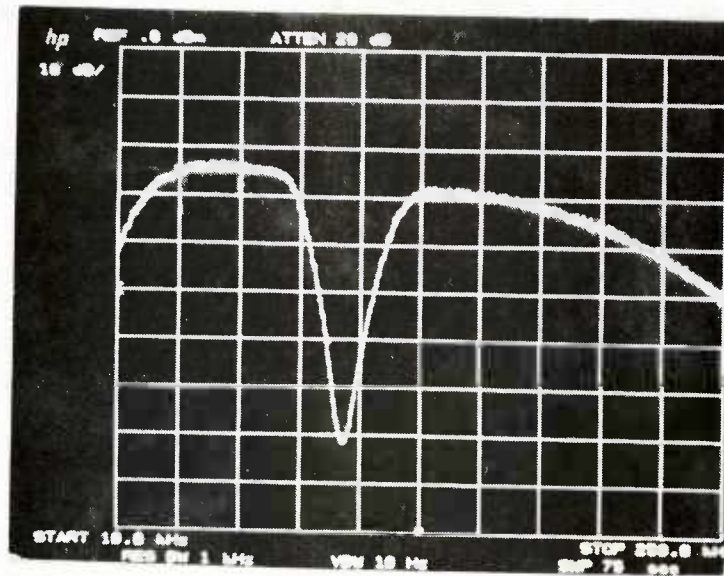
Figure 12

Spectral characteristics of the noise source, notched at 100 kHz frequency, used to drive the A/D-D/A system.
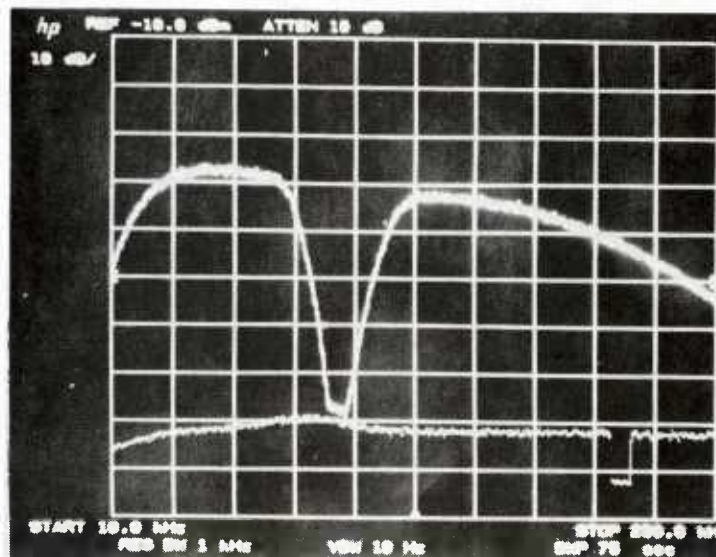
Figure 13

Response of the A/D-D/A system to the notched noise input at a sampling rate of 1.236 MHz (upper trace). Lower trace is the noise spectrum of the A/D-D/A system with no external input. The noise level of the spectrum analyzer is indicated by the rectangular drop-out toward the right side of the lower trace.
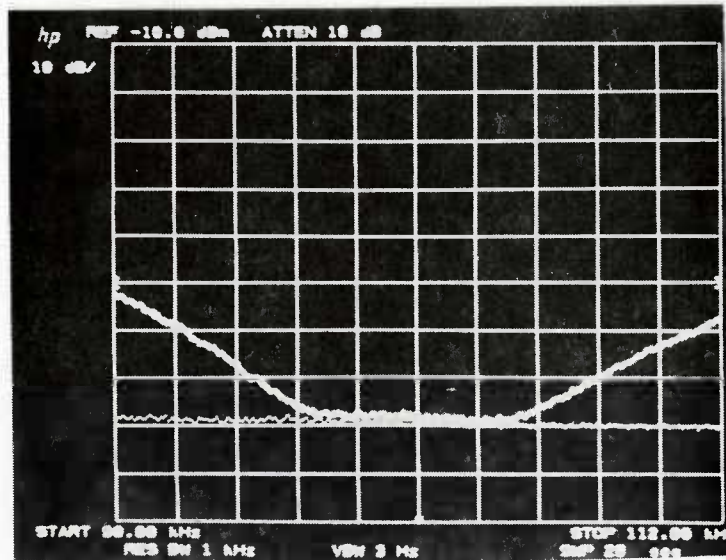
Figure 14

Expanded spectral display of the output of the A/D-D/A system near the notch. The notch depth is limited by system noise rather than by intermodulation noise.

Figure 15

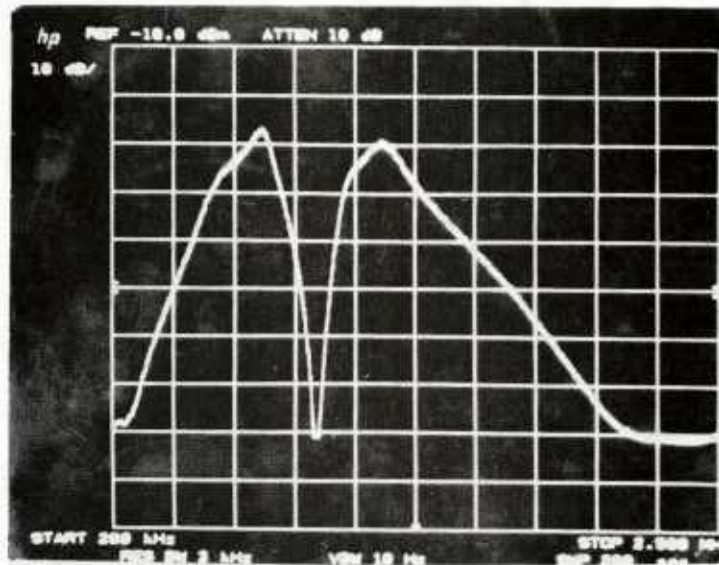Spectral characteristics of the noise source, notched at 1 MHz frequency, used to drive the A/D-D/A system.

Figure 16

Response of the A/D-D/A system to the notched noise input at a sampling
rate of 5 MHz. Depth of the notch is limited by system noise rather
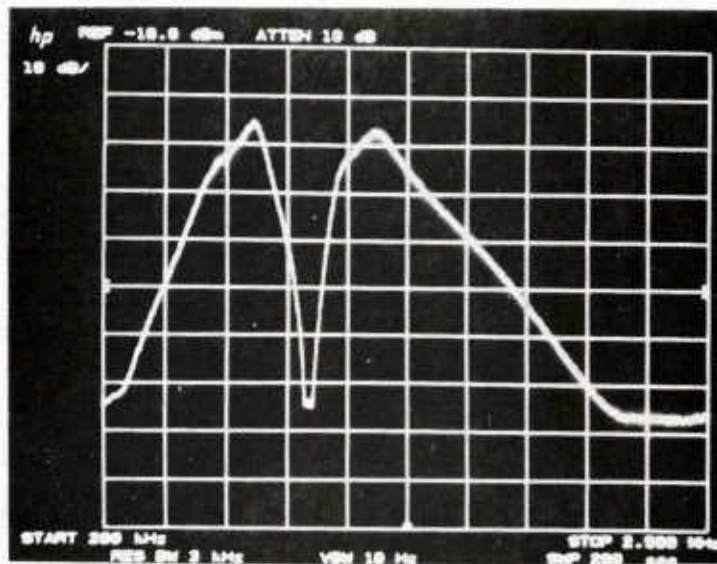than nonlinear effects.

Figure 17

Response of the A/D-D/A system to the notched noise input for a 3 dB
increase in source power. Approximately 6 dB loss of notch depth is
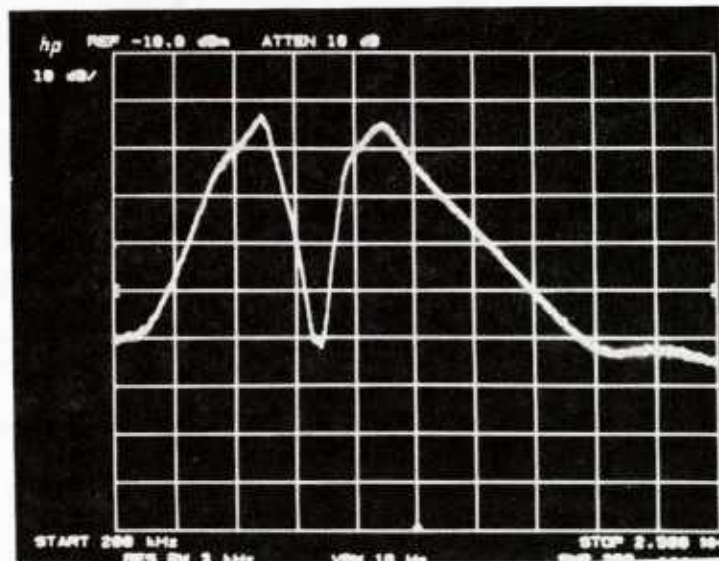attributed to intermodulation noise.

Figure 18

Response of the A/D-D/A system for a 5 dB increase in source power. The notch depth has been reduced to 48 dB indicating a very large increase in intermodulation noise.

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