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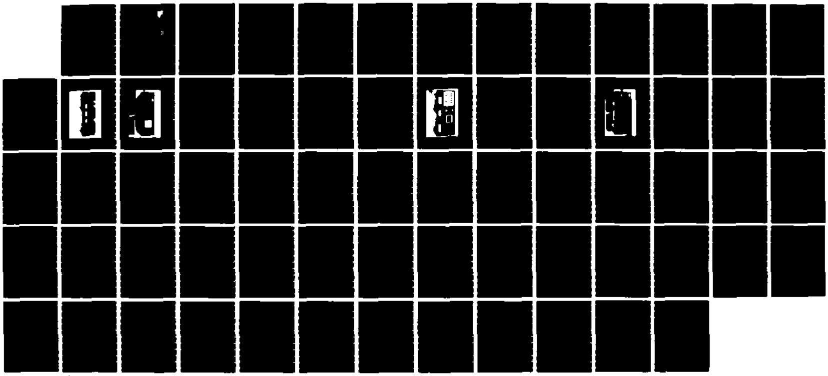

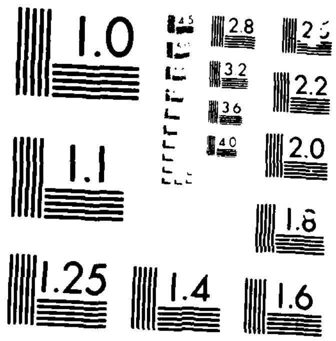
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RADC-TR-86-22
Final Technical Report
March 1986

PASSIVE SURVEILLANCE DATA PROCESSOR/ANALYZER

Syracuse Research Corporation

Dr. Samuel Craig

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ROME AIR DEVELOPMENT CENTER
Air Force Systems Command
Griffiss Air Force Base, NY 13441-5700

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This Final Report is a compendium of the work done over the last two years on Contract F30602-83-C-0090, and it fulfills Data Item [illegible] of that contract. This project involved the design and fabrication of a flexible computer-controlled microwave receiver and associated signal data recording system, given the name Passive Surveillance Data Processor/Analyzer. The system can gather real signal data from a variety of sources and record the digitized signal data for later playback and analysis in a general-purpose computing facility. The receiver can be operated either in a laboratory environment or on an airborne platform.

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SECTION 1

INTRODUCTION

This Final Report is a compendium of the work done over the last two years on Contract No. F30602-83-C-0090, and it fulfills Data Item A010 of that contract. This project involved the design and fabrication of a flexible computer-controlled microwave receiver and associated signal data recording system, given the name Passive Surveillance Data Processor/Analyzer. The system can gather real signal data from a variety of sources and record the digitized signal data for later playback and analysis in a general-purpose computing facility. The receiver can be operated either in a laboratory environment or on an airborne platform.

A number of supporting documents were prepared under this contract, the most significant being the Design Plan, the Test Plan, the User's Manual, and the Program Maintenance Manual. This report extracts material from those documents to provide the basis for the system's design philosophy and the manner in which the design was implemented in hardware. The following sections justify the system as a research tool, discuss the major issues that influenced the design, and describe the system hardware in detail. Some suggestions for follow-on work are also given.

SECTION 2

BACKGROUND INFORMATION

Since its inception, the use of radar has been a key factor in the collecting of tactical information. Its value as a surveillance tool is beyond question. Radar technology has been pressed continually to achieve more accurate detection of smaller and faster targets at greater ranges, while minimizing the probability that its own transmissions can be detected and utilized by the other side. At the same time, the implementation of increasingly complex countermeasures presents an even more hostile operating environment to the radar.

The advancement of radar technology is aided greatly by a flexible, wideband signal analysis capability for the experimental testing of new antenna concepts and signal processing techniques. However, the analysis of wideband RF signals presents two generic problems to the investigator. First, it can be very inconvenient to move the signal analysis equipment out into the field, where the signals are. This problem becomes more acute when signal analysis is to be conducted on an aircraft. Second, it is often useful or necessary to run the signal analysis equipment at slower than real-time data rates. Both problems are solved by a general-purpose receiver coupled with a wideband data recorder.

In response to this need, Rome Air Development Center (RADC) conceived the Passive Surveillance Data Processor/Analyzer (PSDP/A). The PSDP/A combines a very wideband receiver with a high-density digital tape recorder in a compact, transportable package. From a control standpoint, the system is extremely flexible, permitting a wide choice of operating bandwidth, center frequency, and dynamic range. This equipment will become a key element in the testing and development of new antennas and passive radar techniques. To maximize its utility as a development tool, it was designed with flexibility and adaptability as primary goals, along with dependability and ease of use. The high cost of airborne testing dictates emphasis on securing as much reliable data as possible from each mission. The intent of the PSDP/A system is to provide a high-quality instrument to serve this function.

SECTION 3

DESCRIPTION OF PSDP/A SYSTEM

3.1 GENERAL

The PSDP/A consists of five separate parts: a tuner, a controller, a spectrum monitor, a tape recorder, and a data entry terminal. Figure 3-1 shows the components of the system.

The tuner unit and the controller unit make up the surveillance receiver. The tuner was built as a separate unit so that it can be placed near the antenna in cases where transmission line losses are to be minimized. The controller is capable of remotely setting all of the tuner's operating parameters, so it can be placed conveniently for the operator. The received signals are carried from the tuner to the controller/digitizer at an intermediate frequency (IF) of 540 MHz, where line losses are more tolerable and do not impair the sensitivity of the receiver. If desired, the tuner can be disconnected from the remainder of the system and used independently.

The controller is the heart of the system. It has a microprocessor-based front panel where the operator can tune the receiver, select its bandwidth, and set the tuner's input attenuator. These parameters can be recorded on the tape, along with a text header prior to the digitized signal data. The front panel provides an indication of the received signal level to aid in setting the input attenuator. It also has a sub-panel for remote control of the tape transport and includes an indicator for the remaining tape footage.

In addition to its front panel functions, the controller includes that portion of the receiver that converts the IF to a quadrature baseband pair, samples the two signals simultaneously, and digitizes the samples. The system bandwidth is determined by a post-sampling digital lowpass filter that also reduces the sample rate, keeping a constant ratio of sample rate to bandwidth. The controller establishes the tape speed, maintaining a constant data

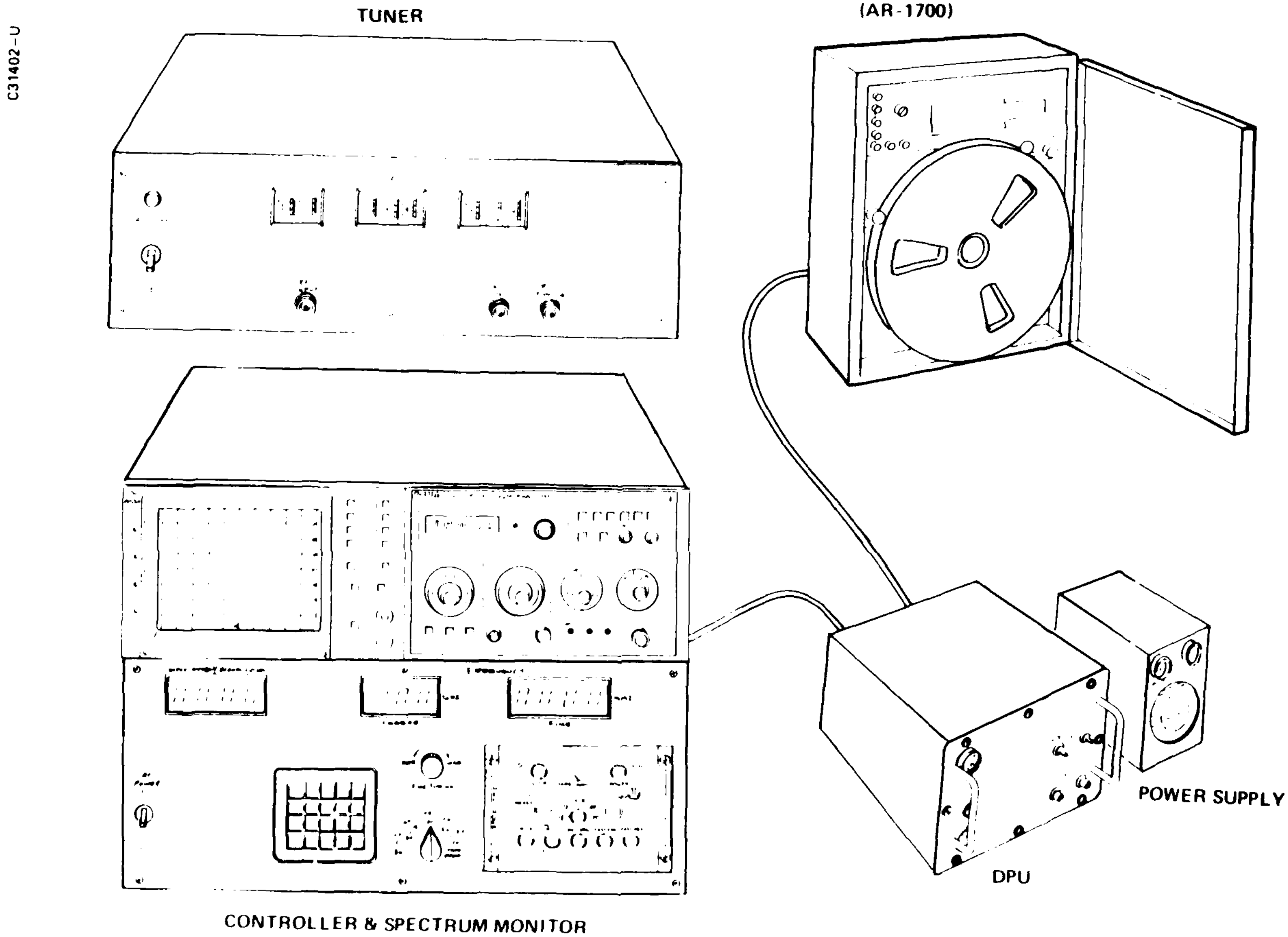

Figure 3-1. Components of the Passive Surveillance Data Processor/Analyzer

density on the tape so recording time is maximized for the bandwidth selected.

A Hewlett Packard spectrum analyzer was furnished with the system for use as a tuning aid and spectrum monitor, as well as a piece of dedicated test equipment. The spectrum analyzer can examine the IF bandpass of the receiver with a marker at band center, or it can check the input to the tuner to locate strong signals in the radio frequency (RF) spectrum.

The fourth major part is the Ampex AHBR-1700 tape recorder subsystem, which includes a tape transport, a digital process unit (DPU), a DPU power supply, and reproduce monitor test unit. A 28V dc power supply was furnished by the contractor for the tape transport. The tape transport and its DPU are standard Ampex products, suitable for airborne use. The tape recorder was purchased in a 28-track, record-only configuration with error-correcting code (ECC). The reproduce monitor unit can be switched to any single track to ensure that data exists on the tape. The DPU does not process the data in the signal-processing sense, but merely puts it in the proper format to allow the 24 synchronous data tracks to be de-skewed on playback.

The Radio Shack TRS-80 Model 100 portable computer was the most cost-effective device found for use as a data entry terminal. It has text-editing and communications utilities that can be used to create a header for the tape recording and to transfer it to the controller unit.

3.2 THE TUNER UNIT

The tuner is the "front end" of the PSDP/A receiver. It accepts signals in the range of 500 MHz to 6 GHz and converts them to a common IF of 540 MHz. It can be controlled locally by means of thumbwheel switches on its front panel, although its normal mode of operation is by remote control originating in the controller unit of the system. The tuner is designed to have good sensitivity, very wide instantaneous dynamic range, and flexibility of control.

Figure 3-2 shows the front panel of the tuner. Most of the items in view are self-explanatory. The two type-N jacks at the lower right connect to the controller unit. Figure 3-3 shows a rear view. The square opening at the left is a honeycomb air inlet filter that maintains the RF shielding provided by the welded aluminum case. The serial data control connector is centered under the rear access cover. The connector at the lower right accepts a standard ac power cord. The loop of semi-rigid coaxial cable at the left side makes the tuner usable with an external high-purity synthesized local oscillator.

A block diagram of the RF and IF portions of the tuner is given in Figure 3-4. The input signal is routed either to one of two preselector/preamplifier branches or to the RF monitor jack on the front panel. Selection is made by a three-position coaxial switch. The preselector in each branch is a three-section YIG-tuned filter. Two separate branches are used because the YIG-tuned filters cannot cover a frequency range of more than two octaves. One branch tunes from 500 MHz to 2 GHz, and the other from 2 to 6 GHz. The use of separate channels for these two ranges also results in better preamplifier performance than would be realized by a single amplifier covering the entire range of the tuner.

A first IF of 2680 MHz is used for the low-band branch, and 1600 MHz for the high-band branch. These values allow a single second local oscillator at 2140 MHz to convert each branch to a common second IF of 540 MHz. They also provide good image rejection and freedom from spurious responses caused by intermodulation products in the mixers.

The first local oscillator is a YIG-tuned device followed by a tracking YIG-tuned filter to reduce the harmonic content of the oscillator output. The first local oscillator tunes on the high side of the signal in the low-band branch. It also tunes on the high side over the lower half of the high-band branch, and on the low side on the upper half. Table 3-1 shows the relationships between these frequencies. This choice was made to minimize the tuning range of the YIG-tuned oscillator so that a device having rela-

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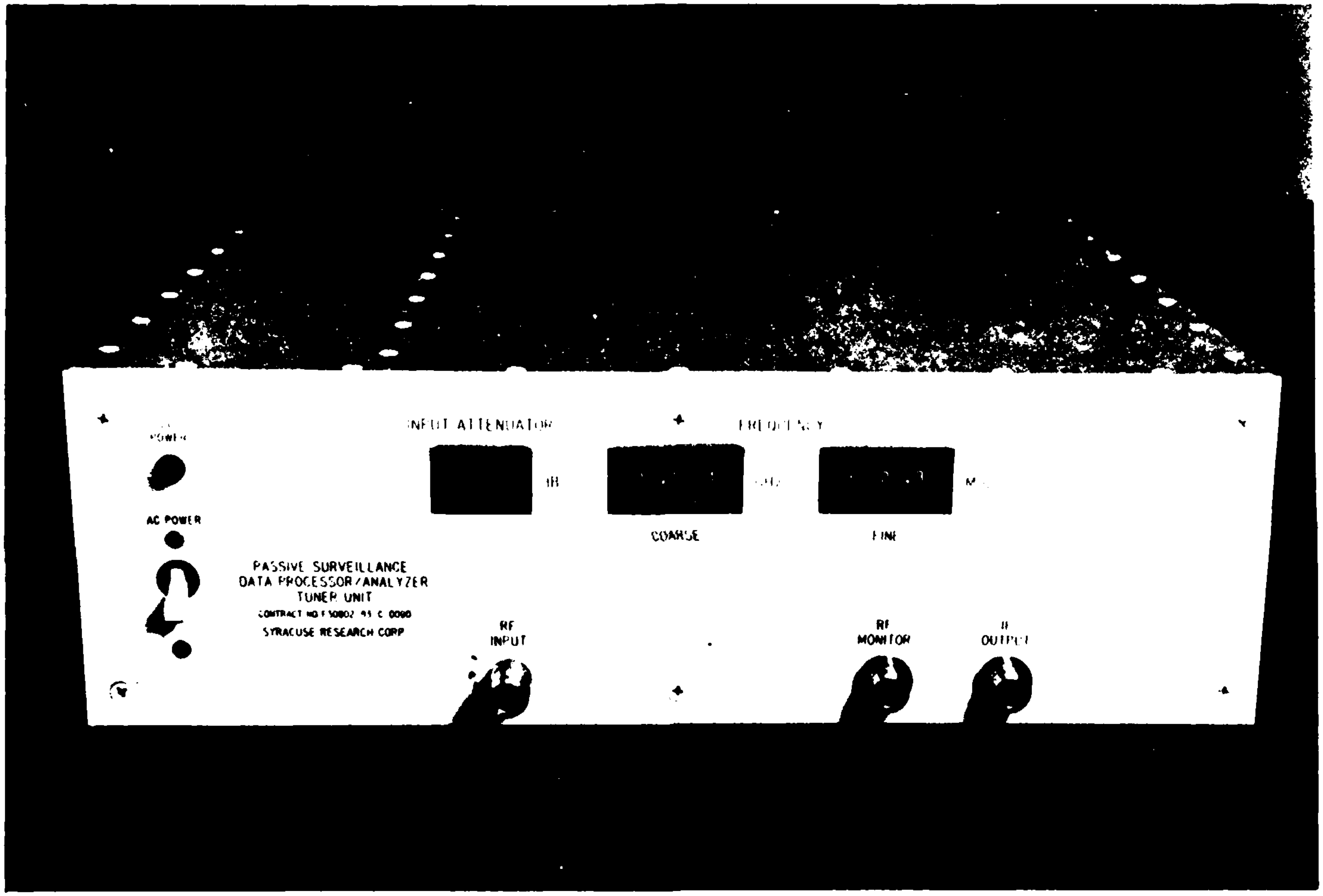

Figure 3-2. Front View of the PSDP/A Tuner Unit

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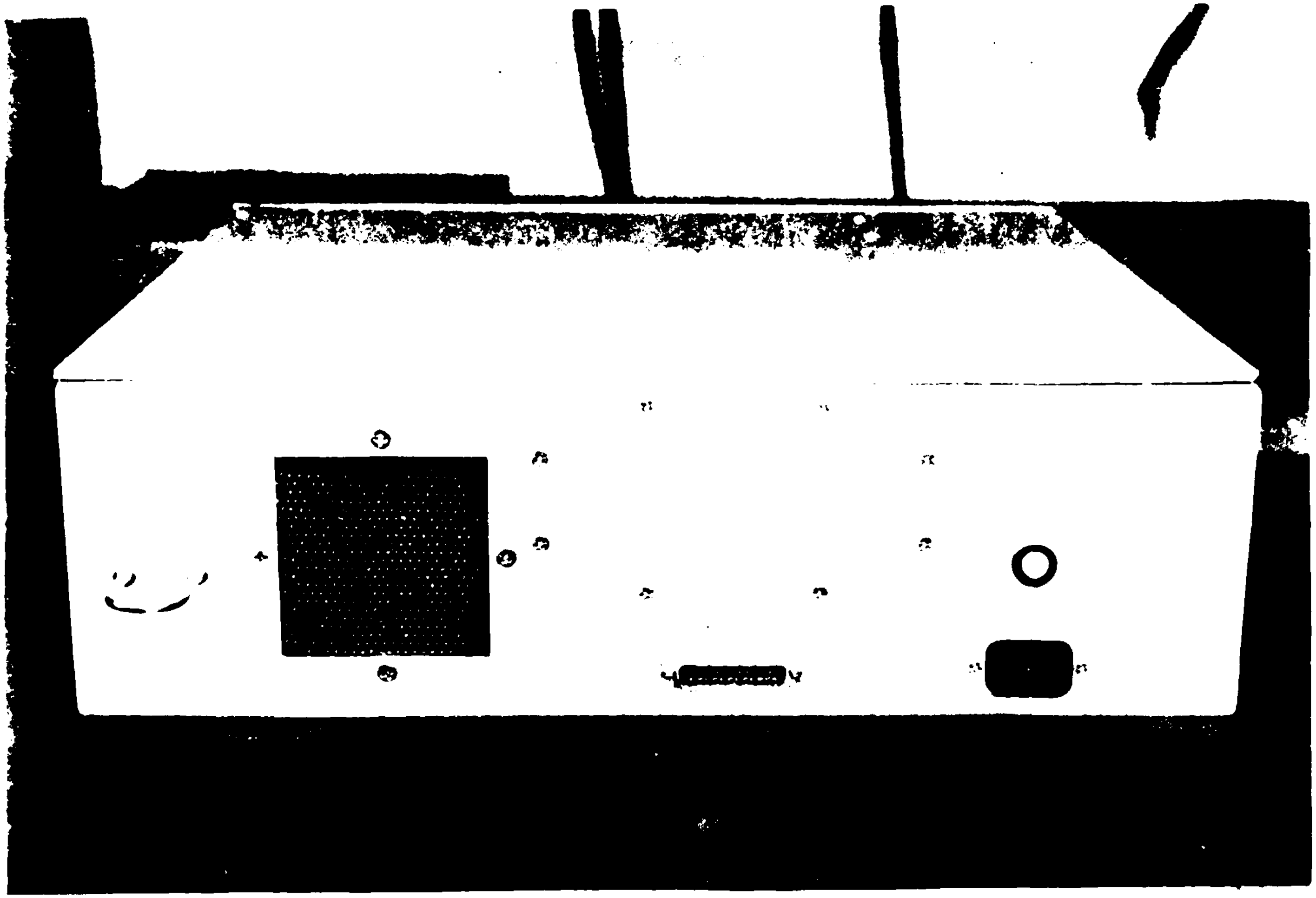

Figure 3-3. Rear View of the PSDP/A Tuner Unit

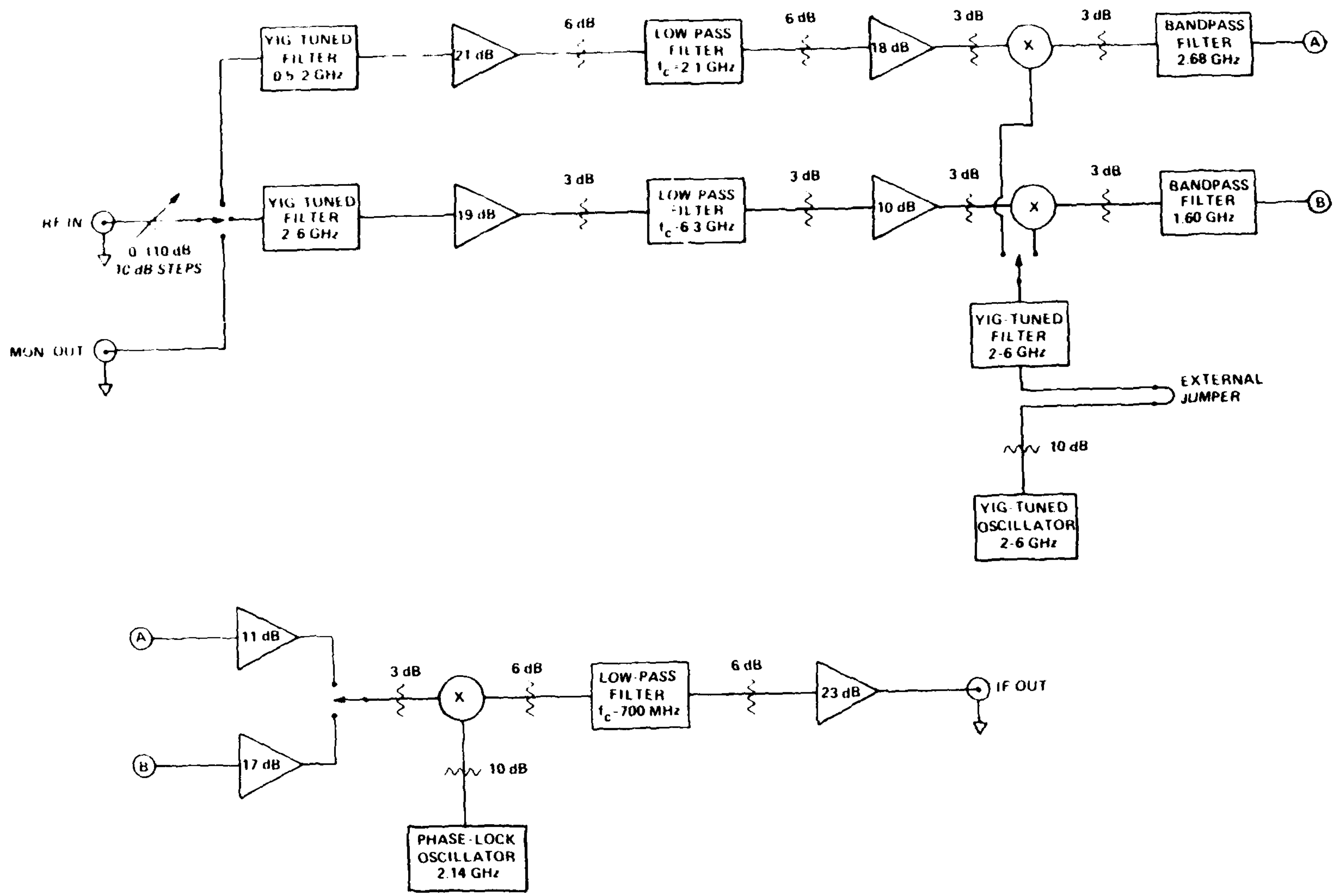

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Figure 3-4. Block Diagram of the RF and IF Portions of the Tuner Unit tively low noise could be used.

Table 3-1. Relationships between Signal and Local Oscillator Frequencies in PSDP/A Tuner

Signal Frequency Range (GHz)	First Local Oscillator Range (GHz)	First Intermediate Frequency (GHz)
0 - 0.5 - 1.99	3.18 - 4.67	2.68
2.00 - 3.99	3.60 - 5.59	1.60
4.00 - 6.00	2.40 - 4.40	1.60

Fixed attenuator pads are dispersed throughout the signal chain to optimize the distribution of gains and losses so that the dynamic range of the tuner is maximized, as well as to improve the impedance match seen by the mixers, amplifiers, and filters. Better matching improves the filter responses and the intermodulation performance of the mixers.

A single-board microcomputer tunes the YIG filters and oscillator, as well as controlling the input attenuator and coaxial switches. The YIG devices are tuned directly by analog voltages that are produced by D/A converters on a separate circuit board. The computer also scans and reads the thumbwheel switches in the local mode and handles the serial communication with the controller in the remote mode. Figure 3-5 shows this arrangement.

The tuner's integral power supplies are located in a separate isolated compartment that occupies the left side of the case. An internal wall separates this compartment from the main part of the case. All leads passing through this wall are filtered, and the cooling air drawn into the power supply compartment passes through a honeycomb filter into the main compartment. The intent of this design is to protect the tuner itself, as well as external devices, from RF interference (RFI) caused by the power

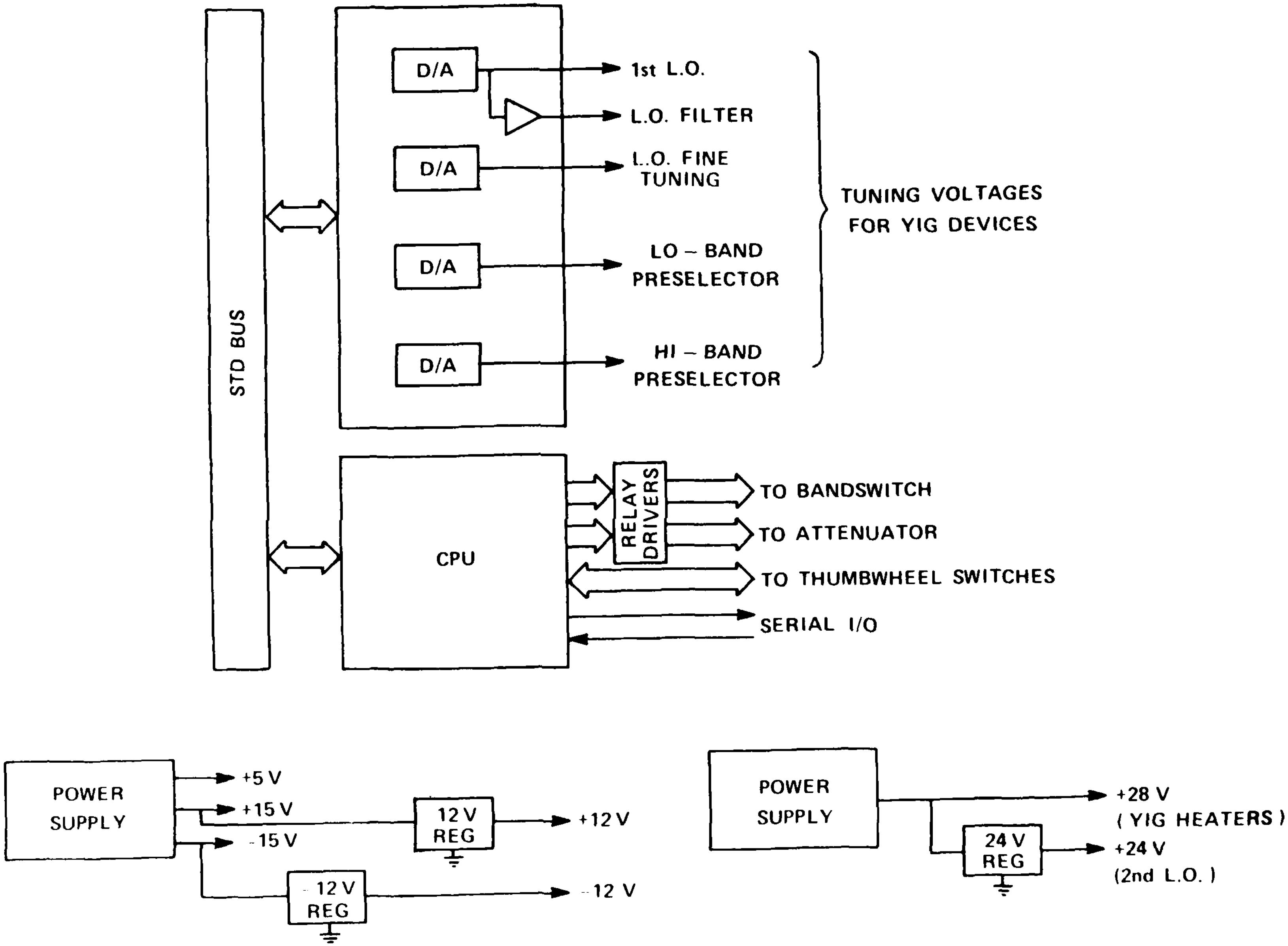

Figure 3-5. Block Diagram of Control Subsystem in PSDP/A Tuner

supplies.

3.3 THE CONTROLLER UNIT

The controller accepts the IF output from the tuner, translates it to baseband, and samples and digitizes the baseband signal. The digitized signal samples are sent to the tape recorder subsystem, either directly or through the digital post-sampling lowpass filter. The controller is also the operator's workstation, with a keypad for entering tuning commands and digital readouts that indicate the status of the tuner. The front panel of the controller includes a remote control unit for the tape transport. Figure 3-6 shows a front view of the controller.

Figure 3-7 is a functional block diagram of the controller. Like the tuner, it includes a single-board microcomputer, which forms the interface between the operator's controls and the functions of the system. The computer scans and interprets the keypad and the fine tuning control, which is a rotary shaft encoder. It also drives the digital LED displays, takes care of the serial data communications, responds to interrupts from error conditions, and (if one has been entered) formats a text header for the tape recording.

Figure 3-8 shows a rear view of the controller with its top cover removed. The two serial data connectors are mounted on a removable cover that allows access to the microcomputer and digital filter circuit boards. The connector on the left (in this view) is for the portable data entry computer, a Radio Shack TRS-80 Model 100. The other data connector is the control interface to the tuner. The two circular connectors at the right side mate with cables to the tape recorder, with the top one carrying the signal data and the one beneath it carrying the remote tape transport control signals. The two coaxial connectors accept the IF output and the input sample signals from the tuner.

The baseband converter coherently translates the IF output from the tuner to a center frequency of zero to minimize the bandwidth prior to A/D

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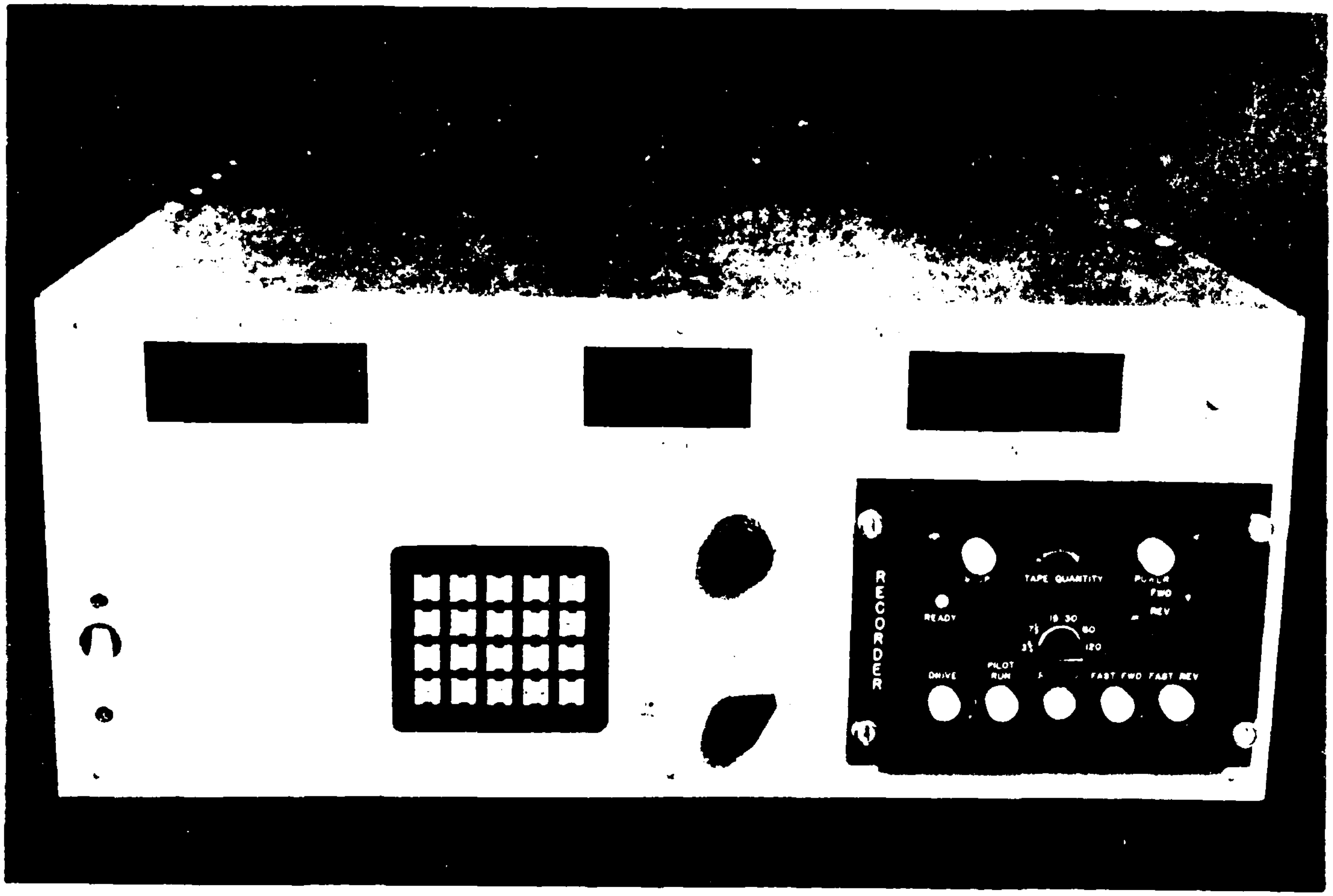

Figure 3-6. Front View of the PSDP/A Controller Unit

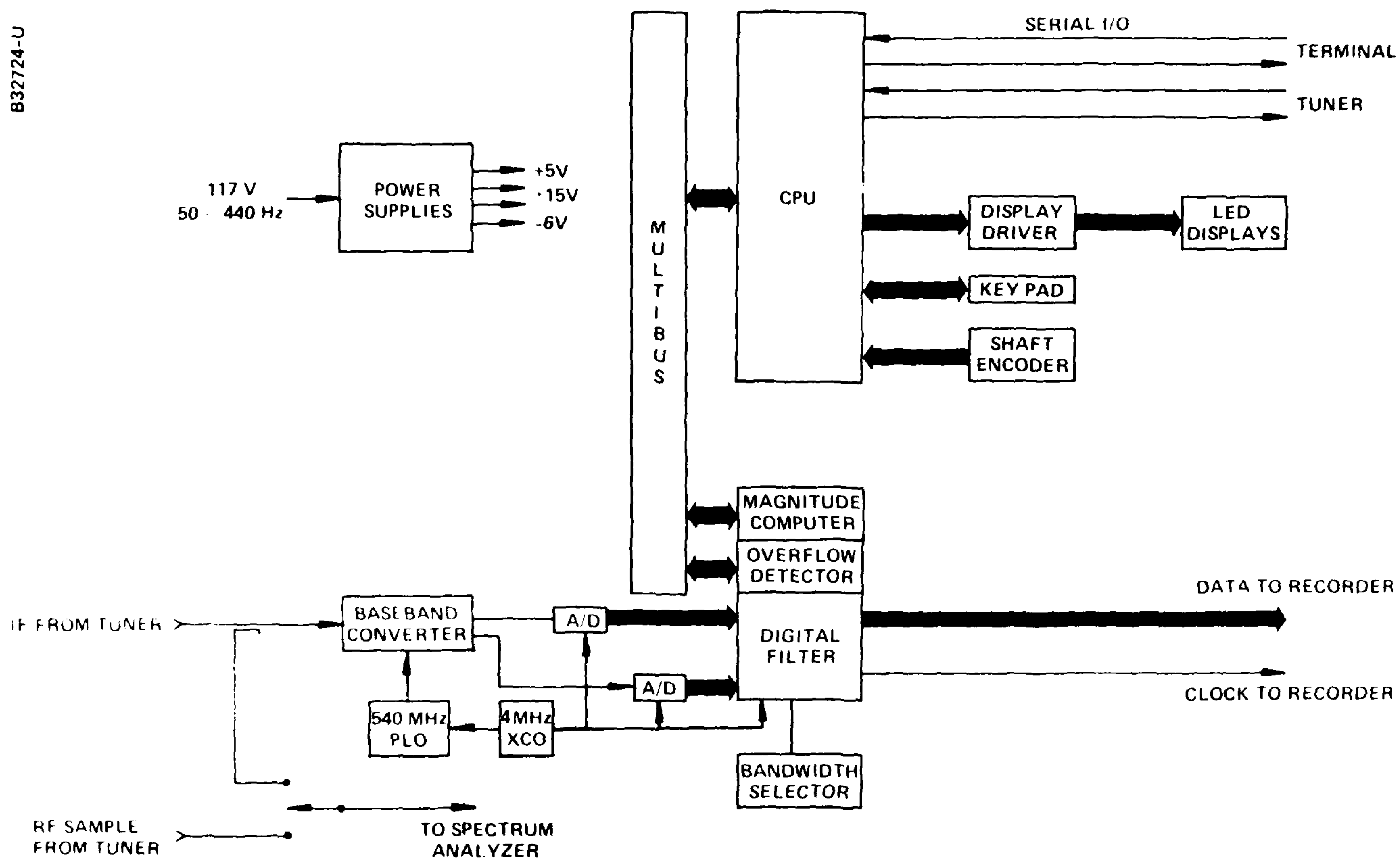

Figure 3-7. Block Diagram of PSDP/A Controller

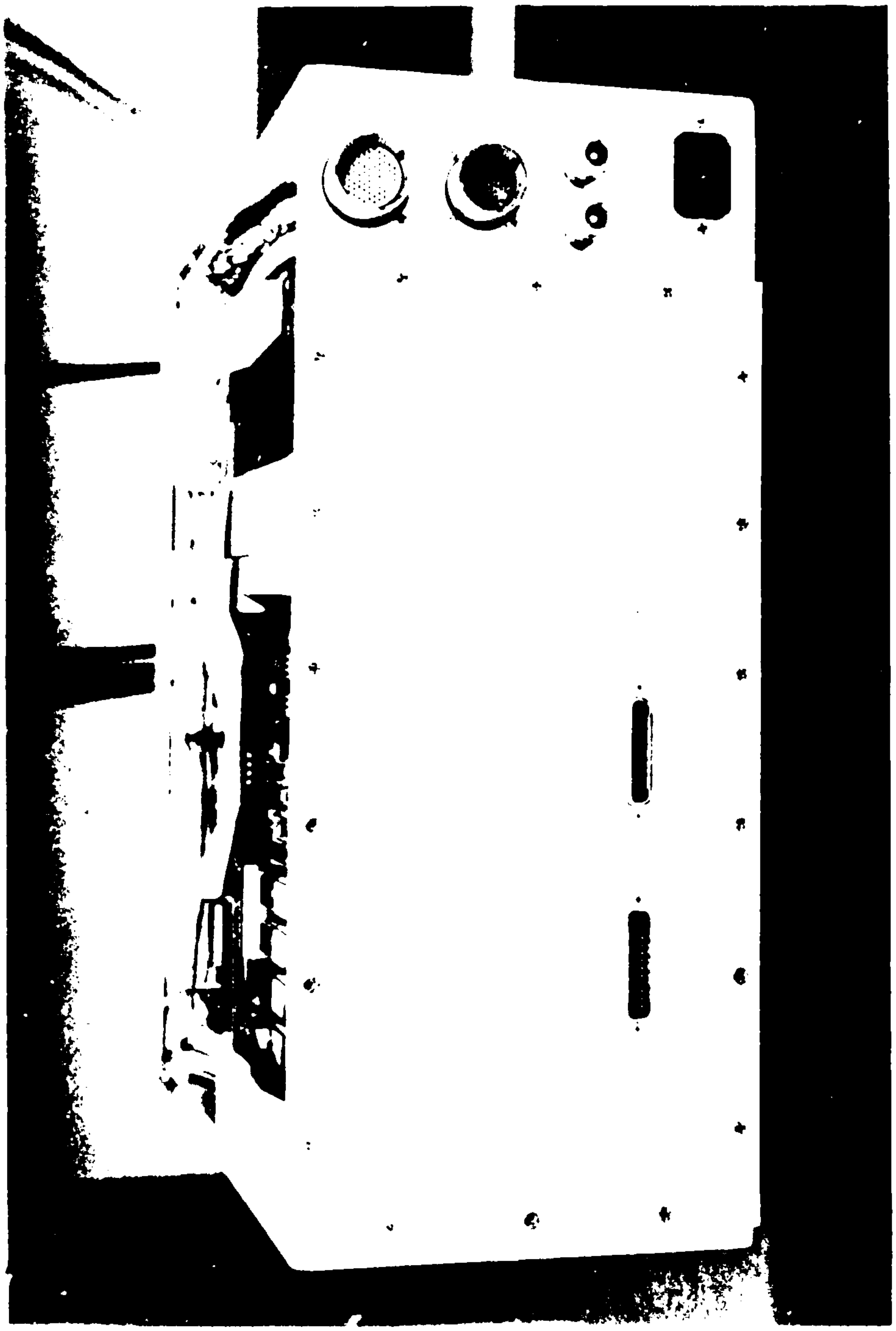

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Figure 3-8. Rear View of the PSDP/A Controller Unit conversion. To preserve the relative phase of the signal, it is split into two branches in phase quadrature, as Figure 3-9 shows. These two branches are matched as closely as possible and are sampled simultaneously for conversion to digital form. Each pair of simultaneous samples can be treated as a complex number when the signals are processed. Each branch is sampled at a 4 MHz rate and digitized with a resolution of 12 bits. The pairs of 12-bit samples are recorded on the tape as 24 bits in parallel. The 540 MHz local oscillator that is fed to the baseband converter is phase-locked to the 4 MHz signal sampling clock.

Figure 3-10 is a functional block diagram of the digital lowpass filter. Its function is to reduce both the effective bandwidth and the output data rate of the system. Although Figure 3-10 shows four separate filters in cascade, the implementation is that of a single filter, through which the data can be passed as many as four successive times. The filter is time-shared between the I and Q data.

Each pass through the filter reduces the bandwidth and data rate by a factor of two. Recirculating the data through the filter on a time-shared basis also provides bandwidth and data rate reductions of either 1/4, 1/8, or 1/16. The final bandwidth is selected by a five-position rotary switch on the front panel. The data output clock, which determines the speed of the tape transport, changes in proportion to the selected bandwidth to maintain a constant recording density. The relationships between the bandwidth, data rate, tape speed, and total available recording time are shown in Table 3-2.

Table 3-2. Record Time and Receiver Bandwidth versus Tape Speed

TAPE SPEED	RECORD TIME*	SAMPLE RATE	RF BANDWIDTH
120 ips	15 minutes	4.00 MS/s	3.2 MHz
60 ips	30 minutes	2.00 MS/s	1.6 MHz
30 ips	60 minutes	1.00 MS/s	800 kHz
15 ips	2 hours	0.50 MS/s	400 kHz
7.5 ips	4 hours	0.25 MS/s	200 kHz

*Assumes 14" Reel Size, Holding 9200' Of 1" Tape

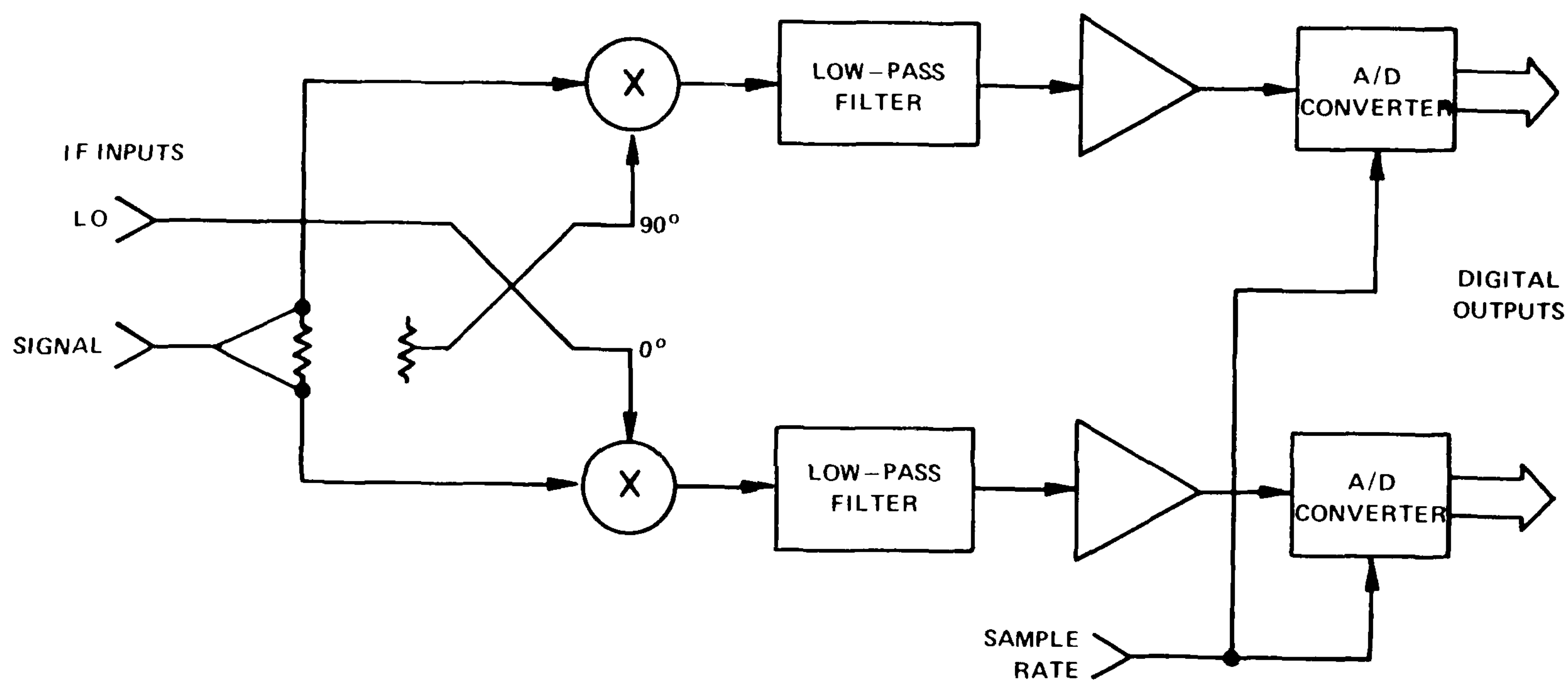

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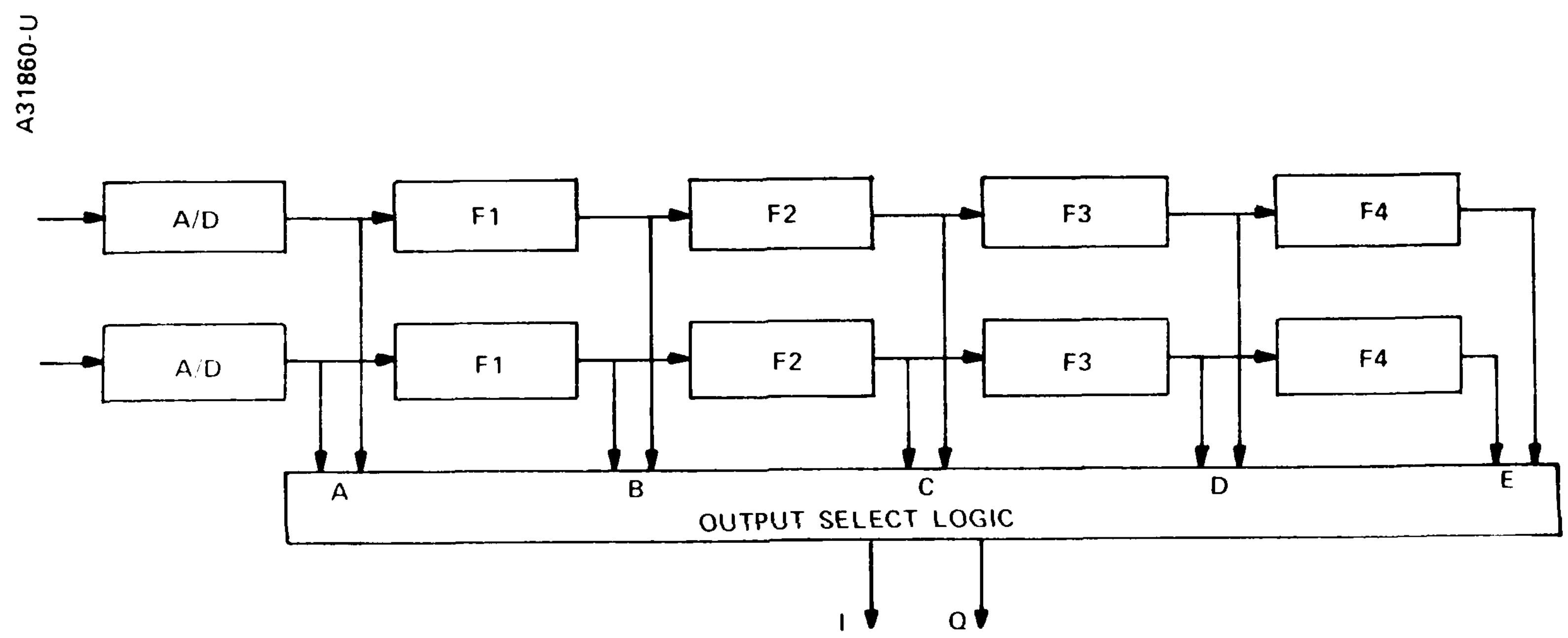
Figure 3-9. Block Diagram of Coherent Baseband Converter



OUTPUT	BANDWIDTH	SAMPLING RATE
A	3.2 MHz	4 MHz
B	1.6	2
C	0.8	1
D	0.4	0.5
E	0.2	0.25

Figure 3-10. Functional Block Diagram of Digital Low-Pass Filter

Saturation of the A/D converters or arithmetic overflow in the digital filter causes an interrupt to the microcomputer, which then displays an error code to the operator. An approximation to the absolute magnitude of each signal sample at the filter output is formed by a part of the digital filter control logic, and the largest value of the magnitude in a one-second period is made available to the microcomputer. It converts this value to dB, scales it appropriately, and displays it as the peak received signal level.

Like the tuner, the controller has its own integral power supplies. However, in this case, they are not physically isolated from the remainder of the circuitry.

3.4 THE SPECTRUM ANALYZER

The spectrum analyzer furnished with the PSDP/A system is a Hewlett Packard Model 8559A, in the 853A display unit. This instrument covers a frequency range of 10 MHz to 21 GHz, considerably in excess of that required to support the system. This analyzer was selected because of its high quality and ease of use. Its primary use in the system is to monitor the IF passband to give a visual indication of the signals being recorded. It can also be used to examine the spectrum of the input to the receiver. In this mode, the spectrum is not limited by any of the receiver's filtering.

3.5 THE TAPE RECORDER SYSTEM

The Ampex AR1700 tape recorder system consists of several separate units, all of which are designed for airborne applications. The largest of these is the tape transport itself, which accommodates a 14-inch reel holding 9200 feet of one-inch tape. The tape recorder system also includes a DPU, a DPU power supply, a reproduce/monitor/test unit, and a power supply for the tape transport, which nominally requires 28V dc. All of these items were made by Ampex, except the tape transport power supply, which was furnished by the contractor.

3.6 THE PORTABLE DATA ENTRY COMPUTER

This system accessory is a Radio Shack TRS-80 Model 100 portable computer, which is a self-contained item that can be used separately if desired. Although most Radio Shack products are for the consumer market, the Model 100 was the most cost-effective unit that could be found for use as a portable data entry terminal. Its included software and provisions for interfacing make it a very useful addition to the system. Its primary purpose is for the entry, storage, and uploading of text files to be recorded as headers on the signal data tape. It can also be used to control the tuner unit and to set its time/date clock. The Model 100 also has its own time/date clock, but it has no relation to the one in the tuner.

SECTION 4

DESIGN PHILOSOPHY

4.1 GENERAL

The design and construction of a receiver having at once a wide tuning range, wide bandwidth, and high instantaneous dynamic range involves a number of complex issues, some of which may be at variance with others. A goal in the design of the PSDP/A system was to resolve these conflicts in a way providing the highest level of performance that could be realized within the scope of the effort. It was therefore necessary that the issues and their interactions be well understood. This section discusses the most significant of these issues in terms of their effects on the performance of the receiver system and on its electrical and physical configuration.

4.2 PERFORMANCE ISSUES

4.2.1 Digital Dynamic Range versus Bandwidth

The digital storage or processing of signal data is limited in performance by some function of bandwidth and dynamic range and, within limits, one can be traded for the other. The tradeoff function is not a simple product relationship but is more like the product of bandwidth and the logarithm of dynamic range. For the case at hand, this limiting function depends ultimately on the tape recorder. The recordings made by this system were to be compatible with the Ampex HBR-3000 tape recorder in place at RADC. This requirement for compatibility pointed to the Ampex AR1700 recorder as the only logical choice. Fortunately, that device represents the current state of the art.

These machines can accept a maximum of 24 bits of parallel digital data, using the 28-track configuration and 1-inch width tape. The maximum recommended data rate is 4 megabits/second on each track, at a tape speed of

120 inches/second (ips). The data output rate of the PSDP/A receiver is, therefore, constrained to a maximum of 96 megabits/second.

In principle, one could use any sample size and sample rate that result in a product of 96 megabits/second. However, the data must be recorded as 24 bits in parallel. For example, the original proposal suggested complex sample pairs of 8 bits each, packed 1½ samples to a 24-bit word. The corresponding sample rate would be 96/(8+8), or 6 MHz. Pairs of 6-bit samples would be packed two to a word and would allow an 8 MHz sample rate. Some other combinations would involve a somewhat awkward packing scheme. The most convenient combination is that of 12-bit samples taken in complex pairs at a 4 MHz rate, which requires no packing at all. This is the combination chosen for the PSDP/A design, although not for reasons of convenience alone.

The effects of the tradeoff between sample rate and size (or, more properly, resolution) can be seen as follows. The signal bandwidth that can be recorded is directly proportional to the sample rate. The constant of proportionality depends on the attenuation slope of the low-pass filters that must be used to prevent aliasing when analog signals are processed as discrete time samples. In a practical system, the ratio of sampling rate to bandwidth must exceed unity when the samples are taken in quadrature pairs.* For this system, let us assume that the ratio is 1.25.

On the other hand, the dependence of dynamic range is not linear. As a rule of thumb, every bit in the sample represents 6 dB in dynamic range, as long as sample sizes of at least several bits are considered. On this basis, we can tabulate the available dynamic range, arbitrarily referred to 1 bit, for different values of recorded RF bandwidth. Table 4-1 shows this relationship.

It is evident that the allowable digital dynamic range can be increased significantly by accepting a somewhat smaller RF bandwidth. For

*Note that the Sampling Theorem is not violated, since each complex sample consists of two independent parts.

Table 4-1. Digital Dynamic Range Relative to 1 Bit vs. Recorded RF Bandwidth

Bandwidth (MHz)	Sample Pair Rate (MHz)	Bits/Sample	Relative Dynamic Range (dB)
3.2	4	12	72
4.8	6	8	48
6.4	8	6	36
9.6	12	4	24

example, reducing the bandwidth by a factor of 2/3 from the originally proposed rate of 4.8 MHz provides an additional 24 dB of digital resolution. Because of the emphasis on wide dynamic range in this system, the tradeoff was pressed in that direction. Therefore, this design is based on sample pairs of 12 bits each, taken at a rate of 4 MHz. The corresponding maximum RF bandwidth is 3.2 MHz. The elimination of any need for data packing is an added benefit.

In addition to the maximum bandwidth, a post-sampling digital low-pass filter can be used to reduce both the recorded bandwidth and the sampling rate by successive factors of 2. The use of a digital filter maximizes the available recording time when the maximum bandwidth is not needed. Table 3-2 in the preceding section shows the relationship between recording time and bandwidth. For the maximum bandwidth of 3.2 MHz, the digital filter is bypassed.

In the same way as the digital tape recorder, the analog-to-digital (A/D) converter places a limit on the number of bits per second that can be processed. Fortunately, this limit is less restrictive than the one imposed by the recorder. Figure 4-1 shows the trend of conversion rate versus bits per sample for state-of-the-art products at the time the design was undertaken. With 12-bit resolution and a 5 MHz conversion rate, the MOD-1205 made by Analog Devices is a very close fit to the limits set by the tape recorder. It was the only A/D converter found that met those specifications.

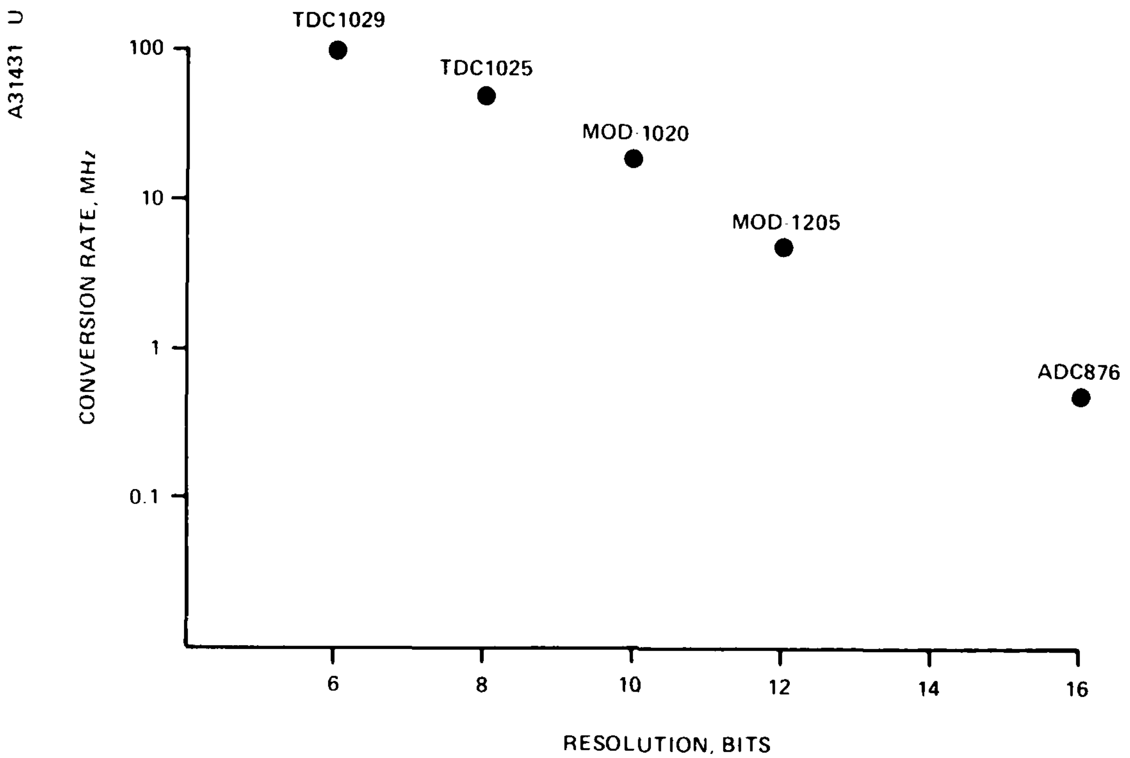


Figure 4-1. Performance of the State-of-the-Art A/D Converters

It should be recognized that digital resolution is only one of many factors that can limit the dynamic range of the system as a whole. The objective was to ensure that lack of digital resolution would not be a significant one of those factors. The use of 12-bit samples provides a dynamic range goal for the RF and IF portions of the receiver. That goal will now be developed.

4.2.2 RF and IF Dynamic Range

Having settled on digital quantization to the 12-bit level, it remained to arrive at specifications for the RF and IF sections of the receiver that take best advantage of the 12-bit resolution. To begin with, the largest voltage that can be represented in a 12-bit system corresponds to 2^{11}-1 levels, since one bit is reserved for the sign (polarity). At the other end, it is well known that for signals uniformly distributed in amplitude over at least a few quantization intervals, the rms quantization noise voltage is

$1/\sqrt{12}$ of the quantization interval. Therefore, the voltage ratio of maximum peak signal without saturation to rms quantization noise is

$$\frac{V_p}{V_{qn}} = \frac{2^{11}-1}{\frac{1}{\sqrt{12}}} = 7091$$

or 77 dB for a 12-bit system. For a single continuous frequency, the peak signal voltage will exceed its rms value by 3 dB. However, real-world signals are rarely steady sinusoids, so the allowance for crest factor should be at least 6 dB.

It is common practice to let the receiver's sensitivity be limited by thermal and electronic noise at the front end, rather than by quantization noise. The total noise power is the sum of the front end noise and the quantization noise, which are assumed to be independent. However, the front-end noise power is proportional to the effective receiver bandwidth, while the quantization noise power is constant. To maintain the condition where front-end noise dominates as bandwidth is reduced, it is necessary to increase the receiver's noise gain by 3 dB for each factor of 2 by which bandwidth is reduced. This gain adjustment could be made most conveniently by incorporating it into the coefficients of the digital post-sampling filter that comes into play when a reduced bandwidth is selected.

However, another consideration in the scaling of the filter coefficients was that they should make the best use of the available number of bits, so that quantization of the coefficients did not become a significant source of error. To give the filter a gain of exactly 3 dB, it would have been necessary to make the largest coefficient only 58 percent of the maximum 12-bit 2's complement value. Therefore, the choice made was to increase this coefficient value to the maximum possible, scaling all other coefficients in the same proportion. The signal data was then shifted down one bit position to prevent arithmetic overflow in the filter. This arrangement gives each pass through the filter a gain of 1.65 dB. The result is an overall dynamic range that is nearly independent of the selected bandwidth, and a receiver sensitivity limited by front-end noise at any bandwidth.

If front-end noise power exceeds quantization noise power by a factor of 4 (6 dB), then the total combined noise will be higher than quantization noise by a factor of 4+1 = 5 (7 dB). The result is that quantization noise degrades the receiver's sensitivity by 1 dB. This is considered to be a good compromise. For example, to achieve a degradation of only 0.5 dB, front-end noise would have to be 9 dB higher than quantization noise (instead of 6 dB), and the overall dynamic range would be less by the difference of 3 dB.

Allowing 6 dB for crest factor and 7 dB for the total noise floor, we find that the overall dynamic range can be 77-6-7 = 64 dB. However, that range will exist only if it is supported by the RF and IF portions of the receiver. In any case, the figure of 64 dB was held as a goal.

4.2.3 Local Oscillator Phase Noise

Early in the process of designing the PSDP/A system, the level of phase noise generated by the three local oscillator sources was recognized to be a more important factor than at the beginning. Other research undertaken concurrently at SRC emphasized the fact that the detection performance of a bistatic radar was critically dependent on the spectral purity of its receiver local oscillators, which in turn affected its ability to reject clutter. The I/Q sampling method used in the PSDP/A receiver preserves the relative phase of each signal sample, and local oscillator phase noise adds directly to the signal phase. Excessive noise generated by the local oscillator sources would, therefore, limit the utility of the system in bistatic radar experiments. This problem pertains only to the first local oscillator, since the other sources are fixed in frequency and phase locked to a crystal-controlled reference.

The basic conflict here was between the wide tuning range of the receiver and the desire for a spectrally pure local oscillator source. Microwave oscillators that are continuously tunable depend on some method for varying the resonant frequency of a cavity, as is done in a YIG-tuned device. The noise generated by the oscillator's active device is confined by the Q of the cavity. Even though this Q may be quite high, an oscillator operating at, for example, 5 GHz may have significant noise 100 kHz away from its center frequency. Frequency synthesizers overcome this problem by phase-locking the source to a highly stable crystal reference, but they can be tuned only in steps rather than continuously.

Frequency synthesizers operating in the microwave region are large and expensive, and the use of such an instrument as a local oscillator source would have had a significant impact on the cost and size of the PSDP/A system. Therefore, the approach taken was to use a YIG-tuned oscillator in a receiver configuration that divided the tuning range into separate but contiguous bands, with a choice of intermediate frequencies that minimized the tuning range of the local oscillator source. This design allowed the selection of a YIG-tuned oscillator specified for relatively low phase noise and operating in a lower frequency range than would have been possible with the receiver configuration considered originally. Although the spectral purity of this local oscillator source is not comparable to that of a synthesizer, it provided the best compromise available within the anticipated cost of the system.

4.2.4 Error Effects

4.2.4.1 Harmonic and Intermodulation Distortion

Two types of distortion can occur in the baseband converter, and both types can produce spurious responses. The goal of the design was to keep these spurious responses at least 64 dB below the largest signal.

Harmonic intermodulation distortion takes place in the mixer when a harmonic of the LO frequency combines with a harmonic of the signal frequency to create a product term falling within the base-bandwidth. Since this bandwidth is much smaller than the IF, the only products of interest arise from signal and LO harmonics of the same order. For example, the second harmonic of the LO mixes with the second harmonic of the signal, and the difference in frequency is twice the difference between the signal and LO frequencies. This type of distortion in mixers has been treated recently in the literature[1,2] with some useful results. The suppression of various harmonic intermodulation products is given as a function of the difference between the LO power and the signal power at the mixer. Specifically, Table 1 of Reference [1] (Table 4-2 of this report) indicates that the second harmonic product ($2f_S - 2f_{LO}$) is suppressed by approximately ΔP-39 dB, where $\Delta P = P_S(\mathrm{dBm}) - P_{LO}(\mathrm{dBm})$. This result leads to a ΔP of -25 dB to obtain 64 dB of suppression. Similarly, the level of the third harmonic product is $2\Delta P$-18, requiring ΔP = -23 dB.

Using the more restrictive value, the design should allow a maximum signal level no greater than 25 dB below the LO power at the mixer. Then, the 2x2 product will be suppressed by 64 dB and the 3x3 product by 68 dB. Terms of higher order will be insignificant.

The other source of spurious responses is mixing of signal frequency components and their harmonics with each other. This effect is commonly referred to as two-tone intermodulation distortion because of its method of measurement. It can occur in the amplifiers, as well as in the mixers. The effect has been discussed extensively in the technical literature. (See, for example, Reference [3].) The two most fearsome varieties are second-order and third-order intermodulations, created by second- and third-order terms in the power series expansion of the transfer function of a mixer or amplifier.

With a test signal consisting of two tones at equal levels and at frequencies f_1 and f_2, the second-order intermodulation product shows up at f_1+f_2 and increases by 2 dB for every dB by which the test tone levels are raised. The third-order products occur at $2f_1-f_2$ and $2f_2-f_1$, and they increase three times as fast as the test tone levels on a dB scale. If these levels are plotted in dB as a function of the test tone level, their extensions to higher levels generally intersect at a common point, as Figure 4-2

Table 4-2. Approximate Suppression of Certain Intermodulation Products

(LO) n	(RF) m	S_{nm} Suppression (dBc)
1	1	0
1	2	ΔP-41
1	3	$2\Delta P$-28
2	1	-35
2	2	ΔP-39
2	3	$2\Delta P$-44
3	1	-10
3	2	ΔP-32
3	3	$2\Delta P$-18
4	1	-35
4	2	ΔP-39
5	1	-14
5	3	$2\Delta P$-14
6	1	-35
6	2	ΔP-39
7	1	-17
7	3	$2\Delta P$-11

n = high-level (LO) input
m = low-level (RF) input
$\Delta P = P_{RF}(dBm) - P_{LO}(dBm)$

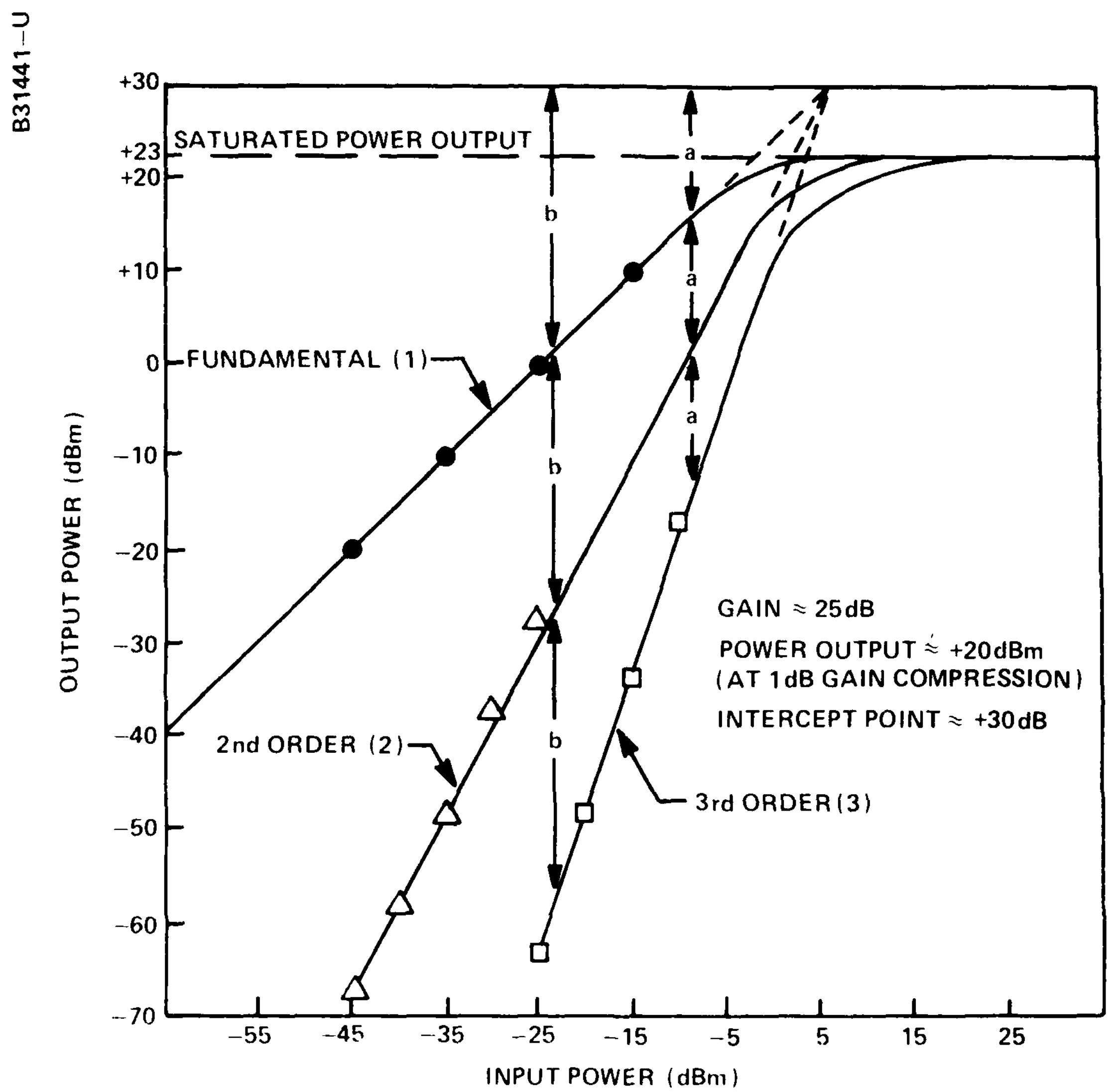


Figure 4-2. Intercept Point Response for a Typical Amplifier

shows. This point, called the intermodulation intercept, specifies the amount by which two-tone intermodulation products will be suppressed in a given device for any input or output signal level.

Two-tone second-order intermodulations in the mixers are of no concern, since they appear at twice the IF. Third-order products, however, will be translated to baseband by the LO and cannot generally be removed by filtering. From Figure 4-2, we see that to keep these responses 64 dB below the desired signal, the intercept point must be 32 dB higher than the maximum signal level. We can consider the specific example of a mixer such as the Anzac MD-148 with an LO power of +10 dBm. Recalling from the preceding section that we need $\Delta P \leq -25$ dB, the maximum signal level should not exceed -15 dBm at the input to the mixer. It follows that the third-order intermodulation intercept should be at +17 dBm or higher. Although the intercept point for that mixer is not stated specifically, we can infer from the catalog data that intermodulation products will be more than 60 dB down from the desired signal at -15 dBm.

In any case, spurious responses owing to all causes had to be measured and tabulated before making a final decision on the components used in the delivered system. Published component data is measured under certain conditions that may differ considerably from those encountered in practice.

Two-tone distortion measurements are a necessity for the baseband amplifiers, since these data are rarely given by the manufacturers of wideband integrated circuits. Second-order, as well as third-order, intermodulations are of interest here, since both can fall inside the passband. The requirements for these amplifiers are established by the baseband frequency range, the input voltage range of the A/D converter, and the necessary gain, based on the maximum output level of the mixers. We also need assurance that the noise contribution of these amplifiers will be insignificant compared to the front-end noise of the receiver.

Several candidates for amplifier components were identified and tested in breadboard circuits. They included the NE5539 made by Signetics; the 1322, 1342, and 1437, all made by Teledyne-Philbrick; and the LH0032, which is multi-sourced. The 1342 showed significantly lower distortion than the other types, had ample bandwidth and output range, and was relatively easy to apply.

4.2.4.2 Baseband Conversion Errors

The processing of the signal samples as complex number pairs is based on the orthogonality of the "real" and "imaginary" parts of the sample. If the LO in Figure 4-3 is not split in exact phase quadrature, or if the two signal branches are not matched exactly in phase, then the two parts of each sample pair will not be truly orthogonal, and the subsequent processing will be in error. A similar effect is brought about by imperfect gain matching in the two signal branches of Figure 4-3. Although gain mismatch does not affect the orthogonality of the sample pair, it does introduce distortion in the representation of the signal.

Theoretically, a complex-valued signal representation generated in this way has a spectrum that is a translated replica of the positive (or negative) IF signal spectrum. It can be shown that either phase error or gain mismatch in the baseband receiver produces image components that overlay the desired baseband spectrum, as Figure 4-4 shows. The effect is analogous to incomplete suppression of the unwanted sideband in a single-sideband (SSB) modulator.

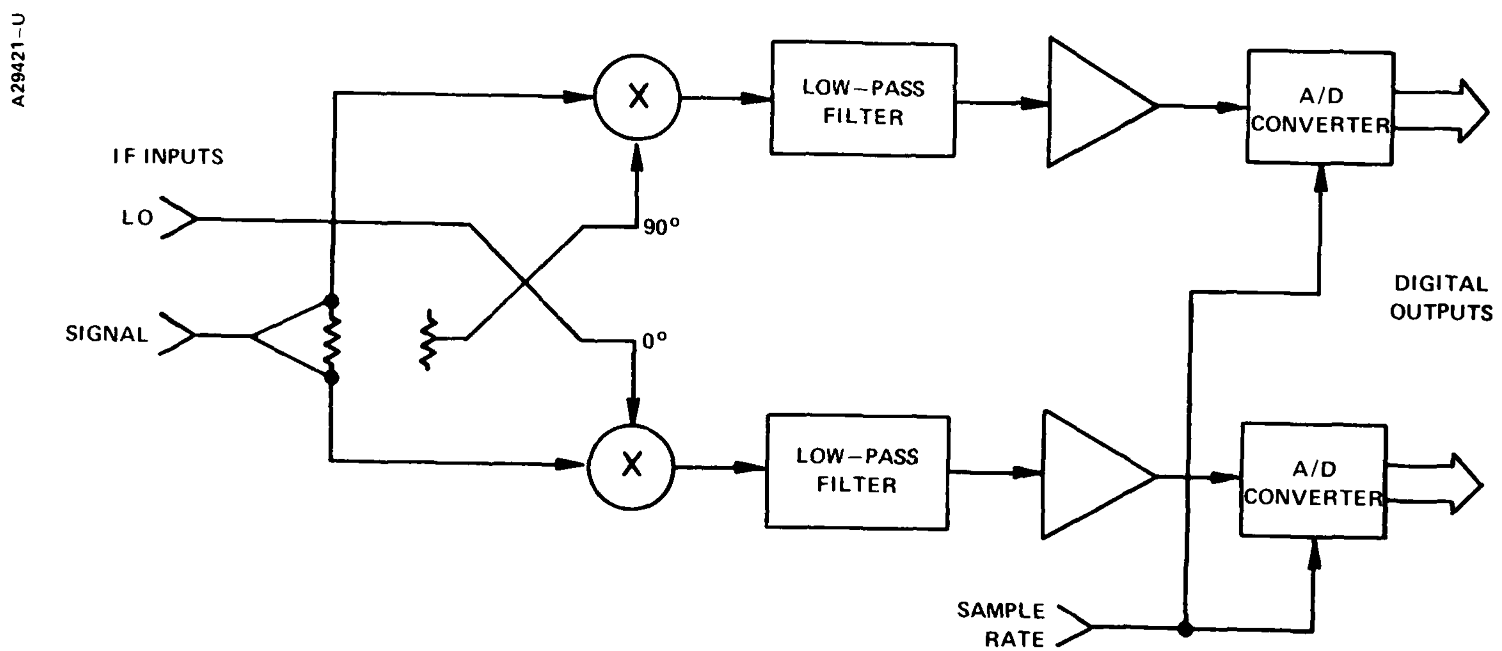
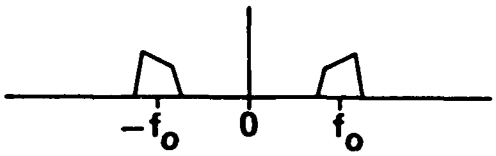
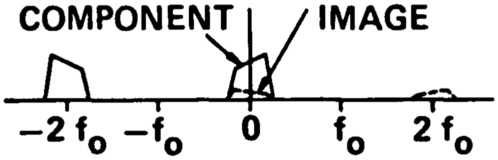


Figure 4-3. General Form of Baseband Converter and Digitizer



(a) INTERMEDIATE – FREQUENCY SIGNAL SPECTRUM



(b) SIGNAL SPECTRUM AFTER TRANSLATION BY A COMPLEX LOCAL OSCILLATOR OF THE FORM EXP $(-j\,2\pi f_o t)$ IN AN IMPERFECT QUADRATURE MIXER

Figure 4-4. Spurious Images Created by Gain Mismatch of Phase Quadrature Errors in Baseband Converter

To realize an instantaneous dynamic range of 64 dB, these spurious image components must be kept more than 64 dB down from the desired signal. Analysis[4] of these error effects has shown that the relative level of the spurious image is given by

$$\frac{S_u}{S_d} = \tfrac{1}{2}(\varepsilon^2 + \sin^2\delta)^{\frac{1}{2}}$$

where S_u/S_d is the voltage ratio of undesired image to desired signal, ε is the gain mismatch error, and δ is the phase quadrature error. To keep the spurious images below the -64 dB level requires a gain mismatch error not exceeding 0.008 dB with a phase quadrature error less than 0.05°. It is doubtful that these tolerances can be met in practice, even with the most careful design and construction, since the errors arise from a variety of sources. For example, the power divider, the mixers, the filters, and the baseband amplifiers can all contribute to both gain mismatch and phase quadrature error. In addition, any time skew in the sampling of the two branches is equivalent to a frequency-dependent phase error.

This situation, however, is not as hopeless as it might appear. First, the errors can be measured and adjusted to a relative level of about -40 dB. Second, the residual error can be corrected by subsequent signal processing, to the extent that it is not frequency-dependent. The procedure would be to record a continuous wave (CW) test signal at several different frequencies. For each frequency, a general-purpose computer can resolve the error into its orthogonal amplitude and phase components. The correction function can be based on an average over the test frequencies used, or it can be optimized for any given frequency within the passband. This correction is then applied to each signal sample when real signals are processed.

Third, the effect of the error is predictable, even if the error function itself is unknown. For example, a spectral analysis of the receiver output in the presence of errors will show an image component corresponding to each frequency component in an actual signal. Each image component is coherent with its counterpart, making it recognizable so that it can be ignored.

Offset or dc bias represents still another potential source of error in the baseband converter and digitizer. The mixers are the most serious contributors to this form of error, followed by the baseband amplifiers. In this design, dc offsets are eliminated by ac coupling prior to the signal sampler. The effect of the ac coupling is to introduce a very narrow notch in the frequency response of the receiver at the center of the IF passband. This notch causes a negligible loss of signal information because of its size relative to the IF bandpass. The only remaining sources of dc offsets are the sampling and A/D modules, which have trimming adjustments that can reduce those errors to the level of the quantization noise.

4.2.4.3 Frequency Aliasing

Prior to sampling and A/D conversion, the pair of outputs from the baseband receiver must be low-pass filtered, as shown in Figure 4-3, to sufficiently attenuate frequencies higher than half the A/D conversion rate. Figure 4-5 depicts the requirements for these filters. To maintain a dynamic range of 64 dB, the response of the filters at $f_s - f_c$ in Figure 4-5 must not exceed -64 dB, so that the aliased frequencies are excluded from the desired passband.

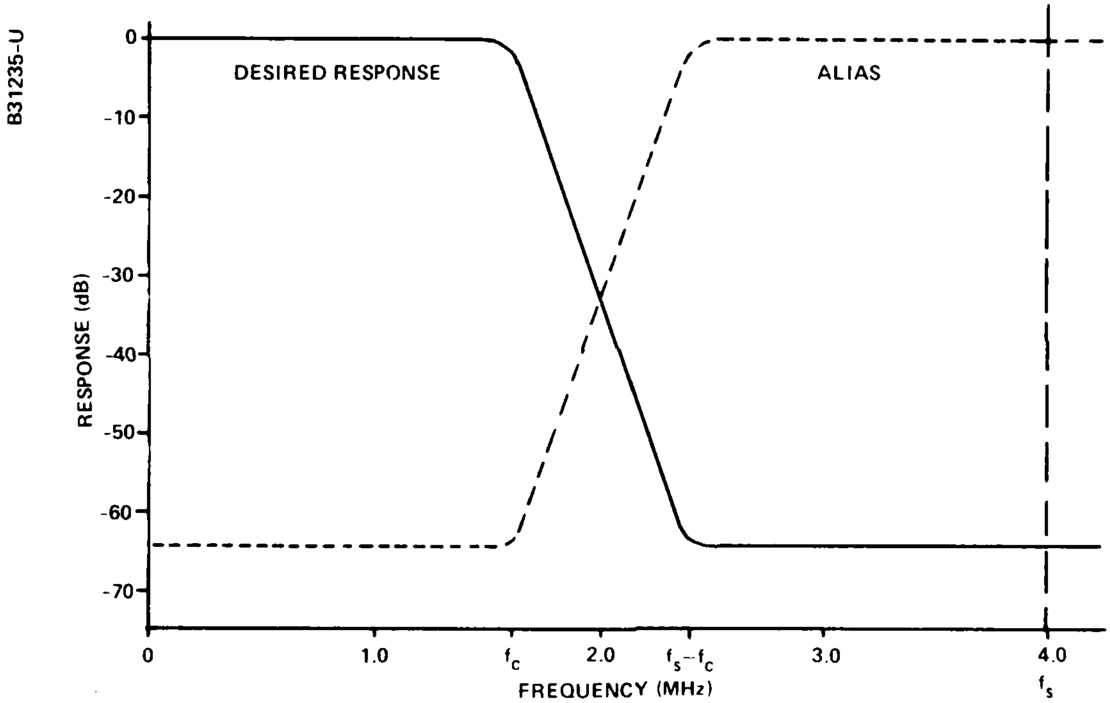


Figure 4-5. Response Specification for Pre-Sampling Anti-Alias Filters

The highest usable A/D conversion rate is ultimately limited to 4 MHz by the maximum rate at which the tape recorder will accept data. To maximize the utility of the PSDP/A system, we wanted to make the filter passband as wide as possible. However, the cut-off frequency (f_c) is limited in a practical sense by the complexity (number of poles) of the filter and by the allowable passband ripple. For example, a 10-pole filter with 0.05 dB of passband ripple has more than 60 dB attenuation at a frequency that is 1.5 times the -3 dB cut-off frequency. With $(f_s-f_c)/f_c = 1.5$, the cut-off frequency would be 1.6 MHz for $f_s = 4$ MHz. This arrangement represents a good compromise. A further increase in the cut-off frequency requires a rapid increase in the complexity of the filter, since the skirt becomes steeper. In addition, amplitude and phase matching becomes more difficult. It should be noted that any mismatch between the two low-pass filters in Figure 4-3 produces the same effects as gain mismatch and phase quadrature errors in the baseband receiver itself.

4.2.4.4 A/D Conversion Errors

The A/D converters following the baseband receiver (Figure 4-3) are constrained by the signal sampling and conversion rate of 4 MHz and the desired resolution of 12 bits. Converters using successive approximation techniques can provide the resolution, but they are nearly an order of magnitude too slow for this application. "Flash" converters, such as the TRW 1007J, can function easily at a 4 MHz rate but with only 8-bit or 9-bit resolution. One product, the MOD-1205, appeared to work on a combination of the two principles and had specifications very well suited to this application. It has 12-bit resolution and a maximum conversion rate of 5 MHz. The converter is a complete modular assembly that includes a sample-and-hold circuit, making it relatively easy to use.

One important parameter for the sample-and-hold circuit is its aperture uncertainty, or the uncertainty in the actual time a sample is taken after it is commanded. The specification is 25 picoseconds for the MOD-1205; this value can be compared to the period of the highest baseband frequency of 1.6 MHz, making it equivalent to a phase angle at that frequency of

$$\theta_{au} = 360 \cdot \frac{25 \times 10^{-12}}{\dfrac{1}{1.6 \times 10^{6}}} = 0.014°$$

This value is well below the phase quadrature tolerance required for errors at the -64 dB level, so aperture uncertainty is not a contributing factor.

Even though the aperture uncertainty is very small for a single unit, it is also important that the two samples be taken simultaneously within a comparable tolerance. For an allowable phase error of 0.05° at 1.6 MHz, the corresponding time skew is 87 ps, although that represents the worst case since it applies only at the edges of the passband.

4.3 CONFIGURATION ISSUES

4.3.1 Choice of Intermediate Frequencies

4.3.1.1 Single First IF

The configuration shown in Figure 4-6, which represents the original design concept for the PSDP/A receiver, has some characteristics that are immediately apparent. First, the tunable local oscillator must cover a range of 5.5 GHz, which is the same as the input frequency range. Also, the first IF must be outside the tunable range to prevent direct feed-through of signals at the IF. An IF below 500 MHz would make it difficult to achieve adequate image rejection, and it would require an local oscillator tuning range of more than two octaves. An IF above 6 GHz, however, presents neither of these problems. The first trial design was based on a first IF of 8 GHz to allow the use of a standard 8-18 GHz component for the local oscillator.

Once an IF is selected, it is possible to examine all of the possible harmonic intermodulation products to see if any fall inside the IF passband. The spurious responses caused by these products can be plotted as a function of the local oscillator frequency, as in Figure 4-7. The solid line is the desired response, and the dashed lines show responses resulting from various harmonic relationships. For example, the dashed line labeled "2,3" represents the second harmonic of the local oscillator mixing with the third harmonic of the signal. A vertical line extended upward from any value of

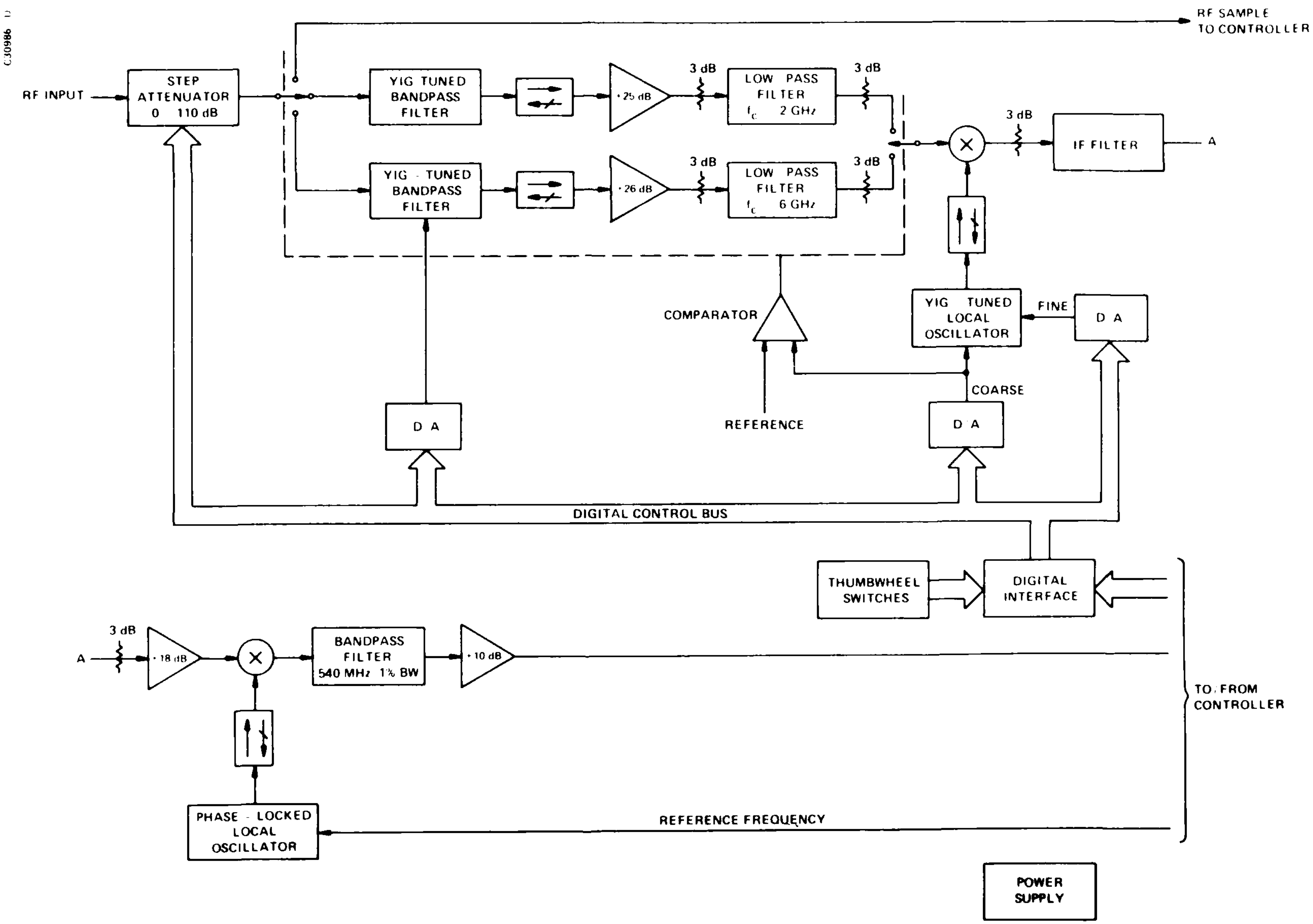


Figure 4-6. Block Diagram of PSDP/A Tuner

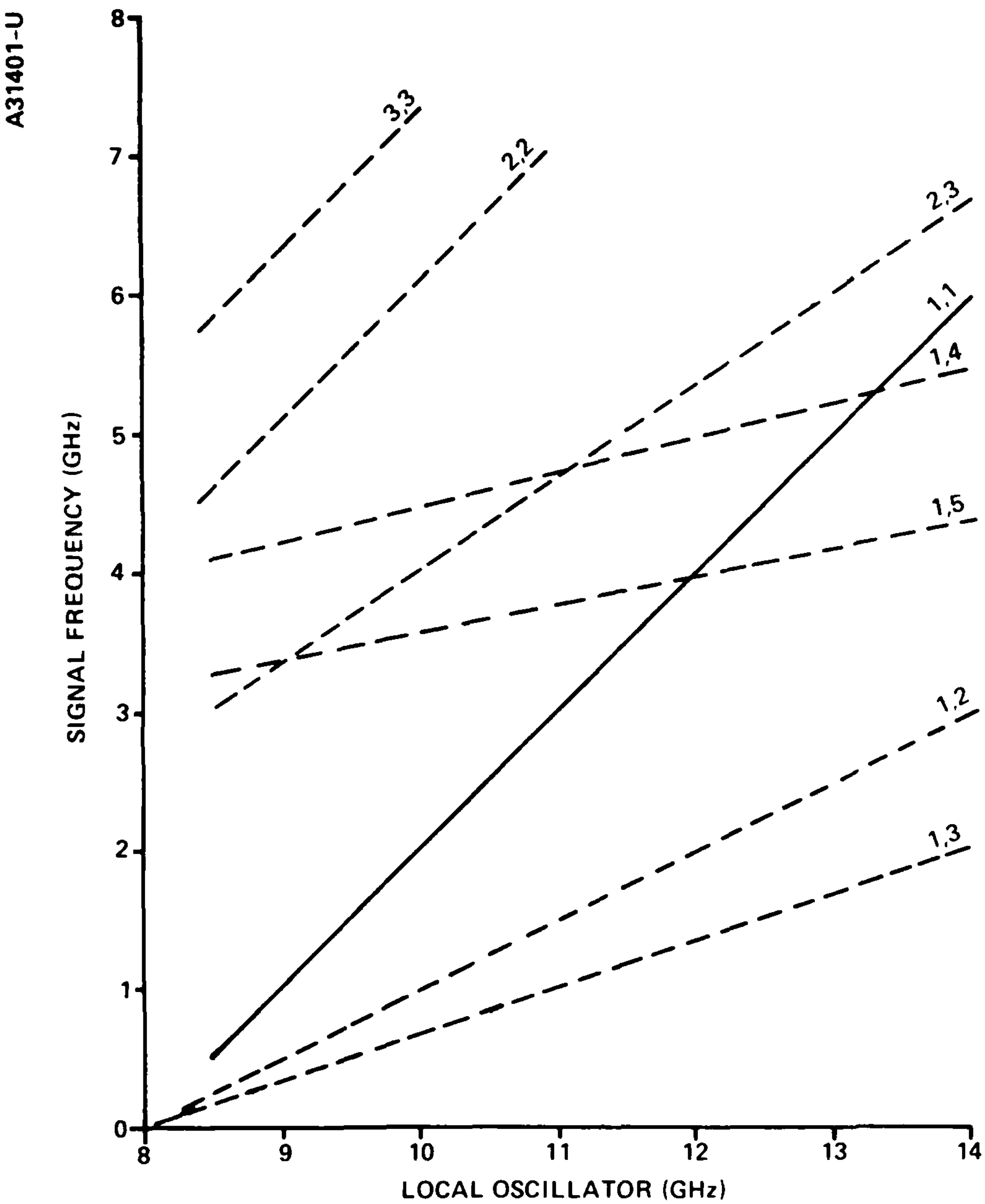


Figure 4-7. Responses Caused by Harmonic Intermodulations with IF of 8.0 GHz; Solid Line Is Desired Response

local oscillator frequency will intersect the dashed lines at the frequency of the corresponding spurious response.

The amount by which these harmonic intermodulations are suppressed in typical cases has been researched recently, and Table 4-2 (from Reference [1]) shows the results. As discussed in §4.2.4.1, the degree of suppression depends on the harmonic numbers and is a function of the difference, ΔP, between local oscillator power and signal power at the mixer.

With the exception of the 1,4 and 1,5 products in Figure 4-7, all of these combinations fall sufficiently far from the desired signal frequency that they can be adequately rejected by the preselector filter. The closest approach is the 1,2 line at the lower end of the tuning range, and that response is 250 MHz away from the desired signal. It is typically suppressed by (ΔP-41) dB, so even with a ΔP of only -15 dB, the preselector filter would need to provide only 8 dB of attenuation at 250 MHz to make that spurious response invisible. As the local oscillator frequency is increased, the 1,2 line moves farther away and should be still easier to eliminate by filtering.

The only apparent problems are the 1,4 and 1,5 responses, which cannot be attenuated by the preselector filter where they cross the desired response at 5.33 GHz and 4.0 GHz, respectively. However, these two harmonic combinations do not appear in Table 4-2, meaning that they are insignificant.

Although this receiver configuration is relatively simple and free of spurious responses, it does have some drawbacks. Its high first and second local oscillator frequencies would make it more difficult to achieve stability and low phase noise. In addition, the high first IF limits the degree of selectivity at that point to something on the order of 50 to 100 MHz if the insertion loss of the filter is to be kept down to a few dB. In a wide dynamic range design, it is a good idea to establish a narrow bandwidth limit as near to the front end as possible.

4.3.1.2 Dual First IF

The alternative configuration (Figure 4-8) uses two separate, lower first IFs but needs only one second local oscillator and mixer to arrive at a 540 MHz output. The two first IFs are separated by twice the second IF of 540 MHz so that the second local oscillator is on the low side in one case and on the high side in the other. Since the two preselector channels are divided by the 2 GHz boundary, switching between branches is simplified by changing IFs on the same boundary. Thus, the first IF for the 0.5 to 2.0 GHz segment should be above 2 GHz, and vice versa for the 2 to 6 GHz segment.

We could have split the difference between the two first IFs at exactly 2.0 GHz, making them 2.54 GHz for the lower input segment and 1.46 GHz for the upper segment. Then a second local oscillator frequency of 2.0 GHz produces an output at 540 MHz in both cases. However, it turns out that some harmonic intermodulation problems can be avoided by moving the two first IFs up to 2.68 GHz and 1.60 GHz, respectively, while keeping their difference at twice 540 MHz. The second local oscillator frequency moves up by the same amount, to 2.14 GHz.

The required tuning range of the first local oscillator can now be reduced, and intermodulation problems can be alleviated further by dividing the upper preselector range again into two branches separated by a 4 GHz boundary. The major difference is that the first local oscillator is on the high side of the signal between 2 and 4 GHz, and on the low side between 4 and 6 GHz. This arrangement confines the frequency range of the first local oscillator to less than two octaves, as Table 4-3 shows. The images are far enough away to be easily rejected by the input filters.

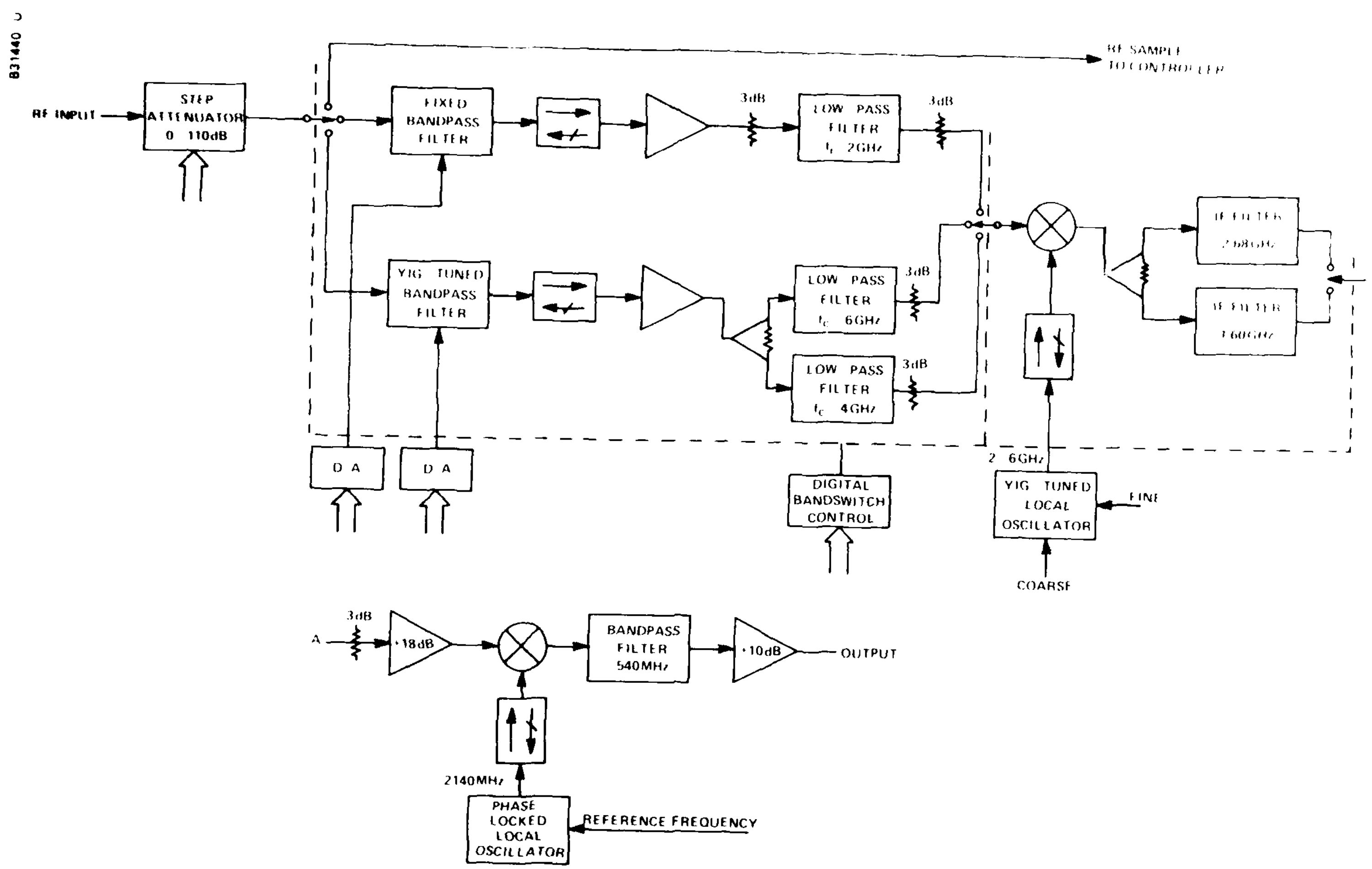


Figure 4-8. Modified Tuner Configuration with Dual First IF's

Table 4-3. Image and Local Oscillator Frequencies Using Dual First IFs

f_{SIG}	f_{IF}	f_{LO}	f_{IMAGE}
0.5 - 2.0	2.68	3.18 - 4.68	5.86 - 7.36
2.0 - 4.0	1.60	3.60 - 5.60	5.20 - 7.20
4.0 - 6.0	1.60	2.40 - 4.40	0.80 - 2.80

The spurious responses caused by harmonic intermodulations are plotted in Figure 4-9 for the input segment covering 0.5 to 2.0 GHz. This situation is nearly ideal. The closest spurious response, the 2,3 product at the high end of the local oscillator range, is 227 MHz away. Neither this one or the 1,2 product is strong, so they can be rejected easily by the tunable preselector filter.

Figure 4-10 shows the corresponding graph for the 2 to 4 GHz and 4 to 6 GHz input ranges. The use of a relatively low IF brings a larger number of harmonic intermodulation products into the region of interest. Two of these cross the desired signal response and cannot be eliminated when they are within the bandwidth of the preselector, which is on the order of 15 to 25 MHz. The 1,2 intermodulation product is relatively weak and should represent no problem in the vast majority of cases. The 2,1 product, however, does not depend on ΔP (see Table 4-2), but rather on the degree to which the mixer is balanced for the local oscillator input. (A perfectly balanced mixer would cancel internally-generated, even harmonics of the local oscillator completely.) The value of -35 dB given in the table is a typical one, but in practice the measured value could easily range between -25 and -50 dB for different units.

To put the problem into perspective, however, we should recall that this intermodulation is apparent only when a relatively strong interfering signal appears near a desired signal at 4.8 GHz.

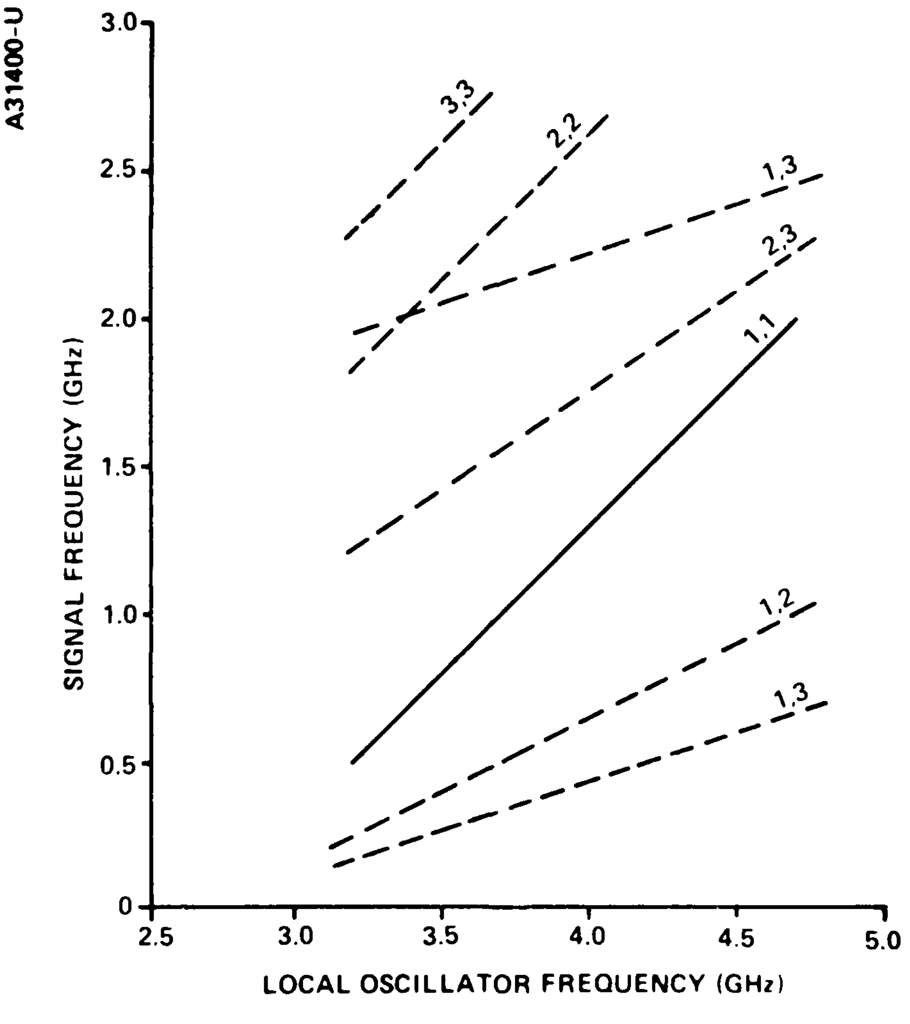


Figure 4-9. Responses Caused by Harmonic Intermodulations with IF of 2.68 GHz; Solid Line Is Desired Response

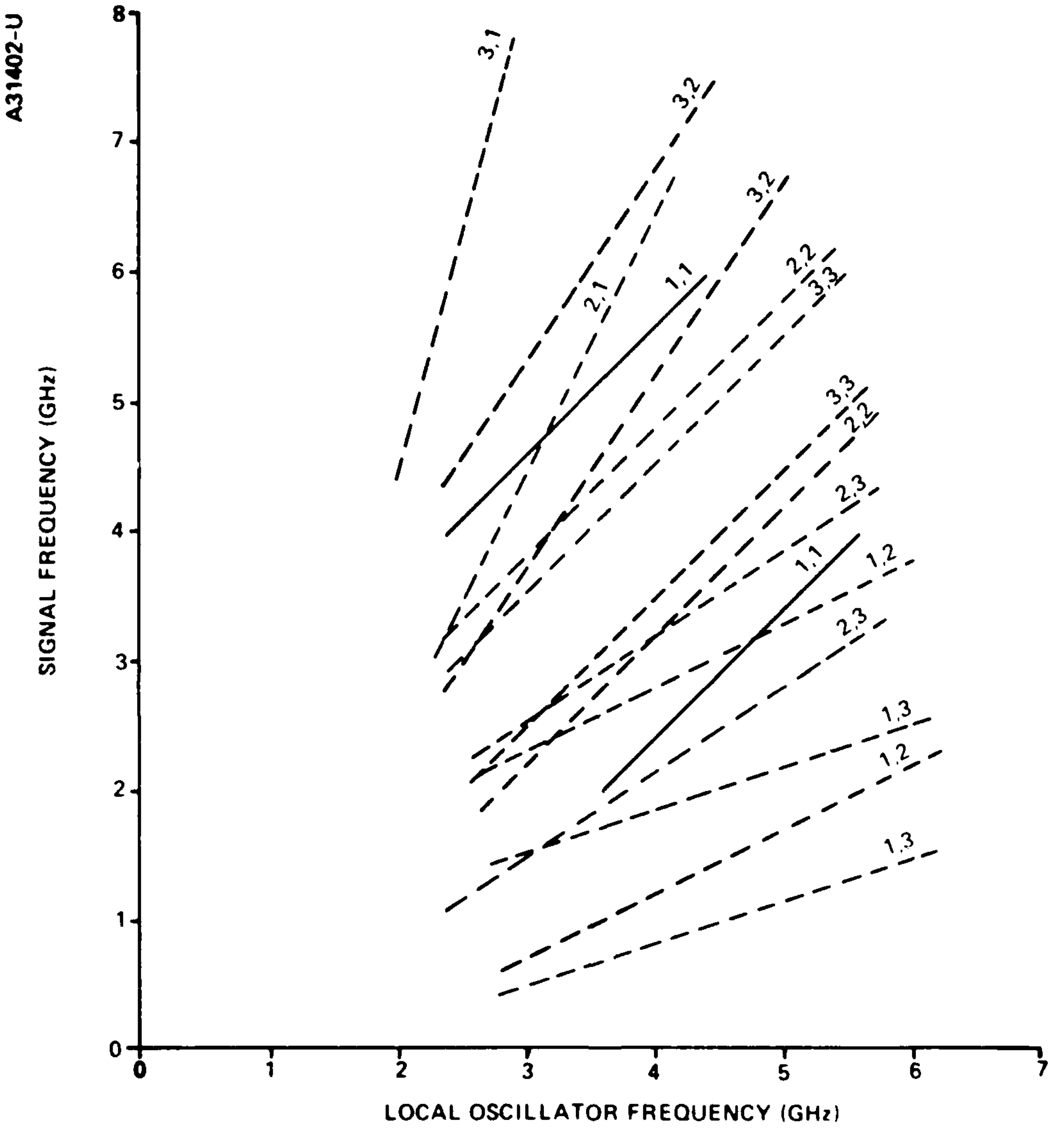


Figure 4-10. Responses Caused by Harmonic Intermodulations with IF of 1.60 GHz; Solid Lines Are Desired Responses

The remaining spurious responses can be adequately rejected by the preselector. The closest one is the relatively weak 2,3 product, which is separated by 133 MHz at its closest approach, when the receiver is tuned to 2.0 GHz.

The added complexity of the dual IF approach is offset by the lower operating frequencies of the first IF amplifier and the first and second local oscillator sources. Those components are less expensive, while offering better stability, better tuning accuracy, and lower phase noise, allowing a "cleaner" representation of the received signals.

4.3.2 Bandwidth Control

To provide an added measure of flexibility for the PSDP/A system, the receiver has five selectable RF bandwidths differing successively by a factor of 2. The maximum RF bandwidth of 3.2 MHz is twice the baseband cut-off frequency, which is determined by the anti-alias filters discussed in §3.2.4. When smaller bandwidths are used, the digital sample rate can be reduced in proportion. A lower sample rate allows, in turn, a slower tape speed, maintaining a constant data density on the tape. In this way, the recording time is maximized for the bandwidth selected.

The conventional way of accomplishing this would be to use a separate pair of low-pass anti-alias filters for each desired bandwidth and to change the sample rate to correspond to the selected filter pair. However, this arrangement makes the problem of filter matching five times worse, and it creates other opportunities for receiver errors because of the switching circuitry required.

A more attractive alternative is the use of a single analog anti-alias filter pair for the maximum bandwidth, followed by a programmable post-sampling digital filter for the subsequent reduction of both bandwidth and sample rate. This is the approach that was taken in the design of the PSDP/A system.

4.3.2.1 Digital Filter Requirements

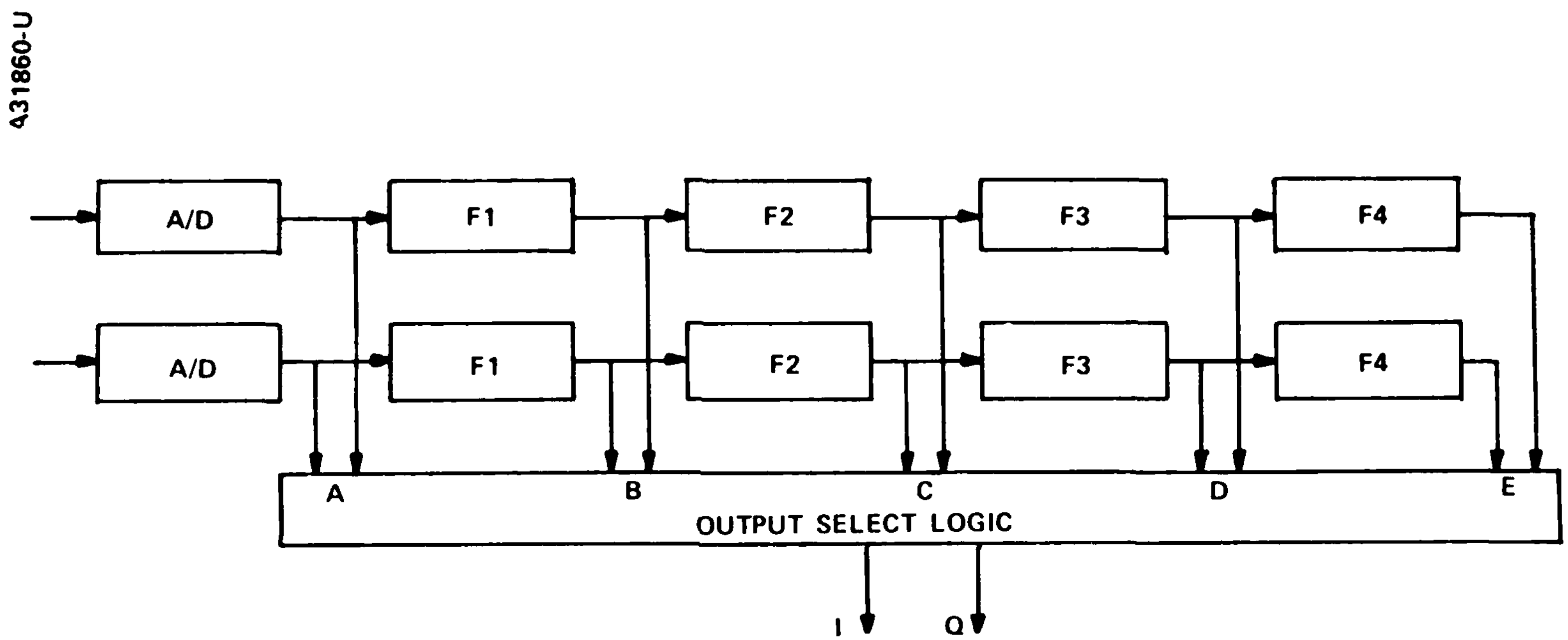
A functional block diagram of the digital filter is shown in Figure 4-11. Note that all of the filter blocks (F1-F4) are identical, in a sampled-data sense. Each block performs a 2:1 bandwidth and sampling rate reduction on the output of the previous block. Even if there were an infinite cascade of such blocks, the total digital computation rate is strictly bounded. Filter F2 has half the computation rate of F1, etc. The total computation requirement is always less than twice the requirement for the first stage.

Figure 4-12 illustrates the specification for any of the functional filter blocks (F1-F4). The frequency axis has been normalized by the filter input sampling rate. Thus, filter F1 has an input sampling rate of 4 MHz and a passband of ±0.2 x 4 MHz, or 1.6 MHz. If the filter stopband commences at a normalized frequency of 0.3, then the sampling rate at the filter output can be reduced by a factor of two without creating alias problems. The passband ripple specification is ±0.5 dB, maximum, with respect to the nominal passband response. A stopband specification of 60 dB maintains the 60 dB system dynamic range when sampling rates are reduced.

4.3.2.2 Filter Design Options

Digital filters are divided into two classes, and the first step in digital filter design is a decision between them. Infinite impulse response (IIR) filters are recursive structures, representing a translation of analog pole-zero specifications into the sampled-data domain. Finite impulse response (FIR) filters are direct-convolution structures, restricted to synthesizing all-zero transfer functions. As a rule, FIR filters are easier to design and more difficult to implement than the corresponding IIR filters. for a typical frequency-shaping design, FIR filters require more hardware than IIR filters.

In filters involving a sampling rate reduction, however, the FIR filter is favored over the IIR filter. In an IIR filter, the entire filter



OUTPUT	BANDWIDTH	SAMPLING RATE
A	3.2 MHz	4 MHz
B	1.6	2
C	0.8	1
D	0.4	0.5
E	0.2	0.25

Figure 4-11. Digital Filter Functional Block Diagram

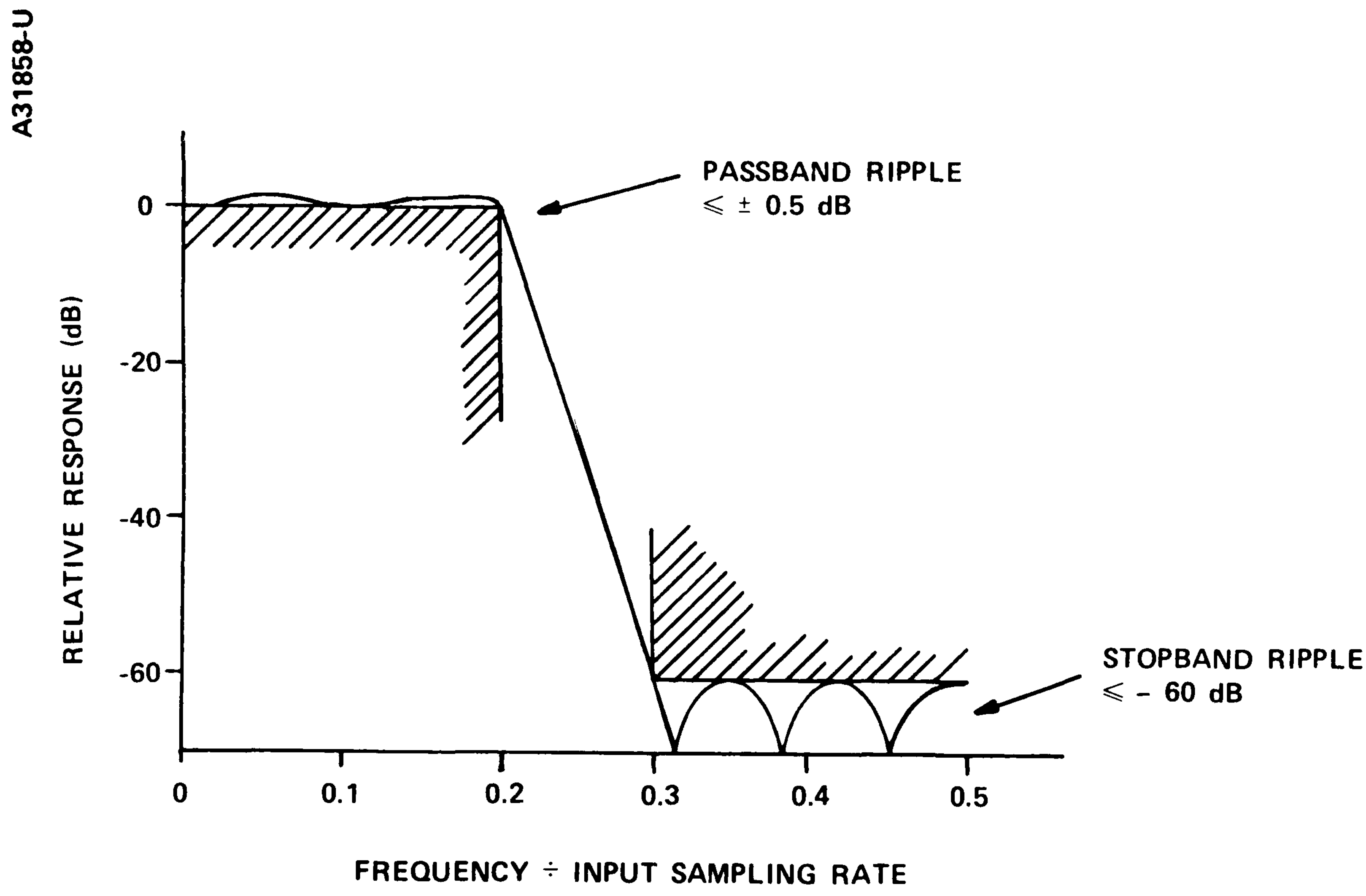


Figure 4-12. Digital Filter Specification

computation must be performed for each input sample. A reduction in sampling rate is accomplished by throwing away samples at the filter output, samples that we have paid to compute. In a FIR filter, only the reduced-rate output samples need be computed in the first place. From a computational point of view, the sampling rate reduction occurs at the filter input, rather than the output, although the higher sampling rate must be maintained at the filter input. For this reason, a FIR architecture was selected for the PSDP/A application.

4.3.3 Tuning Control

Remote tuning of the tuner makes digital control a necessity. An analog tuning voltage sent over a long cable would be too susceptible to noise and interference pickup, making the tuning inaccurate and unreliable.

The digital control format is dependent on the characteristics of the tuning element (in this case, a YIG-tuned oscillator) and on the effective bandwidth of the receiver. The oscillator has a range of 2 to 6 GHz, although only 80% of that range is actually used (see Table 4-3). A typical YIG-tuned device covering that range has a current-to-frequency transfer function that is very linear, but residual tuning errors can cause an uncertainty of about ±10 MHz. Therefore, it makes little sense to display the tuned frequency to a greater precision. For example, a tuning control that can be set in 1 MHz increments would be misleading to the operator. However, since the effective bandwidth of the system can be made as narrow as 200 kHz, the tuning control must have a resolution less than that value.

In view of these factors, the tuning system has both coarse and fine controls. The coarse tuning is digital, in increments of 10 MHz to correspond to the accuracy of the local oscillator source. In the local mode, the coarse frequency can be set by thumbwheel switches. When the tuner is remotely controlled, keypad entries at the front panel set the coarse frequency, and it is displayed by an LED readout in the format X.XX GHz. In this way, the tuned frequency indication is correct within one count. Fine tuning is accomplished digitally also, with a resolution of 10 kHz. However, the fine tuning control is a continuously-rotating knob attached to an incremental shaft encoder. Internal logic interprets the direction and amount of rotation and adjusts the frequency accordingly. The fact that the fine tuning control is implemented digitally is not apparent to the user.

Although only the local oscillator source has been mentioned in this discussion, the preselector filters use the same method of control. The necessary switching between filters takes place automatically when the 2 GHz and 4 GHz boundaries are crossed.

SECTION 5

SUGGESTIONS FOR FURTHER RESEARCH

5.1 EVALUATION OF RECORDED SIGNAL DATA QUALITY

Certain tests of the PSDP/A system as a whole require the recording of signal data and subsequent playback and analysis of the data. These tests include verification of the text header recording capability, overall signal fidelity tests, and measurement of weak signal sensitivity in the presence of quantization noise. Since the system has no playback capability of its own, tests of this kind depend on the availability of other playback facilities.

The basic method for conducting such tests is to record a signal having accurately-known properties and then transfer the recorded data upon playback to a general-purpose computer. The data is then analyzed to verify the parameters and fidelity of the original signal.

For example, a CW test signal could be used to test the error level of the baseband converter in the controller unit. The signal data transferred to the computer consists of pairs of 12-bit signal samples, the mean and variance of which can be computed individually for each part of the sample pair. Ideally, the means will be zero, and the variances will be exactly equal. Any difference represents a gain mismatch error, which can be removed either by circuit trimming or by further signal processing. Next, the normalized correlation coefficient between the two parts of the signal samples can be computed. Ideally, this value will be zero, and any departure therefrom represents a lack of phase quadrature or a time delay mismatch between the two branches of the baseband converter. These tests can be made at different frequencies to separate the time delay mismatch from the phase quadrature error.

After these errors have been removed from the signal data, the absolute magnitude of each complex sample can be computed, followed by the mean and variance of the magnitude. The ratio of these two quantities represents signal-to-noise ratio. Two-tone tests can also be made, using Fourier analysis to identify intermodulation products and their relative levels. Any tests of this sort must be well designed to ensure that errors inherent in the processing do not overshadow the quantities to be measured.

5.2 SIGNAL DATA PRE-PROCESSING

The PSDP/A system produces digital signal data in pairs of 12-bit words at rates as high as 4 MHz. When the data is played back and processed in a general-purpose computing facility, the data would be initially represented as pairs of 16-bit integers, since 12-bit words are not commonly encountered in computers. Looked at in this way, the system generates data at a maximum rate of four bytes every 250 nanoseconds, or sixteen megabytes per second. Although the rate can be reduced on playback to allow the data to be transferred to the computer, the quantity of data corresponding to one second in real time remains fixed. Since a typical recording event might involve a period of several minutes, this amount of data can tax the memory and mass storage capacities of even the most respectable installations.

The data quantity problem is aggravated by the fact that a large fraction of the recording may contain nothing of interest. When recording radar signal data reflected from targets, for example, the useful information is a very small part of the total recording. For that reason, a fruitful area for further work would be some effort relating to the pre-processing of the data to eliminate those samples containing no signal information. A pre-processor that could operate on the data in real time as the recording is played back would make the data analysis function much more efficient. It would relieve the computer of the initial signal processing tasks that are relatively simple but time-consuming because of the very large quantity of data involved.

In its elementary form, such a pre-processor might take the form of a thresholding circuit that would retain only those samples whose magnitude exceeds some selected level, along with an identifier such as the number of the sample, counting from the beginning of the playback. The retained samples could then be passed to the computer at a lower rate.

A pre-processor would also open the door to handling the signal data in real time, without the recording function as an intermediary. One could imagine a PPI display constructed solely from signals received by the PSDP/A system, where a suitable illuminator was at hand.

5.3 SOFTWARE FOR SIGNAL DATA PROCESSING

Another possible focus for further research would be the development of software for the analysis of the signal data, using a general-purpose facility. The most obvious of these programs would perform frequency and time filtering to reveal patterns in the data, such as pulse repetition rates and Doppler shifts. More advanced processing would attempt to validate and track individual radar targets, for example.

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