Memo No. 18

SPECTRAL PROCESSOR

Project 2.625

Summary of Scientific Specification Meeting

1 December 1982

National Radio Astronomy Observatory

Green Bank, West Virginia

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11/2/84 - D. Backer
1/2/87 - R. Foster

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SUMMARY

This meeting was held to finalize the scientific specifications of the spectral processor (formerly known as the pulsar/signal processor) and to consider combining this project with the 300-foot spectrometer project. Participants included 5 non-NRAO pulsar observers, R. L. Brown of NRAO as a representative of the spectral line community, and most of the other engineers and scientists at the NRAO who are involved in this project.

There was a strong consensus that combining the pulsar processor with the 300-foot spectrometer project improved the scientific performance of each instrument, made maximal use of the limited engineering and technical manpower at Green Bank, was less expensive than constructing the instruments separately, and would provide the long-term benefits of high usage and maintenance familiarity. The NRAO Electronics Division reviewed this recommendation and decided to formally combine the two projects. W. D. Brundage summarizes the discussion on this topic in Section 11.

The scientific specifications were finalized to the satisfaction of the pulsar observers, who formed a good representation of the U.S. pulsar community. Particular progress was made in the dynamic range specification, number of spectral channels required, and in deciding on the necessary time resolution. R. L. Brown presented a clear set of specifications for spectral line observing, considering a large class of past observations and examples of new observations that would fully exploit the front-end improvements at the 300-foot. Realizing that some spectral line requirements may not have been included in this discussion, we are encouraging spectral line observers to contact R. L. Brown or R. J. Lacasse, spectral processor project manager. The scientific specifications, as updated at this meeting, will be issued shortly as a separate memo.

Two developments have occurred since this meeting that affect the reading of these notes. The selection of a control computer, which had been envisioned to be an LSI-11 based system, has now been narrowed down to two 68000 based "super-micros" for reasons of improved cost/performance and long-term expandability. Secondly, R. J. Lacasse has produced a detailed program plan for this project (Memo 17, in this series) that calls for its completion in July, 1986. There has been discussion that this completion date could be moved up by almost a year if additional manpower were made available for the project. Potential users of this system are urged to comment on the proposed time table and the relative priority of this project with respect to other Green Bank projects. M. Balister, head of the NRAO Electronics Division, and R. J. Lacasse, head of Electronics in Green Bank, would be the best people to contact.

Finally, the editors of these notes would like to let the contributors off the hook. This memo has been assembled from a variety of sources including formal contributions, copies of viewgraphs used at the meeting, notes taken at the meeting, and notes scrawled on napkins and coffee cups used at the meeting. The price paid for using such a variety of sources is a slight unevenness of presentation, some redundancy, and occasional errors of fact or interpretation. We apologize for any inadvertent misrepresentations and hope that the main content of this pleasant and productive meeting is reported here.

We want to thank all participants, particularly those who travelled long distances, for the time and effort they expended in preparing for and taking part in this meeting. We would like to extend special thanks to M. P. Haynes, Site Director at Green Bank, who provided much of the impetus and assistance that made this meeting possible.

1. FUNDAMENTAL RELATIONS

D. R. Stinebring

Some useful formulae are collected here for reference. The most important points are that:

(1) Each spectral estimate consists of only one independent sample per spectral channel, making frequency and time averaging a necessity.

(2) The dispersion smearing time has a range of 5000:1 for a fixed bandwidth, necessitating a wide range of total bandwidths in order to achieve milliperiod time resolution for most pulsars.

(3) Dispersion curvature is an important consideration over these large fractional bandwidths and must be corrected for.

Two useful references for the topics covered at this conference are: T. H. Hankins and B. J. Rickett, "Pulsar Signal Processing" and J. Ball, "Computations in Radio-Frequency Spectroscopy." Both can be found in Methods in Computational Physics, 14, Academic Press (1975) [QA 401.M514].

SPECTROMETER RELATIONS

$\Delta f = \frac{f_B}{N_Q}$ $\qquad$ $f_B = f_s$ = total bandwidth

$t_s = \frac{1}{f_s}$ = sampling interval

$\Delta t = \frac{N_Q}{f_s}$ $\qquad$ N_Q = channels/spectrum/Quad

= # samples/spectrum/quandrant

$\Longrightarrow \Delta f \Delta t = 1$

For example: f_B = 20 MHz $\qquad$ t_s = 50 nsec

N_Q	Δf	Δt
64	312 kHz	3.2 μsec
256	78	12.8
1024	19	51.2

Spectrometer mode:

N_Q points/Δt ==> f_B data rate

Dedisperser mode:

1 point/Δt ==> f_B/N_Q data rate

DISPERSION

DM = dispersion measure $[\mathrm{cm}^{-3}\ \mathrm{pc}]$

$$\Delta t_d = 8.30 \frac{\Delta f_{MHz}\ (DM)}{f^3_{GHz}} \quad [\mu\mathrm{sec}]$$

= time for pulse to sweep through frequency band Δf

Let $\beta \equiv \dfrac{DM}{f^3_{GHz}}$

Extremes:

f_{GHz}	DM	β
1.0	3	3
0.3	400	1.5×10^4

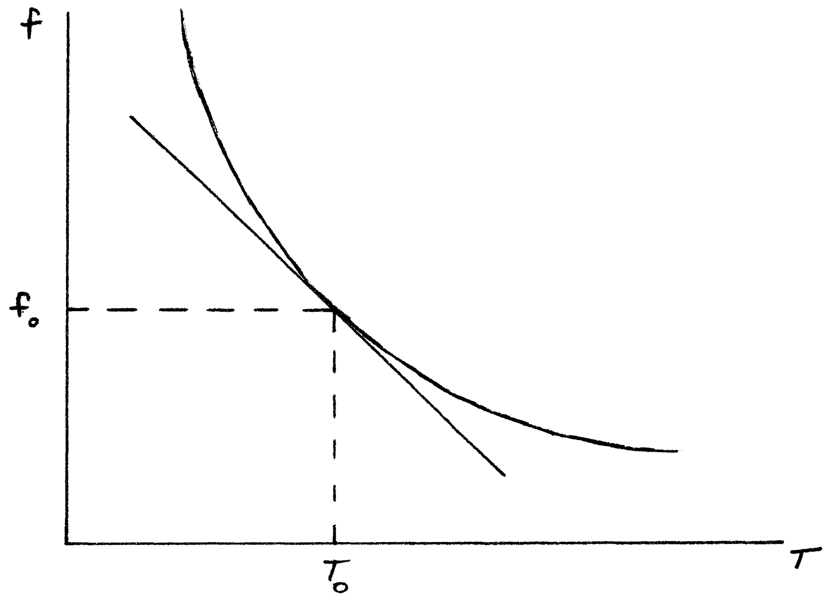
Best time resolution when:

$\Delta t = \Delta t_d$

$$\Delta f = 350 \left(\frac{f^3_{GHz}}{DM}\right)^{1/2} \mathrm{kHz}$$

2.85 kHz $< \Delta f <$ 200 kHz

DISPERSION CURVATURE



$$\alpha \equiv \left|\frac{df}{dt}\right|_{f_o} = 0.12\,\frac{f_{GHz}^3}{DM} \quad \left[\frac{MHz}{\mu sec}\right]$$

Arrival time:

$$T(f) = T_o - \frac{(f-f_o)}{\alpha} + \frac{3\,(f-f_o)^2}{2\alpha\, f_o}$$

Require:

$$\frac{3(f-f_o)^2}{2\alpha f_o} < \frac{\Delta t_d}{2} \qquad \text{assumes } \Delta t = \Delta t_d$$

But, $\Delta t_d = \Delta f/\alpha$

Curvature constraint:

$$\frac{f_B}{f_o} \leqslant \frac{4}{3N_Q}$$

2. PULSAR SPECTROSCOPY

J. M. Cordes

Pulsar spectroscopy is unique in its need for both small time resolution (Δt) and small frequency resolution (Δf). At minimum, the spectrometer must be gated in synchronism with the pulse period such that one or more "on" spectra and a separate "off" spectrum be accumulated from distinct subintervals of the period over an arbitrary number of pulse periods. The period range of known pulsars is now 1.6 ms - 4.3 s. Such gating takes full advantage of the natural temporal switching of pulsar signals which produces spectra with baselines superior to those obtained by frequency or position switching. Below, I list values of Δt and Δf needed for four areas of pulsar investigation in order of increasing temporal resolution and data rate.

1. Absorption Measurements of Interstellar Gas

   HI measurements, for example, require the above mentioned gating with accumulation times of minutes. "On" windows of $\Delta t \sim 1$ ms - 500 ms are needed. Frequency resolution $\Delta f \sim 1$ kHz.

2. Interstellar Scintillations

   Dynamic spectra with gating as in 1, accumulation times of $\geqslant 1$ sec, and Δf from 0.1 kHz to 10 MHz. These can be statistically analyzed for information on the power spectrum of the interstellar electron density.

3. Use of Interstellar Scintillations to Study the Geometry of Pulsar Emission Regions

   Gating with at least two "on" windows + one "off" window is required. Accumulation times as in 2. Different pulse components may scintillate differently if they originate from different locations.

4. Spectra of Individual Pulse Features

(e.g. micropulses with $\Delta t \sim 1$ ms)

This is the most demanding type of observation for Δt: need to record a spectrum once every $\Delta t \sim 1$ ms. Such spectra may yield clues about the emission process via statistical studies as a function of frequency and time.

Comments:

1. Number 4 in my opinion is not likely to be a common use of the spectrometer because previous studies suggest the pulsar signal is consistent with time variable Gaussian noise.
2. However, if achieving 4 is not costly it should be designed in for the sake of future flexibility.
3. Others (V. Boriakoff and D. Stinebring) disagreed with comment number 1. But it was generally agreed that type 4 studies could be achieved in software if necessary and should not be implemented in hardware if the incremental cost is large. [Ed.: It may be possible to provide this capability by including an array processor in the design.] Autocorrelations of the individual spectra could be performed in real-time and accumulated in the array processor. The result would be output after many pulse periods. Similar uses of an array processor, to slow down the data rate by performing high-speed statistical processing and accumulation can be imagined, e.g., pulsar searching and high time resolution dedispersing.

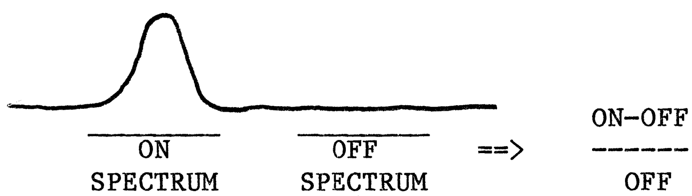
PULSAR SPECTROSCOPY EXAMPLES

I. Interstellar scintillations

<u>Dynamic Spectra</u>

A. Studies of interstellar electron densities (spatial distribution and power spectrum.)

Need to gate the spectrometer



Window size $>$ 5 ms

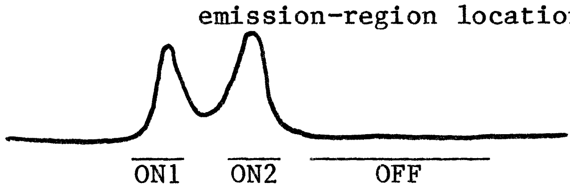
Accumulation over many pulses OK.

B. Studies of pulsar magnetospheres

see if different pulse components scintillate differently

==> measurements of

emission-region locations



Accumulation OK.

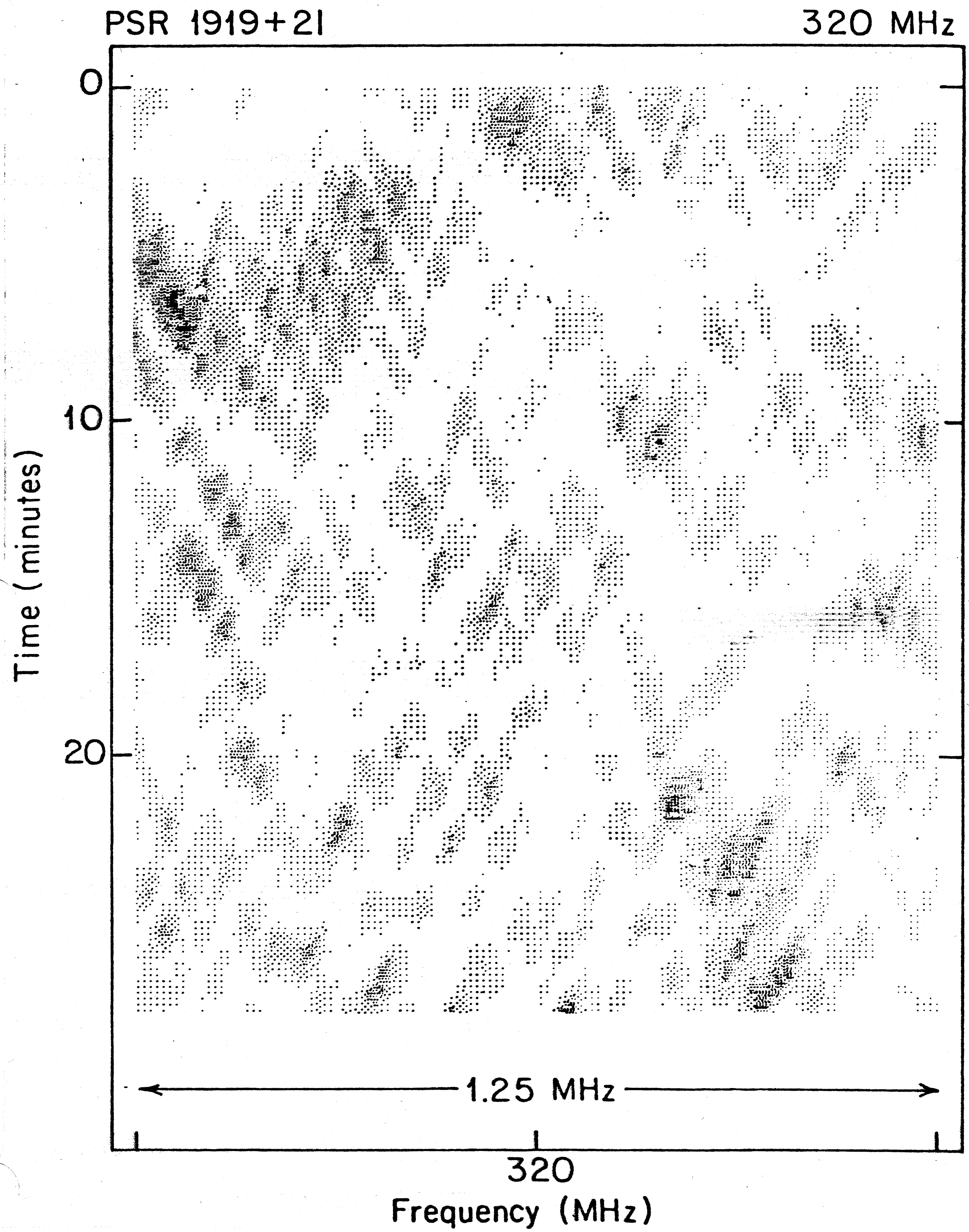


Figure 2: Interstellar scintillation pattern

II. Spectra of Individual Pulse Features e.g., micropulses $\Delta t \sim 1$ ms

Need to record a spectrum every millisecond during the on pulse

```
Q' |-|-|-|------------------. . .----------------|-|-|-|

U' |-|-|-|------------------. . .----------------|-|-|-|
    1 2 3                                              N
```

RICKETT, HANKINS, AND CORDES
Ap. J. 201, 425 (1975)

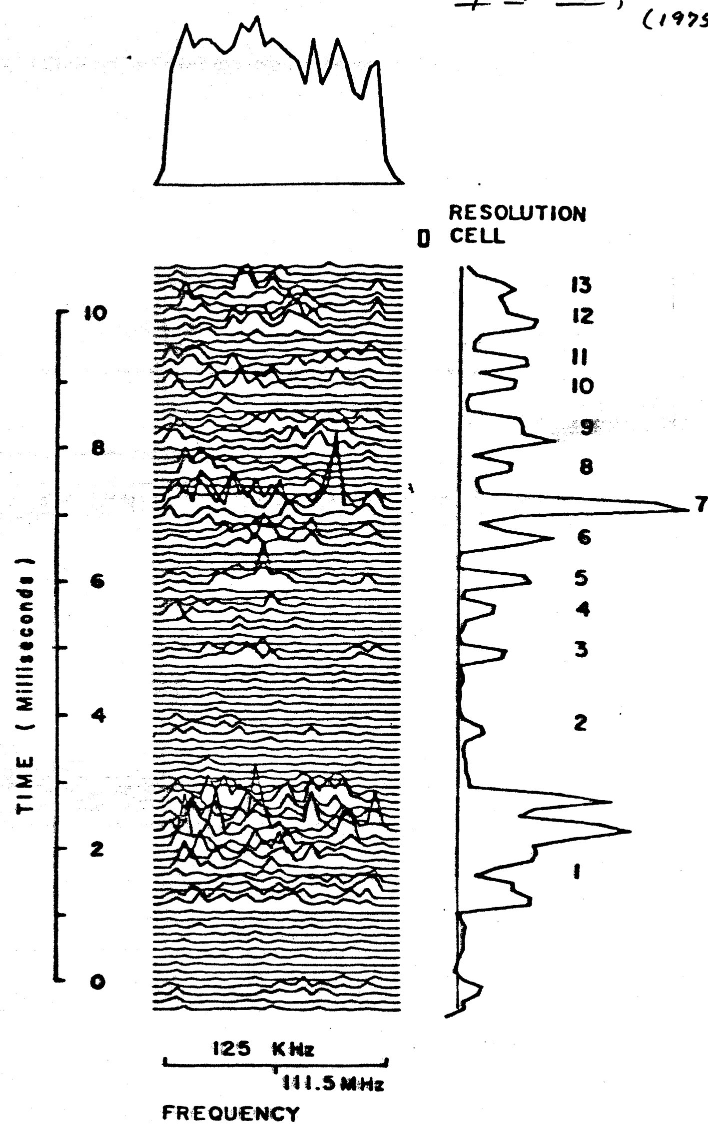


Figure 3

3. SPECTRAL LINE OBSERVATIONS

R. L. Brown

As a result of my experiences in doing spectroscopy on the 300-foot in the frequency range 500-1400 MHz, I see two principle needs that should be addressed when we consider any new spectrometer for the 300-foot. First, we need more than the present 384 spectral channels to properly exploit the sensitivity of our new dual-channel, low-frequency receivers. Second, the spectrometer must be able to perform better in the presence of narrow-band interference than do the present autocorrelators in which strong interference with a bandwidth less than 1 kHz will "ring" and render useless an entire 10 MHz spectrum.

Since the Spectral Processor may have the potential to alleviate both these difficulties and, with some modification, to serve as the 300-foot spectrometer, I enthusiastically support your considering this option. To this end, I summarize below what I see as the scientific considerations that drive the design of the next 300-foot spectrometer.

I. BANDWIDTH

1. Galactic HI

Over that part of the Galaxy accessible to the 300-foot, the width of the Galactic HI emission profile rarely exceeds 300 km s^{-1} (FWHI). At 1420 MHz this implies a bandwidth of only 1.5 MHz.

2. Extragalactic HI and Recombination Lines

These lines can easily be 500 km s^{-1} broad. If we allow 1000 km s^{-1} for line plus baseline region, this means a total spectrometer bandwidth of 5 MHz at 1420 MHz. The same lines redshifted to 750 MHz (say) are even less restrictive since they require, in this case, only 2.5 MHz bandwidth.

3. Line Searches (in frequency)

When searching in frequency the more bandwidth one can sample the better. The only limits that apply have to do with spectral resolution--the channel spacing should not exceed the expected line width--and the characteristic frequency interval between interfering signals that are too strong for the spectrometer to gracefully handle.

The present Model III autocorrelator has a maximum bandwidth of 10 MHz. The Spectral Processor is specified for 20 MHz; this is better, but 40 MHz would be necessary for fine-structure line searches.

II. NUMBER OF CHANNELS: RESOLUTION

1. Galactic HI

All published surveys of Galactic HI have been done with velocity resolution of 2.5 km s^{-1} or worse. But the HI emission profiles have a wealth of detail--self-absorption features, etc.--on a scale of 1.5-2.0 km s^{-1}. To study such features we need to use a velocity resolution of ~0.5 km s^{-1}; this means that in order to see the whole line profile and establish the spectral baseline we need ~300 km s^{-1}/0.5 km s^{-1} = 600 spectral channels. Hence,

reasonable: 512 channels/half at 1.5 MHz BW/half.

desirable: 1024 channels/half at 1.5 MHz BW/half.

2. Line Searches

Let us assume that we are searching for redshifted HI absorption lines. We want our channel spacing to be no wider than the expected line width so as not to dilute the signal. The HI absorption lines toward 0235+164 have widths on the order of 5 km s^{-1}. If we adopt this as the resolution, then in order to search 20 MHz bandwidth at 1000 MHz sky frequency requires 1200 channels. With a dual-channel receiver one would search the same interval with both receivers and combine the two spectra to increase the sensitivity by $\sqrt{2}$. Thus 1024 channels/half at 20 MHz BW/half would be reasonable, but more would be desirable.

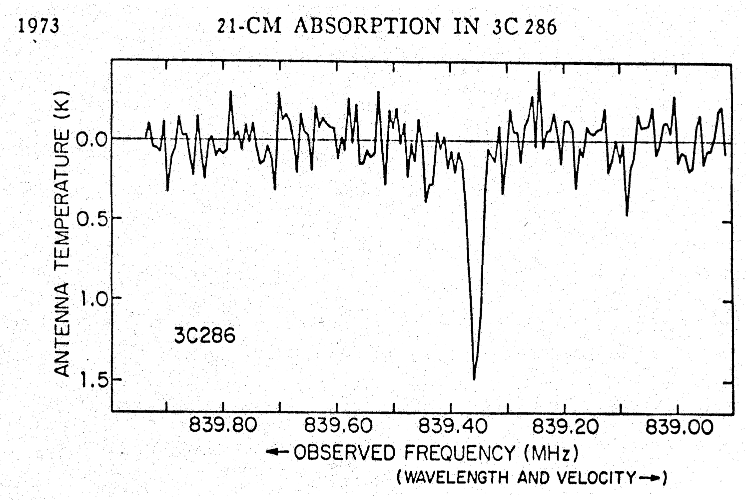
Fine-structure lines, although very broad (tens of megahertz), have little or no spectral structure. Thus 1024 channels/half at 40 MHz BW/half would be reasonable.

III. SAMPLING TIME

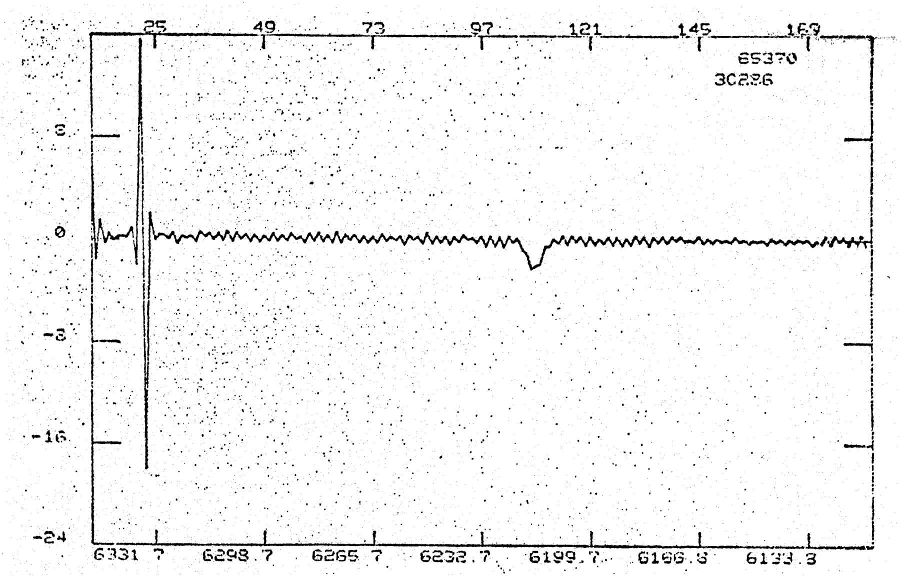
The present Model III and Model IV autocorrelator sample once every 20 seconds (minimum integration period). For most observations this is adequate. But it is highly desirable to be able to sample more rapidly for a few specialized experiments (maser variability/scintillation). Any (optional) capability to sample more rapidly than 1/20 Hz would be a useful feature to incorporate in the design.

Discussion:

M. Damashek remarked that present off-line Modcomp and POPS are limited to 1024 spectral points <u>total</u> in each of signal, reference, signal + cal and reference + cal arrays. Handling 2048 spectral points total (1024 points/half) will require more memory and a revised POPS--difficult but possible. [Ed.: This can be seen as part of the Green Bank computer upgrade now underway.]



Data taken 1972



Data taken 1982

Figure 4: 21-cm Absorption in 3C 286

Two points to be made here:

1. Interference environment, while still very good in Green Bank, is deteriorating--I suppose this is inevitable.

2. Modern receivers are much better than old versions. (For the spectra shown the 1982 observations were ~20 times more sensitive than the 1972 observations.) But in order to make use of this improvement we must address the interference problem seriously.

4. PULSAR SEARCHING WITH THE SPECTRAL PROCESSOR

M. Damashek

The Spectral Processor will excel at high-speed data-taking, with high spectral resolution as well. However, apart from de-dispersion, there is no current plan to include any facilities for searching in the processor hardware. Since the projected off-line processing capabilities are modest, the Spectral Processor itself does not promise a significant increase in searching ability.

This situation might change at some point in the future if an array processor were added to the resident computer. Such array processors are currently available at a cost of approximately $5500, and are directly compatible with the LSI-11/23. Combined with the very large memory that the computer can support (up to 4 megabytes), it should then be possible to search, either on-line or off-line, at a significantly higher sampling rate than is now possible (60 Hz in the Modcomp).

5. POLARIMETRY

D. R. Stinebring and J. M. Rankin

Multiplying vs. Adding Polarimetry

There is an alternative to the "analog (adding) polarizer" shown in Figure 1 of Memo 12. It is possible to do polarimetry on the output spectra of the FFT pipeline. If the amplitude spectra are A(f) and B(f), where A and B come from orthogonal linear inputs, then the Stokes parameters can be formed as:

$$I = A_R^2 + A_I^2 + g\ (B_R^2 + B_I^2)$$

$$Q = A_R^2 + A_I^2 - g\ (B_R^2 + B_I^2)$$

$$U = g^{1/2}\ (A_R B_R + A_I B_I)$$

$$V = g^{1/2}\ (A_R B_I - A_I B_R)$$

where g is a gain constant that can be determined before or after the observations and R and I denote the real and imaginary parts of the spectra. To get full polarization information in this manner requires the calculation, dedispersion, and accumulation of the quantities $(A_R B_R \pm A_I B_I)$, $(A_R^2 + A_I^2)$, and $(B_R^2 + B_I^2)$.

The main advantages of this multiplying polarimetry scheme are:

1. It uses half as many FFT blocks as the analog polarizer since no phase information is discarded.
2. It allows the correction of Faraday polarization across the passband (see below).
3. It allows the possibility of correcting gain and phase errors introduced by imperfections in the front-ends.

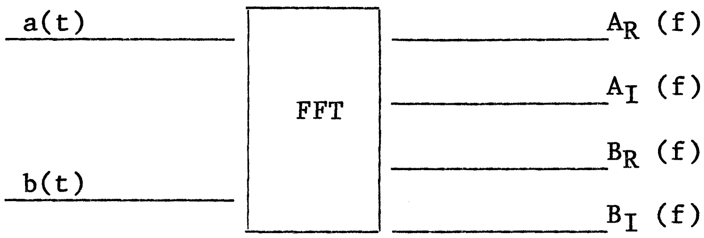
4. It simplifies the calibration procedure (only one gain constant is needed rather than 4) and eliminates a source (the analog polarizer) of possible gain and phase inaccuracy.

Correction for Faraday Depolarization

At 300 MHz the maximum usable bandwidth for polarimetry is 150 MHz/RM, where the rotation measure, RM, often takes on values $|RM| > 50$. This is a severe limitation for low-frequency polarimetry. If quantities proportional to the Stokes parameters Q and U are being separately accumulated, as they are in the multiplying polarimetry scheme above, differential Faraday rotation across the band can be corrected by rotating each $Q(f_i)$, $U(f_i)$ pair by an appropriate angle $\Delta\chi_i$. This needs to be done before dedispersion although it can take place after an initial stage of time merging. This Faraday correction technique may also be useful for spectral line polarization work since it allows wide-band polarimetry with depolarization losses smaller by a factor of 1/N than conventional filter-bank techniques.

Faraday correction could be introduced into the design in a modular fashion, either as an additional multiplier/adder block in the overall "pipeline" or as a slight embellishment to the "fast adder" idea that Hankins has introduced (see Section 10).

MULTIPLYING POLARIMETRY



<u>Total Power</u>

$$I_A = A_R^2 + A_I^2$$

$$I_B = B_R^2 + B_I^2$$

Polarimetry (A, B orthogonal linears)

$$I = I_A + I_B \quad = A_R^2 + A_I^2 + g\ B_R^2 + B_I^2$$

$$Q = I_A - I_B \quad = A_R^2 + A_I^2 - g\ (B_R^2 + B_I^2)$$

$$U = \mathrm{Re}\ (A\ B) \quad = g^{1/2}\ (A_R B_R + A_I B_I)$$

$$V = \mathrm{Im}\ (A\ B) \quad = g^{1/2}\ (A_R B_I - A_I B_R)$$

<u>Advantages</u>

gain and phase stable

no back-end gain calibration necessary

can reduce Faraday depolarization by $\frac{1}{N}$

can correct for gain and phase mismatch

FARADAY DEPOLARIZATION

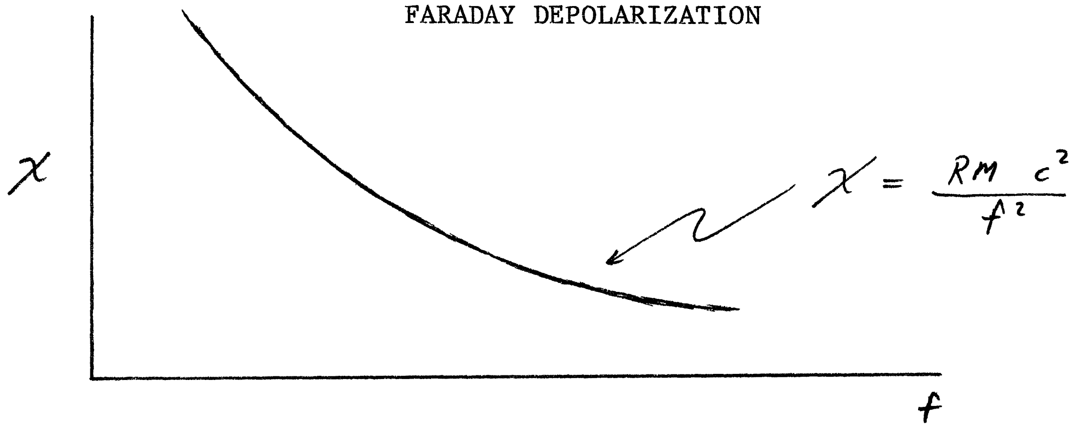


Figure 5. RM = rotation measure [rad m^{-2}]

$$\Delta\chi = \beta \, \Delta f_{MHz}$$

$$\beta = \frac{|1.8 \times 10^{-4}\ RM|}{f_{GHz}^3}$$

<u>Require</u> $\Delta\chi < 1$

$$\Delta f_{MHz} < \frac{5.6 \times 10^3 \ f_{GHz}^3}{RM}$$

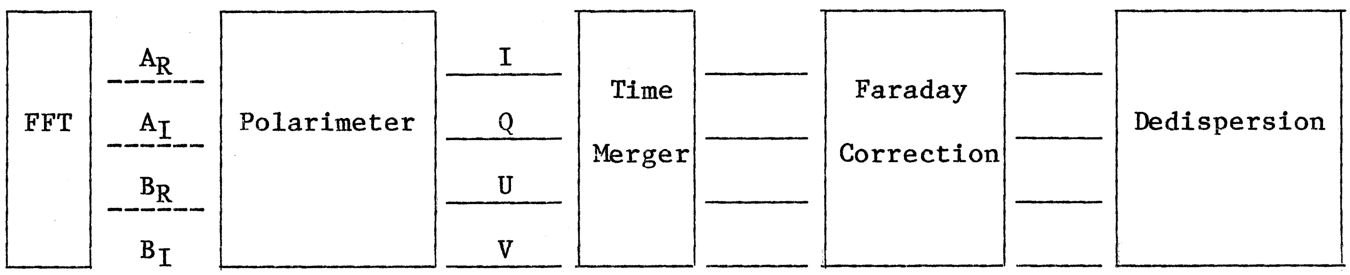
At 300 MHz:

$$\Delta f_{MHz} < \frac{150}{RM}$$

Many lines of sight have $|RM| > 50$

FARADAY CORRECTION

```
Q'  |-|-|-|-----------------.  .  .----------------|-|-|-|
U'  |-|-|-|-----------------.  .  .----------------|-|-|-|
     1 2 3                                               N
```



For i = 1 to N

$$Q_i = c_i Q_i + d_i U_i$$

$$U_i = -c_i Q_i + d_i U_i$$

$$c_i = \cos \chi_i$$

$$d_i = \sin \chi_i$$

$$\chi_i = \chi_o + \frac{RM\ c^2}{f_i^{\,2}}$$

6. DYNAMIC RANGE CONSIDERATIONS

J. R. Fisher

There are basically two definitions of dynamic range for the pulsar processor. One specifies the wideband noise power variations which can be accommodated and the other determines the required adjacent frequency channel isolation.

The noise power dynamic range is set by the largest instantaneous pulsar power relative to the system temperature. There seems to be agreement that 10 T_{sys} will be sufficient for the 300-foot telescope. Stronger interference that would overload the system will be excised before integration.

The processor should be capable of operating in the presence of narrowband interference by dropping a number of frequency channels before dedispersion. The retained channels should not be contaminated by this interference due to insufficient channel to channel isolation. The consensus of the group at this meeting was that no more than five channels on either side of an interfering signal should need to be thrown away when this signal's power is equal to the system noise power. For a one minute integration, with 256 channels covering a 20 MHz bandwidth, this criterion would require an isolation of 57 dB.

The attached figure shows the frequency response of an FFT spectrometer under three input data weighting conditions. The 57 dB specification will require a taper somewhere between the Hamming and Kaiser functions. Considerable work needs to be done to determine the intensity resolution required in the weighting coefficients and the effects of digitization at various stages of the FFT.

Interference excising logic in the processor will probably allow high speed, automatic elimination of wideband pulses and manually controlled, relatively continuous elimination of groups of frequency channels.

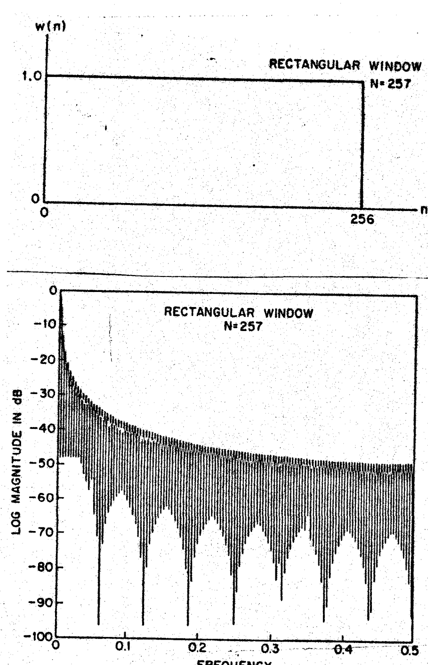


Fig. 3.12. Frequency response of a 257 point rectangular window.

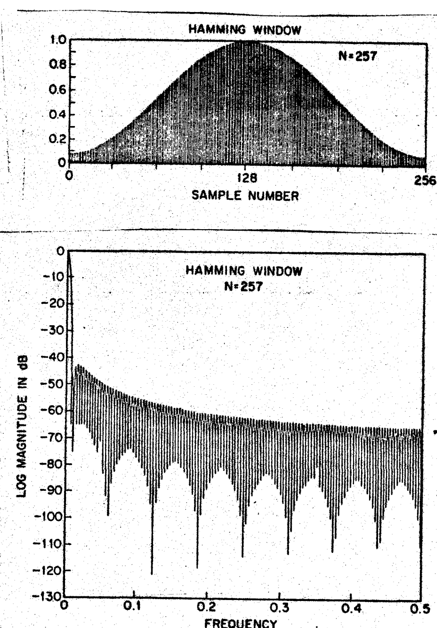


Fig. 3.14. Frequency response of a 257 point Hamming window.

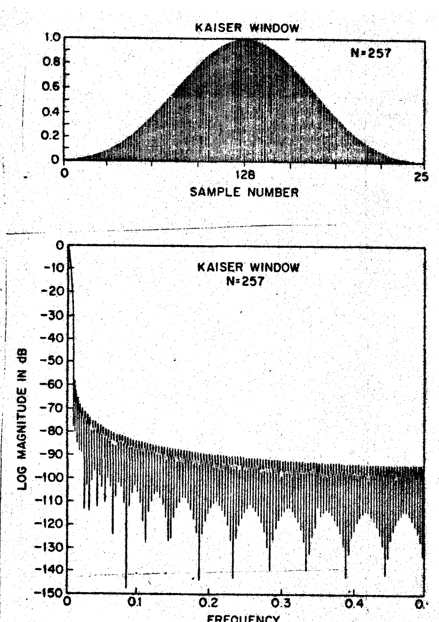


Fig. 3.16. Frequency response of Kaiser window.

Figure 6

7. PULSAR TIMING

T. H. Hankins

Pulsar timing will be done in at least three modes:

I. Pulsars with known pulse phase (routine timing noise, proper motion, polarization, scintillation studies).

II. Pulsars with unknown pulse phase (period, position, phase, DM, determination and study of pulsars without needing to know phase).

III. Pulsars with unknown period (searches, spectral studies).

I. Known Phase

It is nice to start an observing session in phase, and, in fact, most observing programs could use this capability. One could merely specify the data taking window length, off and on pulse and noise cal positions and go.

Advantages:

1. If pulsar is not in window, you know at once that you have done something wrong, the pulsar has glitched, or it is too weak to see. In the first two cases you should take action.
2. Data reduction is simpler and more efficient.
   a. fewer data points to record and correlate.
   b. expected vs. measured arrival time is immediately available.

Disadvantages:

1. You must know pulse phase. (ephemeris)
2. If a glitch occurs, you may well miss it.
3. Requires very precise ephemeris for phase prediction and Doppler correction.

II. Unknown Phase

This requires sampling the whole period. For observing efficiency and computational ease one would like an integer number of samples across the assumed period. If this can always be a power of 2, then doing CCF for pulse phase computation is easier, but this is not essential.

Advantages:
1. No interpolation required at end of sampling window.
2. No need to know phase, a priori, so automation is easier for long term or absentee timing runs.

Disadvantages:
1. Integer number of samples per period restricts sampling intervals backwards in the signal path. This need not be a problem, but it may require some rethinking or earlier stages.
2. For longer pulsars, the average profile record length becomes very long, i.e. at 100 μs resolutions, get 10,000 samples for a 1 second period. This need be only a conceptual problem.

III. Unknown Period

This mode would be used for pulsar searches where a Doppler shifted sampling frequency would be desirable, or for non-pulsar work where an arbitrary fixed sampling rate is desired.

Timing Modes:

1. Start at a particular epoch T_o, i.e., first sample at $T = T_o$ (unknown phase case).
2. Start at first pulse arrival time (minus offset to include beginning of pulse and some baseline) after a given epoch T_o (known phase case).

3. Start as in 1 or 2 above, but cause T_o to be the next 1 second tick after carriage return or button push.

4. Stop at epoch, after N periods, or button push. Auto restart possible.

Doppler Correction Method:

1. Frequency method: as in Joe Taylor's timing system, use a synthesizer without phase jumps continuously updated by the computer. Synthesizer costs $3,100 from Rockland/Wavetek. Allows integer number of samples per period.

2. Time method: As in MPIFR timing systems: A counter is preset by computer to the nearest number of microseconds for the current period. Timing error is never more than 1 μs. Sample intervals are then <u>not</u> Doppler shifted, as they are obtained from a 1 MHz standard.

Advantages:

1. The $3,100 synthesizer is not needed. All timing can be obtained from a few 8254 (or equivalent) programmable timers.

2. The time of the first sample of <u>each</u> window is known exactly to 1 μs and can be recorded with single pulses, for example, to avoid any ambiguity of pulse identification in multi-observatory experiments. This is not directly obtained in the frequency method.

Disadvantages:

1. It is nearly impossible to get an integral number of uniformly spaced samples per period.

2. For the 1.6 ms period pulsar the period calculation and update may take too long.

Implementation:

I recommend a stand-alone Doppler shift predicter with built-in pulsar and observer ephemerides. It needs to know only time, date, and pulsar name. Then it can be started and can continue independently of the rest of the system. This is often very desirable, particularly during new program development. The task can easily be done by an IBM personal computer or any 8086/8087/8088 single board computer.

The "frequency method" appears more flexible than the "time method" of Doppler correction. Either method could be fully controlled by a menu or cursor-driven CRT terminal/display. No other manual controls are necessary. This permits easy inclusion of a file-driven set-up with greatly reduced probability of set-up error.

In all three modes of operation provision for shifting pulse by a given amount would be almost essential, for example, for centering a pulse in a window. Then to be able to shift from mode II to mode I after phase determination would be very desirable.

8. DEDISPERSING I

V. Boriakoff

1. Dispersion Expressions.

The dispersion phenomenon in the propagation of pulsar pulses in the interstellar medium is known since the discovery of the first pulsar, CP1919. It consists of a slightly different group velocities for different frequencies of observation, as a result we see a larger delay for the arrival of the same pulsar pulse at lower frequencies than at higher frequencies. It is due to the free electrons in the interstellar plasma. The delays are of the order of fractions of second to a few seconds for the usual range of observing frequencies (100 MHz - 2 GHz). They can be computed from:

$$t = \int_0^z \left(\frac{1}{v_{gr}} - \frac{1}{c} \right) dz$$

where t is the delay between infinite frequency propagation time and the radiofrequency propagation time, v_{gr} is the group velocity, and the integral is done over the propagation path length.

$$t = \frac{1}{c} \int_0^z (n_{gr} - 1)\, dz = \frac{1}{2cf^2} \int_0^z f_p^2\, dz = \frac{e^2}{2\pi mcf^2} \int_0^z N_e\, dz$$

where n_{gr} is the group index of refraction, e and m are the electron charge and mass, f_p is the interstellar plasma frequency, f is the radio frequency of observation, and N_e is the electron density along the propagation path.

The arrival time difference at two radio frequencies f_1 and f_2 is

$$\Delta t = t_2 - t_1 = \frac{e^2 \int_o^z N_e dz}{2\Pi mc} \left(\frac{1}{f_2^{\,2}} - \frac{1}{f_1^{\,2}}\right)$$

The quantity DM $= \int_o^z N_e da$ is usually called dispersion measure, with units of parsec. cm^{-3}.

If we make the value $K = \frac{2\Pi mc}{e^2} = 2.410528 \times 10^{-4} \frac{\text{parsec}}{\text{cm}^3 \text{MHz}^2 \text{sec}}$

and f_1 and f_2 are the observed radio frequencies in MHz, the expression becomes

$$\Delta t(\text{sec}) = t_2 - t_1 = \frac{DM}{K} \left(\frac{1}{f_2^{\,2}} - \frac{1}{f_1^{\,2}}\right)$$

Both the receiver frequencies f_1 and f_2 and the time interval have to be corrected for Doppler effects, in this case the time interval at the solar system barycenter is

$$\Delta t_B = \frac{DM}{K} \left(\frac{1}{f_{1B}^2} - \frac{1}{f_{2B}^2}\right) = \frac{DM \left(1+\frac{v}{c}\right)^2}{K} \left(\frac{1}{f_1^{\,2}} - \frac{1}{f_2^{\,2}}\right)$$

where f_1 and f_2 are the receiver frequencies at the observing site. Since we measure Δt at the site and not at the barycenter

$$\Delta t = \frac{\Delta t_B}{\left(1+\frac{v}{c}\right)} = \frac{DM \left(1+\frac{v}{c}\right)}{K} \left(\frac{1}{f_1^{\,2}} - \frac{1}{f_2^{\,2}}\right) \tag{1}$$

which is to be our working relationship. Here v is defined as $v = \vec{v} \cdot \vec{r}$ where $\vec{r}$ is the direction to the source from the observing site ($v > o$ when moving towards the source). In practice $v << c$ ($\frac{v}{c} \leq 10^{-4}$) and the correction may be neglected for small $(f_2 - f_1)$ differences. The gradient of the curve is

$$\frac{\Delta t}{\Delta f} = - \frac{2.DM\,(1+\frac{v}{c})}{K} \frac{1}{f^3} \tag{2}$$

In dedispersing systems for small $(f_1 - f_2)$ one can approximate the variable with frequency group delay to be corrected by a linear approximation with a gradient computed from (2), but when the bandwidth $(f_1 - f_2)$ becomes larger than a few percent of f_1 special corrections have to be implemented.

Computations of t, $\Delta t/\Delta f$ for two pulsars (P0950+08, smallest DM known and P1859+03, one of the largest DM known) for a series of typical frequencies are included.

A typical frequency-time plot for two pulses is shown in Figure 1.

2. Filter Relations.

The rise-time of a typical filter of bandwidth B Hz is

$$\tau \approx \frac{1}{B} \text{ (sec)}$$

The dispersion gradient of equation (2) for a radio frequency f gives us a sweep rate of:

$$\frac{\Delta f}{\Delta t} = S = -\frac{K\ f^3}{2DM(1+\frac{v}{c})}$$

Making $\Delta f = B$ we will have $\frac{B}{\Delta t} = S$

For maximum sensitivity to a transient signal the length of time the signal is in the filter passband should be at least equal to the filter risetime. Hence, equating τ and Δt we obtain for the "best" signal-to-noise ratio:

$$B^2 = S \quad \text{or} \quad S = B^{1/2} \tag{3}$$

However, special considerations may dictate quite different B values.

3. Dedispersion.

Two main dedispersion procedures are in use: pre-detection and post-detection.

a. The pre-detection is based on the Fourier Transform pair

$$\phi(t-t_0) \rightleftarrows \Phi(\omega) \exp(-jt_0\omega)$$

where ϕ is a variable (e.g. electric field of the pulsar pulse) function of time, t_1 and t_0 is a time delay (due to interstellar plasma). In our case t_0

is also function of frequency. Inclusion of it's expression in the Fourier transform gives the mathematical operation to be performed on the signal. For several reasons this is not a convenient procedure to apply in this design.

b. The post-detection method is illustrated in Figure 2. It consists of taking the detected outputs of the consecutive filters of a filterbank (in our case sum of squares of real and imaginary parts of the individual frequencies of the FFT), and delaying them by different amounts function of the filter center frequency so that after the delay all the output signals become simultaneous again. At this point direct sumation of all the filter outputs provides a $N^{1/2}$ improvement of the signal-to-noise ratio (N is the number of channels) over a single channel (if interstellar scintillation decorrelation bandwidth is much larger than the total filterbank width). To obtain the intensity of the pulsar pulse usually two orthogonal polarizations are filtered by two identical filterbanks, detected, and summed channel pair by channel pair. The output of each summed pair is then dedispersed. The final output time sequence of such a device is

$$\begin{aligned} y(t) &= \sum_{i=1}^{N} x_i\,(t-(N-i)\Delta t) = \\ &= x_N(t) + (x_{n-1}(t-\Delta t) + (x_{N-2}(t-2\Delta t) + (\ldots\ldots+(x_1(t-(N-1)\\ &\qquad \Delta t)))\ldots) \end{aligned} \tag{4}$$

where $x_i(t)$ is the detected output of the i-th filter, N is the total number of filters and filter 1 is the uppermost filter in frequency. Notice that y(t) has the same timing as the lowest frequency filter ($i = N$), an important consideration for pulsar timing measurements.

4. Post-detection Dedispersion Implementation.

In one practical implementation of this dedisperser (Boriakoff, NAIC publication No. 38, "Pulsar Radiofrequency Observations with a Digital Pulsar Processor," Cornell Univ., 1973; also Boriakoff, "A Digital Pulsar Processor," Astron. and Astrophys. Supplement, 15, 479, 1974), the original idea due to Orstein (Rev. of Sci. Inst., 41, 7, 957, 1970) (Figure 3) was implemented in digital form. Individual channels are not delayed independently by the full amount necessary to make them simultaneous with the lowest frequency filter output, but are delayed only the necessary amount to make them simultaneous with the next lower in frequency filter output (second expression of eq. 4). The outputs of an analog filterbank were digitized by fast A/D converters, and fed to a series of summing stations separated by a shift register dealy. These shift register delays allow the storage of up to 8 intermediate samples between the summing stations. The variable length is produced by running a complex pattern of pulses as the shift register clock (elastic shift register). Availability of this extra storage permits a much higher sampling rate of the $x_n(t)$ signals. There are three main reasons why this is desirable: a) sampling rate, b) fixed filter bandwidths and c) curvature compensation.

a. Although eq. 3 is fulfilled in some pulsar observations, in many occasions (e.g. when fine time-dependent pulse details are not sought) the sweep rate S is set to $S < B^{1/2}$. Time constants then usually follow the detectors, with a common value for the time constant τ equal to $\tau = B/S$. It is considered that since the same signal (pulsar pulse feature) is present in the passband for B/S seconds (and also to increase the τB product of the output signal) the time constant should be equal to B/S. Application of the Nyquist criterion usually leads to sampling the output signal at a rate of 2 S/B Hz. Hence two samples (A/D conversions) should be taken in the time B/S, that is the time taken by the signal to sweep through the bandwidth of one filter channel, which is also the time taken to sweep from the center frequency of one filter to the next. This implies that one (and only one) extra digital word has to be kept in storage between the summing stations. (Orstein's machine and it's successors have been designed with one intermediate storage state between summing stations.) There are two incorrect assumptions here: that sampling taken at the Nyquist rate allows an exact or at least very good replica of the original waveform, when it is well known in engineering practice that only sampling at 5 S/B Hz or higher will allow good waveform reproduction. The second incorrect assumption is that the spectrum of $x_i(t)$ after passage through a τ time constant contains no frequencies above τ^{-1}, where as in reality for the most common analog time constants (RC) signals of frequency $f = \tau^{-1}$ are attenuated only ~ 6 times as compared to the low-frequency components constituting a non-negligible part of $x_i(t)$. When sampling at 2 S/B these high frequency components are aliased and add unwanted noise.

b. The second reason for having extra storage between the summing stations is that the filter bandwidth B and time constants usually have only a small set of discrete values. To adjust proper sampling of $x_i(t)$ (at 5 S/B) and a summing of the different $x_i(t)$ along a dispersion line an increased amount of versatility is given by this extra storage.

c. Another use of multiple storage stations between summing stations is the ability to change the slope of the dedispersion line maintaining a single digital clock operating the whole machine. This is achieved by splitting the dedisperser in two and running the part with the highest frequency filters with m delay stages between the summing stages and running the lower frequency filter section with m + 1 stages between summing stages. Because this procedure actually changes the dedispersion slope the deviation of the dedispersion line in the frequency-time plane from the true dispersion is smaller than that of the other curvature-compensating method, namely adding one delay stage between two specific summing stages (Figure 4) as is done in the Orstein and succeeding machines.

PULSAR=,JP 1858 WITH DISPERSION MEASURE=, 402.000000000PC/CM3

FILTER BANK= 32 FILTERS OF .0200000MHZ EACH FILTER
CENTER FREQUENCY OFFSET FROM 30.0MHZ BY .3500000MHZ

CENTER FREQUENCY MHZ	73.80000	111.50000	196.50000	318.00000	430.00000	606.00000	834.00000	1409.00000	1667.00000	2380.00000
TRUE CENTER FREQ.MHZ	74.15000	111.85000	196.85000	318.35000	430.35000	606.35000	834.35000	1409.35000	1667.35000	2380.35000
MHZ/SECOND	.12223	.41053	2.28608	8.67321	23.89580	66.83844	174.14096	838.29128	1389.74927	4043.69685
MILLISECONDS/MHZ	8181.0755	2383.6139	437.2578	103.3783	41.8484	14.9615	5.7425	1.1915	.7196	.2473
MILLISECONDS/CHANNEL	163.62151	47.67228	8.74516	2.06757	.83697	.29923	.11485	.02383	.01439	.00495

PULSAR=,CP 0950 WITH DISPERSION MEASURE=, 3.000000000PC/CM3

FILTER BANK= 32 FILTERS OF .0200000MHZ EACH FILTER
CENTER FREQUENCY OFFSET FROM 30.0MHZ BY .3500000MHZ

CENTER FREQUENCY MHZ	73.80000	111.50000	196.50000	318.00000	430.00000	606.00000	834.00000	1409.00000	1667.00000	2380.00000
TRUE CENTER FREQ.MHZ	74.15000	111.85000	196.85000	318.35000	430.35000	606.35000	834.35000	1409.35000	1667.35000	2380.35000
MHZ/SECOND	16.37956	56.51716	306.45534	1296.21037	3202.03704	8056.35030	**********	**********	**********	**********
MILLISECONDS/MHZ	61.0528	17.7882	3.2631	.7715	.3123	.1117	.0429	.0089	.0054	.0018
MILLISECONDS/CHANNEL	1.22106	.35576	.06526	.01543	.00625	.00223	.00086	.00018	.00011	.00004

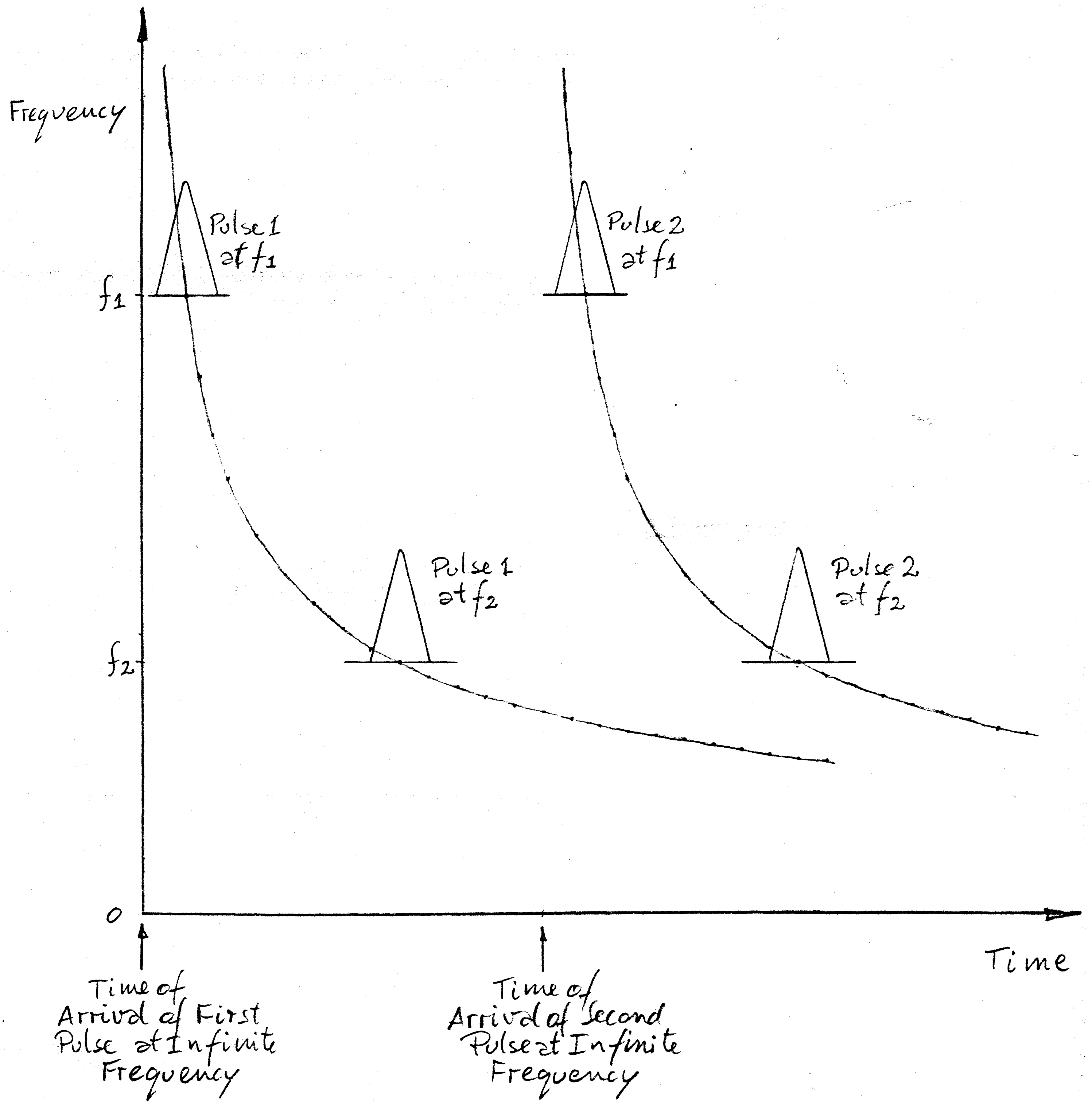
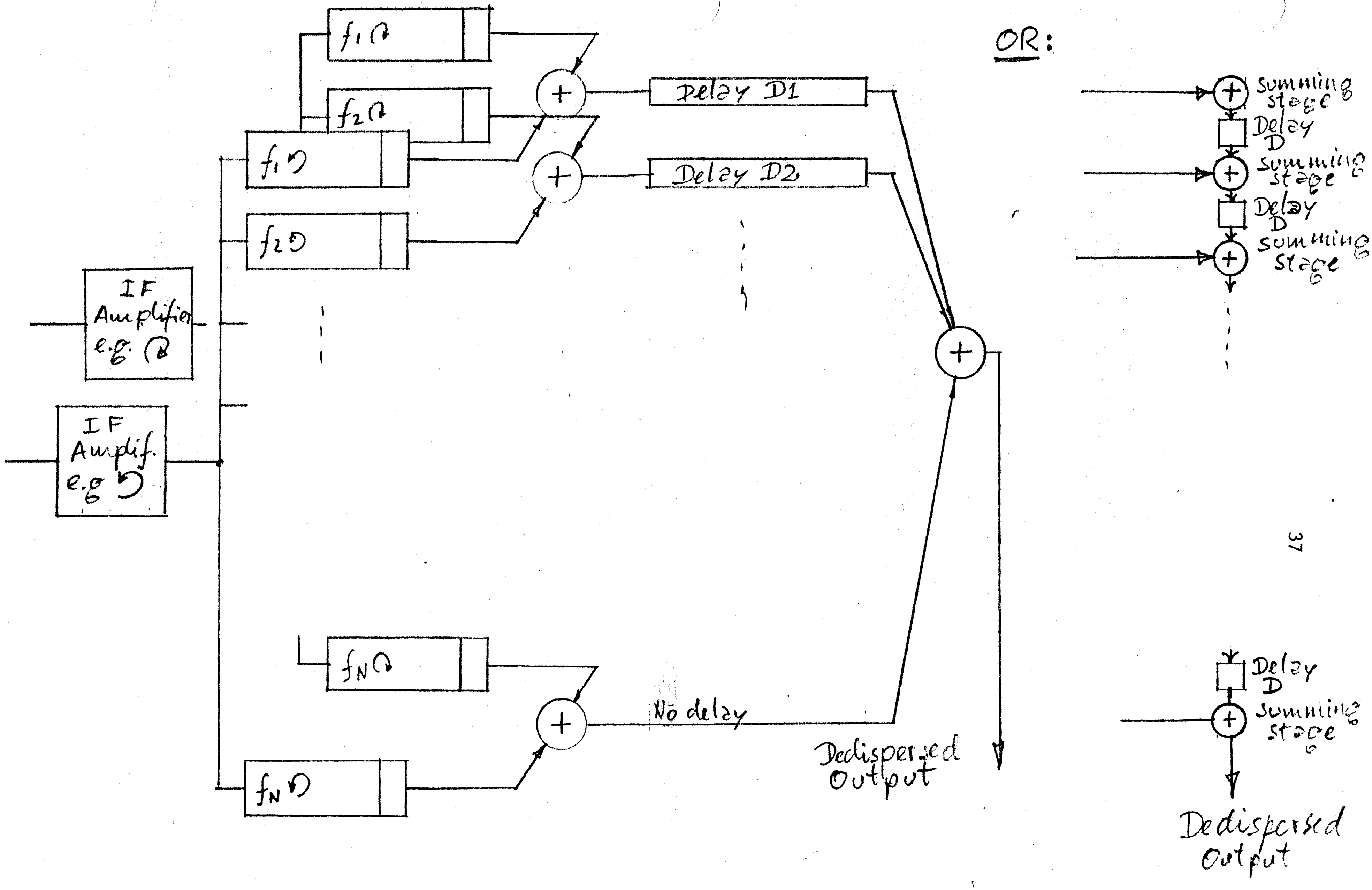


Figure 7
V. Boriakoff, Feb. 1983



$f_1 > f_2 > \ldots > f_N$

$D_1 > D_2 > \ldots > 0$

$D = D = \ldots = D$

Figure 8
Dedispersion by Two Implementations of Equation (4)
V. Boriakoff, Feb., 1983

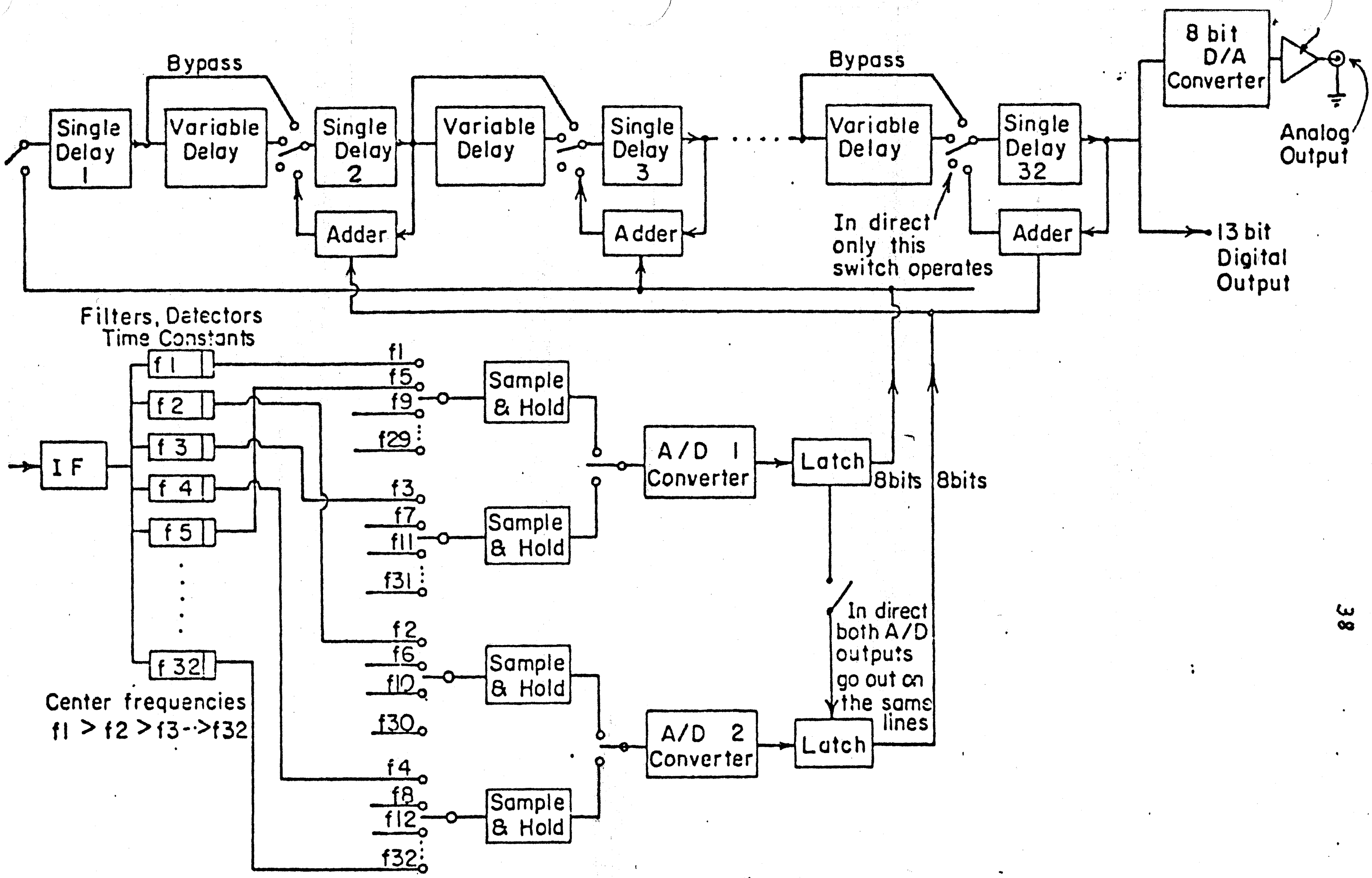


FIGURE 9 Block diagram configuration of the d.p.p. in the Post detection dedispersion mode.
From Borickoff, NAIC publication No. 38, 1973.

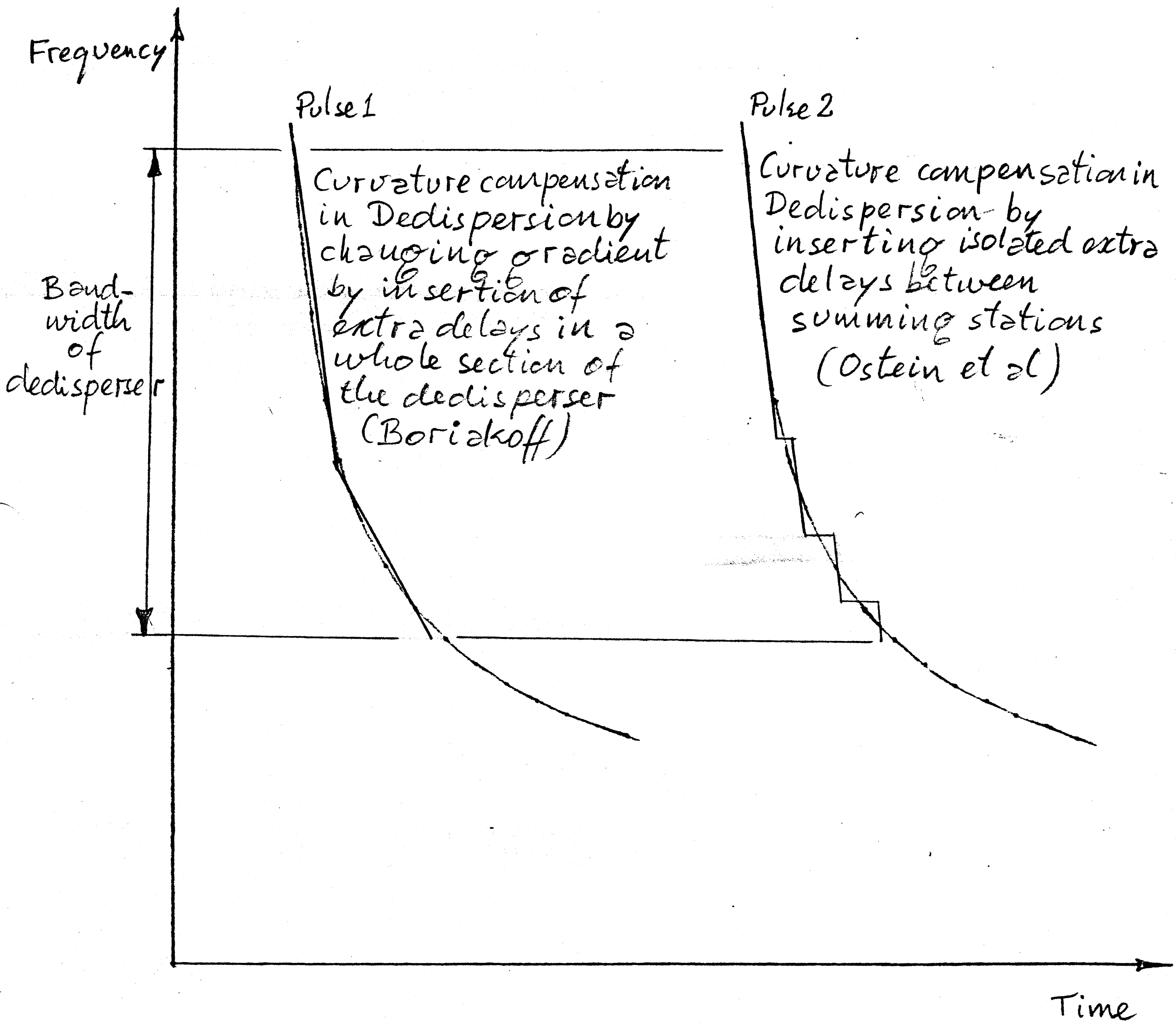


Figure 10

V. Boriakoff, Feb. 1983

9. DEDISPERSING II

J. M. Taylor

Dedispersion can provide a means to search for pulsars with unknown dispersion measure (DM). The present Princeton/NRAO system searches for dispersed pulses over 8 frequency channels at a 60 Hz sample rate (17 ms time resolution) with the possibility of improving this to 32 frequencies at a 120 Hz sample rate. The ideal search would require ~100 μsec time resolution Δt with ~20 to ~256 frequency points.

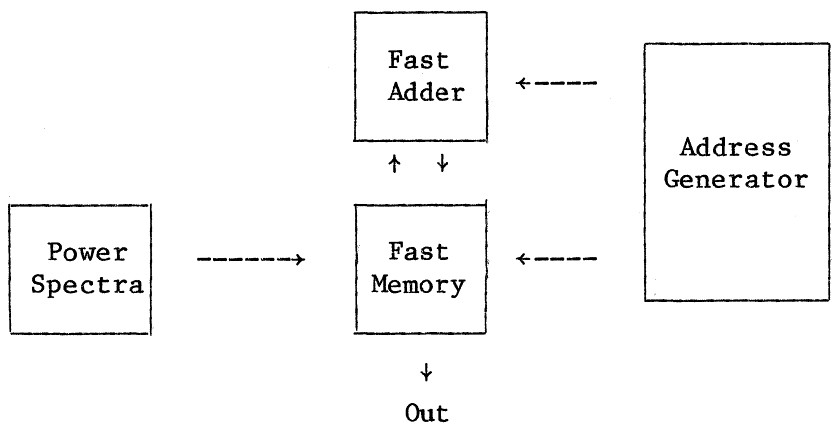
A digital delay and adder matrix could provide outputs of nD dispersions which could feed a software period search. Skipping selected delay cells D in the matrix would provide dedispersion curvature.

A software delay and adder matrix has limited speed. If a hardware matrix could be made for reasonable cost, DM and period searches could be made for short period (1-30 μsec) pulsars. The FFT, dedisperser, and time merging require a nearly continuous and wide-range of clock rates. The output rates generally are non-integer multiples of the input clock rates, so a digital-to-analog-to-digital system may be helpful.

10. DATA MERGING AND DEDISPERSING

T. H. Hankins

Consider an architectural variation where all dedispersing and merging operations are combined into one very large and very fast adder handling ~20 Mega Adds/sec.



Digital polarimetry requires additional butterflies, i.e., multiplier and adders. Multipliers are arrays of fast adders.

Fast adding architecture could accomplish

1. Dedispersing.
2. Time and frequency merging.
3. Signal averaging.
4. Real-time search in dispersion measure.
5. Corrections for Faraday rotation of polarization.

Some notes on the implementation of such an adder, drawn from an NAIC memo by T. H. Hankins dated 10 April 1981, follow. The overall design is identical to the one we are considering, but some of the operational constraints differ. An engineering design of this module has been done by

T. Catlender, at the Arecibo Observatory, who found no problem with the speed of operation or the address generation scheme.

I. Introduction

I will present below an algorithm for a device which will take care of the fast buffer, signal averager, box-car integrator and dedisperser processing needs that have been discussed. I have written it in structured FORTRAN for an unambiguous definition. In its actual execution some of the operations can be paralleled for speed; for example, the inner loop can probably be executed in 2 or 3 clock cycles. I have not included any double buffering of the memory, since it looks like all of the scientific uses anticipated can tolerate a few milliseconds "off" time for modes 1-3 which require a full memory dump. For simplicity I have specified that all loops and memory lengths be powers of 2.

I will describe the four modes separately with their limitations, then describe the input values required for each mode, and then present the algorithm.

The device can be considered as a black box with the FFT block on its input and a 16-bit bus to the computer as its output. It consists of a 1 K by 16-bit RAM and a fast arithmetic logic unit which can fetch a value from a memory location pointed to by the address register, add the current input sample to the value, store the resultant sum into the same location, and increment the address register. Some initialization, control, and output logic are also required. It is assumed that four identical units will be built for polarization measurements.

II. Modes

Mode 1: <u>Fast buffer</u>. In this mode 2^f ($1 \leq f \leq 10$) samples are read from the input and stored sequentially in memory. When the memory is full, the input is ignored and the samples are output to the computer. This mode is exactly the same as mode 2 (signal averaging) where the number of periods averaged is set to 1.

Mode 2: Signal Averager. The objective of this mode is to enhance the signal to noise ratio of a periodic signal of known period in the presence of noise. Each period the signal is sampled NF times and added into NF memory locations. After NP periods, averaging is halted, the contents of memory are output to the computer, and the memory is cleared. Timing of the sampling pulses must be controlled from outside the device, i.e., if NF samples are taken in less than a period, then the sampler must be halted until the next valid sample time.

The average obtained will have NF values where $NF = 2^f$ and $1 \leq f \leq 10$. The number of periods which can be averaged will be NP where NP = 2 and $1 \leq p \leq 10$. Caution should be observed to avoid overflow, since the sum of 2^{10} 8-bit samples could exceed the 16-bit memory word capacity.

Mode 3: Box-car integrator. The objective here is to add together NF (where $NF = 2^f$, $1 \leq f \leq 10$) adjacent samples and store the sum, then add together the next NF adjacent samples, etc., to form a smoothing filter whose time domain impulse response is perfectly rectangular. In practice the signal is oversampled by a factor NF and pNF samples are added into a single accumulation cell. After NP (where NP = 2, $1 \leq p \leq 10$) cells have been filled, the memory is dumped and cleared to make ready for the next cycle.

Mode 4: Pulsar dedisperser. In this mode the input will be connected through an analog multiplexer to a spectrometer with NF (where $NF = 2^f$, $1 \leq f \leq 6$) detected outputs. The objective is to sample the spectrometer outputs and add subsequent samples together such that a sample from filter channel 1 will be added to the sample from filter channel 2 which was taken at a later time interval equal to the dispersion sweep time of the pulsar signal from the center of channel 1 to the center of channel 2, and so forth, for all NF filters. A sum is then output after the pulsar signal contributions from all of the filters have been added together. In earlier de-dispersers this procedure had been implemented by a tapped shift register with "low numbered" filter bank channel signals being added near the beginning of the shift register and "higher numbered" filter bank channel signals being added near the end. The dispersion delay compensation was accomplished by propagating the signals down an appropriate length of the shift register delay line. In the current device, rather than shifting all the data through a shift register memory, the addresses of both the input and output will be shifted appropriately.

There is an added complication because the detector time constant, dispersion sweep time, filter rise time and sampling interval almost never optimize both signal to noise ratio and time resolution. Therefore, we would like to over-sample the detected outputs by a factor NS (where $NS = 2^s$, $1 \leqslant s \leqslant 4$), i.e., sample NS times during the dispersion sweep time between the center frequencies of adjacent filters. This is easily accomplished with little increase in complexity of the de-disperser.

The principal difference in the operation of the de-disperser and modes 1-3 is that the output is continuous, i.e., one value at a time becomes ready and must be dumped immediately because its memory location is required very soon after dumping. By contrast, for modes 1-3 the whole array of NF samples can be output simultaneously. The continuous output mode will permit convenient connection of a DAC for oscilloscope monitoring of the output.

III. Algorithm

A. Initial values: The variables given below represent the contents of the registers or counters used in the device. They must be initialized as follows:

Variable	Description	Mode 1 Fast Buffer	Mode 2 Signal Averager	Mode 3 Box-car Integrator	Mode 4 Dedisperser
NF	-	no. of samples	no. of bins per period	no. of samples in box-car	no. of filters
NS	-	n.a.	n.a.	n.a.	no. of samples per sweep time
NP	-	1	no. of periods to average	no. of bins or range gates	1
NQ	no. of words output for	NF	NF	NP	n.a.
MMAX	maximum memory address	NF	NF	NP	NF*NS
JM	current memory address	0	0	1	MMAX-NS
KS	address increment	1	1	0	NS-1
KS1	address increment	1	1	0	NS
KS2	address increment	1	1	10	NS-1
WFLAG	write during inner loop?	FALSE	FALSE	FALSE OR TRUE	TRUE

B. Constraints

$NF = 2^f, \ 1 \leqslant f \leqslant 10$

$NS = 2^s, \ 1 \leqslant s \leqslant 4$

$NP = 2^P, \ 1 \leqslant P \leqslant 10$

$MMAX, \ 2^2 \leqslant MMAX \leqslant 2^{10}$

C. Algorithm

```
LOOP
  DO                                  $ STAY HERE UNTIL COMPUTER ORDERS
  UNTIL (RUNFLG)                      $ BEGIN
FOR JF = 1, MMAX                      $ CLEAR MEMORY
 AM (JM) = 0
END FOR
WHILE (RUNFLG) DO                     $ REPEAT UNTIL RUNFLG SET FALSE
 LOOP (NP)                            $ REPEAT LOOP NP TIMES
  FOR JF = 1, NF                      $ REPEAT NF TIMES
    READ (adc) SAMPLE                 $ OBTAIN A SAMPLE FROM INPUT
    JM = JM + KS                      $ INCREMENT ADDRESS,
    IF (JM .GT. MMAX) JM = JM-MMAX    $ MODULO MMAX.
    KS = KS1                          $ RESET ADDRESS INCREMENT
    AM (JM) = AM (JM) + SAMPLE        $ ADD SAMPLE INTO MEMORY
    END FOR
    IF (WFLAG) THEN                   $ DEDISPERSION OUTPUT
      WRITE (output bus) AM(JM)       $ WRITE OUT ACCUMULATED VALUE
      AM(JM) = 0                      $ CLEAR MEMORY LOCATION
    END IF
    KS = KS2                          $ INITIALIZE ADDRESS INCREMENT
END LOOP
IF (.NOT. WFLAG) THEN                 $ MODE 1-3 OUTPUT
  FOR JM = 1, NQ                      $ REPEAT FOR EACH MEMORY LOCATION USED
  WRITE (output) AM (JM)              $ WRITE OUT ACCUMULATED VALUE
  AM (JM) = 0                         $ CLEAR MEMORY LOCATION
  END FOR
END IF
END WHILE
END LOOP
```

D. Notes

1. The memory should be cleared at the beginning.

2. The only control signal is RUNFLG. The device will process as long as RUNFLG is true. RUNFLG could come from the computer or be AND'ed with an external signal.

3. LOOP(NP) is controlled by a binary counter.

4. The incrementing of the address is the trickiest part of the algorithm. In modes 1-3 the increment is always 0 or 1, so a binary counter with $\log_2$ MMAX bits would work well. For mode 4 this counter could be clocked KS times to increment it. Since the counter is only $\log_2$ MMAX bits long, it will automatically "wrap around" to provide the correct address. When NS = 16, however, the clock that advances the address counter will have to run <u>very</u> fast. Therefore it may be more practical to use an adder to increment the address counter.

11. PULSAR AND SPECTRAL LINE COMPATIBILITY

W. D. Brundage

Contrary to the early specifications, one FFT pipeline performs the FFT for 2 quadrants by means of biplexing the input and output. With the addition of a fast memory between the ADC and FFT pipeline, a sampling frequency (40 MHz) which is twice the FFT clocking frequency (20 MHz) can be biplexed into/out of the pipeline to "double" the bandwidth. Thus 2 FFT pipelines of 9 stages each and clocking at 20 MHz maximum can provide 2 channels (polarizations) of 40 MHz bandwidth each for 2 x 1024 point spectral line observations. The same pair of FFT pipelines can be truncated to 8 stages and provided with a different set of biplexing and twiddle factors to produce 4 independent channels (quadrants) of 20 MHz bandwidth each to provide 4 x 256 point pulsar observations.

FFT cost is approximately proportional to the number of butterfly stages S which equals $\log_2 N_Q$

N_Q:	64	128	256	512	1024
N_H:	128	256	512	1024	2048
S:	6	7	8	9	10

where N = # of time samples from one ADC which are transformed into one complete frequency spectrum.

N_Q = # of points in frequency spectrum per quadrant

$N_H = 2N_Q$ = # of points in frequency spectrum per half.

Thus adding one additional stage per pipeline for spectral line capability is a small cost increment, especially considering that the FFT pipeline will cost less than the balance of the spectral processor.

The 300-foot will get a new line spectrometer and new pulsar spectrometer/ processor. Rather than duplicate the Model IV A/C and build a separate pulsar processor, we can combine the projects and get better line specifications. The OOE plan has $180 K total.

The U.S. pulsar community and 300-foot line community agreed at this meeting to combine the projects.

Advantages of the combination are:

- More capabilities than Mod IV A/C.
- Up to 40 MHz BW with 1024 CH in each of 2 polarizations (2048 CH Total).
- Resolution (CH BW) down to 78 Hz.
- $>$ 57 dB "Dynamic Range" for $\tau \lesssim 1$ hour
- Interference excision in time and frequency (increasingly important each year).
- Most efficient use of limited manpower in Green Bank.
- Best pulsar processor in existence or under design.

Disadvantages (?):

- In use at 300-foot ~85% of time

  ~50% Line

  ~25% Pulsar

  ~10% "Narrow-Band" continuum

Additional features for pulsar processing are:

- BW $\leq$ 20 MHz and 256 spectral channels/quadrant.
- Frequency resolution 78 kHz to 78 Hz
- Time resolution down to 12.3 μs
- Dedispersion
- Merging in time and frequency
- Periodic time averaging, P = 300 μs to 10 s
- Pulsar timing in 3 modes
    - Known pulse phase
    - Unknown pulse phase
    - Unknown period
- Polarimetry
- Corrections for Faraday depolarization
- Stand-alone system

Basis of this signal processor:

- Biplexed FFT pipeline per JPL 20 MHz BW, 65 K CH, spectrometer.

Our biplexed FFT pipeline has:

- Two pipelines in system
- One pipeline per 2 quads, each having BW $\leq$ 20 MHz for pulsars, or one pipeline per half having BW $\leq$ 40 MHz for spectral lines.
- Clock at $f_s \leq$ 20 MHz
- Eight stages for 256 CH/Quad for pulsar
- Nine stages for 1024 CH/Half for line

Processor is hardwired with software driven control:

- Array processor may be useful

Super micro-computer (e.g., 32-bit 68000 system) will provide:

- User I/O
- Control
- Data storage
- Data to off-line computer

Completion will require 3-4 years (~1986) <u>if</u>:

- No electronics and programmer staff loss in Green Bank
- No new major digital projects in Green Bank
- ?

		Line		Pulsar
FFT PIPE-LINE	⟶	1024 pt/40 MHz	→	(256 pt spectra/20 MHz)*2
		1024 (16 bit word)/25.6 μs		(256 (16 bit word)/12.8 μs)*2
	⟶	25.6 M word/sec		(12.8 M word/sec)*2

Total for 2 FFT pipelines is

51.2 M word/sec

= 102.4 M Byte/sec

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