NATIONAL RADIO ASTRONOMY OBSERVATORY
SOCORRO, NEW MEXICO
VERY LARGE ARRAY PROGRAM

VLA ELECTRONICS MEMORANDUM NO. 175

# DESIGN CONSIDERATIONS FOR IF RECEIVER SUBSYSTEM

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## 1.0 INTRODUCTION

The primary function of the IF receiver subsystem is to convert the 1-2 GHz IF output of the central modem subsystem (modules T1, T2) to four 0-50 MHz bands for input to the sampler/quantizers (modules D1) of the delay-multiplier subsystem.

More specifically, it is required to

1) separate the four 50 MHz-wide signal bands in the modem output from each other and from the LO signals near 1.2 and 1.8 GHz;
2) convert each such band to 0-50 MHz in such a way that any frequency in the input band can be placed at a selected frequency in the output band. This requires use of the tunable LO signals from the Master LO.
3) provide computer-selectable filters to limit the output bandwidth to a value appropriate to the sampling rate being used (rates of 0.391 to 100 MHz will be available);
4) provide about 28 dB net gain, in order to obtain +18 dBm for driving the sampler modules (much more amplification is actually needed due to various losses);
5) synchronously detect, at the final output, the 3% to 10% noise injected at the front end (it has a 50% duty cycle at a 9.6 Hz rate); and

6) maintain constant drive level to the samplers, except for the variation due to the injected noise, using a synchronously-gated ALC loop.

The IF receiver subsystem used for the first 14 of the VLA antennas met some of the above requirements and was compatible with the prototype delay-multiplier and with the original front ends. This design was described in VLA Technical Reports 3 and 5 (Thompson, 1975). A new design is now required for compatibility with the current-generation delay-multiplier (Electronics Memorandum 138, Shalloway, 1976), especially in providing the narrow bandwidths needed for spectroscopy.

## 2.0 FREQUENCY CONVERSION

### 2.1 Band-pass Sampling

The concept of band-pass sampling will be important in the following discussion, so I introduce it here. The familiar sampling theorem states that a function having a low-pass spectrum $S(f) = 0$ for $|f| \geq W$ is fully described by its samples taken at rate $2W$. We need only remark that the theorem may be generalized in that it holds whenever $S(f) = 0$ for $|f| \leq kW$ and $|f| \geq (k+1)W$, provided that $k$ is a known integer. Thus the sampling rate need only be $2W$ for band-pass signals at integral multiples of baseband, even though the highest frequency in the signal greatly exceeds $W$. It should be stressed that the theorem does not hold for non-integral $k$.

Conversion to baseband with conventional filtering and mixing can be very difficult because the unwanted sideband of the mixer becomes arbitrarily close to the desired one, and hence cannot be rejected with a filter of finite selectivity. The use of band-pass sampling thus becomes attractive.

### 2.2 Filter Q

Using band-pass sampling, a straightforward approach yields a conceptual block diagram like Figure 1, which shows

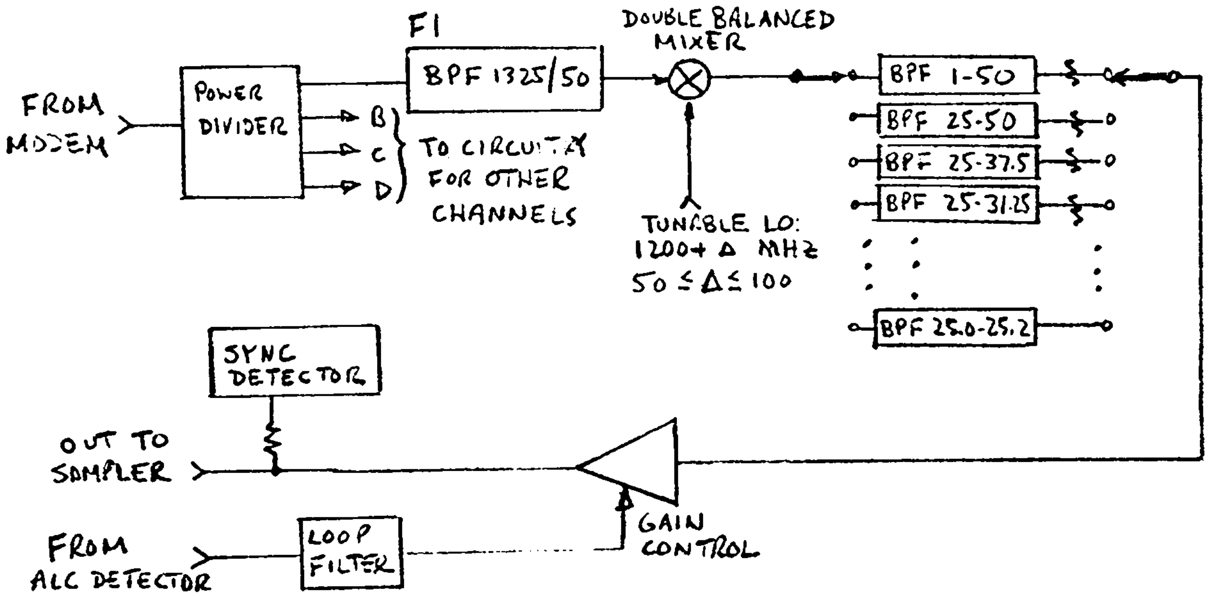


FIGURE 1: BLOCK DIAGRAM FOR BAND-PASS SAMPLING, SINGLE CONVERSION

the channel-A portion of the subsystem. In this design, the lower cutoff of the final filters was set at 25 MHz (except for the 50 MHz filter), because this corresponds to about the -30 dB point on BPF F1 when the LO is tuned to 1300 MHz, assuming that the latter filter is a practical 6-section design. (Actually, we would prefer to be able to use a lower-Q filter at F1 for better temperature stability and matching among antennas; with the Figure 1 design, expensive matched and compensated filters would be needed.) But the narrowest bandwidth final filter must then have a very high Q ($= f_{center}/\Delta f \approx 25$ MHz/.195 MHz = 128) and would be virtually impossible to construct to the required accuracy.[1] This remains true of most of the other filters.

[1]A detailed discussion of the sharpness and matching requirements for the final filters is given in a later Electronics Memorandum.

The $Q$ also affects the filter's temperature stability. For a band-pass filter not using temperature-compensating materials, the temperature coefficient of phase at the center frequency is given accurately by

$$\frac{d\phi}{dT} = \frac{\pi}{2} NQ\varepsilon$$

where $N$ is the number of sections (upper-half-plane poles) and $\varepsilon$ is the temperature coefficient of the materials used. The value is much worse near the band edges. Using an optimistic $\varepsilon = 50$ ppm/$^{\circ}$C, and taking $N = 8$ (required for sufficient sharpness) we find that $d\phi/dT < 0.2^{\circ}/^{\circ}$C implies $Q < 5.6$. The Figure 1 design must therefore be rejected.

## 2.3 Multiple Conversion Approach

To allow use of lower-Q filters, multiple conversion can be used for the narrower bandwidths. Figure 2 shows

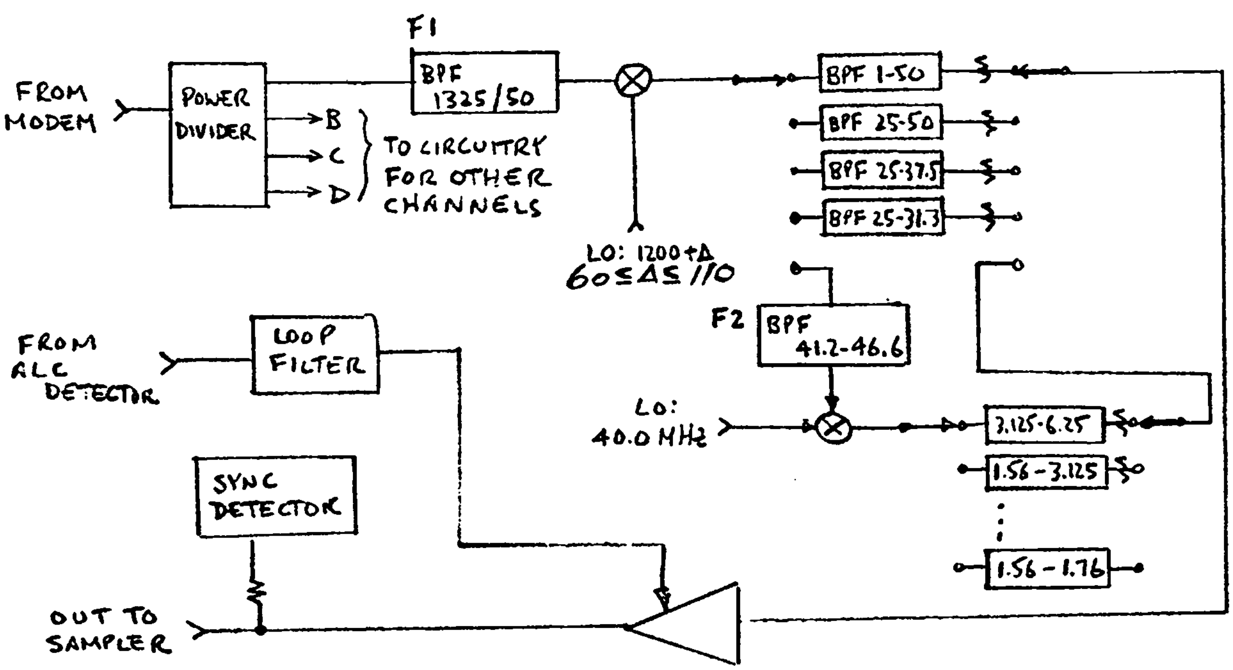


FIGURE 2: BAND-PASS SAMPLING, DOUBLE CONVERSION

a double-conversion compromise design. Here the highest $Q$ among the final filters is about 8.5, still not as low as desired but possibly acceptable with careful filter design and construction. Filter F2 must have a fairly sharp cutoff and would need to be carefully specified and temperature compensated.[2]

## 2.4 Image-Rejecting Mixer Approach

The lowest possible $Q$ for the final filters is obtained if they can all be low-pass, so that the output is strictly at baseband. This is not possible if ordinary double-balanced mixers are used, since an infinite-sharpness filter would be needed ahead of the mixer in order to reject the image band. However, by using an image-rejecting mixer (shown conceptually in Figure 3) it is possible to cancel the unwanted sideband at

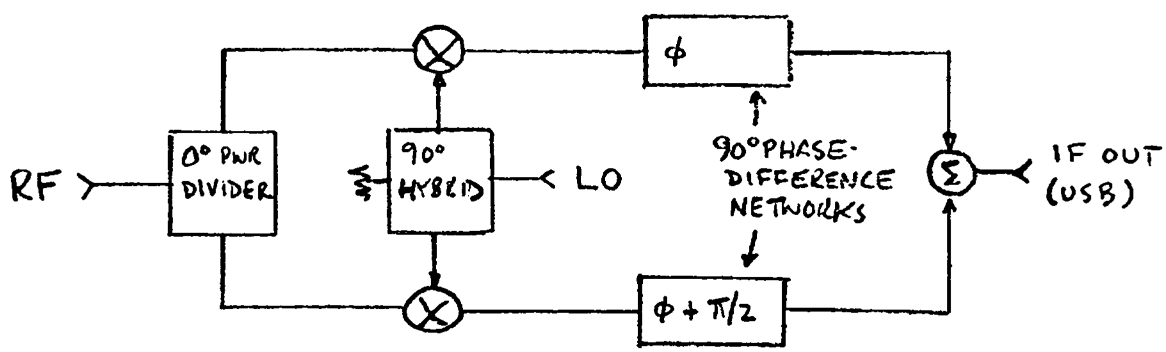


FIGURE 3: IMAGE-REJECTING MIXER

[2]When this design was under active consideration, a specification for F2 was written and proposals were sought from manufacturers (Spec. A13450N4, RFP VLA-209).

frequencies arbitrarily close to the LO, and thus to obtain the desired baseband output. A design based on such a mixer is given in Figure 4.

The use of (mostly) low-pass filters has very significant advantages, in addition to the temperature coefficient considerations discussed earlier. Only half as many components are required in a LPF as in an equivalent-selectivity band-pass, and requirements on the unloaded Q of each component are greatly relaxed. This strongly affects the physical size of the filters, as well as their cost. Furthermore, matching of the passbands of the various antennas will be important for continuum observations, and can be done much more accurately with LPF's than BPF's (cf. footnote 1, earlier).

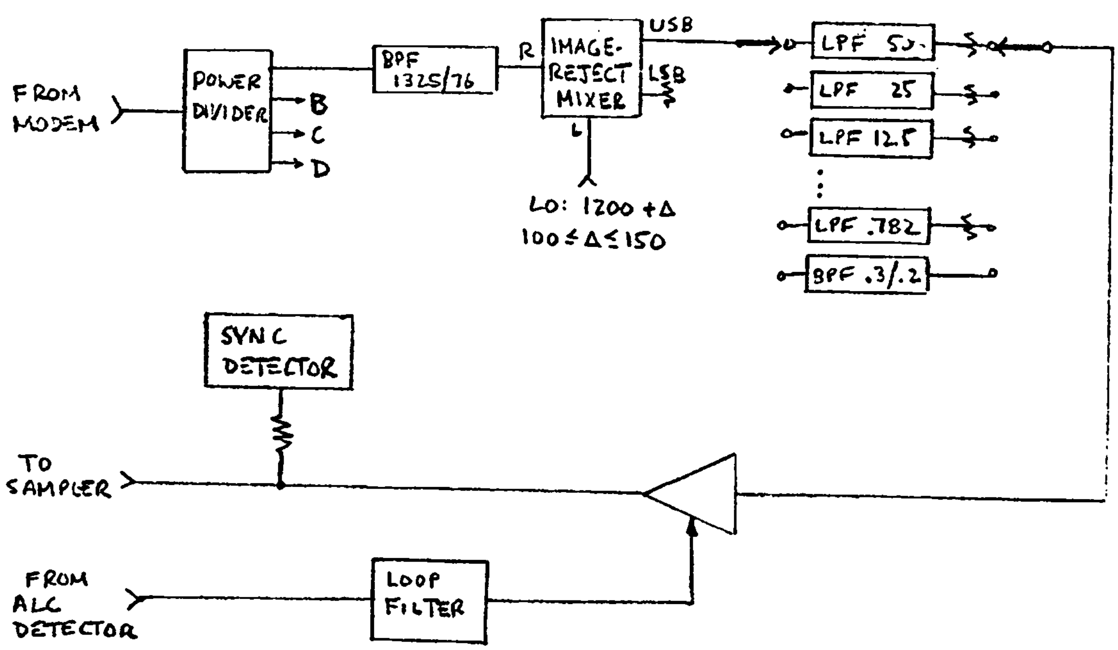


FIGURE 4: SUBSYSTEM DESIGN USING IMAGE-REJECTING MIXER

The amount of unwanted-sideband rejection depends on the accuracy of phasing and of amplitude balance in the two halves of the mixer, as shown in Figure 5. Commercially available image-rejecting mixers (I.R.M.'s) give 15 to 20 dB rejection over octave bandwidths at microwave frequencies, but this is not good enough for our purposes. The amount of rejection required is critical in deciding the feasibility of this design approach.

To determine the required rejection, consider that signals in the unwanted sideband may include (1) noise, uncorrelated between antennas, at the same level as in the desired sideband; (2) correlated broadband noise (continuum signal) from the radio source being observed; (3) correlated narrow-band noise (line signal) from the radio source; and (4) correlated discrete-frequency interference.

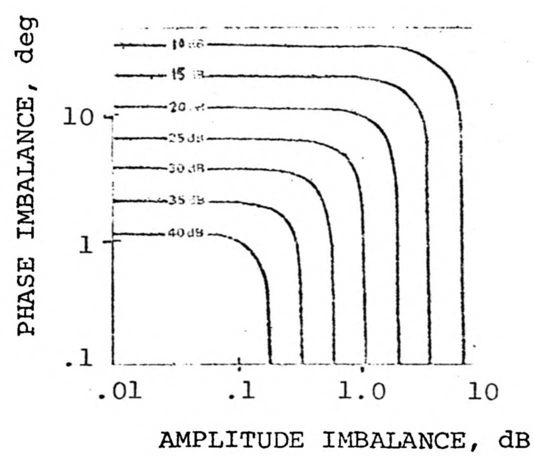


FIGURE 5: UNWANTED SIDEBAND REJECTION IN I.R.M. vs PHASE AND AMPLITUDE ERRORS

We can evaluate each of these separately:

1) Uncorrelated noise affects only the signal-to-noise ratio; a 2% loss in SNR is certainly acceptable, requiring $\geq$17 dB rejection. There is no effect on the gain calibration, since the switched noise signal appears in both sidebands.

2) Correlated noise from the unwanted sideband will interfere with that in the desired sideband in a way which varies with frequency across the IF band, and which depends on interferometer geometry. Let $f_I$ be a frequency in the IF band, let $f_o$ be the observing frequency corresponding to $f_I = 0$, and let $f_R$ be the observing frequency at which the fringe rate is brought to zero by the fringe rotators. Then the phases at the mixer output for upper and lower sideband signals are respectively

$$\phi_{USB} = 2\pi(f_o+f_I-f_R)\tau_g$$

and

$$\phi_{LSB} = 2\pi(f_o-f_I-f_R)\tau_g$$

where $\tau_g$ is the geometrical path delay from the radio source to the mixer. Thus

$$\phi_{USB} - \phi_{LSB} = 4\pi f_I \tau_g .$$

A continuum source will therefore appear to have a spectrum containing ripples whose amplitude depends on the unwanted sideband rejection and whose period is $(2\tau_g)^{-1}$. For the VLA, $0 \leq \tau_g \lesssim 70$ μsec (referenced to the array center), so the ripple period is 7 kHz to infinity. Since this effect would be difficult to calibrate out, and since we desire channel-to-channel relative accuracy of better than 0.1% for line observations, we require $\geq$30 dB rejection; this keeps the ripple amplitude $\leq$0.1% (peak).

3) If a strong spectral line appears in the undesired sideband, the amount of rejection required to render it undetectable depends on the strength of the line, on whether or not it is resolved in frequency, and on the bandwidth and integrating time used. No definite amount of required rejection can be established. Fortunately, the astronomer can distinguish lines in the desired and undesired sidebands by noting their direction of movement at the IF when the LO frequency is changed slightly. We therefore assume that the 30 dB rejection required for 2) will suffice here also.

4) Interfering signals can in principle be at arbitrarily high levels, so as much rejection as possible is desired. But the unwanted sideband is now immediately adjacent to the desired band, and such close-by interference cannot be strongly supressed in any design. Note that the final filter supresses interference in either sideband if it converts to an IF outside the filter's bandwidth. Thus we again assume that 30 dB rejection will suffice.

Can a $\geq$30 dB rejection mixer be constructed? It is somewhat difficult to do with inputs in the 1-2 GHz band, but there is strong evidence that it is possible. With the Figure 4 design, the channel-A mixer must operate at fractional bandwidths of (1350-1300)/1325=3.8% on the LO and (1350-1250)/1200=8.3% on the RF input. Errors of less than $1^{\circ}$ in phase and 0.1 dB in amplitude should easily be possible over such narrow bandwidths, and may even be possible in a single device covering all four channels. The main difficulty may be compensating for mismatch of the mixer diodes.

The IF phase shift networks cannot be made to operate accurately down to zero frequency, but they can come quite close. Figure 6, by John Granlund of NRAO, shows the total number of poles required (both networks) to achieve a given rejection as a function of $R$, the ratio of maximum to minimum frequency. For $R = 50\ \text{MHz}/0.1\ \text{MHz} = 500$, 8 poles should suffice. Although the literature on broadband $90^{\circ}$ phase difference networks is somewhat limited (e.g., Rogers, 1971; Albershein and Shirley, 1969), NRAO has considerable experience in their theory and construction, especially in the VLA Sampler Module (Mauzy and Escoffier, 1976, Electronics Memorandum 132).

## 3.0 LOW FREQUENCY CUTOFF AND IMPLICATIONS FOR SAMPLER MODULE

Operation at baseband for all the desired bandwidths depends on making the low-frequency cutoff sufficiently close to zero. The accuracy of the I.R.M. IF networks affects this limit, but there are additional considerations.

First we want to avoid extensive redesign of the sampler/quantizer module (D1) which follows the IF receiver, and which was originally intended to cover a 1 MHz to 50 MHz passband. The sampler has two (digital) output channels, known as sine and cosine, which ideally are Hilbert transforms of each other when the input signal is in the design passband. Also, the transfer function from the input to either output is ideally the same for any two samplers (connected to different antennas). Now, the digital correlator will be able to operate in various modes (see Electronics Memorandum 138, Shalloway, 1976); the "continuum" modes use both sampler outputs and depend on the Hilbert transform and matching being accurate over a large fraction of the bandwidth actually in use; the "spectral line" modes use only one output of each sampler, so that the Hilbert transform accuracy is not

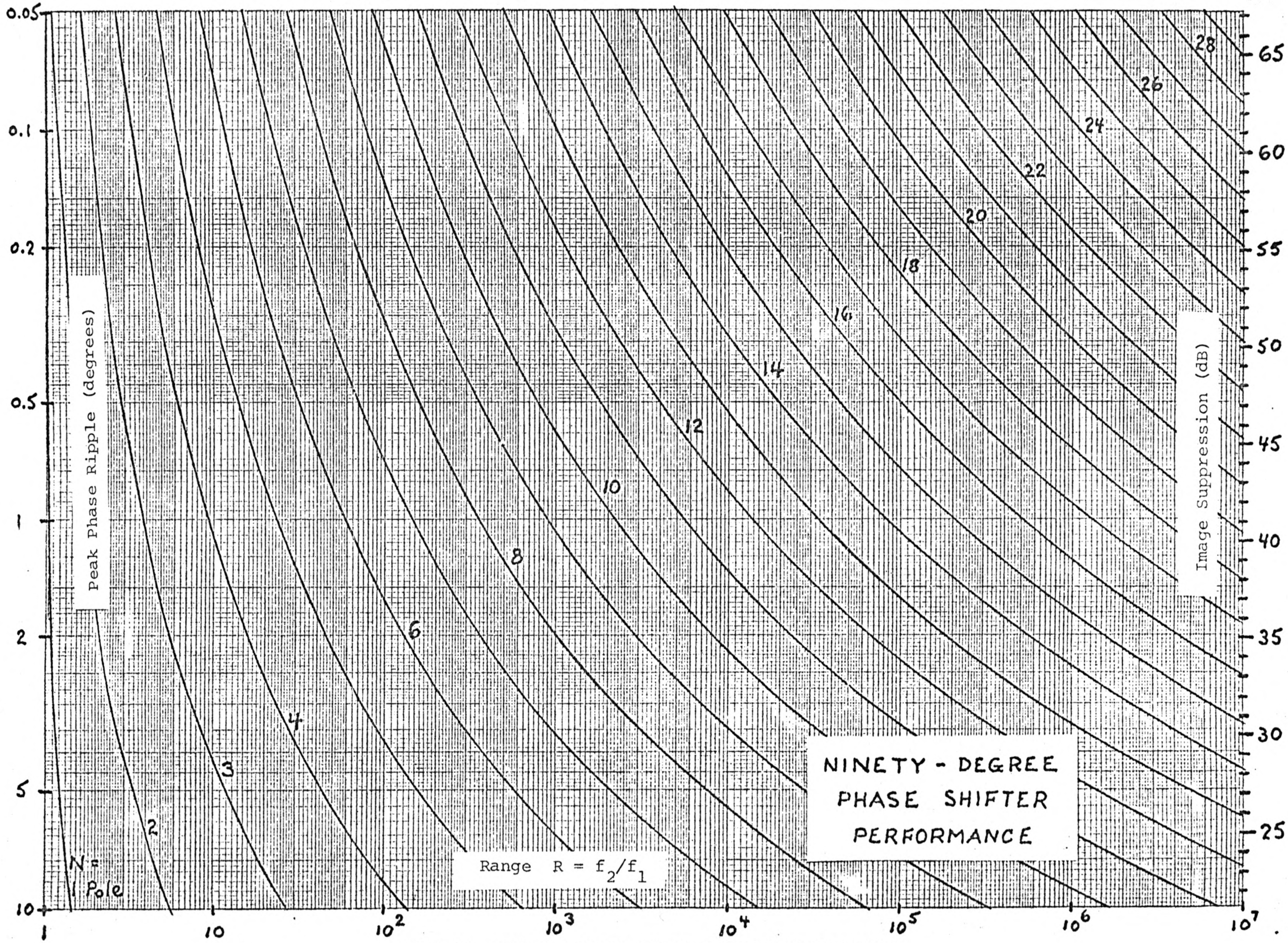


FIGURE 6: NINETY-DEGREE PHASE SHIFTER PERFORMANCE

important, but matching and reasonably flat response across the band in use are required.

The Hilbert transform accuracy depends mainly on an 8-pole $90^{\circ}$ phase difference network similar to that required for the I.R.M.'s; it is designed for $<0.2^{\circ}$ phase error from 0.9 to 55 MHz, and has an all-pass response. The amplitude transmission to each output is determined mainly by an input coupling transformer and several interstage coupling capacitors in amplifiers. We can thus consider the following approach: extend the low-frequency response as far as is practical by replacing the input transformer and coupling capacitors, and leave the $90^{\circ}$ phase difference network unmodified. Tests made to date (by W. E. Dumke) indicate that 0.1 MHz is a practical limit, caused mainly by the transformer. For continuum modes, this will introduce errors in the visibility measurements which are larger at the narrower bandwidths (assuming baseband signals). Rough calculations show that <1% error will be achieved only with bandwidth $\geq$25 MHz if $f_{MIN} = 0.1$ MHz, and bandwidth $\geq$12 MHz if $f_{MIN} = 0.2$ MHz. For bandwidth <12 MHz, operation in continuum mode rapidly becomes inaccurate, but restriction to line mode for such bandwidths is considered acceptable.[3] Nevertheless, better continuum-mode accuracy may be needed, in which case some possible approaches are: (1) cause the IF receiver to have a higher low-frequency rolloff when wide bandwidths are selected; and (2) replace the $90^{\circ}$ network with one having a wider bandwidth, but retaining 8 poles and hence just changing the element values.

Certain components in the IF receiver subsystem also limit the practical low-frequency cutoff. These include diode switches used for filter selection, a voltage-controlled attenuator required for ALC, and ac-coupled amplifiers (dc coupling is not practical because of the high gain required).

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[3] cf. memo, D'Addario to VLA Steering Committee, July 22, 1977.

From all these considerations, a low-frequency limit of 0.195 MHz (half the lowest sampling rate) seems reasonable for the IF receiver output. More precisely, consider the specification:

> For output frequencies from 0.195 to 50.0 MHz, the IF receiver subsystem shall have an image rejection of $\geq$30 dB and shall have a transfer function not significantly affected by components other than the selected filter.

With this lower limit, not all of the available sampling rates can be covered at baseband. The scheme shown in Figure 4 uses band-pass sampling (and hence a BPF) only for the narrowest bandwidth; further discussion of this scheme is given in Section 5.0. Finally, it should be apparent that only the I.R.M. approach will satisfy the above specification, so the remainder of this report assumes that that approach is adopted.

## 4.0 FURTHER DESIGN DETAILS, SIGNAL LEVELS, AND SPURIOUS RESPONSES

A more detailed block diagram of the image rejecting mixer scheme is given in Figures 7(a) and 7(c). Figure 7(b) gives an alternative arrangement in which double conversion is used to allow the I.R.M.'s to operate at lower frequencies. At this writing, prototype development is being pursued for both arrangements. Current plans call for splitting the signal processing into three module types, the first of which would do the frequency conversion for all four channels (T3:IF to Baseband Converter), and the others of which would contain all the baseband circuitry for a single channel (T4:Baseband Filters, T5:Baseband Driver). A separate control and interface module (T6:Baseband Control) then completes the subsystem.

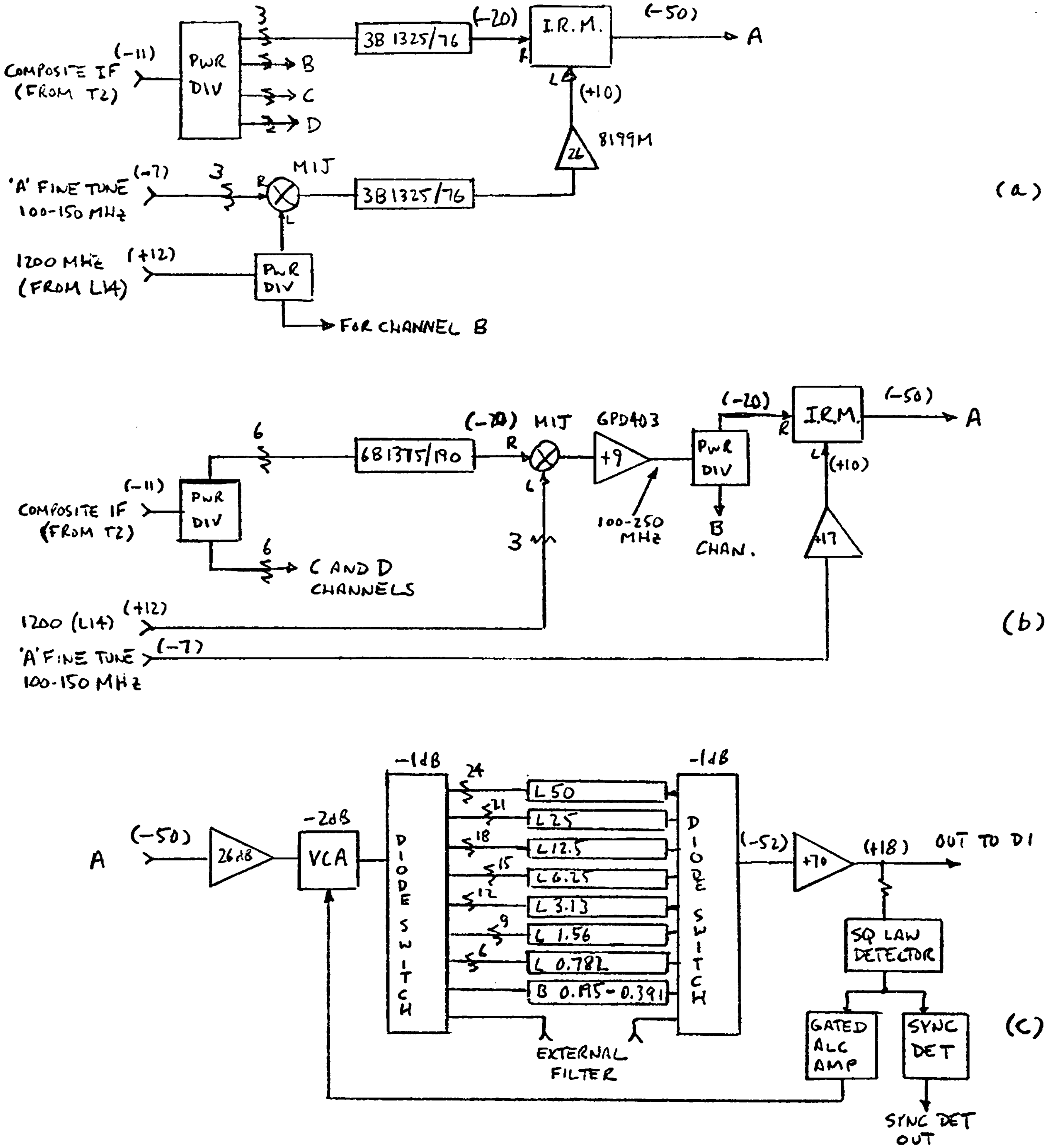


FIGURE 7: IMAGE REJECTING MIXER APPROACHES, SHOWING DETAILS FOR CHANNEL A. (a) Single conversion; (b) double conversion alternative; (c) baseband circuitry. Numbers in parenthesis are desired signal powers in dBm.

It is important to consider possible spurious responses in each arrangement. It turns out that the most serious potential problems occur because of the presence of the 1200 and 1800 MHz LO signals and their side frequencies in the composite IF. Suppression of these signals is the main purpose of the band-pass filters in Figures 7(a) and 7(b). The following discussion concentrates on channel A, but similar considerations apply to the other channels.

In Figure 7(a), the 3B1325/76 band-pass filter provides about 20 dB of rejection at 1200 MHz. The spectrum reaching the I.R.M. consists of

| | |
|---|---|
| IF channel A | -20 dBm |
| 1200 MHz | -41 dBm |
| 1195, 1205 MHz | -51 dBm |
| IF channel B | ~-39 dBm. |

We can imagine two mechanisms for producing undesired outputs from the I.R.M. at 0 to 50 MHz: (1) 5 MHz and 10 MHz outputs due to intermodulation in the I.R.M.; and (2) channel B signal at the I.R.M. R-port mixing with spurious 1400-1500 MHz at I.R.M. L-port, the latter being generated in the double balanced mixer (M1J) as $2f_R + f_L$. The first mechanism is a $2f_R + 0\ f_L$ product in the I.R.M.; we can estimate its magnitude by guessing that the second-order intercept point of the I.R.M. is $P_{int2} = +15$ dBm, whence

$$P_{spur}(5\text{ MHz}) = \frac{P_{1200} \cdot P_{1195}}{P_{int2}} = -42\text{ dBm} + (-52\text{ dBm}) - (+15\text{ dBm})$$

$$= -109\text{ dBm.}$$

If the spur due to 1205 and 1200 MHz adds in phase, the total is -103 dBm. These results are referred to the I.R.M. Report, where the 50-MHz wide channel A signal is -20 dBm.

Mechanism (2) results in extra uncorrelated noise in the output due to channel B, but at a very low level. The M1J intermodulation chart shows typical suppression of $2f_R + 0f_L$ as -61 dBc; conservatively, we can assume -50 dBc. If the fine tuning is set to 100 MHz (the worst case), the undesired signal is at 1400 MHz and the BPF provides less than 3 dB rejection. Nevertheless, the undesired noise will be at least -60 dB with respect to channel A.

In Figure 7(b), mechanism (1) above can occur in the first mixer, and the resulting 5 MHz could appear in the baseband output due to the finite R to I isolation of the I.R.M. In this case, a sharper 6-section filter has been used ahead of the first mixer so as to be 25 dB down at 1200 MHz. Assuming a second-order intercept of +15 dBm for this mixer, we find

$$P_{spur}(5\text{ MHz}) = \frac{P_{1200} \cdot P_{1195}}{P_{int2}} = (-47\text{ dBm}) + (-57\text{ dBm}) - (+15\text{ dBm})$$

$$= -119\text{ dBm}.$$

Adding in the 1205 MHz side frequency and allowing 20 dB R to I isolation in the I.R.M. gives an effective level for this spur of -119+6-20 = -133 dBm.

Additional suppression of such spurs is obtained from the sampler phase switching and from fringe rotation. The latter is not significant at short baselines, so we neglect it here. Calculations show that the former should give about 20 dB suppression.

How small must such spurs be kept? To make them undetectable, we should have

$$P_{spur} < P_{total}\ (TW)^{-\frac{1}{2}} = S_{total}\ (W/T)^{\frac{1}{2}}$$

where $T$ is the integrating time, $W$ is the channel bandwidth, and $S$ is the power spectral density. The smallest total bandwidth in which the 5 MHz spur can occur is 6.25 MHz, in which case the smallest channel bandwidth is $W = 6.25$ MHz/256. Taking $T = 12$ hours and $S_{total} = -20$ dBm/50 MHz, we find

$$P_{spur} < -98.2\text{ dBm}.$$

The calculations of this section are summarized in Table I, which also gives the results of similar calculations for the 500 kHz modulations of the 1800 MHz carrier appearing in channel D.

TABLE I: SUMMARY OF SPURIOUS RESPONSE CALCULATIONS
(all signal powers referenced to I.R.M. input)

| | 5 MHz in Channel A | 500 kHz in Channel D |
|---|---|---|
| Predicted, single conversion (Fig. 7a) | -129 dBm | -134 dBm |
| Predicted, double conversion (Fig. 7b) | -139 | -144 |
| Required (see text) | <-98 | <-104 |

## 5.0 IMPLICATIONS FOR CORRELATOR OPERATION

The digital delay-multiplier subsystem will have the capacity to compute 1024 correlation coefficients per antenna pair at sampling rates of 3.125 MHz or less, and proportionally fewer at higher sampling rates. This provides for up to 512 complex frequency channels, of which only 256 will be retained for further processing. The frequency channels must cover any of ten user-selected bandwidths $W = (50\text{ MHz})\cdot(2^{-K})$, $K = 0, 1, \ldots, 9$. (In correlator modes in which more than one IF channel is processed, the total number of spectral channels is divided among the IF pairs processed.)

From these numbers it should be apparent that, at the narrower bandwidths, the correlator has considerable extra capacity. It has long been planned (Electronics Memorandum No. 138) that some of this extra capacity would be used by sampling at more than twice the input bandwidth, thereby obtaining a slight signal-to-noise ratio improvement. It is now proposed that we use an additional portion of this extra capacity to simplify the filter requirements in the IF receiver. In this way it is possible to include only eight different filters, or correlator input bandwidths, while retaining the ten desired bandwidths at the correlator output. If this approach were not taken, it would be necessary to use high-Q band-pass filters for several of the narrowest bandwidths, in view of the low-frequency cutoff limit discussed in Section 3.0. As it is, we can get away with only one band-pass filter, and its $Q$ is less than 2. The details of one scheme for accomplishing this are contained in Table II. Some points to note about this table are the following: (1) FFT's of length = 512 are required for the three narrowest bandwidths, whereas in the earlier scheme the maximum length was 256; it appears, however, that this is well within the available computing capacity. (2) In two cases, namely $W_{req}$ = .391 and .097 MHz, the filter bandwidth exceeds the requested bandwidth; all 256 channels are then usable, since none need be near the band edges. (3) Oversampling is still obtained for the four narrowest bandwidths. (4) In 2-band and 4-band (or

TABLE II: CORRELATOR NUMBERS FOR SINGLE-BAND LINE MODE

| $W_{req}$ MHz | $f_u$ MHz | $f_l$ MHz | $f_s$ MHz | $\Delta\tau * f_s$ | No. Lags | FFT Length | Total Channels | Usable[4] Channels | Channel BW, kHz |
|---|---|---|---|---|---|---|---|---|---|
| 50 | 50 | (1) | 100 | 1 | 32 | 16 | 16 | 14 | 3125 |
| 25 | 25 | (1) | 50 | 1 | 64 | 32 | 32 | 29 | 781 |
| 12.5 | 12.5 | (1) | 25 | 1 | 128 | 64 | 64 | 57 | 195 |
| 6.25 | 6.25 | (1) | 12.5 | 1 | 256 | 128 | 128 | 111 | 48.8 |
| 3.125 | 3.125 | (1) | 6.25 | 1 | 512 | 256 | 256 | 214 | 12.2 |
| 1.563 | 1.563 | (1) | 3.125 | 1 | 1024 | 256 (2) | 256 | 198 | 6.1 |
| 0.781 | 0.781 | (1) | 3.125 | 2 | 1024 | 256 (2) | 256 | 166 | 3.05 |
| 0.391 | 0.781 | (1) | 3.125 | 2 | 1024 | 512 | 256 (3) | 256 | 1.53 |
| 0.195 | 0.391 | 0.195 | 1.563 | 2 | 1024 | 512 | 256 (3) | 205 | .763 |
| .097 | 0.391 | 0.195 | 0.781 | 2 | 1024 | 512 | 256 (3) | 256 | .381 |

NOTES: (1) Lower limit determined by amplifier rolloffs; approximately 0.19 MHz.

(2) Only alternate lags are processed by the FFT in these cases.

(3) All lags are processed, but half of the FFT output points are discarded.

(4) "Usable" channels assumes lower cutoff of 195 kHz and loss of 10% at filter band edges.

$W_{req}$: requested bandwidth

$f_u$ : filter upper cutoff frequency

$f_l$ : filter lower cutoff frequency

$f_s$ : sampling frequency

$\Delta\tau$ : lag step

FFT length: number of complex output points polarization) modes, the situation for each IF pair can be determined by adjusting the single-band numbers in Table II by factors of 2 and 4, respectively.

## 6.0 AUTOMATIC LEVEL CONTROL AND SYNCHRONOUS DETECTOR

The correlator depends, for accurate measurements of visibility amplitudes, on holding the power supplied to each sampler at a predetermined level, and on knowing the gain from the antenna to the sampler. The first requirement can be met with an ALC loop whose detector is as close to the sampler as possible; the parameters of this loop are discussed below. The second requirement can be met by synchronously detecting the switched noise signal injected at each front end. This detection should also be done as close as possible to the sampler, and the same detector may be used for both this function and ALC. However, in order to measure the switched signal accurately, the ALC loop must either be very slow compared to the switching period, or it must be synchronously gated.

### 6.1 ALC Loop Characteristics

The correlator is designed to operate with the sampler thresholds set at $\pm 0.612\ \sigma$, where $\sigma$ is the RMS signal voltage. At this level, maximum signal-to-noise ratio is achieved (Cooper, 1970). A small deviation from this optimum results in a small, not very significant reduction in SNR; but, if uncorrected, it can also result in significant errors in the measured correlation coefficients. To effect a correction for this, the correlator incorporates self-multipliers on the output of each sampler; these provide a measure of the signal level error relative to the sampling threshold. Recent calculations (J. Granlund, private communication) show that errors of up to 5% in voltage can in this way be corrected to the 0.1% level. We therefore require that the ALC loop maintain the correct signal level, relative to the sampler thresholds, to

±5% (±0.4 dB). If the sampler thresholds and ALC set point are adjusted separately, then this represents the combined error. It also represents the total of the initial setting error and the drift with time and temperature.

The time constant of the ALC loop should be long in order to avoid introducing too much of the detector noise into the signal path, and possibly to allow the switched signal to be detected without ALC gating. On the other hand, the time constant should be short in order to minimize the settling time required when the bandwidth is changed. For the first consideration, it can be shown that the fractional increase in noise introduced by an ALC loop for a Gaussian signal is $(W\tau)^{-1}$ where $W$ is the RF bandwidth and $\tau$ is the loop time constant. This is less than 0.1% for $\tau > 5$ μsec, even at our smallest bandwidth. However, if we want to avoid ALC gating and have the ALC loop attenuate the switched signal by less than 0.1% (in RF power, or detected voltage), we need $\tau > 0.5$ sec. This should result in an acceptable settling time. Nevertheless, a gated loop will probably be required to handle the complete loss of signal during command time.

6.2 <u>Synchronous Detector Characteristics</u>

The switched noise power will normally be in the range 3% to 10% of the total power, and will be on during alternate 19.2 Hz waveguide cycles. We desire to measure its amplitude to better than 0.5% RMS, or to better than .015% of the total power. With no circuit errors or detector errors, this requires $W\tau > 2(.00015)^{-2} = 9 \times 10^7$, or $\tau > 490$ sec at $W = 180$ kHz, but only $\tau > 1.8$ sec at $W = 50$ MHz. The former would produce excessively long settling times, but is unavoidable if this accuracy is to be achieved. We therefore propose to depend on the synchronous computer to provide most of this integration. The integration provided in the analog circuitry associated with each receiver should be an RC time constant of 0.5 to 2 sec, appropriate to a data set sampling period of about 0.62 sec.

The requirements for accuracy in the detector circuitry may be studied in several steps. If we suppose that the total power into the detector during the time that the noise source is off is held perfectly constant by the ALC loop, then the synchronous detector must be capable of monitoring variations in the switched power to a few parts per thousand. This requires that the detector diode's voltage-out vs power-in curve (including any associated dc amplifier) have a stable slope in the vicinity of the operating point. The temperature coefficient of this slope should be $<0.1\%/^{\circ}C$, since temperature variations of a few $^{\circ}C$ can be expected. Also, the slope of the chord of this curve between the noise-source-off point and the noise-source-on point should be constant to 0.1% as the switched noise varies from 1% to 10% of the total (see Figure 8). These requirements

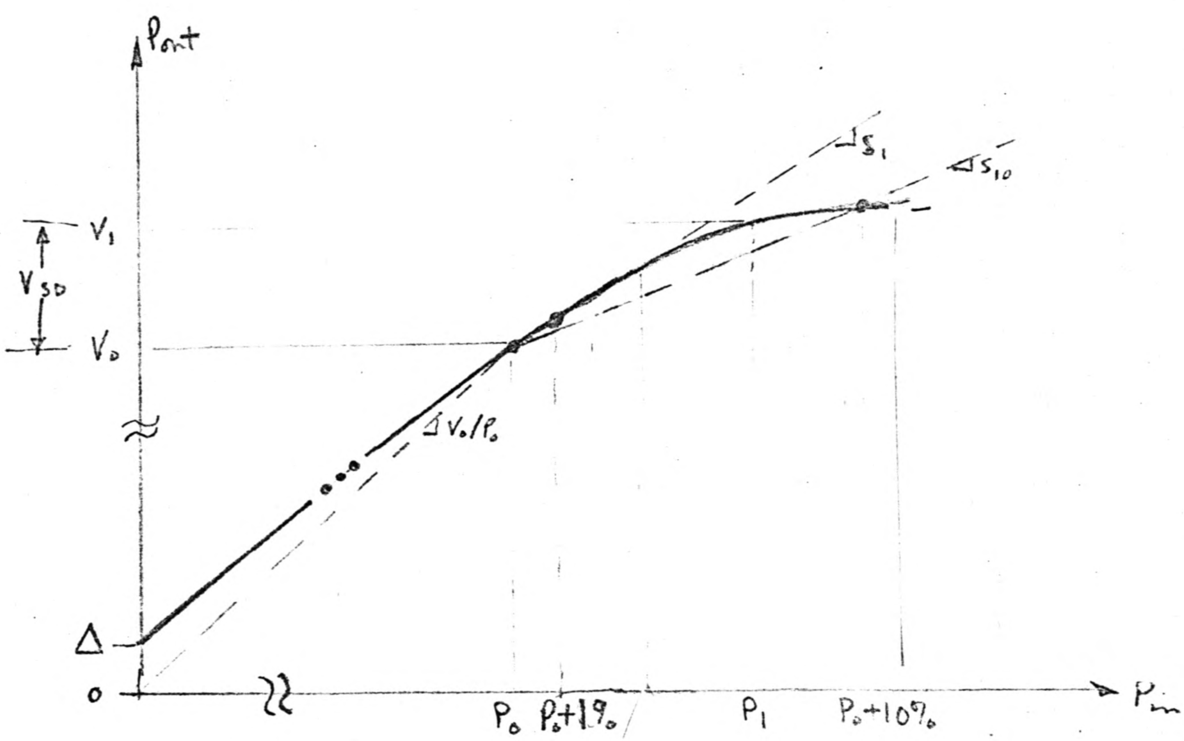


FIGURE 8: DETECTOR LAW SPECIFICATIONS

can be summarized as follows:

$$\frac{1}{S_{10}} \frac{dS_{10}}{dT} < .001\ {}^{o}C^{-1} \quad \text{and} \quad \frac{S_1 - S_{10}}{S_1} < .001.$$

where $S_1$, $S_{10}$ are the slopes of the on-off chord for 1% and 10% switched noise, respectively; and $T$ is temperature. Next, if we relax the assumption that the total power is perfectly constant, but suppose that the same detector is used both for ALC and for measuring the switched power, then we have the following additional requirement on the detector:

$$\frac{1}{V_0} \frac{dV_0}{dT} < .001\ {}^{o}C^{-1}$$

where $V_0$ is the output voltage for constant input power at the nominal operating point. Finally, if a separate detector is used for ALC and if it is allowed to drift so that the total power varies by ±5%, then we need to impose the following additional requirement on the detector used for the switched power measurement:

$$\Delta < .02\ V_0$$

where $\Delta$ is the dc offset of the detector (output for zero input power). This ensures that the ratio $V_{SD}/V_0$ is stable to ±0.1% for variations of ±5% in $V_0$, where $V_{SD}$ is the synchronous detector output. If appropriate corrections for the total power error are made in the correlator, then only this ratio need be accurately measured.

## 7.0 REFERENCES


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