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SIGNAL PROCESSING FOR BEAM POSITION MEASUREMENT

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Abstract

The spectrum of the signals generated by beam position monitors can be very large. It is the convolution product of the bunch spectrum and the transfer function of the monitor including the transmission cable. The rate of information flow is proportional to the bandwidth and the maximum amplitude rating of monitor complex. Technology is progressing at a good pace and modern acquisition capabilities are such that nearly all the information contained in the spectrum can be acquired with a reasonable resolution [1]. However, the cost of such a system is enormous and a major part of the information is superfluous. The objective of a beam position measurement system is generally restricted to trajectory measurements of a portion of the beam that is much larger than the finer details that can be observed with the bare signal generated by the position monitor. Closed orbit measurements are a simple derivation product of the trajectory and will not be considered further. The smallest beam portion that is of practical interest is one bunch. Hence the maximum frequency is in the order of the bunch repetition rate. Lower frequencies than the bunching frequency may be chosen either to obtain better resolution, either because it is technically easier to accomplish. The sensitivity of beam position monitors degrades quickly at low frequencies. Therefore, signals are selected at some convenient multiple of the bunching frequency and are shifted to so called baseband to match the capabilities of the acquisition system. The task of signal processing is to make a selection among the many frequencies that are available and prepare the signals for acquisition. The signal selection is done by filtering, a vast subject but it will not be treated in this paper. Three signal processing techniques will be examined from the point of view of (amplitude) resolution of a single acquisition of the beam position, dynamic range and operational frequency. They are the following: the homodyne receiver, the phase processor and the logarithmic detector. Baseband techniques are also used in practice and will be briefly mentioned to start.

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1. Introduction

The spectrum of the signals generated by beam position monitors can be very large. It is the convolution product of the bunch spectrum and the transfer function of the monitor including the transmission cable. The rate of information flow is proportional to the bandwidth and the maximum amplitude rating of monitor complex. Technology is progressing at a good pace and modern acquisition capabilities are such that nearly all the information contained in the spectrum can be acquired with a reasonable resolution [1]. However, the cost of such a system is enormous and a major part of the information is superfluous. The objective of a beam position measurement system is generally restricted to trajectory measurements of a portion of the beam that is much larger than the finer details that can be observed with the bare signal generated by the position monitor. Closed orbit measurements are a simple derivation product of the trajectory and will not be considered further. The smallest beam portion that is of practical interest is one bunch. Hence the maximum frequency is in the order of the bunch repetition rate. Lower frequencies than the bunching frequency may be chosen either to obtain better resolution, either because it is technically easier to accomplish. The sensitivity of beam position monitors degrades quickly at low frequencies. Therefore, signals are selected at some convenient multiple of the bunching frequency and are shifted to so called baseband to match the capabilities of the acquisition system. The task of signal processing is to make a selection among the many frequencies that are available and prepare the signals for acquisition. The signal selection is done by filtering, a vast subject but it will not be treated in this paper. Three signal processing techniques will be examined from the point of view of (amplitude) resolution of a single acquisition of the beam position, dynamic range and operational frequency. They are the following: the homodyne receiver, the phase processor and the logarithmic detector. Baseband techniques are also used in practice and will be briefly mentioned to start.

2. Classification of processing techniques

The signal of a beam position monitor electrode consists of two parts. One part is proportional to the beam intensity and the other one is proportional to the relative beam position. The signal $E_{1,2}$ of electrode *1* or *2* can be expressed as :

$$E_{1,2} = \frac{\Sigma \pm \Delta}{2} \tag{1}$$

where Σ is the sum of E_1 and E_2 and Δ is their difference. The characteristic of the central region of most monitors is linear over about one third of the geometrical radius. Special split electrode electrostatic pick ups are linear over the full aperture. The following relation between electrode signals and beam position holds in the linear approximation:

$$\frac{x}{A} = \frac{\Delta}{\Sigma} = \frac{E_1 - E_2}{E_1 + E_2}, \tag{2}$$

where x is the beam position and A the normalizing radius of the beam position monitor. The radius A is half the geometrical radius except for the linear electrostatic pick up where it is equal to the geometrical radius. From (1,2) it can be seen that a mathematical operation on the measured quantities $E_{1,2}$, or Δ and Σ, is necessary to derive the beam position *x*. That operation is performed before the acquisition in the *normalizing processors* (phase processor and logarithmic processor) so that only one value per pair of electrodes is digitized. When an integrator or a homodyne receiver are used two values are acquired and the mathematical operation is performed in the software afterwards.

3. Baseband processor

A serious difficulty with the acquisition of the unprocessed signals from a position monitor is the wide range of frequencies that they contain. Passing the signal through a low pass filter before acquisition is a straightforward solution of this problem. That strategy is chosen for RHIC [2]. One of its disadvantages is the important Noise Factor (*NF*) of ~30 dB. It is due to the large bandwidth of the digitizer circuit and makes this technique less interesting in terms of optimum resolution.

The integrator is slightly more complex than the low pass filter as can be judged from Figure 1.

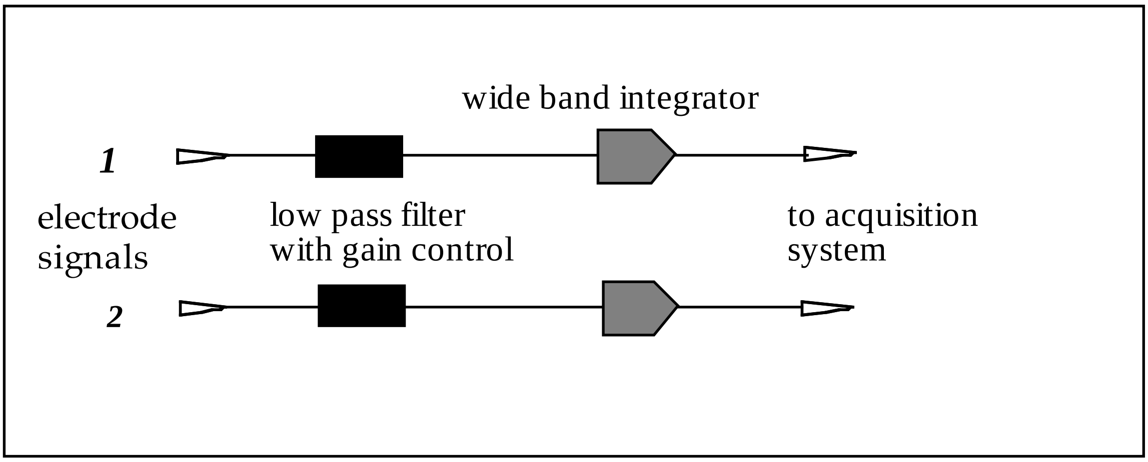

Figure 1 : Principle of integrator processor.

Examples of this processor type can be found in [3,4]. It is used in LEP with a low pass filter cut-off frequency of 57 MHz. It may be worthwhile to take a closer look at the LEP wide band integrator.

The consequence of the integration is an apparent enhancement of the noise factor of the input amplifier. That can be seen as follows.

The integration introduces a transfer function of the form :

$$\frac{1}{\omega\tau_{\text{int}}}, \tag{3}$$

where τ_{int} is the integration scale factor. The average gain of the integrator over bandwidth ω is

$$\frac{1}{\omega\tau_{\text{int}}}\int_{\omega_{rev}}^{\omega}\frac{d\omega}{\omega}=\frac{\ln(\omega/\omega_{rev})}{\omega\tau_{\text{int}}}. \tag{4}$$

The monitor transfer function at low frequencies can be written in general as :

$$Z_t = Z_0 \frac{j\omega\tau}{1 + j\omega\tau} \approx Z_0 j\omega\tau, \qquad (5)$$

where τ is a constant depending on the monitor.

The effective impedance after integration then becomes $Z_0 \tau/\tau_{\text{int}}$. The *signal gain* of this operation is thus $\sim 1/\omega\tau_{\text{int}}$ taking the cut-off frequency of the low pass filter as reference. The relative *gain* of the *noise* signal with respect to the beam signal is

$$\ln(\omega/\omega_{rev}). \qquad (6)$$

This effect is responsible for an increase of 19 dB in noise factor in the LEP wideband system. The total noise nactor is expected to be *NF*=22..30 dB which fits with observations [5].

The performance of the integrator is essentially determined by its bandwidth, the enhanced noise factor of the filter amplifier and the maximum signal that it can digest. A particular difficulty is the fact that the signal to be digitized is dominated by the beam *intensity*. The fraction of the signal that carries the beam *position* information is small compared with the intensity. The dynamic range can be defined as the ratio between the maximum signal and the resolution. To keep the position resolution at a reasonable level a very large dynamic range is needed in the digitizer. The total dynamic range required in practice can be so large that it is split into several levels with the help of programmed pre-amplifiers.

3.1 Discussion

The integrator is a wide band device. It operates from DC to the cut-off frequency of the low pass filter, typically several tens of MHz. The minimum input signal levels ($\Sigma/2$) is -61 dBm assuming a 50 MHz bandwidth for example. The lower signal level is a hard limit in the sense that it defines the absolute limit for resolution. If a minimum resolution of 10% of the aperture is required then the practical lower limit of the signal is -44 dBm in this particular case. The resolution increases to 71 dB (0.03% of the aperture) when the maximum signal (~ 10 dBm limited by electronics) is reached. The practical dynamic range in terms of beam intensity is then 51 dB.

4. Homodyne receiver

Signals detected at some multiple of the bunching frequency are demodulated by the homodyne receiver into baseband so that they can be digitized. The primary high frequency signals permit simple signal manipulations with passive hybrid circuits. Advantage is taken from this possibility and the two electrode signals are transformed into a *sum* (Σ) and a *difference* (Δ) signal eliminating to a great deal the problems with dynamic range that are encountered with baseband systems.

The principle of the homodyne receiver can best be explained with the help of the diagram of the 200 MHz type receiver used in the SPS COPOS system [6,7] shown in Figure 2.

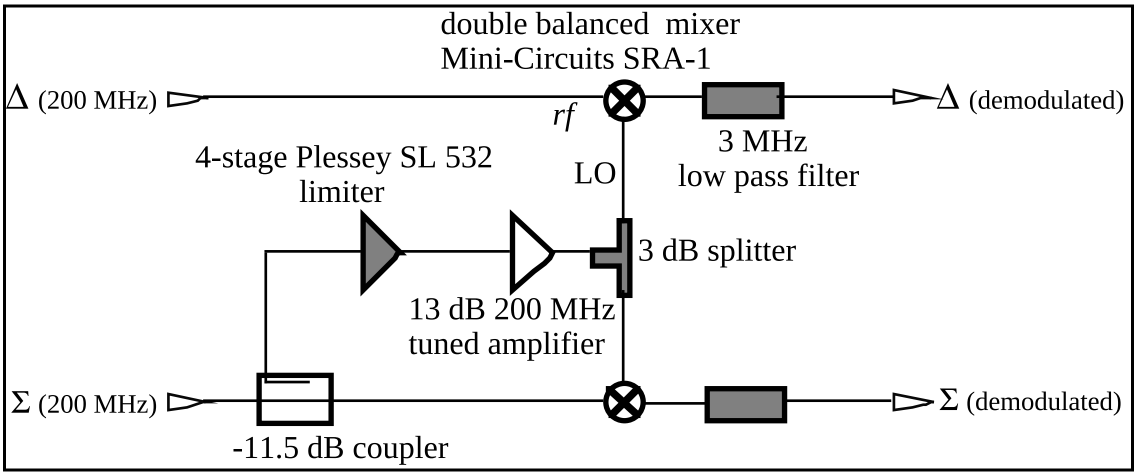

Figure 2 : Diagram of homodyne detector.

The frequency shift is performed by signal multiplication in a non-linear element called mixer. The carrier frequency or Local Oscillator (LO) is derived from the (strong) intensity or sum signal. It demodulates both itself (homo) and the difference signal. A low pass filter at the output blocks the high frequency multiplication products of the mixing process. The circuit performance is complicated by the string of non-linear elements: the limiter and the mixer. In the following the different parts are analyzed in some detail to understand better the limitations of this kind of demodulator.

4.1 Limiter [8,11]

The limiter has a noise factor of 11 dB when operating with an input impedance of 50Ω. Figure 3 shows the output power as a function of the input for a 6-stage Plessey SL 532 limiter. The limiting action is clearly visible when the input is larger than -60 dBm. It can not be lower than -90 dBm since that corresponds with the noise level of of the device. The gain of the limiter as function of the input power is shown in Figure 4. The small signal gain is 14.5 dB per stage.

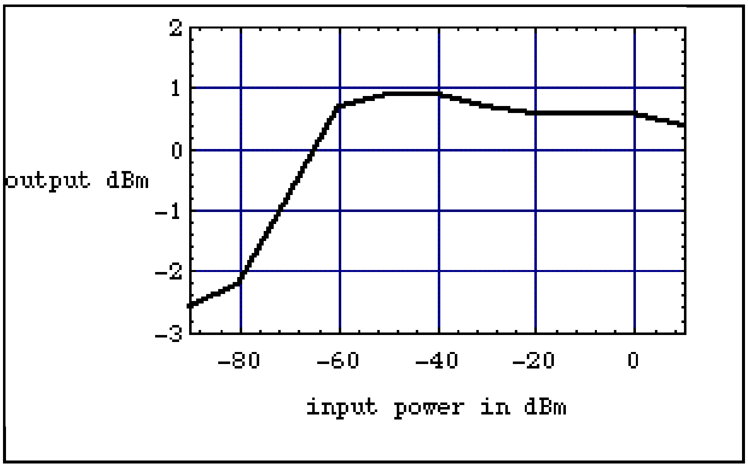

Figure 3 : Output versus input power for 6-stage limiter

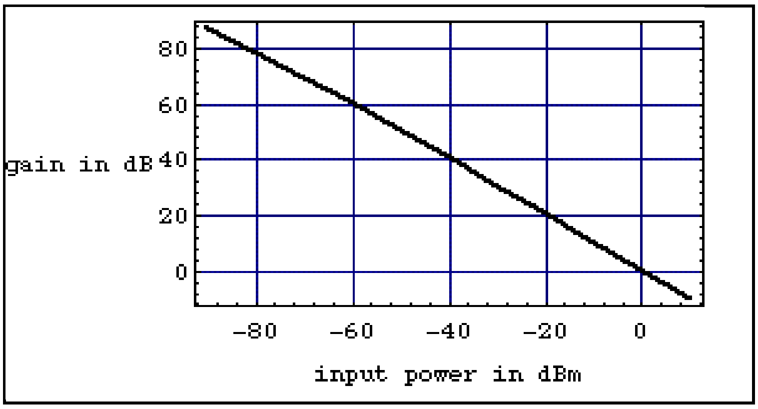

Figure 4: Gain versus input power of 6-stage limiter

The limiter used in Figure 2 consists of 4 stages. The gain characteristic can be derived from Figure 3 and 4 and is shown in Figure 5.

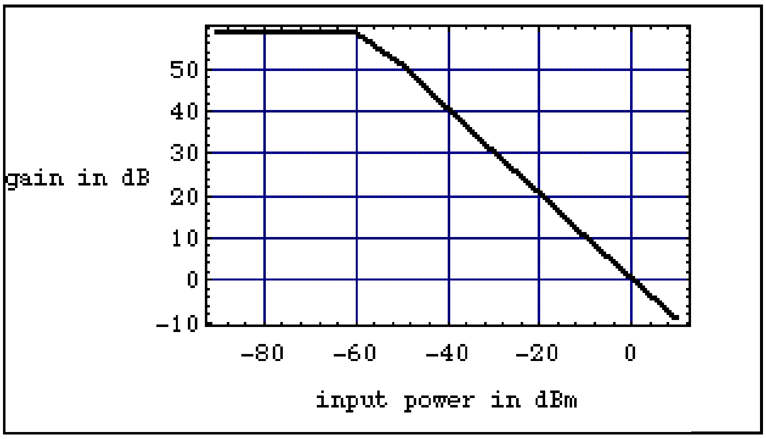

Figure 5 : Gain characteristic of 4-stage limiter

The quoted gain and noise factor were confirmed by direct measurements. An example is shown in Figure 6. It is to be noted that the output of the limiter was passed through a 200 MHz 13 dB amplifier before being fed into a spectrum analyzer.

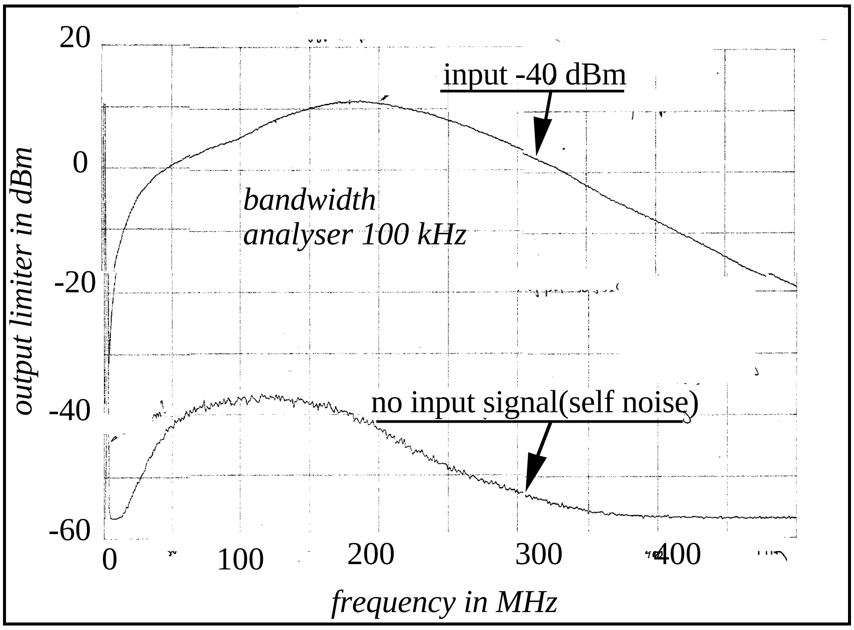

Figure 6: Frequency characteristic of limiter plus tuned amplifier

4.2 Mixer [9]

The insertion loss of the mixer is 3 dB when the LO drive power is between 4 and 10 dBm. The insertion loss increases linearly with decreasing LO power. The signal input should be kept 10 dB below the LO level to conserve linearity in the amplitude response. The isolation between the LO input and the mixer output in the same frequency band varies from 60 dB at low frequencies to 40 dB at high frequencies according to the data sheets.

The noise of the mixer alone was measured with and without generator at its *rf* input as shown in Figure 7.

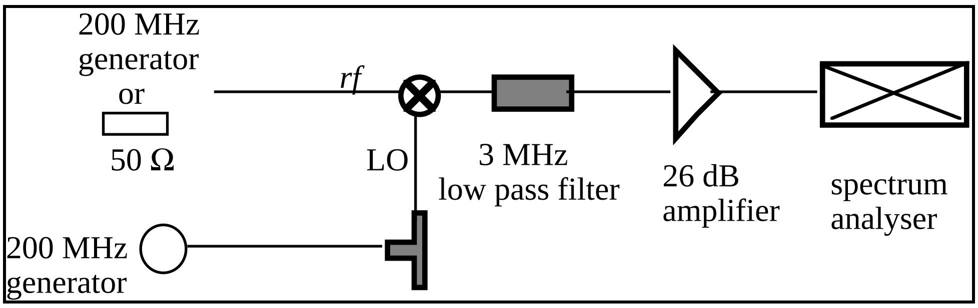

Figure 7 : Measurement set up of mixer noise.

The result of the measurement is shown in Figure 8.

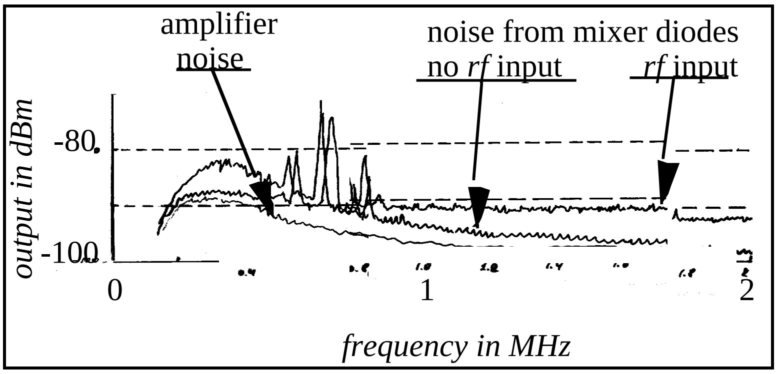

Figure 8 : Noise of mixer.

The mixer contains, apart from transformers, 4 diodes connected in a bridge circuit. The noise generated by a solid state diode can be calculated as follows [10] :

- diode noise current is : $i_n^2 = 2eI\Delta f$, where I is the current through the diode and Δf is the bandwidth.
- the noise power generated by one diode in a load R is : $P_n = 2eRI\Delta f = 2eV\Delta f$, where V is the voltage in the circuit.
- without signal only 2 diodes are conducting, generating a power of $P_n = 4eV\Delta f$.
- P_n = -89 dBm assuming a 10 dBm LO input and a bandwidth of 3 MHz.

The theory for noise in diodes predicts a flat spectrum up to some (very) high frequency. What is observed here is a concentration of noise below 1 MHz. When the *rf* entry of the mixer is connected to a powerful 0 dBm signal, a flat noise spectrum is indeed obtained as shown in Figure 7, approaching very well the theoretical model. In that case the diodes do not carry the same conduction current. It is not clear why this should redistribute the noise power more evenly across the spectrum. While keeping in mind this special feature, it is possible to attribute a noise factor to the mixer. It is simply determined by following ratio :

$$NF = 10 \log\left(\frac{V}{4kT/e}\right). \tag{7}$$

It turns out that the noise factor is 8.5 dB for a LO level of 10 dBm and reduces to 5.5 dB for a LO level of 4 dBm.

4.3 Performance of limiter and mixer combined

The lowest noise level that can be hoped for is determined by the noise factor of the mixer. That situation can be achieved when the input signal of the limiter is larger than a certain limit, in other words, when the gain of the limiter is low enough that its noise is not dominating the mixer noise. For smaller signals a much higher noise level is observed. That is shown in Figure 9.

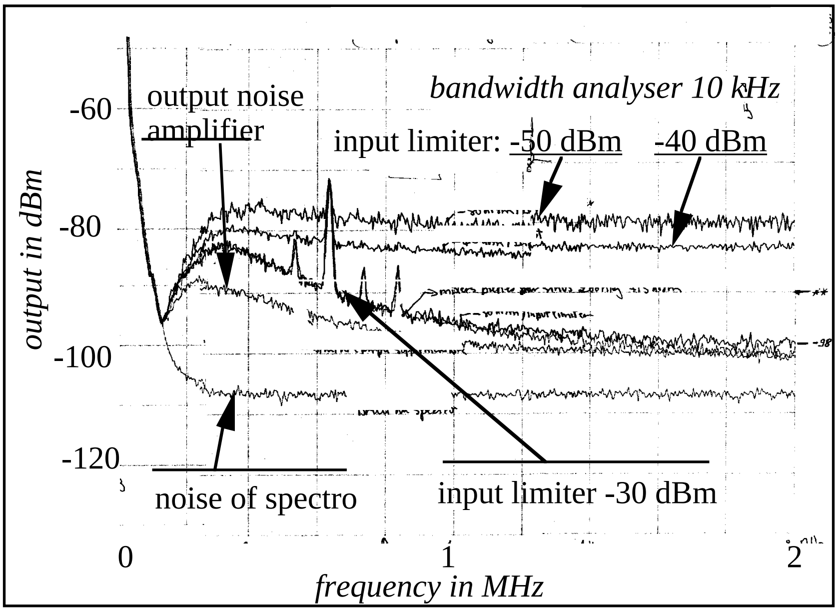

Figure 9 : Output noise of homodyne detector.

This behavior may be understood as follows. In a perfect mixer noise does not leak from the LO source to the output. The isolation is perfect. In a real mixer some finite leakage occurs. The degree of isolation is quoted in *dB* as a function of frequency in the data sheets. For example the isolation of the mixer SRA-1 that was tested drops from 60 dB at frequencies below 50 MHz to 40 dB at ~200 MHz. However the isolation that is needed in the application of the homodyne receiver is between *high* frequencies at the input (200 MHz) and *low* frequencies at the output (less than 3 MHz). No data are available from the manufacturer for this quantity. From the measurement in Figure 9 a *cross isolation* of 48 dB can be deduced which *seems* to be some average between the quoted low and high frequency isolation of the mixer.

4.4 Dynamic range and resolution

Consider the receiver of Figure 2. The noise factor associated with the demodulated Δ signal is constant and equal to the noise factor of the mixer (8.5 dB) when the gain of the limiter stays below 35.5 dB, assuming an average *cross isolation* of 48 dB. The noise increases quickly at a rate of 1 dB per dB additional gain. For the traditional 3 dB margin a gain limit of 38.5 dB is obtained. That gain limit defines a lower limit for the Σ signal at -27 dBm. On the other hand the maximum level that the mixer can digest is 0 dBm and a dynamic range of 27 dB is obtained for a noise factor of at most 11.5 dB. This dynamic range is modest to say the least and can be improved by taking following measures:

- *reduce* the fixed gain in the limiter branch. For example replace the 13 dB tuned amplifier by a 6 dB one. That reduces the LO demodulator level from 10 to 4 dBm which is still acceptable and improves the mixer noise factor from 8.5 dB to 5.5 dB. The dynamic range increases with 9 dB.
- *replace* the -11.5 dB coupler by a 3 dB splitter. The lower limit of the signal decreases with 8.5 dB. The higher limit drops by 3 dB to satisfy the condition that the input signal should not be larger than 10 dB below the LO level. The dynamic range in which the performance is excellent now becomes 41.5 dB for the same noise factor as before of less than 11.5 dB. The processor performance degrades for smaller signals and the absolute lower intensity limit where it stops to work is -57 dBm.

Note that the operating frequency is 200 MHz. It can not be chosen much higher for this particular mixer since the isolation degrades very quickly. However, at a lower operating frequency the isolation gets better, hence also the dynamic range. For example at 80 MHz a dynamic range of ~48 dB may be expected.

4.5 Discussion

In order to obtain more practical figures on resolution a filter bandwidth of 5 MHz is assumed. A noise factor of 11.5 dB will produce a noise power of -89.5 dBm. Intensity signals above -44.5 dBm will be faced with this noise factor and even somewhat less. This yields a resolution of 45 dB (0.56% of the aperture). The resolution increases to 86.5 dB(0.005%) at the maximum intensity signal level (-3 dBm). For the 80 MHz receiver these figures increase by ~6 dB (algebraic factor 2). The absolute lower signal level (100% resolution) is reached with -57 dBm at the input.

5. Phase processor [4,13-18]

The electrode signal of a beam position monitor can be considered as an *amplitude* modulated signal whereby the intensity is the carrier and the relative beam position is the modulation. Straightforward amplitud*e* demodulation of two electrode signals will separate them. The absolute beam position can be computed after detection by dividing (normalizing) the relative position by the intensity (previous chapter). The normalizing operation can be performed prior to detection by transforming the *amplitude* modulation (AM) of the intensity by the position into *phase* modulation (PM). The basic difference between AM and PM is the fact that the modulation vector is *in phase* with the carrier in AM and *π/2 out of phase* in PM. The π/2 phase shift is an essential feature when combining two filtered high frequency electrode signals to perform an AM/PM conversion or normalization. The demodulation is executed with a standard non-linear device that shifts one of the high frequency filtered side-bands to DC so that it can be acquired by suitable sampling techniques. The phase modulated signals are confined to a fixed amplitude by limiters or comparators so that a constant scaling factor is obtained that relates beam position and detected signal level.

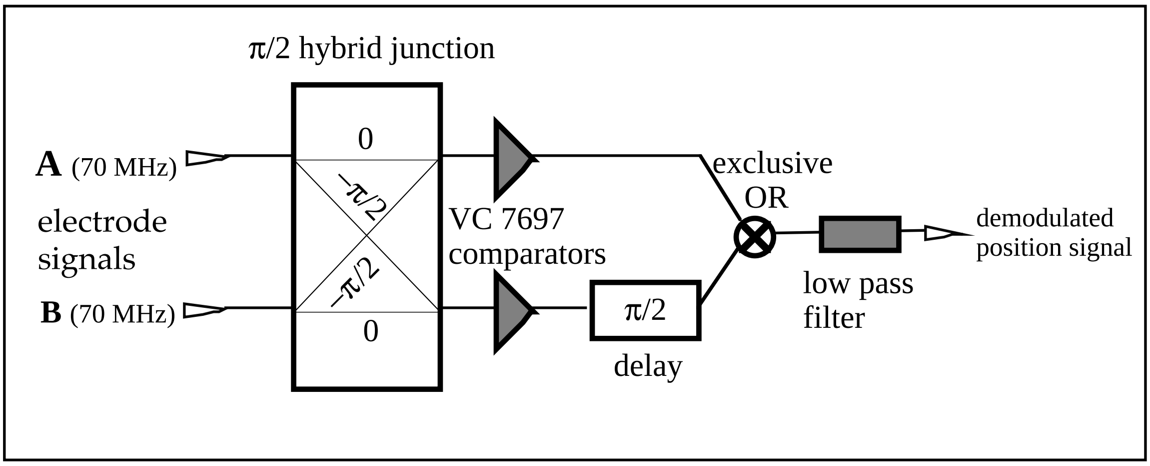

Figure 10 : Diagram of phase processor

Typical features of the phase processor can be recognized in the LEP 70 MHz receiver [4] shown in Figure 10.

5.1 Resolution and dynamic range of the phase processor

The resolution of this system is determined by the noise properties of the first amplifier that the electrode signal encounters. That device is the limiter or comparator. Any noise at their input will cause phase noise on the phase modulated signal which translates directly into position noise. The transformation of the input noise into position noise can be computed as follows.

A limiter or comparator produces a constant output level (apart from the sign) whenever the input level exceeds a certain value. This level is called the hysteresis (*h*) in the case of comparators. The action of the limiter and the effect of the noise is indicated in Figure 11.

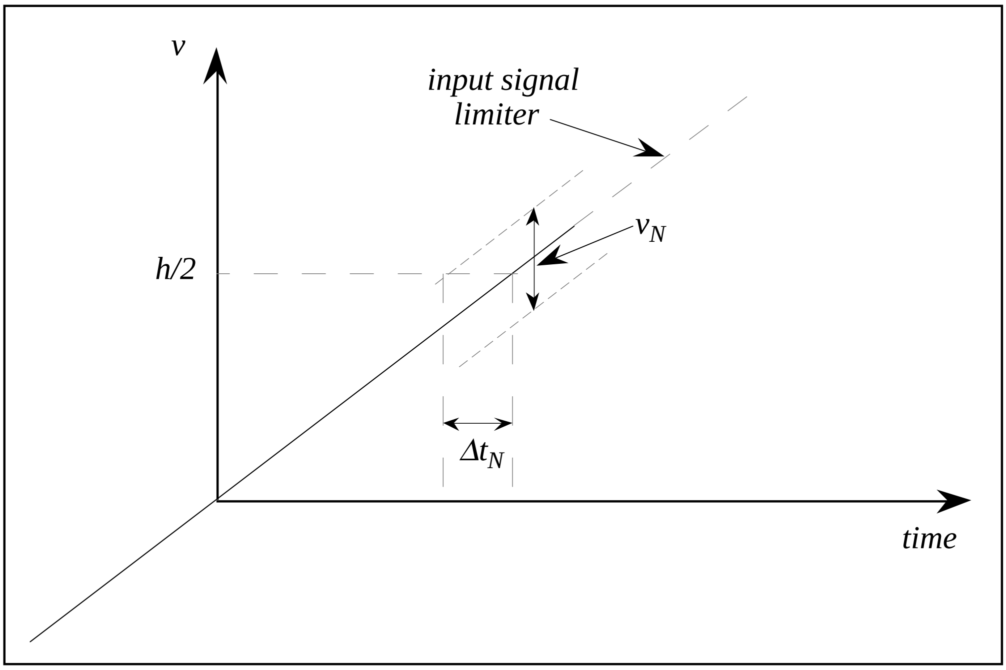

Figure 11 : Signal and noise at input limiter.

The signal at the input of the limiter is: $v_i = V_i \sin(\omega t)$, where $\omega = 2\pi f$ is the central frequency of the detector filters. A noise excursion v_N on top of the signal causes an error in time and phase :

$$\Delta t_N = \frac{v_N}{V_i \omega}; \quad \theta_N = \frac{v_N}{V_i}. \tag{8}$$

The noise voltage of the limiter is defined by its bandwidth. It is larger than the operating frequency *f* by a factor *W*. The output of the detector accumulates the noise of two limiters and of two zero crossings per period so that the phase error (noise) doubles to

$$\theta_N = 2\,\frac{v_N(Wf)}{V_i}. \tag{9}$$

The non-linear detection shifts the frequency of the signals by an amount f . That will concentrate the noise that was originally spread from $-Wf$ to Wf into two bands $-f$ to f and f to $3f$. The low pass filter will only retain the first band around DC reducing the noise power by a factor 2. For a low pass filter bandwidth of $\Delta f < f$, the phase noise at the output of the detector becomes :

$$\theta_N = \sqrt{2}\,\frac{v_N(W\Delta f)}{V_i} \approx \frac{\Delta}{\Sigma} = \frac{x_N}{A}. \tag{10}$$

In fact the output noise signal V_{oN} will be proportional to $tg(\theta_N)$. A calibration of the system will yield a signal level V_{ref} for $\Delta = \Sigma$ so that finally one finds :

$$V_{oN} = V_{ref}\,tg(\theta_N) \approx V_{ref}\,\theta_N = V_{ref}\,\frac{x_N}{A} = \sqrt{2}\,\frac{V_{ref}}{V_i}\,v_N(W\Delta f). \tag{11}$$

The noise voltage v_N is constant and is determined by the noise factor *NF* and the bandwidth of the limiter (*Wf*). As a consequence the resolution of the system is inversely proportional to the signal level, hence with the beam intensity.

The discrimination level or lower limit of the input signal V_i is *h/2*. When V_i approaches h/2 it is more appropriate to write :

$$V_{oN} = \sqrt{2}\,\frac{V_{ref}}{\sqrt{V_i^2 - (h/2)^2}}\,v_N(W\Delta f). \tag{12}$$

The resolution of the phase processor is determined by the noise factor *NF* of the limiter increased by $3\,dB$ (factor $\sqrt{2}$) and $(10\log(W))\,dB$.

In Figure 12 measurements on the LEP phase processor are compared with the proposed model.

The saturation effect at higher signal levels is due to the noise level of the output amplifier used in the measurement. The computed data fit the measurements for a *global* noise factor of 20 *dB* implying *NF=11 dB*. The model shows that the input signal should be at least 6 dB higher than the discrimination level to avoid problems with linearity.

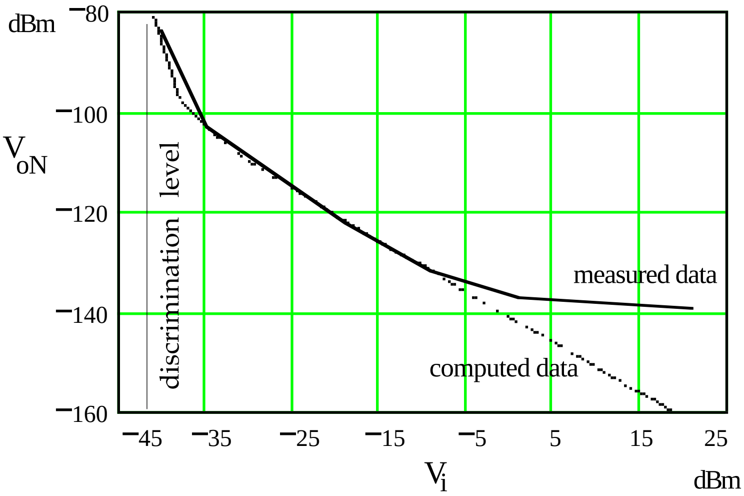

Figure 12 : Output noise of phase processor.

In Table 1 some of the data collected for various limiters [12] used in phase processors [4,13,14] are summarized.

limiter/comparator		AM685	SL532	VC7697
		Tevatron	test SPS	LEP
NF	*dB*	16	11	11
Wf	*MHz*	100	125	250
h/2 (discrimination level)	*mV*	1.5	1.2	2.5
	dBm	-46	-48	-42
f (operating frequency)	*MHz*	50	25	70
W		2	5	3.6
Δf	*MHz*	5	2	20
NF+3+10log(W)	*dB*	22	21	20
max. input	*dBm*	20	10	18
dynamic range	*dB*	60	52	54
output noise level normalized for $V = V_{ref}$	*dBm*	-79	-84	-73
observed level normalized for $V = V_{ref}$	*dBm*	no data	-86/-88	-71±1

Table 1 : Performance data of several phase processors

One of the characteristics of a non-linear detection system is the fact that the discrimination level is much higher (>30 dB) than the basic noise level.

5.2 Systematic errors

5.2.1 Small argument error

The output of the processor is proportional to $tg\alpha \sim \alpha = \Delta/\Sigma$. The error is :

$$\frac{\delta x}{A} = tg\alpha - \alpha . \qquad (13)$$

It is shown in Figure 13.

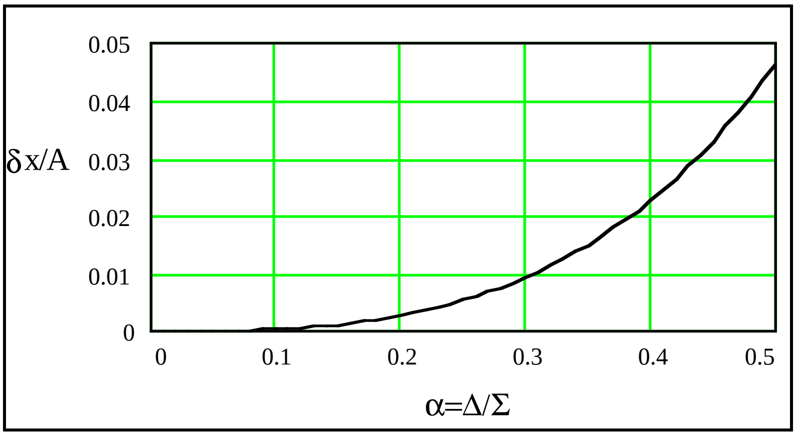

Figure 13 : Small argument error in phase processor.

This error can be corrected by software.

5.2.2 Systematic errors induced by limiter/comparator

The output of the limiter/comparator saturates when its input is larger than the limiting threshold. The output is ideally independent of the input as long as the limiting condition is fulfilled. It may not a surprise that in practice the ideal situation is not achieved. Hence, a systematic perturbation signal will appear at the output of the multiplier (mixer, exclusive OR). It is systematic, very wide banded and it depends on the level of the input signal (intensity). The error tends to be larger for smaller input levels [4].

The consequences may be very important as was the case at Fermilab. Their phase processors were designed to deal with a continuous bunched beam. The signal is strong (integration over several bunches in the bunch train) and the systematic error is therefore relatively small and constant and can go unnoticed or is eliminated with proper calibration. The situation changes drastically with single bunches. The error, as before, depends on intensity but now changes continuously during the signal pulse duration as well. Simple calibration is no cure in this case. Calibration pulses which simulate the beam signal in amplitude and duration can help (LEP). Narrowband filters (~5 MHz) after detection will reduce these error signals correspondingly. This unfortunately was <u>not</u> done in Fermilab or LEP where filters of 15 and 20 MHz were used.

5.3 Discussion

The resolution of the phase processor within its dynamic range varies from about 30 dB (3% of aperture) at minimum intensity (-36 dBm) to 90 dB (0.003%) at maximum intensity (18 dBm, electronics limit). That is clearly adequate for orbit trajectory work on the condition that the minimum signal level in practice is not too close to the discrimination level (-42 to -48 dBm). The operating frequency is less than ~100 MHz.

The detector is limited by its noise factor of 20 dB in its possible use as a very low noise detector. Low noise pre-amplifiers may reduce the noise factor to less than 3 dB at the expense of the dynamic range. Indeed the reduction of the dynamic range will be equal to the reduction of the noise factor. It is important to use an adequate low pass filter after demodulation to safeguard the performance of the detector.

6 The log-ratio detector [16,19,20]

The beam position is found by making the ratio of difference and sum signals of the two electrodes of a beam position monitor. The phase processor performs this operation by changing the modulation from AM to PM. In the logarithmic amplifier (logamp) this normalization is done by taking the difference of the logarithm of the signals. The principle of such a detector that was proposed for the SSC is shown in following Figure 14.

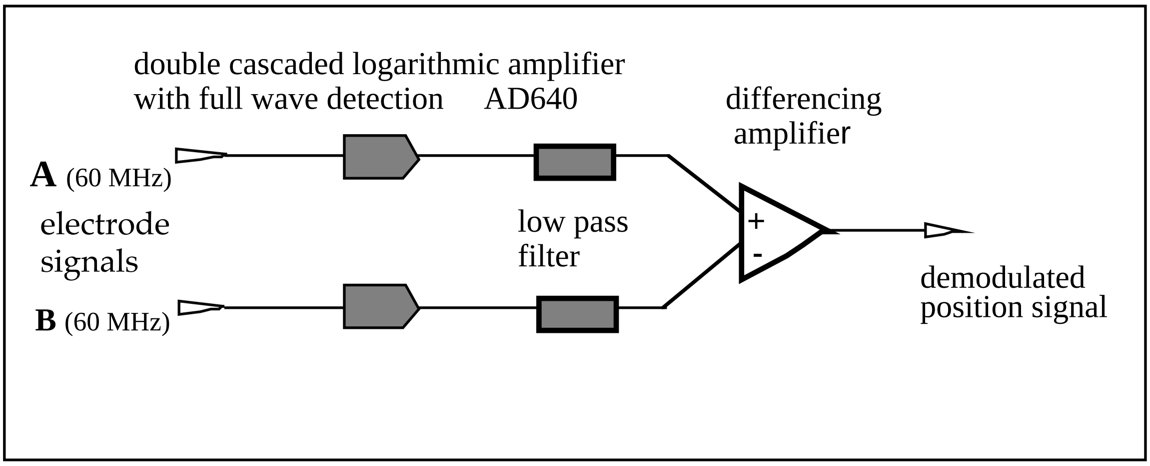

Figure 14 : Diagram of *logamp.*

The electrode signals can be written as (see (1)):

$$E_{1,2} = \frac{1}{2}(\Sigma \pm \Delta). \tag{14}$$

The output of the difference amplifier after passing through the logarithmic detector is :

$$\begin{gathered} \ln\left(\frac{\Sigma+\Delta}{\Sigma-\Delta}\right) \approx 2\frac{\Delta}{\Sigma} = 2\frac{x}{A} \\ \frac{\delta x}{A} = \frac{1}{2}\ln\left(\frac{1+\alpha}{1-\alpha}\right) - \alpha \qquad 0 < |\alpha| < 0.5 \end{gathered} \tag{15}$$

6.1 Resolution

The equivalent signal at the input of the logamp is $\Sigma/2 \pm V_N$ where V_N is the noise voltage at the input. The noise voltage is transformed into position noise :

$$\frac{x_N}{A} = \frac{2V_N}{\Sigma}. \tag{16}$$

Applying this relation to published data [21] the following can be found:

- input signal $\Sigma/2$ = -40 *dBm*.
- noise signal $x_N/A = 0.0017$ which yields a noise power of -95 *dBm*.
- band width 2 MHz so that $\underline{NF = 10\ dB}$, compatible with published limiter noise factors [8].

6.2 Dynamic range of log-ratio detector.

The logamp detector consists of a sequence of identical limiting amplifiers. The output of each amplifier is rectified and then summed. The AD640 contains 5 limiters. Two chips can be cascaded and their output can be summed as well. It can be seen that the limiters define discrete output levels that increase linearly for an exponential rise (the gain of one amplifier) of the input signal. The connection of these points is strictly linear. The transfer characteristic is shown in Figure 15.

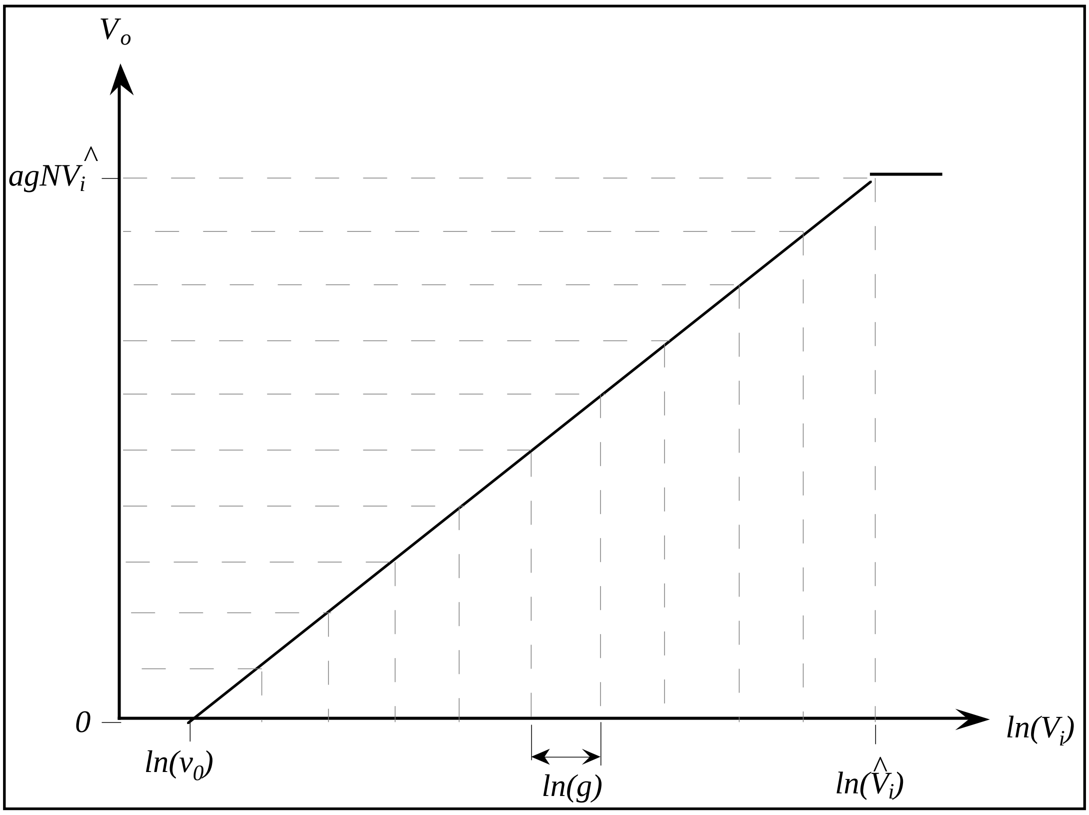

Figure 15 : Transfer function of logamp

The meaning of the symbols is :

- $\hat{V}_i$ is the voltage at the input of a limiter that drives it into saturation and it corresponds typically to ~0 dBm.
- *g* is the small signal algebraic gain of the limiter.
- *N* is the total number of limiters in the logamp (*N*=10 for the SSC detector).
- v_0 is the *zero intercept* voltage of the detector.
- *a* is the gain or attenuation of the logamp output signal.

The theoretical maximum dynamic range (in *dB*) in the *absence of noise* is then:

$$20\log\left(\frac{\hat{V}_i}{v_0}\right) = 20N\log(g) = GN, \tag{17}$$

where G is the gain in *dB* of a single limiter. The logarithmic approximation error at the zero intercept voltage v_0 is very high. Therefore the input signal must always be larger than this value even in the absence of noise to maintain the logarithmic characteristic.

The lower signal limit has to be compared with the <u>*full bandwidth*</u> noise voltage of the logarithmic amplifier. This is computed with:

$$(v_N)_{dBm} = 10\log(4kT) + 30 + NF + 10\log(f_L) = -158 + 10\log(f_L), \tag{18}$$

where f_L is the bandwidth of a single limiter amplifier and V_N the corresponding wideband noise voltage expressed in units of *dBm*.

The value of the dynamic range is further reduced for off-centered beams. The minimum intensity signal for a centered beam is :

$$\Sigma_{\min} = 2gv_L. \tag{19}$$

For an off-centered beam this becomes :

$$\Sigma_{\min} = 2gv_L + \Delta. \tag{20}$$

If the position measurement must be able to explore the central pick up region up to 1/4 of the monitor radius then $\Delta \le \Sigma/2$ and the *low* edge of the dynamic range is reduced by 6 *dB*. The same reasoning leads to a similar reduction at the *upper* edge of the range of 4 *dB*. The total loss of dynamic range for off-centered beam is <u>10 *dB*</u>. Saturation roll-off will cause a further reduction of dynamic range at the upper edge independent of beam position.

The analysis can be checked with the published data on the SSC logamp[20]:

- the total gain can be found to be 82.5 *dB* yielding G=8.25 *dB*.
- $\hat{V}_i = -2dBm$ yielding $v_0 = -84.5dBm$.
- $f_L = \sqrt{5} * 145MHz$ yielding $v_N = -73dBm > v_0$.
- the maximum value of the dynamic range, ignoring saturation roll-off of the limiter, is $DR = \hat{V}_i / v_N = 71dB$ for a centered beam and $61dB$ for an off-centered beam. The measured dynamic range ranges from 66dB to 57dB indicating that the roll-off is responsible for 4*dB*.

6.3 Systematic errors

6.3.1 Small argument approximation

The relative position error $\delta x/A$ is given by (15) and is shown in Figure 16.

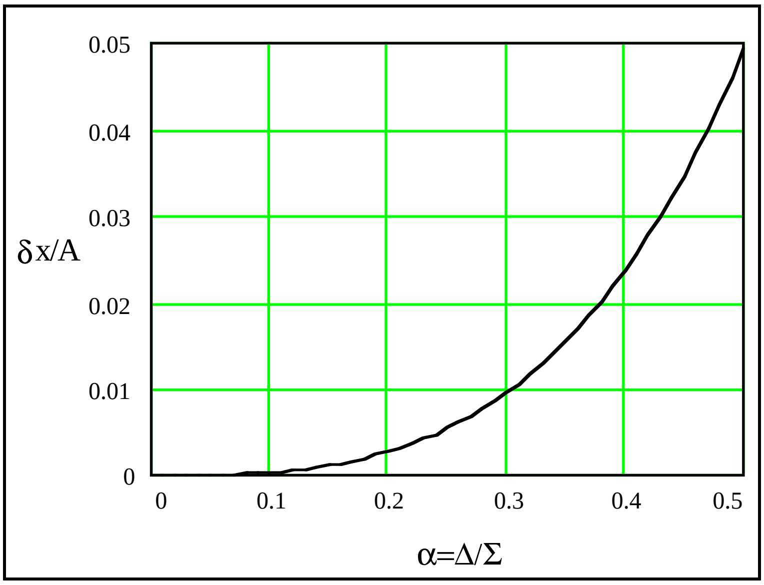

Figure 16 : Small argument error in logamp.

This error is large enough to call for a software correction *a posteriori*.

6.3.2 Linear approximation error

One of the 10 interpolation intervals of Figure 15 is shown in Figure 17 where *z* stands for $ln(V_i)$ and *y* stands for V_o. A linear interpolation between two cross-over points is assumed.

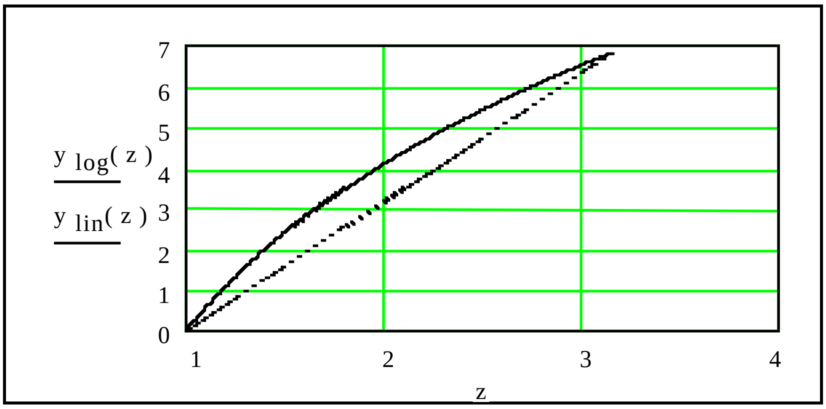

Figure 17 : Difference between linear and logarithmic interpolation.

The interpolation error is obviously zero for centered beams. For off-centered beams it can be seen that the error is a function of the intensity and beam position. It reaches a maximum when one of them is on the edge of the interval while the other is near the middle. Figure 18 illustrates the effect It is assumed that one of the electrode signals corresponds with $z=1$ while the other one increases from $z=1$ to $z=\sqrt{10}$. It is an unlikely situation where beam intensity and beam position both change in a particular way but it shows the nature of the error and its magnitude.

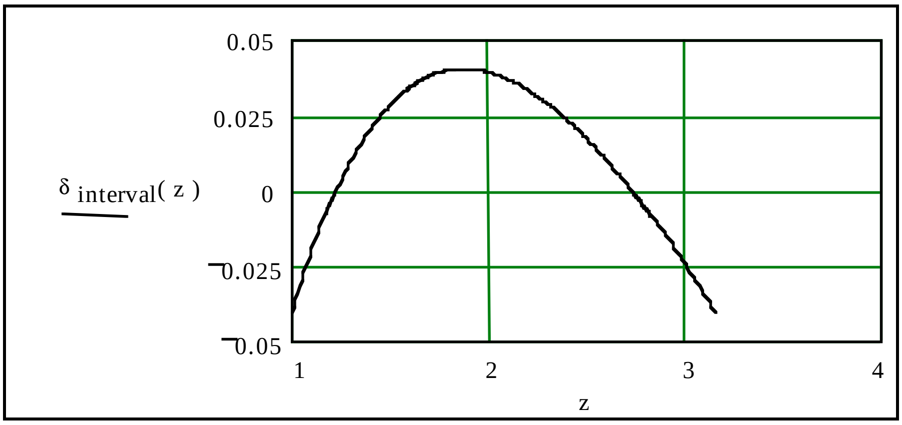

Figure 18 : Evolution of interpolation error across a logamp gain interval.

The peak to peak error is ~8% (0.7 dB) of the normalizing aperture. The peaks occur close to the cross-over between the successive limiting stages. In the real logamp an effort has been made to depart from the linear transfer function in the interval and approach the logarithmic function. However, a variation can clearly be seen in the measurements. The observed effect is ~*2 times less* than the effect calculated here. Apart from the logarithmic approximation it is to be expected that the saturation roll-off of the limiting amplifiers has a beneficial effect this time and will attenuate the sharp rise of the error at the edge of the interval. No simple correction is possible as for the previous type of error.

6.3.3 Bandwidth limitation

The bandwidth of the logamp decreases with the number of non-saturating limiter stages: it is large for strong signals and small for the signals near the noise threshold. The bandwidth reduction at small signals causes a reduction in gain which in turn produces a proportionality error in the measured beam position and that is clearly seen in published data. The effect has been computed for the SSC case and is shown in Figure 19. The working frequency is 60 MHz and the individual limiter bandwidth is 350 MHz. Due to the dependence on a single parameter (power of input signal) it may be possible to correct for this.

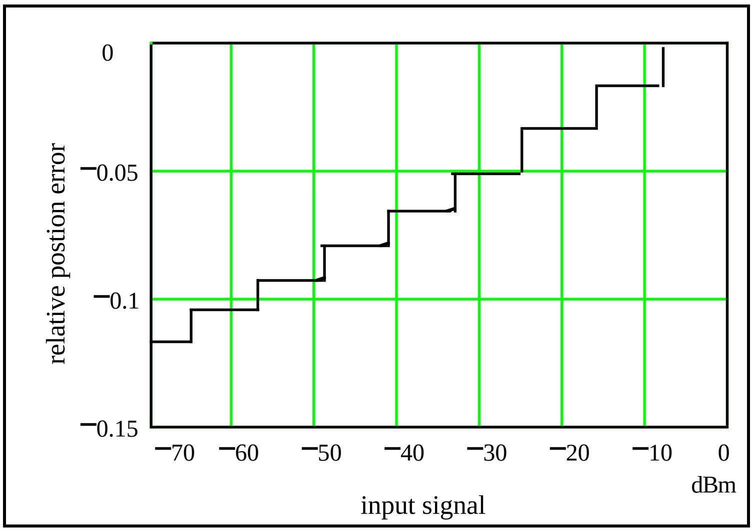

Figure 19 : Relative position error due to progressive bandwidth limitation.

6.4 Discussion

The minimum and maximum signal levels ($\Sigma/2$) defining good operational resolution are -67 and -10 dBm. Assuming a 5 MHz bandwidth the resolution ranges from 24 dB (6.3% of the aperture) to 81 dB (0.009% of the aperture). The hard lower signal level limit is -73 dBm. The operating frequency is limited by present hardware to ~100 MHz. Systematic errors which are difficult to compensate may introduce relative position errors in the order of 10%.

7. Summary

In Table 2 the expected performance of the four different types of processors is summarized. A bandwidth of 5 MHz is assumed for ease of comparison except for the integrator where 50 MHz is taken. The resolution is expressed relative to the aperture of the beam position monitor.

		Integrator	Homodyne	Phase proc.	Log-ratio
max. frequency	MHz		~200	~100	~100
hard low limit($\Sigma/2$)	dBm	-64	-60	-42/-48	-73
low limit($\Sigma/2$)	dBm	-44	-47	-36	-67
high limit($\Sigma/2$)	dBm	~10	-6	18	-10
dynamic range (operational)	dB	~54	41	54	57
dynamic range (resolution limit)	dB	~74	54	60/66	63
resolution low	%	10	0.6	5	6.3
resolution high	%	0.03	0.005	0.005	0.009

Table 2 : Performance summary of processors.

It may be worthwhile to note that the lower limit of the phase processor and the log-ratio detector are rather sharp. That is not the case for the integrator nor for the homodyne detector where the performance degrades more gradually for decreasing beam intensity. The full range of possible signals emanating from a beam position monitor, modified by the signal cable and band filter, will fit nicely in the signal level acceptance of the processor in a well adapted design. Reasonable safety margins should be planned early in the design stage both at the higher and lower limits of the processor. Insufficient acceptance can be enhanced with low noise pre-amplifiers or attenuators.

8. Acknowledgments

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