Pre-A/D Filter and AGC Requirements for Multimegabit Telemetry Data Detection

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This article presents a candidate pre-A/D filter bandwidth versus data rate design for the Multimegabit Telemetry Demodulator/Detector, which is based on considerations of A/D bias, quantization errors, and automatic gain control (AGC) effects, as well as the deleterious effects of filtering.

Two methods of gain control of the input level to the A/D converter are considered. The first method uses a particular value of gain, according to which bandwidth is selected. The second method uses a second narrowband noncoherent AGC (in the LPF bandwidth) to attempt to keep the A/D input level constant. This second method reduces the BER degradation slightly but appears to be more difficult to implement.

I. Introduction

The purpose of this article is to develop a table of data rates vs the pre-A/D low-pass filter requirements for data detection (as an initial design) for the Multimegabit Telemetry Demodulator/Detector assembly, and to specify the types and the number of automatic gain controls (AGCs) needed. To do this task, it was required to consider the bit error rate (BER) degradation due to the effects of AGCs, A/D quantization, A/D bias, phase detector saturation, and the number of AGCs required. The system under consideration is shown, in a simplified form, in Fig. 1. Since data demodulation was the primary concern, the upper arm of Fig. 1 is the system under primary investigation.

The system philosophy is to arrange the sample rate, via the rate buffers, so that the internal sample rate out of the rate buffers is at four times the data symbol rate. Further, internal sample rates (data handling rates) are reduced to one fourth the symbol rate into the loop filter.

The noncoherent AGC is important also since the signal component variations out of the phase detectors (multipliers) can tolerate only so much dynamic variation before they are affected by dc offsets at weak signal conditions or, at the other end, limiting occurs at strong signal plus noise power conditions.

The narrowband noncoherent AGC, if used, can better control the signal and noise over the data rate extremes for a given filter bandwidth than can the gain set approach that fixes the gain for a particular bandwidth (see Fig. 1). However, the advantage in reduction of BER degradation is rather small and the implementation advantage of the filter-determined gain select might well be the overriding factor favoring the latter approach.

Based on the functional requirement description, the symbol error rate degradation is 0.5 dB from theoretical below 10 Msps and 1 dB from theoretical above 10 Msps, for symbol $SNR\ E_S/N_o \geq -4$ dB.

II. Dynamic Range Considerations

First we consider the dynamic range requirements of the wideband noncoherent AGC [see Fig. 1] having pre-AGC noise bandwidth B. From Ref. 1, the ratio of the signal power component out of a noncoherent AGC at an input SNR given by $SNR_0 = \alpha_0 S_0/N B$, to the signal power component out of the AGC at an input SNR given by $SNR_1 = \alpha_1 S_1/N_o B$ was shown to be (see Fig. 2).

$$\frac{P_1}{P_0} = \left[\frac{1}{\dfrac{N_o B}{\alpha_1 S_1} + 1}\right]\left[1 + \frac{N_o B}{\alpha_0 S_0}\right] \tag{1}$$

where N_o is the one-sided noise spectral density, B is the noise bandwidth of the BFP preceding the noncoherent AGC, $\alpha_i (i = 0, 1)$ is the filter loss through the BPF, and S is the unfiltered input signal power. This result, Eq. (1), applies to both the wideband and narrowband noncoherent AGCs.

Now we shall determine the dynamic range of signal power out of the noncoherent AGC. First determine the minimum wideband noncoherent AGC input SNR. The minimum value of E_s/N_o is given by

$$\left(\frac{E_s}{N_o}\right)_{min} = -4 \text{ dB} \tag{2}$$

So that the minimum AGC input SNR is given by

$$\frac{\alpha_0 S_0}{N_o B} = \left(\frac{E_s}{N_o}\right)_{min} (\alpha_0)\left(\frac{R_s}{B}\right) = -31 \text{ dB} \tag{3}$$

where the bandpass filter has a bandwidth of $B = 60$ MHz, and the minimum symbol rate (R_s) of 125,000 sps through it, and $\alpha_0 \geq 1$.

The maximum value of E_s/N_o is given by:

$$\left(\frac{E_s}{N_o}\right)_{max} = +12 \text{ dB} \tag{4}$$

Therefore the maximum value of the AGC input SNR is given as

$$\frac{\alpha_1 S_1}{N_o B} = \left(\frac{E_s}{N_o}\right)_{max} (\alpha_1)\left(\frac{R_s}{B}\right) \geq 8.7 \text{ dB} \tag{5}$$

for the maximum E_s/N_o value and maximum symbol rate of 32 Msps and with $\alpha_1 = 0.87$. Therefore, the ratio of maximum-to-minimum wideband noncoherent signal power is, by Eq. 1,

$$\frac{P_1}{P_0} = 31.2 \text{ dB} \tag{6}$$

It follows that the dynamic range of the phase detectors (multipliers) must be 31.2 dB. If this dynamic range is too large, one option to reduce it would be to narrow the bandpass filter bandwidth at the lower data rates by, say, 10 dB so that the effective dynamic range would be only about 21 dB.

III. Approximate BER Degradation due to Quantization and dc Offsets

The Computer Labs MOD-4100 A/D converter was chosen for the multimegabit system. It has a dc accuracy of 10 MV ± 1/2 of the least significant bit. It has four-bit resolution so that we have four bits per sample. We consider the worst case data rate range[1] when the rate buffer sets $k = 1$, and therefore the statistic (a sample and dump filter), based on sampling four samples per bit on which bit decisions are made, is

$$e_o = \sum_{i=1}^{4} \left\{ g(\sqrt{\alpha} A_i + N_i) + b_i + \epsilon_i \right\} \tag{7}$$

where A_i is the sampled signal voltage, N_i is the sampled noise voltage, b_i is the sampled dc bias, α is now the power filtering loss through the low-pass filters (having bandwidth B_{LP}), g is the product of AGC voltage gains, and ϵ_i is the sampled quantization error. Based on this statistic, we will derive an expression for the matched filter output SNR in terms of the quantization spacing, ΔL, and the bias b. Since our analysis is

[1] See Table 1.

only meant to be approximate, we neglect intersymbol interference, which will be (approximately) accounted for by a separate calculation.

First note that (letting $E(\cdot)$ denote the ensemble average)

$$E\left[N_i^2\right] = N_o B_{LP} \tag{8}$$

with B_{LP} being the low-pass filter bandwidth preceding the A/D converter. Now the mean output signal of the sample and dump filter (just before dumping) is given by

$$E[c_o] = \sum_{i=1}^{4} (g\sqrt{\alpha}A_i + b_i) = 4g\sqrt{\alpha}A + 4b \tag{9}$$

assuming the voltages satisfy $A_i = A$ and $b_i = b$ with α being the power loss due to filtering of the LPF (see Fig. 4)[2]. The variance of the output statistic is given by

$$\mathrm{Var}(c_o) = E\sum_{i=1}^{4} g^2 N_i^2 + E\sum_{i=1}^{4} \epsilon_i^2 \tag{10}$$

If ΔL is the quantizer spacing, then (assuming a uniform distribution)

$$E\left[\epsilon_i^2\right] = \frac{(\Delta L)^2}{12} \tag{11}$$

Therefore, the output SNR is given by

$$SNR_o \cong \frac{(4gA\sqrt{\alpha} + 4b)^2}{4g^2 N_o B_{LP} + \dfrac{(\Delta L)^2}{3}} \tag{12}$$

Rearranging, and using the fact that the symbol duration $T = 4T_s$, with T_s being the time between samples, we obtain

$$SNR_o \cong \frac{2E_s}{N_o}\left[\frac{\alpha\left(1 + \dfrac{b}{g\sqrt{\alpha}A}\right)^2}{1 + \dfrac{(\Delta L)^2}{12 N_o B_{LP} g^2}}\right] \tag{13}$$

[2] This is a first-order correction for the effect due to filtering.

where we have used $2T_s \cong 1/B_{LP}$. The bracketed term is the degradation in terms of the quantizer spacing ΔL and the quantizer bias voltage b for a given bit and given bias.

The Computer Labs Mod-4100 A/D Converter has a 4-bit resolution and a maximum input range of ± 2.1 volts, so that

$$\Delta L = 0.2625 \text{ volts} \tag{14}$$

IV. Candidate Filter Bandwidth Versus Data Rate Design

As was mentioned in the Introduction, the basic philosophy in the digital demodulator/detector design is to minimize the analog filter count and to set the internal sample rate (processing rate) at four times the data symbol rate out of the rate buffers (see Fig. 1). This way everything after the rate buffers can be scaled according to data rate.

The two basic requirements of operation into the A/D converter are, one, that the signal plus n_o times the rms noise be less than the one-sided A/D input voltage range, or

$$g_{WB} g_{NB} \sqrt{\alpha}A + n_o\, g_{WB} g_{NB} \sqrt{N_o B_{LP}} < V \tag{15}$$

The second requirement is that the signal voltage into the A/D converter shall always be bigger than b, or:

$$g_{WB} g_{NB} \sqrt{\alpha}A \geq b\, n_b \tag{16}$$

where g_{WB} is the noncoherent wideband voltage AGC gain, g_{NB} is the noncoherent narrowband voltage AGC gain, V is the one-sided quantizer input range, and n_o and n_b are confidence safety factors (numbers greater than one). To keep the BER degradation down to around 0.5 dB or less, below 10 Msps, $B_{LP}T$ should be greater than about 1.7 (Ref. 2). Above 10 Msps, $B_{LP}T$ should be greater than 0.8 to hold the BER degradation to about 1 dB (Ref. 2).

Table 1 illustrates a candidate filter design that requires six analog low-pass filter pairs (B_{LP}'s) and should keep the symbol BER just about within specifications.

V. Comparison of the Filter-Controlled Gain Approach With the Coherent AGC Approach

In this section we will compare the relative performance for the case that the signal plus noise into the A/D is controlled by a fixed gain, set according to the low-pass filter bandwidth on one hand, and to the narrowband noncoherent AGC to control the signal plus noise into the A/D on the other hand. We will consider the 26.7-MHz filter data rate range since it operates over the largest data rate range.

Case I: filter bandwidth controlled gain

First we consider the case when the selection of a particular low-pass filter pair specifies a fixed gain into the A/D converter based on achieving acceptable bit error rate degradation.

At the highest data rate (32 Msps) and at an E_s/N_o (3 dB), which will provide a nominal maximum bit error rate of 1×10^{-5}, we require that Eq. (15) be satisfied as follows:

$$g_{WB_1} g_o \left[\sqrt{\alpha_1} A_1 + n_o \sqrt{N_o B_{LP}}\right] < V = 2.1 \text{ volts} \tag{17}$$

where g_o is the filter-controlled gain value fixed to the 26.7-MHz filter. This gain g_o is in lieu of the narrowband gain g_{NB}.

At the lowest data rate (4 Msps), for the same filter, and the same gain setting, g_o, and with the minimum E_s/N_o (-4 dB), we require that Eq. (16) satisfy:

$$g_{WB_2} g_o \sqrt{\alpha_2} A_2 = n_b b \tag{18}$$

where $b = 0.01$ volts and n_b should be at least 3 and preferably larger, and the subscript 2 denotes a different level of that variable. Now using $E_s/N_o = 2.0$ (3 dB) we conclude that the SNR into the AGC (see Fig. 1) satisfies

$$SNR_{LP} = \frac{\alpha_1 A_1^2}{N_o B_{LP}} = \frac{2.0}{\left(\dfrac{B_{LP}T}{\alpha_1}\right)} \tag{19}$$

Letting $n_o = 3.0$, which is equivalent to considering a 3σ noise amplitude, we obtain from Eqs. (17) and (19)

$$g_{WB_1} g_o \sqrt{N_o B_{LP}} \left[3 + \sqrt{\frac{2.0}{\dfrac{B_{LP}T}{\alpha_1}}}\right] = 2.1 \text{ volts} \tag{20}$$

Now since $B_{LP}T = 0.83$ at this maximum symbol rate (see Table 1) we can determine the filtering loss to be $\alpha_1 = 0.87$, assuming an ideal low-pass filter preceding the A/D converter. It follows from Eq. (20), with the fact that $B_{LP}T = 0.83$ and $\alpha_1 = 0.87$, that the rms noise voltage into the A/D converter is given by

$$g_{WB_1} g_o \sqrt{N_o B_{LP}} = 0.472 \text{ volts} \tag{21}$$

From Eq. (17), it follows that the signal voltage into the A/D converter is given by

$$g_{WB_1} g_o \sqrt{\alpha_1} A_1 = 0.684 \text{ volts} \tag{22}$$

Now it is easy to check to see if the condition of Eq. (18) has been satisfied. We have

$$n_b = \frac{g_{WB_1} g_o \sqrt{\alpha_1} A_1}{b} = \frac{0.684}{0.01} = 68.4 \tag{23}$$

which, as we shall see, causes a negligible degradation.

From Eqs. (13), (14), (21), and (23), the resulting degradation[3] due to bias and quantization only (the filtering losses will be considered below) is

$$DEGR(A_1) = \frac{\left(1 + \dfrac{b}{\sqrt{\alpha_1} A_1 g_o g_{WB_1}}\right)}{1 + \dfrac{(\Delta L)^2}{12 N_o B_{LP} g_o^2 g_{WB_1}^2}}$$

$$= \frac{\left(1 - \dfrac{1}{68.4}\right)^2}{1 + \dfrac{1}{12}\left[\dfrac{0.2625}{0.472}\right]^2} \tag{24}$$

or

$$DEGR(A_1) = 0.24 \text{ dB} \tag{25}$$

Now consider the degradation due to filtering, which includes power loss, distortion, and intersymbol interference effects.

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[3]This equation is somewhat a worst case since it assumes the bias is always "bucking" the signal.

Since the baseline LPFs are 2-pole filters with $B_{LP}T = 0.83$, we have that the loss, from Ref. 2,[4] is about 0.6 dB. Therefore, the total degradation of about 0.84 dB is to be expected at the maximum data rate (32 Msps). Hence using $B_{LP} = 26.7$ MHz and a symbol rate of 4 Msps at $E_s/N_o = 3$ dB we have

$$DEGR_{\mathrm{TOT}} = 0.84 \text{ dB} \tag{26}$$

This degradation does not include carrier loop or bit synchronization loop tracking inaccuracies. When $E_s/N_o = -4$ dB again at the highest data rate it can be shown that the degradation increases less than 0.1 dB so that the total degradation is about 0.9 dB.

Equations (21) and (22) apply to the case the highest data rate is in effect along with the assumption that the symbol-to-noise spectral density ratio is 3 dB and the gains, g_o and g_{WB_1} are set to satisfy Eq. (17) at $E_s/N_o = 3$ dB.

Now consider the case that the lowest admissible data rate (4 Msps) of the 26.7 megahertz filter is used (see Table 1) and the E_s/N_o ratio is minimum (-4 dB). Since the same LPF bandwidth is used, the gain g_o will be unchanged. However, the wideband noncoherent AGC will change gain since the signal power has been reduced. To compute the new gain level of the noncoherent AGC, we equate the noncoherent output power under strong and weak signal conditions so that

$$g_o^2 g_{WB_1}^2 \left[\alpha_1 A_1^2 + N_o B\right] = g_o^2 g_{WB_2}^2 \left[\alpha_2 A_2^2 + N_o B\right] \tag{27}$$

From Eq. (27) we can obtain the ratio of gains:

$$\frac{g_{WB_2}}{g_{WB_1}} = \sqrt{\frac{SNR_1 + 1}{SNR_2 + 1}} \tag{28}$$

where the SNRs out of the filter preceding the wideband AGC is given by

$$SNR_1 = \frac{\alpha_1 A_1^2}{N_o B} \tag{29}$$

$$SNR_2 = \frac{\alpha_2 A_2^2}{N_o B} \tag{30}$$

[4] The degradations computed here were at a BER = 10^{-5}, and were the only ones available.

We can now compute the SNRs in the bandwidth B:

$$SNR_1 = \left(\frac{E_s}{N_o}\right)_1 \alpha_1 \left(\frac{R_{s_1}}{B}\right) = 2(0.87)\left(\frac{32}{60}\right) = 0.9289 \tag{31}$$

$$SNR_2 = \left(\frac{E_s}{N_o}\right)_2 \alpha_2 \left(\frac{R_{s_2}}{B}\right) = 0.398(0.99)\left(\frac{4}{60}\right) = 0.0263 \tag{32}$$

It follows from Eqs. (28), (31), and (32) that

$$\frac{g_{WB_2}}{g_{WB_1}} = \sqrt{\frac{1 + 0.9289}{1 + 0.0263}} = 1.372 \tag{33}$$

We can now compute the new noise power associated with the lower symbol rate and lower E_s/N_o. From Eqs. (21) and (33), we obtain

$$g_o g_{WB_2} \sqrt{N_o B_{LP}} = g_o g_{WB_1} \sqrt{N_o B}\ (1.372) = 0.648 \text{ volts} \tag{34}$$

illustrating the fact that the noise level has increased from the strong signal result of Eq. (21).

From Eq. (22) we can compute the new signal voltage into the A/D converter

$$\left(g_o g_{WB_2} \sqrt{\alpha_2} A_2\right)^2 = \left(g_o g_{WB_1} \sqrt{\alpha_1} A_1\right)^2 \downarrow 7 \text{ dB}$$
$$\downarrow 10 \log\left(\frac{32}{4}\right) \uparrow 10 \log\left(\frac{0.99}{0.87}\right)$$
$$\uparrow 20 \log(1.372) \tag{35}$$

where, for example, ↓ 7 dB means reduce the previous quantity by 7 dB. We have

$$g_o g_{WB_2} \sqrt{\alpha_2} A_2 = 0.158 \text{ volts} \tag{36}$$

From Eqs. (13), (34), and (36) the resulting degradation (neglecting the filtering loss) is

$$DEGR\,(A_2) = \frac{\left(1 - \frac{0.01}{0.158}\right)^2}{1 + \frac{1}{12}\left[\frac{0.2625}{0.648}\right]^2} = 0.63 \text{ dB} \qquad (37)$$

Now the filtering losses with the low data rate corresponding to $B_{LP}T = 6.68$ obtained from Ref. 2 is 0.10 dB, so that the total degradation at $E_s/N_o = -4$ dB and at a symbol rate of 4 Msps is given by

$$DEGR_{TOT} = 0.73 \text{ dB} \qquad (38)$$

By comparing Eqs. (26) and (38), it is clear that the high data rates suffer greater BER degradation.

Case II: narrowband noncoherent AGC controlled gain

Now we consider the case where a narrowband noncoherent AGC is used to control the signal-plus-noise level into the A/D converter, rather than a fixed gain in addition to the wideband noncoherent AGC. Using the same initial gain conditions at the high E_s/N_o (3 dB) as in the fixed gain method, we have from Eqs. (21) and (22),

$$g_{NB_1} g_{WB_1} \sqrt{N_o B_{LP}} = 0.472 \text{ volts} \qquad (39)$$

and

$$g_{NB_1} g_{WB_1} \sqrt{\alpha_1} A_1 = 0.684 \text{ volts} \qquad (40)$$

and, further, the same degradations occur as before since the signal voltage and noise voltage are the same. Therefore, the total loss is as before (Eq. (26))

$$DEGR_{TOT} = 0.84 \text{ dB} \qquad (41)$$

Now we consider the effect of both the wideband and narrowband noncoherent AGCs at the lowest symbol rate (4 Msps) and minimum E_s/N_o (-4 dB) for the 26.7-MHz filter. The wideband noncoherent AGC sets the gain as before, from Eq. (33), and increases with weaker input signal by the ratio

$$\frac{g_{WB_2}}{g_{WB_1}} = 1.372 \qquad (42)$$

Notice that the narrowband AGC output power is determined by Eqs. (39) and 40), i.e.,

$$\left(g_{NB_1} g_{WB_1}\right)^2 \left(N_o B_{LP} + \alpha_1 A_1^2\right) = 0.831 \text{ volts}^2 \qquad (43)$$

Now since the wideband noncoherent AGC gain has increased by the factor 1.372 (Eq. 42), we find that the new noise voltage and signal voltage into the narrowband AGC are given by

$$g_{WB_2} g_o \sqrt{N_o B_{LP}} \doteq (1.372)(0.472) = 0.648 \text{ volts} \qquad (44)$$

and

$$\left(g_{WB_2} g_o \sqrt{\alpha_2} A_2\right)^2 = (0.684)^2 \Big(\downarrow 7 \text{ dB} \downarrow 10 \log \left(\frac{32}{4}\right)$$
$$\uparrow 10 \log \left(\frac{0.99}{0.87}\right)$$
$$\uparrow 20 \log (1.372)\Big).$$

or

$$g_{WB_2} g_o \sqrt{\alpha_2} A_2 = 0.158 \text{ volts} \qquad (45)$$

The result of Eq. (45) can be verified using Eq. (1) directly. Now the narrowband noncoherent AGC gain must be set so that the total power given by Eq. (43) is maintained at the output; therefore, it must be true, using Eqs. (43), (44), and (45), that

$$\left(\frac{g_{NB_2}}{g_{NB_1}}\right)^2 \left[(0.648)^2 + (0.158)^2\right] = 0.831 \qquad (46)$$

or

$$\frac{g_{NB_2}}{g_{NB_1}} = 1.37 \qquad (47)$$

Therefore, the signal voltage out of the narrowband noncoherent AGC is given by

$$g_{WB_2} g_{NB_2} g_o \sqrt{\alpha_2} A_2 = 0.158 \times 1.37 = 0.216 \text{ volts} \qquad (48)$$

$$g_{WB_2} g_{NB_2} g_o \sqrt{N_o B_{LP}} = 0.648 \times 1.37 = 0.888 \text{ volts} \qquad (49)$$

From the degradation result, Eq. (13), this new signal and noise level produce a degradation of

$$DEGR\,(A_2) = \frac{\left(1 - \dfrac{0.01}{0.216}\right)^2}{1 + \dfrac{1}{12}\left(\dfrac{0.2625}{0.888}\right)^2} = 0.44 \text{ dB} \qquad (50)$$

The degradation due to filtering when $B_{LP}T = 6.68$ was already found to be 0.10 dB so that the total degradation is

$$DEGR_{TOT} = 0.54 \text{ dB}$$

VI. Conclusions

An initial low-pass pre-A/D converter filter bandwidth versus coded symbol rate table has been constructed and shown to meet the essential BER degradation specification while requiring only six low-pass filter pairs.

Further refinements in the BER degradation will be considered in a later DSN article.

References

1. J. K. Holmes, "Stability Analysis of the DSN Multimegabit Telemetry Demodulator/ Detector Design," *DSN Progress Report* 42-51 (this issue).
2. J. J. Jones, "Filter Distortion and Intersymbol Interference Effects on PSK Signals." *IEEE Transactions on Communication Technology*, *Com-19*, No. 2, April 1971.

Table 1. Candidate filter BW vs symbol rate

Data rate R_S, Msps[a]	LPF BW B_{LP}, MHZ	Sampling rate R_s, Msps[a]	Rate buffer divisor k	R_s/B_{LP}[a]	Samples per Symbol
16 → 32	26.7	64 → 128	1	2.4 → 4.8	4
8 → 16	26.7	64 → 128	2	2.4 → 4.8	8
4 → 8	26.7	64 → 128	4	2.4 → 4.8	16
2 → 4	13.35	32 → 64	4	2.4 → 4.8	16
1 → 2	6.68	16 → 32	4	2.4 → 4.8	16
0.5 → 1	3.34	8 → 16	4	2.4 → 4.8	16
0.25 → 0.5	1.67	4 → 8	4	2.4 → 4.8	16
0.125 → 0.25	0.83	2 → 4	4	2.4 → 4.8	16

[a]The notation 16 → 32 denotes the range of 16 to 32 Msps. For example, a data rate of 32 Msps corresponds to a sampling rate of 128 Msps and a ratio of data rate to low-pass filter BW of 4.8.

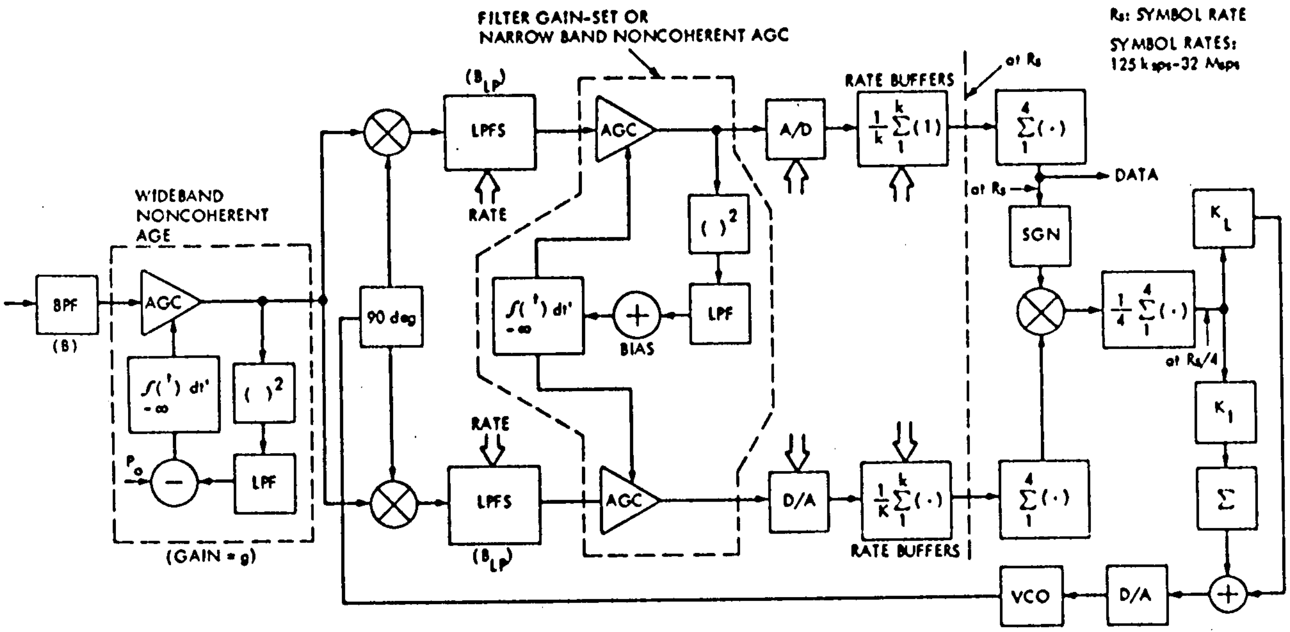

Fig. 1. Partial functional block diagram of the feasibility model, Multimegabit Demodulator/Detector Assembly

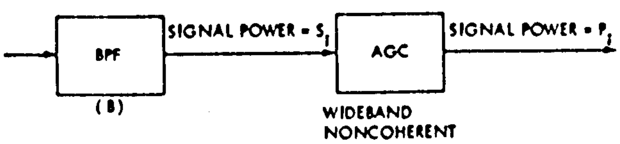

Fig. 2. Block diagram showing AGC input signal power, S_i, and its output signal power, P_i