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The readout of the LHC beam luminosity monitor: accurate shower energy measurements at a 40 MHz repetition rate*

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The LHC beam luminosity monitor is based on the following principle. The neutrals that originate in LHC at every PP interaction develop showers of minimum ionizing particles in the absorbers placed in front of the separation dipoles. The shower energy, measured by suitable detectors in the absorbers is proportional to the number of neutral particles and, therefore, to the luminosity. The principle lends itself to a luminosity measurement on a bunch-by-bunch basis. However, to make such a measurement feasible, the system must comply with extremely stringent requirements. Its speed of operation must match the 40 MHz bunch repetition rate of LHC. Besides, the detector must stand extremely high radiation doses. This paper discusses the solutions adopted to comply with these requirements.

1. Introduction

The instrument described here, whose purpose is the luminosity optimization of the colliding beams at LHC, is based on the following idea [1]. Neutrals created at every PP interaction develop showers of minimum ionizing particles in the absorbers that are placed in front of the separation dipole to shield them from the energy deposition by the neutral flux. If detectors are located inside the absorbers close to where the maximum of the shower is expected, the luminosity can be estimated from the energy released in the detectors. The four quadrant detector segmentation, fig. 1, provides additional information like beam-beam separation at the interaction point and transverse beam shape and size.

The monitoring system is intended for pulsed operation, on a bunch-by-bunch basis, which sets the requirement of an operational speed compatible with the LHC bunch crossing rate of 40 MHz. A further requirement is related to the extremely high radiation dose the detector must stand before a replacement is possible, up to 1 GGy, a dose exceeding by at least two orders of magnitude that expected for detectors in LHC experiments. A preliminary version of the instrument has been realized and tested on the beam at CERN in 2000 and 2001. Though very successful, the tests have clarified a few limitations that have been removed in what is considered to be the final design and whose description is the main purpose of this paper.

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2. Design guidelines

The best detector choice, as far as radiation tolerance goes, appears to be an ionization chamber where the gas is continuously fluxed to prevent a large charge buildup by positive ions.

The sensitive volume of the chamber is split into thin gaps to reduce the distance traveled

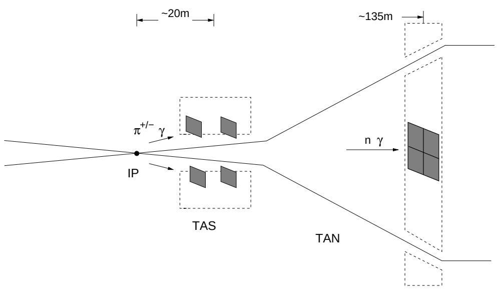

Figure 1. Neutral absorber to protect the separation dipole and segmented detector.

by the charge carriers, electrons, in their motion. The gas mixture must feature an adequately high electron drift velocity, without using organic molecules that may polymerize under the effect of radiation, thus endangering the radiation tolerance of the detector.

A mixture which meets these requirements is Argon added with a few percent of Nitrogen [2].

The charge yield in the shower detection can be raised to the desired value by acting on the pressure of the filling gas.

Very demanding are also the requirements for the front-end electronics.

First, the noise issues are not trivial because the response of the analog channel to a δ-impulse detector current must have a peaking time t_P of a few nanoseconds, as constrained by the 25 ns interbunch period. Secondly, a suitable dynamic range is required. The system must be able to detect from a minimum of one to at least 20 PP interactions. Fluctuations in the energy deposited by the shower may actually extend the range beyond these limits. Thirdly, there are design constraints related to the harsh environment. No active device could survive the radiation field foreseen for the detector site. Therefore, the front-end electronics must be located a few meters away, where the radiation is already reduced to a level which can be tolerated by a *suitably chosen* $front-end\ device$. As shown in fig. 2, where C_D is the detector capacitance and $i(t)$ its current pulse, the coupling between detector and front-end preamplifier, modeled as a charge-sensitive loop with feedback capacitance C_F, is made via a cable, a special radiation hard unit in the actual case.

An active termination with resistive component R_C=$(1/g_m)\cdot C/C_F$, where g_m is the transconductance of the input device, C_F the feedback capacitance and C the internal bandwidth limiting capacitance, is provided by the charge-sensitive loop to match the cable. The following reasons suggest to use bipolar transistors as front-end devices.

- At such short t_P values the signal-to-noise ratio is largely governed by the voltage noise and to a limited extent by the parallel noise associated with the base current.

- The bipolar transistor, whose transconductance g_m is predictable in value and mildly temperature - dependent lends itself to the realization of an accurate active termination.

- The bipolar transistor can be considered radiation hard to a suitable extent. The dominant effect of radiation is a decreased β. This results in an increased parallel noise, whose effect, as previously stated, is not of paramount importance.

The root mean square noise voltage at the output of an analog channel of the type of fig. 2, employing a bipolar transistor as a front-ed element, reaches a minimum at the optimum $I_C t_P$

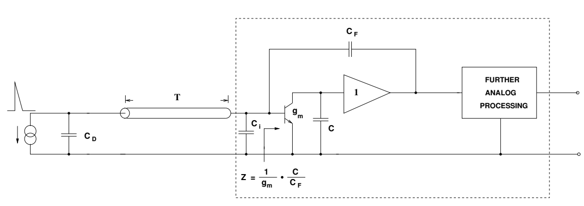

Figure 2. Front-end simplified schematics and detector-front-end coupling.

product given by (1):

$$I_C t_P = \frac{kT}{q} \cdot (C_D + C_i + C_F) \cdot \sqrt{\frac{A_1}{A_3} \cdot \beta}. \quad (1)$$

where q is the electron charge, k is Boltzmann's constant, T the absolute temperature, I_C the collector current, β the dc base-to-collector current gain of the bipolar transistor, C_i its base-emitter capacitance, t_P the peaking time in the δ-impulse response and A_1, A_3 are the coefficients of the analog processor for the noise components with respectively f_0 and f^{-2} frequency dependence at its input. For the sake of simplicity, the thermal noise contribution due to the base spreading resistance $r_{BB'}$ has been neglected. The signal-to-noise ratio is:

$$\eta^2_{opt} = \frac{\lambda^2 \sigma^2 Q^2}{4kT \cdot (C_D + C_i + C_F) \cdot \sqrt{\frac{A_1 A_3}{\beta}}} \quad (2)$$

where the factor σ^2<1 accounts for the ballistic deficit in the case of a detector signal of finite width and λ^2<1 for the degradation due to the cable. To maintain λ^2 within acceptable limits, the detector time constant $C_D R_0$, the cable delay T_{delay} and the peaking time t_P must be such that the ratio $t_P/C_D R_0$ is consistently less than one and the ratio $T_{delay}/C_D R_0$ is as small as possible. These are two important additional design guidelines.

3. Evolution in the detector design

If Q_{GAP} is the induced charge, assumed to be the same for all the gaps by neglecting the statistical fluctuations, a chamber made of m×n gaps of capacitance C_{GAP}, associated with a bipolar transistor front-end optimized according to eqn. 1 provides a signal-to-noise ratio:

$$\eta^2_{opt} = \frac{\lambda^2 \sigma^2 mn Q^2_{GAP}}{4kT C_{GAP} \cdot \sqrt{\frac{A_1 A_3}{\beta}}}. \quad (3)$$

independent of the way the gaps are configured to produce the output signal [4]. The first version of the detector, employed in the beam tests carried out at CERN in years 2000 and 2001 consisted of 60 square gaps of 4 cm side and 0.5 mm thickness, yielding a 28 pF C_{GAP} [5]. The large number of gaps is explained by the endeavor to recognize one PP interaction in one bunch crossing. At a pressure of 1 atm the value of Q_{GAP} corresponding to 1 PP interaction is about 550 electrons. With the values A_1=1.33, A_3=0.39, β=120 relevant to the transistor and the analog processor employed and σ^2=0.2 as determined by the 22 ns basewidth of the triangular detector signal and the 5 ns t_P, the signal-to-noise ratio according to (3) is around 3. An easy way to improve it consists in raising the gas pressure.

A tolerable λ^2 was assumed to be about 0.8, which with the cable parameters T_{delay}=15 ns and R_0=50 Ω, according to the analysis of ref. [3] would occur at a detector capacitance C_D around 50 pF. The gaps in the chamber were accordingly configured as the parallel connection of ten branches, each consisting of six gaps in series, as shown in fig. 3 a). The tests showed that the signal at the output of the analog processor was longer than expected and this was attributed to a capacitance larger than the design value and stray inductances, both aspects related to the very complicated detector layout. At the same

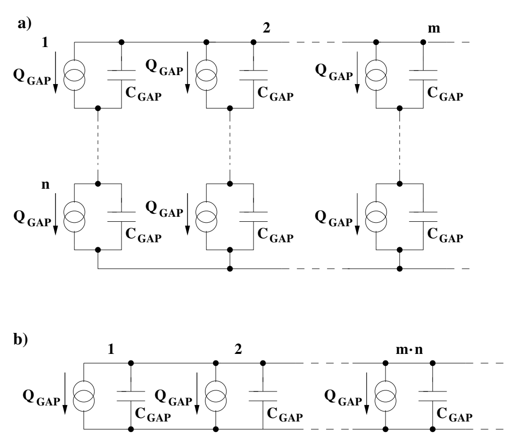

Figure 3. a) Series-parallel connection of the gaps in the first version of the chamber. b) Gap configuration in the second version of the chamber.

time it became also clear that luminosity measurements in the actual LHC operation should be averaged over a certain number of bunch crossings to smooth down the statistical fluctuations in the shower. This consideration relaxes the requirements in terms of signal-to-noise ratio and opens the way to a detector design based on the considerably simplified configuration of fig. 3 b), consisting of six gaps in parallel, which leads to a much simpler layout. Each gap is a square of 4 cm side and 1 mm thickness. Q_{GAP} is twice as much as in the previous case, while C_{GAP} is reduced by a factor of 2. The increased gap thickness would result in a longer signal, but this is partially offset by the use of a faster gas mixture, employing the same gases as before, but in a slightly different proportion. Use of a deconvolution algorithm after the analog signal processing makes the remaining increase in width irrelevant, as pointed out in the next section. The theoretical η_{opt} relevant to the new chamber configuration is only very marginally worse than the previous one, a degradation which can be easily cured by increasing the gas pressure. As an important advantage, the new chamber configuration is expected to be affected to a considerably less extent by the stray capacitances and inductances that impaired the performance of the previous detector.

4. The front-end electronics

The charge-sensitive loop, which hasn't changed from the initial development of the system, is based upon a high transconductance, low spreading resistance input configuration made of four bipolar transistors in parallel, each working at a collector current of 1.25 mA. The transistors are high frequency specimens, BFP540, featuring a β of 120 and $r_{BB'}$ of less than 3 Ω.
The analog processor following the charge-sensitive loop is based upon the following principle. Through iterated pole-zero cancellations it converts the signal at its input into a short pulse, as close as possible to a δ-impulse. Cascaded RC integrations bring this signal to a semigaussian shape, whose basewidth is close to the 25 ns interbunch period. The outstanding cable matching features of the active termination are apparent in fig. 4, result of a bench test done to compare the output signal in the case of detector directly connected to the front-end amplifier and the output signals obtained when cables of two different lengths are inserted between them. The input referred voltage noise density $(\overline{de_N^2}/df)^{1/2}$ of the four transistor connection is about 0.22 $nV/Hz^{1/2}$ and its input capacitance is 1.2 pF. The resulting $C_i\overline{de_N^2}/df$ product is about $5.8{\cdot}10^{-32}$ $J{\cdot}s$, a value difficult

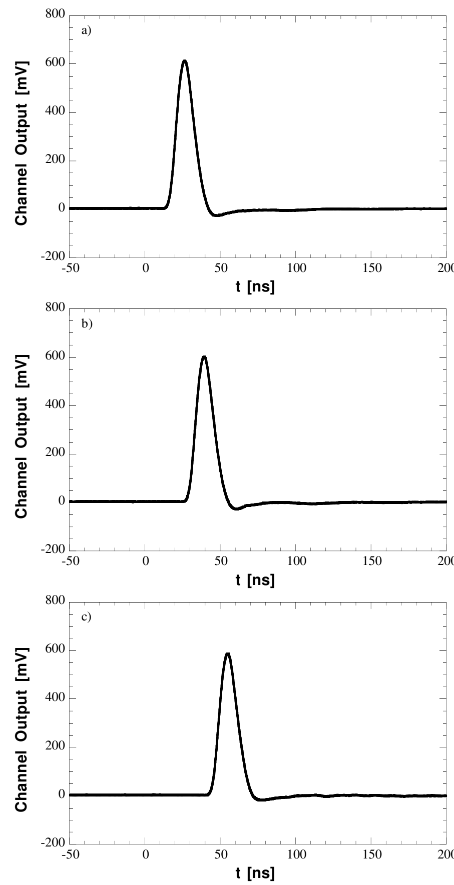

Figure 4. Signals at the processor output in response to a 5 ns long injected long and 47 pF C_D a) absence of cable b) 3 m cable, c) 6 m cable between detector and charge-sensitive loop.

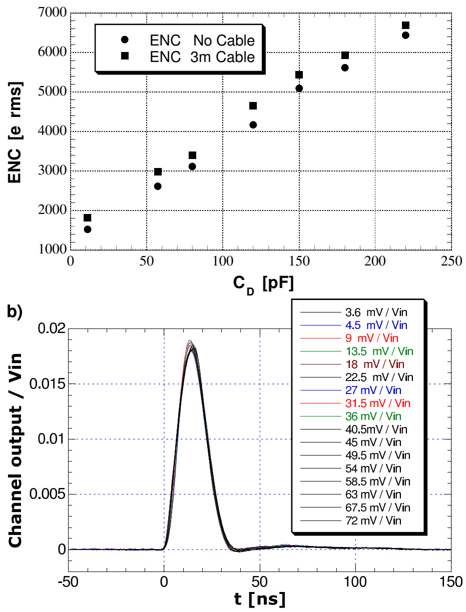

Figure 5. a) Equivalent noise charge as a function of the detector capacitance; b) dependence of the shape of output pulse on the detector charge.

to achieve in a monolithic circuit, which justifies the present discrete implementation. In passing, in its function as an active termination, the simplified charge sensitive loop of fig. 2, adjusted to match a 50 Ω cable, behaves noisewise as a physical resistor of the same value cooled to about 30 K.
The Equivalent Noise Charge ENC is plotted in fig. 5 a) as a function of the detector simulating capacitance C_D in two cases, C_D and charge-sensitive loop are directly connected or a 3 m cable is introduced between them. The cable-induced ENC degradation is around 10%.
The shape of the output signal has a little dependence on the value of the injected charge, as apparent from fig. 5 b). The plots of fig. 5 come from a bench test obtained by reproducing the case of the first detector version, where C_D was increased to about 180 pF by the stray contributions and the basewidth of the detector current was about 20 ns. In the same condition the integral non linearity over the range of charge values spanning 1 through 35 PP interactions has been measured and found to be less than 1%.
The little charge-dependence of the output signal and the satisfactorily linear behavior of the analog channel suggest that the highly demanding situation, where the signal determined by one PP interaction occurs in a time slot following a sequence of a few events corresponding to the largest number of PP interactions, be processed by a deconvolution method. This has been proven by the simulations to be an efficient way to disentangle the small signal from the baseline perturbed by the piling-up of the previous tails. An important advantage is that the constraint on the duration of the detector signal is relaxed, and this supports the decision of doubling the thickness of the gaps, made in the design of fig. 3 b).

5. Conclusions

The system designed for luminosity monitoring at LHC was beam-tested twice. After that a substantial improvement was introduced. The detector was upgraded by the use of a better gap structure and a faster gas mixture. Front-end optimization was completed by the study of a deconvolution algorithm to disentangle potentially ambiguous situations.

6. Acknowledgment

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