

Czech Technical University in Prague

Faculty of Electrical Engineering

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Fast Beam Intensity Measurements for the LHC

Doctoral Thesis

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DEDICATION

Writing this thesis was a long term competition with myself. Competition full of slow and painful learning, searching for correct answers, and frustration in times when the inspiration betrayed me. The winners are however those who sustain.

I want to dedicate this thesis to my wife, Barbora, without whom this thesis might not have been written, and to whom I am greatly indebted. You have been with me all the distressful way through, seamlessly taking care of our family affairs, and providing me with all those essential little things which allowed me to finish the task. Thank you for your love, support, and source of encouragement and inspiration that you have always given me. Thank you as well for your insistence, which boosted me in the rainy days.

I want to dedicate this thesis to my son Erik as well. Thank you for your patience. I promise to spend more time with you.

I love you both

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CONTENTS

1	List of Symbols	7
2	List of Abbreviations	11
3	Introduction	15
4	Current state of the problem	17
4.1	Beam theory as seen by an electronics engineer	18
4.1.1	Beam Current	18
4.1.2	Wall Image Current	23
4.1.3	Particle beams and noise sources	27
4.2	Non-intercepting beam intensity measurements	30
4.2.1	Wall current monitors	30
4.2.2	DC current transformers	32
4.2.3	The fast beam current transformers	33
4.3	Fast beam intensity measurements at CERN	48
4.3.1	Types of the beams produced at CERN	49
4.3.2	LINAC and LEIR FBCTs	50
4.3.3	FBCTs in the transfer lines and circular accelerators	52
4.4	The LHC intensity measurement	56
4.4.1	Functional specification of the intensity measurement using FBCTs	57
4.4.2	The LHC injection scenarios	60
5	Detailed objectives for this work	63
6	Analysis of the requirements for the intensity measurement	65
6.1	Determination of the required dynamic ranges	65
6.1.1	Input signal definition	65
6.1.2	Dynamic range of the beam signal	67
6.2	Determination of the transformer bandwidth	68
6.2.1	The LHC pilot bunch	69
6.2.2	Injection of multiple batches into the LHC	74
6.2.3	The high frequency cut-off estimation	84
7	The FBCTs	91
7.1	Mechanical design of the FBCT	91
7.1.1	Supporting elements	93
7.1.2	Magnetic Circuit of the Measurement device	95
7.2	The vacuum chamber fabrication technological processes	98
7.2.1	The vacuum chamber	98
7.2.2	Titanium coating	100
8	Analysis and implementation of the electronic chain	105
8.1	Acquisition system	107
8.2	RF signal distribution	108
8.2.1	Principle of operation	108

8.2.2	Filtering	111
8.2.3	Hardware implementation	113
8.2.4	Electrical properties	114
8.3	Machine protection system - beam circulating flag detector	117
8.3.1	Hardware implementation	117
8.3.2	Algorithm of detection	118
8.3.3	Physical realisation	119
8.4	The calibration	119
8.4.1	Calibration turn	119
8.4.2	Pulse Current Generator	121
8.4.3	Implementation of the Calibrator	131
8.4.4	Measured results	132
9	Achieved results	135
10	Conclusions and outcomes from the dissertation thesis	141
10.1	My contribution to the work	146
10.2	Theses for future work	147
11	PhD student publications	149
11.1	Publications related to the dissertation	149
11.2	Other publications	149
12	References	151
13	Appendices	157
13.1	Background information	157
13.1.1	Particle accelerators	157
13.2	Simplified mechanical acceptance calculation for the BCTRs	163
13.3	Photographs	167

1 LIST OF SYMBOLS

$\mathcal{L}$	Laplace Operator	
$\mathcal{L}^{-1}$	Inverse Laplace Operator	
ϕ	Unity vector of angular component in cylindrical coordinate system	
ρ	Unity vector of radial component in cylindrical coordinate system	
$d\mathbf{B}$	A vector element of magnetic induction	T
$\mathbf{e}_r$	Unity vector in the radial direction of the cylindrical coordinate system	—
$\mathbf{e}$	Unity vector in Cartesian coordinates: $\langle \mathbf{x}, \mathbf{y}, \mathbf{z} \rangle$	
$\mathbf{j}$	Vector of current density	Am^{-2}
$\mathbf{v}$	Vector of velocity	ms^{-1}
$\mathbf{z}$	Unity vector of z component in cyl. and Cartesian coordinate systems	—
Φ	Unity vector in angular direction of the cylindrical coordinate system	—
$\mathbf{F}_s$	Vector of Lorentz force	N
$\mathbf{B}$	Vector of magnetic flux density	T
$\mathbf{E}$	Vector of electric field	Vm^{-1}
$\mathbf{F}_1, \mathbf{F}_2$	Vectors of force	N
$\mathbf{F}$	Force	N
$\hbar$	The reduced Planck's constant	eVs
α	Generalised angle	rad
β	Relativistic factor	v/c
$\beta_{x,y}$	Optical beta-function	m
δ	Dirac pulse	
δ_p	Momentum offset at the injection	—
ϵ	Time constant	s
ϵ	Transverse emittance at specific energy	$mrad$
ϵ_0	Permittivity of vacuum	Fm^{-1}
ϵ_n	Normalised transversal emittance	$mrad$
γ	Relativistic factor	—
λ_N	Number of charges per unit length	m^{-1}
μ	Absolute permeability	Hm^{-1}
μ_0	Permeability of vacuum	Hm^{-1}
μ_r	Relative permeability	—
ν	Time constant of a first order low-pass filter	s
ω	Angular frequency	$rad s^{-1}$
ω_0	Angular revolution frequency	$rad s^{-1}$
ω_s	Synchrotron oscillation angular frequency	$rad s^{-1}$
$\phi(t)$	Gaussian distribution function	
σ	Longitudinal RMS beam size	m
σ_x	Standard deviation of x	
$\sigma_{x,y}$	Transversal RMS beam size	m
τ	Displacement operator in a convolution process	

τ	Time constant of RC circuit	s
$\tau(t)$	Beam lifetime	s
τ_s	Current amplitude of Synchrotron oscillation	A
θ	Angular deflection	rad
$\theta(t)$	Rounding function	
ζ	Elevation	rad
a	Vacuum chamber radius or general distance	m
a, b	Constants related to bunch width	s^{-1}
a, b	Toroid core outer and inner diameter	m
c	Speed of light	ms^{-1}
c_k	Calibration constant	—
dl	A length element	m
e	Elementary charge	C
$e(k)$	Measurement relative error in bunch slot k	%
$e_{c,0}$	Relative error for the first bunch injected into the LHC	%
$e_{p,0}$	Relative error for the LHC pilot bunch	%
$\bar{x}$	Mean value of x	
f_0	Revolution frequency	Hz
f_c	Low frequency cut-off of a filter or toroid transformer	Hz
f_t	Cut-off frequency of a filter	Hz
$g(s)$	Laplace image of first order low-pass filter.	
$g_1(t), g_2(t)$	Partial solution of the convolution integral calculating response of the measurement device to the excitation by a beam.	
h	Capacitive pickup electrode length or toroid core width	m
$h(s)$	Laplace image of a second order filter transfer function	
$H(t)$	A second order filter transfer function	
$i_b(t)$	Instantaneous beam current	A
$i_g(t)$	Instantaneous bunch current	A
$i_o(t)$	Instantaneous beam current including synchrotron oscillation effects	A
$i_s(t)$	Instantaneous beam current orbiting on a circular trajectory	A
$i_{mb}(t)$	Instantaneous beam current of multi-bunch beam	A
j	Imaginary number	
k	Harmonic number or bunch slot number	—
k_β	Beta-beating factor	—
k_b	Boltzmann constant	JK^{-1}
k_D	Parasitic dispersion factor	—
l	Length	m
m	Bunch number	—
m	Mass	kg
m_e	Electron mass	kg
m_p	Relativistic mass of the proton	eV
n	Period number	—

n_1, n_2	Collimator aperture sizes	m
p	Momentum	Ns
q	Charge	C
q_n	Generalised charge	C
q_{eb}	Charge present at capacitive pickup plates	C
r	Distance or radius	m
s	Laplace operator	
s_d	Total mechanical displacement	m
s_f	Surface enclosed by the shape of the ideal bunch	C
s_{CO}	Closed orbit beam excursion	m
$s_{p,0}$	Surface enclosed by the shape of the pilot bunch	C
t	Time	s
u	Generic type B uncertainty	
u_A	Type A uncertainty	
u_B	Type B uncertainty	
u_C	Combined type A and B uncertainty	
v	Velocity	ms^{-1}
Δf	Bandwidth	Hz
Γ	Calculated calibration constant	$bins^{-2}$
∇	Nabla, mathematical operator	
Φ	Magnitude of magnetic flux	Wb
$\Phi(\omega)$	Fourier transform of Gaussian distribution function	
Φ_T	Magnitude of magnetic flux generated in a toroid	Wb
Θ	Calculated calibration offset	—
A	Amplifier gain	—
A	Measured binary value corresponding to a calibration current	$bins$
A_l	Magnetic material specific constant, Inductance per square root	H
A_m	Relative charge	—
B	Magnetic induction	T
C	Capacitance	F
$D_{parasitic}$	Parasitic dispersion of the beam	m
E	Energy	eV
E	Magnitude of electric field	Vm^{-1}
$E(k)$	Absolute error of the measurement in bunch slot k	—
E_0	Particle rest energy	eV
E_i	Injection efficiency	%
E_p	Beam energy	eV
E_x	X component of vector of electric field	Vm^{-1}
E_y	Y component of vector of electric field	Vm^{-1}
E_z	Z component of vector of electric field	Vm^{-1}
$E_{c,0}$	Absolute error for the first bunch injected into the LHC	—

$E_e(t)$	Absolute error caused by transient response of the transformer after the injection	—
E_{kin}	Kinetic energy of particle	eV
$E_{p,0}$	Absolute error for the LHC pilot bunch	—
$E_{p,k}$	Absolute error for the bunch slots outside of the LHC pilot bunch position	—
F_c	Magnitude of centrifugal force	N
I	Magnitude of current	A
I_b	Equivalent DC beam current	A
$I_g(\omega)$	Fourier transform of an instantaneous bunch current	A
$I_o(\omega)$	Fourier transform of $i_o(t)$	
I_s	Magnitude of the beam current transformed at secondary winding of measurement transformer	A
$I_s(\omega)$	Fourier transform of current $i_s(t)$	
I_w	Wall image current	A
I_{avg}	Modeled average beam current	A
$I_{mb}(\omega)$	Fourier transform of $i_{mb}(t)$	
I_{peak}	Modeled bunch peak current	A
$J_m(x)$	m^{th} order Bessel function	
K	Number of bunches in a single injection	—
L	Inductance	H
L_k	Inductance of k^{th} toroid winding	H
N	Number of bunches or periods	—
N_k	Number of turns of k^{th} transformer winding	—
N_p	Number of charges	—
O_{acq}, O_k	Offsets	
P_s	Signal power of the beam current transformer at secondary winding of measurement transformer	W
P_{eff}	Thermal noise power	W
Q	Charge	C
R_0	Trans-impedance amplifier feedback resistance, load resistance	Ω
R_i	Input resistance of trans-impedance amplifier	Ω
R_s	Resistance of toroid winding	Ω
T	Temperature	K
T_0	Revolution period or bunch slot length	s
T_m	Time of arrival of bunch m in multi-bunch beam	s
U	Extended uncertainty	
U_{rel}	Relative extended uncertainty	%
V_{eff}	Voltage noise	V
Z_b	Magnitude of the beam impedance	Ω
V_m	Voltage measured on calibrator's reference resistance	V
$V_{m,h}$	V_m for high sensitivity range	V
$V_{m,l}$	V_m for low sensitivity range	V

2 LIST OF ABBREVIATIONS

AC	Alternate Current
AD	Anti-proton Decelerator
ADC	Analog to Digital Converter
ASIC	Application specific integrated circuit
BCF	Beam Circulating Flag
BCT	Beam Current Transformer
BCTR	Beam Current Transformer for the LHC ring
BCTD	Beam Current Transformer for the LHC dump
BNL	Brookhaven National Laboratory
BCF	Beam Circulating Flag
BST	Beam Synchronous Timing
CERN	European Organisation for Nuclear Research
CFD	Constant Fraction Discriminator
CFOA	Current-feedback Operational Amplifier
CLIC	Compact Linear Collider
CPLD	Complex Programmable Logic Device
CTE	Coefficient of Thermal Expansion
CTF3	CLIC Test Facility 3
CNGS	CERN Neutrinos to Gran Sasso
CPU	Central Processing Unit
DAC	Digital to Analog Converter
DAB	Digital Acquisition Board
DC	Direct Current
DCCT	DC current transformer
DIRAC	DI-meson Relativistic Atom Complex, a CERN experiment
DTL	Drift Tube Linac
ECL	Emitter Coupled Logic
EASTB	Beam produced for DIRAC experiment
EASTC	Beam produced for radiation test facility
EB	Electron-Beam
EMI	electromagnetic interference
FBCT	Fast Beam Current Transformer
FPGA	field programmable gate array
FT	Fourier Transform
FS	full scale
FSR	full scale range

FWHM	Full Width at Half Maximum
GSPS	giga-samples per second
HF	high-frequency
HIGAIN	high gain
HIBW	high bandwidth
HV	high voltage
IBMS	Individual Bunch Measurement System
IP4	LHC access point 4
IP6	LHC access point 6
ISOGPS	Beam produced for ISOLDE General Purpose Separator experiment
ISOHRS	Beam produced for ISOLDE High Resolution Separator experiment
ISOLDE	On-Line Isotope Mass Separator
LEIR	Low Energy Ion Ring
LEP	Large Electron and Positron collider
LF	low-frequency
LHC	Large Hadron Collider
LHCb	LHC beauty, a CERN experiment
LHCb2002	An improved version of the LHCb integrator chip
LINAC	Linear Accelerator
LOBW	low bandwidth
LOGAIN	low gain
LUT	look-up table
MDPRO	Machine Development Probe, one of test beams used at CERN for machine developments
MPS	Machine Protection System
NEG	non-evaporable getter
nTOF	Beam produced for the neutrino Time of Flight experiment
OA	operational amplifier
OFE	Oxygen Free Electronic Copper
OFS	Oxygen Free Copper with Silver
OTR	Optical Transition Radiation
PET	Positron Emission Tomography
PCB	Printed Circuit Board
PPC	Power PC
PS	Proton Synchrotron
PSB	Proton Synchrotron Booster
PWM	Pulse Width Modulation
RF	Radio Frequency

RFQ	Radio Frequency Quadrupole
RHIC	Relativistic Heavy Ion Collider
RMS	Root Mean Square
RTOS	Real-time operational system
SEM	Secondary Emission
SPS	Super Proton Synchrotron
STD	Standard Deviation
TDR	time-domain reflectometer
TS/MME	Technical Support/Mechanical and Material Engineering
UHV	Ultra High Vacuum
VFOA	Voltage-feedback Operational Amplifier
VHDL	Very High Speed Integrated Circuit Hardware Description Language
VME	Virtual Machine Environment
VME64x	64 bit version of the VME
WCM	Wall Current Monitor
YAG	Yttrium Aluminium Garnet

3 INTRODUCTION

Particle accelerators are constructed and operated for a wide variety of applications. In particle physics - the branch of physics that studies the elementary constituents of matter and forces between them - high energy accelerators are used to look deep into the structure of matter. Medical particle accelerators are used for example in medicine to treat tumours [31], in imaging techniques such as Positron Emission Tomography (PET) [24], or for the radio-isotopes production. They also serve in many other industrial branches, e.g. geology, radiocarbon dating [39], molecular complex spectroscopy, lithography, food preservation etc. The field of accelerator technology draws knowledge and expertise from a wide range of scientific disciplines and uses the latest technical knowledge. The incomplete list of covered disciplines includes mathematics, physics, electronics, computing, electromagnetic field technology, microwave technology, cryogenics, vacuum technology, special materials, mechanical engineering or civil engineering.

Effective control of an accelerator requires numerous types of diagnostic tools. The tools providing an information about the beam parameters are called “beam diagnostics”. They include many measurement techniques which could be grouped into two major branches:

- *Intercepting measurements*, which are destructive for the beam, or they result in absorption of a significant amount of its energy. These include e.g. wire scanners (monitors which detect longitudinal profile of the beam of particles) [6], Secondary Emission (SEM) grids (measurement of transversal beam profile) [9], Optical Transition Radiation (OTR), scintillator or Yttrium Aluminium Garnet (YAG) screens (measurement of beam size and position) [75], and few techniques of beam intensity measurements as Faraday cup measurement [29].
- *Non-intercepting measurements*, which use electric or magnetic field coupling of beam to the measuring instrument. These include e.g. a beam position measurement [12], emittance and acceptance measurements, beam loss measurement, luminosity [7] [8] and tune measurements [13], and capacitive or inductive beam intensity measurements (either AC or DC).

Currently, a new and unique accelerator is being commissioned at the European Organisation for Nuclear Research (CERN), Switzerland - the Large Hadron Collider (LHC).

The LHC firstly started the beam production at the end of 2008. A year later, in November 2009, it had already set the new record in the achieved beam energy - the protons were accelerated to 1.2 GeV. The LHC’s unique parameters - circumference of 27 km, extreme collision energy (14 TeV), very fine spatial bunches distribution and separation (beam longitudinal RMS size from 280 ps to 680 ps repeated at 25 ns) - require substantial changes in the fast beam intensity measurement methods currently used at CERN.

This doctoral thesis aims to design, optimise and implement the fast (AC) beam intensity measurement system to measure the intensity of the LHC circulating beams.

4 CURRENT STATE OF THE PROBLEM

Fast beam intensity measurement is a process whose result is an information about a number of particle beam's charges. The directly measured quantity is a beam current, number of charges is a calculated value. An integral of the measured beam current during a specific time results in a beam charge. Number of charges is expressed as the beam charge divided by the elementary charge. As the measurement monitor is sensitive to the beam current it is called a fast beam current monitor, or a Beam Current Transformer (BCT) if using a toroid to acquire the beam signal.

The beam current monitors fall into the category of non-intercepting measurements. The monitors are inductively coupled to the beam providing an electrical current proportional to the beam current. Two types of beam current monitors are recognised: DC and AC. The DC monitors use a magnetic feedback to measure the beam current. They are slow compared to the AC monitors and used exclusively for circulating beam measurements. The DC monitors provide excellent measurement precision. The AC monitors are fast, usually have wide bandwidth and provide bunch by bunch measurements. They were primarily used only on the transfer lines due to problems related to a DC signal restoration (see later). By using new fast electronics and advanced digital algorithms this limitation was overcome and the Fast Beam Current Transformers (FBCTs) could be used also to measure the circulating beam.

The beam intensity measurement is used to define the intensity loss at the injection, acceleration and the extraction (transfer efficiency). Slow charge loss measurement provides an information about lifetime of the circulating beam. The information can be used as well to protect the machine or humans against machine malfunctions [65]. Number of charges is also very important for the physicists.

Different accelerators operate with different beams resulting to a wide range of constraints put on the measurement equipment. Most critical demands are related to the accuracy and the dynamic range of the measurement. High accuracy and high bandwidth of the measurement is required to measure individual bunches. Precision and speed is required to calculate the beam lifetime.

A great variety of beam current monitors covering wide dynamic ranges and frequency bands exists [15], [14]. Currents ranging from several kilo-amperes in induction Linear Accelerators (LINACs) [2] to some micro amperes e.g. in CERN's anti-proton decelerator [36], [44] are measured. The bandwidth of the measurement device is given by properties of the measured current. The bunch width and the bunch spacing mainly affect the high-frequency (HF) cut-off, the revolution period defines low-frequency (LF) cut-off requirements. A 600 MHz bandwidth was required to measure sub-nanosecond pulses generated at the CERN Large Electron and Positron collider (LEP) (≈ 200 ps). The CLIC Test Facility 3 (CTF3) requires bandwidth greater than 7 GHz [54]. Duty cycle affects the measurement bandwidth as well. Low duty cycle particle beams were accelerated e.g. in LEP (bunch spacing $11.1 \mu\text{s}$) [19],[38],[21]. High duty-cycle beams e.g. for fixed

Machines	LHC	SPS	CPS	PSB
Extraction Energy [GeV]	7000	450	26	1.4
Injection Energy [GeV]	450	14	1.4	0.31
Circumference/ 2π [m]	4242.89	1100	100	25
Revolution period T_{rev} [μ s]	88.92	23.1	2.1	0.6
Revolution frequency [kHz]	11.25	43.3	475	1700
Typ. bunch length 2σ [ns]	0.28-0.68	2	14	55
Number of charges per bunch	11.5	11.5	100	150
N [$\times 10^{10}$]				
Max. number of bunches [-]	2808	4620	20	5
Min. bunch spacing T_{per} [ns]	25	5	not fixed	300-600

Table 4.1: *Partial parameter list for various machines at CERN. Parameters of LHC correspond to optics version 6.4 from 15. October 2003*

target at the Super Proton Synchrotron (SPS) (5 ns bunch spacing, ≈ 2 ns bunch length). The LHC, with its 25 ns bunch spacing and 0.6 ns bunch length, provides the high duty cycle particle beams as well.

Table 4.1 summarises parameters related to the beam intensity measurement for various CERN accelerators. For detailed information on other parameters of the CERN machines refer to [20] and [56].

4.1 Beam theory as seen by an electronics engineer

Following sections present some mathematical aspects of the fast beam current measurements. The emphasis is put on discussion of the beam signal and properties of corresponding measured signal.

4.1.1 Beam Current

A moving charged particle creates a current called *the beam current*. Two approaches are presented to describe mathematically the beam current. The first approach does not take into account an inner structure of the beam. The beam is treated as continuous flow of charges with an average DC current. Is used as an approximative model, mainly for Linear Accelerators (LINAC), as their beam current can be liken to a simple pulse current source. The second approach takes into account the bunch structure and periodicity of the signal. This is required due to beam currents present in the circular accelerators.

a) DC beam as a current source:

The flow of charged particles can be seen as a current source with almost infinite internal resistance. The amplitude of the current source is proportional to the number of charges passing through the monitor. This can be described by following set of implications:

$$Q = I_b \cdot t = N_p \cdot e \rightarrow I_b = \frac{N_p e}{t} = \frac{N_p e \beta c}{l}, \quad (4.1)$$

where e is the elementary charge, N_p number of charges, l is the length over which we express the current, c is speed of the light and relativistic β is defined as:

$$\beta = \frac{v}{c} \quad (4.2)$$

The equation expresses a DC current of a number of charges longitudinally and uniformly distributed over the length l . The beam current, normalised to the longitudinal parameter, can be expressed as:

$$I_b = e \lambda_N \beta c, \quad (4.3)$$

where λ_N is the number of charges per unit length. In high-energy accelerators the relativistic β could be neglected as it converges to one.

The impedance of the source is defined as $\frac{\Delta V}{\Delta I}$, where ΔV represents change in the voltage due to an external perturbation and ΔI is a corresponding current change. A relativistic beam¹ represents nearly perfect current source as change in current due to voltage change is very small. The voltage used to generate an accelerating electric field changes the beam's energy hence also the velocity. The magnitude of the effective impedance of the beam as a current source can be expressed in terms of change of the energy of the particles as [40], [41]:

$$Z_b = \left(\frac{di}{dV} \right)^{-1} = \left[(e \cdot \lambda_N) \cdot \left(\frac{d\beta c}{dE} \right) \cdot \left(\frac{dE}{dV} \right) \right]^{-1} = \frac{\beta^2 \cdot \gamma^3 \cdot E_0 / e}{I_b}, \quad (4.4)$$

where e is the elementary charge and E_0 is the proton rest energy. The β and γ are the relativistic parameters, and γ is defined as follows:

$$\gamma = \frac{1}{\sqrt{1 - \beta^2}} = \frac{c}{\sqrt{c^2 - v^2}} \quad (4.5)$$

The impedance Z_b is non-linear and inversely proportional to the beam current itself.

¹A relativistic beam is assumed when $\gamma > 10$ (see Eq. 4.5).

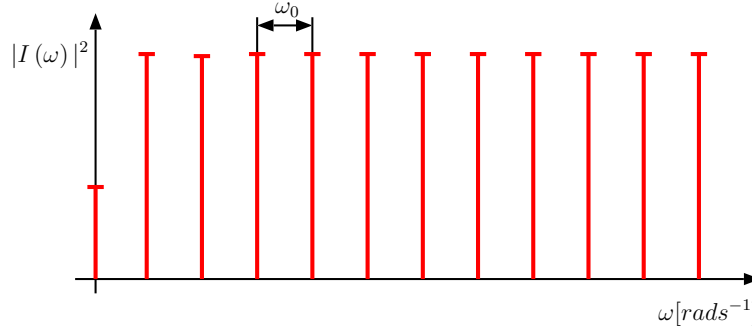


Figure 4.1: Single particle orbiting on optimal circular trajectory generates infinite comb type spectrum.

b) Bunched beam and its representation in time and frequency domain:

In this section a mathematical model describing properties of the circulating beam current signal is shown. Detailed explanation can be found in [32].

Let us consider a particle circulating in a storage ring on an optimal circular trajectory. The revolution period T_0 is linked to the revolution frequency: $f_0 = 1/T_0$. The current observed at a fixed point can be described as:

$$i_s(t) = e \sum_n \delta(t - nT_0) = e\omega_0 \sum_n e^{jn\omega_0 t}, \quad (4.6)$$

where δ represents the Dirac pulse and $\omega_0 = 2\pi f_0$ angular revolution frequency. The equation can be interpreted in terms of Fourier Series:

$$i_s(t) = ef_0 + 2ef_0 \sum_{n=1}^{\infty} \cos(n\omega_0 t), \quad (4.7)$$

The signal spectrum is obtained by applying the Fourier Transform (FT) to Eq. (4.6) and (4.7) respectively:

$$I_s(\omega) = e\omega_0 \sum_k \delta(\omega - k\omega_0) \quad (4.8)$$

$$I_s(\omega) = e\omega_0 \left(\delta(0) + 2 \sum_{k=1}^{\infty} \delta(\omega - k\omega_0) \right) \quad (4.9)$$

where k is a harmonic number (e.g. [88]). The resulting signal spectrum has a form of infinite comb as shown in Fig. 4.1.

To precise the model an effect called synchrotron oscillation must be introduced. Synchrotron oscillation is caused by a presence of an Radio Frequency (RF) accelerating voltage and results in a modulation of the arrival time by synchrotron angular frequency. Equation (4.6) becomes:

$$i_o(t) = e \sum_n \delta(t - nT_0 + \tau_s \cos(\omega_s t)), \quad (4.10)$$

where τ_s and ω_s are the synchrotron oscillation amplitude and angular frequency. Applying the FT on Eq. (4.10) results in the frequency spectrum:

$$I_o(\omega) = e\omega_0 \sum_n e^{-jn\omega_0(t + \tau_s \cos \omega_s t)} \quad (4.11)$$

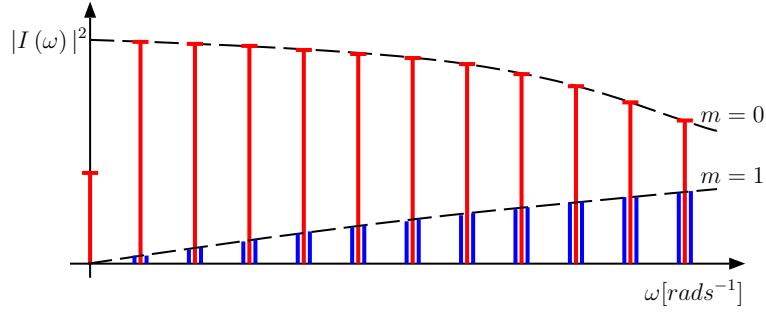


Figure 4.2: Signal spectrum of a single particle affected by the synchrotron oscillation. Particle is orbiting on a circular trajectory.

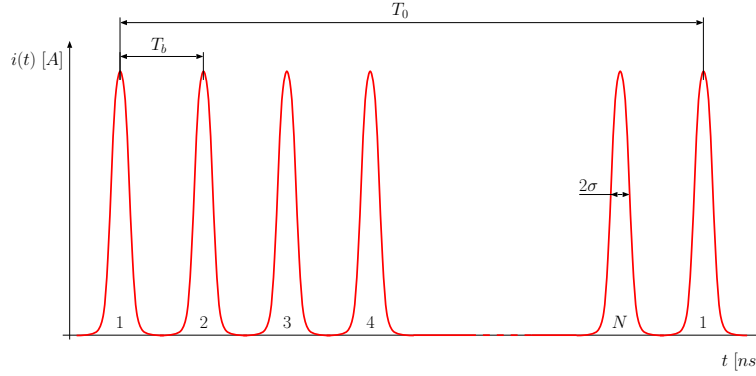


Figure 4.3: Longitudinal distribution of bunches as seen by a BCT in a circular accelerator. T_0 corresponds to the revolution period, T_b is bunch spacing and σ is the longitudinal RMS size of the beam.

The expression $e^{jx \cos \Phi}$ is replaced by:

$$e^{jx \cos \Phi} = \sum_m j^{-m} J_m(x) e^{jm\Phi}, \quad (4.12)$$

where $J_m(x)$ is the m^{th} order Bessel function. Equation (4.11) becomes:

$$I_o(\omega) = e\omega_0 \sum_m j^{-m} J_m(\omega\tau_s) \sum_k \delta(\omega + m\omega_s - k\omega_0) \quad (4.13)$$

The frequency spectrum is shown in Fig. 4.2. The synchrotron oscillation acts as an amplitude modulator. It adds side-bands to the harmonics of the revolution angular frequency. The amplitude of the harmonics follows the 0^{th} order Bessel function while the amplitude of the side-bands is determined by higher order Bessel functions. Therefore a frequency at which the revolution frequency harmonics disappears exists.

The accelerated beams never contain only a single particle. The particles form bunches with a normal distribution as shown in Fig. 4.3. This simplification is widely used as it approximates the distribution of the beam in the longitudinal plane².

²Sometimes a “cos²” model is used to avoid problems caused by undefined limits of the Gaussian curve.

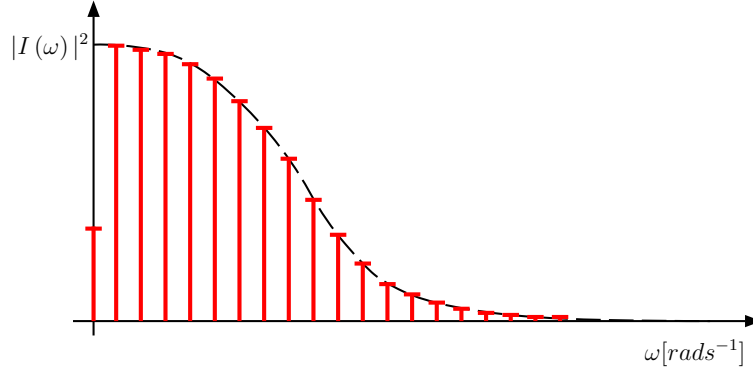


Figure 4.4: The envelope of the harmonics in the signal spectrum is given by a Fourier image of the distribution function.

The current produced by the beam is a convolution of a distribution function with the current itself:

$$i_g(t) = \int_t \phi(\tau) i(t - \tau) d\tau, \quad (4.14)$$

where $\phi(t)$ is the Gaussian distribution function. Signal spectrum is defined by the Fourier transform:

$$I_g(\omega) = e\omega_0 \Phi(\omega) \sum_k \delta(\omega - k\omega_0), \quad (4.15)$$

where $\Phi(\omega)$ corresponds to the Fourier image of the distribution function.

The signal spectrum is given by a comb structure as in the previous cases. The amplitude of harmonics is determined by the Fourier image of the distribution function. Assuming Gaussian distribution function in Eq. (4.15) results in a concentration of the signal energy into a relatively narrow region as depicted in Fig. 4.4.

Should the BCT see individual bunches, higher harmonics of the signal must be captured. The number of required harmonics increases with decreasing longitudinal RMS size of the bunch. The bunches “shrink” with the acceleration hence the worst case scenario corresponds to the extraction of the accelerated beam. In high energy accelerators usual pulse widths are some hundreds of picoseconds and the bandwidth required of the BCT therefore exceeds 1 GHz.

The accelerators work often in multibunch modes. The total current can be then expressed as a sum of the currents passing the observation point (for single particle) [32]:

$$i_{mb}(t) = e \sum_{m=1}^{m=N} \sum_n A_m \delta(t - nT_0 - T_m) = e \sum_{m=1}^{m=N} \sum_n e^{jn\omega_0 t} e^{jn\omega_0 T_n}, \quad (4.16)$$

where N is number of bunches, A_m is relative charge and T_m is the time of arrival of bunch m . The FT of (4.16) for symmetric filling results in:

$$I_{mb}(\omega) = Ne\omega_0 \sum_k \delta(\omega - kN\omega_0) \quad (4.17)$$

The signal spectrum is the same as that of a single bunch, but the fundamental harmonic frequency is factor of N higher with respect to the revolution period.

4.1.2 Wall Image Current

The wall image current is used by several measurement methods to evaluate various parameters of the beam. However, the wall image current disturbs the BCTs. Following paragraphs clarify occurrence and properties of the wall image current.

a) Electric field of moving particle:

The Coulomb's law describes the force acting on two charges:

$$\mathbf{F}_1 = \frac{1}{4\pi\epsilon_0} \frac{q_1 q_2}{r_{12}^2} \mathbf{e}_{12} = -\mathbf{F}_2, \quad (4.18)$$

where $\mathbf{F}_1$ represents the vector of the force applied to q_1 , $\mathbf{e}_{12}$ is the unit vector in the direction from q_2 to q_1 and r_{12} is the distance between charges q_1 and q_2 . ϵ_0 is the permittivity of the vacuum. The force applied to q_2 has the same magnitude but an opposite direction. If there are more than two charges in the system, the principle of superposition applies - the force on the charge is given by a vector sum of the forces from each charge present.

The space surrounding an electric charge has a property called an electric field. The electric field $\mathbf{E}(1)$ is equal to the force acting on q_1 belonging to the unity charge. Hence electric field for spatial coordinate of q_1 is obtained by dividing equation (4.18) by the charge q_1 :

$$\mathbf{E}(1) = \frac{1}{4\pi\epsilon_0} \frac{q_2}{r_{12}^2} \mathbf{e}_{12} \quad (4.19)$$

This is also valid when the charge q_1 does not exist, provided that all other charges remain at their places. We say that $\mathbf{E}(1)$ is the electric field at spatial coordinate '1' [85].

The equipotential surface of the electric field forms a sphere which is concentric with the centre of mass of the charge. The direction of the electric field produced by charge q is radial, having a unity vector $\mathbf{e}_{12}$ perpendicular to the unity surface of the sphere as seen in Fig. 4.5.

The electric field of the particle travelling at non-relativistic velocity is not significantly affected by the motion. This is not the case for the high energy accelerators as

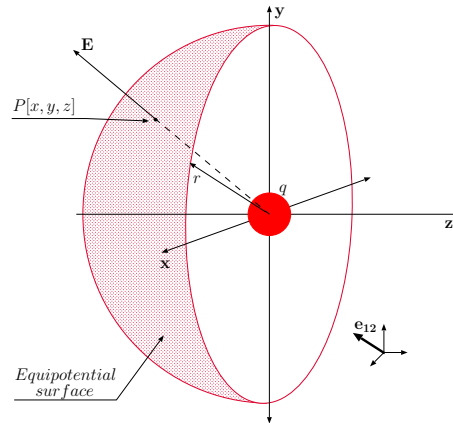


Figure 4.5: Electric field of a particle having a charge q

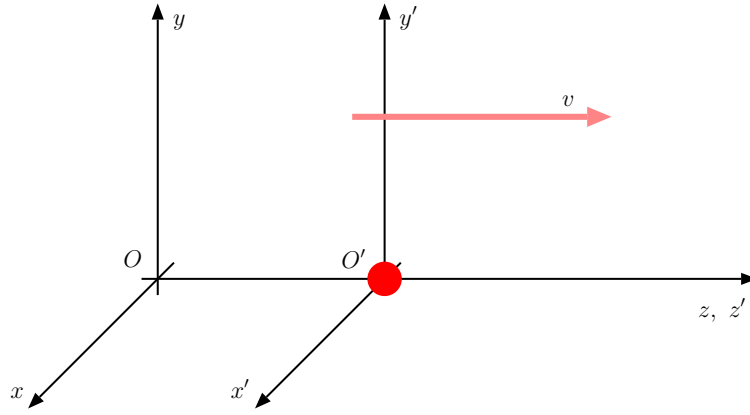


Figure 4.6: Lorentz transformation

velocity of the particles quickly approaches the speed of light and the Einstein's special theory of relativity and Lorentz transformation must be taken into account.

The special theory of relativity defines different observation frames for an object moving at relativistic velocity. Figure 4.6 depicts the situation. Assume an elementary particle moving at constant velocity v along the axis z . Particle is observed by observers from different frames. The rest frame O is still. The frame O' is moving with the same velocity and direction as the particle.

The particle is observed from the moving frame O' . It is in a rest state hence observer "sees" a particle producing no magnetic field. The electric field is given by Eq. (4.20):

$$\mathbf{E}' = \frac{e}{4\pi\epsilon_0} \cdot \frac{\mathbf{r}'}{r'^3}, \quad (4.20)$$

where $\mathbf{r}'$ is the vector pointing from the beginning of the coordinate system of the moving frame O' to the particle.

Observed from the frame O , the particle moves. The Lorentz transformation must be applied. It is defined by following set of equations:

$$\begin{aligned} E_x &= \gamma (E'_x + vB'_y) & B_x &= \gamma \left(B'_x - \frac{vE'_y}{c^2} \right) \\ E_y &= \gamma (E'_y - vB'_x) & B_y &= \gamma \left(B'_y + \frac{vE'_x}{c^2} \right) \\ E_z &= E'_z & B_z &= B'_z \end{aligned} \quad (4.21)$$

where v is the velocity of the particle and γ and β are relativistic factors.

Note the result of Eqns (4.21): Pure electric or magnetic fields in one frame do not necessarily create adequate separate fields in the other one. At relativistic speeds the resulting fields are a mixture of both of them.

To provide Eq. (4.19) in the rest frame O a magnitude of vector $\mathbf{r}'$ must be defined:

$$r' = \left(x'^2 + y'^2 + z'^2 \right)^{\frac{1}{2}} \quad (4.22)$$

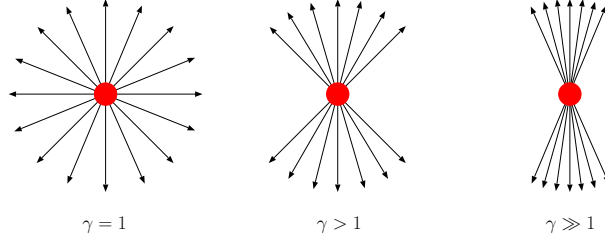


Figure 4.7: *Deflection of the electric field seen from the rest frame*

It can be expressed by parameters of the rest frame O as well:

$$x' = x \quad (4.23)$$

$$y' = y \quad (4.24)$$

$$z' = \gamma(z - vt) \quad (4.25)$$

hence

$$r' = (x^2 + y^2 + \gamma^2(z - vt)^2)^{\frac{1}{2}} \quad (4.26)$$

Assuming there is no magnetic field produced by the particle observed in moving frame O' the “magnetic part” of the equation set (4.21) can be neglected as $\mathbf{B}' = \mathbf{0}$:

$$\mathbf{E} = \langle \gamma, \gamma, 1 \rangle \circ \mathbf{E}' \quad (4.27)$$

The symbol “ $\circ$ ” denotes Hadamard product of the column vector $\mathbf{E}'$ with the row vector of Lorentz transformation factors. Substituting Eqns (4.26), (4.20) and (4.27) we obtain equation for electric field of moving particle:

$$\mathbf{E} = \frac{e}{4\pi\epsilon_0 [x^2 + y^2 + \gamma^2(z - vt)^2]^{\frac{3}{2}}} \cdot \left\{ \langle \gamma x, \gamma y, \gamma(z - vt) \rangle \circ \mathbf{e} \right\}, \quad (4.28)$$

where $\mathbf{e}$ is a unity vector in Cartesian coordinates. At $t = 0$ Eq. (4.28) becomes:

$$\mathbf{E} = \frac{\gamma e}{4\pi\epsilon_0 [x^2 + y^2 + \gamma^2 z^2]^{\frac{3}{2}}} \mathbf{r}, \quad (4.29)$$

where $\mathbf{r}$ is the vector pointing from the beginning of the coordinate system O to the moving particle.

The analysis of the equation shows that the motion of the particle affects the electric field generated. The field keeps the symmetry in the transversal plane, however the component in the direction of movement is contracted. The contraction factor depends on the speed of the particle observed in a stationary frame O . For non-relativistic speeds this effect becomes negligible as seen in Fig. 4.7.

The coordinate system can be redefined in terms of a distance from the particle (r), and an elevation angle (ζ) as shown in Fig. 4.8. Using following substitution:

$$\begin{aligned} x^2 + y^2 &= r^2 \sin^2 \zeta \\ z &= r \cos \zeta \end{aligned} \quad (4.30)$$

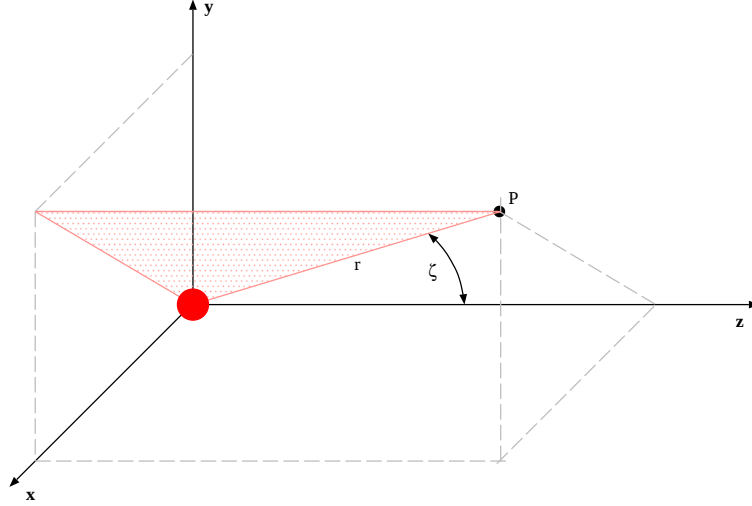


Figure 4.8: Coordinate transformation used to calculate the electric field of moving particle from the relative velocity β .

the electric field is expressed by the relative velocity β :

$$\mathbf{E} = \frac{e}{4\pi\epsilon r^3} \cdot \frac{1 - \beta^2}{[1 - \beta^2 \sin^2 \zeta]^{\frac{3}{2}}} \mathbf{r} \quad (4.31)$$

b) Magnetic field of moving particle:

The magnetic field of a moving particle is calculated using the Lorentz transformation as well. Presence of no magnetic field in the moving frame simplifies the transformation to:

$$\mathbf{B} = \langle -\gamma \frac{v}{c^2} E'_y, +\gamma \frac{v}{c^2} E'_x, 0 \rangle \circ \mathbf{e}, \quad (4.32)$$

where the electric field components magnitudes are described using Eqns (4.26) and (4.20). Applying the same set of the substitutions defined in (4.30) and performing the calculation for the observer in the rest frame O at $t = 0$, the magnetic induction can be defined:

$$\mathbf{B} = -\frac{\mu v e}{4\pi r^3} \cdot \frac{1 - \beta^2}{[1 - \beta^2 \sin^2 \zeta]^{\frac{3}{2}}} \cdot \left\{ \langle y, x, 0 \rangle \circ \mathbf{e} \right\} \quad (4.33)$$

This can be transformed into the cylindrical coordinates:

$$\mathbf{B} = \frac{\mu \sin \zeta}{4\pi r^2} \cdot \frac{v e (1 - \beta^2)}{[1 - \beta^2 \sin^2 \zeta]^{\frac{3}{2}}} \phi \quad (4.34)$$

c) Wall Image Current and its distribution function:

Assume an ideal vacuum chamber with zero longitudinal resistance. The particles moving inside the vacuum chamber are always accompanied by an electromagnetic field. It couples the beam and the vacuum chamber. This results in an attraction of a charge on the inner wall of the vacuum chamber. The attracted charge is of an opposite sign to the beam

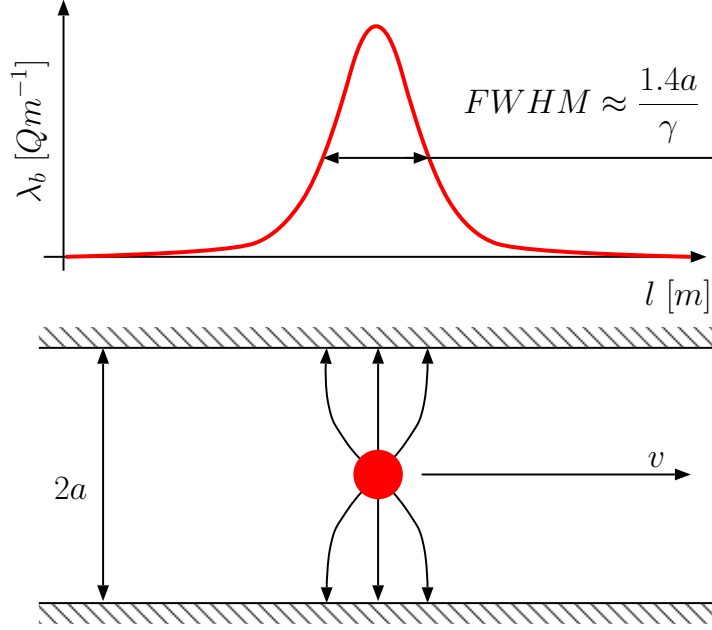


Figure 4.9: *Distribution of the mirror charge on the inner wall of the vacuum chamber*

charge. Its longitudinal distribution depends on the shape of the electric field. The electric field deflects as the particle's speed approaches the speed of light hence the longitudinal distribution of the mirror charge mimics the distribution of the beam current (Fig. 4.9). The Full Width at Half Maximum (FWHM) of the pulse created by a moving mirror charge can be expressed in terms of the relativistic factor γ as:

$$FWHM \approx \frac{1.4a}{\gamma}, \quad (4.35)$$

where a is the vacuum chamber radius.

As the particles travel, they are always accompanied by this mirror charge. The mirror charge is called a “wall image current” and it forms an inseparable counterpart to the beam current:

$$I_w = -I_b \quad (4.36)$$

4.1.3 Particle beams and noise sources

The theory described in the previous sections does not take into account other interactions affecting the beam. So far a perfect Gaussian shaped beam producing a limited number of spectral lines was assumed. However, the beam is exposed to various sources of noise, e.g. a power supply noise in the magnets, various RF noise sources, noise due to improper grounding or even mechanical vibrations. This results in a complex signal spectrum. An example to mention is the noise of the kicker magnet power supplies, affecting the trajectory of the beam, and leading to an emittance growth [30]. Noise signals can be

used as an additional treatment of the beam [25]. A controlled noise injection is used e.g. for tune measurements [42].

The noise sources in the intensity measurement chain, either affecting the beam or the processed signal, can be classified into two main categories: deterministic noises, and non-deterministic noises generated by an operation of an accelerator. The noises imposed on the beam have almost no effect on the intensity measurement. The measurement devices are beam position insensitive and the measured signal is integrated so the beam noise is largely suppressed. In contrary, noise sources affecting the processed signal influence the intensity measurements.

The deterministic noise sources can be mathematically evaluated. They include e.g. thermal and amplifier noise which can be treated by statistical methods, and any other source which can be modelled. A good example is a noise of an acquisition chain (ADC quantisation noise and impurities, integration errors etc.).

Although the white noise could be removed by statistic methods, in some cases this is not possible, e.g. when not enough measurements is available. In the case of transfer lines there is only one “shot” available. This is as well a case of the intensity measurement during an injection.

In the case of systematic errors there is no general rule how to treat them. Developer can face various sources of digital noise (e.g. clock coupling), impurities (ADC conversion), electromagnetic interference (EMI), clock jitter related noise, sample and hold circuitry etc. These effects are well described in the literature [84].

Second category of the noise sources includes all *noise sources generated by an operation of the accelerator*. One of the most noticeable is the Schottky noise [13]. Its source is a random variation of the position caused by motion of each individual particle around the mean orbit. The variation of the position manifests itself as a noise superimposed on the beam current. The measurement of this noise provides an additional information about the accelerator machine itself in form of a tune measurement [26]. Additional information about the Schottky spectra can be found in [48]. The Schottky noise is acquired by the intensity measurement devices however it is not a source of the measurement error providing that sufficiently wide-band measurement system is used.

Another noise source to be considered is the RF noise. This is a high frequency noise imposed on the beam by the RF acceleration process and the RF procedures. Any non-linearity in the accelerating RF voltage produces a variation of the synchrotron frequency [82]. This could cause e.g. “filamentation” in the phase space. In the case of the noise spectrum which is not extending into the RF frequencies the beam is affected either by the phase or amplitude noises, but the particles do not mix in the phase space [11]. Some studies of the RF noise and its influence to the lifetime were done in the SPS at CERN [5].

One of the noise sources affecting the intensity measurement is shown in Fig. 4.10. Figure depicts a single bunch beam measured by an intensity measurement device in

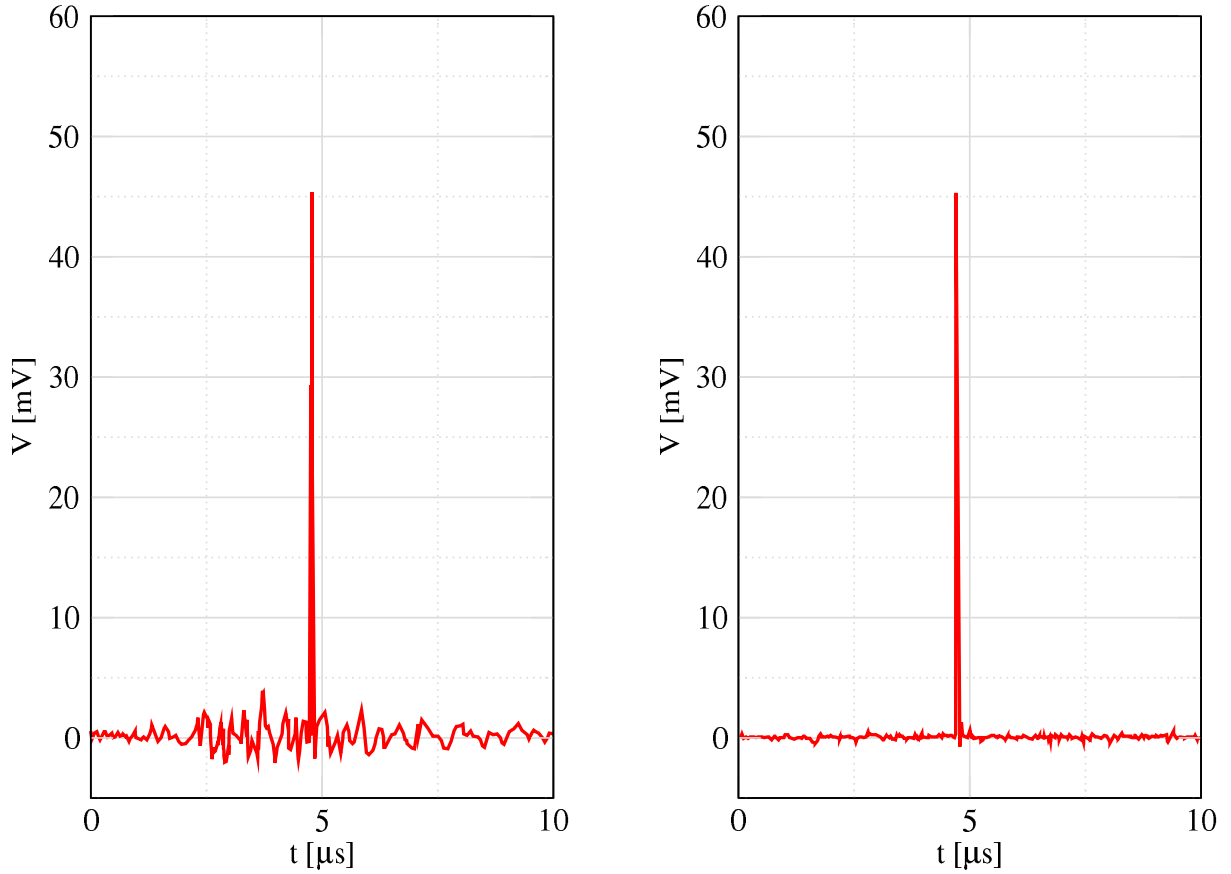


Figure 4.10: *Signal measured at the BCT in the SPS. The bandwidth of the transformer exceeds 700 MHz. Figure on the left shows a beam signal measured during an injection cycle. The figure on the right shows the same signal. This time the injection kicker is not active.*

the SPS. The BCT is situated in the SPS ring (see Fig. 4.32 in Sec. 4.3.3) about 60 m from the injection kicker. On the right the reference measurement is shown. This was taken during the acceleration when the injection kicker was not active. The left picture depicts a signal measured during an injection cycle when the particles enter the SPS. The bunch is surrounded by a noise raised by the injection kicker. The noise is transferred via the vacuum chamber and cannot be easily filtered away. Precautions have been taken to filter at least the LF noise travelling through the vacuum chamber. This was accomplished by putting capacitive insertions between the vacuum chamber flanges. Although this solution also prevents the DC current flow, it is not sufficient to suppress the RF noise. It is worth mentioning that due to the non-existence of capacitive coupled vacuum chamber connections in the LHC one would expect much higher noise levels to be measured. This may potentially degrade the resolution of the beam intensity measurement system.

Figure 4.10 also shows another potential problem - gating of the beam intensity integrator. As it is difficult to implement a digital acquisition and integration for short pulses, the measured signal is integrated by an analogue integrator. The integrator has a finite integration time. The signal having white-noise like properties, like the one depicted in Fig. 4.10 left, does not have zero mean value when integrated over the period of interest. This appears as a random offset fluctuation of the integrated signal.

4.2 Non-intercepting beam intensity measurements

The non-intercepting devices measure the charge by integration of the beam current or the wall image current. Corresponding signal is inductively or capacitively coupled to the measurement device.

Capacitive coupling is mostly used by electrostatic pick-ups. To obtain a measurable signal, the pick-ups use a conductive electrode inserted into the vacuum chamber, as shown in Fig. 4.11. The electrode is isolated and it is subjected to a charge deposit caused by moving particles. The charge difference between the electrode and the vacuum chamber creates a current flowing through a load resistor R .

The load resistor R and the capacitor C form a high pass RC filter. The value of capacitor is given by sum of a cylinder capacitance and a capacitance of connected cables. Therefore the cable length influences considerably the transfer impedance. For example 900 MHz pick-ups installed in the SPS have a transfer impedance of $13\ \Omega$. Connected cables limit their transfer impedance to $\approx 1.1\ \Omega$ and the bandwidth to 600 MHz [4].

Improvements in the magnetic measurements caused that the capacitive pick-ups are nowadays used only for complementary intensity measurements. However, they are still widely used as beam position detectors. Additional information can be found e.g. in [69] and [52].

Following sections give an overview of intensity measurement methods using inductive coupling. Three major methods are discussed, with the emphasis on the intensity measurement using BCTs.

4.2.1 Wall current monitors

The Wall Current Monitor (WCM) measures the intensity (e.g. [16]) and the longitudinal profile of the beam. The bandwidth of these devices often exceeds 5 GHz.

The WCM intercepts the wall image current. Only HF components of the image current provide an useful information. Therefore the WCM must act as a current divider

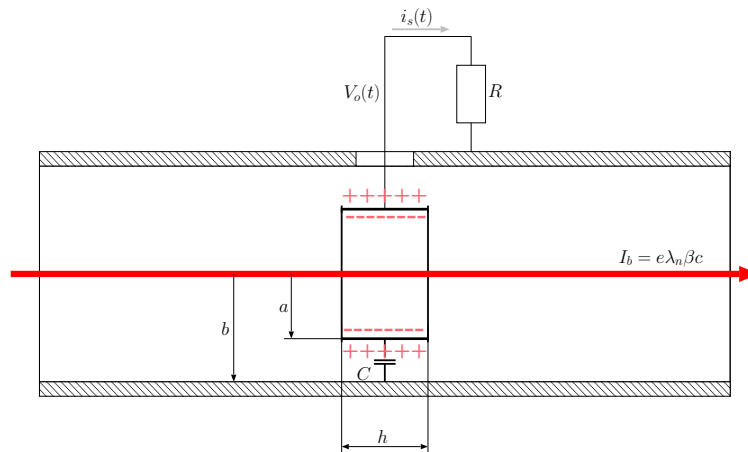


Figure 4.11: Principle of operation of capacitive pick-up

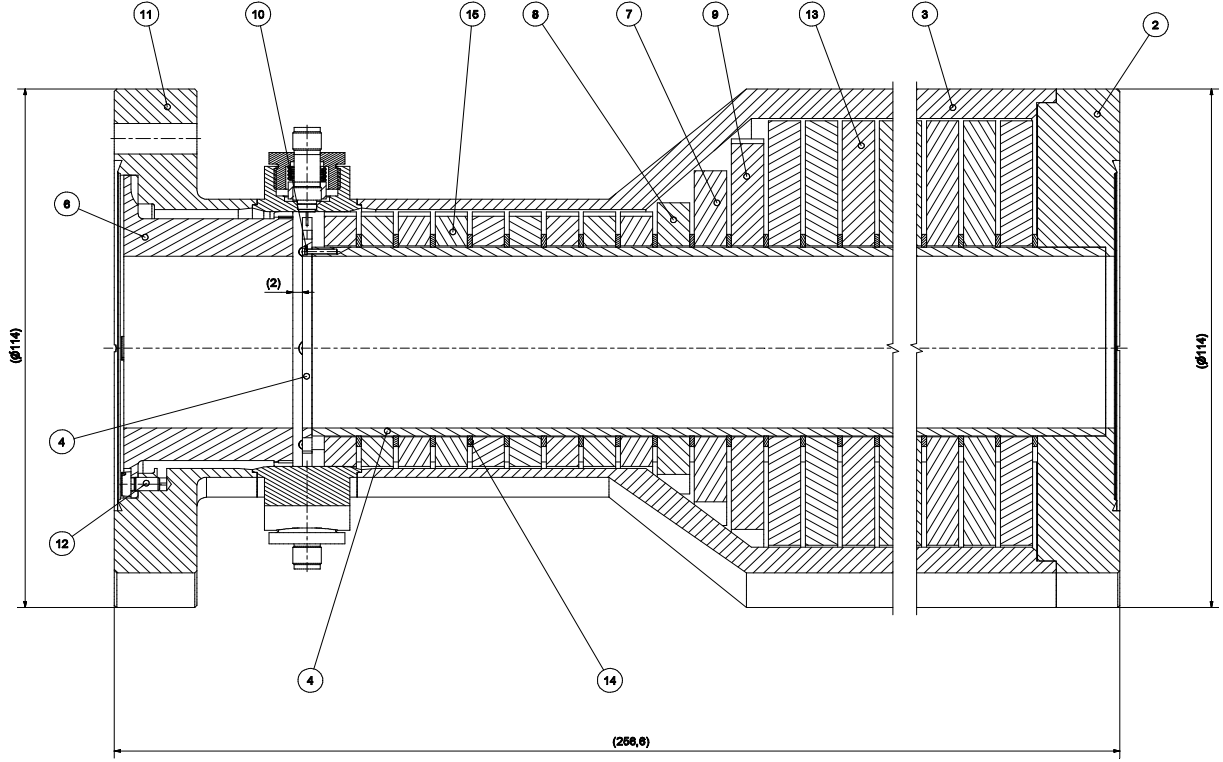


Figure 4.12: Cross-section of the WCM installed in the CTF3. The 2 mm air gap is depicted at the left side of the flange (4). The WCM is filled with ferrite rings to increase the inductance of the shielding box.

providing separate paths for HF and LF current components. The HF current passes through a load resistor connected in series with the vacuum chamber. To install the resistor the vacuum chamber must be split, and separated by an isolating gap. To avoid the LF current to pass through the resistor a low-impedance bypass must be installed over the gap.

An example of WCM is shown in Fig. 4.12. An air gap, 2 mm wide, separates both sides of the vacuum chamber. The load resistance is formed using 8 feedthroughs installed over the gap. The interior of the WCM is filled with a ferrite. It increases the inductance of the screening box hence it forces the HF current to pass through the feedthroughs. The inductance is affecting the LF cut-off of the device as well. Due to that, the WCMs have the LF cut-off in the range of ten to hundreds of kilohertz. The WCM shown in Fig. 4.12 provides a bandwidth of 250 kHz to 10 GHz.

For the intensity measurements a calibration of the WCM is necessary. A DC current transformer (DCCT) can be used as a calibration reference. Another possibility is to inject a known current through a calibration turn [33]. Should the WCM be used to measure a circulating beam, a signal baseline restoration must be done. The intensity measurement using WCM calibrated by a DCCT is used e.g. in Relativistic Heavy Ion Collider (RHIC) at Brookhaven National Laboratory (BNL) [34]. At CERN a high bandwidth WCM has been developed e.g. to observe the longitudinal profile of the CTF3 beam [54].

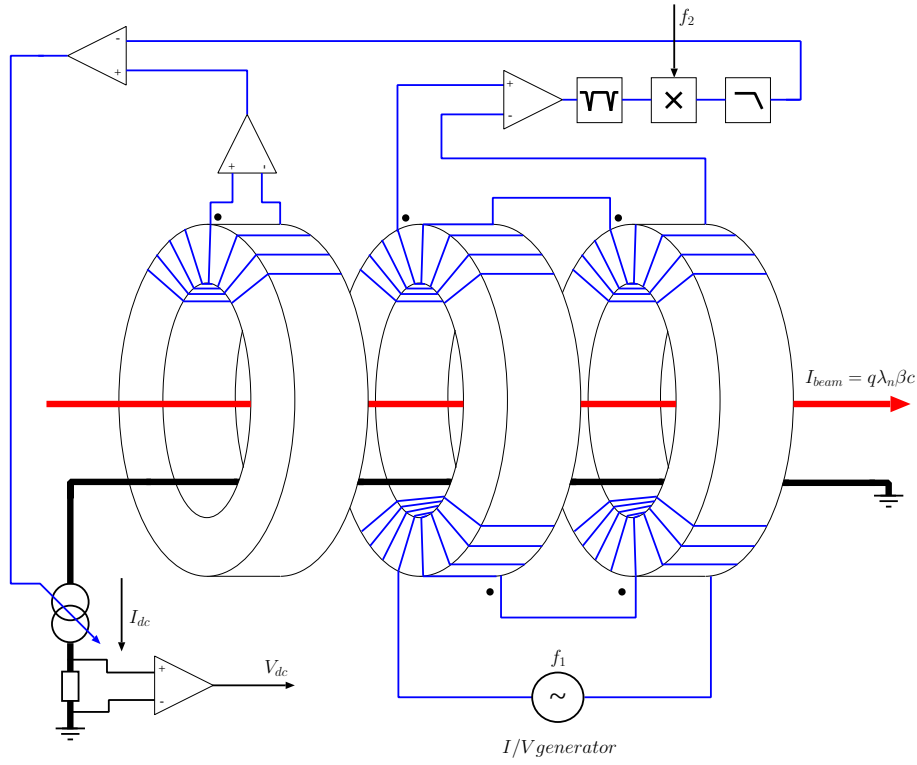


Figure 4.13: A DCCT - principle of operation

4.2.2 DC current transformers

The DCCT measures an intensity of a circulating beam. The principle of operation is shown in Fig. 4.13. The measurement device uses a “magnetic modulator” to obtain a signal proportional to the DC current of the beam. Magnetic modulation is done by superposition of the magnetic flux generated by a signal of a generator with the beam signal. The modulation flux develops from injection of a current into windings of two magnetic cores. The windings configuration is such, that the current generates in both cores a flux of opposite direction.

A common sensing winding provides an output signal. Assuming perfectly matched cores the second harmonic in the output signal cancels due to symmetric excitation by modulation current (Fig. 4.14 middle). A beam passing through the aperture of the cores creates an additional flux. This unbalances the excitation hence even harmonics appear in the output signal, particularly the second harmonic. The amplitudes of even harmonics are proportional to the beam current.

The second harmonics of the output signal mixed to the base-band provides a driving signal for a current generator. The generator acts to suppress the flux generated by the beam current. The amplitude of the needed current corresponds to the beam current [3].

Additional AC transformer extends the bandwidth of the demodulated signal. This results in faster response of the measuring device to current changes.

The reaction time of the feedback loop is usually some tens of milliseconds. Due to that, the DCCTs cannot measure on transfer lines. Installed in the rings, DCCTs provide

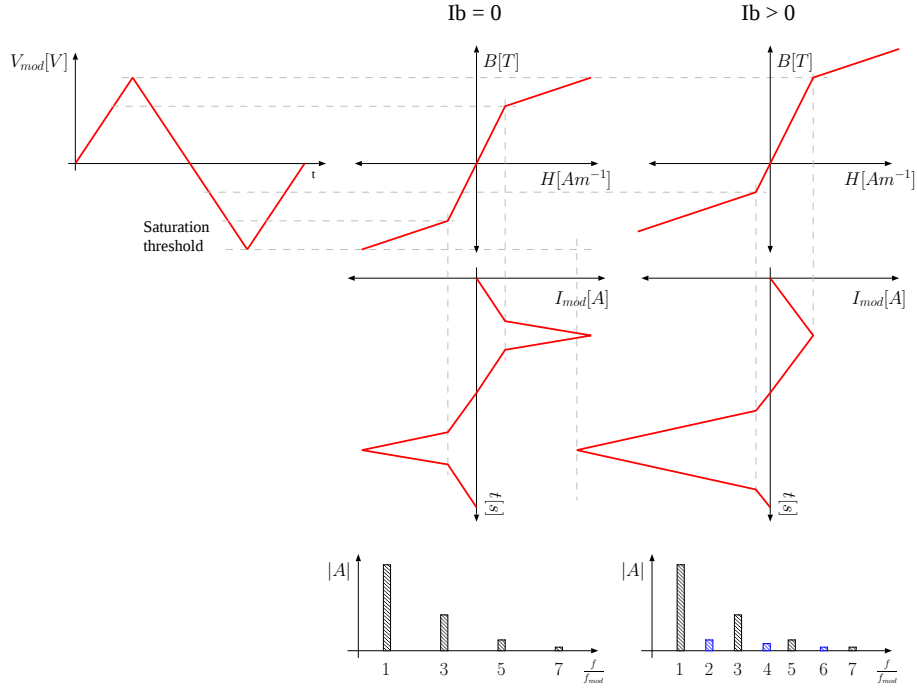


Figure 4.14: General principle of 2nd harmonic production in magnetic modulator

measurements of exceptional sensitivity ($1 \mu\text{A}$) and high dynamic range ($>100 \text{ dB}$). This is advantageous for beam lifetime measurements.

At CERN, the DCCTs are Currently used in CERN Proton Synchrotron Booster (PSB), Proton Synchrotron (PS), Anti-proton Decelerator (AD), Super Proton Synchrotron (SPS) and recently built Low Energy Ion Ring (LEIR) machines. A DCCT with a sub-micro ampere resolution was constructed to measure a circulating beam in the LHC rings. DCCTs are available also from industry [74].

A summary of DCCT technology can be found e.g. in [61], performance limitations are discussed in [63] and [62].

4.2.3 The fast beam current transformers

The FBCTs provide complementary measurements to the DCCTs. The FBCTs are typically used in the transfer lines, to measure injection and extraction efficiency, or provide measurements of low energy and intensity beams. Improvements of readout and integration techniques allowed them to enter the domain of circulating beam measurements. This extends their usability to e.g. evaluation of fast beam losses which is vital for accelerator protection system.

The principle of measurement is shown in Fig. 4.15. A toroidal transformer is used to measure the current. It couples inductively to the beam current. To make the transformer to “see” the beam current the vacuum chamber must be split into two parts and separated by an isolator. Contrary to the WCMs, the gap is wider, and formed by a ceramic ring brazed to both parts of the vacuum chamber. In order to provide a conducting path for the wall image current the toroid is bypassed from outside by a low

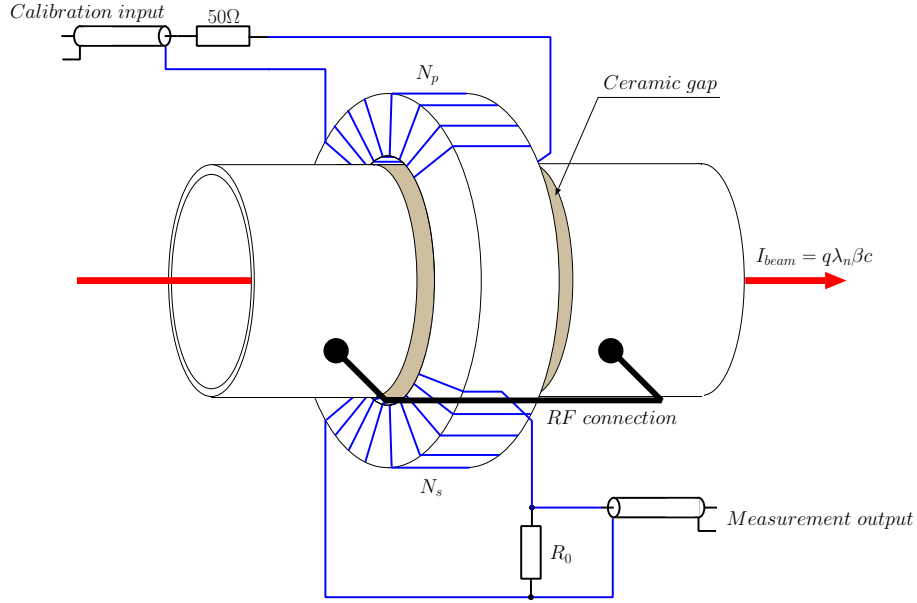


Figure 4.15: *Principle of fast current transformer*

impedance RF connection. Such configuration permits to integrate a current induced in the winding and evaluate the number of charges.

The quality of obtained signal depends on bandwidth of the transformer. Larger bandwidth permits to copy more accurately the beam current. The LF cut-off of the transformer is given by an inductance of the winding. The HF cut-off is limited by a capacitive coupling between the turns, stray and eddy currents, the energy loss in the core material and the loss of permeability with the frequency.

Classical transformer theory defines an inductance of the transformer from an equivalence of voltages created by a magnetic flux and a time-varying voltage across an inductor [40, 81]:

$$-N_k \frac{d\Phi_k}{dt} = L_k \frac{dI_k}{dt} \quad (4.37)$$

where Φ_k is the total flux in the core, N_k is number of turns of winding k carrying a current I_k . L_k is a self-inductance of the winding. The magnetic flux can be expressed as a product of a magnetic induction passing an area element:

$$\Phi_k = \int_S \mathbf{B}_k \cdot d\mathbf{S} \quad (4.38)$$

where $\mathbf{B}_k$ is vector of magnetic induction. Its magnitude at a distance r from an indefinitely long straight wire carrying current I is expressed as:

$$B(r) = \frac{\mu I}{2\pi r} \quad (4.39)$$

The inductance of the winding is obtained by putting together Eqns. (4.37), (4.38) and (4.39). For a toroidal core it is defined as follows:

$$L_k = A_l N_k^2 \quad (4.40)$$

$$A_l = \frac{\mu h}{2\pi} \ln \frac{b}{a} \quad (4.41)$$

where the parameters a , b and h specify dimensions of the core: a the internal diameter, b the external diameter and h is the width of the core. Constant A_l is often referred by manufacturers of the toroids.

The primary side of the transformer is represented by the beam which acts as a one-turn winding. The FBCTs often use two windings. The first winding, referred as a calibration turn, is used to inject a known current into the transformer. The second winding is connected to an acquisition system and provides the measurement signal. As the calibration of the device and the intensity measurement are mutually exclusive tasks, the total magnetic flux can be expressed by assuming one primary and one secondary winding:

$$\Phi_T = \frac{L_p I_p}{N_p} + \frac{L_s I_s}{N_s} \quad (4.42)$$

$$\frac{d\Phi_T}{dt} = -\frac{V_p}{N_p} = -\frac{V_s}{N_s} \quad (4.43)$$

Primary winding has by definition only one turn and carries the beam current I_b :

$$N_p = 1 \quad (4.44)$$

$$I_p = I_b \quad (4.45)$$

Switching into Laplace domain a current induced in the secondary winding can be described:

$$\Phi_T = A_l I_b + \frac{L_s I_s}{N_s} \quad (4.46)$$

$$s\Phi_T = -I_s \frac{R_s}{N_s} \quad (4.47)$$

Equations (4.46) and (4.47) define the current at measurement winding:

$$I_s = \frac{s\tau}{s\tau + 1} \cdot \frac{I_b}{N_s}, \quad (4.48)$$

where the time constant $\tau = L_s/R_s$ defines the LF cut-off.

The first term of Eq. (4.48) describes a high pass filter. The FBCTs thus do not transfer the DC component of the beam. Missing DC component in the output signal causes its “droop”. For transfer line intensity measurements the droop represents a minor problem which can be dealt with by an appropriate selection of the time constant. The FBCTs installed in the rings must inevitably provide a DC baseline restoration system to assure correct functionality.

For frequencies well above the LF cut-off the current can be approximated by:

$$I_s \approx \frac{I_b}{N_s} \quad (4.49)$$

The signal power of the measurement signal can be calculated as:

$$P_s = R_s I_s^2 = \left(\frac{I_b}{N_s} \right)^2 R_s = \left(\frac{I_b}{N_s} \right)^2 \frac{A_l N_s^2}{\tau} = \frac{I_b^2 A_l}{\tau} \quad (4.50)$$

Proportional relation between the P_s and the constant A_l determines the signal to noise ratio. To achieve the best results toroids made of high permeability materials, small in diameter and having large cross-section are recommended.

Providing there is no noise source induced in the winding, the noise floor of the measurement signal is given by a thermal noise of a load resistor:

$$V_{eff} = \sqrt{4k_b T R \Delta f}, \quad (4.51)$$

where T is a temperature of resistor, k_b the Boltzmann constant and Δf the bandwidth of the transformer. The thermal noise power corresponds to:

$$P_{eff} = 4k_b T \Delta f \quad (4.52)$$

a) The transformer bandwidth influence to the intensity measurements

The required bandwidth of the transformer is given by properties of the beam and a type of the measurement provided. Proper choice of LF and HF cut-offs is a matter of compromise, especially when considering a single transformer covering large bandwidth as usually required for large circumference machines.

The HF cut-off: The minimum HF cut-off of the device is given by the beam's bunch width and bunch spacing. If a bunch by bunch measurement is required, the HF cut-off is proportional to the spectra width as discussed in Sec. 4.1.1 b). Maximum achievable HF cut-off is limited by losses in the magnetic materials and permeability loss. Modern materials as amorphous cobalt-based alloys permit to achieve frequencies up to 2 GHz [74]. If even higher frequencies are required a WCM must be used.

The LF cut-off: The intensity measurement measurement error is affected by the droop of the measurement signal. The amount of droop depends on the LF cut-off of the measurement device defined by the inductance Eq. (4.40). The higher is the number of turns, the higher inductance of the winding is obtained. The inductance influences the time constant as defined in Eq. (4.48).

The droop causes a displacement of the base line of the signal. The effect can be well controlled in the transfer lines, where the repetition rate of the measurement is small compared to the time constant. In this case the droop compromises only an accuracy of the measurement. This is a case of e.g. measurements in the transfer lines at CERN PS complex. The period of the super-cycle (Sec. 4.3.1) is 1.2 seconds and the particles are extracted from the PSB within 4 to 15 μ s. Hence in the worst case one

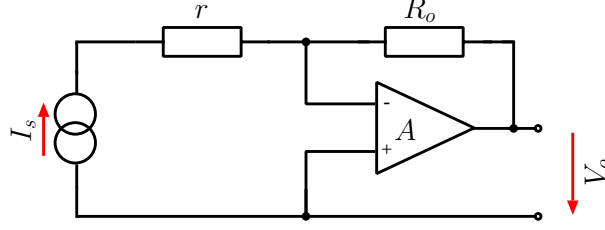


Figure 4.16: Transimpedance amplifier decreases the time constant by the factor of amplifier gain

integration lasting $15 \mu\text{s}$ from 1.2 s is done. Using measurement transformers having droop better than $0.1 \text{ } \%/ \mu\text{s}$ assures the measurement error better than $2 \text{ } \%$ for all types of the beams extracted from the PSB. Additional DC restoration techniques permit to improve the measurement accuracy. Typically, the LF cut-off of the transformers for the transfer lines varies between 100 Hz and 1 kHz .

For the repetition rates comparable to the time constant, the base line of the measured signal does not fully recover. In the circular machines the repetition rate corresponds to the bunch spacing. Bunch spacing is usually much shorter than the time constant and the base line sets such to get a zero mean value of the signal. Theoretically, integral of this signal results in zero as well. In practise, small differences between measured bunches give an “error” signal, which is proportional to the difference of the beam charge to the mean charge within a bunch slot. DC signal restoration techniques are mandatory in order to perform the measurement. Selection of proper method depends much on characteristics of the beam. Circulating beam measurement using FBCTs is performed e.g. in the SPS. The accelerated LHC type beams have a bunch separation of 25 to 100 ns .

b) The LF cut-off improvement techniques

Transfer line intensity measurements: The simplest method to lower the LF cut-off is to increase number of turns of the measurement winding. This can be applied for beams generating sufficient current. Higher number of turns reduces the amplitude of the measurement signal and lessens the HF cut-off due to higher stray capacitance of the winding. Therefore the signal rise time is affected. Both problems can be partially suppressed using pre-amplifiers and choosing different type of winding technique.

Another solution is to use an operational amplifier to lessen the load impedance. This technique is called “an active transformer”. To minimise the droop the transformer is loaded by a transimpedance amplifier instead of a resistive load R_s . Situation is depicted in Fig. 4.16. The resistor r represents resistive impedance of the cable and R_o sets the gain. Assuming non-ideal operational amplifier (OA) the equivalent input impedance R_i (from Eq. (4.48)) is given by the input impedance of the amplifier circuit:

$$R_i = r + \frac{R_o}{A + 1} \quad (4.53)$$

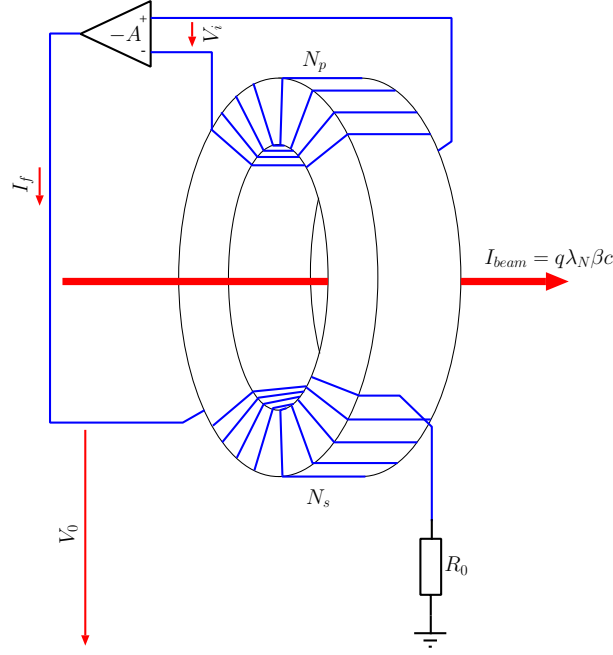


Figure 4.17: *AC transformer with Hereward feedback*

Hence the time constant of the circuit becomes:

$$\tau_1 = \frac{L_s}{r + \frac{R_0}{A+1}} \quad (4.54)$$

As gain $A \gg 1$, the time constant of the circuit is mainly affected by the impedance of the connecting cable:

$$\tau_{1,A \rightarrow \infty} = \lim_{A \rightarrow \infty} \tau_1 = \frac{L_s}{r} \quad (4.55)$$

The current I_s is amplified by a gain of transimpedance amplifier. Assuming $r \ll R_0$ the transimpedance gain is:

$$\frac{V_o}{I_s} = -R_0 \frac{A}{A+1} \approx -R_0 \quad (4.56)$$

Low bandwidth OAs are recommended due to the amplifier noise. This worsens the HF response of the measured signal, particularly the rise time. Attention must be paid to the dynamic range of the input signal as well. The electronics must be installed in the vicinity of the toroid. This is necessary as longer cables cause ringing due to an impedance mismatch. Some studies were presented for CERN LEIR FBCTs concerning a possible cure of the effect of long cables [60], but an acceptable solution has not been found.

Another type of LF cut-off compensation is a Hereward feedback. Compensation is made using an additional winding of the transformer as depicted in Fig. 4.17.

Assume a high impedance at the input of the amplifier. The voltage at the input of the amplifier is given by a sum of voltage induced by the beam and voltage induced due to a current change in the second winding having N_s turns:

$$V_i = N_p A_l s (I_b + N_s I_f) \quad (4.57)$$

Following the theory presented e.g. in [14] an approximate sensitivity can be expressed:

$$S = -\frac{R_0}{N_s} \cdot \frac{1}{1 + (1/s\tau_h)}, \quad (4.58)$$

where the time constant τ_h is defined as:

$$\tau_h = A_l N_p N_s \frac{A}{R_0} \quad (4.59)$$

Compared to already discussed techniques, the Hereward transformer offers higher degree of freedom. By changing the load impedance R_0 and the number of turns N_s the overall sensitivity can be changed without affecting the time constant. Parameters N_p , A_l and gain A set the time constant.

This technique can be used only when the OA can handle the output current I_f and output voltage V_0 :

$$I_f = \frac{I_b}{N_s} \quad (4.60)$$

$$V_0 = \left(\frac{R_0}{N_s} + sA_l N_p \right) I_b \quad (4.61)$$

In large machines the peak current I_b can reach several amperes hence the signal must be correspondingly attenuated. This results in poor rise time. Also, lack of a DC feedback path causes a base line shift when no beam is present. However, this arrangement allows to achieve the time constant of an hour or more.

Intensity measurements in circular accelerators: Restoration methods used for the transfer lines are not suitable for circular accelerators because in storage rings the beam can circulate for several hours. Hence even the largest achievable time constant would not be sufficient to keep the droop of the measured signal at acceptable level. For monitors measuring circulating beams the techniques based on the properties of the beam signal must be used. One such technique, based on the principle of a sample and hold circuit, is depicted in Fig. 4.18.

The input, labeled '*on/off*', controls a buffer generating high voltage signals to drive the sample and hold circuit. The sample and hold is implemented using Schottky diodes in a bridge connection. When a positive voltage is applied between points 4-3, a virtual connection is created between points 1-2. This charges the output capacitor to a voltage given by a difference between *INPUT 2* and point 1. Ideally, the point 1 should be connected to ground. In reality it is fixed to a DC voltage, which permits to compensate for various offsets. Reverse voltage applied between the points 4-3 disconnects the DC voltage present at point 1 from the capacitor. As the capacitor is charged, the voltage appearing at its terminals is added to the voltage of *INPUT 2* when observed at *OUTPUT 2*. During the period of hold, the capacitor is discharged through a load impedance connected to the *OUTPUT 2*. Hence only high-impedance loads must be used.

Yet another type of the DC restoration was used for the SPS ring intensity measurements. The beam intensity was estimated by an integration of Gaussian-like functions fitted to the measured signal [45]. This removed the dependence of the measurement on the position of the base line in the measured signal. The system required a precise clocking to obtain reasonable results. Due to that a dedicated 200 MHz beam synchronous RF timing was used. Although the timing provided excellent clock source, another problems were discovered. They were mostly related to varying phase shift of the beam and the RF clock source due to fixed cable lengths and a variable time-of-flight of the accelerated particles. The method was sensitive to the bunch length and was not able to measure ghost bunches.

Nowadays, the analogue techniques of the DC restorers are replaced by a digital ones. This is especially case of long bunches measurements as the properties of the measured signal permit an easy analogue to digital conversion. Using fast samplers, having a sampling rate of several GSPS, the data are treated mathematically. Base line is restored by summing the acquired data with a linear fit of the data measured at the positions where is no beam. This technique replaced the Gaussian fit method used in the SPS intensity measurement.

Sampling measurement techniques provide a continuous flow of the values measured. At high sampling rates, the data however cannot be processed on-fly. Hence a storage must be available. Product of a storage size and a sampling period defines the length of a single-shot sampling window. Additional limitations are given by a sampling rate and a dynamic range of the sampler. The measured signal must be oversampled. To have a good resolution at least 10 measurements per bunch width must be obtained. This does not represent a problem for turn based intensity measurements.

c) Calibration techniques

Theoretical background: For proper operation the intensity measurement system must be calibrated. Calibration is a process of matching a reference signal measurement to the reference value. If the reference value is not known, it must be measured. Reference signal measurement must be done with better precision than offers the measurement system being calibrated. Properly calibrated system provides a compensation for systematic errors, e.g. gain changes, errors caused by fabrication processes and tolerances of components.

Basic type of calibration is a gain correction using multiplicative constant. *For each type of measurement* a calibration constant is calculated:

$$c_k = \frac{S_{theoretical,k}}{S_{calib,k}} \quad (4.62)$$

$$S_{corrected,k} = S_{measured,k} \cdot c_k \quad (4.63)$$

where S represents a variable describing signal. It can be voltage, current, digital value or even mixture of all those. In the case of intensity measurement $S_{theoretical,k}$ and $S_{corrected,k}$

PSB	O_k
User Name	[Number of charges $\times 10^{10}$]
ISOGPS	106
ISOHRS	38
MDPRO	56
nTOF	-28
EASTB	3
EASTC	8

Table 4.2: Calibration offsets O_k for an intensity measurement at the PSB extraction ($FS=4 \times 10^{12}$ charges). Average of 100 measurements was done. Approximately 96×10^{10} charges of front-end electronics offset was measured ($T_{int} = 4 \mu s$, $V_{in,ofs} \approx 13 mV$).

are given in number of charges whereas $S_{measured,k}$ and $S_{calib,k}$ in digital form (ADC bins). This is to convert the sampled signal directly into the number of charges. Index k denotes the type of the measurement. At CERN, this could represent e.g. different users in the super-cycle.

Second frequently used calibration is offset suppression. *Constant offsets*, e.g. ADC offset error, could be compensated by subtracting a value from the measured signal. *An analogue front-end offsets* originate in amplification and integration of the signal measured. Preferably, they should be suppressed directly at the source as any offset entering the integrator will be multiplied by its gain. In worst case such an offset could saturate the integrator. Due to finite-time integration the *offsets* can *originate in a noise* as well. Noise signals have an origin in the electronics, or they are intercepted by the monitor. This is typically case of currents induced on the vacuum chamber (pulsing kicker magnets) or a noise coupled to the cables. These noise sources are not systematic, are different for each type of measurement, and difficult to remove before the integration. They also vary with machine settings.

The offset correction enters Eq. (4.63) as follows:

$$S_{corrected,k} = S_{measured,k} \cdot c_k - O_{acq} \cdot t - O_k, \quad (4.64)$$

where $O_{acq} \cdot t$ is a front-end offset recalculated to the output of the integrator, and O_k is an offset due to noise. Should the time of integration be constant, the terms $O_{acq} \cdot t$ and O_k merge. Table 4.2 shows typical offset values measured on the beam extracted from the PSB for different users. The calibration was performed in a “noiseless” environment (the PSB was in the operation shutdown). The values shown in the table were taken later with machine fully operational, however with no beam being accelerated. Every user exhibits a different offset value. This is caused by different settings applied while accelerating the beams. Changes in the settings result in changes in the offsets, thus a periodical calibration is necessary.

The calibration factors c_k , O_{acq} and O_k must be found by measuring the signals

using multiple calibration standards. If the integration time is constant, two measurements are necessary to perform the calibration. In case of variable time of integration, at least three measurements are necessary. This is to separate the O_{acq} and O_k components of the offset error.

If needed, additional data treatment can be applied. Most frequent is a linearisation of the acquired data. Implementation of the algorithm depends on available hardware/software resources. For example a look-up table (LUT) is a good solution if enough memory is available. The LUTs provide excellent conversion speed and permit to create complex correction curves. Another possibility is to implement the algorithm to a field programmable gate array (FPGA), or CPU. Complex corrections using multiple input variables or even feedback can be made at almost no expenses in terms of memory. These algorithms are however slower and have higher latency.

Due to required measurement dynamic range the systems often work with multiple input ranges. In each range a measurement is performed individually and the resulting intensity is calculated using the best suitable dynamic range. The dynamic ranges overlap, hence a linking of the extremities of each range must be assured to provide a consistent measurement in overlapping regions. This process is called a cross-calibration.

Calibration standards: The intensity measurement calibration methods are based on injecting a defined number of charges into a calibration turn of the toroid transformer. The charge can be generated by different means, however the most frequently used are two methods: a current pulse generator, or a capacitor acting as a charge storage [43]. The example implementation of a calibrator using the capacitor storage method is shown in Fig. 4.20.

The calibrator uses a capacitor of a known value (connected to $J6$), which is charged to a known voltage ($J1$). A fully charged capacitor contains a charge corresponding to a product of its capacitance and the charging voltage. Once charged, the capacitor is discharged into the calibration turn of the transformer creating a response in the measurement winding, as seen in Fig. 4.21. The number of “calibration” charges contained in a single pulse expressed in convenient units is:

$$N_p[\times 10^{10}] = \frac{V[V] \cdot C[pF]}{1602} \quad (4.65)$$

The capacitor discharge time is limited by the length of the integrator’s gate in the measurement path, what limits the maximum value of the capacitor. In order to store amount of charge needed to simulate correspondingly the type of a beam, the charging voltage must be increased. With high intensity beams, this results in unrealistic values of charging voltage. E.g. given a fixed integration window of $4 \mu s$ as required in the PSB, the capacitor of maximum $10 nF$ could be used. In order to calibrate the FS response (4×10^{13} ch.) a voltage source exceeding $600 V$ is needed. Should the FS bunch by bunch PSB measurement be calibrated, the integration time decreases to some $600 ns$. Despite of mentioned

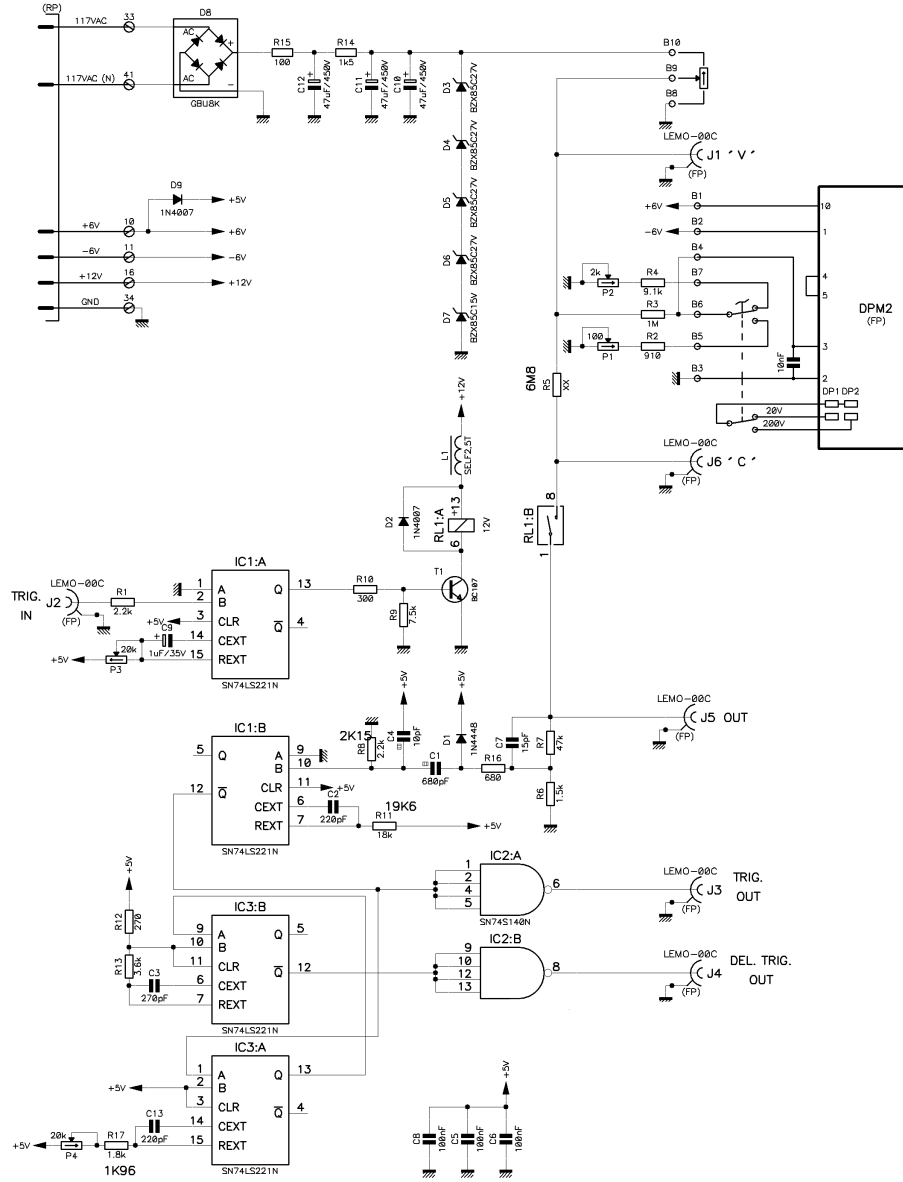


Figure 4.20: *The FBCTs calibrator used in the PS complex*

limitations, the implemented calibrator (Fig. 4.20) provides an absolute accuracy better than 5 %.

Figure 4.22 shows a schematic diagram of a calibrator used in the SPS intensity measurement. The calibrator generates $5 \mu\text{s}$ long 128 mA current pulses. In order to achieve an acceptable signal rise time, the current generator provides the current continuously into a dummy load. When calibration is requested, the current is diverted by a fast switch into the calibration circuit. Typical rise time achieved is 150 ns. In order to suppress the temperature dependence, the calibrator is enclosed in a box. The power dissipated in the dummy load is transformed to a heat which increases the interior temperature, thus alleviates the effects of external temperature changes.

The FBCTs can also be calibrated indirectly. For this, the DCCTs or Faraday cups could be used. In both cases, the calibration is possible only with both devices installed in the vicinity of each other. Otherwise the calibration would become unreliable due to

d) Integration techniques

The fast intensity measurements involve usage of integrators of different speeds. Simplest form of integrator is an operational amplifier with a capacitor in the feedback. This circuit was used e.g. for measurements in the LEP [17], or is still operational in the PS complex [46]. The integrator's reset is realised using resistance connected in parallel to the integrating capacitor. It is advantageous as this solution does not introduce errors caused by a charge injection due to used switching transistors. On the other hand, a much longer time is needed to perform the reset than to integrate, what limits the measurement repetition rate. Dynamic range of the integration lessens with the speed of integrator as HF OAs generally support only lower amplitude signals.

For bunch trains with the bunch separation less than 50 ns this simple technique is not suitable. The integrator must reset quickly, what results in additional problems. Just to mention few: charge injection due to MOSFET switching, design of fast drivers for MOSFET transistors, increased input offset of the HF OAs, increased noise etc. Dedicated integrated circuits partially simplify the design what was the case in the SPS and LEP where a BCT acquisition front-end was made using a combination of the ACF2101 integrator and an AD1376 16 bit ADC [90]. The front-end was implemented in the VME crate.

In 2005 a new acquisition card for the PS complex FBCTs was developed. The measurement method uses two 200 MHz ADCs working in a time multiplex to achieve a virtual 400 MHz sampling rate [70]. This configuration permits to have a temporal resolution to perform a numerical integration in the hardware. The signal is integrated digitally in an FPGA what permits to read the intensity values directly by the front-end computer. In order to provide a complete measurement front-end, both capacitor and current generator calibrators are implemented on board. The capacitor calibrator permits to charge a 10 nF capacitor up to 300 V. Current generator generates pulses 200 ns to 100 μ s long with maximum peak value 4 A (max. duration 5 μ s).

Due to high repetition rate of the measurement which requires to reset the integrator within few nanoseconds it is not possible to use these integration methods in the LHC. The reset acts as a dead time, and the integrator does not register any input signals. This represents a minor issue when the accelerator is correctly set-up, however in the case of the machine start-up a presence of numerous anomalies must be expected (e.g. ghost bunches, debunched beam, improper timing setup).

Various resetting techniques have been evaluated. In the LHCb an integrator processing a signal from a photo-multiplier is required. It was proposed to reset an integrator using transmission lines, as shown in Fig. 4.23 [37]. Signal propagates from the negative to the positive input of a buffer through a transmission line. The length of the line is such that it permits to integrate for 25 ns a signal appearing at the negative input of the OA. When the delayed signal arrives at the positive input, the output of the integrating OA must be acquired as the signal at positive terminal starts to discharge the capacitor. After another 25 ns the integrator output approaches back to zero (Fig. 4.24). The transmission

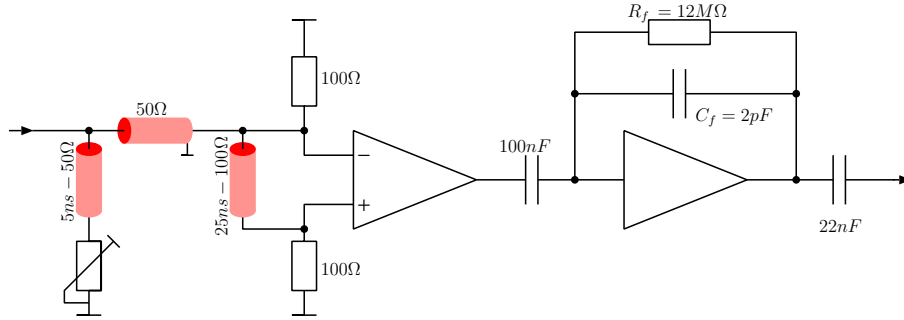


Figure 4.23: *Integrator electronics for LHCb calorimeter.*

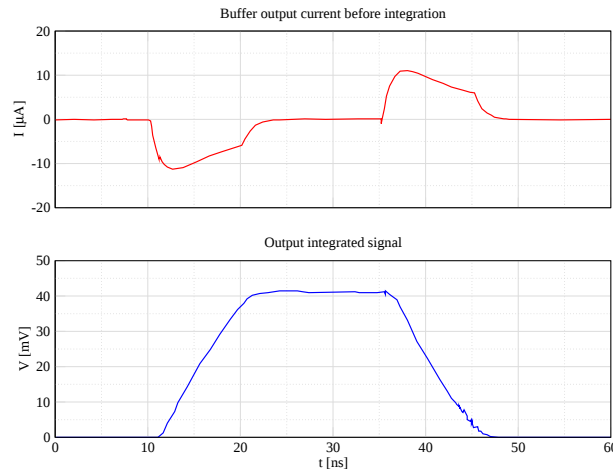


Figure 4.24: *LHCb calorimeter integrator signals [37]*

line behaves as an integrator reset. The reset phase requires the input signal to be zero. Hence a 5 ns 50 Ω transmission line clips the input signal after 10 ns. This is done by summing the source and reflected signal. The integrator is a BiCMOS 0.8 μm IC (known as the LHCb chip) developed in collaboration with university at Clermond-Ferrand, France.

The integrator repetition rate is 50 ns what was not sufficient for the LHC operation, hence an improved version of the integrator chip was developed. The new version (LHCb2002) implements two integrators working in duplex mode. This allows continuous 25 ns integration using one integrator while the other one performs reset. The reset is implemented on-chip. The buffer and the delay lines are of no use anymore.

As the LHCb2002 was suitable also for the SPS intensity measurements, the FBCT system was redesigned to implement integration using this chip (IBMS) [45].

4.3 Fast beam intensity measurements at CERN

The CERN accelerator complex is a chain of accelerators connected via transfer lines as shown in Fig. 4.25. The particles are produced in two linear accelerators - protons in the LINAC_{II}, and lead ions in the LINAC_{III}.

Particles produced by the LINACs are injected into the Proton Synchrotron Booster (PSB). The PSB accelerates from an initial energy of 50 MeV (or 4.2 MeV/nucleon for $^{208}\text{Pb}^{82+}$ ions) to 1.4 GeV using four acceleration rings. Accelerated particles can be extracted to three different areas: the PSB dump line (not shown), the ISOLDE experimental area, and the PS accelerator complex which provides further acceleration.

The PS has a circumference of 628.3 m and accelerates all stable and electrically charged particles (electrons, protons), anti-particles (positrons, anti-protons) and heavy ions (oxygen, sulphur, lead) up to 26 GeV/c. Each acceleration cycle takes 3.6 seconds. The PS complex also works as an accumulation ring. Successive injections fill the PS with equidistantly spaced bunches. As the circumference of the PS is exactly four times the circumference of the PSB, particles extracted from the four PSB rings fill the PS entirely. The particles are used by several experimental areas (e.g. East Area (EASTB, EASTC) and

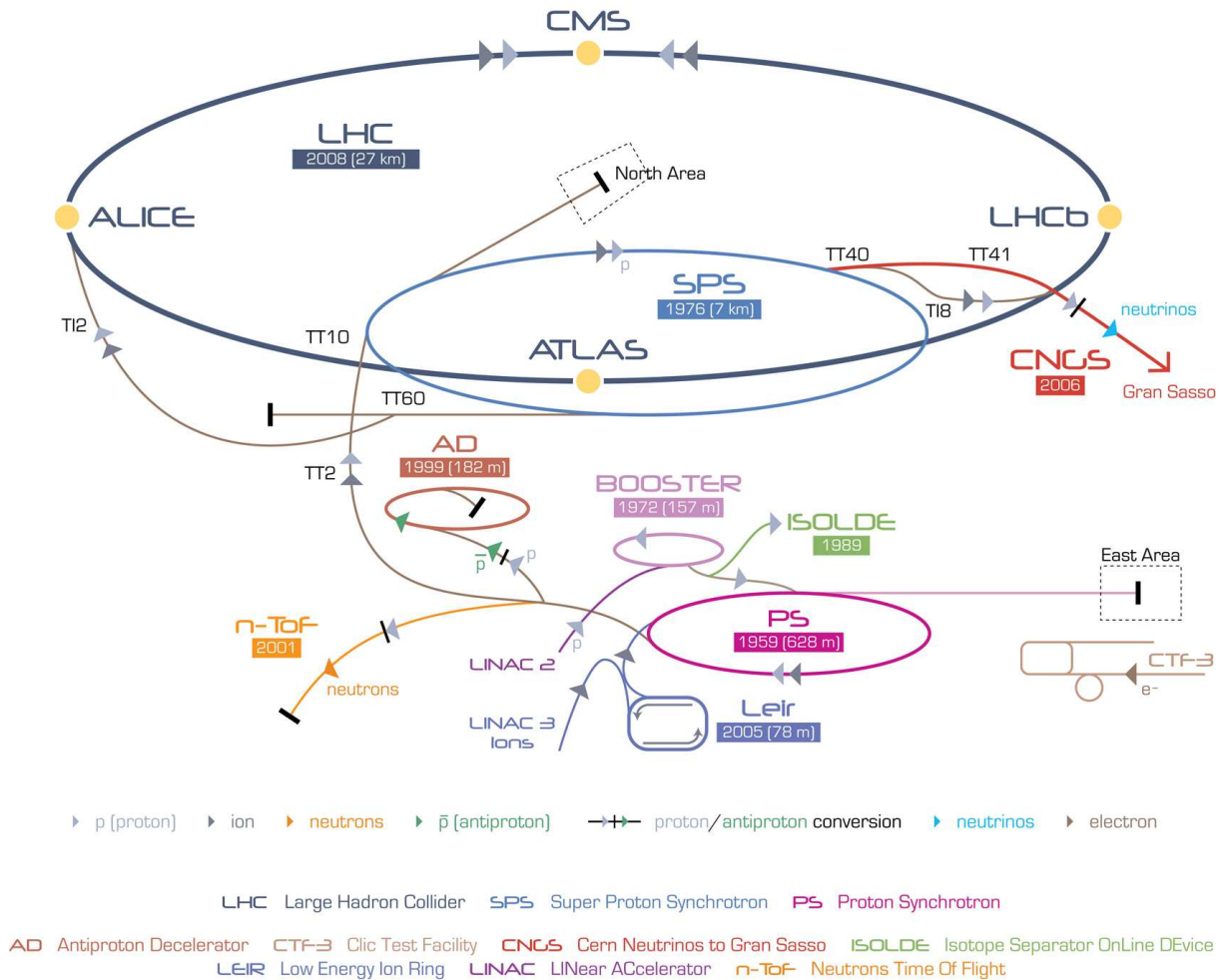


Figure 4.25: CERN Accelerator Complex (courtesy C. Lefevre, CERN)

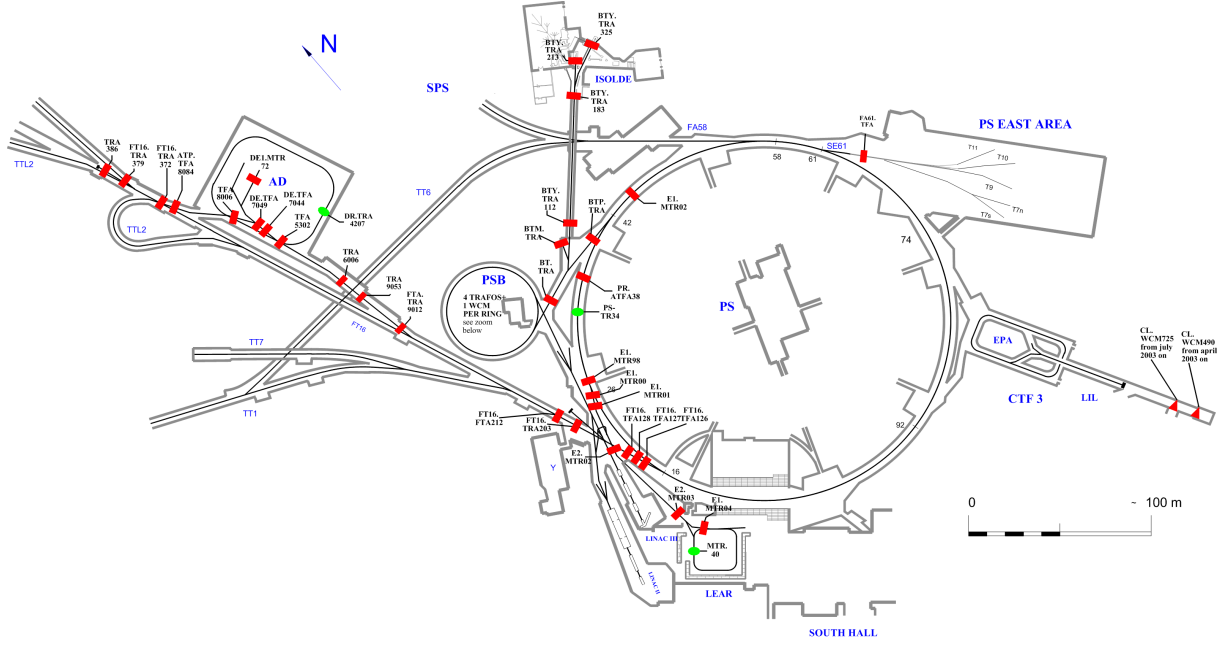


Figure 4.26: CERN PS complex. Each mark depicts location of a FBCT.

the AD) or extracted to the SPS.

The SPS has a circumference of 6.9 km. It accelerates protons up to 450 GeV/c, and lead ions to 400 GeV/c proton equivalent energy. The particles are extracted either to fixed target experiment in the North Area, or to the CERN Neutrinos to Gran Sasso (CNCS) experiment. The SPS is also a part of the injector chain for the LHC. The particles are transported into both LHC rings via the TI2 and TI8 transfer lines.

4.3.1 Types of the beams produced at CERN

Nowadays, the beam is produced in the CERN PS complex (Fig. 4.26) in cycles with period of 1.2 s^3 . Every cycle a beam is sent to an user. Users represent physical experiments, each having different beam properties and targeting different CERN experimental areas. All currently active users are referenced in a list called a “super-cycle”. The list defines a sequence in which the users receive the accelerated particles.

During the acceleration process the particles are bunched. The bunch lengths of produced beams vary between 50 ns to some hundreds of picoseconds. The CERN PSB can provide up to 16 bunches having a total length of 4 to 15 μs in a single acceleration cycle. Larger accelerators in the chain provide longer bunch trains, but they require longer filling time.

The beam intensity is measured during complete acceleration process. Many variants of the BCTs are used. Table 4.3 summarises properties of those devices.

³The limit is given by the length of the PSB acceleration cycle. Studies were done to decrease the acceleration time to 900 ms [55], which would increase the rate of the LHC filling. The upgrade is foreseen to the time of full performance of the LHC.

Transformer Type / Availability	Signal Acq. Time	Dynamic Ranges	Absol. Acc. Error	Calibration Type	Frequency Range
DC	300 ms	4 ranges			DC
$\approx 100 \mu s$	$\vdots$	$1 \mu A - 1 A$	$<1 \%$	each cycle	$\vdots$
after injection	3 s				10 kHz
AC	$2 \mu s$	1 range			7 Hz
LINAC _{II}	$\vdots$	300 mA	$<5 \%$	each cycle	$\vdots$
Immediately	$120 \mu s$				350 kHz
AC	typ.	1 range			6 Hz
LINAC _{III}	$560 \mu s$	150 mA	$<10 \%$	each cycle	$\vdots$
Immediately					700 kHz
AC PSB	600 ns	3 ranges			500 Hz
AC PS	$\vdots$	2 mA-300 mA	$<5 \%$	once per year	$\vdots$
Immediately	$15 \mu s$	peak-peak			200 MHz
AC AD		1 range			500 Hz
	$4 \mu s$	50mA	$<5 \%$	once per year	$\vdots$
Immediately		peak-peak			200 MHz
AC SPS	25 ns bunch				500 Hz
	by bunch		$<2 \%$	on demand	$\vdots$
Immediately					750 MHz

Table 4.3: *Properties of the BCTs used at CERN.*

4.3.2 LINAC and LEIR FBCTs

Detailed history of the CERN LINACs can be found in [49]. Functional description of the particle sources can be found in [22]. The particle source types are discussed e.g. in [18] and [10]. At CERN the particles are currently produced by a proton and ion LINAC.

Ten BCTs measure the intensity and the acceleration efficiency of LINAC_{II}. The LI.TRA02 (Fig. 4.27) is a commercial type, made by Pearson (type 110). It has 100 secondary turns and a sensitivity of 100 V/A. It measures the beam intensity directly at the output of the duoplasmatron source. All other FBCTs in the LINAC_{II} beam chain were developed at CERN. The magnetic core material is Ultraperm 10 containing 80 % NiFe alloys. It reaches the highest permeability ($\mu \approx 300000 @ 50 Hz$) of all commercially available Permalloy materials. The B-H curve can be shaped by a thermo-magnetic treatment. The material saturates at a relatively low magnetic induction ($B_s \approx 0.8 T$) compared to nanocrystalline materials (e.g. VITROPERM with $B_s=1.2 T$). For the intensity measurement this is not an issue as long as the FBCT is installed far from kicker, focusing or bending magnets. The Ultraperm 10 is produced by VacuumSchmelze GmbH, Hanau, Germany, in form of torroid rings. The torroid winding was fabricated at CERN. Using

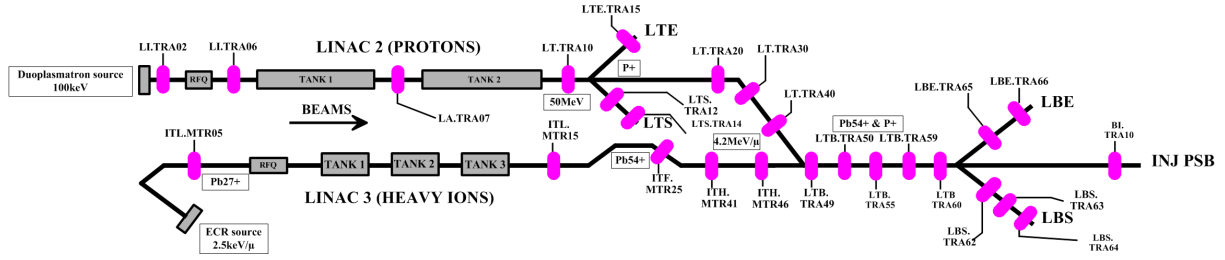


Figure 4.27: $LINAC_{II}$ and $LINAC_{III}$ injector chain. The violet marks depict locations of the intensity measurement devices.

ten measurement turns the total sensitivity of 25 V/A is achieved. A typical bandwidth of the installed transformers is 7 Hz to 350 kHz.

Five FBCTs are installed in the $LINAC_{III}$ beam chain. The toroid's are made of cobalt based amorphous alloy VITROVAC 6025 FXCF 505 (VacuumSchmelze GmbH). It has an advantage over the NiFe alloys in its low coercivity and magnetostriction and good mechanical properties. While it is a soft-magnetic material, it is mechanically rigid and can be machined without loss of its magnetic properties. It is however less resistant to high temperatures due to a tendency to crystallise [83].

The FBCTs installed in the common transfer line of the LINACs must process signals of high dynamic ranges and different lengths. This is due to common processing of protons and ions, and measurement of ions injected and ejected from the LEIR storage ring. The LEIR accumulates low intensity ion beams. They are injected in long trains ($\approx 40 \mu A$ in $250 - 500 \mu s$, multi-turn injection), accumulated up to $160 mA$, recaptured and ejected in 4 bunches with 100 ns bunch spacing [60, 71]. The toroids are fabricated using Ultraperm 10 with a single 10 turn winding. In order to cope with the dynamic range there are pre-amplifiers installed close to each transformer. All pre-amplifiers provide three outputs with FS of 160 mA, 16 mA and $100 \mu A$. The amplified signals are converted into a digital form and the beam intensity is calculated by numerical integration. Haar wavelets are used to lessen the noise on the $100 \mu A$ dynamic range [53]. The bandwidth of the installed transformers is 6 Hz to 700 kHz.

Mechanical design of the common part of the LINAC FBCTs is shown in Fig. 4.28. The toroid is installed over the vacuum chamber which is electrically isolated by the Alumina (Al_2O_3) ring. Three layers of high permeability material (μ -metal) form an external magnetic shielding. The μ -metal is isolated from outer shielding made of Armco, acting as an additional protection against external magnetic fields.

Each FBCTs system is equipped with a constant current pulse generator serving as a calibrator. The amplitude of the current injected into the calibration turn is fixed to 100 mA. As the particle train leaves the LINACs only each 1.2 seconds there is enough time to perform the calibration before every measurement cycle. Calibration takes approximately $20 \mu s$. The figure 4.29 shows the intensity measurement for the ISOGPS cycle measured by the LTB.TRA50 installed in the common LINAC transfer line.

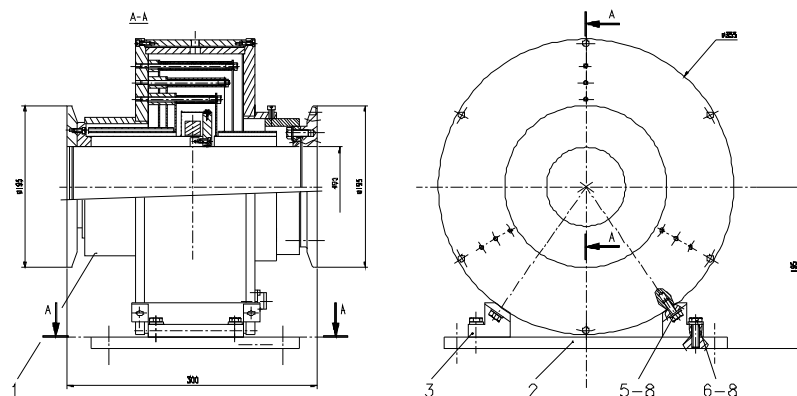


Figure 4.28: *Mechanical design of LINAC_{II} and LINAC_{III} FBCTs*

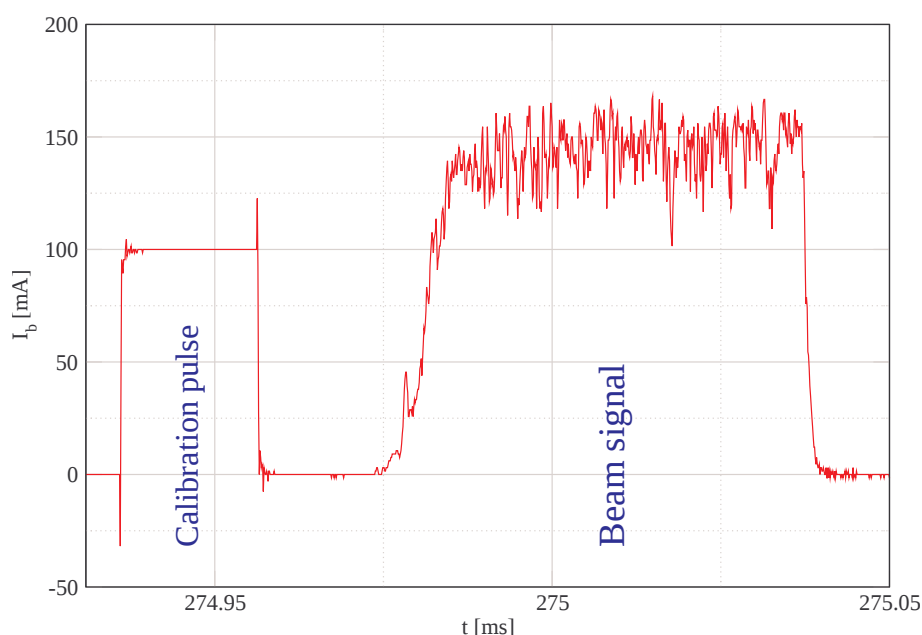


Figure 4.29: *Measurement signal of the ISOGPS beam on LTB.TRA50 LINAC transformer.*

4.3.3 FBCTs in the transfer lines and circular accelerators

The particle train of 20 to 200 μs leaving the LINAC is sliced by a PSB distributor and accelerated in four separate PSB rings (Fig. 4.30). Each slice is maximum 50 μs long. One PSB revolution period takes approximately 2 μs , hence the 50 μs slices are “layered” in each of the four rings. This process is called multi-turn injection and results in an increase of the bunch intensity produced by the PSB.

To calculate the injection efficiency an intensity measurement is performed for each of the four injection lines (BIn.TRA20). The measurements are compared to the total intensity measured before the particle train is sliced (BL.TRA10). This permits to estimate the losses caused by the beam distribution into the rings. Measurements are performed by the same type of the FBCTs as used in LINAC transfer lines.

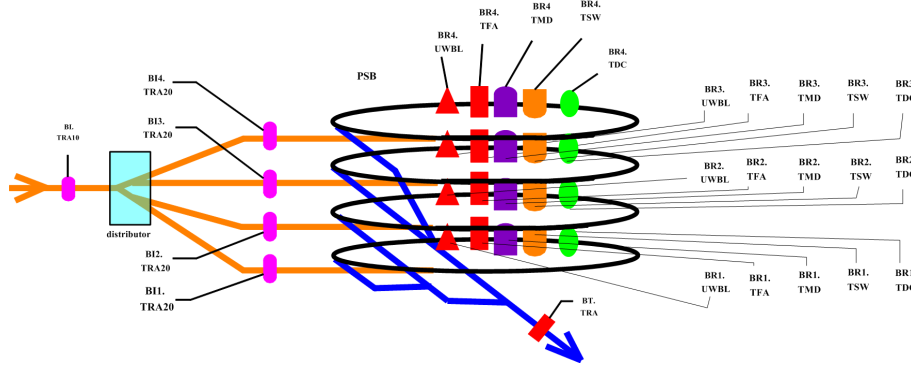


Figure 4.30: The PSB and the transfer lines. The violet and red marks depict the locations of the FBCTs. The green marks identify the DCCTs.

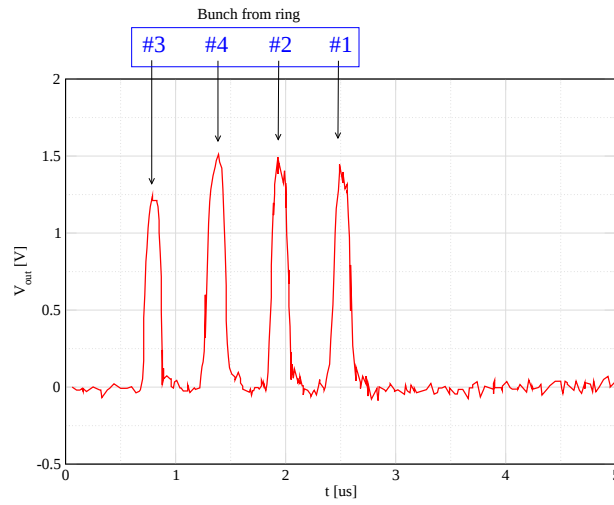


Figure 4.31: Measurement signal of the ISOGPS as observed during ejection from the PSB.

The debunched and sliced beam undergoes an "RF treatment" during the acceleration, which results in a formation of particle bunches in each of the four PSB rings. At the end of the acceleration cycle the bunches are consecutively ejected from the PSB (Fig. 4.31). Depending on the user, the beam consists of 1 to 8 bunches. Their intensity differs from 5×10^9 charges, as in the case of the LHC pilot beam, through some 1600×10^{10} charges for anti-proton production, to more than 3000×10^{10} charges for the ISOLDE. Lead ions with maximum intensity 4×10^{10} charges have also been accelerated in the PSB.

The shape of the beam varies during the acceleration cycle hence the constraints put on the FBCTs change. A bandwidth of more than 100 MHz is required to measure individual ejected bunches. The length of the particle train changes from $\approx 200 \mu\text{s}$ to $\approx 15 \mu\text{s}$. The multi-turn injection further increases the required dynamic range for both ions and protons.

The intensity measurement in the PS ring is provided by a single FBCT. As the FBCT is not equipped with the DC signal restoration, the measurement is valid only for few revolution periods after the injection, what is enough to estimate the PS injection efficiency.

The FBCT uses a toroid made of Ultraperm 10. A six turn winding loaded by a

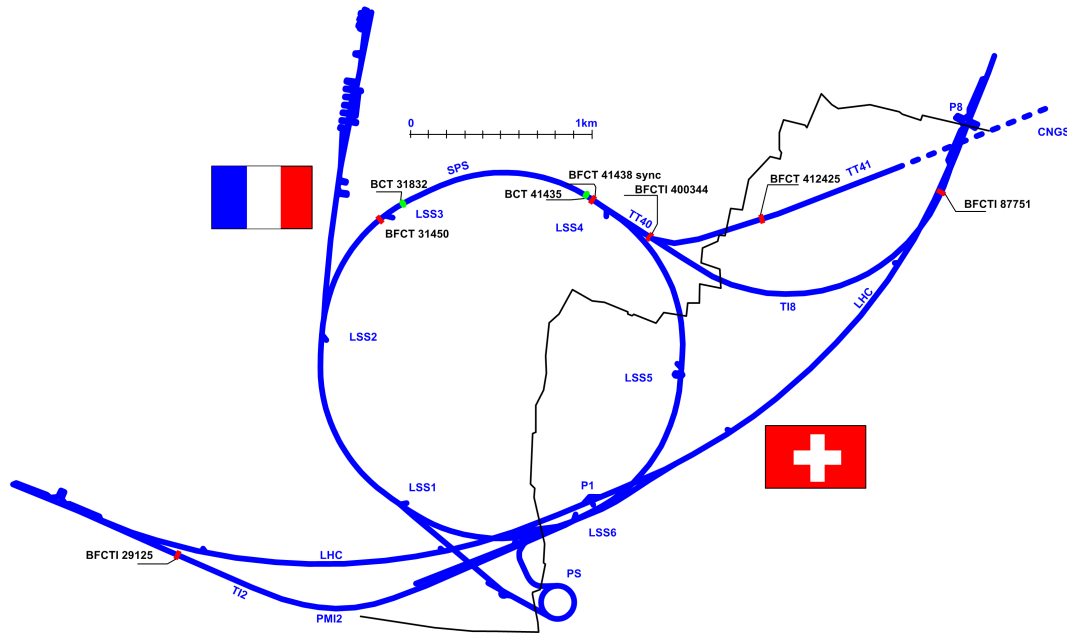


Figure 4.32: *The CERN SPS and adjacent transfer lines*

combined $5\ \Omega$ load is a compromise between the sensitivity of the FBCT ($S = 0.75\ \text{V/A}$) and the LF cut-off. Droop of the transformer is kept during the measurement below 1 %. The measured signal is pre-amplified by an amplifier installed near the FBCT. Long cables connect the pre-amplifier with electronics installed outside of the accelerator tunnel. The electronics is based on an OA analogue integrator which provides a FS output voltage of $\pm 5\ \text{V}$. The integrated signal is sampled by an ADC from Pentland (type MPV908). The MVP908 provides 64 single-ended inputs, which are sampled by a 12-bits ADC at maximum sampling rate of 100 kHz. The acquisition system is installed in a VME crate. A PPC controller running LynxOS v4.0 RTOS provides all necessary software facility to read and correct the measured data, and to publish them to the control system.

Additional one turn wound on the toroid provides a calibration turn. The calibrator is shown in Fig. 4.20. The calibration is performed manually once or twice per year.

The described installation is very common for the majority of the FBCTs used in the PSB and the PS complex [27]. The cable lengths vary from ≈ 30 metres to more than 600 m. This justifies the use of pre-amplifier. Dynamic range of the measurement is assured using multiple dynamic ranges. Generally three measurement ranges per FBCT are available. The FS of 4×10^{10} charges is used to measure the ions and two dynamic ranges, 2×10^{12} and 4.5×10^{13} charges, measure the protons.

Compared to the PS complex, the SPS does not deal with a large variety of beams, however their energies and the intensities are much higher. This imposes additional constraints on the intensity measurement. To measure bunch by bunch a signal bandwidth greater than 300 MHz is required and a large circumference of the machine imposes the LF cut-off to be lower than 1.5 kHz. This is to assure a droop lower than 1 % per revolution period. As these requirements are common for the SPS ring and adjacent transfer lines,

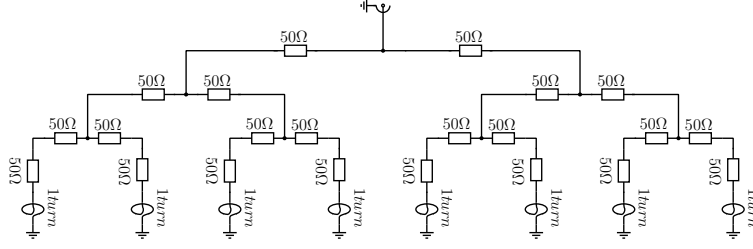


Figure 4.33: Calibration turn for the SPS FBCTs. Eight parallel turns form a $50\ \Omega$ single turn winding. This configuration provides better distribution of the magnetic flux in the core, hence better transmission of the calibration signal can be achieved. The calibration turn is realised on a 0.4 mm thick FR4 material

the same measurement devices are used to perform the intensity measurement in the entire SPS complex.

The used toroid is a commercial product, fabricated by Bergoz Instrumentation [74], St. Genis-Pouilly, France. The 40 turns measurement winding already includes a $50\ \Omega$ back-matching resistance. The sensitivity is $1.25\ \text{V/A}$. The calibration winding of the FBCT is formed by a group of 8 parallel turns forming a $50\ \Omega$ resistive load (Fig. 4.33).

The acquisition electronics is realised using the LHCb2002 integrator discussed in Sec. 4.2.3 d). The integrator and a 14 bit ADC are mounted on a mezzanine (IBMS, [45]) which forms together with a Digital Acquisition Board (DAB) an acquisition system. Two mezzanines are required for each FBCT as bunch by bunch and average turn intensity measurements are needed. The first mezzanine measures a full-bandwidth signal, while the other a signal limited in bandwidth to $\approx 2.5\ \text{MHz}$. Both mezzanines are installed on a DAB, which collects the measured data using an FPGA, calculates the intensities, and stores them for later analysis. It also provides a communication channel between the IBMS mezzanines and the 64 bit version of the VME (VME64x) bus interface. The DAB is commonly used for the beam instrumentation of the LHC and the SPS [47] [50].

Mechanical construction of the SPS transformer is shown in Fig. 4.34.

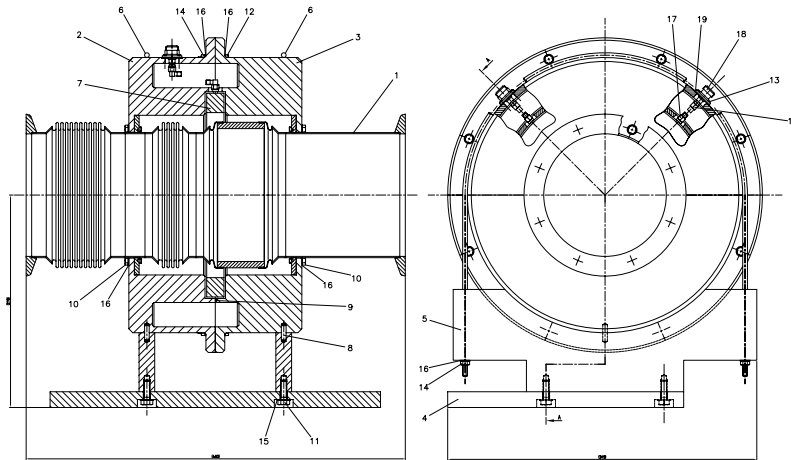


Figure 4.34: The FBCT for the SPS ring and adjacent transfer lines.

4.4 The LHC intensity measurement

The LHC, the largest and the most powerful particle accelerator ever built. It firstly started the beam production at the end of 2008. A year later, in November 2009, the LHC has already set the new record in the achieved beam energy - the protons were accelerated to 1.2 GeV. Currently (June 2010), stable proton beams accelerated at 3.5 GeV are delivered to the experiments and first collisions were produced.

The LHC's circumference reaches nearly 27 km (revolution frequency is 11.245 kHz) and it accelerates two counter rotating proton beams from an initial energy of 450 GeV to 7 TeV. As two contradirectional beams are accelerated simultaneously, 14 TeV centred mass collisions will be produced. The proton ultimate intensity is 1.7×10^{11} charges per bunch⁴ [56] with maximum of 2808 bunches per beam accelerated.

The LHC also accelerates fully stripped $^{208}\text{Pb}^{82+}$ lead ions. The ions are accelerated from an initial energy of 177 GeV/nucleon to the energy of 2.76 TeV/nucleon. The number of ions per bunch is 7.0×10^7 , which corresponds approximately to 5.6×10^9 charges. To achieve such intensity the ions are first accumulated in the LEIR before being injected into the LHC [87] via the PS and the SPS.

For the LHC a new type of FBCTs had to be developed. Several problems needed to be addressed, e.g. achieving the required bandwidth, choice of an integration method, and design of an algorithm for real-time intensity calculation. In addition to transfer line measurements, the FBCTs must also measure the circulating beams and hence a suitable DC restoration system had to be implemented. Several technological issues needed to be solved, e.g. protection of the FBCTs against high temperatures due to the vacuum bake-out procedure, a nonstandard connection of the vacuum chambers to the ceramics, and deposition of various coatings on the vacuum chamber and the ceramics for beam impedance optimisation. An appropriate integration of the intensity measurements into the Machine Protection System (MPS) was also required. The FBCTs are installed in the LHC tunnel while the electronics is located in a neighbouring equipment gallery. To improve the measurement algorithms on-fly a maximum attention was paid to develop remotely configurable and controlled electronics and remote observation system.

In this introductory section a summary of requirements is presented. Their collection and optimisation was an iterative process between the system designers (author of the thesis and his colleagues) and the users. It resulted in a technical specification, which defines the precision of the measurement for various operational modes, interfaces to the accelerator machine and serves as a basis for the development. Details of the specification are summarised in [66]. The information of possible injection scenarios is discussed in [59]. Following sections provide an extract of the important information from the above referenced documents.

⁴abbreviated using "ch/b"

Particle type	Number of charges per bunch	Number of bunches	Bunch spacing [ns]	Average beam current [mA]	RMS length [ns]
proton	2×10^9	1	88925	0.0036	0.68
	↓	↓	↓	↓	
	1.7×10^{11}	2808	24.95	860	⋮
$^{208}\text{Pb}_{82}^+$	$5.6 \times 10^8 \rightarrow$	$1 \rightarrow$	1350	$0.001 \rightarrow$	⋮
	$5.6 \times 10^9 \rightarrow$	$62 \rightarrow$	↓	$0.01 \rightarrow$	
		592	100	6.5	0.28

Table 4.4: The LHC beam parameters. The 5.6×10^9 charges correspond to a bunch population of 6.8×10^7 ions.

4.4.1 Functional specification of the intensity measurement using FBCTs

Table 4.4 summarises the beam parameters applicable to the intensity measurements. Some key beam intensity values are recognised:

- 5×10^9 charges - an intensity of the LHC “pilot” *proton* bunch. It corresponds to the maximum beam current that can be lost without inducing a magnet quench⁵.
- 5.6×10^9 charges - a “nominal intensity” *ion* bunch
- 1.15×10^{11} charges - a “nominal intensity” *proton* bunch
- 1.7×10^{11} charges - an “ultimate intensity” *proton* bunch

The primary observable is a number of charges corresponding to a “fraction” of the beam measured by the FBCT. The fraction is defined as an interval of ≈ 10 RF periods⁶ forming a bunch. Bunch by bunch intensity measurement is required what permits to see the imperfections of the beam production and acceleration. Other inherited information must be provided as well. Amongst others the injection and the extraction efficiency, information of a current loss, and a status of whether the beam circulates in the machine.

Tables 4.5 and 4.6 summarise required absolute accuracy and resolution for the protons and lead ions intensity measurements. The absolute accuracy represents a closeness of the measured value to the absolute value of the measured intensity. The resolution represents the smallest detectable increment of the measurand. As the specification of the FBCT measurement modes uses those terms to express the quality of the measurement, the same convention is adopted throughout this dissertation thesis.

⁵The tests in September 2008 discovered that even 5×10^9 ch/b can trigger the quench protection system hence an another “probe” beam (2×10^9 ch/b) was defined and is used during the set-up of the LHC. In this document the minimum “functional” intensity of 5×10^9 ch/b is still assumed as this was a valid definition at the time of the project design phase.

⁶RF period is one period of the RF frequency driving the beam, i.e. $1/400789650$ Hz

							Pilot bunch		Nominal bunch	
Monitor type	Operation mode	Beam Scenario	Number of charges per bunch	Ob-servation mode	Ob-servation rate	Integration time per acquisition	Accuracy	Resolution	Accuracy	Resolution
Single pass to few pass	bunch charge						$\pm 20\%$	$= \pm 20\%$	$\pm 3\%$	$= \pm 1\%$
							$\pm 10^9$ p	$\pm 10^9$ p	$\pm 3 \times 10^9$ p	$\pm 10^9$ p
							$\pm 20\%$	$= \pm 10\%$	$\pm 3\%$	$= \pm 1\%$
Single pass to few pass	injection efficiency	pilot bunch to ultimate SPS batch	5×10^9 to 1.7×10^{11}	turn by turn	bunch by bunch	1 turn	$\pm 10\%$	$= \pm 10\%$	$\pm 2\%$	$= \pm 1\%$
							$\pm 5 \times 10^8$ p	$\pm 5 \times 10^8$ p	$\pm 2 \times 10^9$ p	$\pm 10^9$ p
							$\pm 10\%$	$= \pm 10\%$	$\pm 2\%$	$= \pm 1\%$
Circulating bunch monitor	extraction efficiency						$\pm 10\%$	$= \pm 10\%$	$\pm 2\%$	$= \pm 1\%$
							5×10^2 turns	10^8 p	2×10^9 p	10^9 p
							$\pm 10\%$	$= \pm 10\%$	$\pm 1\%$	$= \pm 1\%$
Circulating bunch monitor	normal mode	pilot bunch to ultimate beam	2×10^9 to 1.7×10^{11}	turn by turn	bunch by bunch	2×10^2 to 10^4 turns	$\pm 10\%$	$= \pm 10\%$	$\pm 1\%$	$= \pm 1\%$
							5×10^8 p	10^8 p	10^9 p	10^9 p
							$\pm 10\%$ of pilot bunch	$= \pm 1\%$ of nominal bunch	$\pm 2\mu A$	$= \pm 1\%$
Cross calibration	absolute calibration						10^4 turns to 10^5 turns			
							$\pm 10\%$			
							$\pm 10\%$			

Table 4.5: Required absolute accuracy and resolution of the proton intensity measurement using FBCTs

	Number of charges			
	Nominal/10		Nominal	
	7×10^6 ions=		7×10^7 ions=	
	5.7×10^8 charges		5.7×10^9 charges	
	Accuracy	Resolution	Accuracy	Resolution
Injection efficiency Integration: 1 turn	$\pm 50\%$	$\pm 50\%$	$\pm 10\%$	$\pm 10\%$
Extraction efficiency Integration: 1 turn	$\pm 50\%$	$\pm 50\%$	less than $\pm 10\%$	less than $\pm 10\%$
Cross-calibration	$< \pm 50\%$ $< \pm 3 \times 10^8$ ch.		$< \pm 10\%$ $< \pm 6 \times 10^8$ ch.	

Table 4.6: Absolute accuracy and resolution of the Pb ions intensity measurement using FBCTs

Four basic measurement modes must be provided:

- **Capture** - a snapshot of the intensity measurement in each bunch slot for a specified number of turns. Results in 3564 measured values per turn.
- **Turn Sum** - measurement of a turn intensity. Results in a single intensity value per turn corresponding to sum of intensities measured in each bunch slot within that turn.
- **Slot Sum** - sum of bunch slot intensities measured over specified number of turns. Results in 3564 measured values.
- **Sum Sum** - measurement of *Turn Sum* averaged over 225 turns. Results in a single value each 225 turns.

The **injection efficiency** is defined as a ratio of number of charges measured at the end of the injection insertions $N_{p,i}$, and number of charges measured at the corresponding transfer lines $N_{p,t}$ (Fig. 4.32, BCTFI 29125 and BCTFI 87751):

$$E_i = 100 \frac{N_{p,i}}{N_{p,t}} \quad (4.66)$$

The absolute accuracy and the cross-calibration of the transfer lines and the rings FBCTs must be better than ± 4 % of the *nominal intensity*. This implies an absolute accuracy of the FBCTs better than $\pm 3.5 \times 10^9$ charges (± 3 %). A relative accuracy of ± 1 % is required. These constraints permit to provide a reliable measurement of the fast current loss⁷.

⁷The magnet quench level was estimated to 10^{10} p. lost in 10 metres at 450 GeV [72]. Satisfying an absolute accuracy of $\pm 3.5 \times 10^9$ charges a safety factor of 2 to the quench level is provided when injecting a nominal intensity bunch.

The absolute accuracy of the pilot bunch measurement is relaxed to $\pm 1 \times 10^9$ charges ($\pm 20\%$). The same accuracy applies to the measurement of the nominal charge of the lead ions.

The Extraction efficiency is defined as a ratio of intensity measurements of particles extracted to the dump lines, and the last measurements performed in the rings before the beam was extracted. The required absolute accuracy of $\pm 2.3 \times 10^9$ charges is more restrictive due to the importance of clean extraction at high energy. A measurement resolution better than $\pm 1\%$ for the proton nominal beam, and $\pm 10\%$ for the pilot beam is needed.

The beam circulating flag detector (BCF) informs of whether a beam circulates in the LHC. It interlocks an injection of high intensity beams. The BCF signal generation is based on a direct comparison of the beam current to a threshold. The required threshold of $2 - 4 \times 10^9$ charges measured over one revolution period permits to detect the circulating pilot beam.

4.4.2 The LHC injection scenarios

As already discussed, DC recovery techniques must be applied to the signal provided by the FBCTs. In case of the LHC this is valid not only for the rings measurements but also for the dump lines. A fully loaded LHC ring represents a bunch train $\approx 88.9\ \mu\text{s}$ long, which in case of need will be entirely extracted into the beam dump in one shot. This puts a considerable constraint on the droop of the used FBCTs. It was shown (Sec. 4.2.3) that a key information to solve the droop issue is hidden in the beam parameters and related injection scenarios. Six baseline injections were standardised [59]. Two of them describe the operation with the lead ions, the others the operation with the protons. Next sections discuss two extreme cases applicable to the intensity measurement.

a) The proton injection scenario

The proton full-injection scenario with 25 ns bunch spacing is depicted in Fig. 4.35. The particle production for the LHC starts in the PSB, where totally 6 relatively long bunches are produced. The bunches are extracted in two batches, containing 4 and 2 bunches. After being injected into the PS they undergo an RF treatment. Twelve new bunches are created from each injected bunch by a bunch splitting. First splitting (triplication) starts at 1.4 GeV, the second (double-double) at 25 GeV. 72 bunches spaced by 25 ns are obtained and circulate in the machine after the PS acceleration cycle.

The PS extraction kicker then sends the particles to the SPS. To keep the losses at the minimum, there has to be no particles passing the kicker aperture during the magnetic field ramp-up. To satisfy this condition a gap at least 12 bunch lengths long must exist in the bunch train. The gap is generated by omitting the bunch at 7th position during the second injection from the PSB (Fig. 4.35, black bunch).

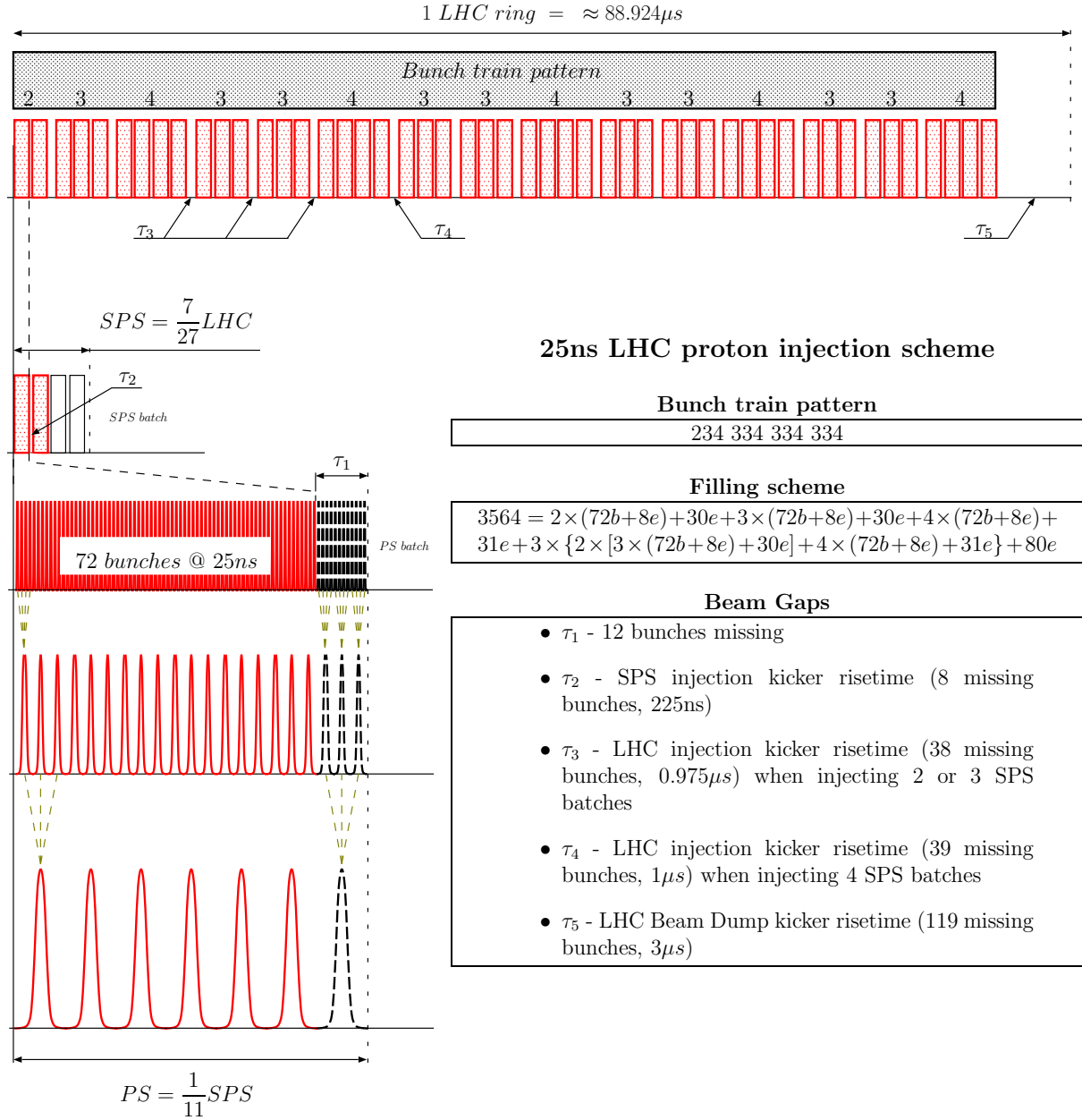


Figure 4.35: The 25 ns proton filling scheme for the LHC

The SPS injection kicker requires a gap at least 8 bunch lengths long. By adjusting the timing of injection and extraction kickers the PS kicker extraction gap vanishes. Only 8 empty slots separate each PS batch in the SPS. Up to 4 PS batches can be injected to the SPS. After acceleration the beam is injected into the LHC. An another gap is needed in the bunch train to assure a correct injection.

The batches are transferred from the SPS into the LHC in a specific pattern in order to get the bunches colliding in the experiments at points 1, 2, 5, 8, and providing equal collision rate for the main experiments ATLAS and CMS. For the 25 ns proton scheme the pattern is “234 334 334 334”. Each number corresponds to a number of PS batches transferred from the SPS into the LHC in a single injection. When injecting four batches, a 39 bunch lengths gap is left between consecutive SPS batches. In all other cases

a 38 bunches long gap is sufficient. An extra $\approx 3 \mu s$ abort gap (119 bunches) must be left free of any particles due to the LHC beam dump kickers ramping. Using this injection scenario a total of 2808 slots out of 3564 available could be physically used.

b) The ion injection scenario

The ion injection scenario is based on the same injection principle as the proton one. The LHC pre-injectors generate sets of 4 bunches spaced by 100 ns, which are accumulated in the SPS in either 8, 12 or 13 batches spaced by 225 ns. The filling pattern is “8 13 13 12 13 13 12 13 13 12 13 13”. The total number of the bunch positions is 891, from which 592 can contain ion bunches. One SPS filling cycle takes approximately 54 seconds and entire LHC ring can be filled in 10 minutes.

5 DETAILED OBJECTIVES FOR THIS WORK

The work presented is in the domain of beam diagnostics related to particle accelerators. The results of the work are applied to fast beam intensity measurements of the LHC beams. The proposed method is foreseen to be optimised and refined as more experience will be gained in the first years of LHC operation. It is also foreseen to upgrade the SPS system to use the LHC measurement method.

To introduce the problem of the beam intensity measurements a basic theory was presented. Currently available technical solutions were discussed. It was shown that the LHC cannot simply re-use the SPS system, but a new more sophisticated solution must be designed and implemented. One of the reasons is the dynamic range of the measured signals and their temporal properties, which implies an implementation of several measurement channels using different bandwidths. This introduces a problem of cross-calibration and problems related to a relative calibration of multiple FBCTs, not necessarily of the same type. Another limitation exists due to the large circumference of the LHC, which puts tougher constraints on lowering the FBCTs droop to minimise the measurement errors during the injection.

The LHC FBCT system is also special in several other details:

- a robust system protecting the machine against damage has been developed (Machine Protection System (MPS)) [67]. The MPS expects the FBCTs to deliver information about any beam loss. Another information provided by the FBCTs is a status of whether there is any beam circulating in the LHC [64].
- to ensure Ultra High Vacuum (UHV) (10^{-10} to 10^{-11} mbar) in the beam pipe a chemical pumping (e.g. [23]) is used. A non-evaporable getter (NEG) helps to absorb residual gases present in the vacuum system. The NEG is applied on the inner wall of the vacuum chamber [58] and is activated before pumping by high temperature baking ($> 250^{\circ}\text{C}$). Temperatures exceeding 60°C are already damaging to the magnetic material used in the transformer toroid hence an appropriate protection had to be developed.

Development of an FBCT satisfying the measurement specification is an iterative process of finding the compromises between the measurement requirements, and limitations given by real implementation. In order to successfully fulfil the task the project was fragmented into specific topics which were studied separately. Following list shows the most important subjects discussed in this dissertation thesis:

1. Analysis of the requirements for the intensity measurement
 - Determination of the required dynamic ranges
 - Determination of the transformer bandwidth
2. Technological and implementation analysis
 - System analysis with respect to tough machine impedance and aperture policy
 - Technological analysis of the FBCT's UHV compatibility

- Analysis of the vacuum chamber fabrication processes
 - Study of issues related to the equipment integration into the LHC tunnel
 - Analysis of design's compatibility with industry's manufacturing capabilities
3. Analysis and implementation of the electronic chain
- RF distribution circuits
 - Machine protection system - circulating beam detector
 - Calibrator design
4. Analysis of the obtained results

6 ANALYSIS OF THE REQUIREMENTS FOR THE INTENSITY MEASUREMENT

6.1 Determination of the required dynamic ranges

6.1.1 Input signal definition

Signal function: The particle distribution density of the beam core is often approximated by a Gaussian function (e.g. [35]). For some types of calculations this is a very convenient approximation, however for other calculations this approximation causes problems due to its infinitesimal properties. For “electrical” calculations the more suitable crest of cosine is preferred. This approximation is defined on a specific interval and hence simplifies the mathematical operations. However it models precisely either bunch amplitude or its charge, but not both at the same time with reasonable accuracy. To achieve a better approximation of both parameters a higher order function is required. A good candidate is the following function:

$$f(t) = \left\{ \begin{array}{ll} a[\cos(b(t - kT_0))^2 - 1]^2 & t \in \langle kT_0; kT_0 + \frac{\pi}{b} \rangle \\ 0 & t \in \langle kT_0 + \frac{\pi}{b}; (k+1)T_0 \rangle \end{array} \right\}_{k \geq 0} \quad (6.1)$$

where T_0 is the bunch period. Parameter kT_0 aligns the core function, naturally periodic to an interval $\frac{2\pi}{b}$, to the period T_0 . Equation (6.1) can be rewritten to a form more suitable for numeric computations:

$$f(t) = \left\{ \begin{array}{ll} a [\cos(b(t - \zeta(t)))^2 - 1]^2 & 0 \leq t - \zeta(t) < \frac{\pi}{b} \\ 0 & otherwise \end{array} \right. \quad (6.2)$$

where $\zeta(t)$ is an equivalent of the element kT_0 :

$$\zeta(t) = \text{round} \left(\frac{t}{T_0} - \frac{1}{2} \right) T_0 \quad (6.3)$$

The function *round* trims its argument to the nearest integer number.

Equation (6.1) is generated by shifting and trimming a “ $\cos(t)$ ”⁴ function. It represents a bunch like shape on an interval $\langle 0; \frac{\pi}{b} \rangle$, as seen in Fig. 6.1. The function is causal as for negative values of k it returns zero.

Equivalent pulse widths: The constants a and b define the bunch amplitude and width. They can be expressed in terms of the Gaussian function defined by its surface and the parameter σ . The surface is related to a bunch charge, and the σ to the bunch width. As the bunch charge represents a multiplicative constant applied to both surfaces of the Gaussian and the base function, the surface of the base function can be normalised to ‘1’. This implies that every calculation of an absolute error must be ‘normalised’ to the

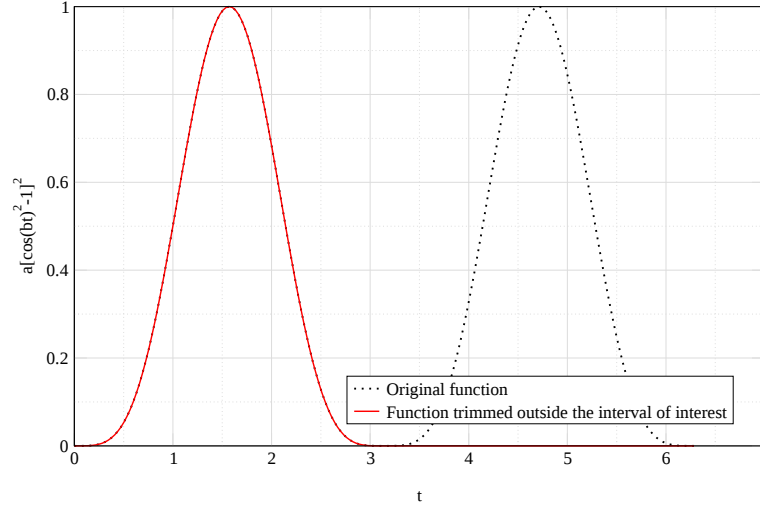


Figure 6.1: The $\cos(t)^4$ approximation of the Gaussian function. The black trace shows a natural periodicity of the approximation. The red trace depicts the piecewise function defined in the Eq. (6.1). The function is defined on the interval $\langle 0; \frac{\pi}{b} \rangle$. The plot was made using $a = 1 \text{ s}^{-1}$, $b = 1 \text{ s}^{-1}$ and $\zeta = 0$

number of charges in the bunch by an appropriate multiplicative constant. The surface of the base function is expressed as:

$$s_f = \int_0^{\frac{\pi}{b}} a [\cos(bt)^2 - 1]^2 dt = \frac{3\pi a}{8b} \quad (6.4)$$

Hence the constant a can be defined:

$$s_f \stackrel{!}{=} 1 \longrightarrow a = \frac{8b}{3\pi} \quad (6.5)$$

As σ is an unique property of the Gaussian function it must be transformed into a more general parameter - the Full Width at Half Maximum (FWHM). The FWHM of the Gaussian function, denoted using subscript “G”, is defined as:

$$FWHM_G = 2\sqrt{2}\sqrt{\ln(2)}\sigma \quad (6.6)$$

To define the FWHM of the base function (6.1) a time t at which the base function has a magnitude equal to the half of the maximum must be found. Two such solutions exist and their absolute difference defines the FWHM as:

$$FWHM_C = \frac{\pi - 2 \arccos\left(\frac{1}{2}\sqrt{4 - 2\sqrt{2}}\right)}{b} \quad (6.7)$$

The constants a and b are expressed in $[\text{s}^{-1}]$ units and are found by solving Eqns. (6.5), (6.6) and (6.7):

$$a = \frac{2\sqrt{2}}{3\pi\sqrt{\ln(2)}\sigma} \left(\pi - 2 \arccos\left(\frac{1}{2}\sqrt{4 - 2\sqrt{2}}\right) \right) \approx \frac{0.41226}{\sigma} \quad (6.8)$$

$$b = \frac{\sqrt{2}}{4\sqrt{\ln(2)}\sigma} \left(\pi - 2 \arccos\left(\frac{1}{2}\sqrt{4 - 2\sqrt{2}}\right) \right) \approx \frac{0.48569}{\sigma} \quad (6.9)$$

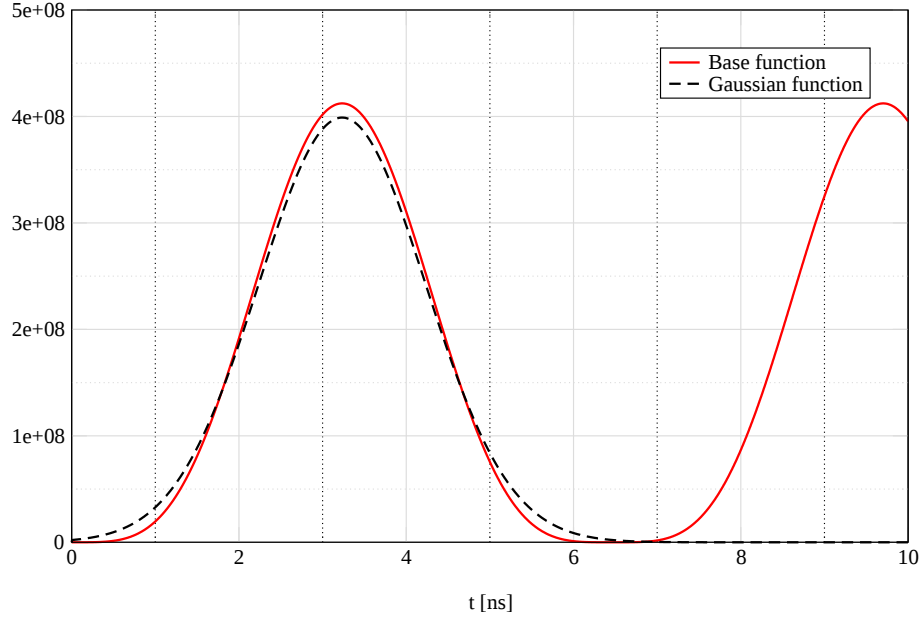


Figure 6.2: The black trace depicts the Gaussian function ($\sigma = 1 \times 10^{-9}$ ns). Equivalent naturally periodic base function is shown in red. The constants a and b are calculated using Eqns. (6.8) and (6.9).

Particle type	Number of charges per bunch [-]	Peak beam current [mA]	
		injection	extraction
protons	1.7×10^{11}	16511	40098
$^{208}\text{Pb}_{82}^+$	5.74×10^9	54	132

Table 6.1: Peak currents for the proton and ion LHC operation mode

A plot of the the base function and the Gaussian equivalent is shown in Fig. 6.2. Compared to a simple cosine approximation a better match of both surface and amplitude criteria is achieved. An exact match of a preferred parameter can be obtained by optimising the constants a and b . Using Eqns. (6.8) and (6.9) the base function exhibits higher peak amplitude than the corresponding Gaussian function, making the approximation safe for dynamic range calculation.

6.1.2 Dynamic range of the beam signal

Quick inspection of Eq. (6.1) reveals that the constant a is related to the peak value of the base function, and Eqns. (6.8) and (6.9) normalise its surface to unity. Hence to obtain the peak beam current the constant a must be multiplied by a total bunch charge:

$$I_{peak} = a \cdot N_p \cdot e, \quad (6.10)$$

where N_p is a number of charges in the bunch, and e is the elementary charge.

Dynamic range	FS range per bunch
High Gain	1.0×10^{10} charges
Low Gain	1.7×10^{11} charges

Table 6.2: *Proposed full-scale ranges for defined dynamic ranges*

Table 6.1 summarises the properties of the beam signal at both extremities of the dynamic range. The maximum signal amplitude is generated by the ultimate LHC proton beam whereas the minimum by the probe beam. The peak currents are calculated using Eq. (6.10) for both injection and “flat-top” (extraction) because the σ shrinks during the acceleration from 680 ps to some 280 ps.

The dynamic range of the beam signal exceeds 40 dB with the peak current of the ion beam being very low. The situation is made worse by the scaling factor introduced by the number of secondary turns of the FBCT. A single dynamic range would not provide satisfactory measurement results for low currents, hence two dynamic ranges had to be implemented. The high gain (HIGAIN) dynamic range covers the measurement of the ion beams and the pilot proton beams. The low gain (LOGAIN) is suitable for the measurement of higher intensity proton beams. The proposed theoretical per-bunch FS ranges for both dynamic ranges are summarised in Tab. 6.2.

6.2 Determination of the transformer bandwidth

The goal of this chapter is to find an answer to the question: “What measurement error is caused by the droop of the transformer”. The worst case error estimation for the LHC ring FBCTs is obtained when measuring the injection of four SPS batches. In case of the LHC dump it is the measurement of the beam being extracted from a fully loaded LHC ring. In order to model these two cases the relation between the FBCT’s LF cut-off and the measurement error must be expressed. The LF cut-off is limited by the periodicity of the bunch population, hence the worst-case errors are generated by the proton injection and extraction scenarios. This is understood as injection of *4 batches of 72 bunches spaced by 25 ns* into the LHC, or extraction of *2808 bunches spaced by 25 ns* to the beam dumps.

The measurement error can be referred to using two forms:

- An absolute error, denoted using the capital letter “E”, defined as a difference of the measured (or inferred) value and the true value. It is expressed in number of charges, either in a “ $\pm$ ” notation, or as a RMS value.
- A relative error, defined as an absolute error divided by the true value. The result is unit-less and can be expressed as a percentage error. The relative or percentage error is denoted using the lowercase letter “e”.

The use of either form depends on the relevance of the given error value to the measurement being undertaken.

6.2.1 The LHC pilot bunch

The orbiting LHC pilot bunch represents the simplest LHC operational mode. A correctly constructed mathematical model of the measurement error for this case can be re-used to describe other, more advanced filling scenarios.

The beam signal can be modelled using Eq. (6.1). The FBCT can be thought of as a second order band pass filter. In the case of wide-band FBCTs the measurement error due to the device's HF cut-off can be neglected. Hence an even simpler model - a first order high pass filter - can be used. The measurement signal can be modelled by letting the beam signal pass the filter. The difficulty of this method lies in the analytic solution of the resulting equations. It seems that by using a convolution integral the problem can be partially circumvented.

The causal response of the FBCT allows the use of a finite form of the convolution integral:

$$y(t) = \int_0^t f(t - \tau)h(\tau)d\tau \quad (6.11)$$

The function $h(t)$ is the impulse response of the FBCT, which is calculated from the filter's transfer function:

$$h(s) = \frac{\epsilon s}{\epsilon s + 1} \Rightarrow h(t) = \mathcal{L}^{-1}\{h(s)\} \Rightarrow h(t) = \delta(t) - \frac{1}{\epsilon}e^{-\frac{t}{\epsilon}} \quad (6.12)$$

where $\mathcal{L}^{-1}$ denotes the inverse Laplace operator and $\delta(t)$ the Dirac delta operator. The constant ϵ is the time constant of the LF cut-off of the device:

$$\epsilon = \frac{1}{\omega_c} = \frac{1}{2\pi f_c} \quad (6.13)$$

where ω_c is the cut-off angular frequency.

The resulting measurement signal is obtained injecting Eq. (6.12) into Eq. (6.11):

$$y(t) = \int_0^t f(t - \tau) \left[\delta(\tau) - \frac{1}{\epsilon}e^{-\frac{\tau}{\epsilon}} \right] d\tau \quad (6.14)$$

As the repetition period of the pilot bunch is much longer than its FWHM, the signal's periodicity can be neglected and Eq. (6.1) can be thought of as if it was not periodic, and written as:

$$f(t) = \begin{cases} a[\cos(bt)^2 - 1]^2 & t \in \langle 0; \frac{\pi}{b} \rangle \\ 0 & otherwise \end{cases} \quad (6.15)$$

The situation is depicted in Fig. 6.3. The blue trace depicts the filter's impulse response (Eq. (6.12)). The Dirac pulse is shown as a thin peak at $t = 0$. The red trace represents the base function defined by Eq. (6.15). The base function is already mirrored and shifted due to the substitution performed in Eq. (6.11). The product of the functions' surfaces defines the desired impulse response. As the base function is bounded in time it is sufficient to perform the calculation for non-zero parts of the product.

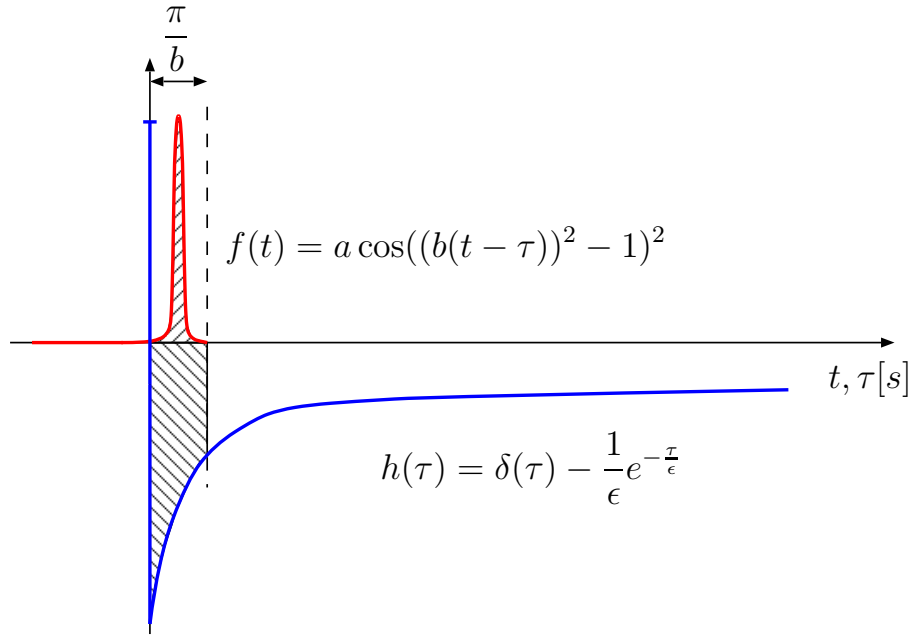


Figure 6.3: Graphical representation of the convolution equation (6.11)

The impulse response for $t < \frac{\pi}{b}$ involves a multiplication of the Dirac pulse with the base function, which is not the case for $t \geq \frac{\pi}{b}$. Hence the calculus must be split into two parts.

The convolution integral defined by boundary condition $t < \frac{\pi}{b}$ can be expressed as:

$$y_{(t \leq \frac{\pi}{b})}(t) = \int_0^t a[\cos(b(t - \tau))^2 - 1]^2 \left[\delta(\tau) - \frac{1}{\epsilon} e^{-\frac{\tau}{\epsilon}} \right] d\tau \quad (6.16)$$

The properties of the Dirac pulse permit Eq. (6.16) to be rewritten in the terms of the base function (Eq.(6.15)):

$$y_{(t \leq \frac{\pi}{b})}(t) = f(t) - \frac{1}{\epsilon} \int_0^t a[\cos(b(t - \tau))^2 - 1]^2 e^{-\frac{\tau}{\epsilon}} d\tau \quad (6.17)$$

where the following substitution for the integral part can be used:

$$-\frac{1}{\epsilon} \int_0^t a[\cos(b(t - \tau))^2 - 1]^2 e^{-\frac{\tau}{\epsilon}} d\tau = \frac{a}{8\epsilon} \left(\frac{8b \sin(2bt) + \frac{4}{\epsilon} (\cos(2bt) - e^{-\frac{t}{\epsilon}})}{\frac{1}{\epsilon^2} + 4b^2} - \frac{4b \sin(4bt) + \frac{1}{\epsilon} (\cos(4bt) - e^{-\frac{t}{\epsilon}})}{\frac{1}{\epsilon^2} + 16b^2} + 3\epsilon (e^{-\frac{t}{\epsilon}} - 1) \right) \quad (6.18)$$

For $t > \frac{\pi}{b}$ the convolution is formed using the exponential part of the impulse response:

$$y_{(t > \frac{\pi}{b})}(t) = -\frac{a}{\epsilon} \int_{t-\frac{\pi}{b}}^t a[\cos(b(t - \tau))^2 - 1]^2 e^{-\frac{\tau}{\epsilon}} d\tau \quad (6.19)$$

The lower boundary of the integral is defined by the fact that the base function is zero outside of the interval $0 \leq t < \frac{\pi}{b}$. The response of the measurement device to the pilot

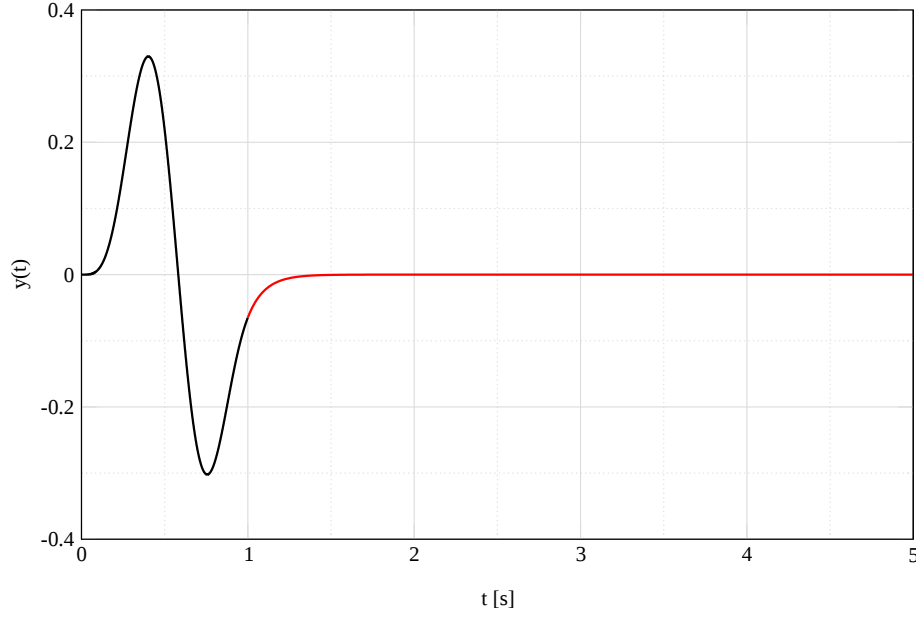


Figure 6.4: Convolution of a single bunch input signal with an impulse response of a high pass filter. The black trace shows the result of Eq. (6.17) and the red trace of Eq. (6.20). The following parameters were used: $a = 1 \text{ s}^{-1}$, $b = \pi \text{ s}^{-1}$ and $\epsilon = 0.1 \text{ s}$

beam for the case $t > \frac{\pi}{b}$ can be expressed by solving Eq. (6.19):

$$y_{(t > \frac{\pi}{b})}(t) = -\frac{24ab^4\epsilon^4 (e^{\frac{\pi}{b\epsilon}} - 1) e^{-\frac{t}{\epsilon}}}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} \quad (6.20)$$

The final equation is given by the superposition of Eqns. (6.4) and (6.17). The plot of the measurement signal is depicted in Fig. 6.4.

a) The error caused by the transformer droop

The convolution can be thought of as a black box fed by a bunch-like signal as shown in Fig. 6.5. In bunch by bunch measurements the absolute error for a single bunch slot is given as the difference between the surface of the output and the original signal within that slot.

The convolution process gives a response lasting longer than the input signal. This results in the presence of the output signal in following bunch slots, which should be empty according to the input signal. In such cases the relative error is not defined and hence only an absolute error can be expressed.

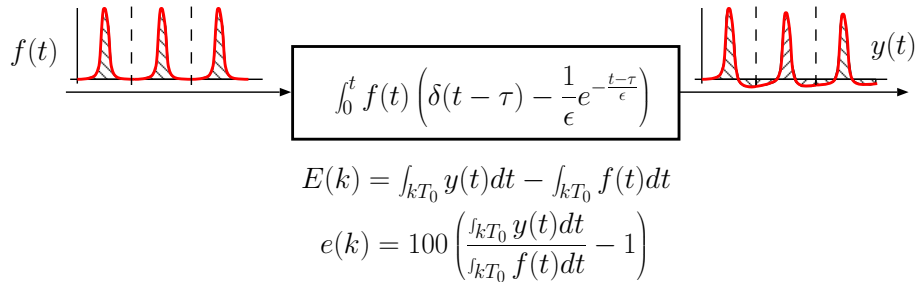


Figure 6.5: The absolute error for each bunch slot is calculated as a difference between its surface at the output and the input of the convolution.

Let us assume the time constant ϵ such that the filter's response fits in a single bunch slot T_0 . The surface defined by this curve for the first slot can be expressed as a sum of the surfaces defined by Eqns. (6.17) and (6.20):

$$s_{p,0} = \int_0^{\frac{\pi}{b}} y_{(t \leq \frac{\pi}{b})}(t) dt + \int_{\frac{\pi}{b}}^{T_0} y_{(t > \frac{\pi}{b})}(t) dt \quad (6.21)$$

Evaluation of Eq. (6.21) determines the “charge” contained in the first bunch slot:

$$s_{p,0} = \frac{24a\epsilon^5 b^4 \left(e^{\frac{\pi}{b\epsilon}} - 1\right) e^{-\frac{T_0}{\epsilon}}}{64b^4 \epsilon^4 + 20b^2 \epsilon^2 + 1} \quad (6.22)$$

The surface of the excitation signal is defined by Eq. (6.4). The absolute error, given by the simple difference of the surfaces, is normalised to unity:

$$E_{p,0} = s_{p,0} - s_f = \frac{24a\epsilon^5 b^4 \left(e^{\frac{\pi}{b\epsilon}} - 1\right) e^{-\frac{T_0}{\epsilon}}}{64b^4 \epsilon^4 + 20b^2 \epsilon^2 + 1} - 1 \quad (6.23)$$

The scaling factor $N_p \cdot e$ must be applied to express $E_{p,0}$ in number of charges.

The relative error for the first bunch slot can be expressed as:

$$e_{p,0} = \frac{s_{p,0}}{s_f} - 1 = \frac{64b^5 \epsilon^5 \left(e^{\frac{\pi}{b\epsilon}} - 1\right) e^{-\frac{T_0}{\epsilon}}}{\pi (64b^4 \epsilon^4 + 20b^2 \epsilon^2 + 1)} - 1 \quad (6.24)$$

For $t > T_0$ no excitation exists, however the output signal is non-zero due to the convolution process. The measured output phantom “charge” for the k^{th} bunch slot can be calculated as:

$$s_{p,k} = \int_{kT_0}^{(k+1)T_0} y_{(t > \frac{\pi}{b})}(t) dt = -\frac{24\epsilon^5 b^4 a \left(e^{\frac{\pi}{b\epsilon}} - 1\right) \left(e^{-\frac{kT_0}{\epsilon}} - e^{-\frac{(k+1)T_0}{\epsilon}}\right)}{64b^4 \epsilon^4 + 20b^2 \epsilon^2 + 1} \quad (6.25)$$

The relative error is not defined, and the absolute error simplifies to:

$$E_{p,k} = s_{p,k} \quad (6.26)$$

b) Evaluation of the error for the LHC pilot bunch

Figure 6.6 shows the dependence of the relative error on a chosen LF cut-off of the FBCT. The data were calculated for the LHC pilot bunch present in the first bunch slot. An error of 1 % is achieved using an FBCT with LF cut-off of ≈ 100 kHz. As the intention is to use a toroid with much lower cut-off, the influence of the droop on the pilot bunch is negligible. Figure 6.7 shows the values of the absolute error in specific bunch slots. An interesting fact is demonstrated: by increasing the cut-off frequency the absolute error in the first few bunch slots increases, but in all other slots drops quickly (red and orange traces). This is caused by a filtering effect. A beam-like signal passing through a filter with a high cut-off frequency converts, in the extreme case, to the shape depicted in Fig. 6.4. The integrated value converges to zero already within the first slot, resulting in 100 % relative error. As

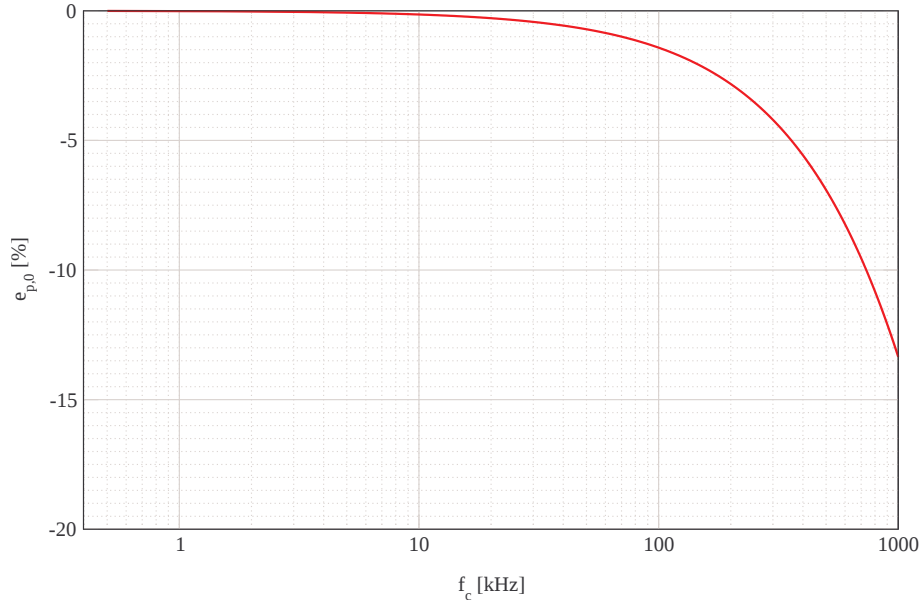


Figure 6.6: The plot of the relative error caused by the droop of the transformer for the pilot bunch. Negative value means that the charge of the base function signal defined for first T_0 interval at the output of the transformer is lower compared to the nominal number of charges.

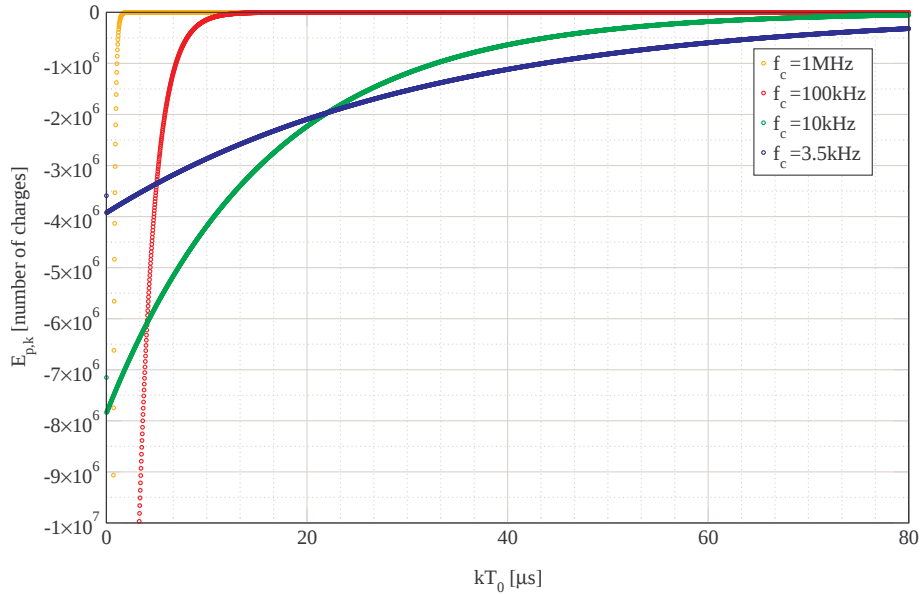


Figure 6.7: Absolute error calculated for the LHC pilot bunch using different FBCT cut-off frequencies.

the signal is zero in all other bunch slots, the absolute error drops quickly. Lowering the cut-off frequency decreases the absolute error in the first slot, however the absolute error, or droop, for all other bunch slots deteriorates. This is the valid working condition for the FBCTs.

6.2.2 Injection of multiple batches into the LHC

The basis calculation from the previous section can be extended to model a more complex injection. Up to 4 batches of 72 bunches spaced by 25 ns can be injected into the LHC from the SPS. A gap of ≈ 225 ns separates the batches. To simplify the calculation the gap will not be taken into account. This introduces a small error, but as the calculation is incorrect in a “good” sense (the calculated results exhibit a worse error than in reality) it will set an upper bound on the error expected.

a) Mathematical background

The convolution method is used again, but this time it is the impulse response which is mirrored and shifted:

$$y(t) = \int_0^t f(\tau) h(t - \tau) d\tau \quad (6.27)$$

where $f(t)$ is the base function (Eq. (6.2)) periodic on the interval T_0 and $h(t)$ is the impulse response as defined in Eq. (6.12). The process, shown in Fig. 6.8, can be described by:

$$y(t) = \int_0^t f(\tau) \left[\delta(t - \tau) - \frac{1}{\epsilon} e^{-\frac{t-\tau}{\epsilon}} \right] d\tau \quad (6.28)$$

The blue trace depicts the filter's impulse response and the red trace the input signal. The situation gets complicated compared to the previous calculation because the periodicity T_0 must be taken into account. Should the time t lie on the interval $\langle kT_0 + \frac{\pi}{b}; (k+1)T_0 \rangle$, the Dirac pulse term in the calculation vanishes due to the zero value of the base function for this interval. Additional complications arise due to calculations on the interval $\langle \zeta(t); t \rangle$, which contains only a fraction of the bunch.

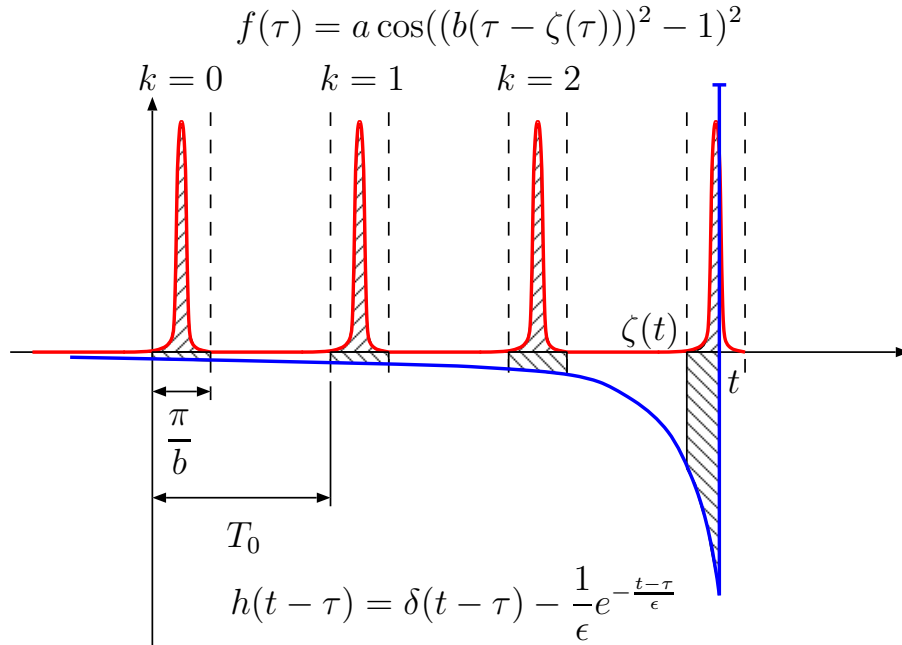


Figure 6.8: Graphical solution of the convolution integral with shifted impulse response

Equation (6.28) can be rewritten as:

$$y(t) = \int_0^t f(\tau) \delta(t - \tau) d\tau - \frac{1}{\epsilon} \int_0^t f(\tau) e^{-\frac{t-\tau}{\epsilon}} d\tau \quad (6.29)$$

The left term of the equation can be simplified using the properties of the Dirac pulse:

$$y(t) = f(t) - \frac{1}{\epsilon} \int_0^t f(\tau) e^{-\frac{t-\tau}{\epsilon}} d\tau = f(t) + g(t) \quad (6.30)$$

And the right must be evaluated separately:

$$g(t) = g_1(t) + g_2(t) = -\frac{1}{\epsilon} \int_0^{\zeta(t)} f(\tau) e^{-\frac{t-\tau}{\epsilon}} d\tau - \frac{1}{\epsilon} \int_{\zeta(t)}^t f(\tau) e^{-\frac{t-\tau}{\epsilon}} d\tau \quad (6.31)$$

The first term of Eq. (6.31) is graphically denoted in Fig. 6.8 as the hashed region of the bunches labeled $k = [0, 1, 2]$, and associated fractions of the impulse response function. The second term, depicted by the hashed region between $\zeta(t)$ and t , evaluates a non-periodic interval between the last multiple of period T_0 and the time t of the calculation.

The base function is defined by regions. Closely inspecting the interval of the integration of the function $g_2(t)$ it can be concluded that $g_2(t) = 0$ on the interval $t - \zeta(t) > \frac{\pi}{b}$. Hence Eq. (6.1) can be substituted into the right integral of Eq. (6.31) as follows:

$$g_2(t) = \begin{cases} -\frac{a}{\epsilon} \int_{\zeta(t)}^t (\cos(b(\tau - \zeta(\tau)))^2 - 1)^2 e^{-\frac{t-\tau}{\epsilon}} d\tau & 0 \leq t - \zeta(t) < \frac{\pi}{b} \\ 0 & \text{otherwise} \end{cases} \quad (6.32)$$

The variable $\zeta(\tau)$ is *always constant* on the interval of integration, and its value is always equal to the value of the lower integration boundary $\zeta(t)$. Hence the integral can be evaluated as:

$$\begin{aligned} -\frac{a}{\epsilon} \int_{\zeta(t)}^t (\cos(b(\tau - \zeta(\tau)))^2 - 1)^2 e^{-\frac{t-\tau}{\epsilon}} d\tau = \\ \frac{a}{\epsilon} \left[\frac{\cos(2b[\zeta(t) - t]) - 2b\epsilon \sin(2b[\zeta(t) - t]) - e^{\frac{\zeta(t)-t}{\epsilon}}}{2\epsilon \left(\frac{1}{\epsilon^2} + 4b^2\right)} + \right. \\ \left. - \frac{\cos(4b[\zeta(t) - t]) - 4b\epsilon \sin(4b[\zeta(t) - t]) - e^{\frac{\zeta(t)-t}{\epsilon}}}{8\epsilon \left(\frac{1}{\epsilon^2} + 16b^2\right)} + \right. \\ \left. - \frac{3}{8}\epsilon \left(1 - e^{-\frac{t-\zeta(t)}{\epsilon}}\right) \right] \end{aligned} \quad (6.33)$$

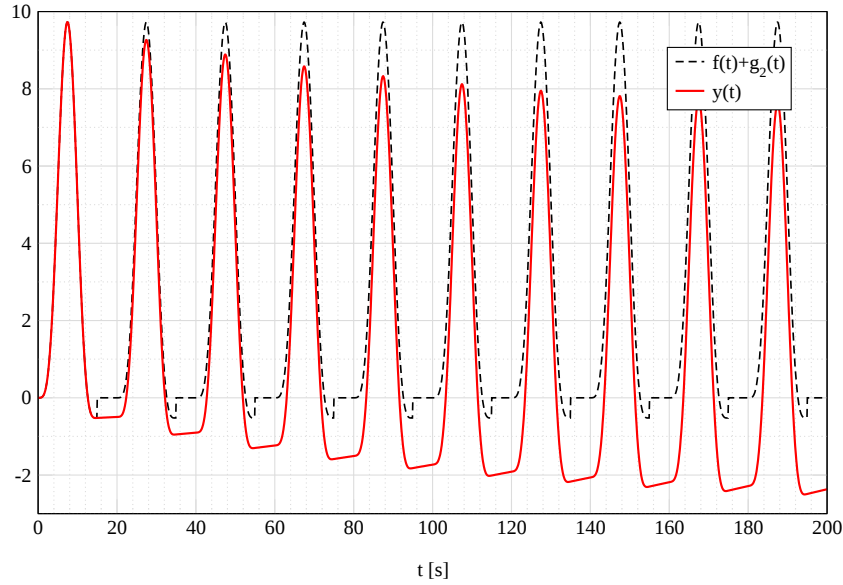


Figure 6.9: Convolution of multiple bunches with an impulse response of a first order high pass filter. The red trace shows the result calculated using equation (6.29). The black trace depicts the results of the $f(t) + g_2(t)$. The parameters used are arbitrarily chosen to highlight the effect of filtering: $T_0 = 20$ s, $b = \frac{\pi}{15}$ s⁻¹, $a = 10$ s⁻¹ and $\epsilon = 100$ s.

Such piecewise definition of the base function also applies for the $g_1(t)$ term of Eq. (6.31). $g_1(t)$ can be derived by inspecting the intervals at which the base function is not zero:

$$g_1(t, N) = -\frac{1}{\epsilon} \int_0^{\zeta(t)} f(\tau) e^{-\frac{t-\tau}{\epsilon}} d\tau = -\frac{1}{\epsilon} \sum_{\substack{k=0 \\ N \geq 0}}^N \int_{kT_0}^{kT_0 + \frac{\pi}{b}} a (\cos(b(\tau - kT_0))^2 - 1)^2 e^{-\frac{t-\tau}{\epsilon}} d\tau \quad (6.34)$$

where N is the number of full bunches appearing in the integration interval. The integral term of Eq. (6.34) can be substituted by:

$$\int_{kT_0}^{kT_0 + \frac{\pi}{b}} a (\cos(b(\tau - kT_0))^2 - 1)^2 e^{-\frac{t-\tau}{\epsilon}} d\tau = \frac{24b^4\epsilon^5 a (e^{\frac{\pi}{b\epsilon}} - 1)}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} e^{-\frac{t-kT_0}{\epsilon}} \quad (6.35)$$

and the sum of the exponential in Eq. (6.34) by:

$$\sum_{k=0}^N e^{-\frac{k-kT_0}{\epsilon}} = \frac{e^{\frac{T_0(N+1)-t}{\epsilon}} - e^{-\frac{t}{\epsilon}}}{e^{\frac{T_0}{\epsilon}} - 1} \quad (6.36)$$

Equation (6.34) can be rewritten using (6.36) and Eq. (6.35) as:

$$g_1(t, N) = -\frac{24b^4\epsilon^4 a (e^{\frac{\pi}{b\epsilon}} - 1) \left(e^{\frac{T_0(N+1)-t}{\epsilon}} - e^{-\frac{t}{\epsilon}} \right)}{(64b^4\epsilon^4 + 20b^2\epsilon^2 + 1) \left(e^{\frac{T_0}{\epsilon}} - 1 \right)} \quad (6.37)$$

The solution of Eq. (6.37) is valid on the interval $t \geq \frac{\pi}{b}$. This is denoted in Eq. (6.34) using condition $N \geq 0$. The number of periods N can be determined as:

$$N = -\text{round} \left(\frac{\pi - tb}{bT_0} + \frac{1}{2} \right) \quad (6.38)$$

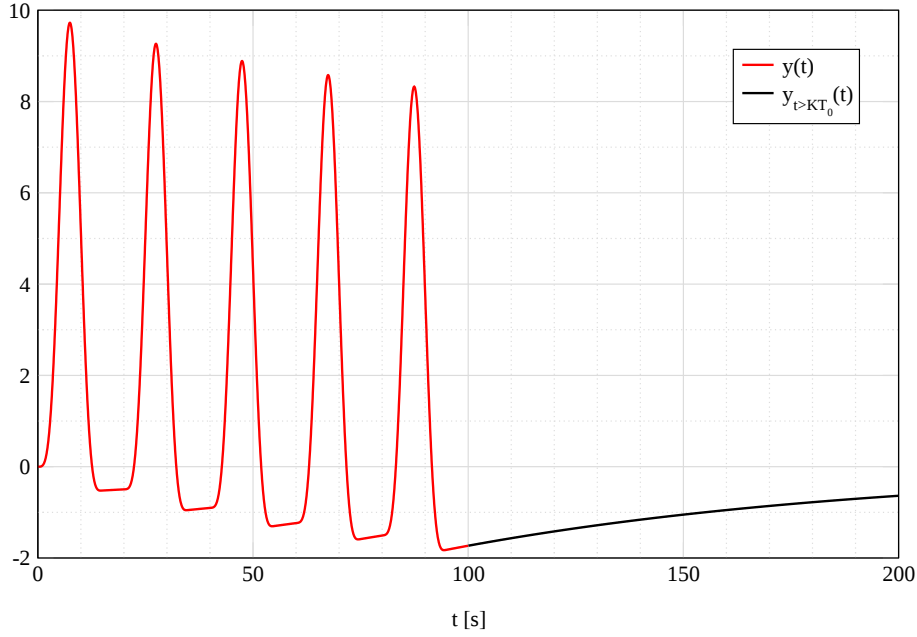


Figure 6.10: The result of the convolution for $K = 5$. The same parameters as used to produce Fig. 6.9 were used. The black trace shows the result of Eq. (6.42).

The described calculation is valid only for $b < \frac{\pi}{T_0}$. This is due to the assumption the base function is periodic on the interval T_0 . Figure 6.9 shows the result of the convolution calculation using arbitrarily chosen parameters.

b) Exact number of bunches

The previous calculations are based on the idea of injecting an infinitely long train of equidistantly spaced bunches. This does not permit to evaluate the error at the end of an inserted batch, hence the calculus must still be improved. A nice feature of Eq. (6.37) is that it evaluates the term of the convolution corresponding to full periods of the base function. Thus the injection of a specific number of bunches is equivalent to the following set of substitutions:

$$f(t) = 0 \quad (6.39)$$

$$g_2(t) = 0 \quad (6.40)$$

$$g_1(t, N) = g_1(t, K - 1) \quad (6.41)$$

where K is the number of injected bunches. The “tail” of the injection is defined as:

$$y_{(t \geq KT_0)}(t) = -\frac{24b^4\epsilon^4 a \left(e^{\frac{\pi}{b\epsilon}} - 1 \right) \left(e^{\frac{KT_0-t}{\epsilon}} - e^{-\frac{t}{\epsilon}} \right)}{(64b^4\epsilon^4 + 20b^2\epsilon^2 + 1) \left(e^{\frac{T_0}{\epsilon}} - 1 \right)} \quad (6.42)$$

Figure 6.10 depicts the function shown in Fig. 6.9 for the injection of five bunches.

c) The error caused by the transformer droop

The result of the convolution is split into three terms: $g_1(t, N)$, $g_2(t)$ and the function $f(t)$ itself. While the functions $f(t)$ and $g_2(t)$ are defined on the periodic interval

$0 \leq t - \zeta(t) < \frac{\pi}{b}$, the function $g_1(t, N)$ is valid on the interval $t > \frac{\pi}{b}$ and is not periodic. From Fig. 6.9 it can be concluded that the surface given by the sum $f(t) + g_2(t)$ is a constant for each bunch slot.

The corresponding base function surface was defined in Eq. (6.4). The area of $g_2(t)$ can be expressed by substituting $\zeta(t) = 0$ into Eq. (6.32) and reusing the solution of the integral from Eq. (6.33):

$$s_{g_2} = \int_0^{\frac{\pi}{b}} g_2(t) dt = -\frac{3a}{8b} \cdot \frac{64b^5\epsilon^5 (e^{-\frac{\pi}{b\epsilon}} - 1) + 64\pi b^4\epsilon^4 + 20\pi b^2\epsilon^2 + \pi}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} \quad (6.43)$$

Summing up equations (6.4) and (6.43) the constant term of the surface for all the bunch slots is obtained.

The calculation of the surface generated by the function $g_1(t, N)$ is the last piece of the puzzle. Due to the various intervals involved in the definition the final calculation is composed of four terms. The function parameter N , defined in Eq. (6.38), splits the solution of Eq. (6.37) into two terms due to the step located at $t = kT_0 + \frac{\pi}{b}$. Additionally, an equation for the first bunch slot must be solved separately due to the constraint $t > \frac{\pi}{b}$ put on Eq. (6.37). It also splits into two terms due to the rounding at $t = \frac{\pi}{b}$. However, as Eq. (6.37) is zero for the interval $t < \frac{\pi}{b}$, only the solution in the interval $t \in \langle \frac{\pi}{b}; T_0 \rangle$ must be found.

Substituting $N = 0$ into Eq. (6.37) and integrating it with respect to t for the interval $\langle \frac{\pi}{b}; T_0 \rangle$ the surface of the function $g_1(t, N)$ for the first bunch slot is calculated as:

$$s_{g_{1,0}} = \int_{\frac{\pi}{b}}^{T_0} g_1(t, 0) dt = -\frac{24\epsilon^5 b^4 a (e^{\frac{\pi}{b\epsilon}} - 1) (e^{\frac{T_0}{\epsilon}} - e^{\frac{\pi}{b\epsilon}}) e^{-\frac{T_0 b + \pi}{b\epsilon}}}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} \quad (6.44)$$

Summing up Eqns. (6.44), (6.4) and (6.43) the total surface for the first bunch slot can be calculated, and hence its absolute error defined as:

$$E_{c,0} = s_{g_2} + s_{g_{1,0}} = -\frac{3a}{8b} \left(\frac{64b^5\epsilon^5 (e^{-\frac{T_0}{\epsilon}} - e^{-\frac{T_0 b - \pi}{b\epsilon}}) + \pi}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} \right) \quad (6.45)$$

and the relative error as:

$$e_{c,0} = \frac{E_{c,0}}{s_f} = -\frac{64b^5\epsilon^5 (e^{-\frac{T_0}{\epsilon}} - e^{-\frac{T_0 b - \pi}{b\epsilon}}) + \pi}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} \quad (6.46)$$

Assume $k = 1$ for the bunch slot starting at $t = T_0$. Quick inspection of the rounding function (6.38) reveals that N changes in the interval $t \in \langle T_0; 2T_0 \rangle$ from -1 to 0 at $t = T_0 + \frac{\pi}{b}$. Matching the variables N and k allows the following substitution to be proposed:

$$s_{g_{1,k}} = \int_{kT_0}^{kT_0 + \frac{\pi}{b}} g_1(t, k-1) dt + \int_{kT_0 + \frac{\pi}{b}}^{(k+1)T_0} g_1(t, k) dt \quad (6.47)$$

which can be re-written as:

$$s_{g_{1,k}} = -24b^4\epsilon^5 a \cdot \frac{e^{-\frac{\pi}{b\epsilon}} + e^{-\frac{kT_0}{\epsilon}} - e^{-\frac{bkT_0-\pi}{b\epsilon}} - e^{\frac{T_0b-\pi}{b\epsilon}} + e^{\frac{T_0}{\epsilon}} - e^{-\frac{(k+1)T_0}{\epsilon}} + e^{-\frac{bT_0(k+1)-\pi}{b\epsilon}} - 1}{64b^4\epsilon^4 \left(e^{\frac{T_0}{\epsilon}} - 1\right) + 20b^2\epsilon^2 \left(e^{\frac{T_0}{\epsilon}} - 1\right) + e^{\frac{T_0}{\epsilon}} - 1} \quad (6.48)$$

Hence the absolute and relative errors for a bunch slot k can be defined as:

$$E_{c,k}(k) = s_{g_2} + s_{g_{1,k}} = -\frac{3a}{8b} \left(\frac{64b^5\epsilon^5 \left(e^{\frac{\pi-bT_0(k+1)}{b\epsilon}} + e^{-\frac{kT_0}{\epsilon}} - e^{\frac{\pi-bkT_0}{b\epsilon}} - e^{-\frac{(k+1)T_0}{\epsilon}} \right)}{(64b^4\epsilon^4 + 20b^2\epsilon^2 + 1) \left(e^{\frac{T_0}{\epsilon}} - 1 \right)} + \pi \right) \quad (6.49)$$

$$e_{c,k}(k) = \frac{E_{c,k}(k)}{s_f} = -\frac{64b^5\epsilon^5 \left(e^{\frac{\pi-bT_0(k+1)}{b\epsilon}} + e^{-\frac{kT_0}{\epsilon}} - e^{\frac{\pi-bkT_0}{b\epsilon}} - e^{-\frac{(k+1)T_0}{\epsilon}} \right)}{\pi (64b^4\epsilon^4 + 20b^2\epsilon^2 + 1) \left(e^{\frac{T_0}{\epsilon}} - 1 \right)} - 1 \quad (6.50)$$

The absolute error for the empty bunch slots is equivalent to the surface under the exponential shown in Fig. 6.10 (black trace), and corresponds directly to the surface calculated for a specific bunch slot k :

$$E_e(k) = \int_{kT_0}^{(k+1)T_0} g_1(t, K-1) dt = \frac{24\epsilon^5 b^4 a \left(e^{\frac{KT_0}{\epsilon}} - e^{\frac{KbT_0+\pi}{b\epsilon}} + \frac{\pi}{b\epsilon} - 1 \right) e^{-\frac{(k+1)T_0}{\epsilon}}}{64b^4\epsilon^4 + 20b^2\epsilon^2 + 1} \quad (6.51)$$

where K is the total number of injected bunches and k is the bunch slot number for which the error is calculated. Equation (6.51) is valid for $k > K$. The relative error is not defined.

d) Evaluation of the error for a complex injection

Assume an empty LHC ring prepared for injection. It was shown in Sec. 4.4.2 that the 25 ns proton injection scenario starts with the injection of two SPS batches. Theoretically up to 4 batches could be however injected.

Every batch contains 72 bunches and 8 empty bunch slots. Assuming the empty slots are also filled with the bunches one can consider it as the injection of a continuous bunch train (320 bunches, equivalent to 8 μs). All other slots in the machine are empty ($\approx 81 \mu s$). This assumption introduces a small error into the calculation, however as the absolute error for this case is pessimistic with respect to the real injection scenario, it leads again to an upper bound of the error.

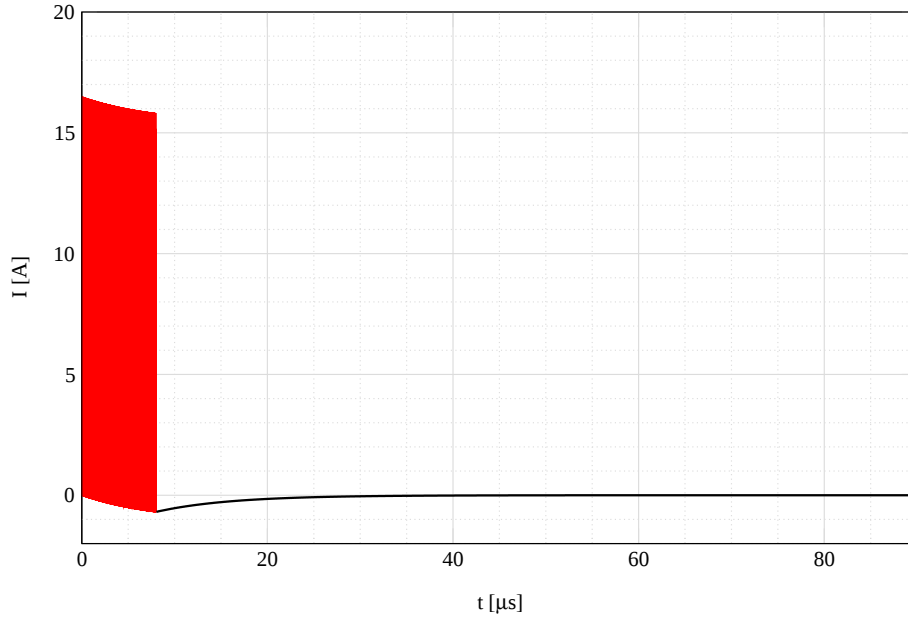


Figure 6.11: *The calculated droop of the FBCT. An arbitrary LF cut-off of 20 kHz was chosen to highlight the droop. The red trace shows the values calculated using Eq. (6.29). The black trace shows the decay at the end of the injection calculated using Eq. (6.42). Four SPS batches with 1.7×10^{11} charges per bunch are modelled (320 bunches).*

Figure 6.11 shows the calculated droop of the FBCT. The LF cut-off is set to 20 kHz to highlight the influence of the missing DC component in the signal. Each simulated bunch contains 1.7×10^{11} charges. This is calculated by substituting Eqns. (6.8) and (6.9) into Eq. (6.1), and multiplying by $N_p \cdot e$, where N_p is the number of charges in a single bunch. Using $\sigma = 680$ ps the vertical scale depicts the beam peak current at injection.

Figure 6.12 depicts the relative error for each injected bunch. It demonstrates that at the end of injection the signal droop of the FBCT with a LF cut-off of 20 kHz causes a relative error exceeding 60 %. The error is negative because the output charge is lower than the nominal one. This is in agreement with Fig. 6.11.

Figure 6.13 shows the absolute error of the signal depicted in Fig. 6.11. The maximum error occurs at the end of injection. Attention must be paid to the first empty slot after the end of injection. Its corresponding absolute error is calculated using Eq. (6.51). Increasing the cut-off frequency reduces the error in this slot with respect to the error calculated for the slot containing the last injected bunch. This is caused by the filtering effect mentioned in Sec. 6.2.1 b). Equation (6.49) calculated for the last bunch injected can be used to express the worst-case error for the injection. The result of this calculation is shown in Fig. 6.14.

e) Multiple batch injection - conclusion

The droop of the FBCT has a direct influence on the measurement error. In the case of injection the measured data cannot be electronically corrected. To comply with the LHC

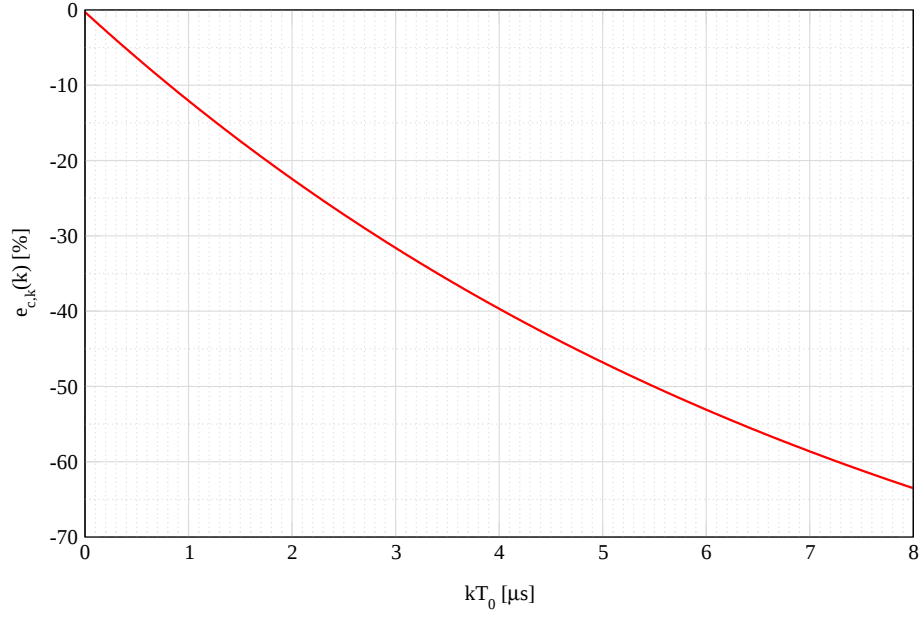


Figure 6.12: *The percentage error calculated for each bunch slot of the four batches injected into the LHC.*

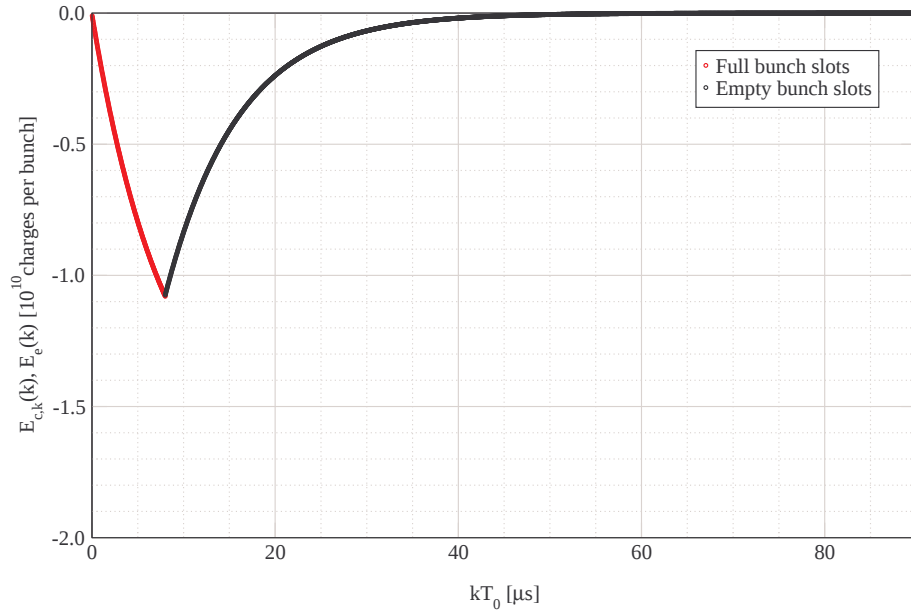


Figure 6.13: *Absolute error calculated for the injection of four batches of nominal intensity (1.7×10^{11} charges per bunch). The red trace shows the calculation for the case of bunches present in the bunch slots. The black trace shows the error caused by the exponential decay at the end of injection.*

FBCT specifications the error on a single-pass bunch by bunch measurement must not exceed 2 % ($\approx 2 \times 10^9$ charges) for a nominal beam and 10 % ($\approx 5 \times 10^8$ charges) for the pilot bunch. Hence an FBCT with droop lower than 2 % at the end of injection is required. Looking at Fig. 6.14 the LF cut-off of at most 400 Hz is needed. All the necessary plots for this case are depicted in Fig. 6.15.

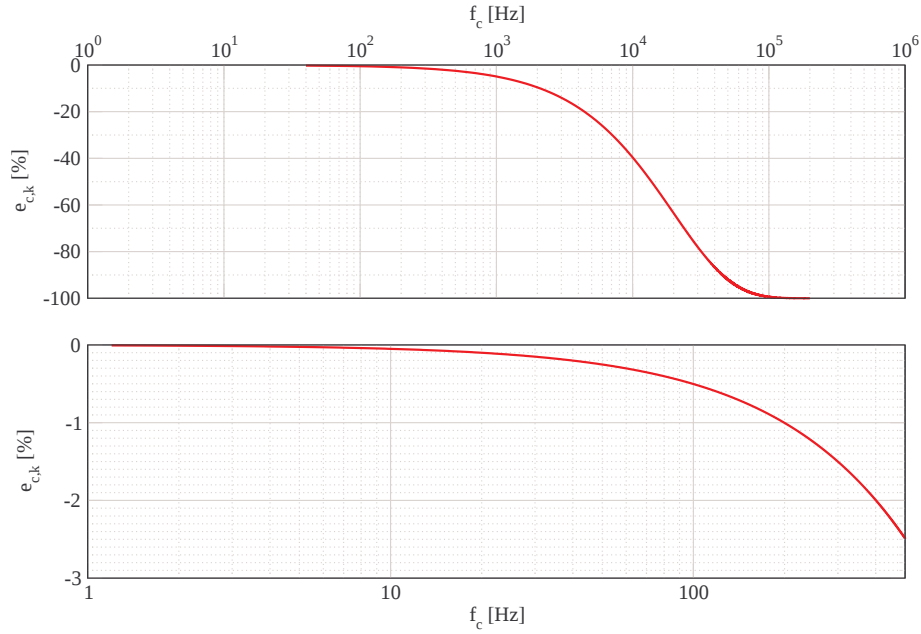


Figure 6.14: Percentage error calculated for the last bunch of four batches injected into the LHC as a function of cut-off frequency (lower plot is a zoom for $f_c < 500$ Hz). The parameters $\sigma = 680$ ps and $T_0 = 25$ ns were used.

f) Extraction in the dump lines

A single injection to the LHC lasts at most ≈ 8 μ s. In contrast the dump line measurement chain must deal with much longer times as in the extreme case the beam dump will empty a complete LHC ring. This is equal to an extraction of 2808 bunches or - if the mathematical simplifications are applied - 3564 bunches. The result of such calculation is shown in Fig. 6.16.

g) Dump line extraction - conclusion

The constraints put on the value of the LF cut-off are much tougher than in the case of multiple batch injection. Should the relative error be kept below 2 %, a FBCT LF cut-off of ≈ 40 Hz must be attained. Such a low cut-off increases the induced noise due to the first and second harmonic components of the mains power supply. It was therefore decided to acquire the droop information using a secondary toroid instead of lowering the LF cut-off of the primary one. In the future this will permit either electronic restoration of the DC component of the signal using the information from the secondary toroid or use of off-line algorithms to calculate the extracted beam intensity using only information from primary toroid. The latter is preferred as it eliminates the noise problems.

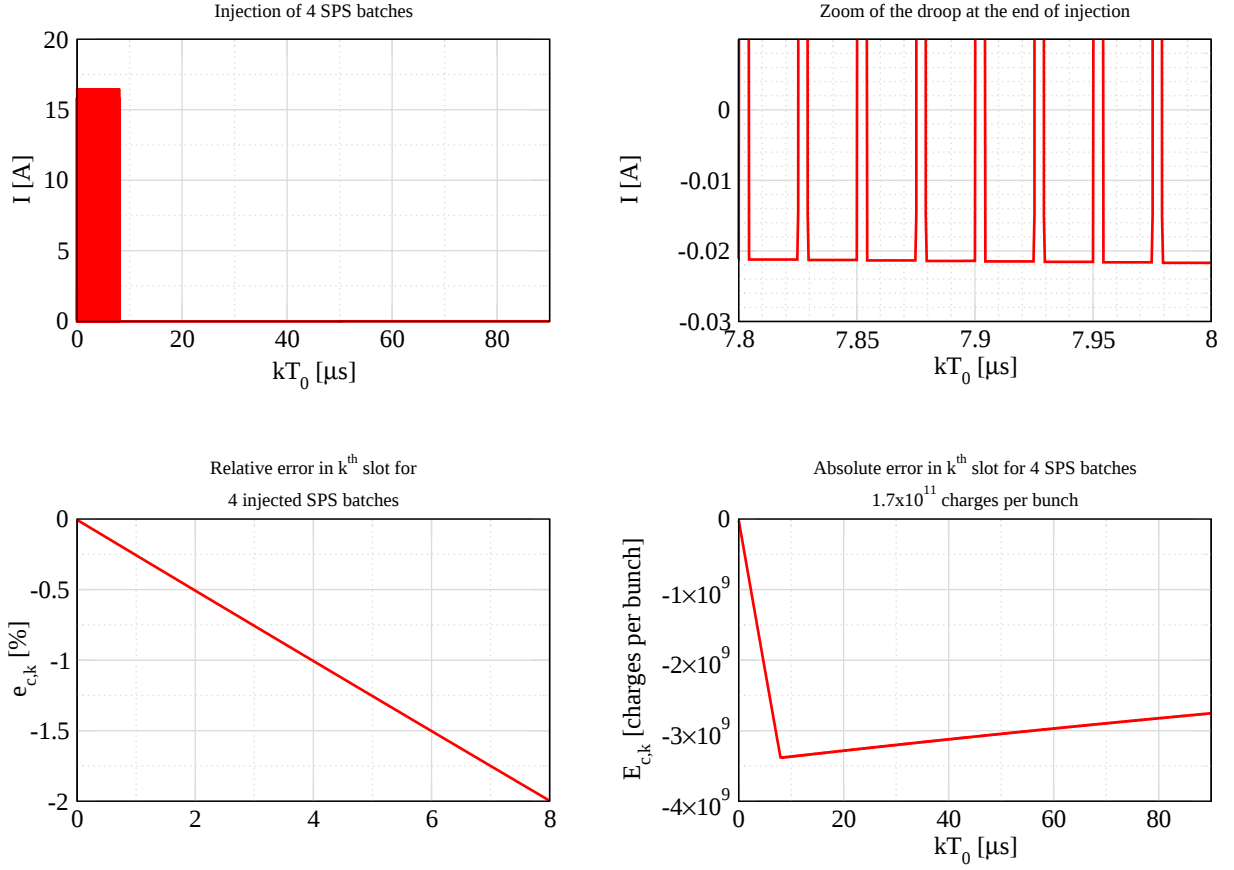


Figure 6.15: Complete error calculations for $f_c = 400$ Hz. The top-left plot shows the 320 bunches injected into the LHC. The plot top-right emphasises the droop of the last injected bunches. The magnitude of the droop at the end of injection is 21 mA compared to ≈ 16 A for the peak current. The bottom graphs depict the percentage and the absolute errors for each bunch slot k . The absolute error is calculated for ultimate beam (1.7×10^{11} charges per bunch).

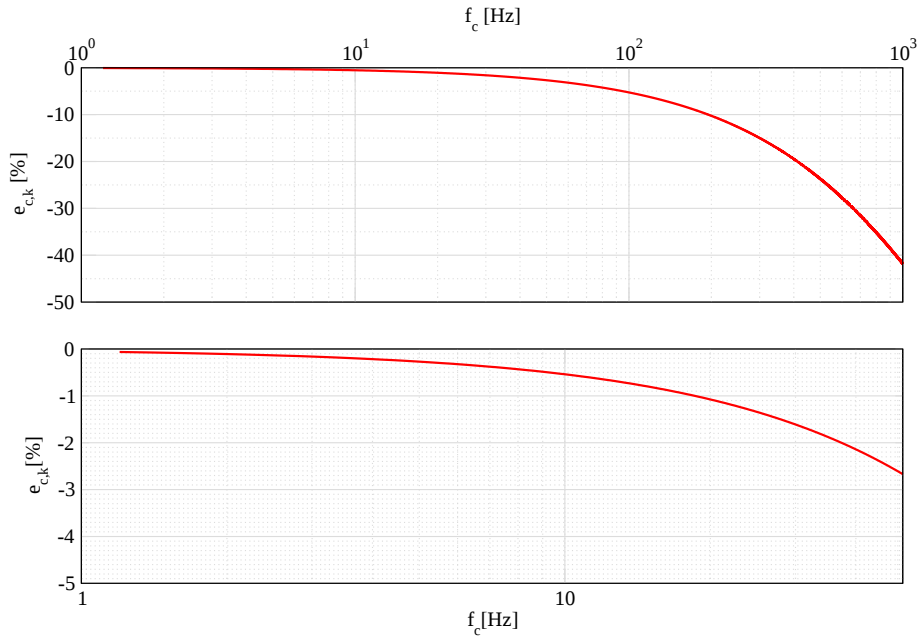


Figure 6.16: Relative error calculated for the last bunch of the beam ejected into the dump line. The bottom graph emphasises the error at low frequencies

6.2.3 The high frequency cut-off estimation

a) Full bandwidth requirements

The requirement for the HF cut-off of the system is determined by the need for bunch by bunch measurement. To measure correctly, the width of a bunch signal transmitted through the RF front-end to the integrator input should not exceed one bunch slot. Failing to satisfy this results in a leakage of the measured bunch into the subsequent bunch slots hence corrupting the bunch by bunch measurement. A security margin must be introduced due to a relative position of the bunch to the bunch slot. Other uncertainties should be also considered: a jitter of the Beam Synchronous Timing (BST) (300 ps_{rms}) and the time needed to reset the integrator ($2-3 \text{ ns}$). Lastly, the amplitude of the filtered bunch signal must decay to zero before the end of the bunch slot. Considering these constraints it was decided to keep the bunch width to less than one-half of a bunch slot.

The FBCT behaviour for high frequencies can be modelled by filtering the measured beam signal. The bunch function Eq. (6.1) is convolved with an impulse response of a first-order low-pass filter, defined by its transfer function in the Laplace domain:

$$g(s) = \frac{1}{1 + s\nu} \quad (6.52)$$

where ν is the time constant of the filter. The impulse response can be defined as:

$$g(t) = \frac{1}{\nu} e^{-\frac{t}{\nu}} \quad (6.53)$$

Equation (6.53) together with Eq. (6.1) (for $k = 0$) form the convolution integral defined in Eq. (6.11):

$$y(t) = \frac{a}{\nu} \int_0^t (\cos(b(t-\tau))^2 - 1)^2 \cdot e^{-\frac{\tau}{\nu}} d\tau \quad (6.54)$$

This situation is depicted in Fig. 6.17. The red trace is the bunch signal, mirrored and stripped to a single bunch. The blue trace denotes the impulse response of the filter.

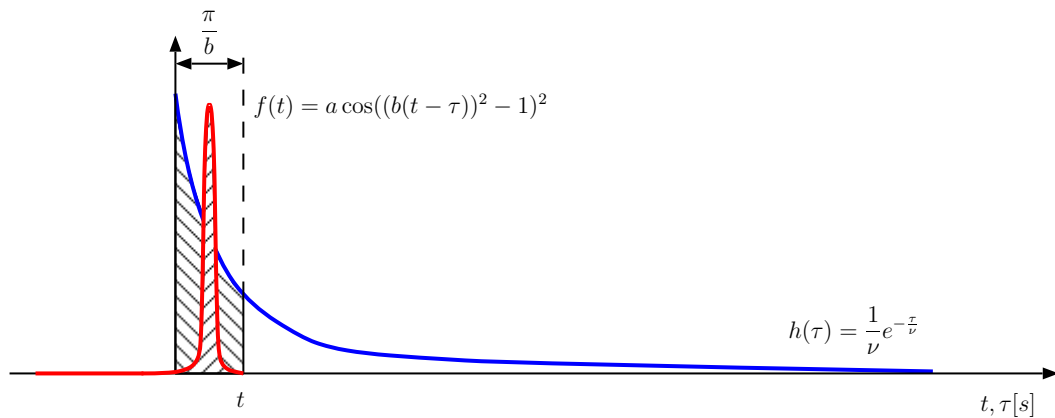


Figure 6.17: Graphical representation of the convolution Eq. (6.54)

The definition of the base function requires the convolution integral Eq. (6.54) to be split into two terms. The first term, valid for $t \leq \frac{\pi}{b}$, can be expressed as:

$$y_{t \leq \frac{\pi}{b}}(t) = \frac{ae^{-\frac{t}{\nu}}}{64b^4\nu^4 + 20b^2\nu^2 + 1} \cdot (\varsigma \cos(bt)^4 - 2\varsigma \cos(bt)^2 + \varsigma + 16\varsigma b^2\nu^2 + 4\varsigma b^2\nu^2 \cos(bt)^4 - 20\varsigma b^2\nu^2 \cos(bt)^2 + 4\varsigma b\nu \sin(bt) \cos(bt)^3 - 4\varsigma b\nu \sin(bt) \cos(bt) + 16\varsigma b^3\nu^3 \sin(bt) \cos(bt)^3 - 40\varsigma b^3\nu^3 \sin(bt) \cos(bt) + 24\varsigma b^4\nu^4 - 24b^4\nu^4) \quad (6.55)$$

where $\varsigma = \exp(\frac{t}{\nu})$ and the constants a and b are defined by Eqns. (6.8) and (6.9). The second term defines the response for $t > \frac{\pi}{b}$. Equation (6.54) transforms to:

$$y_{t > \frac{\pi}{b}}(t) = \frac{a}{\nu} \int_{t-\frac{\pi}{b}}^t (\cos(b(t-\tau))^2 - 1)^2 \cdot e^{-\frac{t}{\nu}} d\tau = \frac{24ab^4\nu^4 (e^{\frac{\pi}{b\nu}} - 1) e^{-\frac{t}{\nu}}}{64b^4\nu^4 + 20b^2\nu^2 + 1} \quad (6.56)$$

Substituting Eqns. (6.8), (6.9) and

$$\nu = \frac{1}{2\pi f} \quad (6.57)$$

into Eq. (6.56) a set of plots for specific HF cut-offs can be drawn, as shown in Fig. 6.18.

Keeping in mind the required criterion of a bunch width less than half a bunch slot, a cut-off frequency higher than $\approx 60 \text{ MHz}$ is seen to satisfy the condition. However the HF cut-off of the FBCT must be much higher due to additional filtering effect of the following signal transmission line. Attention must also be paid to provide excellent impedance matching of the signal path as any resulting ringing must not perturb the following bunch slots.

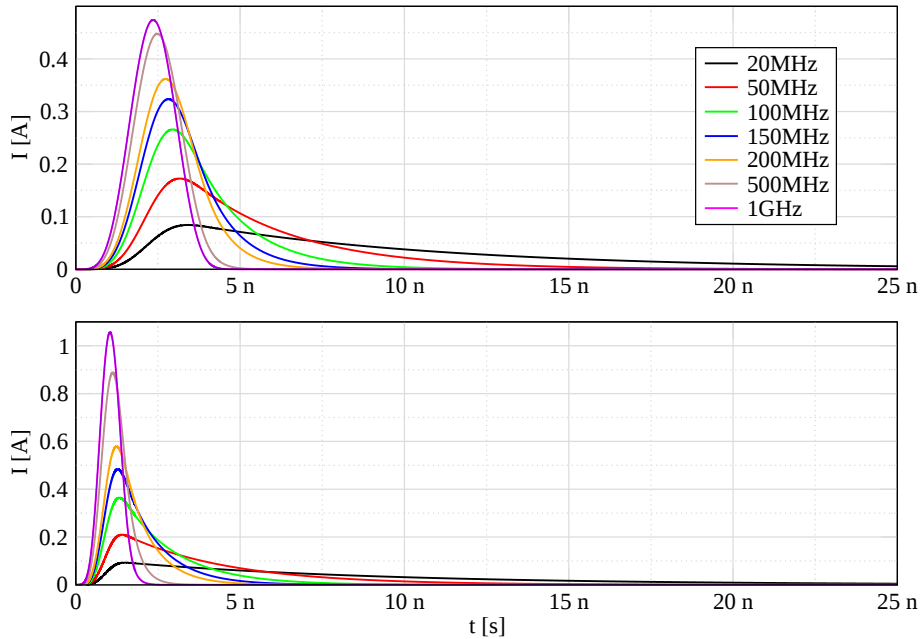


Figure 6.18: Filtering effect for the pilot bunch. The top and bottom graphs depict the bunch at injection and flat-top respectively.

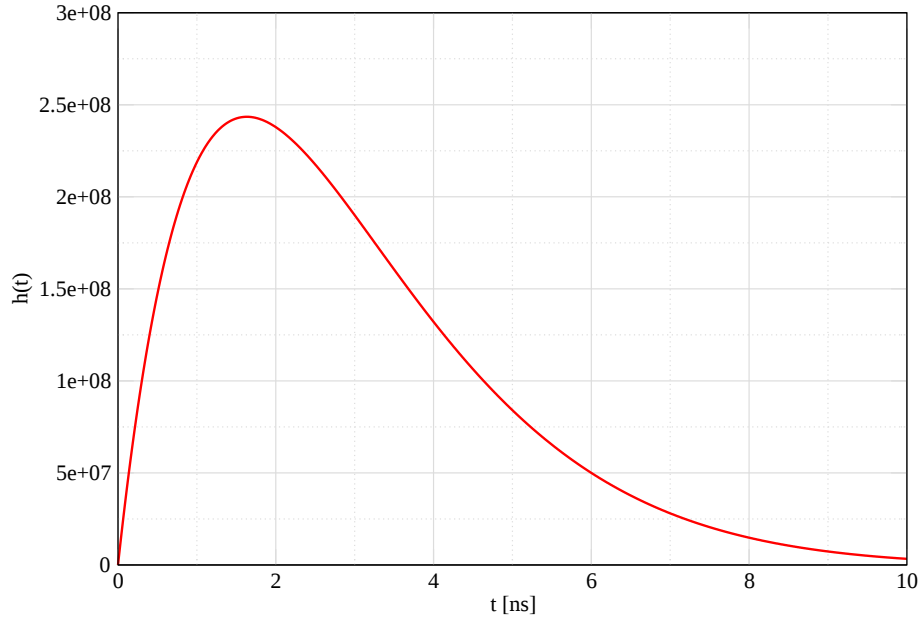


Figure 6.19: *Impulse response of a Gaussian filter with a cut-off frequency of 100 MHz*

b) Reduced bandwidth requirements

The bunch by bunch measurement can only work when a phase shift between the arriving bunch and the BST clock is adjusted to fit the bunch signal entirely into the bunch slot, taking into account the reduced bunch slot size due to the integrator reset. This is certainly not the case when the machine is being set-up and hence an additional measurement, which is time insensitive, must be available. This is provided by a turn by turn measurement, which gives the total intensity measured on a turn basis. The measurement principle is the same as for the bunch by bunch measurement, however the beam signal is low-pass filtered to integrate only the signal envelope. The rise and fall times of the filtered signal must allow DC signal restoration in the $3 \mu s$ long abort gap. Integrating the envelope of the signal minimises the errors caused by a mismatched timing as the integrator “sees” an average value of the beam signal, while all imperfections of the integrator are calibrated away.

The crucial question is what cut-off frequency the filter should have, and what type of filter response should be used. Filtering the signal to its envelope calls for a low frequency filter, certainly below 10 MHz . To ensure a ring free response to the input signal a Gaussian filter was selected.

The principle described in Sec. 6.2 can be used to calculate the cut-off frequency of the filter. The bunch-like signal function is convolved with the impulse response of the Gaussian filter. As higher order filters are difficult to solve analytically, a filter of second order is chosen, taking into account that higher order filters result in a lower output signal ripple.

The transfer function of a second-order low-pass filter can be written as:

$$h(s) = \frac{K \cdot f_c^2}{s^2 + Lf_c s + K} \quad (6.58)$$

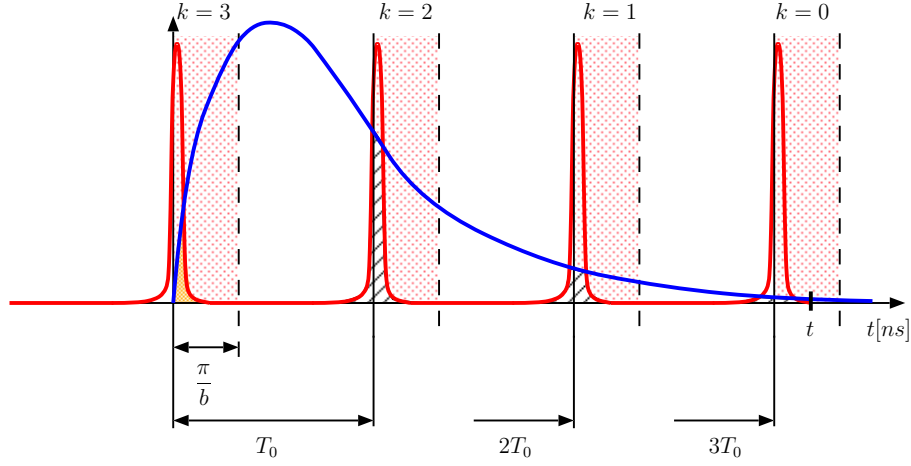


Figure 6.20: Convolution of a bunch-like signal with a Gaussian filter impulse response. The blue trace depicts the impulse response, the red trace the bunch-like signal.

where constants $K = 39.48$ and $L = 11.61$ define the Gaussian filter. An inverse Laplace transform of the transfer function using substituted constants K and L formalises the impulse response:

$$H(t) = \mathcal{L}^{-1} \{h(s)\} = 16.418 f_c e^{-5.805 f_c t} \sin(2.404 f_c t) \quad (6.59)$$

where f_c defines the cut-off frequency of the filter. Figure 6.19 shows the impulse response for a 100 MHz Gaussian low-pass filter.

Convolution of Eq. (6.59) with Eq. (6.1) characterises the filter response. Two calculation ranges must be considered separately to enforce the periodicity T_0 due to Eq. (6.1): $(kT_0 \leq t < kT_0 + \frac{\pi}{b}) \quad \forall k \in \mathbb{N}$ and $t \geq \frac{\pi}{b}$. This is demonstrated in Fig. 6.20.

The interval $(kT_0 \leq t < kT_0 + \frac{\pi}{b})$ is highlighted using a red dotted hash. The impulse response for the time t matching this interval consists also of a fraction of a new bunch appearing on the interval $0 \leq t < \frac{\pi}{b}$, emphasised using an orange dotted hatch. Outside of this interval the bunch takes part of the second interval comprising all complete bunches present on the interval $\langle \frac{\pi}{b}; t \rangle$.

The convolution process on the first interval can be defined as:

$$g_1(t) = \begin{cases} G_1(t) & kT_0 \leq t < kT_0 + \frac{\pi}{b} \\ 0 & \text{otherwise} \end{cases} \quad (6.60)$$

where

$$G_1(t) = \int_0^{t-kT_0} 16.418a [\cos(b(t-kT_0-\tau))^2 - 1]^2 f_c e^{-5.805 f_c \tau} \sin(2.404 f_c \tau) d\tau \quad (6.61)$$

Substituting kT_0 into Eq. (6.61) takes into account, that only a fraction of the first bunch is included. Convolution on the interval $\langle \frac{\pi}{b}; t \rangle$ gives the following equation:

$$g_2(t) = \begin{cases} G_2(t) & t > \frac{\pi}{b} \\ 0 & \text{otherwise} \end{cases} \quad (6.62)$$

$$G_2(t) = \sum_{\eta=0}^k \int_0^{\frac{\pi}{b}} 16.418a [\cos(-b\tau)^2 - 1]^2 f_c e^{-5.805f_c(t+\tau-\xi)} \sin(2.404f_c(t+\tau-\xi)) d\tau \quad (6.63)$$

where ξ defines the “displacement” function:

$$\xi = \frac{\pi}{b} + \eta T_0 \quad (6.64)$$

and k is the number of complete bunches in the interval $\langle \frac{\pi}{b}; t \rangle$:

$$k = \text{trunc} \left(\frac{t - \frac{\pi}{b}}{T_0} \right) \quad (6.65)$$

The complete convolution is given by a sum of the terms evaluated over their specific intervals:

$$y(t) = g_1(t) + g_2(t) \quad (6.66)$$

To find an optimal filter frequency for both ions and protons a set of filter responses, depicted in Fig. 6.21, was produced. 25 ns proton (top), and 100 ns ion scenarios were simulated using 5×10^9 charges per bunch. Apparently, the choice between the ripple and the rise time is a matter of compromise, hence two cut-off frequencies were chosen. The 2.5 MHz cut-off is a best choice for all 25 ns bunch space scenarios, and a compromise for the 75 ns proton scenario to be used during the commissioning of the LHC. For the 100 ns ion scenarios a cut-off of 1.5 MHz was chosen. Using 4th order filters the ripple can be partially suppressed.

Filtering scatters a single bunch over multiple bunch slots. A set of filter responses for the LHC pilot bunch is shown in Fig. 6.22. Using a 2.5 MHz cut-off frequency spreads the pilot bunch over 16 integration periods.

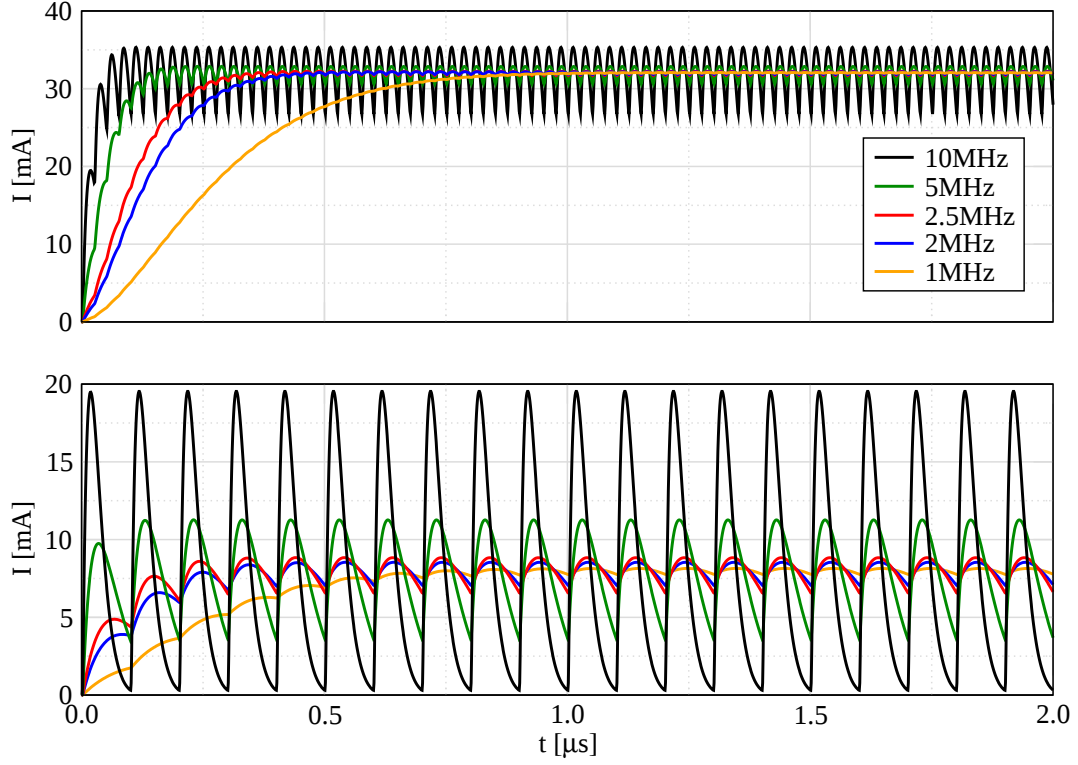


Figure 6.21: Gaussian filtering of the bunch train using various filter cut-off frequencies. The top graph depicts the LHC proton injection scenario while the bottom graph shows the train of $^{208}\text{Pb}^{82+}$ bunches spaced by 100 ns. The plots are calculated using $\sigma = 680$ ps and 5×10^9 charges per bunch.

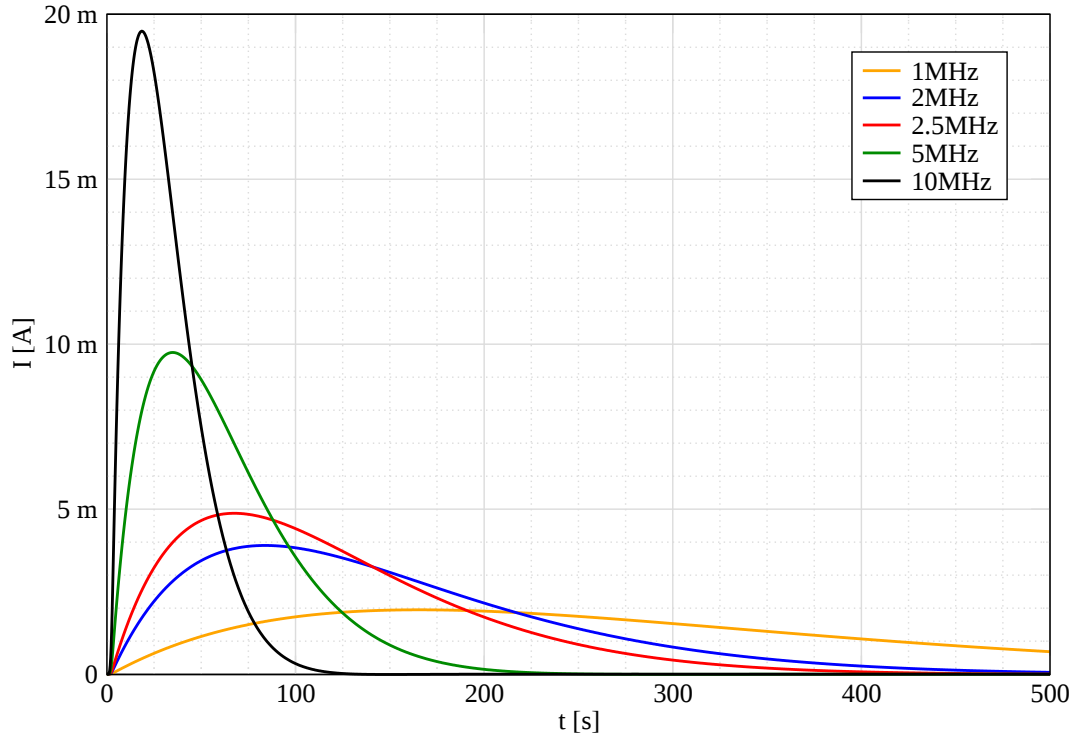


Figure 6.22: Gaussian filtered pilot beam equivalent signal.

7 THE FBCTs

7.1 Mechanical design of the FBCT

The fast beam intensity measurements for the LHC are provided by eight FBCTs. Four FBCTs installed in the LHC rings (BCTRs) provide both bunch by bunch and total turn by turn beam intensity information. Another four FBCTs (BCTDs), two in each of the LHC dump lines, measure the total extracted beam intensity. The BCTRs are installed in the straight section at the LHC access point 4 (IP4) (see Fig. 7.1) near Echenevex, France, where a majority of the beam instrumentation for the LHC circulating beam is situated. The BCTDs are installed on the vacuum chambers heading towards the beam dumps located at the LHC access point 6 (IP6), geographically situated between Versonnex, France and Bossy, Switzerland.

A common issue for both installations is a position fine-tuning (3D and tilt). The vacuum chambers and the FBCTs must be aligned to the centre of the beam with a sub-millimetre precision. An uneven surface of the tunnel floor (Fig. 7.2) and various fabrication inaccuracies however result in position deviations often exceeding 20 mm, what must be reflected in the design the FBCT support and its girder.

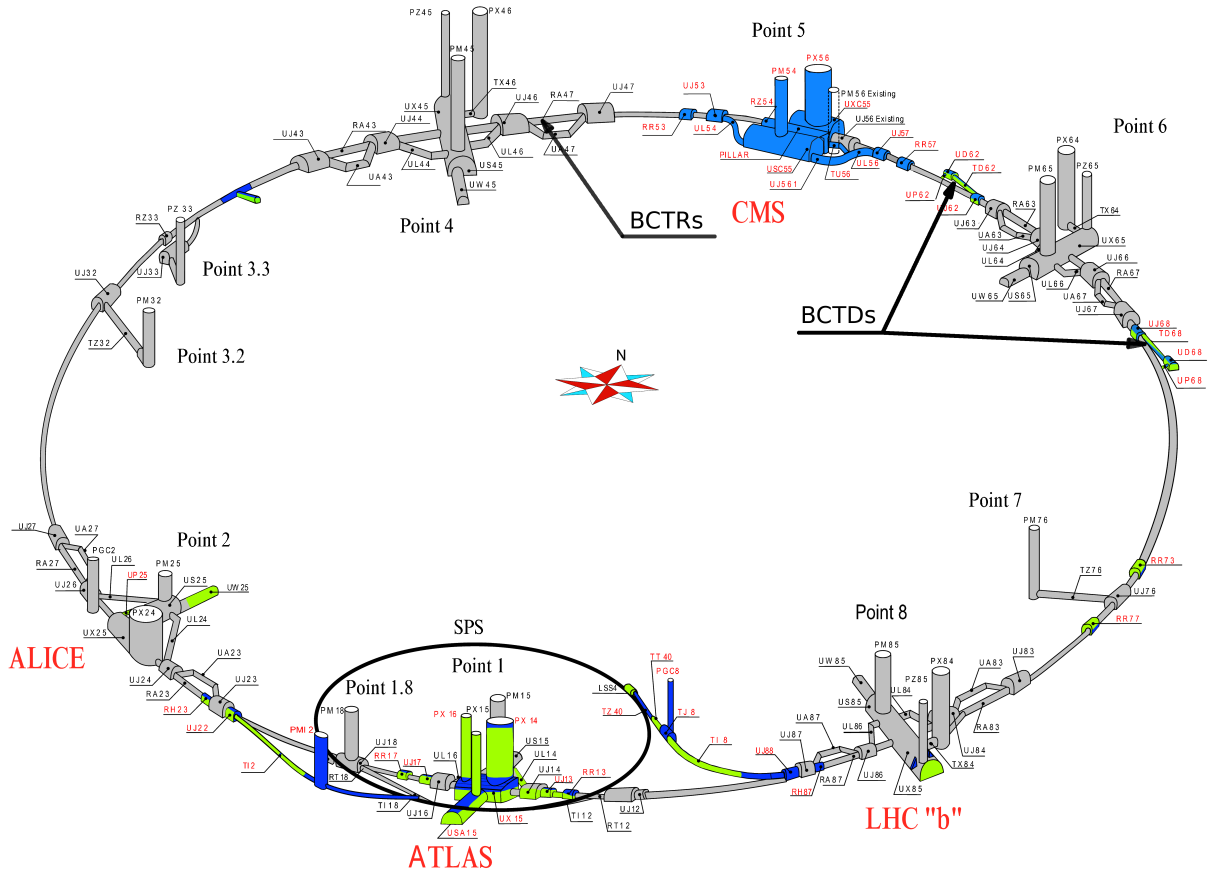


Figure 7.1: The LHC underground installation (courtesy ST-CE/JLB-hlm) showing the location of the FBCTs.

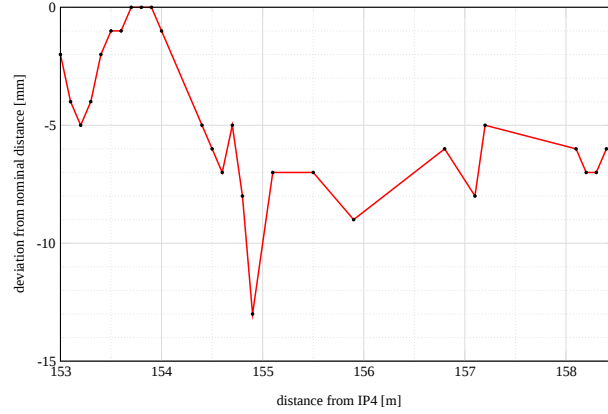


Figure 7.2: Deviation of the floor flatness from the nominal beam position (1100 mm) at IP4.

All LHC ring vacuum system is considered as Ultra High Vacuum¹ (UHV) hence a non-evaporable getter (NEG) coating is required for each installed vacuum chamber including the chambers for the beam instrumentation. The vacuum pumping is preceded by bake-out which activates the NEG. The activation temperature is approximately 200°C, nominal baking temperature used at CERN is 250°C. The magnetic material used in the FBCTs starts to change the crystal structure at the temperatures well below the baking temperature. Hence cooling circuit and proper interlock mechanism avoiding FBCTs damage is needed.

The FBCTs must be protected also against external magnetic field perturbations. This is especially important for the BCTDs due to their vicinity to a high field neighbouring magnets². Magnetic shielding is efficient only for perturbations caused by aerial EM coupling and has only minor effects in suppression of the signals induced by varying currents travelling through the vacuum chamber due to kicker transitions.

The BCTRs are installed in the vicinity of the DCCTs hence a common design of the girder, flanges, water cooling and the vacuum chambers was used. This remarkably simplified the design and reduced the cost of the development. It also brought additional advantages, e.g. possibility to cross-calibrate the two systems and allow a qualified decision about the intensity value sent to the MPS. In order to increase the availability of the measurement there are two independent FBCTs per vacuum chamber installed. A space was reserved in the LHC integration database to make possible an upgrade to a future three measurement devices. The third FBCT would increase the reliability of the measurement as averaging and majority voting could be applied on the measured data.

The installation (Fig. 7.3) consists of a 12 metres long girder. The monitors are mounted on the girder in four sections. Adjacent sections are connected via an LHC standard 84 mm vacuum chamber. Each section is composed of two monitors of the same type and their vacuum chambers. The chambers are installed symmetrically around a flange fixing them longitudinally. Bellows are installed at both extremities of each section

¹pressure 10^{-10} to 10^{-11} mbar [57]

²septum magnetic induction is 1.17 T at flat-top, kicker magnets generate ≈ 0.35 T. Septum magnets provide vertical deflection to raise the beam above the LHC machine towards the beam dumps. [57]

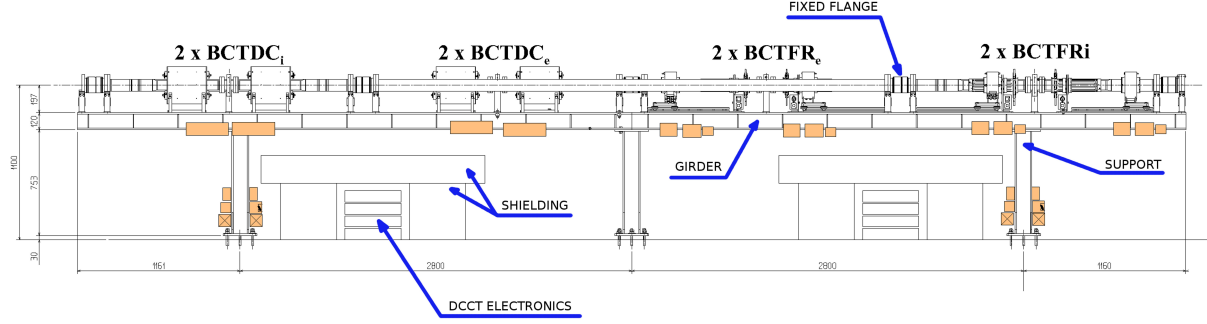


Figure 7.3: The LHC DCCT and BCTR installation. There are four sections of the BCTs, each containing two devices connected by a fixed flange.

to suppress mechanical forces seen by the ceramics. “RF fingers” inside of each bellow maintain constant longitudinal impedance. The DCCT electronics is shielded against the radiation by desk-like structures, each consisting of two concrete blocks and a thick iron desk. Orange zones in Fig. 7.3 depict a placement of water distribution components.

The high frequency signals are transmitted from each FBCT to its front-end electronics placed 25 metres away in the service gallery using 7/8” Heliflex air dielectrics cables (type HCA78-50). The high quality cables assure low losses over large frequency range (3.94 dB/100 m at 1 GHz). An attention was paid to match their electrical lengths and to choose their placement away of the cables transmitting high power pulsed signals. The chosen BCTRs location minimises the distance from the front-end electronics. Measurements shown the longest cable length to be 34 metres.

The installation of the BCTDs is symmetrical. Devices are located on both sides of the IP6, approximately 120 m away from the IP6 centre. The location choice is restricted by the dump lines vacuum chambers diameter, which increases towards the dumps from “standard” 84 mm to some 600 mm. Towards the injection insertions the dump vacuum chambers approach the ring vacuum chambers, not leaving enough space for the BCTDs. The chosen IP6 location permits usage of the same vacuum chamber and device in both IP4 and IP6. The girder is however different due to presence of both the dump lines and the circulating beams vacuum chambers in the IP6. Lower requirements on the vacuum quality in the dump lines simplify the BCTD design: no bake-out hence the cooling circuit is not needed. No NEG and other coatings are necessary. The IP6 installation is a simplified version of the IP4 installation.

Following sections describe the IP4 system mechanical construction optimisation, which is due to very tight technical constraints the most demanding part of the intensity measurement development process.

7.1.1 Supporting elements

The FBCTs have to be cooled during baking. Permanent installation of the heat-sink and the heating element does not permit to maintain short and low-impedance RF path for the wall image current and would result in an increase of the toroid internal diameter.

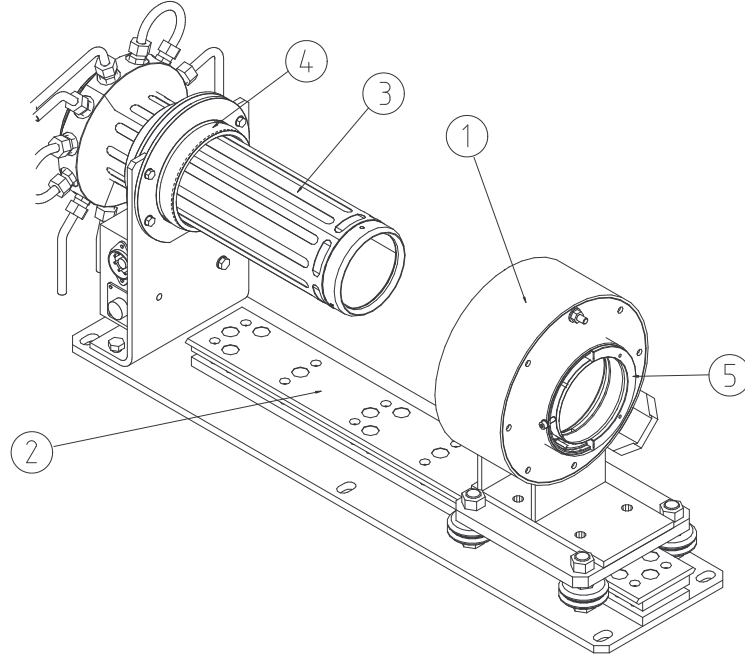


Figure 7.4: *The assembly of the BCTR.*

Having the FBCT permanently installed over the cooling tank would increase complexity of the repairs, especially when working in radiation areas.

Considering various implementations it was decided to design a movable support permitting to displace the FBCT over a heat-sink and heating element during baking. During the measurement is the FBCT installed over the ceramics, as close as possible to the vacuum chamber, minimising the path for the wall image current. The proposed solution keeps the FBCT compact what is important in order to shift the resonant frequencies due to a cavity formed by the FBCT's housing. Resonant modes driven by the beam spectral lines (every 40 MHz) would trap EM energy yielding to energy dissipation and heating of the toroids. In an extreme case the heat could lead to a loss of their permeability. The resonant modes would be also seen by the beam, increasing the machine impedance what could drive beam instabilities.

The assembly of the FBCT, support and the cooler in the measurement position is depicted in Fig. 7.4. The FBCT (pos.1) is fixed to HEPCO³ rails (2) using rolling wheels what permits a precise realignment of the transformer during the position change. The transformer is displaced between two the positions manually. Fixation of the rails by micro-metric threaded screws allows also to adjust the vertical FBCT position. Four half-collars (5) provide an RF path for the wall image current. These must be removed when placing the FBCT over the cooling tank (3). A fibre glass flange and RF contacts switch (4) serve as an interlock not permitting bake-out of incorrectly placed FBCT. The interlock signal is connected to the AT/VAC⁴ baking controller.

³www.hepcomotion.com

⁴AT/VAC: group responsible for the vacuum system

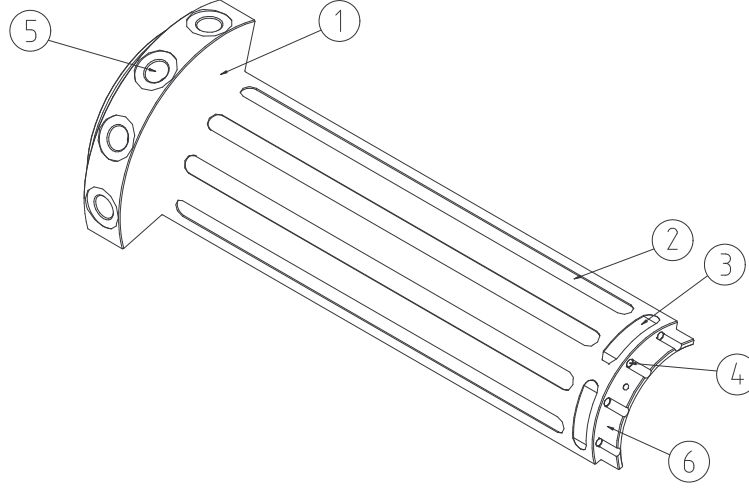


Figure 7.5: *One third of the BCTRs heat-sink. Model and realisation.*

The heat-sink is composed of three parts fabricated from a single block of copper by turning, milling, electron beam welding and electro-mechanical cutting. One third of the heat-sink is shown in Fig. 7.5. The cooling water is distributed by meander shape channels (2,3), milled into the heat-sink body (1) and re-sealed using half millimetre thick copper flat sheets. Drilled holes (4) going through the body connect the longitudinal and transversal channels. The copper parts (4) drilled away of the body are re-sealed brazing stoppers into the created holes. The meanders are connected in series to a common water outlet using 1/4" gas threaded fittings and stainless steel flexible tubes. In order to prevent bake-out when no coolant circulates in the heat-sink, the bake-out controller is interlocked by a water flow switch installed at the outlet.

The complete heat-sink assembly is fixed to the vacuum chamber on one side using a copper collar mounted over the comb-like top of the heat-sink (6) and by stainless steel tightening strips on the other side. The heat-sink is installed over a heating element, made of a 5 mm wide 50 μm thick stainless steel tape. The length of the heating tape wound on the vacuum chamber represents approximately 30 Ω load for the regulation system. This corresponds to an available heating power of 1.8 kW. A PWM regulator keeps the temperature of bake-out at 200°C. A 3 mm thick Microtherm Quilted panel separated by Kapton foil fills the volume between the heater and the heat-sink providing necessary thermal insulation needed to protect the toroid from the bake-out heat.

Experimental measurements from the first bake-out proved that a 1 l/min water flow maintains the temperature of the heat-sink at 37°C during whole 200°C baking cycle.

7.1.2 Magnetic Circuit of the Measurement device

The measurement device is composed of two toroids differing by sensitivity and bandwidth coverage. The primary toroid (HF) is a commercial product available from Bergoz Instrumentation [74]. It has 40 turns, droop of 0.14 %/ μs , and it is suitable for radiation areas.

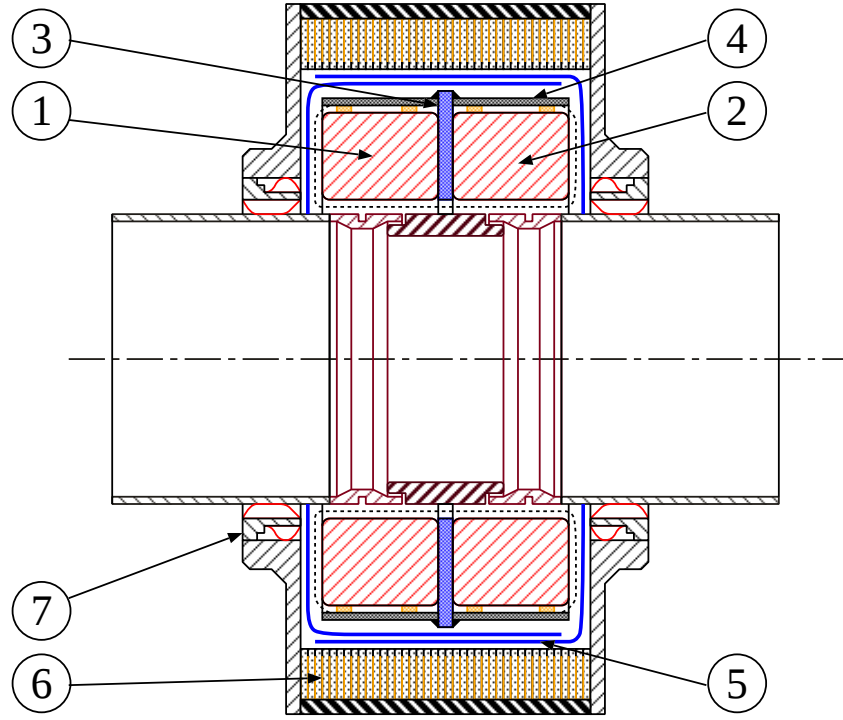


Figure 7.6: *Assembly of the FBCT. The toroids (Fig. 13.9 in the appendix), denoted using positions 1 and 2, are installed as close as possible to the ceramics, minimising the RF path for the wall image current.*

Its bandwidth extends from ≈ 250 Hz to more than 1 GHz. The decision to add a second toroid (LF) results from a calculation of the required BCTDs bandwidth, and from the fact, that the second toroid could be used in the future for other measurements at no additional cost. The second, LF toroid, is fabricated using a Fe-based nano-crystalline material Vitroperm 500F, produced by Vacuum Schmelze, having permeability $\mu_r \approx 10000$ @ 1 MHz what perfectly fits the needs to cover the frequency range up to 100 kHz. The core type is T60004-L2130-W630 with effective cross-section of 2.74 cm^2 . Nominal A_L at 100 kHz is $16.5 \mu\text{H}/\text{N}^2$ hence winding of 125 turns will provide a LF cut-off below 40 Hz. The winding of the toroid is protected by two layers of a fibre-glass tape, which is highly resistive to the radiation. Physical dimensions of the assembly ($135 \times 95 \times 30 \text{ mm}$) optimally satisfy the vacuum chamber design (outer diameter 84 mm).

Dimensions of both cores are matched to simplify their integration into the FBCT housing. Figure 13.9 in the appendix depicts the cores ready for the installation.

Assembly of the FBCT is shown in Fig. 7.6. Positions (1) and (2) depict the toroids mechanically fixed by a 5 mm thick fibre-glass insulating pad (3) and a calibration circuit (4), wound at exterior of the toroids. The dashed lines show the position of the calibration winding.

This assembly is inserted into a shell (5) composed of two elements which fit together. The shell forms first part of the Mu-metal static magnetic protection of the FBCT. The material exhibits high permeability provided the alignment of the material's grain with the magnetic field is correct. The alignment is however partially broken during ma-

chining and an annealing must take a place in order to restore the permeability. The two elements (Fig. 13.10 in the appendix), have been annealed at 1080°C during 1 hour under vacuum and let naturally cool down. The cuts in the shell parts provide a room to install SMA connectors.

The volume between the outer housing and the Mu-metal is filled by 3 mm thick PCBs (pos. 6 in Fig. 7.6) aligning the Mu-metal and the toroids to the device's geometrical centre. The PCBs are plated by a 35 μm thick copper foil. Electro-gilding is applied to the copper surface to avoid oxidation and to maintain a good electrical connection between the PCBs and the housing, reducing the volume of the resonant cavity.

The outer housing as a secondary part of the FBCTs's magnetic shielding is made of Armco material due to its low permeability and higher magnetic field saturation level compared to the Mu-metal. It protects the Mu-metal shell against a saturation caused by magnetic field perturbations. The material was annealed at 980°C during 1 hour in the vacuum and cooled down at constant rate of 100°C per hour. To polish the material a sand blasting followed the thermal treatment. The housing was plated by 10 μm of silver to improve an electrical contact between the internal parts of the FBCT, the housing, and the vacuum chamber.

Connection of the housing to the vacuum chamber, providing path for the wall image currents, is realised using RF collars (pos. 7 in Fig. 7.6). The collars are made of copper coated by 10 μm silver layer and equipped at both interior and exterior with beryllium copper RF contacts. The contacts maintain low-impedance connection to the vacuum chamber providing that appropriate parts of the chamber are silvered. Completely assembled measurement device is shown in Fig. 13.11 in the appendix.

7.2 The vacuum chamber fabrication technological processes

7.2.1 The vacuum chamber

Figure 7.7 depicts the vacuum chamber final design. It was decided to reduce the vacuum chamber diameter at the location of the heater and the heat-sink to maintain the toroids' internal diameter similar to the standard vacuum chamber diameter (84 mm). It also permits to have the bake-out elements permanently installed.

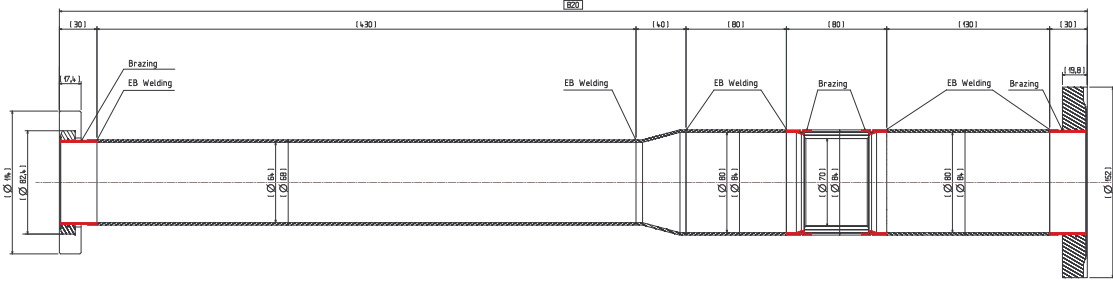


Figure 7.7: Design of the LHC FBCT vacuum chamber. Four brazed and six electron-beam welded connections are required to assembly the vacuum chamber. The OFE material is shown in red.

The vacuum chamber is 820 mm long, length of the permanent bake-out section is approximately 430 mm. The standard external diameter of 84 mm was reduced to 68 mm. This value is determined from the machine *mechanical acceptance*, defining a maximum beam aperture for a given location. A simplified calculation, specific for the BCTRs, is shown in the appendix 13.2.

In order to avoid sudden longitudinal impedance changes the vacuum chamber aperture is tapered using a short 15° cone section.

Fabrication process of the BCT vacuum chamber involves a realisation of an UHV proven ceramics to the stainless steel tube connection. Direct connection of the two materials is problematic so an additional buffer interface, acceptable for both materials, is needed. Typical connection usually uses a Kovar or Dilver⁵ in form of collars brazed to moly-manganese cemented extremities of the ceramics. The collars are Electron-Beam (EB) welded to the stainless steel vacuum chamber. Low conductivity of the stainless steel must be improved by depositing a copper layer to the chamber interior, introducing an additional technological complications.

In order to eliminate the copper deposition process it was decided to fabricate the vacuum chamber out of the Oxygen Free Copper with Silver (OFS). The annealing point of the OFS ($\approx 320^\circ\text{C}$) assures the material resistivity to a mechanical softening caused by

⁵KovarTM and DilverTM are special steels containing less than 2 % of carbon, and are doped with nickel and cobalt. Resulting alloys have Coefficient of Thermal Expansion (CTE) similar to borosilicate glasses. [80]

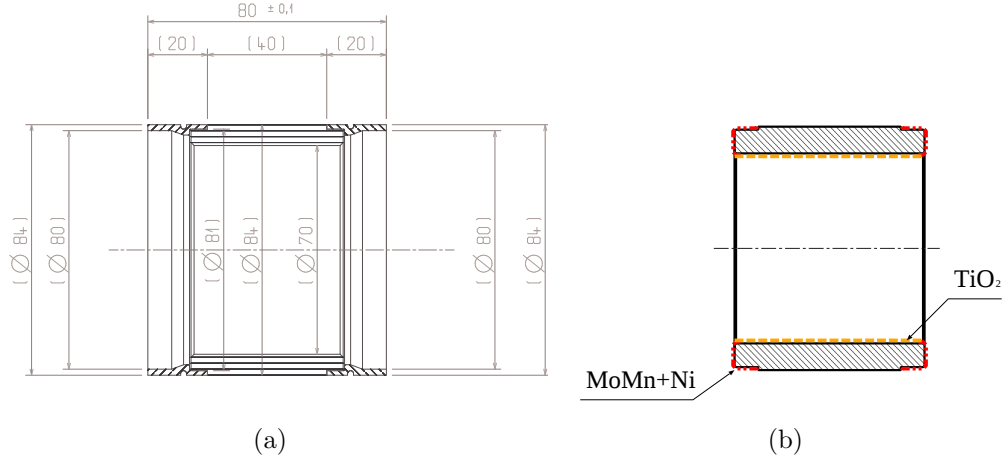


Figure 7.8: The FBCT vacuum chamber ceramics. Figure on the left shows the connection assembly. Figure on the right highlights MoMn+Ni and TiO_2 resistive coating.

the baking temperatures. The 68 mm copper tube is fabricated out of a 2 mm thick OFS copper sheet, rolled to the corresponding diameter and EB welded. The 84 mm tube is a standard LHC chamber fabricated from OFS by hot-rolling.

The vacuum chamber is equipped at both extremities with stainless steel flanges. The material guarantees creation of an UHV connection using a soft-material gasket elastically deformed when clamping flanges. The flange depicted on the left in Fig. 7.7 is demountable. Its external diameter is 114 mm. When disassembled the fixed part of the flange does not interfere with the toroids internal diameter so the FBCT can be quickly dismantled off the vacuum chamber.

The OFS still has a relatively high content of oxygen (≈ 10 ppm) hence is not suitable for direct brazing to the stainless steel flanges. The oxygen is released during brazing at high temperatures into a brazing additive and initiates an oxidation process resulting in poor vacuum connection. In addition, the brazed material is exposed to temperatures exceeding the OFS softening point. An Oxygen Free Electronic Copper (OFE) is used whenever brazing takes place. The OFE contains ≈ 5 ppm of oxygen. High temperature brazing in the vacuum without any flux and additives creates the UHV proven connection. The OFE has lower annealing temperature (250°C) hence is not suitable for baking. Combination of small OFE collars EB welded to the OFS makes the vacuum chamber resistant to baking temperatures. As seen in Fig. 7.7, there are 4 short sections of the vacuum chamber made of OFE: two at the extremities to braze the flanges and two collars connecting the ceramics to the OFS tubes.

Connection of the ceramics to the OFE collars is shown in Fig. 7.8(a). The ceramics type is Alumina AL300 (Al_2O_3) from WESGO Ceramics. The brazing extremities (Fig. 7.8(b)) are metallized using a thin molybden-manganese layer ($13\text{--}38 \mu\text{m}$). To improve wetting and adherence during the brazing an additional layer of pure nickel ($2.5\text{--}6 \mu\text{m}$) is deposited over the MoMn. Both layers are prepared by ceramics manufacturer as the process of metallization involves firing of the ceramics in a wet hydrogen atmosphere. An eutectic AgCu filler (28 % Cu) is used during diffusion brazing. A series of

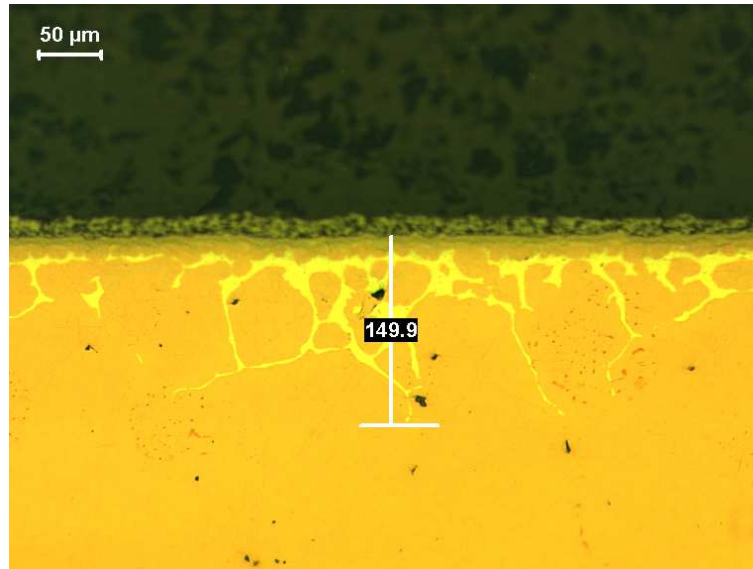


Figure 7.9: Photography of the ceramics connection to the OFE copper. Top dark part is the OFE, followed by MoMn and Nickel layers. The bottom amber part is the ceramics. Yellow paths show the AgCu alloy. Courtesy S. Mathot, CERN, TS/MME

destructive tests was performed at CERN TS/MME in order to find an optimum thickness of the filler layer. An excellent materials connection is achieved brazing in the vacuum at 795°C during 12 minutes. A 12 μm thick AgCu filler diffuses 150 μm deep into the ceramics (Fig. 7.9).

Due to a different CTE of copper (16.7 ppm/°C) and Alumina (6-7 ppm/°C) the parts of collars pulled over the ceramics detach during brazing. A molybdenum wire (CTE=5.5 ppm/°C) was wrapped around the copper collars to avoid this effect.

7.2.2 Titanium coating

Any charge deposited at the ceramics extremities during the beam passage must be removed using a resistive layer diffused at the interior of the ceramics. The resistive layer is depicted in Fig. 7.8(b) using orange colour. The MoMn layer covers a fraction of the ceramics interior as well and connects the resistive layer to both extremities. The resistance value is determined from the frequency characteristics of the vacuum chamber and FBCT assembly. A test bench, composed of the FBCT and its vacuum chamber, was developed to measure the frequency characteristics.

The ceramics of the used chamber is free of a resistive layer. Connection for the wall image current is realised using only external FBCT housing. The beam-like signal can be injected via an antenna inserted into the vacuum chamber, as seen in Fig. 7.10. The assembly forms a 50 Ω coaxial transmission line.

The antenna is composed of two parts fitting together (pos. 1, 2 in Fig. 7.10). Conical sections at the extremities hold central pins (7) of N-type female connectors, and realise a smooth diameter change while maintaining the 50 Ω impedance. Two flanges (5, 6) close the vacuum chamber and provide threads for the N connector housings (type

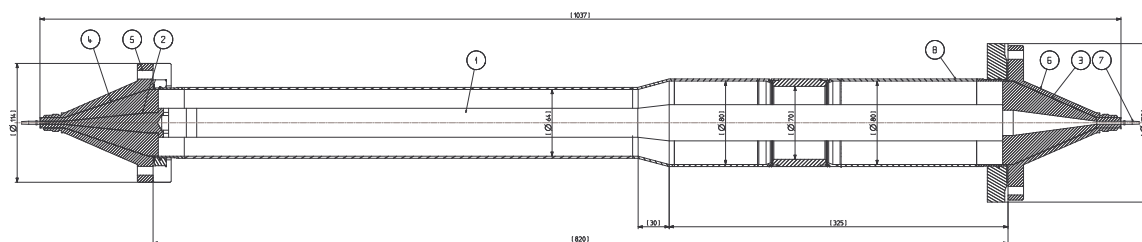


Figure 7.10: *The assembly of the vacuum chamber and the antenna.*

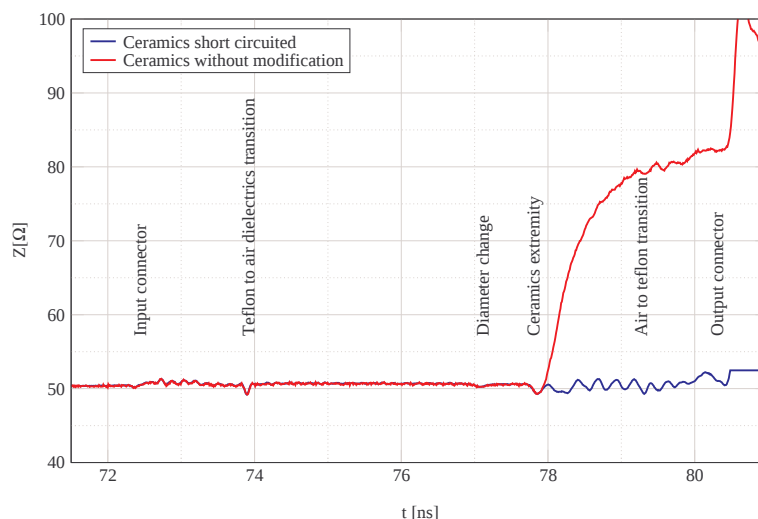


Figure 7.11: *TDR measurement of the vacuum chamber equipped with antenna.*

21 N-50-7-8C from Huber-Suhner). In order to minimise number of transitions from different materials the Teflon insertions (3, 4) extend from the connectors into conical parts of the vacuum chamber. The antenna and the flanges are made of an OFE copper, electro-plated by 10 μm thick layer of silver. The entire assembly is shown in Fig. 13.12 in the appendix.

To evaluate the electrical properties of the assembly the ceramics extremities were short-circuited and the antenna impedance was measured using a time-domain reflectometer (TDR). The result of the measurement is shown as a blue trace in Fig. 7.11. At the location where the ceramics is brazed the impedance starts to “wave”, however the peak-to-peak impedance does not exceed $50 \pm 1.5 \, \Omega$. The impedance mismatch is probably caused by a type of the used short circuit, fabricated using a copper scotch mounted at the exterior of the vacuum chamber. This changes the impedance because the signal travelling in the antenna should “see” a conducting surface at the interior of the vacuum chamber. The composite of the ceramics and the air dielectrics, and non-homogeneous mechanical structure also contribute to the impedance change. The red trace in Fig 7.11 shows the results of the same measurement without the short circuit installed. The impedance of the disconnected part is uncontrolled due to lack of an electrical connection to the ground potential. A short impedance drop common to both measurements identifies the location where the dielectrics type changes, i.e. where the conical section goes straight.

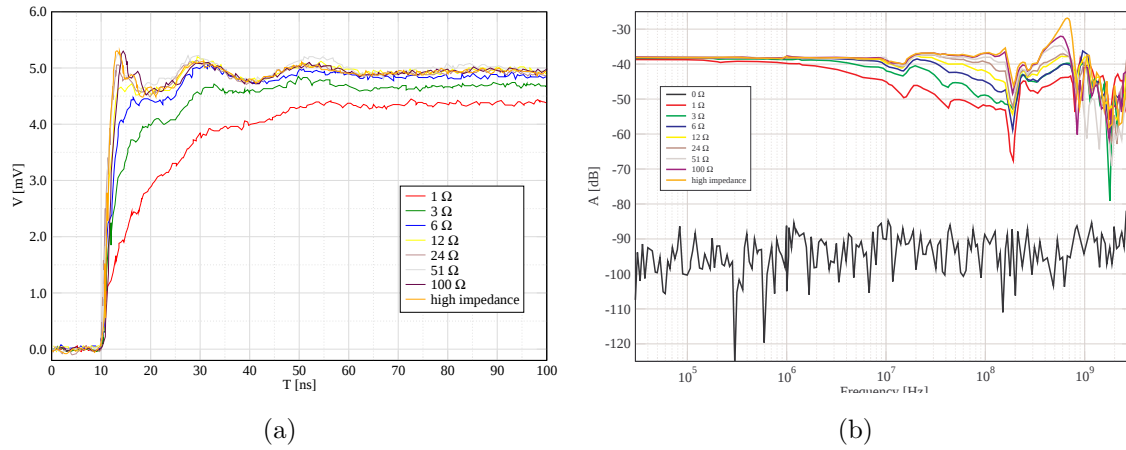


Figure 7.12: The primary toroid pulse-response measurement (a) showing influence of various impedances installed over the ceramics. Corresponding transfer function is shown on the right. The signal was injected in both measurements using antenna.

To perform the measurement a wide-band impedance controlled connection between the two extremities of the ceramics had to be provided. It was found that the best results were obtained using two sheets of copper ($w \times t = 20 \times 0.2$ mm) wound over the ceramics and fixed at both sides of the vacuum chamber by a copper scotch. The sheets were separated by 1 mm wide gap what permitted to solder eight 0603 resistors over the ceramics circumference. A set of such connections was fabricated using different values of resistors. Each connection was mounted over the ceramics, and a frequency and time domain transmission measurements were performed. The time-domain results are shown in Fig. 7.12(a), corresponding frequency domain measurements in Fig. 7.12(b).

The graph 7.12(b) shows a resonance present at the frequencies around 600 MHz, causing a ringing in the FBCT output signal. It is suppressed by lowering the impedance value to less than 24 Ω. A further damping is caused by ≈ 50 metres of a cable connecting the FBCT to the front-end electronics. Final value of the coating impedance is 20 – 25 Ω. Choosing lower impedance is not advantageous due to a decrease of the signal amplitude.

The resistive layer was created using a Titanium Oxide layer. The treatment was performed by specialists at CERN TS/MME. Specific resistance value was achieved controlling the deposited layer thickness. The titanium (Ti) was produced by a cathodic sputtering in a low pressure argon atmosphere (7×10^{-2} mbar). Method uses argon ions produced in a plasma to bombard a titanium cathode. The ions are accelerated to the cathode using static electric field created by a potential difference between the cathode and the ceramics collars connected to a high voltage (HV) source. The acceleration provides the ions with enough kinetic energy to liberate the titanium atoms from the cathode, so they can drift towards the ceramics.

Resulting resistance value is highly affected by an oxidation process, started by withdrawing the ceramics from the argon atmosphere. Oxidation increases the resistance of the deposited layer. Thinner the Ti layer is, the more significant increase of the resistance is observed during the time. For example a requested 10 kΩ resistance of the DCCT

Time	Voltage [kV]	R [Ω]	Note
0'	1.35	∞	Deposition starts
11'	1.17	720 k	
20'	1.25	135 k	
30'	1.35	620	
51'	1.37	260	
1h00	1.34	208	
1h10	1.4	169	
1h20	1.42	14.2	
1h30	1.44	12.6	
1h38	1.45	11.5	Treatment finished. Left in vacuum
$\approx 26h$	N/A	13.2	Under vacuum, before putting into air
27h		11.8	Air: atmospheric pressure
28h		13.3	Oxidation process
29h		13.7	
$\approx 48 h$		15.1	

Table 7.1: *Resistance measurement of a single BCTR ceramics at various stages of fabrication and aging [93].*

vacuum chamber ceramics, measured 7 $k\Omega$ after the deposition, increases to 100 $k\Omega$ in 24 hours and several megohms in two months after the treatment. This is caused by a thin Ti thickness (≈ 30 nm).

A much thicker layer is needed for desired 25 Ω resistance (≈ 200 nm) hence the effect of oxidation is less significant. Table. 7.1 shows the resistance evolution of a single FBCT ceramics. The 15.1 Ω resistance measured 48 hours after treatment increased to 21.6 Ω in 10 months. Similarly the resistance of the other BCTR ceramics increased to 26.2 Ω , 25.5 Ω and 28.5 Ω .

8 ANALYSIS AND IMPLEMENTATION OF THE ELECTRONIC CHAIN

Implementation of all necessary fast intensity measurement subsystems is separated into two units. The first, measurement unit, implements the signal distribution from the FBCT to the acquisition system, the acquisition system, protocols to integrate the measured data into the LHC controls system, and the BCF detector. The second unit provides the measurement calibration, namely its hardware part and software calibration procedures.

All FBCT hardware is implemented into the VME64x crates chosen as a standard in the CERN beam diagnostics group. The measurement chain schematic is shown in Fig. 8.1. Each turquoise block represents a physical equipment (a card). The green blocks symbolise parts of other systems using the FBCTs provided information.

The crates are installed in the LHC tunnel hence the electronics is not physically accessible during the LHC operation. This remarkably increases the hardware complexity as remote operation and supervision must be implemented. The remote supervision relies on LeCroy WaveRunner 64Xi oscilloscopes installed in the racks together with the crates. Ethernet connectivity of the oscilloscopes permits remote observation of desired signals. As number of observed signals exceeds the number of oscilloscope channels, a remote controlled RF multiplexer was developed extending each oscilloscope input into additional eight.

The Digital Acquisition Board (DAB), the calibrator, and the RF multiplexer are remotely controlled and reprogrammable using implemented VME64x interface. The RF distributor is passive, using only VME power supplies.

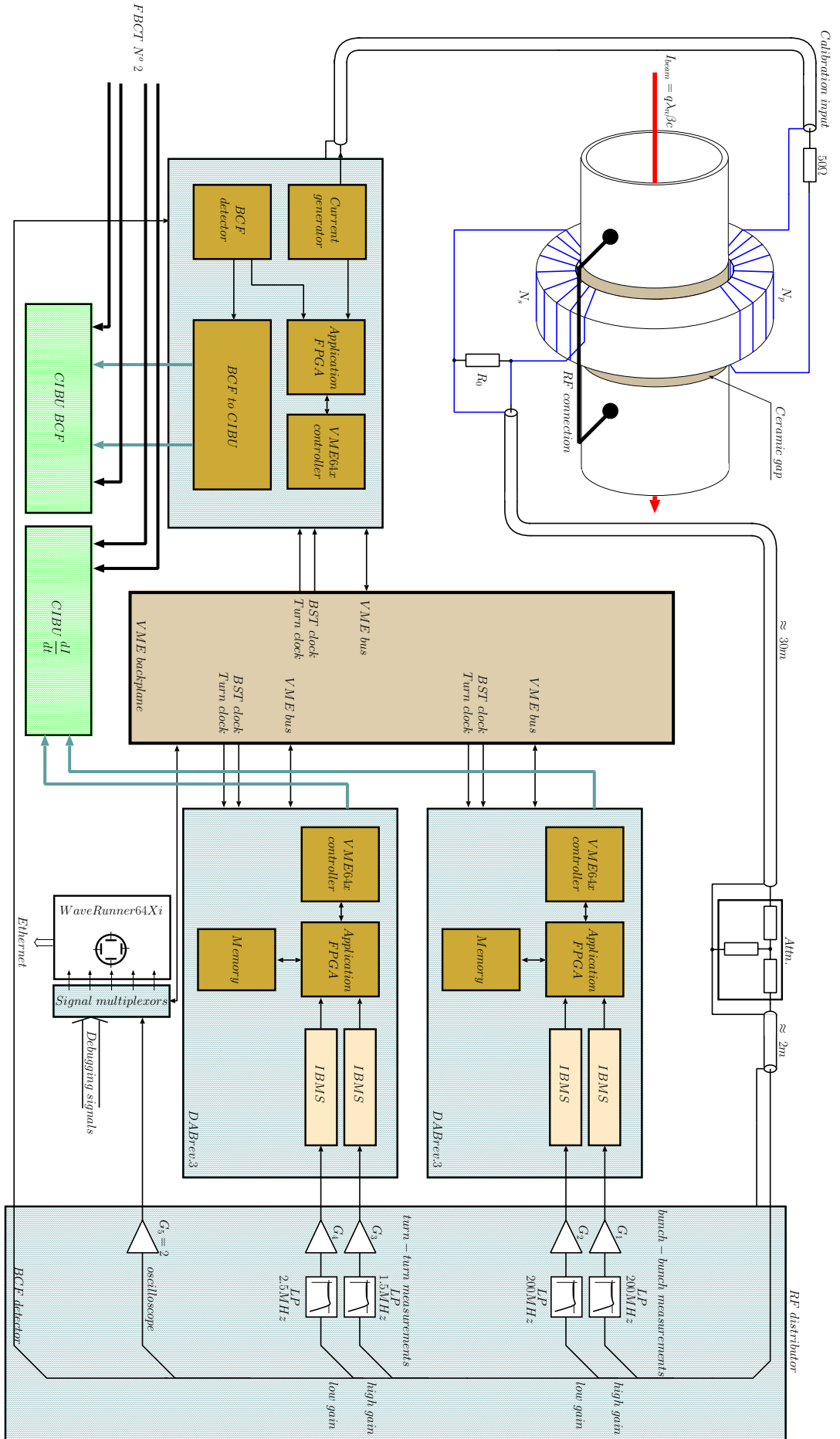


Figure 8.1: Building blocks of the intensity measurement system for the LHC

8.1 Acquisition system

A simplified block diagram of the fast intensity measurement system emphasising the acquisition part is depicted in Fig. 8.2. In order to maintain the compatibility of the CERN-SPS and the LHC FBCT systems it was decided to use an upgraded version of the digital processing chain based on the CERN-SPS system. This uses a DAB64x acquisition card [51] developed by TRIUMF(Canada) on which two IBMS mezzanine cards are mounted. The mezzanine cards integrate the incoming analogue signal using a 40 MHz integrator ASICs developed for the LHCb experiment (LHCb2002). The integrated signal is then digitised and processed on the DAB64x card to provide bunch-by-bunch intensity values.

Four channels (2 DAB64x boards) are used to provide measurements in two dynamic ranges as specified in Sec. 6.1.2. For each dynamic range the measurements can be acquired with a high bandwidth (HIBW) for bunch by bunch measurements and a low bandwidth (LOBW) for a timing insensitive total intensity measurements.

The LHCb2002 implements eight integration channels, each of them internally composed of two integrators working in a time multiplex. While one integrator integrates the incoming signal, the other performs reset. At the end of each 25 ns long integrating period a certain time is required to inverse the operation (a turn-off time). The two integrators exhibit different turn-off times hence they have different integration gate lengths. The average integration gate length is $\mu = 20.85 \text{ ns}$ with a standard deviation $\sigma = 400 \text{ ps}$ measured on a lot of 18 LHCb chips required for the FBCT measurement system.

Difference between the integration gate lengths transforms to a gain error. In order to compensate this effect a system which was successfully applied in the SPS was used. After the conversion of the integrated signal the data pass to the numeric value through a LUT. The table consists of a separate sector for each integrator. While the first sector is a “pass-through”, the second sector contains values compensating the offset of the integrator 1 and the gain difference between the two integrators. Hence the LUT corrected data stream looks as if it was produced by the integrator 0. In case of the LHC the algorithm was extended to suppress offsets of both integrators while matching the

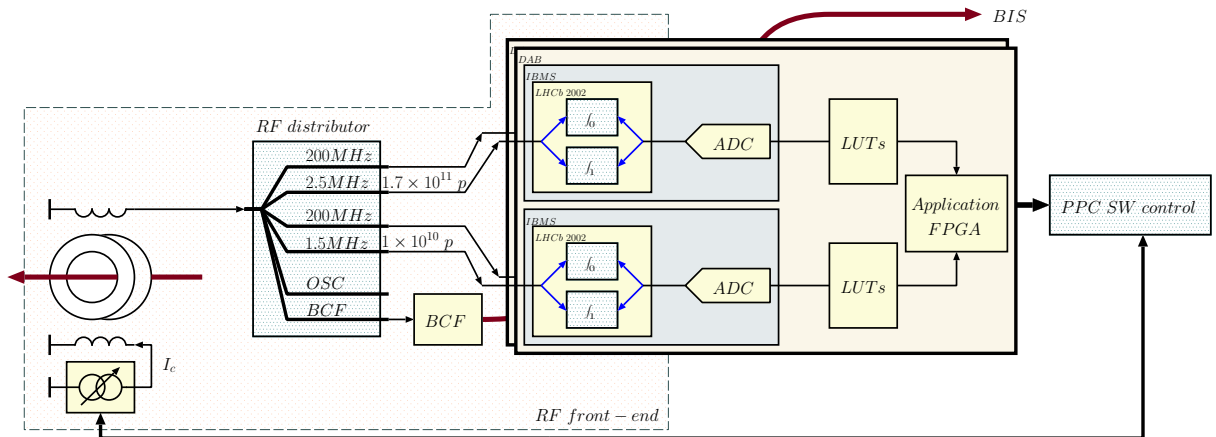


Figure 8.2: Schematic of the intensity measurement emphasising the acquisition system.

gain of integrator 1 to its counterpart. This is mandatory in order to provide correct functionality of the *sum sum* measurement mode.

The SPS correction calculation is based on the gate lengths measurement of each internal integrator. The difference of the integration time corresponds to a charge difference when a DC voltage is applied at the input of the integrator. This information is used to calculate a correction constant for the integrator 1.

For the LHC an another approach was developed. Method uses known properties of the input signal. If the LHCb chip input is fed by a square pulse of a given amplitude, the output signal, composed of consecutive train of 25ns integrations, must also have a form of a square pulse. As the applied pulse does not fluctuate (with the exception of a noise) the output signal should be fluctuation free as well. This is not a case of an uncompensated integration channel. The noise fluctuations can be partially cancelled by averaging on particular ranges of interest.

The calibrator can provide an input signal for the compensation. Each integrator channel was compensated using data from a 16 samples averaged current pulse produced by the calibrator. The integrated data stream was demultiplexed to separate the data produced by the integrators 0 and 1. Two ranges of interest were identified in each demultiplexed stream - a “low” range defining an average integrator offset and a “high” range defining an average value of the calibration pulse (in ADC bins). The four calculated values were transformed by linear fits used to suppress the offsets of both integrators, and match their gains. The linear fits were evaluated for all 16384 possible values returned by the ADC, and are reloaded into the LUTs on every reboot of the VME crate. The LUT compensation is depicted in Fig. 8.3.

When applying a DC signal to the integrator input there is always a fraction of charge lost due to the integrator’s turn-off time. In case of a pulsed signal no charge is lost providing that such signal fits entirely into the gates of both internal integrators. The measured charge difference between pulsed and DC input signal manifests itself as a gain difference. As the calibration pulse is treated as a DC signal, and the beam signal is a pulse-like signal the gain must be corrected for both HIBW measurement channels. The compensation factor is calculated using known integration times of both integrators. The data measured by the HIBW mezzanines must be multiplied by a ratio of the physical integrator gate lengths to the bunch period. As the correction factor is constant for each LHCb chip, it could be incorporated to the LUTs for HIBW measurement channels.

8.2 RF signal distribution

8.2.1 Principle of operation

The purpose of the RF distributor is to split the measured beam signal into a total of six outputs. Four of these outputs used for the intensity measurement are connected via

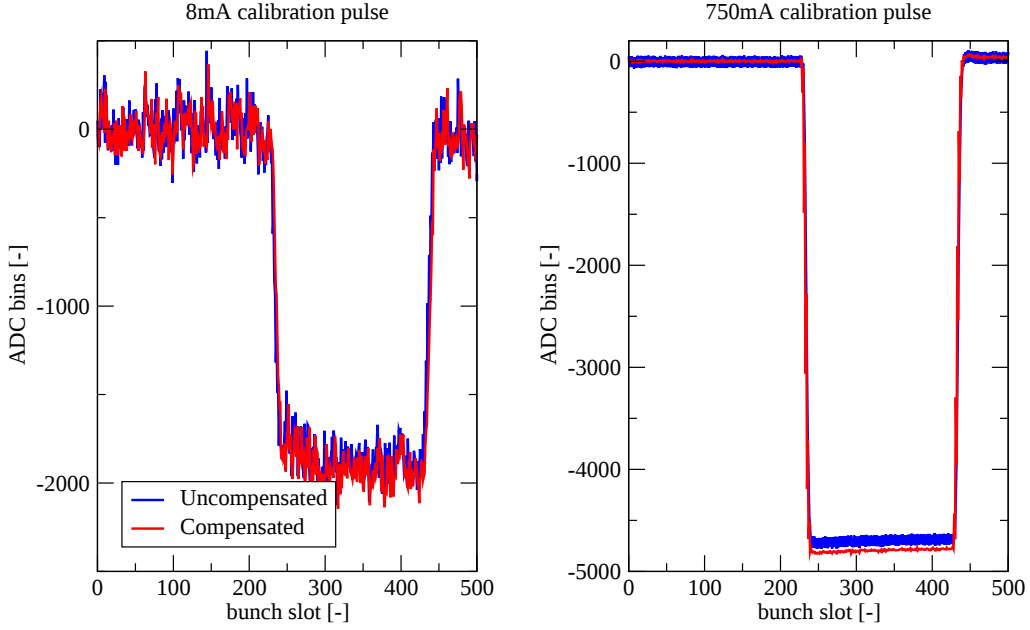


Figure 8.3: The LUT compensation of the LOBW measurement channels. Raw calibration pulses of 8 mA and 750 mA are generated using 1:1 LUT filling (blue traces). The LUT tables were produced using 30 mA (HIGAIN) and 800 mA (LOGAIN) calibration pulses. They were loaded into the DABs and an another set of measurements at 8 mA and 750 mA was produced (red traces). The calibrator is a current sink hence a negative signal is observed.

an appropriate protection circuitry to the IBMS mezzanines. The remaining two outputs are used for the BCF detector and an oscilloscope. As the fast intensity measurement is a HF system a good impedance matching at the RF distributor input and outputs must be assured. The four measurement channels must provide an offset compensation in order to maintain high measurement dynamic range. The RF distributor LF cut-off must not limit the transmission bandwidth of both installed toroids hence a value lower than 40Hz is required. A very wide band resistive splitter provides necessary beam signal splitting.

The input signal properties can be estimated using the base function (Eq. (6.1)):

$$i(t) = N_p e \cdot s_f(t) = N_p e \cdot a (\cos(bt)^2 - 1)^2 \quad (8.1)$$

$$I_{avg} = \frac{1}{T_0} \int_0^{\frac{\pi}{b}} i(t) dt = \frac{3}{8} \cdot \frac{\pi N_p e a}{T_0 b} \quad (8.2)$$

$$I_{peak} = N_p e \cdot a \quad (8.3)$$

The calculation was performed for two extreme cases: the ultimate beam at the flat-top ($N_p = 1.7 \times 10^{11}$ charges, $\sigma = 280$ ps) and the pilot beam at the injection ($N_p = 5 \times 10^9$ charges, $\sigma = 680$ ps). Table 8.1 summarises the signal properties recalculated to the output of the 1 : 40 turn ratio primary toroid. Its in-housing back-matching resistance results in a twice smaller signal amplitude compared to theoretical expectations.

The peak currents produced by the LHC ultimate beam have sufficient amplitude to be split however the pilot beam signal with a very low amplitude is at the limit of detection. This is especially the case of the average beam current. Furthermore it must

	Pilot injection 5×10^9 p.	Ultimate flat-top 1.7×10^{11} p.	Description
I_{pp}	6.07 mA	501.2 mA	peak beam current
I_{avg}	400.5 μ A	13.6 mA	average beam current
V_{pp}	303.5 mV	25.1 V	peak voltage at 50 Ω load
V_{avg}	20 mV	0.68 V	average voltage at 50 Ω load
Q	10.0 pC	340.4 pC	charge per bunch

Table 8.1: Various beam parameters for 25 ns proton beam scenario. The values are specified at the output of 1 : 40 measurement transformer equipped with 50 Ω back-matching resistance.

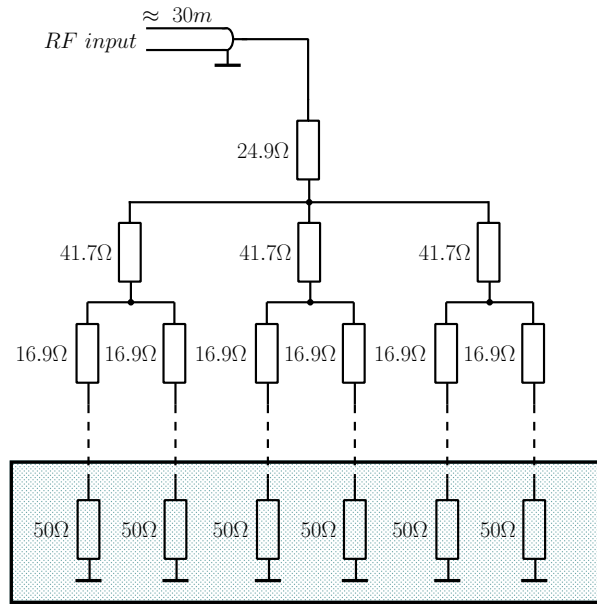


Figure 8.4: Resistive splitter part of the LHC RF signal distributor.

not be forgotten, that the average current for the standard ion scenario is still $4 \times$ lower due to longer bunch spacing.

The resistive splitter implementation is shown in Fig. 8.4. Usage of resistive splitter requires a good impedance matching at its input and outputs. If the condition is not met, the signal is not equally distributed into splitter branches and reflections may appear. In order to satisfy the condition it has to be assured that the loads, highlighted in Fig. 8.4 using a box, represent a wide-bandwidth 50 Ω loads. This is only achievable loading each splitter branch by a high-impedance buffer 50 Ω terminated at its input. High voltage operating amplifiers usually do not provide sufficient bandwidth for this application. Hence the idea is to use bipolar emitter followers, as shown in Fig. 8.5.

Transistors with 5 GHz transition frequency provide sufficient speed, low feedback capacitance, and high bandwidth for the follower to track the input signal. Their maximum C-E voltage is however limited to 15 V hence to avoid their damage the input peak voltage must not exceed approximately 6 V. This condition is met for all types of the LHC beams.

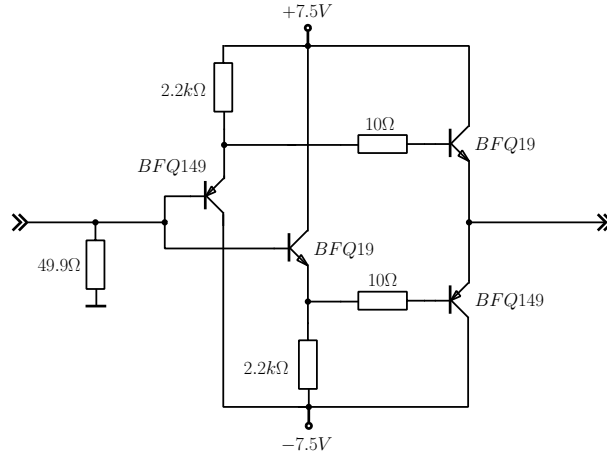


Figure 8.5: Bipolar emitter follower used as a unity gain buffer.

8.2.2 Filtering

As stated in Sec. 6.2.3, two types of low pass filters must be implemented (Fig 8.1):

- a high-order filter with $f_c = 2.5$ MHz and $f_c = 1.5$ MHz for time insensitive intensity measurements
- a filter with $f_c \geq 60$ MHz to keep the filtered bunch within a half of the bunch space

Both filter types must be implemented in the analogue part of the acquisition chain. To suppress an output signal overshoot and ringing a Gaussian type filters must be used.

The **high order low frequency** filters are used to provide a turn-basis measurement of the bunch current. As the filtered signal has extremely low amplitude it must be amplified to use a full dynamics of the analogue integrator. In case of the LOGAIN LOBW channel, the amplification factor of ≈ 3 dB is required mostly due to back-matching resistors installed at the outputs of each RF distributor measurement channel. In order to use full dynamic range of the analogue integrator an amplification higher than 30 dB is required for HIGAIN LOBW channel. The Current-feedback Operational Amplifiers (CFOAs) provided necessary gain. The channel output offsets were compensated using active feedback. The used amplifier configuration is shown in Fig. 8.6. The CFOA gain is set by appropriate selection of the resistor R_g . Suitably chosen value of the capacitor C_g in combination with the high-pass filter composed of C_p and R_p set the gain at low frequencies to 1 so the amplifier does not amplify neither the DC component of previous block (Sallen-Key Gaussian filter) nor input offset of the CFOA. The selected CFOA can exhibit input offset up to 3 mV and without further correction circuitry the HIGAIN LOBW channel integrator would saturate.

To reduce the effect of the CFOA input offset voltage (3 mV input offset of THS3001 corresponds to ≈ 2.5 % of integrator's full-scale) an active low-pass filter composed of OPA627, R_i and C_i was designed. The filtered output signal is re-injected with

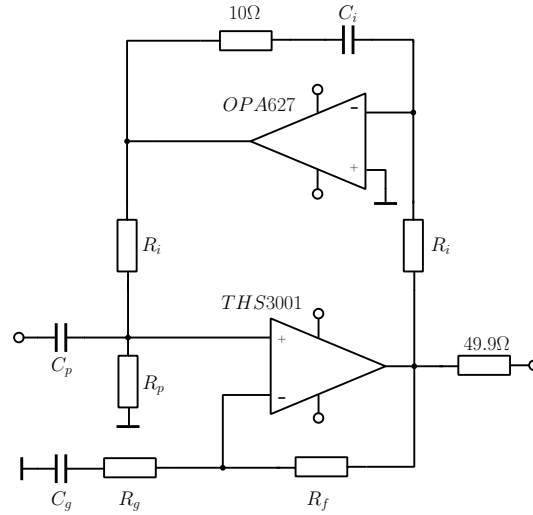


Figure 8.6: *High gain amplifier with offset suppressor*

an opposite phase to the input of the CFOA to compensate for its offset. Choosing correct cut-off frequencies for both HF and LF circuit paths the input offset of the CFOA can be suppressed to the value of the input offset of the OPA627.

The analogue integrator cannot process signals with amplitude higher than 2 V. Measuring on the HIBW channels the beam signal peak amplitude exceeds this value and it must be attenuated. The **high frequency** filter was designed to provide this attenuation (see Sec. 6.2.3). Its cut-off frequency was calculated such that the attenuation of the filter fits the required signal attenuation still complying to $f_c \geq 60$ MHz condition.

Equation (6.66) from Sec. 6.2.3 is re-used to find for a given filter's cut-off a maximum of the signal amplitude for specific bunch width. The relative data expressed in decibels are plotted into a graph. Using suitable approximation a single equation per bunch width can be used to calculate the attenuation factor.

Thiele rational interpolation was used. Changing the number of interpolation points permits to control the degree of the resulting polynomial. Using four points a rational polynomial of second order is obtained, which perfectly satisfies our needs. The Thiele interpolation for the entire set of bunch sizes is depicted together with the values numerically calculated in Fig. 8.7. The values shown are independent of number of charges. The most important interpolations correspond to $\sigma = 680$ ps and $\sigma = 280$ ps as these represent the LHC pilot beam at injection and the ultimate beam at the flat-top. Simple mathematical evaluation of the system RF path shows that the LOGAIN HIBW channel must be attenuated by ≈ 6.8 dB. This condition satisfies any filter with a cut-off frequency lower than 310 MHz. In the case of the HIGAIN HIBW channel an attenuation of ≈ 6 dB is required, what corresponds to a filter with a cut-off frequency lower than 350 MHz. A more practical value of 200 MHz, resulting in higher values of capacitors, was used to implement the HIBW channels filters.

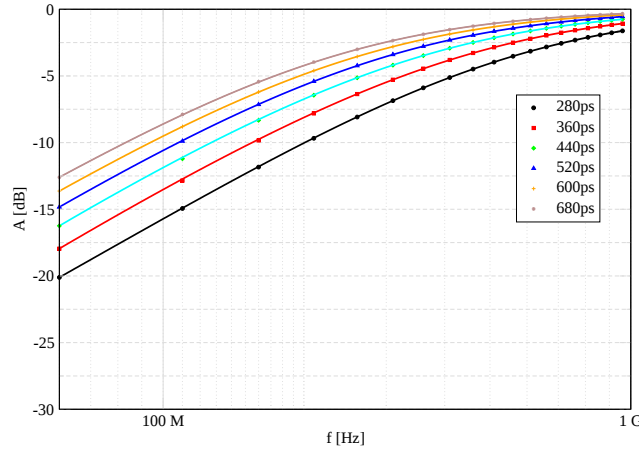


Figure 8.7: Attenuation of a single bunch beam-like signal, defined by its bunch width σ , passing through a second order low-pass Gaussian filter. Dots correspond to the attenuation factors numerically calculated from the maximum of the filtered bunch signal. Thiele interpolation is used to develop the attenuation equation for specific bunch width. The interpolation is shown as set of traces connecting the calculated values.

8.2.3 Hardware implementation

The resistive splitter as well as other parts of the RF distributor require a precise PCB tracks impedance control. Due to that a standard FR4 material, whose dielectric constant is not defined and widely spreads over various manufacturers and fabrication batches could not be used. In addition the design requires external shielding due to noisy environment in the VME crate. As a compromise between the RF performance and a circuit complexity it was decided to use a composite four layer PCB. Instead of using exclusively the RF material for all four layers (which would be mechanically fragile) a composite of an RF substrate and a regular FR4 material was used.

As a base RF substrate a NY9220 material from Park Electrochemical Corp. (formerly known as Nelco) was used. The dielectric constant of $D_k = 2.2$ together with a very low dissipation factor of $D_f = 0.0009 @ 10 GHz$ and low PCB thickness permit to keep the track widths below 3 mm for the entire design. 0.254 mm thick material was laminated with 0.5 mm prepreg together with 0.8 mm FR4 material to form a PCB plate approximately 1.7 mm thick. The top layer serves as a low RF loss and an impedance controlled environment. Power supplies and ground are routed exclusively on the internal and bottom FR4 layers. The NY9220 was chosen due to relatively similar coefficient of thermal expansion to FR4 (NY9220 CTE=25/35/260 ppm/°C, FR4 CTE=14/13/175 ppm/°C). Unfortunately no better match was found for materials having similar dielectric constant and targeting the same price category. As the match of the CTEs is not perfect the resulting PCB slightly bends when exposed to an excessive heat (e.g. wave soldering). This is not harming for our application but it denies usage of fine-pitch components in the design (eg. BGA or LQFP packages).

SMA edge mount connectors were used in the first prototypes (type Huber-

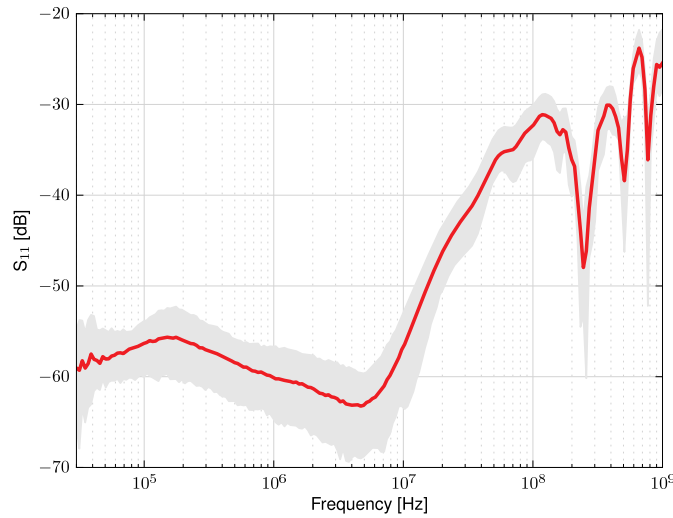


Figure 8.8: An average reflection coefficient S_{11} of the RF distributor input, measured on a lot of 20 produced cards. The grey area represents minimum and maximum measured values.

Suhner, 82_SMA-S50-0-45/111). Their combination with the RF substrate was found problematic due to mechanical fragility of the connection, high risk of the PCB material breakage, and complex shape of the front panel cut-out. Therefore next generation of the RF distributor was equipped with SMA connectors having through-hole ground connection from Johnson/Emerson (type 142-0761-821). The RF distributor is realised as a VME form-factor card as shown in Fig. 13.13 in the appendix.

8.2.4 Electrical properties

In total, twenty RF distributors were produced. Average frequency characteristics of all channels are shown in Fig. 8.8 and 8.9. Due to the PCB stacking and installation of the brass shielding it was possible to achieve a good average matching (measured reflection coefficient $s_{11} = -25 \text{ dB}$ at 800 MHz and $s_{11} = -20 \text{ dB}$ at 1.2 GHz) and maintain excellent electrical characteristics reproducibility of all produced cards. The LF cut-off lower than 30 Hz was measured for all measurement channels of the fabricated cards.

The measurement of the output DC offset, performed over all measurement channels on a lot of 20 fabricated RF distributors, is shown in Fig. 8.10. The average offset value of $47 \mu\text{V}$ with $\sigma = 145 \mu\text{V}$ fully agrees with predictions when using an “A” grade chips having $V_{io} = 500 \mu\text{V}_{max}$.

The standard deviation measurement of the high-frequency noise having a low amplitude was technically difficult to realise. Therefore to characterise the noise of each RF distributor channel a different approach was used. A test setup using full acquisition chain (RF input terminated RF distributor connected to the DAB card via a single mezzanine) was constructed. A single IBMS mezzanine was characterised in terms of gain difference between the two integrators, offsets, mean value and standard deviation of the ADC converted output noise for 50Ω terminated input.

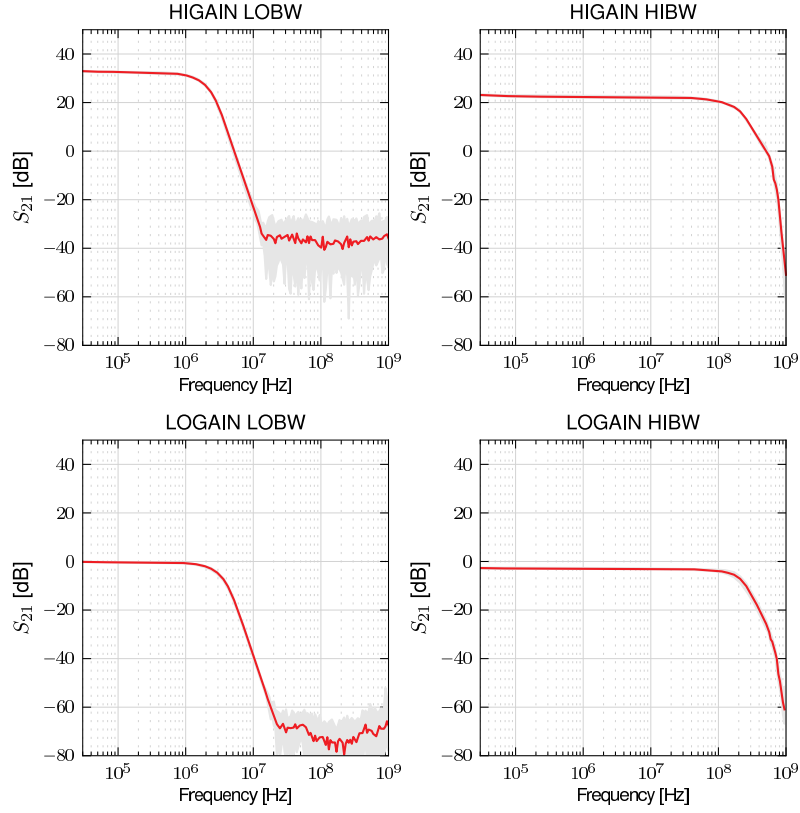


Figure 8.9: Average forward transmission measurement (S_{21}) of all four measurement channels. The resistive attenuation is excluded (15.5 dB). Gray area represents minimum and maximum measured values.

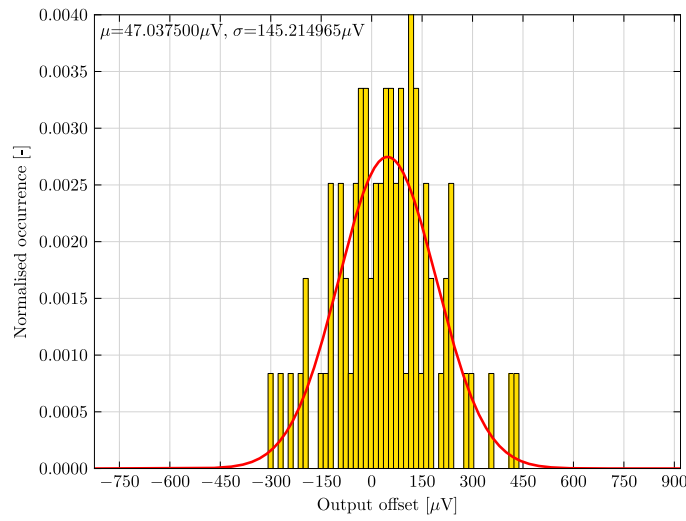


Figure 8.10: DC offset measured over four measurement channels on a lot of 20 fabricated RF distributors

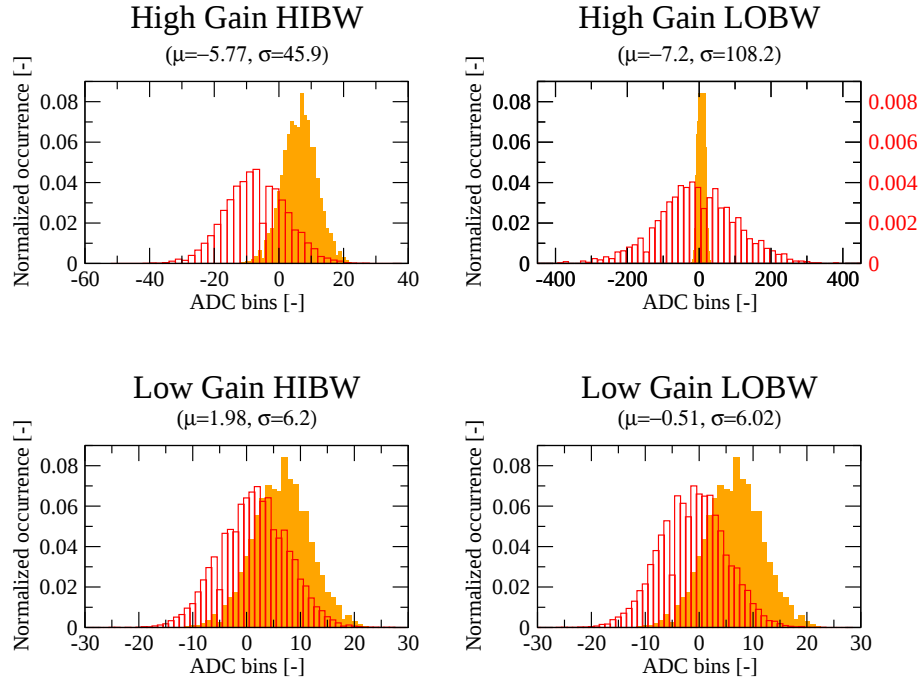


Figure 8.11: Typical noise measurement of RF distributor outputs (red histograms). The orange histograms show the noise reference measurement (input of the IBMS mezzanine terminated by 50Ω). The reference measurement is characterised by $\mu = 7.02$ and $\sigma = 5.33$ ADC bins.

Such mezzanine was used to perform comparative measurements. By connecting successively all the outputs from a single RF distributor to the reference mezzanine, and acquiring the integrated value, the noise behaviour could be characterised in terms of a standard deviation. Such measurement does not parametrise the noise at the particular output of the distributor. It rather describes the properties of a noise integrated over a bunch slot. Evidently, such measurement is of a higher importance, as it describes the intensity measurement signal noise floor determining the measurement precision.

The measurement results are shown in Fig. 8.11. The reference mezzanine noise measurement is depicted using orange colour. Red histograms show the measurements of each RF distributor channel. The shifts in the average values correspond to the RF distributor output offsets. Assuming normal distribution the calculated standard deviation equals to the measurement RMS value.

Known number of charges per ADC bin permits to recalculate the measured RMS noise into a number of charges. The constant must be found experimentally measuring the calibration pulse by the reference mezzanine for every RF distributor output.

The time of integration and the number of ADC bins corresponding to a known calibration current define the number of charges per ADC bin. Two calibration currents (650 mA and 30 mA) were used to calibrate all channels. The LUT correction factors were calculated and applied to the measured calibration signal. Table 8.2 shows the number of charges corresponding to an ADC bin together with the RMS noise recalculated to the number of charges for the particular channel. The noise threshold is compared to the measurement FS range to express the relative signal to noise ratio.

RF distributor output channel	number of charges [-]	RMS noise [-]	Relative to FS [%]
HIGAIN HIBW	1.45×10^6	6.6×10^7	0.66
HIGAIN LOBW	489.8×10^3	5.3×10^7	0.53
LOGAIN HIBW	27.47×10^6	1.7×10^8	0.01
LOGAIN LOBW	20.83×10^6	1.3×10^8	0.08

Table 8.2: *Number of charges per ADC bin recalculated to the RMS noise measured by the IBMS mezzanine using a single RF distributor.*

The values shown in the Tab. 8.2 are theoretical. They do not take into account noise acquired by the FBCT. This noise is expected to be dominant. Experience from other machines shows, that the captured noise can be orders of magnitude higher (up to 10^9 charges), and hence the values from Tab. 8.2 should be considered with great reserve and they are shown in the thesis only for the sake of completeness.

8.3 Machine protection system - beam circulating flag detector

Significant effort has been invested into the LHC Machine Protection System (MPS). Numerous equipments take a part in the MPS and deliver important information helping to decide whether the LHC works in a desired operational mode. One of the systems is also the fast beam intensity measurement. Its primary role is to detect fast losses (known as dI/dt , abbreviated as $dIdt$), slow losses (lifetime measurements) and generation of the Beam Circulating Flag (BCF).

For the LHC start-up some system simplifications had to be accepted. As both lifetime and $dIdt$ measurements are useful only when the intensity measurement algorithm and the electronics chain performance are tuned to the maximum, theirs usefulness will be limited during the start-up. Implementation of both is scheduled to the second phase of the LHC commissioning.

The critical part of the BCTR MPS implementation is the BCF detector. This system provides an information whether the LHC pilot beam currently circulates in the machine. The information is transmitted to the MPS in form of a single-bit information. Due to security precautions the detector malfunction causes an interlock blocking any further high intensity injection to the machine.

8.3.1 Hardware implementation

For safety reasons the circulating beam detector must be completely detached from any software solution and it is rather “hard-wired” circuit. Hardware implementation of such system would be to use a comparator to compare the measured signal to a fixed threshold, expressed in number of charges. However such implementation would only work for beam

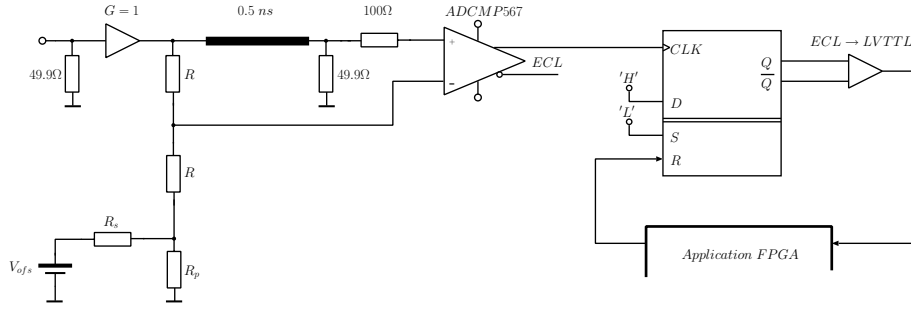


Figure 8.12: *Principal description of the BCF for the LHC*

signals exhibiting negligible droop as referring the threshold value to a fixed potential would result in a non-constant threshold.

To avoid the use of a drifting reference it was decided to implement the BCF using a droop insensitive Constant Fraction Discriminator (CFD). The implementation is shown in Fig. 8.12.

The dedicated output of the RF distributor is connected to a $50\ \Omega$ matching buffer. The CFD is realised as a fast comparator-summarator of a fraction-amplitude beam signal, and full amplitude beam signal delayed by $0.5\ ns$. The summed signal exhibits a zero-crossing exactly at a defined fraction of the original signal's amplitude regardless of the signal DC offset. An offset voltage V_{ofs} added to the part being subtracted displaces the zero-crossing point hence modifies the threshold. As sub-nanosecond pulses are processed, a very fast comparator must be used. A good candidate is ADCMP567 from Analog Devices.

Fast comparators usually use ECL inputs/outputs with switching rates far exceeding the capabilities of FPGAs. In order to properly register comparator output signals a hand-shaking synchroniser through the FPGA must be implemented. The pulse is latched in a fast ECL flip-flop and erased only when registered by the FPGA.

8.3.2 Algorithm of detection

The detection of the circulating beam is based on events counting. The circulating beam creates a CFD generated triggers repetitively each turn. Minimum number of trigger events is one, representing the LHC pilot bunch. A counter incrementing its value every time an event is detected is implemented in the FPGA. The counter is at the same time independently decremented at a rate slower than an LHC revolution period ($100\ \mu s$ vs $\approx 88.9\ \mu s$). When the beam circulates in the machine higher increase event rate saturates the counter in a time proportional to difference between the event period and the decrease period. When the beam stops to circulate the increase events are no more generated, and the counter drifts back to zero. The beam circulating flag is determined from an actual value of the counter. The flag is set to “True” when a counter value exceeds an user defined threshold value. To avoid the fluctuations of the flag value a fail-safe hysteresis was implemented. Decrease of the counter value below a defined threshold resets the

counter to zero hence the BCF is set to “False” more quickly than it takes to set it to “True” providing that the threshold is fixed to a value higher than half of the total number of counts.

This simple method assures that if the flag is set to “True” the beam was circulating at least a number of events determined by the counter threshold. In case of the LHC pilot bunch this corresponds to a number of the LHC revolution periods. For the time being an 8 bit counter with the threshold value set to approximately 75 % of the full scale is used.

8.3.3 Physical realisation

In order to save the development resources it was decided to implement the BCF detector as a part of the calibrator. Detailed implementation is shown in Sec. 8.4.3.

Complexity of the threshold calculation (finding roots of a non-linear equation, filtration effects etc.) does not permit to express analytically the dependence of the threshold in number of charges to the DC offset V_{ofs} . The detection threshold was found experimentally in the laboratory and fixed to $\approx 3.5 \times 10^9$ protons at injection. The system can detect a minimum of $\approx 3.0 \times 10^9$ protons, and is limited by the noise of the electronics.

8.4 The calibration

Measurement absolute accuracy is affected by attenuation and amplification of the measured signal. The gain change cannot be predicted with appropriate accuracy due to a non-deterministic behaviour of the electronic chain. Hence a calibration must be an integral part of the system. Precision of the calibrator influences the measurement precision so there is an interest to develop a calibration method with accuracy and resolution better than the one defined in the specification of the measurement.

The calibration part of the FBCTs is composed of two units - a pulse current generator and a calibration turn. A pulse current of a known charge is sent through the transmission line into the calibration turn, implemented into the FBCT housing (see Fig. 7.6, pos. 4). The calibration turn encompasses both toroidal transformers hence it simulates the beam current. The calibration constants can be calculated for all four measurement channels using measured system response.

8.4.1 Calibration turn

The magnetic flux in the toroids must be distributed in a maximally homogeneous way in order to transmit the maximum amount of energy between the primary and the secondary windings. For a single calibration turn an optimum flux distribution would be obtained injecting the current into a metallic cage encompassing the toroids, however this configuration is impossible to realise in the LHC due to mechanical constraints.

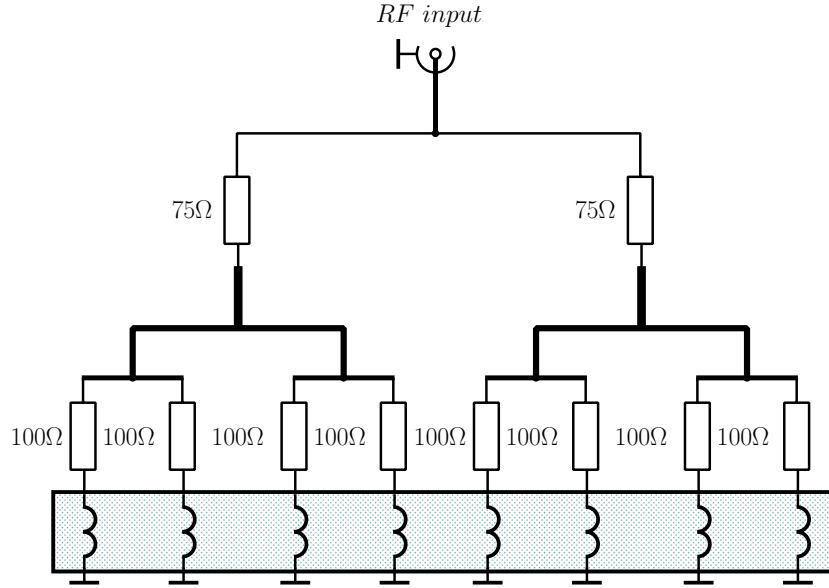


Figure 8.13: *The LHC FBCTs calibration turn.*

number of branches [-]	coefficient of transmission [%]
4	87.2
6	95.2
8	99.3
16	99.8

Table 8.3: *The calculated transmission coefficients for various number of single-turn parallel branches.*

It turns out that nearly similar behaviour could be achieved using parallel connection of sufficient number of turns. Similar method was already implemented in the SPS (Fig. 4.33), however in this case the resistances distribution did not allow impedance matching of the PCB tracks.

In order to find the best possible arrangement of the calibration turn winding a set of them was tested. The schematic drawing of the winning candidate is shown in Fig. 8.13. The test calibration turns were fabricated using 0.381 mm thick NY9220 RF substrate. Four connection types, dividing current to 4, 6, 8 and 16 parallel branches were tested. The charge transmission ratio was calculated comparing the charge measured at the FBCT output to the value calculated from the charge measured at the input of the calibration turn.

Measurement uncertainty due to the FBCT droop was limited injecting short current pulses (100 ns) into calibration winding. Summary of the results is shown in Tab. 8.3.

Unfortunately mechanical constraints do not permit to use the 16-branch version

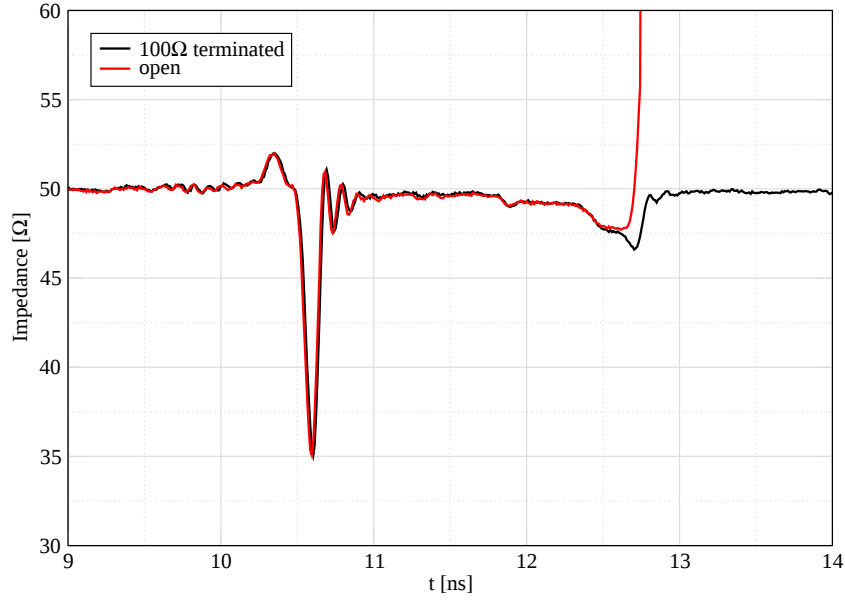


Figure 8.14: *TDR measurement of the 8-branch FBCT calibration turn*

as the calibration turn PCB is not large enough to fit required number of microstrips. Hence a compromise 8-branch version, shown in Fig. 13.14 in the appendix, was used. The circuit impedance was measured using a TDR and the results are shown in Fig. 8.14. There were two measurements done in order to identify the location of 100 Ω terminations. The black trace shows the measurement of the calibration turn including locally grounded 100 Ω termination resistors. The non-terminated calibration turn measurement is shown in the figure using the red trace. It is apparent, that weak point of the circuit is the transition between the SMA connector and the PCB. At this location the impedance drops to $\approx 35 \Omega$. It seems the fabrication reproducibility is excellent as both measurements were done using PCBs from different fabrication batches.

8.4.2 Pulse Current Generator

A gain and offset calibration is needed to match measurements in two dynamic ranges. Offsets can be measured when no beam is present in the machine. To correct for the gain as well a well known current must be measured. As each channel has different gain, there are needed totally three calibration currents to calibrate the FBCT acquisition chain. The calibration currents of 12 mA (for LOBW HIGAIN channel), 55 mA (HIBW HIGAIN), and 750 mA (both LOGAIN channels) have been chosen to excite appropriate channels to approximately 75 % of their FS ranges. The calibration current selection and the calibration procedure is controlled by the software executed in a PPC installed in the first slot of the VME crate.

a) Mathematical background

One calibration turn assures that the current transfer ratio to the FBCT output from the calibration turn is the same for the generated calibration pulse and the beam. The charge

generated by a current pulse could be expressed as:

$$Q = I_x \cdot t_i, \quad (8.4)$$

where I_x is the current sent into the FBCT calibration turn and t_i is the time of integration. Number of charges corresponding to the charge Q is determined using:

$$N_p = \frac{Q}{e} \quad (8.5)$$

where $e = 1.602 \times 10^{-19}$ C represents the elementary charge.

Assume the FBCT front-end and acquisition as a black box. For a given input in terms of number of charges an output in form of ADC bins is obtained. Having two point measurement a linear approximation of the measured number of charges can be defined:

$$N_p = \Gamma \cdot A + \Theta \quad (8.6)$$

where A is a LUT corrected ADC value corresponding to the beam signal integral for given bunch slot. Γ and Θ are wanted gain and offset correction factors.

Two measurements per a measurement channel form a measurement type:

$$N_{p,1} = \Gamma \cdot A_1 + \Theta \quad (8.7)$$

$$N_{p,2} = \Gamma \cdot A_2 + \Theta \quad (8.8)$$

The numeric index indicates number of measurement type performed using different calibration current. It was found that the best correction is obtained when acquiring A_1 using just a noise floor and acquiring A_2 using approximately 75 % full scale range (FSR) current, which corresponds to the currents mentioned at the beginning of section. The correction factors can be estimated using following equation:

$$\Gamma = \frac{N_{p,1} - N_{p,2}}{A_1 - A_2} \quad (8.9)$$

$$\Theta = \frac{N_{p,2}A_1 - N_{p,1}A_2}{A_1 - A_2} \quad (8.10)$$

The accuracy of the correction factors can be further improved using averaging over consecutive bunch slot measurements. Usage of a maximum of 16 averaged samples (400 ns) is recommended in order to keep the errors due to the FBCT droop under a reasonable threshold (0.2 %). The averaging process could be described mathematically using following set of equations:

$$\Gamma = \frac{N_{p,1} - N_{p,2}}{A_{a,1} - A_{a,2}} \quad (8.11)$$

$$\Theta = \frac{N_{p,2}A_{a,1} - N_{p,1}A_{a,2}}{A_{a,1} - A_{a,2}}, \quad (8.12)$$

where

$$A_{a,(1,2)} = \frac{1}{16} \sum_{i=0}^{15} A_{i,(1,2)} \quad (8.13)$$

Averaging consecutive bunch slot measurements $A_{i,(1,2)}$ results in more precise conversion ratio between the $N_{pm,(1,2)}$ and A_i .

b) The calibrator operation

The calibrator operation is depicted in Fig. 8.15. The calibrator consists of a high voltage (HV) source, a variable current sink (I_c) and switches (Q_1 , Q_2). The operation of the calibrator is monitored by an ADC which measures the voltage over the reference resistance R_m using two measurement channels, as described in the following paragraph.

The calibrator functions as follows. In the idle mode the HV switch is turned on with transistors Q_1 and Q_2 in the off state. This permits the capacitors C_c to charge via the current passing through the path R_c - C_c - R_d as shown in the picture by the orange arrow. Both capacitors are charged at the same moment. When the calibration is triggered a set of timings is generated. Firstly, the HV switch is turned off and Q_1 opens (current flows). The disconnection of the HV is relatively time consuming as it takes $8\ \mu\text{s}$ to switch off the 200 V source. Once this is done, switch Q_1 is opened and with the HV disconnected from the R_c , the current starts to flow from the left-hand C_c into the current sink I_c (blue arrow) stabilising the feedback of the I_c and charging all internal capacitances. Once I_c is stabilised, switch Q_1 is quickly closed while opening Q_2 . The current now no longer flows through Q_1 and the dummy resistance R_d , but via the right-hand C_c , R_l and the calibration winding L_x (green arrow). As the current sink is already stabilised, concurrent switching of Q_1 and Q_2 provides a very sharp negative current pulse. Under laboratory conditions, using a short cable to connect a resistive load, a fall time of $4\ \text{ns}$ was measured for a current of $800\ \text{mA}$ ($\text{SR}=200\ \text{A}/\mu\text{s}$). The calibration cycle is completed by closing switch Q_2 and reconnecting the HV. This permits both capacitors C_c to charge again.

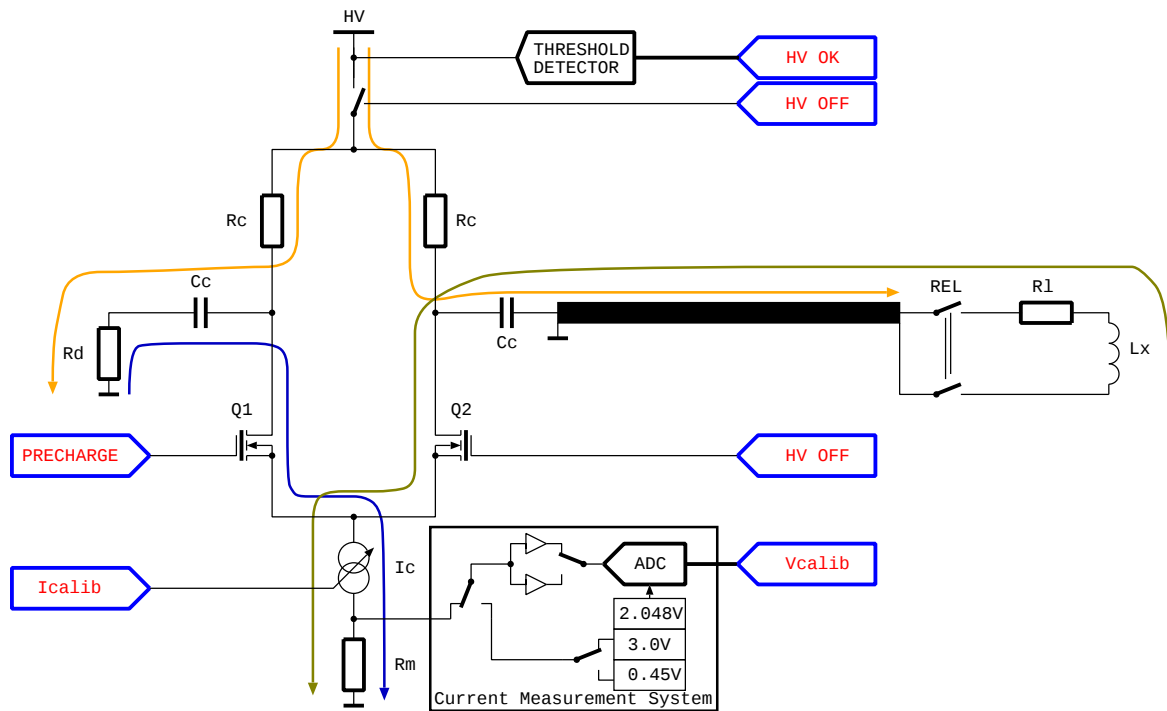


Figure 8.15: The principle of the LHC FBCT calibrator operation

Inspecting closely the principle of operation we can conclude following points:

- Calibrator generates negative current into the load
- Calibrator does not trigger instantly on the request as there is a time needed to turn-off the HV
- Two consecutive calibration shots must be sufficiently distant in time due to time needed to recharge the capacitors C_c after the calibration.
- A precise timing is required to switch between each state in the calibration process.

c) Precision of the calibrator

Absolute accuracy of the current amplitude setting is limited by offset errors in the DAC, on-board amplifiers and by precision of the used resistances. The amplifiers have high bandwidth and high slew rate, thus higher input offset voltages, and contribute significantly to the current setting uncertainty. Their usage also considerably increases a noise in the circuit. To limit their influence a lot of precautions has been done yielding to actual current setting uncertainty of ≈ 1.9 mA.

In order to improve the performance it was decided to measure the current sent to the calibration winding. The current measurement is based on a sampling and conversion of a voltage developed across a 0.1 % sense resistor installed in the feedback of the current source amplifier. Measurement is performed at the time the calibration pulse is precharged.

Situation is depicted in Fig. 8.16. The measured voltage is amplified by two fast Voltage-feedback Operational Amplifiers (VFOAs). The amplifiers amplify the input signal using different gains and provide two sensitivity ranges: the high-sensitivity range performs measurements of currents not exceeding 150 mA in amplitude and the low-

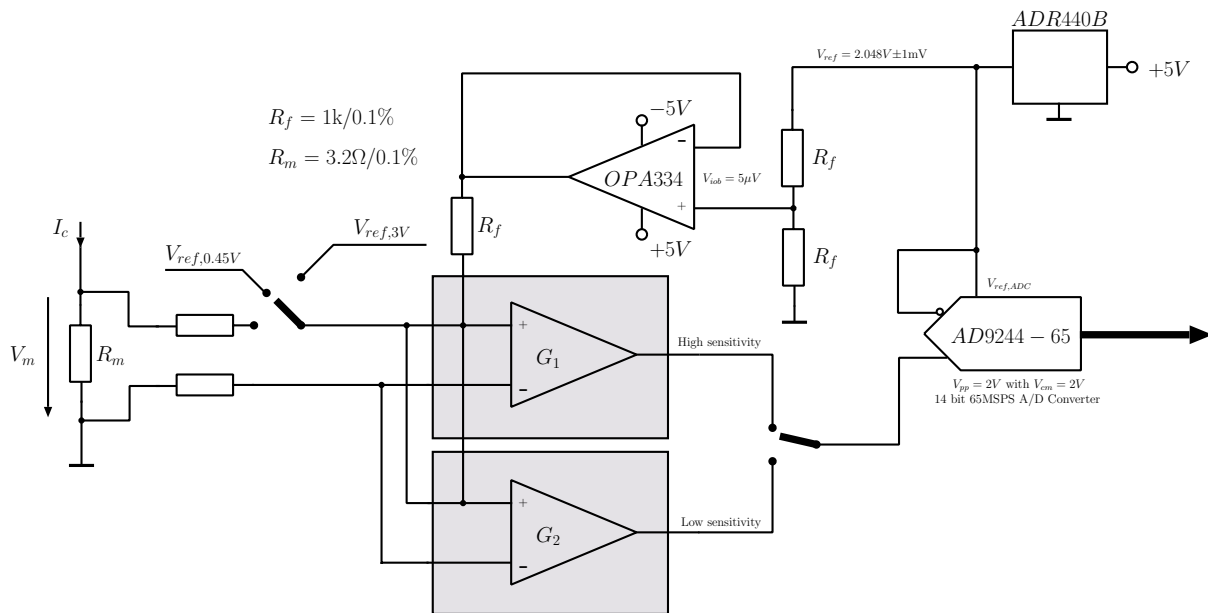


Figure 8.16: Calibration current measurement method principle

sensitivity range provides measurement of full 800 mA current span. The amplified voltage is then sampled using a 14-bit ADC (AD9244) working at 40 MHz sample frequency.

The reference voltage for the ADC is generated by a precision 2.048 ± 0.05 % voltage source. As the ADC's input voltage range is 1 to 3 volts an operational amplifier (OPA334) was used to shift the measured voltage V_m by 2 volts to the operational range of the ADC. The amplifier adds as well a small positive offset so that in all possible cases the ADC generates non-zero positive reading when performing offset measurements.

d) Uncertainty of the current measurement

Each measurement range is calibrated for offset and gain using reference voltages. In case of the low-sensitivity range reference voltages 3 V and 0.45 V are used. The high-sensitivity range is calibrated using “zero” and 0.45 V voltage references, where the “zero” measurement corresponds to a measurement of the noise. The sensitivity range calibration is performed on every current-setting using 4096 averaged samples of appropriate reference voltages. The measured reference voltages construct a two-point linear fit transforming the measured ADC bins into the calibration current.

Once the linear fit is known, the calibrator is ready to produce the calibration pulses. When a calibration is requested the amplitude of the calibration pulse is measured during its production. The length of the calibration pulse ($5\mu s$) permits to average the first 32 measurements. The calculated average ADC value is transformed into the calibration current using previously calculated linear fit.

The type-B uncertainty [73] of the measured current is estimated using the uncertainty propagation equation:

$$u_f = \sqrt{\sum_i \left(\frac{\partial f(x_0, x_1, \dots, x_i)}{\partial x_i} u_{x_i} \right)^2} \quad (8.14)$$

The current is determined from the voltage V_m measured on the resistance R_m , so that the type-B uncertainty of the measured current is determined as:

$$I_c = \frac{V_m}{R_m} \quad (8.15)$$

$$u_{B,I_c} = \sqrt{\left(\frac{\partial I_c}{\partial V_m} u_{V_m} \right)^2 + \left(\frac{\partial I_c}{\partial R_m} u_{R_m} \right)^2} = \sqrt{\frac{u_{V_m}^2}{R_m^2} + \frac{V_m^2 u_{R_m}^2}{R_m^4}} \quad (8.16)$$

where u_{V_m} and u_{R_m} express uncertainties of each parameter.

Providing that the electronics components are specified by manufacturer in terms of various parameters and their percentage tolerances δ , an uniform distribution of these parameters is expected. The conversion of the uniform range $p \pm \delta\%$ to the type-B uncertainty is realised using widely-accepted $\sqrt{3}$ rule:

$$u_p = \frac{1}{100} \cdot \frac{p \cdot \delta}{\sqrt{3}} \quad (8.17)$$

	High sensitivity	Low sensitivity
$V_{ref,1}$	0.45 V	3 V
$V_{ref,2}$	0 V	0.45 V

Table 8.4: *Parameters used to calculate the linear fit for both sensitivity ranges.*

and e.g. the uncertainty u_{R_m} for installed $R_m = 3.2 \Omega \pm 0.1 \%$ can be determined as:

$$u_{R_m} = \frac{1}{100} \cdot \frac{R_m \cdot \delta_{R_m}}{\sqrt{3}} = \frac{1}{100} \cdot \frac{3.2 \cdot 0.1}{\sqrt{3}} = 1.84 \text{ m}\Omega \quad (8.18)$$

Uncertainty u_{V_m} of the voltage value V_m must be determined using uncertainty propagation methods applied to the equation of the linear approximation.

Uncertainty due to the linear approximation equation: Linear approximation equation is used to convert the measured ADC value related to the calibration pulse amplitude into the corresponding calibration voltage. Then using Eq. (8.15) the voltage can be recalculated into the value of the calibration current. The algorithm of linear approximation calculation is realised in an FPGA and in order to simplify the implementation a common calculus is used for both sensitivity ranges. Following equation is used:

$$V_m = (V_{ref,1} - V_{ref,2}) \cdot \frac{ADC_m - ADC_{ref,2}}{ADC_{ref,1} - ADC_{ref,2}} + V_{ref,2} \quad (8.19)$$

where $V_{ref,(1,2)}$ are the reference voltages, and $ADC_{ref,(1,2)}$ are the ADC bins measured for given reference voltages. The value of ADC_m is proportional to the measured voltage V_m , and lies in the interval of $\langle 0; 16383 \rangle$.

High sensitivity range: Substituting values for high sensitivity range from Tab. 8.4 into Eq. (8.19) the linear approximation equation for high sensitivity range is obtained:

$$V_{m,h} = V_{ref,0.45V} \cdot \frac{ADC_m - ADC_o}{ADC_{V_{ref,0.45V}} - ADC_o} \quad (8.20)$$

Uncertainty of $V_{m,h}$ is calculated using Eq. (8.14). Square root of sum of isolated terms u_h , defined by following set of equations, determines resulting uncertainty $u_{B,V_{m,h}}$.

$$u_{h,1}^2 = \frac{V_{ref,0.45V}^2}{(ADC_{V_{ref,0.45V}} - ADC_o)^2} \cdot u_{ADC_m}^2 \quad (8.21)$$

$$u_{h,2}^2 = \frac{(ADC_m - ADC_o)^2}{(ADC_{V_{ref,0.45V}} - ADC_o)^2} \cdot u_{V_{ref,0.45V}}^2 \quad (8.22)$$

$$u_{h,3}^2 = \frac{V_{ref,0.45V}^2 (ADC_{V_{ref,0.45V}} - ADC_m)^2}{(ADC_{V_{ref,0.45V}} - ADC_o)^4} \cdot u_{ADC_o}^2 \quad (8.23)$$

$$u_{h,4}^2 = \frac{V_{ref,0.45V}^2 (ADC_m - ADC_o)^2}{(ADC_{V_{ref,0.45V}} - ADC_o)^4} \cdot u_{ADC_{V_{ref,0.45V}}}^2 \quad (8.24)$$

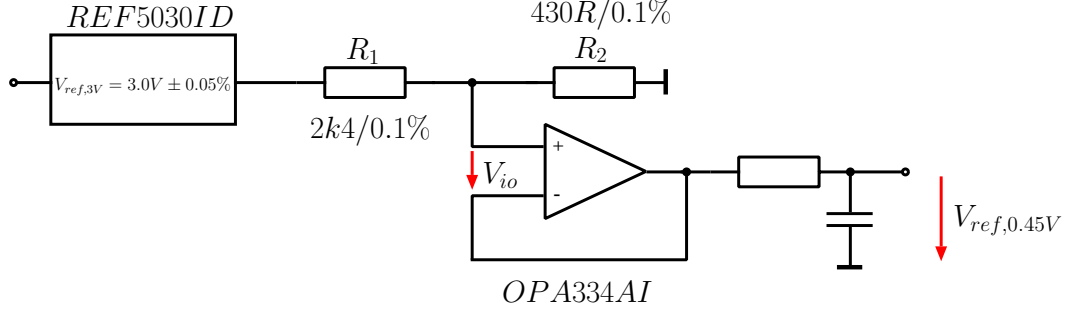


Figure 8.17: Calibration reference voltages

The total type-B uncertainty for the high sensitivity linear equation is determined using following equation:

$$u_{B,V_m,h} = \sqrt{\sum_{i=1}^4 u_{h,i}^2} \quad (8.25)$$

Low sensitivity range: The low sensitivity range linearisation equation is obtained substituting parameters of low sensitivity range from Tab. 8.4 into Eq. (8.19):

$$V_{m,l} = (V_{ref,3V} - V_{ref,0.45V}) \cdot \frac{ADC_m - ADC_{V_{ref,0.45V}}}{ADC_{V_{ref,3V}} - ADC_{V_{ref,0.45V}}} + V_{ref,0.45V} \quad (8.26)$$

The isolated uncertainty terms:

$$u_{l,1}^2 = \frac{(ADC_{V_{ref,3V}} - ADC_m)^2}{(ADC_{V_{ref,3V}} - ADC_{V_{ref,0.45V}})^2} \cdot u_{V_{ref,0.45V}}^2 \quad (8.27)$$

$$u_{l,2}^2 = \frac{(ADC_m - ADC_{V_{ref,0.45V}})^2}{(ADC_{V_{ref,3V}} - ADC_{V_{ref,0.45V}})^2} \cdot u_{V_{ref,3V}}^2 \quad (8.28)$$

$$u_{l,3}^2 = \frac{(V_{ref,3V} - V_{ref,0.45V})^2}{(ADC_{V_{ref,3V}} - ADC_{V_{ref,0.45V}})^2} \cdot u_{ADC_m}^2 \quad (8.29)$$

$$u_{l,4}^2 = \frac{(V_{ref,3V} - V_{ref,0.45V})^2 (ADC_{V_{ref,3V}} - ADC_m)^2}{(ADC_{V_{ref,3V}} - ADC_{V_{ref,0.45V}})^4} \cdot u_{ADC_{V_{ref,0.45V}}}^2 \quad (8.30)$$

$$u_{l,5}^2 = \frac{(V_{ref,3V} - V_{ref,0.45V})^2 (ADC_m - ADC_{V_{ref,0.45V}})^2}{(ADC_{V_{ref,3V}} - ADC_{V_{ref,0.45V}})^4} \cdot u_{ADC_{V_{ref,3V}}}^2 \quad (8.31)$$

determine the uncertainty $u_{B,V_m,l}$:

$$u_{B,V_m,l} = \sqrt{\sum_{i=1}^5 u_{l,i}^2} \quad (8.32)$$

Uncertainties of the reference voltages: Two reference voltages are implemented on-board. The $V_{ref,3V} = 3\text{ V} \pm 0.05\%$ reference is provided using REF5030 from Texas Instruments. This reference is used as well to generate the 0.45 V reference voltage as shown in Fig. 8.17. The type B uncertainty of the 3 V reference voltage is determined using Eq. (8.17):

$$u_{V_{ref,3V}} = \frac{1}{100} \cdot \frac{V_{ref} \cdot \delta}{\sqrt{3}} = \frac{1}{100} \cdot \frac{3V \cdot 0.05}{\sqrt{3}} = 866\text{ }\mu\text{V} \quad (8.33)$$

Parameter	Value	type B uncertainty
V_{io}	5 μV	2.9 μV
R_1	2400 $\Omega \pm 0.1 \%$	1.39 Ω
R_2	430 $\Omega \pm 0.1 \%$	248 m Ω
V_{ref}	3 V $\pm 0.05 \%$	866 μV

Table 8.5: Values substituted into Eq. (8.35). The type B uncertainty is in Eq. (8.35) denoted using lowercase 'u'.

In case of the 0.45 V reference voltage generator an effect of operational amplifier's input offset must be considered. The voltage at the output of the amplifier is expressed as:

$$V_{ref,0.45V} = V_{io} + \frac{V_{ref,3V} \cdot R_2}{R_1 + R_2} \quad (8.34)$$

where V_{io} is the input offset of the operational amplifier and $V_{ref,3V}$ is the 3 V reference voltage. Uncertainty of the resulting $V_{ref,0.45V}$ is determined using Eq. (8.14) as follows:

$$u_{V_{ref,0.45V}} = \sqrt{u_{V_{io}}^2 + \frac{R_2^2 u_{V_{ref,3V}}^2}{(R_1 + R_2)^2} + \frac{V_{ref,3V}^2 R_1^2 u_{R_2}^2}{(R_1 + R_2)^4} + \frac{V_{ref,3V}^2 R_2^2 u_{R_1}^2}{(R_1 + R_2)^4}} \quad (8.35)$$

Using values shown in Tab. 8.5 and recalculating parameters tolerances using Eq. (8.17) the uncertainty of the 0.45 V reference voltage is determined as:

$$u_{V_{ref,0.45V}} = 342 \mu V \quad (8.36)$$

Uncertainty due to the ADC: All the data measured by the ADC exhibit a combined uncertainty of the original signal source combined with an uncertainty of the used ADC. The acquisition chain offset and gain errors are compensated by Eq. (8.19) and hence their influence can be neglected. The only considerable error contributor in the measurement comes from the ADC's integral non-linearity:

$$u_{ADC} = \frac{\sqrt{3} INL \cdot FSR}{3 \cdot 2^N - 1} \quad (8.37)$$

where INL is the integral non-linearity (in ADC bins), N is the number of the ADC bits and FSR is the full-scale voltage. Using $INL = 4$, $N = 14$ and $FSR = 2.048 V$ the uncertainty due to the used ADC is:

$$u_{ADC} = 288.6 \mu V \quad (8.38)$$

Implemented linear approximation algorithm makes use of the ADC to measure the reference voltages and hence their uncertainties must be compensated to reflect the error due to the used ADC.

Parameter	High sensitivity		Low sensitivity	
	Value	Uncertainty	Value	Uncertainty
ADC_o	276 bins	2.3 bins	-	-
ADC_m	0 - 16384 bins	2.3 bins	0 - 16384 bins	2.3 bins
$ADC_{V_{ref},3V}$	-	-	15632 bins	6 bins
$ADC_{V_{ref},0.45V}$	13159 bins	3.58 bins	2577 bins	3.58 bins
$V_{ref,0.45V}$	0.45 V	342 μV	0.45 V	342 μV
$V_{ref,3V}$	-	-	3 V	866 μV

Table 8.6: Values substituted into the uncertainty calculus equations

This is performed using geometrical sum of the appropriate uncertainties:

$$u_{ADC_{V_{ref},3V}} = \sqrt{u_{V_{ref},3V}^2 + u_{ADC}^2} = 912.8 \mu V \quad (8.39)$$

$$u_{ADC_{V_{ref},0.45V}} = \sqrt{u_{V_{ref},0.45V}^2 + u_{ADC}^2} = 337.5 \mu V \quad (8.40)$$

Both uncertainties of ADC_m and ADC_o are determined as:

$$u_{ADC_m} = u_{ADC_o} = u_{ADC} = 288.6 \mu V \quad (8.41)$$

All quantities u_{ADC} must be recalculated to ADC bins using resolution Q :

$$Q = \frac{FSR}{2^N - 1} \quad (8.42)$$

Total type-B uncertainty: Equations (8.25) and (8.32) determine the type-B uncertainty u_{V_m} of the voltage V_m for the high and the low sensitivity channels. Table 8.6 summarises the substituted values into those equations. Further substitution into Eq. (8.16) determines the type-B uncertainty of the measured current.

Type A uncertainty: The type A uncertainty can be estimated from the measurements performed on the calibrator prototype. The uncertainty was determined independently for each sensitivity range from a standard deviation of the current value calculated by the algorithm implemented in the FPGA. Estimation was based on a set of measurements using different values of current. There were 32 measurements per set performed and a standard deviation of the pulse amplitude was determined in each set. It was found that in no case the standard deviation exceeds $\sigma_{A,I_h} = 70 \mu A$ for high sensitivity range, and $\sigma_{A,I_l} = 480 \mu A$ for low sensitivity range. The type-A uncertainty can be found using standard deviation measurement as:

$$u_A(x) = \sqrt{\frac{1}{n(n-1)} \sum_{i=1}^n (x_i - \bar{x})^2} = \frac{1}{\sqrt{n}} \sqrt{\frac{1}{n-1} \sum_{i=1}^n (x_i - \bar{x})^2} = \frac{\sigma_x}{\sqrt{n}} \quad (8.43)$$

where n is number of samples, x_i is the i^{th} sample, σ_x is the standard deviation, and $\bar{x}$ is a mean value of the sample set, defined as:

$$\bar{x} = \frac{1}{n} \sum_{i=1}^n x_i \quad (8.44)$$

Substituting measured standard deviations and number of samples $n = 32$ into Eq. (8.43) the type-A uncertainty for both sensitivity ranges is determined as follows:

$$u_{A,I_h} = \frac{\sigma_{A,I_h}}{\sqrt{32}} = 12.4 \mu A \quad (8.45)$$

$$u_{A,I_l} = \frac{\sigma_{A,I_l}}{\sqrt{32}} = 84.9 \mu A \quad (8.46)$$

Geometrical sum of both uncertainties determines the total uncertainty:

$$u_{C,l} = \sqrt{u_{B,I_{c,l}}^2 + u_{A,I_l}^2} \quad (8.47)$$

$$u_{C,h} = \sqrt{u_{B,I_{c,h}}^2 + u_{A,I_h}^2} \quad (8.48)$$

As the presented model is a simplified version of the real implementation, it certainly contains various inaccuracies. It is common to include these inaccuracies, or other effects which cannot be mathematically described, into the model and provide an extended uncertainty. Using multiplication factor $k = 2$ the extended uncertainty is defined as:

$$U_{I_c} = k \cdot \begin{cases} u_{C,h} & \text{if } I_{out} < 140 \text{ mA} \\ u_{C,l} & \text{otherwise} \end{cases} \quad (8.49)$$

Relative extended uncertainty: Knowing the ADC readings for currents corresponding to the reference voltages at both sensitivity ranges linear equations $ADC_m = f(I_c)$ can be constructed for each sensitivity range. Plugging these equations into Eq. (8.19) determines the voltage V_m as a function of the theoretical current I_c , and when substituted into Eq. (8.16) it determines the relative extended uncertainty as a function of the theoretical current I_c :

$$U_{I_c,rel} = 100\% \cdot \frac{U_{I_c}}{I_c} \quad (8.50)$$

The relative uncertainty was calculated using Eq. (8.50) and results of the calculation are shown in Fig. 8.18.

The graph denotes four important current values summarised in Tab. 8.7. The worst-case uncertainty corresponding to a current pulse of 12 mA is largely lower than 1 %. As this calibration current is used to calibrate low bandwidth high gain measurement range (FS=1 × 10¹⁰ charges) defined by required 10 % absolute accuracy (Tab. 4.5 and Tab. 4.6), reasonable safety margin is obtained.

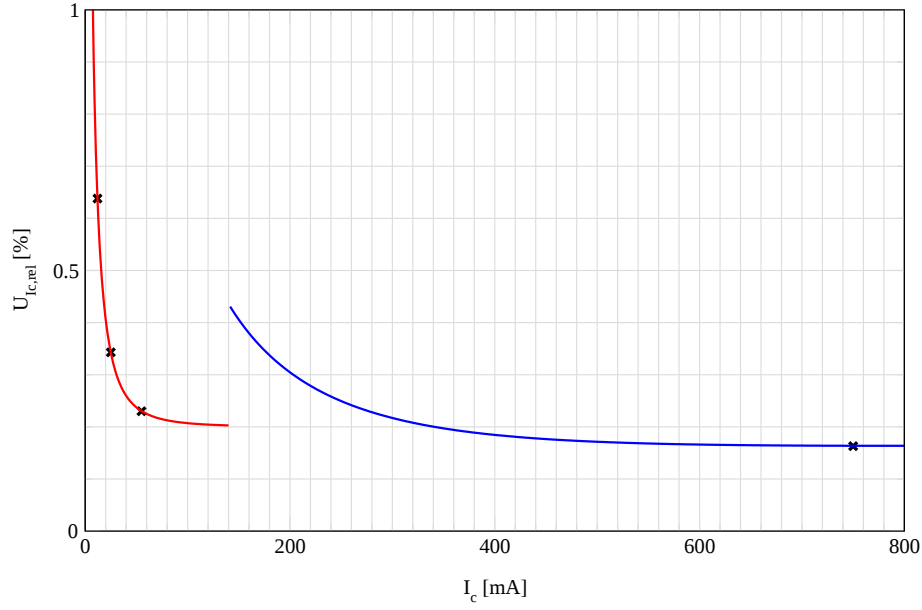


Figure 8.18: *The relative error of the current measurement*

I_c [mA]	$U_{I_c,rel}$ [%]
12	0.64
25	0.35
55	0.23
750	0.16

Table 8.7: *Relative extended uncertainty calculated for specific calibration currents.*

8.4.3 Implementation of the Calibrator

The calibrator together with the BCF detector is built on a single VME64x form factor card. The main application is implemented in hardware using an FPGA (Altera Cyclone). An additional CPLD (MaxII family) implements “common” processing tasks, as:

- providing a bridge between the VME interface and the target application
- a means to remotely program the FPGA using active-serial memory interface
- providing an unique card serial number
- proving a configuration memory which is accessible both from the VME space and the application FPGA

Such hardware implementation permits to re-use the calibrator design for other VME based cards by simply having a different FPGA firmware, and provision of the application specific connections. The design was already re-used e.g. by remotely controlled RF multiplexor, which is a part of the BCTR diagnostics facility.

The calibrator and the BCF detector implementation is shown in Fig. 13.15 in the appendix. A DC-DC converter provides the 200 V supply voltage for the calibrator. The BCF detector delay line is implemented using a 0.5 ns long SMA cable.

Each FBCT installed in the LHC is equipped with one calibrator. The calibration current is sent through approximately 35 metres long 7/8" air-dielectric Heliflex coaxial cable [79] into the calibration turn.

8.4.4 Measured results

Figure 8.19 shows the oscillogram of the calibration pulse measured at a $50\ \Omega$ load resistance connected to the calibrator through 2 metres long coaxial cable (type CKB50). The top graph depicts $5\ \mu\text{s}$ long calibration pulses of 10 mA, 25 mA and 55 mA in amplitude. The bottom graph depicts first third of a $15\ \mu\text{s}$ long 750 mA calibration pulse. The values shown in the legend correspond to the current measurements performed by the calibrator acquisition chain.

Figure 8.20 depicts three calibration pulses used to calibrate all four measurements channels. The graphs show the generated calibration pulses, acquired by the measurement mezzanines, LUT corrected, and recalculated to the number of charges using previously generated calibration coefficients. The top graph depicts measurements using both low gain (52.2 mA) and high gain (12.2 mA) low bandwidth channels. The bottom

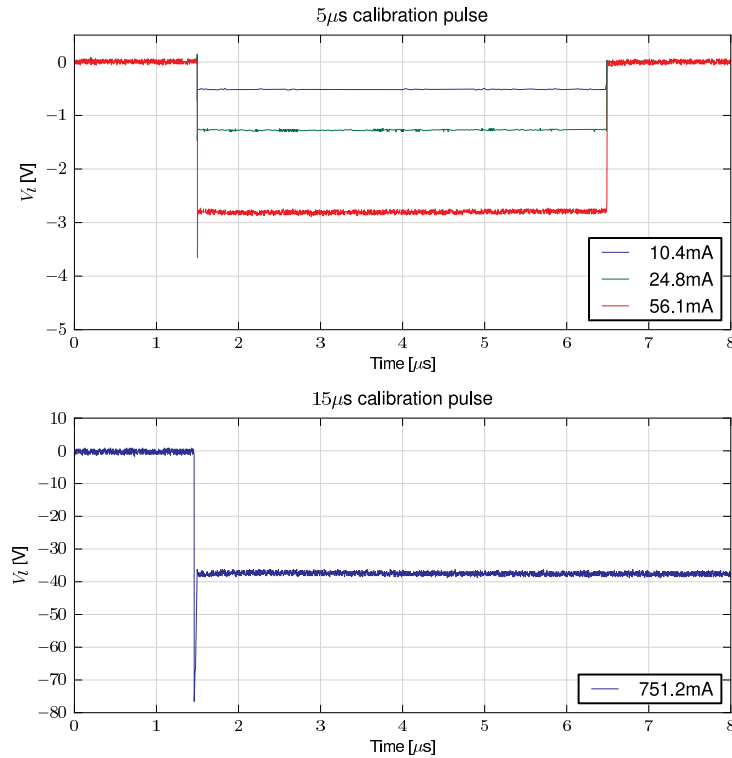


Figure 8.19: Calibration pulse measured at $50\ \Omega$ load resistance

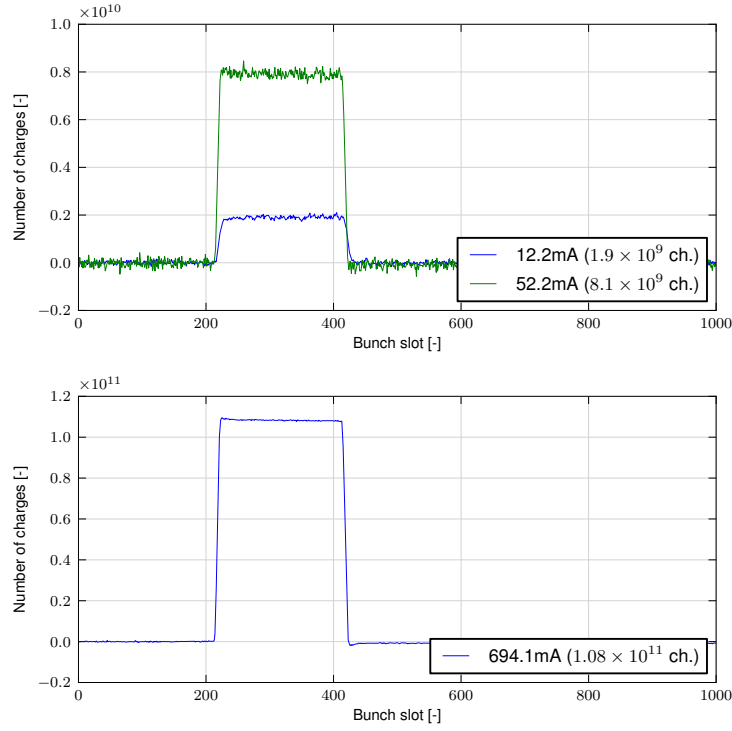


Figure 8.20: Calibration pulse acquired by the FBCT acquisition system and recalculated to the number of charges.

graph depicts the measurement of the low bandwidth low gain measurement channel.

The x-axis denotes the bunch slot. Each bunch slot corresponds to a 25 ns integration time and hence both graphs show an integral of the calibration pulse signal. The information is expressed in number of charges per bunch-slot.

Figure 8.21 shows the calibrator current setting error. A set of 16 samples was taken on selected calibration currents and standard statistical data were calculated. The red trace shows the average current setting error, calculated as a difference of the value measured by the calibrator acquisition chain, and the 'theoretical' value of the current specified by settings of the current-setting DAC. The error in the low current region is caused by various offsets in the system. The minimum current which can be set with reasonable pulse repeatability is ≈ 6 mA. Below this value the output signal is affected by noise and cannot be used for calibration. At high currents ($I_c > 850$ mA) the calibrator saturates. The grey area indicates the minimum and maximum error measured within each set of 16 samples.

The measurement confirms that the direct DAC current setting cannot be used to set the calibration current precisely. Instead, another approach must be used. Either every calibration pulse is generated and measured by the calibrator's ADC while it is acquired by the acquisition chain, and the final calibration coefficients (gain and offset) are determined using the average of calibration coefficients calculated from each calibration pulse current and the measured response of the acquisition chain individually. Or another

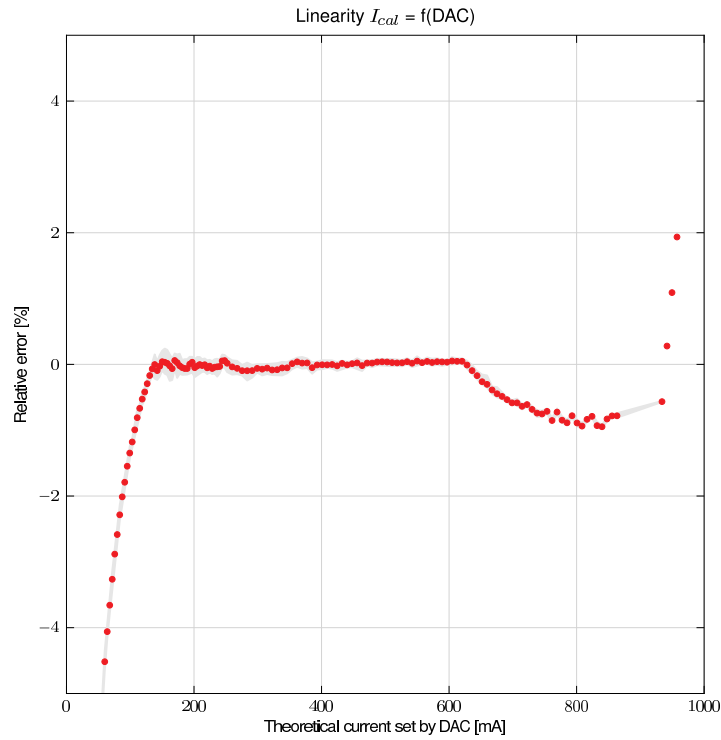


Figure 8.21: *Calibrator error setting measurement*

possibility would be to implement a current-setting procedure, which sets the calibrator current iteratively to produce calibration pulses of requested amplitude, and determine the calibration coefficients using this calibration current and the averaged data acquired by the acquisition chain.

The latter method is implemented to calibrate the LHC FBCTs.

9 ACHIEVED RESULTS

For the LHC operation a total of ten FBCTs has been built. Eight FBCTs were installed in the LHC tunnel and two FBCTs serve as spares and permit further development. Four installed FBCTs provide circulating beam measurements, another four provide measurements of the extracted beams.

The signals from the FBCTs installed in the LHC tunnel are transported using 7/8" Heliflex cables to the analogue front-ends installed in neighbour radiation shielded equipment gallery. Each front-end consists of an RF distributor, an acquisition system and a calibrator. All electronics including the acquisition systems is installed into VME64x crates located in the equipment gallery. Front-end controllers provide a connection to the LHC control system via an Ethernet interface.

The complete system was commissioned with beam during the first few days of the LHC operation in September 2008 and since then it is in a routine operation. The devices were calibrated, however at that time the correctness of their calibration could not be thoroughly verified. A test facility using the SPS type FBCT equipped with the LHC acquisition chain was constructed. As the FBCT was concurrently connected to both SPS and LHC acquisition chains, the measured values could be compared. Using this setup it was discovered that the LHC calibration was systematically optimistic by $\approx 1.9\%$ with respect to the SPS FBCT calibration and also exhibited a mean offset of 183×10^9 charges. It was found that the offset was caused by improper calibration procedure: the FBCT calibration coefficients were calculated using two-point linear fit. The fit was constructed using two calibration currents instead of an offset measurement and one calibration current. Since then the calibration procedure was appropriately changed and a systematic offset of less than 5×10^8 charges is achieved.

Nowadays the LHC DCCT measurements are used to evaluate the performance of the FBCTs. As both measurement devices are installed in the LHC tunnel on the same 12 metres long support, the beam loss on such distance is negligible and hence both systems should see the same number of charges. The DCCTs are the 'ultimate' measurement tool when considering the absolute accuracy (DCCTs: 10 % of the LHC pilot bunch and 1 % of the LHC nominal bunch measured on turn-basis [66]) so if both systems are correctly calibrated the obtained measurements results must be very similar. Such experiments were done after the LHC restart in late 2009. The results showed sometimes a difference more than 30 % between the measurements of the DCCTs and the FBCTs (Fig. 9.1). This behaviour was studied and it was found that the original FBCT calibration currents (55 mA, 120 mA and 850 mA) drove the integrators into the non-linear region and hence the calibration coefficients were incorrectly calculated. The results produced using an FBCT calibrated by the new calibration currents can be seen in Fig. 9.2.

Figure 9.2 shows the results of the DCCT measurements compared with sum-sum measurement mode of the FBCT. Both graphs show excellent matching of the measurements (measurement differ by less than 0.5 %). The top graph depicts the LHC beam 1,

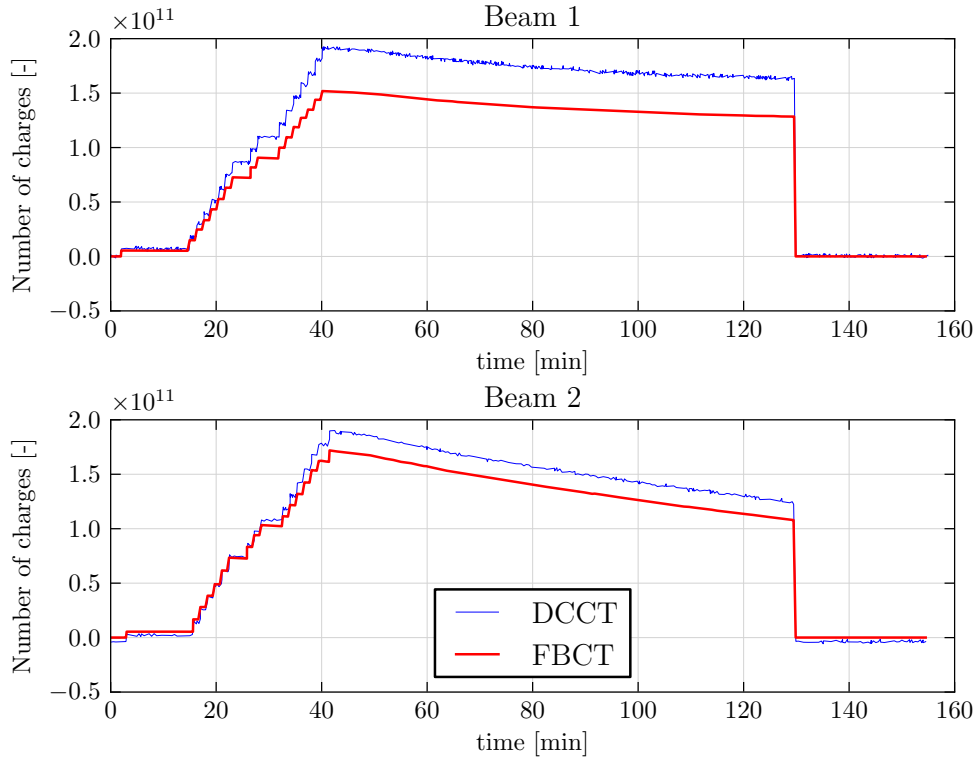


Figure 9.1: Comparison of the DCCT and the FBCT measurement using improperly calibrated FBCTs. Too high calibration currents forced the integrators to work in non-linear region. This introduced inaccuracies in the calibration coefficients calculation.

the bottom graph corresponds to the beam 2. The blue trace shows the measurement of the LHC DCCT. As the pilot bunch is at the limit of visibility for the DCCTs, the top measurement contains a lot of noise. The bottom graph shows measurement of higher intensity ($\approx 4 \times 10^{10}$ charges). Few interesting facts are visible on the bottom figure:

- Until approximately 80 min the system was slightly miscalibrated. This resulted in lower values measured by the FBCT (e.g. measurements performed at 20 min). After the recalibration the FBCT perfectly copies the measurements of the DCCT.
- The event at ≈ 135 min shows the effect of debunching. As the FBCT cannot measure debunched beam, the measured value quickly drops to zero. Comparison of FBCT and DCCT provides indication of the debunched beam circulating in the machine.
- Without beam the FBCT measures a non-zero mean value of the noise ($\approx 3 \times 10^8$ charges). This is an effect of the base-line restoration algorithm which uses minimum bunch slot measurement within a turn to provide information about the base-line. The bunch by bunch measurement is affected by this error as well, however for this mode it does not represent any issue as per-bunch offset is three orders of magnitude lower. The effect is non-cumulative and can be further suppressed using different base-line restoration algorithm.

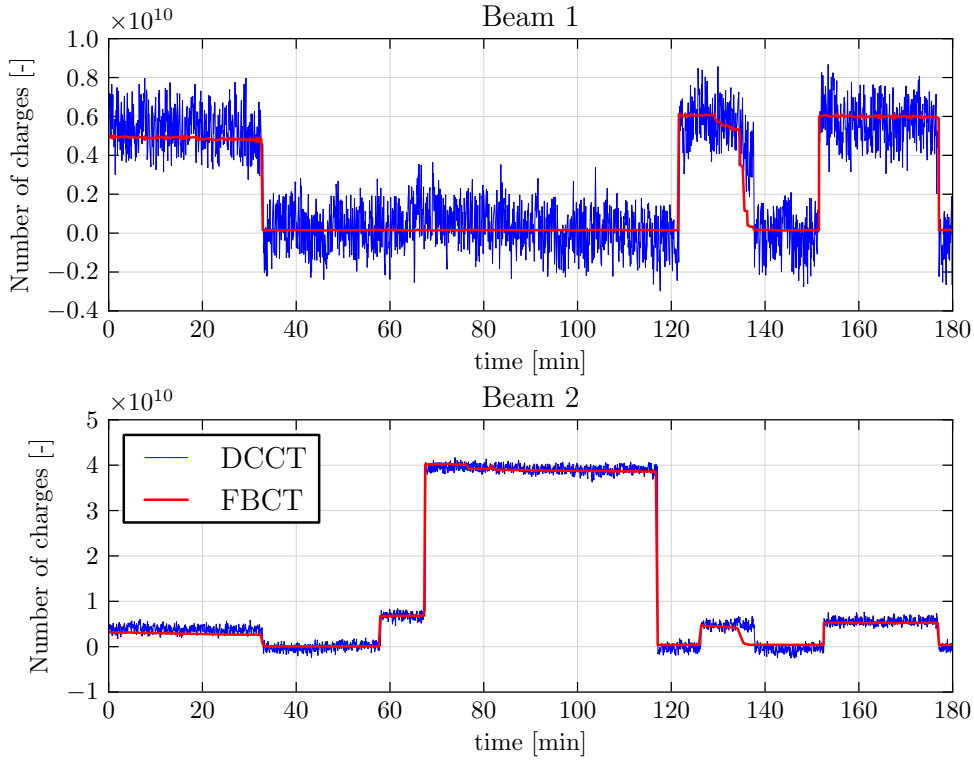


Figure 9.2: Comparison of the DCCT and the FBCT measurement on the LHC beam

Channel	Mean value [-]	Standard deviation [-]	STD relative to real FS [%]
HIGAIN HIBW	-0.7×10^6	1.3×10^7	0.09
HIGAIN LOBW	0.6×10^6	6.2×10^7	0.42
LOGAIN HIBW	10.3×10^6	1.6×10^8	0.1
LOGAIN LOBW	-1.7×10^6	1.4×10^8	0.1

Table 9.1: Noise measurement performed on single calibrated FBCT installed in the LHC. Last column shows the noise standard deviation recalculated to the percentage of the real-measured saturation of HIGAIN channels (1.5×10^{10} and 1.7×10^{11} charges)

Table 9.1 summarises the noise floor measured on a calibrated FBCT installed in the LHC. As expected, due to automatic offset suppression and proper calibration the mean value of the noise is negligible compared to the full-scale readings.

Important part of the FBCT development is a study of interference between FBCTs and other LHC equipment. The measurements of the FBCTs installed in the LHC rings do not show any perturbations due to neighbouring equipment, as seen in Fig. 9.3. However the FBCTs installed in the LHC dumps (BCTD) acquire noise induced on the vacuum chamber by pulsing of the extraction kicker magnets. Figure 9.4 shows the BCTD signal measured during kicker magnets ramping. The traces depict raw signal measured at the output of the BCTD. Blue trace corresponds to measurement using the primary toroid, red

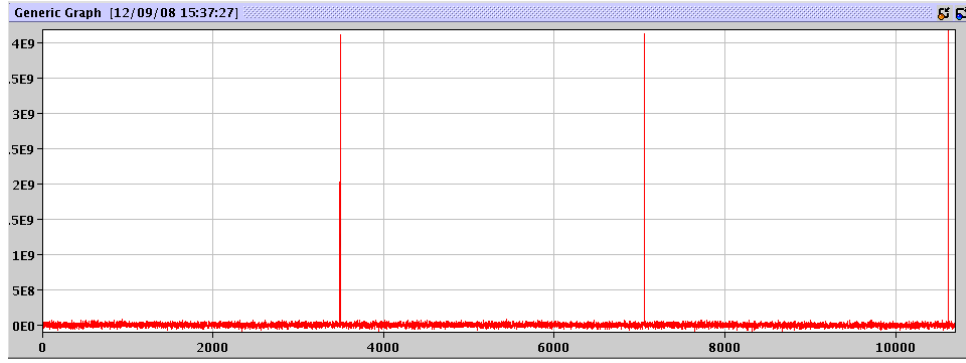


Figure 9.3: The LHC pilot beam measured during the first few days of the LHC operation. Picture is a screen-shot of the FBCT application used by the LHC operators. The x-axis is the bunch-slot number, y-axis shows intensity of the measured beam in number of charges. Three revolution periods are shown (3564 bunch slots per revolution period).

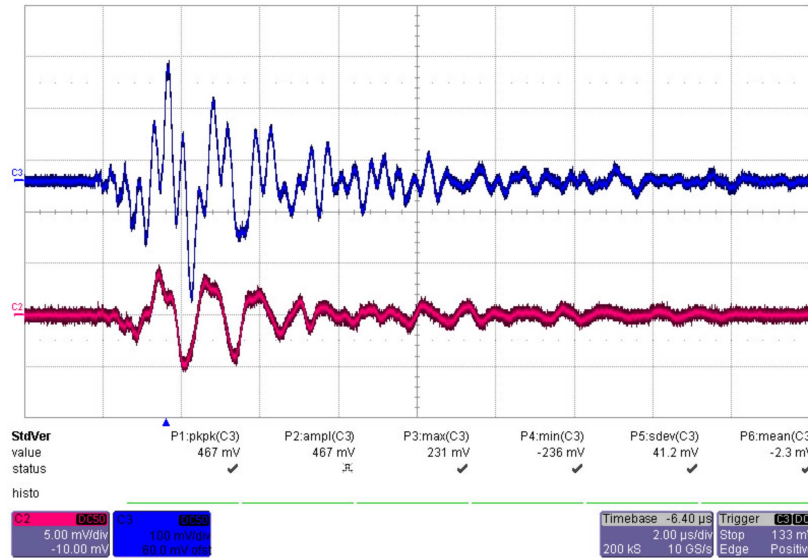


Figure 9.4: Oscillogram of the signal acquired during ramping of the extraction kicker in the LHC dump line. The acquired data show the direct output of the FBCTs

trace the secondary toroid. Peak voltage of ≈ 250 mV measured by the primary toroid is comparable to the voltage generated by the LHC pilot bunch (≈ 300 mV). During the LHC operation it was found that the kicker noise is not critical as the time difference between acquired noise and the first bunch passing through the devices is long enough to avoid the interference. However the BCTDs cabling was found more problematic. The BCTDs share the cable tray with high voltage cables supplying the sputter ion pumps. This results in a non-negligible noise superimposed on the calibration pulses and measurement signals. The noise acquired during the calibration can be suppressed using averaging. This is not the case for measured beam signal, hence the absolute accuracy of the measurement degrades by ≈ 10 %. Problem is currently under investigation and possible remedies are currently being studied. If no satisfactory solution is found using electronic suppression (e.g. ferrite rings), cabling needs to be modified during the next LHC shut-down.

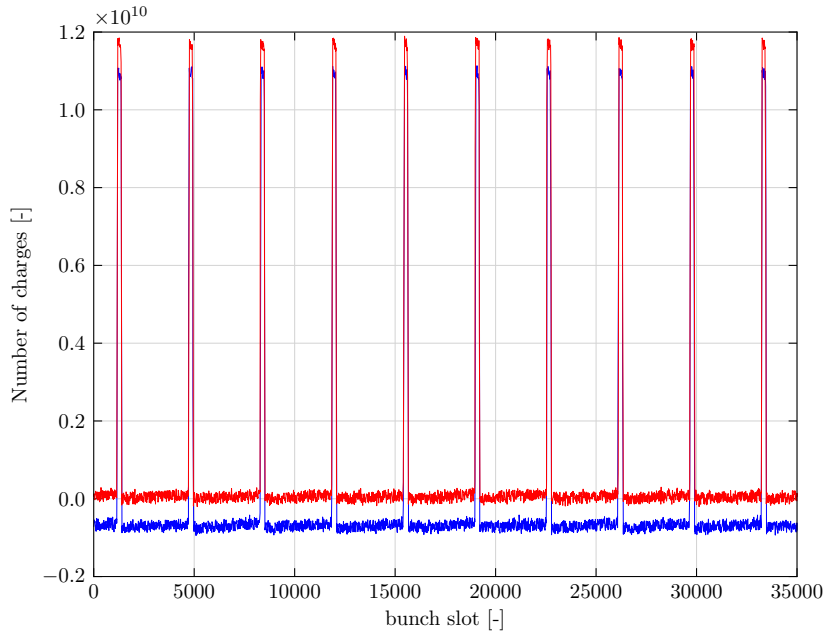


Figure 9.5: *Function of the base-line restorer.*

Figure 9.5 shows the function of the base-line restorer. A repetitive signal from signal generator was injected into the FBCT calibration turn. The data were acquired and the base line was restored. The implemented base line restoration algorithm uses an IIR filter to 'smooth' the measured signal. Each revolution period a minimum value of the filtered signal from preceding turn is determined and added to the bunch measurements from the current turn. The filter time constant is user controllable, however its value is limited to less than $3 \mu\text{s}$ in order to establish correct base-line of the filtered signal within a single revolution period. This base-line restoration method results in a displacement of the 'zero' measurement by an offset which is equivalent to a half of the filtered noise peak amplitude.

The BCF worked flawlessly already with the first LHC beams. The threshold is set to $\approx 3 - 4 \times 10^9$ charges per bunch what leaves enough margin to safely detect the LHC pilot beam. The development of the BCF detector continues to lower the system noise and thus detect beams with less than 1.5×10^9 charges per bunch. It would provide the beam circulating flag detection also for the LHC probe beam.

Apart of use in the LHC control center, the LHC intensity measurements using FBCTs is world-wide broadcasted using the LHC VISTAR page 1 (see Fig. 9.6). The same page shows also the beam circulating flag ('Beam Presence') provided by the FBCT system. For on-line information visit following web page: <http://op-webtools.web.cern.ch/op-webtools/vistar/vistars.php?usr=LHC1>

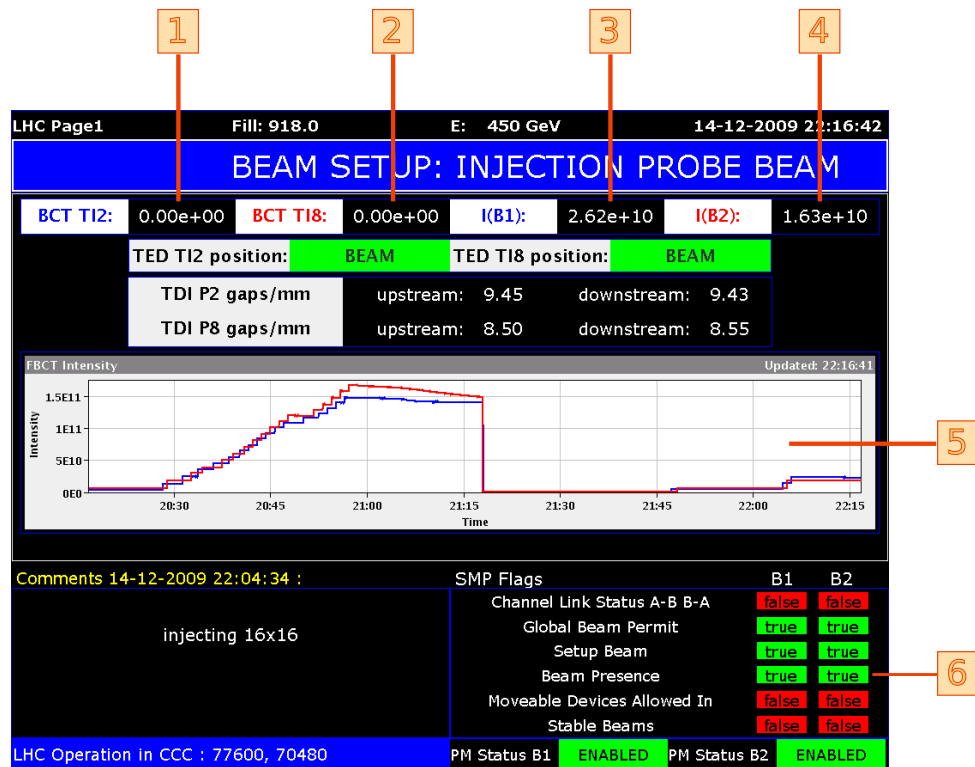


Figure 9.6: LHC VISTAR page 1, world-wide broadcasted information about current status of the LHC machine

Legend:

1. Beam 1 intensity measured in the SPS to LHC transfer line
2. Beam 2 intensity measured in the SPS to LHC transfer line
3. Beam 1 intensity measured by the DCCT in the LHC ring
4. Beam 2 intensity measured by the DCCT in the LHC ring
5. Graph of the FBCT intensity measurement in the LHC rings
6. Beam circulating flag

10 CONCLUSIONS AND OUTCOMES FROM THE DISSERTATION THESIS

The aim of this dissertation thesis was to develop a fast beam intensity measurement system which satisfies the LHC intensity measurement specification. During the development different fields of technology needed to be addressed.

An extensive mathematical model of the transformer droop putting emphasis on the bunch structure of the LHC beams must have been constructed before any hardware development. The base part of the model is definition of an equivalent bunch function (base function) which permits to predict accurately the bunch amplitude, its FWHM and equivalent charge. The defined function (Sec. 6.1.1, Eq. (6.1)) constraints the entire bunch charge into a finite space. This leads to significant calculus simplification with respect to models based on Gaussian function standardly used to model the bunch. It was shown that the base function precisely models the charge distribution and in contrast to the standard Gaussian model allows optimisation of either the bunch peak amplitude or the bunch FWHM depending on the requested priority. For the purpose of the work the FWHM optimisation was used, resulting in an increased signal peak amplitude (Fig. 6.2) introducing an additional security factor for the analogue front-end design.

The base function was used to define the number of required dynamic ranges (Sec. 6.1). Two dynamic ranges were implemented as described in Tab. 6.2. The high gain dynamic range ($FS=1.0 \times 10^{10}$ charges) permits efficient measurement of the low-intensity beams such as the LHC pilot beam and the lead ion beams. The low gain dynamic range ($FS=1.7 \times 10^{11}$ charges) measures nominal and ultimate LHC proton beams.

The measurement error was modeled and used to estimate the required FBCT bandwidth. It was demonstrated that the major error contributor to the intensity measurement is the FBCT droop, which causes 100 % error in the circulating beam measurements if not correctly compensated. DC signal restoration techniques help to overcome this problem, however they are of no use for beam injection/extraction measurement. The worst-case error corresponds to the measurement performed at the end of injection.

The FBCT droop was modeled using a first-order low-pass filter whose impulse response was convolved with the base function appropriately scaled in the time domain. Presented equations permit to estimate analytically the signal droop of a generic bunch train (specified by FWHM, number of charges, total number of bunches) acquired by an FBCT of known LF cut-off. The model is used to calculate the droop for two injection scenarios: the LHC pilot bunch (Eqns. (6.18), (6.20)), and the injection of 4 SPS batches (Eqns. (6.29), (6.42)).

Once the analytical model of the FBCT droop was known the measurement error could have been estimated. The error evaluation methods in the integration measurements require the error calculus to be split into the case where the input signal is present in the bunch slot and where only an excitation decay exists. The latter cannot be quantised in

terms of a relative error. The worst-case error calculation was performed for the case of the LHC pilot bunch (Eqns. (6.23) and (6.26)) and for injection of multiple consecutive bunches (Eqns. (6.45) and (6.49)). It was shown that the injection droop error of the LHC pilot bunch measurements is negligible for the FBCT LF cut-off in a range of few tens of a kHz (Sec. 6.2.1 b)). It was also shown, that this is not case for the multiple batch injection. In order to comply with the specification the LF cut-off lower than ≈ 400 Hz is required (Sec. 6.2.2 e)). Requirements for the single shot dump line extraction are even more restrictive as a full LHC ring can be instantly dumped. This represents a measurement of $\approx 88.9 \mu s$ long bunch train. Should the droop error be kept below the nominal 2 %, the FBCT LF cut-off of ≈ 40 Hz is required (Sec. 6.2.2 g)). Implementation of a single measurement bandwidth in a range of 40 Hz to 1 GHz could lead to a measurement precision worsening due to a noise induced from the mains supply. To overcome this the measurement profits from the fact that the dump line extraction is a single event measurement. The data can be processed off-line what allows to use transformer with only the cut-off lower than 400 Hz.

To estimate the required FBCT HF cut-off an another mathematical model has been developed (Sec. 6.2.3). It is based on the same calculus as the FBCT droop model: a first order low pass filter impulse response is convolved with the base function to estimate the bunch spread in the bunch slot. The model permits to determine analytically the cut-off frequency at which the bunch width equals to one half of the bunch slot size. Using this model it was shown that the HF cut-off higher than ≈ 60 MHz (at the input of the integrator) satisfies the defined bunch length criteria. This does not appear to be critical, however the accuracy of this model is limited by quality and length of the transmission lines connecting the FBCT to the front-end electronics. The FBCT HF cut-off also limits the phase margins between the BST clock and the incoming bunch hence it is desired to provide the measurement with much higher analogue bandwidth. This model has been extended (Sec. 6.2.3 b)) to estimate the low-pass filter cut-offs for the time-insensitive low bandwidth intensity measurements. Low-pass second order Gaussian filter has been used. Various bunch spacings were modeled and the optimum bandwidths determined. It was shown that two different cut-off frequencies must be provided: 1.5 MHz suitable for 75 ns and 100 ns bunch spacing, and 2.5 MHz for the 25 ns bunch spacing scheme.

Based on the results of the mathematical model the FBCT system was finally designed. The FBCT mechanical construction is composed of a vacuum chamber and the measurement device.

According to the information retrieved from the analytical model the FBCT device implements two toroids measuring the beam signal using different bandwidths (Sec. 7.1.2). The primary toroid, commercially available from the Bergoz Instrumentation, provides measurements in the bandwidth of less than 300 Hz to 1.2 GHz. Secondary toroid was fabricated in house using T60004-L2130-W630 core from Vacuum Schmelze and provides measurements in the bandwidth of 20 Hz to ≈ 10 MHz. Both toroids are inserted into

a shell formed by a calibration turn. A composite of μ Metal and ARMCO, appropriately annealed, houses the toroids and protects them against external magnetic field perturbations. In order to shift the parasitic FBCT cavity resonances towards higher frequencies the volume of the housing was filled by 3 mm thick PCBs plated by 35 μ m electro-gilded copper. Furthermore, excellent electrical connection for the wall image current is achieved plating the outer FBCT housing by 10 μ m of silver.

The fabrication of the vacuum chamber (Sec. 7.2.1) involves many intermediate steps including ceramics brazing, electron welding and application of various coatings. The vacuum chamber shape was optimised to provide necessary space for implementation of the heating element and the heat-sink required for bake-out. It also must not represent a constraint for the toroids' internal diameters.

Chemical pumping is used to obtain extremely high vacuum levels in the LHC. A non-evaporable getter (NEG) diffused at the interior of the vacuum chamber helps to absorb residual gases present in the vacuum system. The NEG is activated by a high temperature baking (250°C) which would damage the toroids if they would not be protected. Possible protection scenarios were studied and resulted in a design of a movable support permitting to displace the FBCT over the heat-sink before the bake-out while still maintaining highly reliable and low impedance path for the wall image current when in measurement position (Sec. 7.1.1). This is achieved using removable silver plated copper collars connecting the outer FBCT housing and the vacuum chamber by copper-beryllium RF fingers.

To remove the charge deposited on the ceramics extremities a titanium resistive layer was applied from inside. The resistance had to be experimentally determined from the frequency characteristics of the FBCT system. A coaxial line assembly was set-up to inject a bunch-like signal (Sec. 7.2.2). The shape was studied and designed to form a 50 Ω coaxial transmission line together with the FBCT vacuum chamber (Fig. 7.11). Using this test-bed it was demonstrated that the FBCT not using resistively loaded ceramics exhibits a resonance at ≈ 600 MHz (Fig. 7.12(b)). A resistive layer of ≈ 24 Ω to 28 Ω damps the resonances and still maintains the fast signal rise-time.

Both NEG and Ti coatings were applied in house at CERN. The titanium layer resistance (Tab. 7.1) increases in time due to oxidation. The deposition process duration was studied and optimised to achieve a long-term stable resistance lower than 28 Ω . An average resistance of ≈ 27 Ω was measured one year after the installation of the vacuum chambers into the LHC.

The FBCT acquisition system is based on the CERN SPS FBCT acquisition system. It was upgraded to support four measurement channels using 14bit ADCs (two dynamic ranges providing measurements in two different bandwidths). A LHCb2002 ASIC developed in TRIUMF (Canada) provides necessary 40 MHz bunch by bunch integration using two on-chip integrators working in a time multiplex. Each integrator exhibits different gain and offset so a compensation algorithm has been studied. It was found experimentally that

the best offset and gain correction is obtained measuring a calibration pulse corresponding to $\approx 85\%$ FS and using it to calculate gains and offsets of both integrators, match the offsets to zero and match the second integrator gain to be equal to the first one. The compensation algorithm is implemented as a single LUT applied to the ADC outputs what assures only single clock cycle latency when processing the data stream (Fig. 8.3 in Sec. 8.1).

The four acquisition modules are fed by an RF signal distributor which splits the incoming FBCT signal into four measurement channels, Beam Circulating Flag (BCF) detector channel and an oscilloscope output channel. To keep the signal reflections at minimum the RF signal distributor is fabricated using a composite of a RF substrate and FR4 material. This permits to maintain at the same time impedance controlled board and provide additional power supply layers. The measurement channels are protected by wide-band buffers (Fig. 8.5) providing $50\ \Omega$ termination for the resistive splitter. Filtering is provided by two fourth order active, and two second order passive Gaussian filters for low bandwidth and high bandwidth measurement channels respectively.

Mathematical model predicts that the amplitude of the incoming FBCT signal would saturate the high-bandwidth channels due to the peak signal amplitude. High bandwidth channels filters are used to reduce the signal's amplitude below the saturation level. As the behaviour of the low-pass filtering on a bunch signal was already modeled when calculating the HF FBCT cut-off, the same model was used also to predict dependence of the bunch signal peak amplitude on the filter's cut-off for specific bunch widths (Fig.8.7). Generic analytical solution of this problem is very complex so the signal amplitude for specific bunch σ and filter frequency was determined numerically. A set of Thiele interpolations was produced to provide bunch peak current attenuation as a function of filter cut-off frequency for a given bunch width. It was shown, that both low and high gain HIBW channels must be low-pass filtered using maximum 300 MHz cut-off (Sec. 8.2.2). A practical value of 200 MHz was used to implement the filters.

A lot of 20 RF distributors which has been fabricated exhibits a worst-case input reflection coefficient of $S_{11} = -25\text{ dB}$ at 800 MHz and $S_{11} = -20\text{ dB}$ at 1.2 GHz. The transmission measurements of the measurement channels are shown in Fig. 8.9. To keep the channels output offsets minimal an offset suppressor was studied. Figure 8.6 shows the final implementation, limiting the offset value to theoretical $\approx 500\ \mu V_{rms}$. In practise even better average offset of $47\ \mu V$ with $\sigma = 145\ \mu V$ was obtained, as seen in Fig. 8.10, Sec. 8.2.4.

The RF distributor noise characteristics were measured including the acquisition part of the system as this measurement permits to evaluate the system noise floor in terms of number of charges. The Tab. 8.2 summarises the measurements concluding that the high gain channels RMS noise does not exceed 0.7 % of the full scale reading.

Calibration is an essential part of the system. Calibration is based on measurement of a current pulse applied to the calibration turn. Factors translating the number

of measured ADC bins to an appropriate number of charges are calculated from the measured response and known calibration currents. The calibration system is composed of two parts: the calibrator itself and the calibration turn.

The calibration turn acts as a single turn on the toroids, inducing a calibration pulse sent by the calibrator into the measurement winding. The process is not loss-free hence an optimum transmission system had to be studied. It was shown that the amount of transferred energy from the calibrator to the calibration turn depends much on the uniformity of the current distribution around the magnetic core (Tab. 8.3). 4, 6, 8 and 16 parallel calibration turn branches were implemented of which the transmission ratio was measured. It was shown that the 16 branch calibration turn would be the best choice (giving a transmission ratio of 99.8 %), however it is not technologically realisable. Hence an 8 branch calibration turn (transmission ratio of 99.3 %) was implemented. In order to keep the signal reflections minimal the calibration turn was realised on a RF substrate providing good matching and excellent reproducibility of the fabrication process (Fig. 8.14).

The implemented calibrator (Sec. 8.4) permits to generate pulses of user-selectable amplitude (≈ 6 mA to 800 mA). It is implemented as a charge storage, which is discharged via a current source into either a dummy load or the calibration turn. This permits to achieve high slew rates as the dummy load is used to suppress the transitions during the current source setting-up.

It was shown that uncertainty of the current setting is ≈ 1.9 mA due to the used amplifiers. It was decided to implement a current-readback circuit which measures the amount of current sent into the calibration winding by measuring a voltage drop across a reference resistance installed in the feedback of the current source. A 14-bit ADC measures the voltage. In order to provide appropriate measurement resolution there are two measurement dynamic ranges implemented (≈ 150 mA with a resolution of $8.5 \mu\text{A}$, and 935 mA with a resolution of $\approx 60 \mu\text{A}$). High absolute accuracy of the measurement is assured by calibration of the ADC readout using a long-term “zero” averaging and a measurement of the calibration reference ($3 \text{ V} \pm 0.05 \%$ and derived 0.45 V). These are used to correct the ADC gain and offset errors.

The uncertainty of the calibrator’s pulse current measurement was estimated in Sec. 8.4.2. As both sensitivity ranges use different equations to provide the linear approximation, the uncertainty calculus had to be split into two parts. The results of the calculation are shown in Fig. 8.18. All used calibration currents exhibit the extended uncertainty lower than 0.8 % providing reasonable margin to satisfy the absolute accuracy requirements summarised in Tab. 4.5 and Tab. 4.6.

Totally eight FBCTs were installed in the LHC tunnel. Four FBCTs provide circulating beam measurements. Another four provide measurements of the injected and the extracted beams. The signals from the FBCTs are transported using 7/8” Heliflex coaxial cables to the analogue front-ends. Each front-end consists of an RF distributor (Sec. 8.2),

an acquisition system (Sec. 8.1) and a calibrator (Sec. 8.4). All electronics including the acquisition systems is installed into VME64x crates located in the LHC underground technical areas. Front-end controllers provide a connection to the machine control system via an Ethernet interface.

The measurement chain was successfully commissioned during the first days of the LHC operation in 2008 (Sec. 9). The system development still continues as there is a constant need for measurement algorithm improvements and perfectioning. Nowadays the system performance conforms to the specification. Comparison of DCCT and FBCT measurements for correctly and independently calibrated systems results in less-than 0.5 % measurement difference (Fig. 9.2).

10.1 My contribution to the work

The design of the fast intensity measurement system for the LHC requires wide range of knowledge from various scientific and industrial fields. I am involved in the development as a project leader who is looking for suitable solutions, presents them to appropriate experts and makes iterative steps to converge the system design to its optimal performance. This involves as well making right decisions based on the information from the experts, and the follow up of their recommendations. I am responsible also for the electronic chain design to match the required performance.

I had to solve following project tasks in order to successfully design, build, install and commission the Fast Beam Intensity Measurement System for the LHC:

- analysis and preliminary estimation of the FBCT functional and measurement parameters to be able to provide complete information to the specification committee
- complete mathematical analysis of the measurement chain: development of droop mathematical model, calculation of the signal droop for different injection scenarios, specification of measurement device bandwidth limits, specification of the dynamic ranges and bandwidths required for specific measurement channels
- estimation of optimal resistive layer to be diffused on the ceramics interior
- involved in the development of the measurement devices' mechanical construction:
 - development and implementation of the FBCT's toroids:
 - analysis, design and implementation of the in-house fabricated toroid
 - analysis and design of the toroids' wrapping and their fixation to the centre of the FBCT housing
 - analysis and design of the inner μ Metal shell and its fixation to the FBCT housing
 - iterative FBCT mechanical construction development together with the CERN engineering department
 - mathematical and mechanical study of the coaxial transmission line test-bed
 - iterative design of the vacuum chamber together with CERN engineering de-

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- partment and CERN accelerators and beam physics group
 - vacuum chambers and the FBCT outer housing (e.g. thermal treatment, surface treatment) fabrication follow-up, completion, prototype testing and evaluation, and measurement of the systems in the laboratory
 - complete design of the analogue front-end:
 - mathematical analysis, RF analysis of calibrator, RF distributor, BCF detector and calibration turn, including other system units not covered in the thesis (e.g. RF multiplexor)
 - schematic and PCB design, evaluation of the RF and FR4 PCB feasibility and manufacturability
 - design and implementation of the algorithms in VHDL
 - prototype and series production, fabrication follow up and measurements
 - analysis and specification of the calibration algorithm
 - analysis and implementation of the LUT compensation method using calibration pulse to determine the gain and offset correction factors
 - integration of the measurement system into the LHC machine and the LHC tunnel
 - installation of the measurement systems in the LHC tunnel, including electronics in-situ testing

The project design phases **I did not contribute to**: IBMS mezzanine design, DAB card design, software data treatment (with the exception of the calibration and LUT algorithms analysis and implementation, and tools to verify correct functionality of the hardware), ceramics design and fabrication and its connection to the bellows, resistive layer deposit.

10.2 Theses for future work

- Improvement of the FBCT droop restoration algorithm to suppress the offset generated by the current system implementation.
- Improvement of the BCF detector to be able to detect reliably the LHC probe beam (2×10^9 charges)
- Analysis and implementation of suitable arbiter algorithms permitting on-fly decision on a correctness of the intensity value measured.
- Analysis and implementation of an inter-connection matrix permitting to distribute different intensity values measured by different instruments to the MPS such that the MPS would always receive an intensity value measured with the best absolute accuracy available.

11 PHD STUDENT PUBLICATIONS

11.1 Publications related to the dissertation

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13 APPENDICES

13.1 Background information

Following chapters give a brief introduction to the accelerator technology what should help to better understand constraints put to the beam intensity measurement instrument.

13.1.1 Particle accelerators

A need to look deep inside the structure of the matter led physicists to search for the methods of its deconstruction. One such method is to let the particles collide. Successful collision provokes a reaction between the particles and creates new and different particles. New particles can provide an information about the internal structure of the matter.

In order to create a new particle an energy that corresponds at least to the rest energy of that particle must be produced. The mass-energy equivalence [86] shows that the rest energy is proportional to the mass of the particle. Therefore higher collision energies allow to see “heavier” particles. Another reason to increase the energy is the increase of the resolution when looking at the structure of the matter. This can be liken to the flow of the source particles serving as a ‘light’ which enlightens the target. The wavelength of the probing light corresponds to the spatial resolution of the measurement. Shorter wavelength of the light results in more detailed picture.

The wavelength of the particle (particle wave) corresponds to its energy, and can be expressed using Planck’s constant as:

$$E = \hbar\omega, \tag{13.1}$$

where $\hbar = 6.582 \times 10^{-16} \text{ eVs}$ is reduced Planck’s constant and ω is corresponding angular frequency (rads^{-1}).

The accelerators have limitations in terms of minimal energy of the particles injected, and maximum achievable energy. No single machine could cover the energy range from the particles’ rest energy up to the TeV. Therefore arranged into an accelerator chain the desired energies can be reached.

Speaking in terms of maximum currently achievable energies, the leptons could be accelerated to $\approx 100 \text{ GeV}$. The limit is imposed by synchrotron radiation losses which scale with fourth power of the particle energy [89]. The hadrons could be accelerated to TeV. The limits are given only by an intensity of a magnetic field required, as in the case of the Tevatron at FNAL, Illinois, USA. The LHC is designed to produce collisions at centre-mass energy of 14 TeV. This permits to prove a theoretical existence of the Higgs bosons.

Following sections define various accelerator structures used, and their role in the field of the accelerator technology.

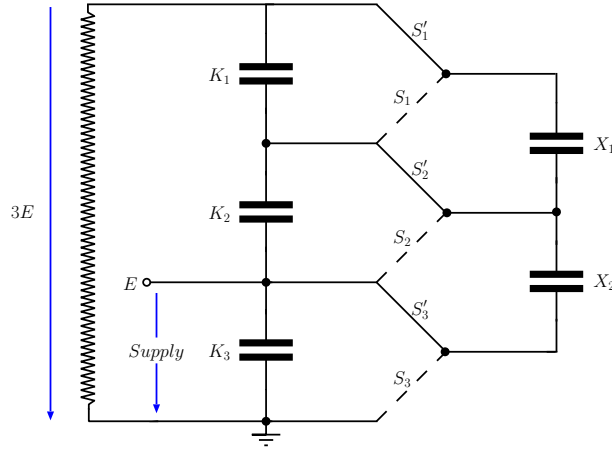


Figure 13.1: Schematic of Cockcroft-Walton's voltage multiplier. Opening and closing the switches S transfer charge from capacitor $K3$ through the capacitors X up to $K1$.

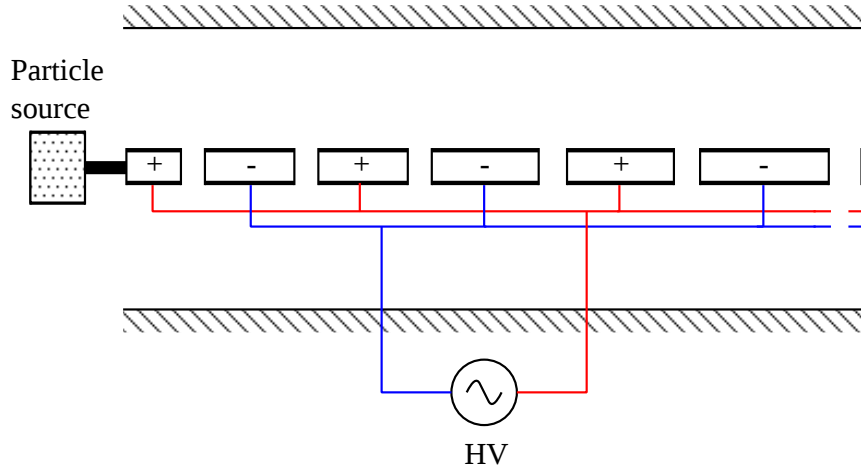


Figure 13.2: Principle of operation of Wideröe's LINAC

a) Linear accelerators

First ideas of particle acceleration were implemented by J. D. Cockcroft and E. Walton in 1930-1932 in Cavendish Laboratory at Cambridge. The protons were accelerated by an 800 kV voltage multiplier (Fig. 13.1) in a 2 metres long vacuum tube [76].

Difficulties to achieve higher voltages led physicists to use different accelerating schemes, e.g. RF acceleration by AC fields. In 1928 R. Wideröe published an article summarising the results of acceleration of ions using RF powered LINAC [1].

The Wideröe's LINAC used low voltage RF generator to create an electric field in the gaps between adjacent electrodes (Fig. 13.2). The particles located in the gaps are accelerated, they enter the electrodes, and drift to the next adjacent gaps. Due to a fixed RF frequency the electrodes must extend with an increasing velocity of the particles. The particle train leaving the LINAC has an internal structure corresponding to the used RF frequency.

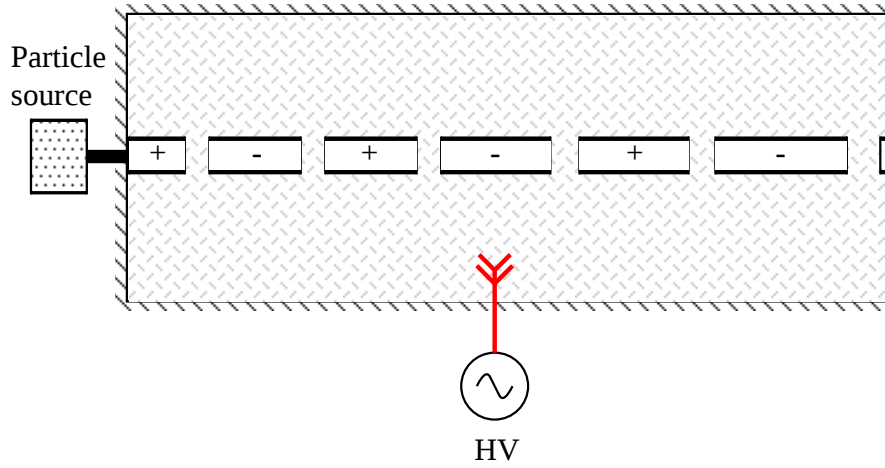


Figure 13.3: *Principle of operation of Alvarez LINAC*

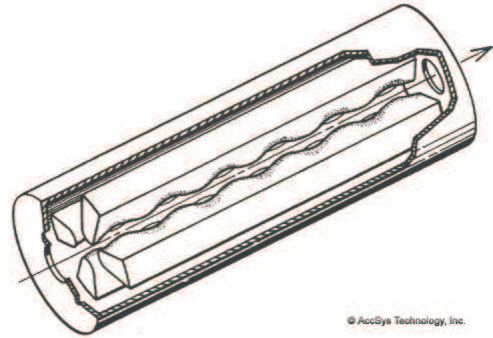
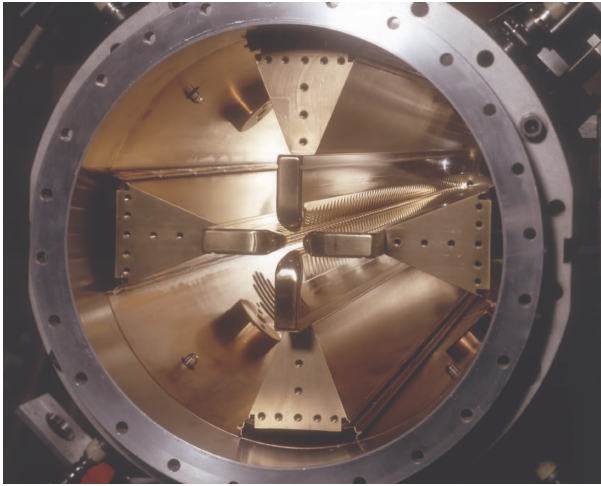


Figure 13.4: *Left image shows the RFQ structure used at CERN in years 1981-1996. It accelerated protons and negative hydrogen ions to 520 keV. Right shows the shape of the electrodes of the RFQ LINAC. (courtesy AccSys Technology)*

This accelerating structure was used in 1931 by Lawrence and Sloan to accelerate Hg^+ ions to 1.26 MeV.

A different LINAC structure was introduced by L. Alvarez in 1946, today known as a Drift Tube Linac (DTL). The principle of operation is shown in Fig. 13.3. This configuration permits to excite a standing wave electric field inside a metallic cavity. If the cavity is longer than the distance a particle travels in half an RF period, the particles are both accelerated and decelerated [91]. A set of hollow electrodes creates field-free regions, which shield the particles during deceleration. Inside the electrodes the particles drift as in case of the Wideröe's LINAC. Typical achieved linear gradients are 1-2.5 MV/m.

In 1970 I.M Kapchiski and V.A. Teplyakov used four electrodes arranged similarly to the RF quadrupole to produce an electric field. Instead of using flat electrodes the structure uses four electrodes forming a shape of hills and valleys (Fig. 13.4). This creates a non-uniform electric field in the direction of the beam propagation. Such field has not

only longitudinal component, but contains also an additional transverse components.

Longitudinal component forces the beam to be accelerated. Particles are accelerated in every gap, but they are not shielded during the drift. Instead, they are focused by a transversal electric field produced by a symmetric geometry of the electrodes. This structure permits to accelerate, bunch and focus the particles by one RF field. The structure is known as a RFQ LINAC.

The three principles are used by LINACs serving as injectors. Additional sources of information can be found e.g. in [78], [68] or [92].

b) Circular accelerators - cyclotrons

Acceleration to high energies would require excessively long linear accelerators. This is impractical therefore there were designed accelerators of particles travelling on a closed circular orbit. During the acceleration the particles re-encountered the same accelerating structure many times.

In order bend trajectory of a particle (to a closed orbit) a force must be applied. In case of circular trajectory with radius r , the magnitude of the force is given by a magnitude of the required centrifugal force:

$$F_c = \frac{mv^2}{r}, \quad (13.2)$$

where m is particle's mass and v is the velocity.

Such force can be generated using the Lorentz force. This is a force acting on a particle exposed to a electro-magnetic field perpendicular to the motion of that particle.

For the case of zero electric field it is defined as:

$$\mathbf{F}_s = q\mathbf{v} \times \mathbf{B}, \quad (13.3)$$

where $\mathbf{v}$ is a vector of velocity and $\mathbf{B}$ vector of magnetic induction of the static magnetic field.

To create a constant circular trajectory the centrifugal and the Lorentz forces must be equal in magnitude:

$$\frac{v}{r} = \frac{qB}{m} \quad (13.4)$$

Figure 13.5 depicts the principle of cyclotron operation. The RF voltage is applied to two half-cylinders separated by a gap, creating an electric field on the gap which accelerates the particles in the same way as the Wideröe's LINAC. Accelerated particles entering the metallic cylinders are shielded from the electric field. The Lorentz force bends their trajectory and forces them to make a circular movement towards the opposite half-cylinder. As the magnetic induction is constant, with every gain of energy the particles travel with larger radius. They make a spiral movement with the v/r ratio constant.

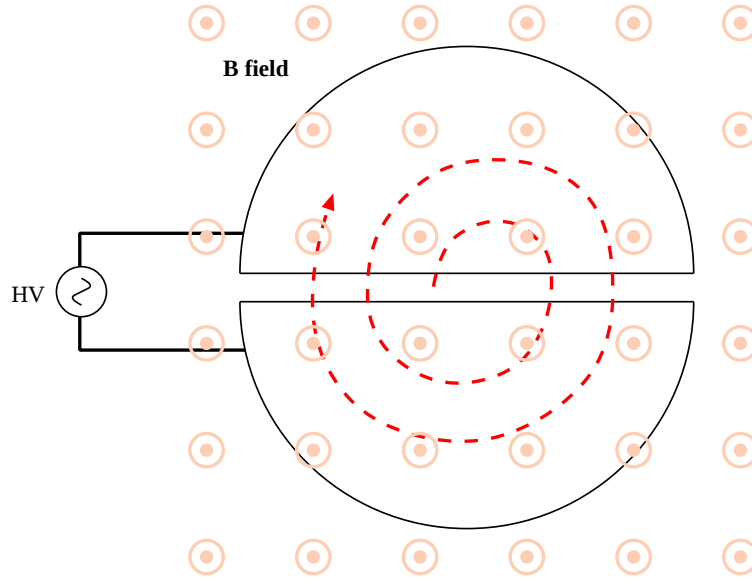


Figure 13.5: *Principle of cyclotron operation*

First cyclotron was built by Lawrence and Livingston. A 1 kV RF voltage generator working at 10 MHz was used to accelerate hydrogen ions to 80 keV in a magnetic field of 1.3 T [77].

c) Circular accelerators - synchrotrons

The maximum achievable energy is given in the cyclotrons by strength and size of the magnet. The magnetic induction of normal conducting magnets is limited to approximately 1.5 T. Hence it is a physical size of the magnet which puts the limits on the achievable energy. To overcome this problem the particles were kept on a circular orbit using many small magnets. First accelerators of this type used a static magnetic field hence the radius of orbiting particles increased with each passage through the accelerating structure. Later, when electromagnetic coils were introduced, it was possible to control the field strength synchronously with the particle energy gain keeping the revolution radius constant (Fig. 13.6 at right).

Using the closed orbit the particles could revolve in the machine as many times as needed. The only limitation was imposed by a magnitude of the magnetic field, and a need for an initial energy of the particles injected.

Electric field in the accelerating RF cavity must be properly synchronised with the orbiting particle revolution frequency. The red dots in Fig. 13.6 indicate the locations where the direction of the electric field coheres with the direction of the moving particle. It is required to capture maximum of the particles at those places. Particles which are not captured cannot be accelerated. They orbit in the machine uncontrollably and they are considered as lost.

The synchrotrons are widely used due to numerous advantages: only a simple vacuum chamber is required, they accelerate high intensity beams to high energy, and they can be used as storage rings as well.

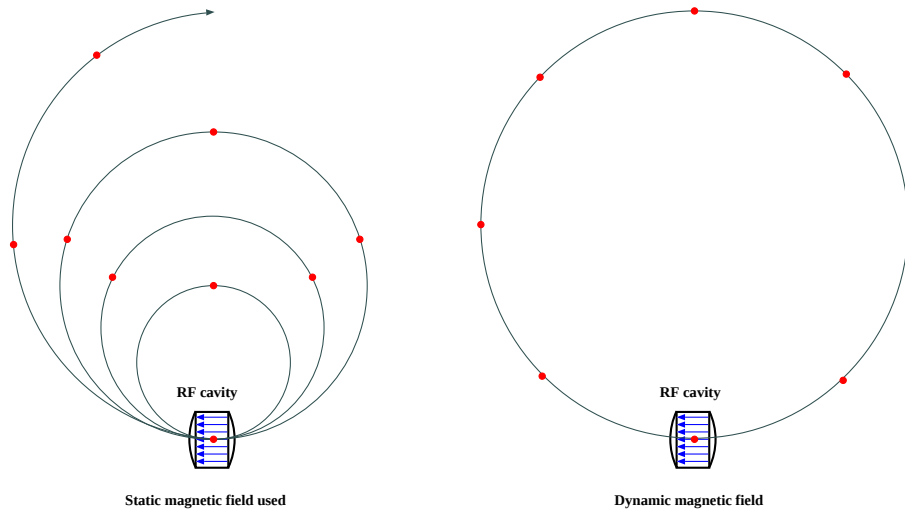


Figure 13.6: *Two operational modes of synchrotrons. The figure on the left shows the synchrotron using static magnetic field. The radius of orbiting is increased in order to compensate the magnetic field bending force and centrifugal force. The figure on the right shows a synchrotron having fixed radius. The forces are compensated by an increase of the magnetic induction.*

13.2 Simplified mechanical acceptance calculation for the BC-TRs

Detailed calculation of the mechanical acceptance is discussed in [57] and [28]. A transversal beam size at the locations of the magnets surrounding the BCTRs must be calculated in order to express the acceptance defining the minimal BCTR vacuum chamber diameter. The beam size is defined from the normalised emittance of the LHC:

$$\epsilon_n = 3.75 \mu\text{m} \cdot \text{rad} \quad (13.5)$$

Normalised emittance divided by relativistic γ determines nominal emittance. As the LHC beam energy varies (450 GeV to 7 TeV), the emittance value for the beam energy of 450 GeV is taken into account as it determines a maximum transversal beam size:

$$\epsilon = \frac{\epsilon_n}{\gamma} = \frac{\epsilon_n}{E_p/m_p} = \frac{3.75 \mu\text{m} \cdot \text{rad}}{450 \text{ GeV}/0.9383 \text{ GeV}} = 7.82 \text{ nm} \cdot \text{rad} \quad (13.6)$$

where E_p is the energy of the beam and m_p is the mass of the proton. The transversal RMS beam size is:

$$\sigma_{x,y} = \sqrt{\epsilon \cdot \beta_{x,y}} \quad (13.7)$$

where $\beta_{x,y}$ is the optical β -function specific for given machine location. Figure 13.7 defines parameters to calculate the mechanical aperture.

Horizontal scale of the graph in Fig. 13.7 shows a distance relative to the LHC “start” in the IP1. The BCTRs are located approximately 10154 metres away of IP1.

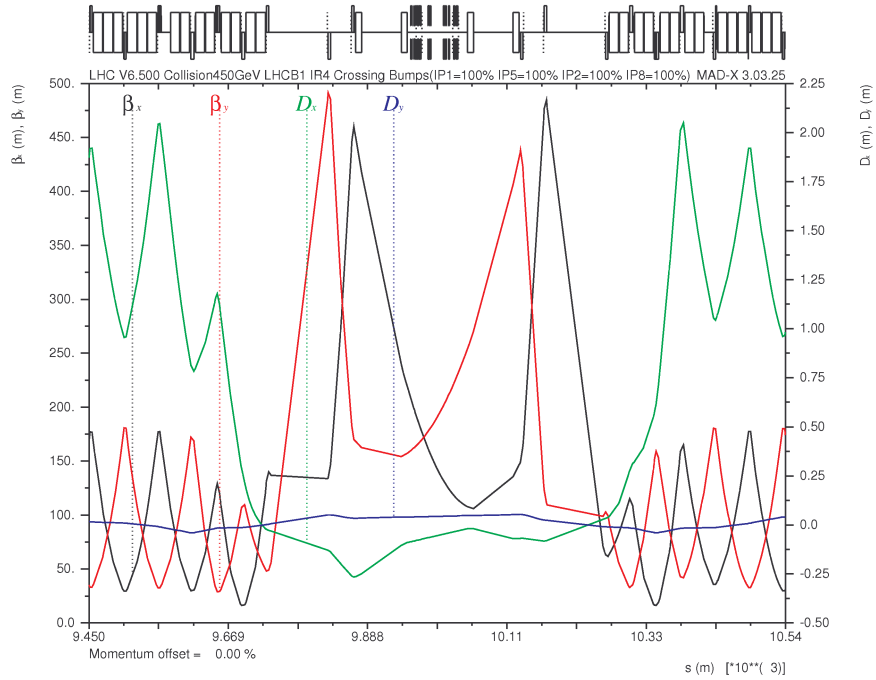


Figure 13.7: The optical β -function and nominal dispersion function $D(s)$ at the IP4. The calculation corresponds to the LHC optics version 6.500 at 450 GeV.

Looking at graph the worst-case β -function could be estimated as:

$$\beta_x \approx 427 \text{ m} \quad (13.8)$$

$$\beta_y \approx 187 \text{ mse} \quad (13.9)$$

The beam size at the injection can be estimated substituting the superior value of Eqns. (13.9) into Eq. (13.7):

$$\sigma_x \approx 1.83 \text{ mm} \quad (13.10)$$

The total mechanical aperture is determined by the LHC collimators, whose aperture decreases with the beam size in order to damp the particles outside of defined transversal range. The aperture of the primary LHC collimators is seven sigma of the beam transverse size [57], secondary collimators are set to 7/6 of the primary collimators aperture:

$$n_1 = 7\sigma_x \quad (13.11)$$

$$n_2 = \frac{7}{6}n_1 \quad (13.12)$$

The secondary halo boundary, where the particles cannot cause a quench, was found by a numerical analysis of the trajectory of the particles colliding with the secondary collimators. This aperture limit is recalculated in terms of n_1 as:

$$n_r = 1.4n_1 \quad (13.13)$$

Equation (13.13) shows that the limiting aperture is approximately ten sigma of the beam transversal size:

$$s_x = 10\sigma_x \approx 18.3 \text{ mm} \quad (13.14)$$

In order to take into an account the beta-beating effect the beam size must be enlarged by a beta-beating factor $1 + k_\beta$. Its value was estimated as 20 percent of the nominal beam size, hence the estimated maximum beam RMS size is:

$$s_{x,\beta} = s_x (1 + k_\beta) = 18.3 \cdot 1.2 = 21.9 \text{ mm} \quad (13.15)$$

The calculated value corresponds to a beam-appearance radius, providing that the beam orbits on a theoretical trajectory. This is not the case due to various errors. Major error contributor is a transversal closed orbit error. The value is specified as a peak closed orbit excursion, and acts as an offset to the calculated beam size aperture. Generally referred value of $s_{CO} = 4 \text{ mm}$ is used. It is deduced by an experience made during the LEP run.

Second error contributor is a transverse quadrupole magnets misalignment, causing a parasitic dispersion, a margin which must be added to the nominal dispersion shown in Fig. 13.7. For the BCTRs a nominal dispersion of $D_{x,nominal} = -0.08 \text{ m}$ applies. The

parasitic dispersion is of the order of 27 % of its value in the arcs of the rings, weighted by the β -functions, i.e. [57]:

$$D_{parasitic} = k_D \sqrt{\frac{\beta_x}{\beta_{arcs}}} \cdot D_{arc} = 0.86 \text{ m} \quad (13.16)$$

with $\beta_{arc} = 170 \text{ m}$, $D_{arc} = 2 \text{ m}$, $k_D = 0.27$ and β_x as defined in Eq. 13.8. The total dispersion contributes to the mechanical displacement as:

$$s_d = (|D_{x,nominal}| + |D_{parasitic}|) \cdot \delta_p = 0.94 \cdot 1.5 \times 10^{-3} = 1.41 \text{ mm} \quad (13.17)$$

where δ_p is the momentum offset at injection. The displacement of the beam centre from its ideal orbit is estimated summing the s_{CO} and s_d together with a mechanical tolerance defined by experience as $s_m = 3 \text{ mm}$:

$$s_{total} = s_{CO} + s_d + s_m = 8.41 \text{ mm} \quad (13.18)$$

Sum of the estimated beam size and corresponding displacements determines a minimum vacuum chamber radius:

$$r_{min} = s_{total} + s_{x,\beta} = 8.41 + 21.9 = 30.3 \text{ mm} \rightarrow d_{min} = 60.6 \text{ mm} \quad (13.19)$$

The indicated value is approximative, a lot of simplifications were assumed compare to [28].

Focusing magnets surrounding the BCTRs use the vacuum chamber with internal diameter of 63 mm. The BCTRs must not be a limiting element in the vacuum chamber aperture, hence 64 mm vacuum chambers were used providing enough space for the installation of the bake-out and cooling circuitry.

13.3 Photographs

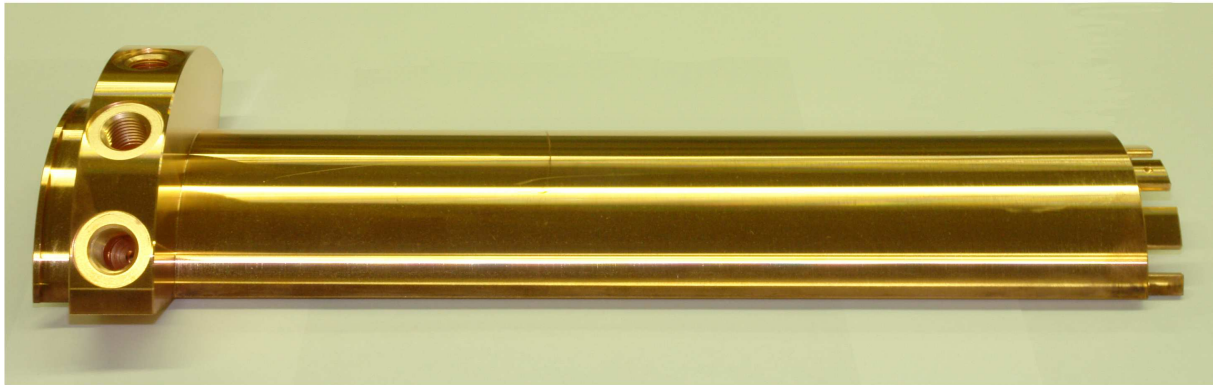


Figure 13.8: *Realised one third of the BCTR's heat-sink.*

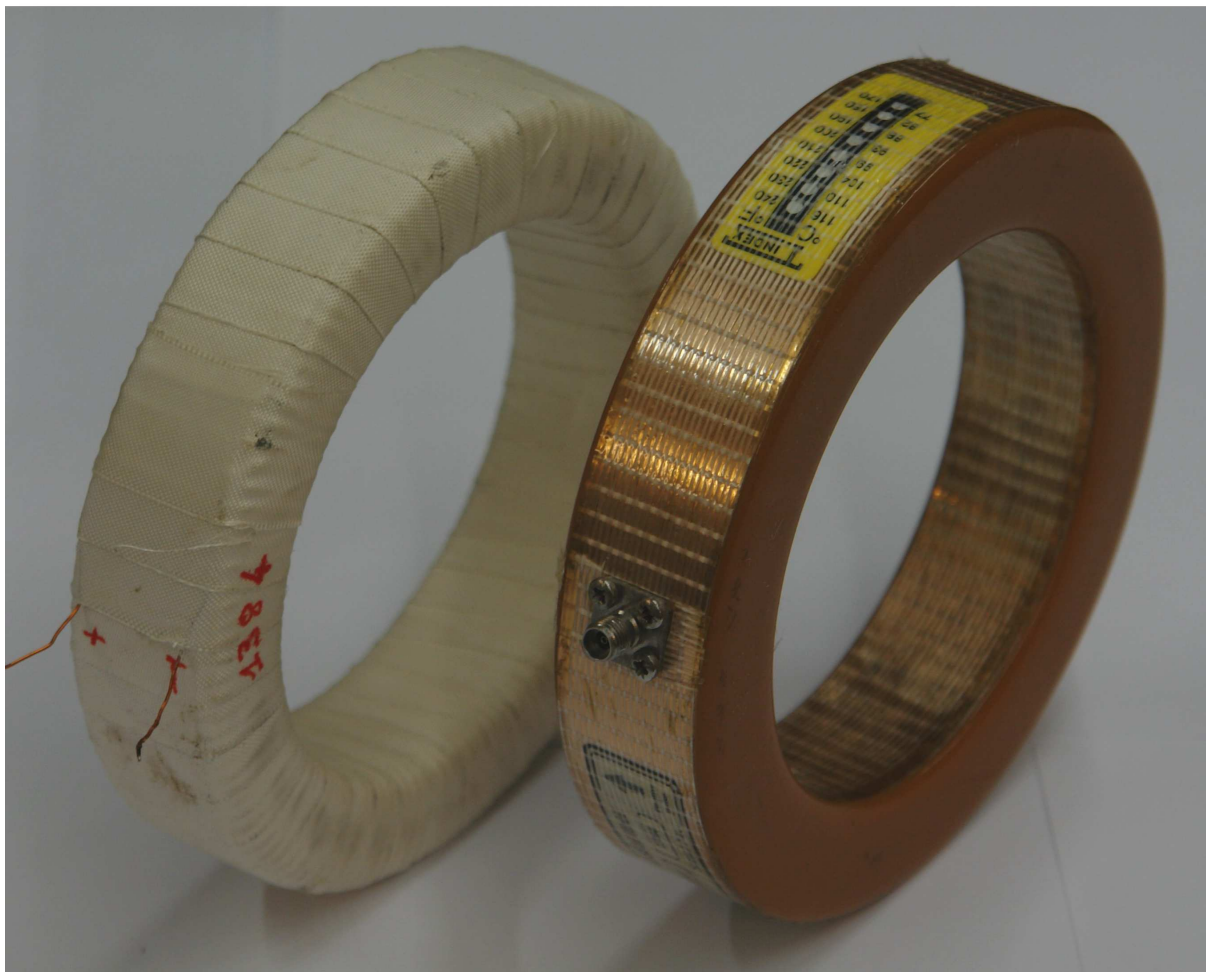


Figure 13.9: *Two FBCT toroids prepared for the installation*



Figure 13.10: *The Mu-metal static magnetic protection*

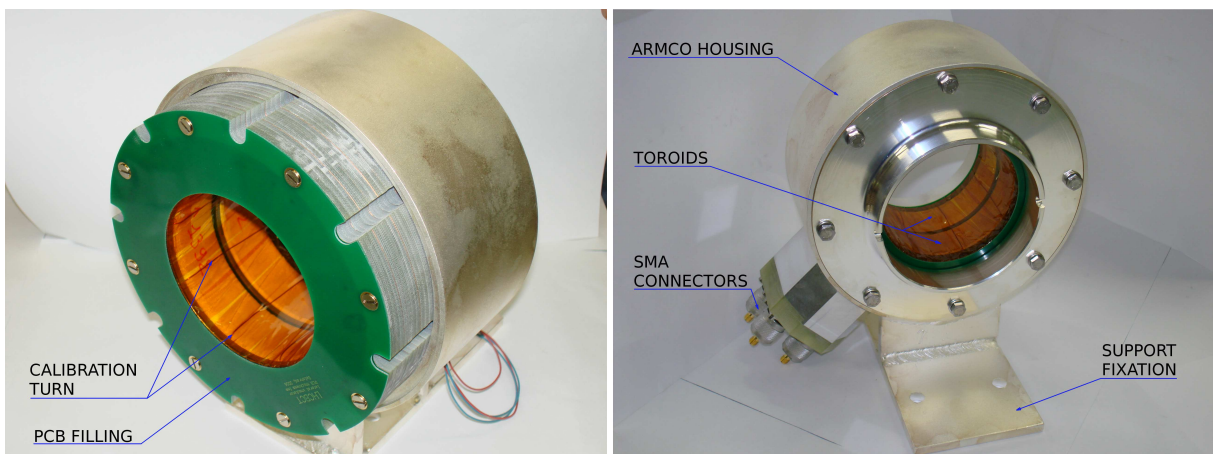


Figure 13.11: *Complete assembly of the FBCT for the LHC. The picture left depicts the internal structure of the assembly. Filling material is visible. Thin wire at the interior of the measurement device is a part of the calibration winding.*

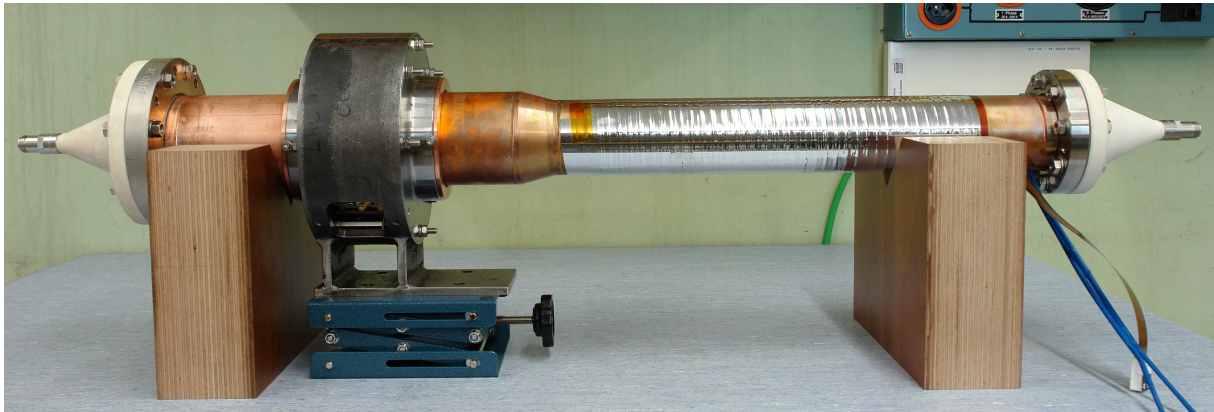


Figure 13.12: *The assembly of the FBCT, vacuum chamber and the measurement antenna. The picture was taken at the time of mechanical tests. During the measurement a final version of the FBCT was used (finished sand-blasting, thermal treatment and electro-plating). The silver coloured tape at the right side of the vacuum chamber protects the installed heating element.*

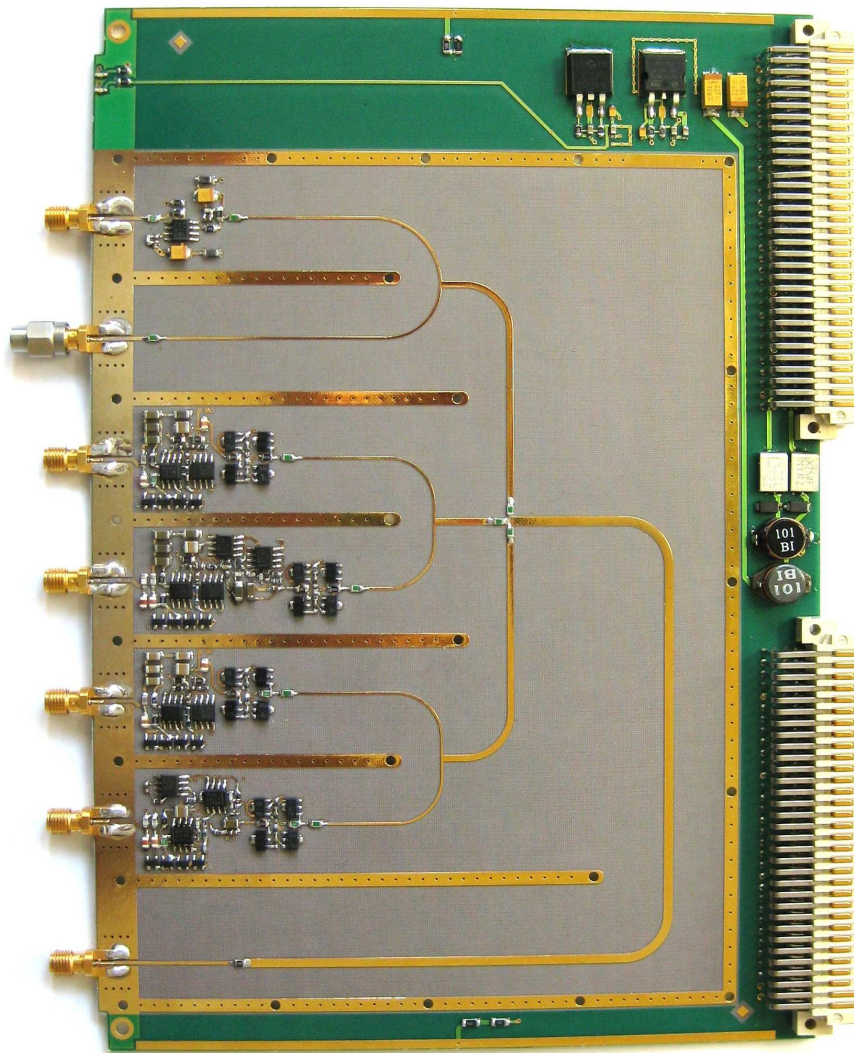


Figure 13.13: *RF distributor. The brass shielding (not shown) is installed over the PCB part not covered by the solder stop mask.*

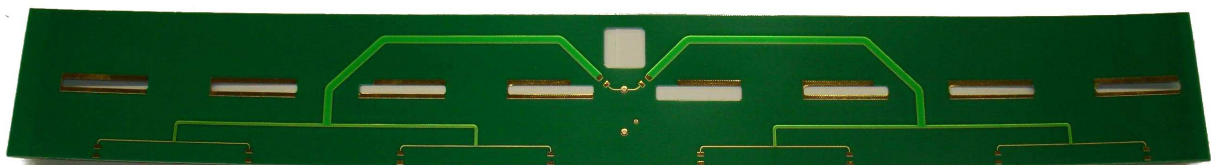


Figure 13.14: *Realisation of the 8-branch FBCT calibration turn*

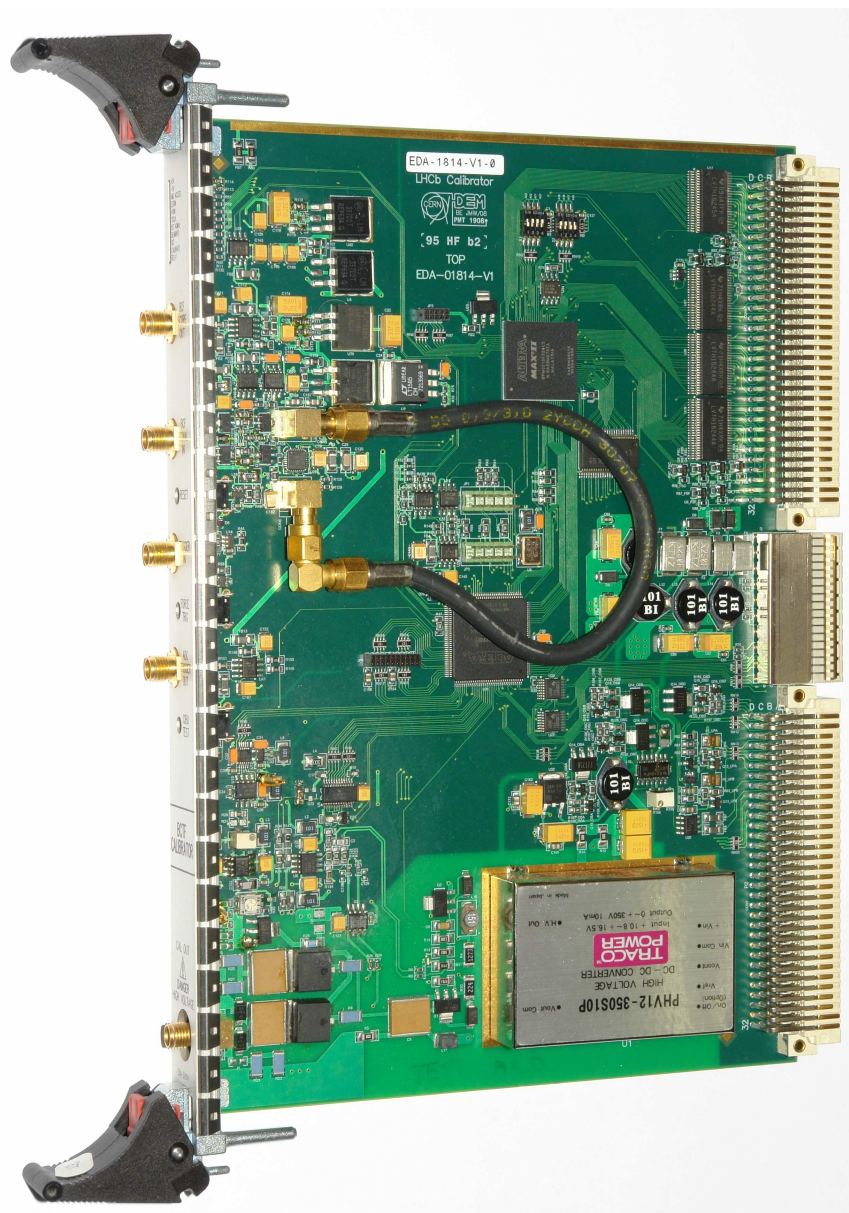


Figure 13.15: *Implementation of the calibrator and BCF detector.*