THEORETICAL AND EXPERIMENTAL ANALYSIS OF INTERACTIONS BETWEEN ELECTROMAGNETIC FIELDS AND RELATIVISTIC ELECTRONS IN VACUUM CHAMBER

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ANNO ACCADEMICO 2011-2012
Abstract

The Free Electron Laser (FEL) is a fourth generation light source that has more stringent specifications with respect to the third generation light sources, such as synchrotrons. The so-called emittance and the beam trajectory determine the beam quality, and must satisfy stringent requirements in FELs. For example, in the undulator hall, the beam position must be measured with the $\mu$m resolution. The control in the beam position can be achieved using a cavity beam position monitor (Cavity BPM). This thesis describes the research performed on the cavity BPM. Precisely, the electromagnetic design, the simulation and the optimization of a cavity BPM have been carried out. Subsequently, 25 cavity BPMs have been manufactured and installed in the undulator hall of the FERMI@Elettra project [1] [2]. A new RF front-end has been set up, and a series of measurements have been performed. The second device studied in this PhD is the travelling wave linear accelerator. Traditional accelerating structures endowed with a single feed coupler cause degradation of the electron beam properties, due to the electromagnetic field asymmetry. A new type of single feed structure with movable short circuit is proposed, where the electric field has been symmetrized. The electromagnetic design, simulation and optimization of the device have been carried out, and a prototype of the accelerating structure has been produced and tuned. The electric field has been measured with the bead-pull method. Finally, in this thesis are described the High Energy RF Deflector (HERFD), which are a fundamental diagnostic tool to measure the electron beam properties, in particular the bunch length and the longitudinal phase space.
Acknowledgments

I would like to warmly thank:

Roberto Vescovo, Paolo Craievich, my supervisors, always ready and helpful to give me the right suggestions, always supporting me in my work, in the conferences and in the accelerators schools.

Alessandro Fabris, Michele Svandrlik, for the support in all my research work, in the conferences and in the accelerators schools. The research presented in this thesis was made possible thanks to the financial support of Elettra - Sincrotrone Trieste S.C.p.A.

Marco Petronio, Alberto Lutman, the previous PhD students, they gave me a lot of precious information about electromagnetic fields, accelerating structures and FEL process.

Raffaele De Monte, Giulio Gaio, Michele Predonzani, Andrea Borga, Mario Ferianis, Claudio Scafuri, for the support in the studies, cables, materials, low level software and graphic user interfaces of the Cavity BPM system, accelerating and deflecting structures.

Defa Wang, Dave Whittum, Eric Colby, Giuseppe Penco, Simone Di Mitri, Enrico Allaria and all the accelerator group for the precious informations about linear accelerators.

Ivan Cudin, Maurizio Barnaba, all the staff of the Elettra mechanical workshops, for the realization of the linac prototype, for all the mechanical support in the linac, the cavity BPM movers and supports.

Dino Zangrando, Davide Castronovo, William Fawley, Claudio Serpico, Federico Gelmetti, Massimo Milloch, Paolo Del Giusto and all the Linac group for all the information, help and support in the and deflector installation, commissioning and in the linac studies.

Luca Rumiz, Furio Zudini and all the vacuum staff for the tools needed in the linac measurements.
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Introduction

This thesis describes the research performed during my PhD in Electromagnetic Fields, supported by Elettra - Sincrotrone Trieste S.C.p.A [1] [2]. The research topics involve the interaction between relativistic electron bunches and electromagnetic fields, for the FERMI@Elettra project [1] [2] [3] [4]. The FERMI@Elettra project is an international facility, with the objective of realizing a Free Electron Laser (FEL) light source. The produced photons have excellent qualities, properties, in terms of coherence, brilliance and photon flux. The emitted wavelength can vary from the extreme ultraviolet (EUV) to the X-ray spectral regions. Such light is used ultra-fast and ultra-high resolution processes in material science and physical biosciences, to analyze molecules and molecular processes with very high resolution. The quality of the microscopy is related to the temporal and spatial coherence of the FEL pulses. The produced light can be also used for medical purposes, to obtain tomographies with higher resolution with respect to the ones obtained with the traditional light sources.

The FEL process requires accurate studies on the machine components. This thesis describes the studies and the research made on the following FEL devices:

1. The cavity beam position monitors (Cavity BPM)
2. The RF linear accelerators (Linac)
3. The high energy RF deflectors (HERFD)

The cavity BPM is a beam diagnostic instrument able to measure the electron beam position in a non-destructive way, within the micrometer resolution. All the development and the commissioning processes of these devices are described, starting with the cavity design, and ending with the evaluation of the measured performances. The design, simulation and optimization of
the cavity BPM have been realized using the simulation softwares Ansoft HFSS [5] and CST Particle Studio [6]. A prototype has been manufactured at Cinel Strumenti Scientifi srl, Padova [7]; it has been tested with a vector network analyzer (VNA), and in the linac tunnel, in presence of the electron beam. Subsequently, a series of 25 cavity BPM has been produced, working at 6.5 GHz. They have been tested at the workbench and successfully installed in the undulator hall of the FERMI@Elettra project. The cavity BPM are assembled on mechanical movers, able to shift the transversal position of the cavities. The whole system is in vacuum, and a frequency tuning has been performed with the vector network analyzer. The research work also involves a detection mode, achieved with a new type of electronic circuit. Previous front-ends use the IQ demodulation to obtain the position signals, with a complex device architecture. In this thesis we introduce an innovative approach, based on a sum and difference detection method, achieved with a 180° hybrid. Such approach has the advantage of reducing the circuit complexity, leading to a simple system, easy to be set-up and commissioned. The entire system has been tested in presence of the electron beam. A real-time data acquisition system allows to simultaneously acquire the cavity BPM data of many electrons bunches. Two graphic user interfaces (GUI) have been developed. One is used to calibrate the cavity BPM, by using the mechanical movers. The other GUI is used to evaluate the resolution of the measurement system.

The second device studied in the PhD is the travelling wave linear accelerator, which is used to accelerate the electron beam. The aim is to design and to realize a prototype of the new “Linac 0” of the FERMI@Elettra project. Traditional accelerating structures endowed with a single feed coupler cause deflection and degradation of the electron beam properties, due to the electromagnetic field asymmetry. The main contribution to the beam deflection is due to the dipolar and quadrupolar component of the electric field. Presently, a dual feed coupler is adopted to symmetrize the coupler field, but such a device requires a power splitter and a dual waveguide feed. So, the cost and the complexity of such device is not negligible. A new type of single feed structure with movable short circuit is proposed, where the dipolar field has been reduced with the symmetrical geometry of the input coupler, which has one arm input and a movable sort-circuit placed on the opposite side. The electromagnetic design, simulation and optimization of the device have been carried out. This linac works at the FERMI@Elettra working
frequency, which is 2998.01 MHz. A prototype of the accelerating structure has been produced by the Elettra - Sincrotrone Trieste mechanical workshop and tuned. By using a VNA, the longitudinal and transversal electric field measurements have been performed with the bead-pull method. The longitudinal field measurement ensures that the linac has been correctly tuned and that it works with the correct phase advance per cell. The transversal field measurements allow to evaluate the electric field asymmetry in the input and output couplers.

The last topic of this thesis concerns the high energy RF deflectors, called HERFD. Such devices are used to measure the electron beam length and its longitudinal parameters, such as the longitudinal phase space. Two HERFDs are placed after the linac 4, and deflect the electron bunch in the YZ and XZ plane, respectively. The working frequency is 2998.01 MHz. This thesis shows the preliminary RF measurements carried out with the vector network analyzer, in order to verify that the devices work with the correct phase advance. The RF measurements have been carried out in the PMB-Alcen manufacturing company [8]. Subsequently, the two HERFDs have been delivered to Elettra, installed and conditioned in the linac tunnel. The conditioning, the breakdown rate and the electron beam measurements have been reported.
Chapter 1

The FERMI@Elettra project

1.1 An introduction to FERMI@Elettra

The FERMI@Elettra project [1] [2] [3] [4] has the objective of realizing a photon source with excellent characteristics concerning the flux, the coherence and the brilliance. The emitted wavelength varies from the UV to the X-rays. Such radiation is emitted by an electron bunch and its duration is of the ps order of magnitude. The FERMI@Elettra light source is a powerful research tool for scientific activities in miscellaneous sectors. Furthermore, the high brilliance allows to study non linear processes. The time properties of the photon pulse will allow the study of quick dynamical processes, concerning the electrons and the nucleus of the atoms. The FERMI@Elettra layout is depicted in Fig. 1.1.

Figure 1.1: Layout of the FERMI@Elettra facilities.
The machine can be represented in the following sections:

1. The photo-injector
2. The linear accelerator
3. The undulators
4. The light beamlines

The photo-injector is the first part of the machine. It is composed by a GUN and by two accelerating sections. The electron bunch is generated with the photoelectric principle. The photoinjector laser produces an UV light beam that hits a copper cathode. The extracted charge can be chosen in the range from 100 pC to 1 nC. The GUN is endowed with a RF standing wave cavity, 1.6 cells long, which accelerates the electron bunch. The input power is around 12 MW, and the bunch is accelerated at 5 MeV. The first two accelerating structures are called also “Linac 0”, and are illustrated in Fig. 1.2.

![Accelerating structures of the Linac 0.](image)

Their purpose is to accelerate the bunch, making it go away from the space-charge regime. After such acceleration, the electron energy reaches about 100 MeV. The photoinjector must not degrade the beam quality, measured with the beam emittance [9]. Therefore the electron bunch must be accelerated without receiving transversal kick or without being deflected. The
factors that must be considered in the cavity design are the space-charge (in the gun and in the machine parts with low energy beam), the field asymmetries and the wakefields. The design of the two accelerating structures of the RF photoinjector is explained in chapter 3. The following part of the machine, called the “Linear accelerator”, must accelerate and compress the electron bunch. The beam quality must be preserved, controlling the emittance and the energy spread parameters. The linear accelerator is composed by four linacs and by two bunch compressors. The four accelerating sections are called “Linac 1”, “Linac 2”, “Linac 3” and “Linac 4”. The final electron energy is 1.2 GeV. In the linear accelerator there are two bunch compressors: one is placed between “Linac 1” and “Linac 2”, and the other between “Linac 3” and “Linac 4”. They are composed by a dipole chicane that longitudinally compresses the electron bunch. At the end of linac 4, there are two high energy RF deflectors. They have the aim of controlling the electron beam quality before entering in the undulator hall. The HERFD is reported in Fig. 1.3 and is explained in chapter 4.

Figure 1.3: High energy RF deflector.

The electron beam is transported from the linac tunnel to the undulator hall with the spreader. The latter is essentially a beamline that connects the end of the accelerator with the undulator entrance. The spreader allows choosing one of the two FEL emitting lines, which are the FEL 1 or the
FEL 2 undulator chains, where the FEL process occurs [10] [11] [12]. In the FERMI@Elettra Project the free electron laser will operate in the 100-20 nm wavelength region in FEL 1 and down to 4 nm in FEL 2. Both FEL lines are based on the principle of high gain, harmonic generation free electron laser amplifier employing multiple undulators, up-shifting an initial seed laser having a 240-300 nm wavelength. In the first part the electron beam will interact with a first undulator called “Modulator”, which will induce an energy modulation. Then the magnetic field of a dispersive section will convert the energy modulation into spatial modulation at the harmonics of the wavelength of the seed laser. Thus the bunched electrons emit coherent radiation in a second undulator (the “Radiator”) tuned at a higher harmonic corresponding to the desired FEL output. During the FEL process, in the undulator hall, the electron beam trajectory must be kept under control with the \( \mu m \) resolution. Devices able to satisfy this constrain are the so called cavity BPMs (Beam Position Monitor). Fig. 1.4 shows a cavity BPM installed in the undulator hall.

![Cavity BPM](image)

Figure 1.4: *Cavity BPM.*

Chapter 2 explains the design, production and installation of the cavity BPM for the FERMI@Elettra project. The FEL radiation is carried on from the undulators to the experimental halls with the PADReS system (Photon Analysis, Detection and Reduction systems). Three beamlines
are designed in order to serve three separate endstations dedicated to different scientific areas: Low Density Matter (LDM), Elastic and Inelastic Scattering (EIS), and DIffraction and PRojection Imaging (DIPROI). The extreme brightness and high coherence of FERMI will enable the collection of single-pulse coherent diffraction patterns, providing structural information of objects. Key experiments will cover a selected range of scientific fields, including topics of physics, material science, catalysis, biochemistry as well as aerosol/atmospheric chemistry.
Chapter 2

The cavity beam position monitors

2.1 Overview of the beam position monitor devices

The beam position monitor devices (BPMs) are used to determine the transversal position of the electron beam in vacuum beampipe. One of the main challenges in the synchrotrons and in the free electron lasers community is to determine such position with high accuracy.

There are several kinds of devices used for this purpose. They are based on the wakefield generated by the transit of a ultra-relativistic electron beam [13]. In fact, an electron bunch travelling at high speed (usually with $v \approx c$), generates an electromagnetic field (Fig. 2.2).

Figure 2.1: Beampipe with electron beam.
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

Figure 2.2: Example of the wakefield generated by an electron beam.

The generated wakefield is then used to determine the position of the electron beam, by determining its trace on a transversal plane XY (Fig. 2.3).

Figure 2.3: X and Y position of the beam.

2.1.1 Buttons BPMs

The buttons BPMs [14] [15] are made with four buttons placed in the beampipe, as shown in Fig. 2.4.

The four buttons are electrodes that receive the beam wakefield. The closer is the distance between the beam and one electrode, the higher is the signal collected by the respective electrode. It is possible to estimate the position of the bunch by elaborating the amplitude of the four signals.

The advantages of the button BPM are:

- they are easy to build;

- they do not affect the electron beam, because the geometry of the vacuum chamber is unchanged.
2.1. OVERVIEW OF THE BEAM POSITION MONITOR DEVICES

The disadvantages are:

- they have low output signal in single bunch;
- they have low resolution in single bunch.

Due to the low resolution in single bunch operation, they cannot be used in a single-pass free electron laser, because such FEL device operates with a single electron bunch, with low repetition rate. However, these devices are well employed in synchrotrons, because the same electron bunch crosses the BPM several times per second (due to the circular path of the electrons). In this way the output signal is emphasized. Moreover all the techniques of average and noise reduction can be applied.

2.1.2 Microstrip BPMs

The microstrip BPMs [16] [17] are made with four microstrip lines located inside the beampipe, as shown in Fig. 2.5.

The four microstrip lines resonate when the electron bunch crosses the BPM. The closer is the distance between the beam and one of the microstrips, the higher is the output signal collected from such microstrip. Elaborating the amplitude of the four collected signals, it is possible to estimate the position of the bunch. This kind of devices can reach a resolution of 10 µm in single shot. A better resolution can be achieved with a new type of BPM, called “Cavity BPM”; where the term “Cavity” indicates the presence of a resonant cavity. This last kind of device is nowadays an important research field. It is described and analyzed in detail in the following chapters.
2.2 The cavity BPM

In order to obtain higher resolutions, in single shot, the cavity BPM has been used. It mainly consists of a resonant cavity and of some waveguides. The resonant cavity is located along the beampipe, as shown in Fig. 2.6.

The electron beam can be modeled as a current distribution. When it passes through the cavity, it excites its resonant modes (Fig. 2.7).

The first four modes in this type of cavity are:

- $TM_{010}$, also called “monopole”;
- $TM_{110}$, also called “dipole”;
- $TM_{210}$, also called “quadrupole”;
- $TM_{020}$, also called “second monopole”.
2.2. THE CAVITY BPM

The electric field magnitude of such modes (simulated with Ansoft HFSS [5]), is depicted as follows:

![Electric field magnitude plots](image)

The first two modes are the most relevant ones in this kind of device, and they must be considered when the cavity is used as beam position monitor. The signal of the resonant modes must be collected. Four rectangular waveguides are coupled with the resonant cavity [18], [19], [20] and receive the signal (Fig. 2.8).
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

2.2.1 The dipole mode

The $TM_{110}$ (dipole mode) is the most important mode of the cavity, and all the working operation of the BPM is related to it. In fact, the intensity of the dipole mode is proportional to the offset of the electron beam from the center axis of the beampipe (in the linearity range). There are two orthogonal polarizations: one is excited when the electron beam has a ‘X’ offset, the other when the electron beam has an ‘Y’ offset. The E-field configuration of the two dipole polarizations is shown in Fig. 2.9.

Figure 2.9: Dipole X(horizontal) and Y(vertical) polarization of the E-field.

The signals of the two polarizations (horizontal and vertical) must be separated in order to have one signal for the ‘X’ displacement and one for
2.2. THE CAVITY BPM

The separation is achieved with the cavity-waveguide coupling, where details are explained later on. The frequency target for the dipole is 6.5 GHz; hence the cavity will be designed and simulated to determine a particular radius that will satisfy such specifications.

2.2.2 The monopole mode

The $TM_{010}$ mode (monopole) is an unwanted mode in the cavity BPM, because its signal voltage is only proportional to the beam intensity and does not depend on the beam position. The magnitude electric field configuration of the monopole is shown in Fig. 2.10.

![Monopole electric field.](image)

The BPM is designed with a dipole resonant frequency of 6.5 GHz, and the monopole resonates at 4.6 GHz. Since the beam travels near the axis of the beampipe, the output signal of the monopole is very strong, while the dipole signal is very weak. Fig. 2.11 gives an example of such inequality.

In order to separate the $TM_{010}$ from the $TM_{110}$ signal, a band-pass filter centered at the dipole frequency (6.5 GHz) can be used. Additional solutions can be exploited for this purpose, in the electromagnetic design of the cavity, through the following expedients:

- Using the high-pass behavior of the rectangular waveguide;
- Using the magnetic coupling in the cavity-waveguide transition.

2.2.3 Rejection of the monopole mode with the waveguide cut-off frequency

The unwanted $TM_{010}$ mode covers the weak signal of the position-sensitive $TM_{110}$ mode. The first way to reject the monopole mode is the use of the cut-off frequency of the four rectangular waveguides. In fact, they behave as
a high-pass filter, and the cut-off frequency for the dominant mode (TE\textsubscript{10}) is given by:

\begin{equation}
 f_L = \frac{c}{2a} 
\end{equation}

where \( c \) is the speed of light in the vacuum, and \( a \) is the maximum transversal dimension of the waveguide (Fig. 2.12). According to (2.1) the cut-off frequency is 5 GHz, so the unwanted TM\textsubscript{010} at 4.6 GHz is rejected.

### 2.2.4 Rejection of the monopole and separation of the two dipole polarizations with the magnetic coupling

The second way for rejecting the monopole exploits the cavity-waveguide coupler. This is called “Magnetic coupler” [18], [19], [20], [21], because only the magnetic field of the dipole (TM\textsubscript{110}) will couple with the waveguide. To analyze this kind of monopole rejection, let us study the field configurations in the cylindrical cavity resonator.
2.2. THE CAVITY BPM

The fields of the $TM_{010}$ mode are given by:

\[ E_z = C J_0 \left( \frac{j_0 \omega r}{R} \right) \]
\[ H_\phi = -i C \frac{\omega e a R}{j_{10}} J'_0 \left( \frac{j_0 \omega r}{R} \right) \]  

while the expressions for the $TM_{110}$ mode are:

\[ E_z = C J_1 \left( \frac{j_{11} \omega r}{R} \right) \cos(\phi) \]
\[ H_r = -i C \frac{\omega e a R^2}{j_{11}} \frac{J_1 \left( \frac{j_{11} \omega r}{R} \right)}{r} \sin(\phi) \]
\[ H_\phi = -i C \frac{\omega e a R}{j_{11}} J'_1 \left( \frac{j_{11} \omega r}{R} \right) \cos(\phi) \]  

where $J_0$ and $J_1$ are the Bessel functions of the first kind and order zero and one, respectively, $R$ the cavity radius, $r$ the radial position, $\phi$ the azimuth angle of a cylindrical coordinate system where z-axis coincides with the axis of the cavity, and $C$ is a constant. These equations are a good approximation of the field configuration of the cylindrical resonator located along the beampipe. The dipole magnetic field lines, for an ‘X’ displacement of the electron beam, are reported in Fig. 2.13. In order to describe the magnetic coupling, Fig. 2.14 shows one coupler placed at $\phi=90^\circ$. It is important to see (Fig. 2.14) that only $H_r$ has the same direction as the $H$ field.
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

Figure 2.13: Dipole magnetic field lines

Figure 2.14: Magnetic coupling.

of the waveguide, in the coupling volume cavity-waveguide. Therefore, a resonant mode in the cavity can generate the $TE_{10}$ mode in the waveguide only if the resonant mode has a radial ($H_r$) component in proximity of the waveguide. Therefore the coupling is called “Magnetic coupling” [18], [19], [20], [21], where $H_r$ (in the cavity) couples with $H$ in the waveguide. There are three important issues to analyze:

1. Effect of the monopole mode;
2. Effect of the dipole mode generated by an ‘X’ beam displacement;
3. Effect of the dipole mode generated by a ‘Y’ beam displacement.

The first case considers the effect of the monopole, which is always present for any displacement of the beam, and has a high intensity. However, it has
only the $H_\phi$ component, but not the $H_r$ component. Therefore, the magnetic coupling that works with $H_r$, will not transfer energy to the rectangular waveguides, as shown in Fig. 2.15.

Figure 2.15: Magnetic coupling with monopole.

In the second case the electron beam has an ‘X’ offset, and the dipole mode is oriented in the horizontal polarization. The magnetic coupling transfers energy only to the two vertical waveguides; therefore only port 1 and 3 will have an output signal, as shown in Fig. 2.16.

Figure 2.16: Magnetic coupling with ‘X’ dipole.

In the third case the electron beam has a ‘Y’ offset, and the dipole mode is oriented in the vertical polarization. The magnetic coupling transfers energy only to the two horizontal waveguides; therefore only ports 2 and 4 will have an output signal.
The magnetic coupling has the following advantages:

- It rejects the monopole mode
- It separates the horizontal polarization (whose signal exits only through ports 1 and 3) from the vertical one (whose signal exits only through ports 2 and 4).

2.2.5 The reference cavity

With reference to Fig. 2.16, the magnitude of the output signal of ports 1 and 3 is proportional to the absolute value of the X position, while the magnitude of the output signal of ports 2 and 4 is proportional to the absolute value of the Y position. When the beam trajectory passes from a positive offset (e.g., $X = 0.1 \ mm$), to a negative offset (e.g. $X = -0.1 \ mm$), the phase has a $180^\circ$ change when the beam is passing through the zero position. In order to measure both the positive and the negative beam offset, this phase inversion must be detected. To this purpose, an additional reference cavity is needed. The reference cavity works with a monopole mode at 6.5 GHz. Such signal is independent of the beam position. An example of the signal generated by the reference cavity, and of the signal generated by the position cavity, with both a negative and a positive beam offset, is illustrated in Fig. 2.17. The electronic system receives three signals: those related to the X and Y position, and the reference signal, as depicted in Fig. 2.18. With the reference signal, the electronic system is able to recognize positive and negative beam offsets. The traditional way to achieve such purpose, is based on the use of the IQ (In-phase, in-quadrature) demodulation [22] [23]. However, in this thesis we will propose an alternative approach that is simpler and totally innovative.

2.2.6 Output voltage on a matched load

When the beam excites the cavity resonant modes, an output signal is generated. With the equations below, it is possible to estimate the signal levels generated by the reference cavity (working with the $TM_{010}$ mode) and by the position cavity (working with the $TM_{110}$ mode). The energy losses [24] of a bunch of charge $q$ that crosses a pillbox cavity, for the $TM_{010}$ mode and the $TM_{110}$ mode, are given, respectively, by Eq. 2.4.
2.2. THE CAVITY BPM

Figure 2.17: Output signals from the reference and the position cavities.

Figure 2.18: Cavities and electronic system.

\[ \Delta U_{010} = q^2 k_{010} \quad (TM_{010}) \]

\[ \Delta U_{110} = q^2 k_{110} x^2 \quad (TM_{110}) \]
where \( k_{010} \) and \( k_{110} \) are the so called “loss factors”, defined as:

\[
k_{i10} = \frac{\omega_{i10}}{2} \left( \frac{R}{Q} \right)_{i10}
\]

where \( i = 0, 1 \) (reference and position cavity, respectively), while the resistance \( R \) is meant to be the circuit resistance defined from the circuit theory, and not by the linac convention.

The output power of the cavity [24] is therefore (for the reference and the position cavity, respectively):

\[
P_{010,\text{ext}} = \frac{\omega_{010} \Delta U}{Q_{\text{ext}}} = \frac{\omega_{010}}{Q_{\text{ext}}} k_{010} q^2
\]

\[
P_{110,\text{ext}} = \frac{\omega_{110} \Delta U}{Q_{\text{ext}}} = \frac{\omega_{110}}{Q_{\text{ext}}} k_{110} q^2 x^2
\]

The output voltage on a matched \( Z_0 \) load is therefore (for the reference and the position cavity, respectively):

\[
V_{010,\text{out}} = \sqrt{2Z_0 P_{010,\text{ext}}} = \sqrt{2Z_0 \frac{\omega_{010}}{Q_{\text{ext}}} k_{010} q}
\]

\[
V_{110,\text{out}} = \sqrt{2Z_0 P_{110,\text{ext}}} = \sqrt{2Z_0 \frac{\omega_{110}}{Q_{\text{ext}}} k_{110} q x}
\]

where the voltage has the meaning of maximum amplitude value, and not of rms value.

### 2.2.7 Linearity range of the position cavity

The cavity BPM works in a satisfactory way, as a diagnostic tool, only in the linearity range, that is, when the electron bunch travels with a small offset with respect to the beam pipe axis. The linearity range can be estimated by analyzing the field functions of the \( TM_{110} \) mode in a cylindrical resonator (2.3). In the linear range the intensity of the dipole mode is proportional to the bunch offset \( r \). Expanding \( J_1(x) \) as a Taylor series, yields:

\[
J_1(x) \approx \frac{x}{2} - \frac{x^3}{16} + ...
\]
We want to evaluate when the cubic term is no larger than 1%, 5% or 10% compared to the linear one. For linearity in the range below 1%, the following inequality holds:

\[
\frac{x^3}{16} < \frac{1}{100} x
\]  

(2.9)

which yields \(x < 0.2\), where \(x = j_{11} \cdot r/R\), \(j_{11} = 3.8\) and \(R = 26.29\) mm. This yields \(r < 1.5\) mm.

Table 2.1 summarizes the maximum beam offset that gives a non-linear term below 1%, 5% and 10% of the signal.

<table>
<thead>
<tr>
<th>Linear range</th>
<th>Linear range [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1%</td>
<td>1.50</td>
</tr>
<tr>
<td>5%</td>
<td>3.37</td>
</tr>
<tr>
<td>10%</td>
<td>4.74</td>
</tr>
</tbody>
</table>

### 2.2.8 Undesired tilt signal

The beam can pass through the cavity in three different ways:

- with an offset (Fig. 2.19a);
- with a trajectory angle (Fig. 2.19b);
- with a beam head-tail angle (Fig. 2.19c).

The cavity BPMs are devices sensitive to the beam position, to the trajectory tilt and to the beam head-tail angle. The electron beam can show a combination of the three crossing ways. We are only interested in the beam offset measurement. Using Eq.(2.7), for the position cavity signal, let us write the output voltage, in the case of only offset, as:

\[
V_{\text{cav,a}} = C q x
\]  

(2.10)

where ‘C’ is a constant depending on the cavity RF parameters. We will compare this result with those of the other two cases.
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

Figure 2.19: Electron beam crossing the cavity in different modes.

**Beam trajectory angle**

When the beam is crossing the cavity with a trajectory angle (Fig. 2.19b), the contribution to the cavity voltage is given by [22] [23]:

\[
V_{\text{cav,b}} = j C q \frac{2 \tan(\alpha)}{k^2 L} \left( \sin\left(\frac{kl}{2}\right) - \frac{kl}{2} \cos\left(\frac{kl}{2}\right) \right)
\]  

(2.11)

This is an imaginary quantity. This means that this unwanted signal is in quadrature with respect to the desired one of Eq. (2.10).

**Beam head-tail angle**

Considering a finite bunch length, a tilted beam bunch excites the dipole mode even when it passes the cavity center with a correct beam trajectory. It also originates from the phase transition of the passing beam, and the contribution to the cavity voltage is given by [22],[23]:

\[
V_{\text{cav,c}} = j C q \omega \sigma_z^2 \tan(\theta)/c \]  

(2.12)

This is an imaginary quantity. This means that also this unwanted signal is in quadrature with respect to the desired one of Eq. (2.10). This important result leads to use the IQ demodulation to separate the undesired in-quadrature signal by the beam offset signal, that is in-phase. However, the IQ demodulation is not needed if the desired signals are negligible with respect to the beam position signal. In the latter case, in order to achieve 1 \(\mu\text{m}\) of resolution, the trajectory tilt angle \(\alpha\) must be less than 0.1 \(\text{mrad}\).
2.3 The electromagnetic design

This section describes the simulations and the improvements performed for the cavity BPM [25] [26]. The reference and the position cavities have been designed to have a high time constant \( \tau \), as the RF frontend needs a long duration signal. In this way the needed sampling frequency can be obtained using low cost electronic devices. On the other hand, \( \tau \) and the quality factors of a cavity are related as follows:

\[
\tau = \frac{2Q_L}{\omega} \tag{2.13}
\]

where \( \omega \) is the angular frequency and \( Q_L \) is the loaded quality factor, given by:

\[
\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{\text{EXT}}} \tag{2.14}
\]

with \( Q_0 \) and \( Q_{\text{EXT}} \) the internal and external quality factors, respectively. By (2.13), a high value of \( \tau \) is achieved with a high quality factor \( Q_L \). This requires a high \( Q_{\text{EXT}} \), by (2.14), since the \( Q_0 \) cannot be changed for a fixed material and frequency.

Both the reference and the position cavities have been designed and analyzed with Ansoft HFSS [5] and CST Particle Studio [6]. The parameters of interest are, for both cavities, the resonant frequencies, the internal and external \( Q \), the \((R/Q)\) ratio, the \( \beta \) coupling factor, the loss factor and the level of the output signals. The HFSS simulator has been used to analyze the RF parameters of the cavities, while the CST Particle Studio has been used to estimate the output signals of the cavities.

2.3.1 The position cavity

The geometry of the position cavity is illustrated in Figure 2.20.

Theoretical design of the position cavity

This paragraph describes the design of the position cavity resonator. In the FERMI@Elettra project, the first prototype worked with a 6.48 GHz resonant frequency. The design starts with the analysis of the cylindrical pillbox, that is the simplest structure of the cavity BPM and can be treated analytically (Fig. 2.21).
Since the cavity L/R ratio is lower than 2.03, the first resonant mode is the $TM_{010}$. It follows the dipole mode $TM_{110}$, which is the mode of interest. From the electromagnetic theory, the resonant frequency of the dipole is given...
by:

\[ f_{TM110} = \frac{c \cdot j_{110}}{2\pi \cdot R} \quad (2.15) \]

By the above relation, the radius \( R \) can be found. For the first prototype, which works at 6.48GHz:

\[
R = \frac{c \cdot j_{110}}{2\pi \cdot f_{TM110}} \approx \frac{3 \times 10^8 \text{m/s}}{6.48 \cdot 10^9 \text{Hz}} = 28.2 \text{mm} \quad (2.16)
\]

This value gives an approximate value of the cavity radius. However, the complete position cavity resonator is inserted in the beampipe and is connected to the four rectangular waveguides. The field is perturbed by these new elements. To correctly design the radius of the device, a simulation performed with Ansoft HFSS [5] is needed. With a radius \( R = 26.4 \text{ mm} \) the simulation with HFSS gives \( f = 6.484 \text{ GHz} \). For the final cavity BPM series of the FERMI@Elettra project, the resonant frequency must be 6.5 GHz. The new radius is derived by considering that, for small variations, the product \( R f \) is constant, giving the new radius: \( R = 26.29 \text{ mm} \).

**Simulation results of the position cavity**

The simulations exploit the symmetry of the structure with respect to orthogonal planes, using the “Eigenmode” and “Driven modal” settings. The eigenmode setting is used to estimate the resonant frequency of the cavity and the internal quality factor \( Q_0 \). The driven modal setting is used to estimate the reflection coefficient \( S_{11} \) of the output port, thus estimating the external quality factor [24] and the output voltage, with Eq. 2.7. The external quality factor \( (Q_{EXT}) \) refers only to one of the output ports, arbitrarily chosen. For a bunch charge of 1 nC, we estimated the RF parameters of the position cavity. The results are listed in Table 2.2.

The level of the output signal has been estimated using CST Particle Studio, assuming a bunch charge of 1 nC, a bunch length of \( \sigma_z = 8 \text{ mm} \), and a beam offset of 1 mm. Figure 2.22 represents the simulated output voltage of the position cavity.

The dimensions of the position cavity are listed in Table 2.3.

The bunch offset measurement exploits the dipole mode that arises in the cavity when the beam passes off-axis. In order to obtain a high quality factor, such mode must be trapped in the resonant pillbox to avoid its propagation.
Table 2.2: HFSS simulations for the position cavity. “E.” stands for “Eigenmode”, “D.” for “Driven Modal”, and “+ant” for “plus antenna”

<table>
<thead>
<tr>
<th>Symmetry planes</th>
<th>90° (E.)</th>
<th>180° (E.)</th>
<th>180° (D.)</th>
<th>N. (D.)</th>
<th>N.+-ant (D.)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{RES}$ [MHz]</td>
<td>6514.1</td>
<td>6507.7</td>
<td>6508.8</td>
<td>6507.7</td>
<td>6507.3</td>
</tr>
<tr>
<td>$Q_0$</td>
<td>7870</td>
<td>7856</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$S_{11}$</td>
<td>-</td>
<td>-</td>
<td>0.907</td>
<td>0.9055</td>
<td>0.908</td>
</tr>
<tr>
<td>$\beta_{PORT1}$</td>
<td>-</td>
<td>-</td>
<td>0.0488</td>
<td>0.0496</td>
<td>0.0488</td>
</tr>
<tr>
<td>$(R/Q)_{110}$ [Ω/mm$^2$]</td>
<td>0.46</td>
<td>0.46</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$k_{110}$ [V/nC/mm$^2$]</td>
<td>9.5</td>
<td>9.5</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$Q_{EXT\ PORT1}$</td>
<td>-</td>
<td>-</td>
<td>161000</td>
<td>158000</td>
<td>161000</td>
</tr>
<tr>
<td>$P_{OUT\ PORT1}$ @ 1 nC [W]</td>
<td>-</td>
<td>-</td>
<td>2.41e-3</td>
<td>2.45e-3</td>
<td>2.41e-3</td>
</tr>
<tr>
<td>$V_{OUT\ PORT1}$ @ 1 nC [V]</td>
<td>-</td>
<td>-</td>
<td>0.49</td>
<td>0.49</td>
<td>0.49</td>
</tr>
<tr>
<td>Convergence</td>
<td>average</td>
<td>good</td>
<td>good</td>
<td>average</td>
<td>average</td>
</tr>
<tr>
<td># Tetrahedra</td>
<td>60000</td>
<td>56000</td>
<td>40000</td>
<td>161000</td>
<td>194000</td>
</tr>
</tbody>
</table>

Table 2.3: Position cavity dimensions

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Cavity radius</td>
<td>26.29 mm</td>
</tr>
<tr>
<td>Cavity gap</td>
<td>10 mm</td>
</tr>
<tr>
<td>Beam pipe radius</td>
<td>10 mm</td>
</tr>
<tr>
<td>Coupling WG</td>
<td>30 x 6 mm</td>
</tr>
<tr>
<td>Distance WG to beam axis</td>
<td>19.39 mm</td>
</tr>
</tbody>
</table>

Figure 2.22: Output signal of the position cavity.
2.3. THE ELECTROMAGNETIC DESIGN

along the beam pipe. The beam pipe radius is $R_{BP} = 10$ mm, so its dominant mode ($TE_{11}$) has the following cut-off frequency:

$$f_{TE_{11}} = \frac{c}{2\pi R_{BP}} \approx 8.78 \text{ GHz}$$  \hspace{1cm} (2.17)

where $j'_{11}$ is the first zero of the derivative of the Bessel function of the first kind and first order. Since the dipole mode oscillates at the resonant frequency $f_{RES} = 6.5$ GHz, by (2.17) it is under the cut-off of the beam pipe. Thus this mode is trapped in the cavity. As shown in Figure 2.20, the signal is extracted from the cavity using four rectangular waveguides, which are coupled with the cavity through the magnetic coupling described in [18], [21], [22] and [23]. The magnetic coupling allows the separation of the two dipole polarizations, one sensing the ‘X’ displacement, and the other sensing the ‘Y’ displacement.

The waveguide-coaxial transition

The electron beam excites the dipole mode signal in the position cavity; thus such signal propagates along the four rectangular waveguides and reaches two coaxial cables of the electronic circuitry, which carry out the two output signals, one for the X polarization and the other for the Y polarization. This section focuses on the waveguide-coaxial cable transition. The latter can be realized by exploiting the electric or the magnetic coupling between the waveguide and the antenna, as discussed below.

Electric coupling

In the waveguide-antenna electric coupling, the antenna is placed in the position corresponding to the maximum of the electric field (Fig. 2.23). To do so, the waveguide is shortened, and the maximum of the E-field is $\lambda/4$ away from the short circuit, thus catching the voltage antinode and the current node, as is shown in Figure 2.23. Furthermore, the antenna must not touch the base surface of the waveguide.

This method requires to calculate three dimensions (Figure 2.24):

1. The antenna-closure distance $l_c$, which is not exactly equal to $\lambda/4$, due to the antenna thickness.

2. The antenna-lateral waveguide distance $l_l$. 


3. The antenna-bottom waveguide distance $l_b$.

Unfortunately, the stringent tolerances on these three parameters make this kind of coupler difficult to realize. As an example, Fig. 2.25 shows the dependence of the $|S_{21}|$ parameter on $l_b$, obtained with an HFSS simulation.

The sensitivity of $|S_{21}|$ with respect to $l_b$ can be completely avoided making use of the magnetic coupling, discussed below.

**Magnetic coupling**

In the magnetic coupling the antenna is connected to the bottom of the waveguide, in a point at which a current antinode and a voltage node occur. The magnetic field variation induces a voltage in the antenna, which therefore receives the signal. The latter is then sent to the electronic circuitry. This alternative approach requires calculating two dimensions (Figure 2.26):

1. The antenna-closure distance $l_c$

2. The antenna-lateral waveguide distance $l_l$.

This coupling approach is more robust, because there are only two tolerances, which are not critical. This transition has been designed and simulated.
2.3. THE ELECTROMAGNETIC DESIGN

Figure 2.25: *Sensitivity with respect to the antenna-bottom waveguide distance $l_b$."

Figure 2.26: *Magnetic coupling transition.*

with Ansoft HFSS. The antenna position has been chosen to maximize transmission from the waveguide to the coaxial cable, in the frequency range of interest. This has been done with a trial-and-error procedure, obtaining the dimensions listed in Table 2.4.

Table 2.4: Magnetic coupling dimensions

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$l_c$</td>
<td>31 mm</td>
</tr>
<tr>
<td>$l_l$</td>
<td>13 mm</td>
</tr>
<tr>
<td>Antenna radius</td>
<td>0.5 mm</td>
</tr>
</tbody>
</table>

Figure 2.27 shows the parameters $|S_{21}|$ and $|S_{11}|$ obtained with such magnetic coupling.

In the previous prototypes the geometry of the antenna connector pro-
produced a spurious resonant frequency, because the complicated interconnection system behaved like an additional resonant cavity [27] [28]. For this reason we introduced a new type of antenna, which is simpler, directly brazed in the copper bulk, and not generating parasitic resonances. Figure 2.28 shows both types of connectors.

Figure 2.27: $|S_{21}|$ (a) and $|S_{11}|$ (b) parameters of the waveguide-coaxial cable transition.

Resonance of the rectangular waveguides

The four rectangular waveguides might behave as rectangular resonators, if mismatching in the waveguide-coaxial transition occurs, due to the residual power reflected by the waveguide-coaxial transition. The spurious modes thus generated have a resonant frequency that must be far away from that of the signal of interest. This can be obtained by suitably choosing the
2.3. THE ELECTROMAGNETIC DESIGN

dimensions of each rectangular cavity. With reference to Figure 2.29, the designed dimensions of the cavity are the following: $X_{MAX} = 53$ mm, $Y_{MAX} = 30$ mm, $Z_{MAX} = 6$ mm.

![Waveguide-coaxial transition](image)

Figure 2.29: Waveguide-coaxial transition.

With this choice, the first four resonant frequencies are listed in table 2.5.

Table 2.5: Resonant frequencies of the first four resonant modes of the four rectangular resonator

<table>
<thead>
<tr>
<th></th>
<th>$TE_{101}$</th>
<th>$TE_{102}$</th>
<th>$TE_{103}$</th>
<th>$TE_{201}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant freq</td>
<td>5.73 GHz</td>
<td>7.53 GHz</td>
<td>9.82 GHz</td>
<td>10.36 GHz</td>
</tr>
</tbody>
</table>

The position cavity dipole frequency is 6.5 GHz, which is located between the $TE_{101}$ and $TE_{102}$ resonant frequencies. Thus, the resonances of the rectangular cavity do not affect the signal of interest, because they will be filtered away by the band-pass filter of the electronic circuitry, centered at 6.5 GHz.

Performing the in-tunnel test of the cavity BPM, in presence of the electron beam, the whole output signal has been acquired with the oscilloscope. The Fast Fourier Transform (FFT) of this signal gives the spectrum of Figure 2.47, where the peak number 2 represents the resonant frequency of the rectangular cavity $TE_{102}$ mode, at 7.7 GHz.

2.3.2 The reference cavity

For the reference cavity, we propose the geometry of Fig. 2.30, which consists of a particular pillbox where the antenna is mounted in the protrusion.

The depth of the antenna is related to the level of the desired output signal. Precisely, the deeper the antenna penetration, the higher the output
signal. With this technique, it is possible to adjust the antenna length to obtain the desired output signal level without affecting the resonant frequency of the cavity. We use HFSS to perform the design of the cavity, working at the 6.5 GHz resonant frequency, using the “Eigenmode” and the “Driven modal” settings. The eigenmode setting is used to estimate the resonant frequency of the cavity and the internal quality factor $Q_0$. The driven modal setting is used to estimate the reflection coefficient $S_{11}$ of the output port, thus estimating the external quality factor [24] and the output voltage. For a bunch charge of 1 nC, the RF parameters of the reference cavity are listed in Table 2.6. As a second step, we evaluated the output signal level with

Figure 2.30: Geometry of the reference cavity, with the cross section.

<table>
<thead>
<tr>
<th>HFSS simulations for the reference cavity</th>
<th>90° (E.)</th>
<th>180° (E.)</th>
<th>180° (D.)</th>
<th>No symmetries (D.)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Symmetry planes</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$f_{RES}$ [MHz]</td>
<td>6481.1</td>
<td>6478.2</td>
<td>6477.4</td>
<td>6471.6</td>
</tr>
<tr>
<td>$Q_0$</td>
<td>6307</td>
<td>6255</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$S_{11}$</td>
<td>-</td>
<td>-</td>
<td>0.731</td>
<td>0.730</td>
</tr>
<tr>
<td>$\beta$</td>
<td>-</td>
<td>-</td>
<td>0.155</td>
<td>0.156</td>
</tr>
<tr>
<td>$(R/Q_{010})$ [Ω]</td>
<td>36.4</td>
<td>36.2</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$k_{010}$ [V/nC]</td>
<td>741</td>
<td>737</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$Q_{EXT}$</td>
<td>-</td>
<td>-</td>
<td>40355</td>
<td>40100</td>
</tr>
<tr>
<td>$P_{OUT}@1\hspace{1pt}nC$ [W]</td>
<td>-</td>
<td>-</td>
<td>0.74</td>
<td>0.75</td>
</tr>
<tr>
<td>$V_{OUT}@1\hspace{1pt}nC$ [V]</td>
<td>-</td>
<td>-</td>
<td>8.62</td>
<td>8.64</td>
</tr>
<tr>
<td>Convergence</td>
<td>very good</td>
<td>good</td>
<td>good</td>
<td>good</td>
</tr>
<tr>
<td># Tetrahedra</td>
<td>13000</td>
<td>35000</td>
<td>16000</td>
<td>52000</td>
</tr>
</tbody>
</table>
2.3. THE ELECTROMAGNETIC DESIGN

CST Particle Studio, setting a bunch charge of 1 nC, with a bunch length $\sigma_z = 8 \text{ mm}$, and assuming the electron beam as travelling along the beam pipe axis. Figure 2.31 shows the simulated output voltage of the reference cavity. All the CST results, for both the position and the reference cavities, are listed in Table 2.7.

![Figure 2.31: Output signal of the reference cavity.](image)

<table>
<thead>
<tr>
<th>$V_{OUT} \text{ [V]}$ ($\sigma_z = 8 \text{ mm}$)</th>
<th>Ref. cavity</th>
<th>Position cavity</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>4</td>
<td>0.4</td>
</tr>
</tbody>
</table>

*Values calculated using the form factor

The dimensions of the reference cavity are listed in Table 2.8.

<table>
<thead>
<tr>
<th>Reference cavity dimensions</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cavity radius</td>
</tr>
<tr>
<td>Cavity gap</td>
</tr>
<tr>
<td>Beam pipe radius</td>
</tr>
</tbody>
</table>

### 2.3.3 Design of the tuners

The mechanical tolerances involved in the fabrication of both the position and the reference cavities do not allow one to measure the resonant frequencies
with errors below $\sim$4 MHz. More precisely, the sensitivities of the resonant frequencies with respect to the radius of each cavity are given by:

\[
\frac{df_{\text{REF}}}{dr} \bigg|_{r=R_{\text{REF}}} = -\frac{c}{2\pi} \frac{j\pi}{r^3} \bigg|_{r=R_{\text{REF}}} \cong -3.5 \text{ MHz}/10\mu m
\]

\[
\frac{df_{\text{POS}}}{dr} \bigg|_{r=R_{\text{POS}}} = -\frac{c}{2\pi} \frac{j\pi}{r^3} \bigg|_{r=R_{\text{POS}}} \cong -2.5 \text{ MHz}/10\mu m
\]

(2.18)

Therefore, additional tuners must be used to calibrate the resonant frequencies of both cavities. Such tuners use screws that squeeze the copper.

The reference cavity has three transverse tuners that increase the resonant frequency, in accordance with the Slater theorem [29]. The position cavity is equipped with four longitudinal tuners that decrease the resonant frequency [29]. Two tuners act on the horizontal polarization, while the other two are used for the vertical polarization. For correct operation of the tuners, the resonant frequency of the reference cavity should be lower than the one of the position cavity. Only in this way the frequencies can be suitably adjusted. Figure 2.32 shows the tuners of both cavities.

A simple justification for the fact that longitudinal tuners decrease the resonant frequency is reported. Fig. 2.33 shows two of the position cavity tuners.

The tuners are placed close to the maximum of the dipole E-field. If the volume of the electrical stored energy decreases, the resonant frequency will decrease. Let us consider the equivalent circuit of the resonant cavity (Fig. 2.34).
The electric energy is stored in the capacity. The capacitance is given by:

\[ C = \frac{Q}{V} = \frac{Q}{E l} \]

By lowering the longitudinal dimension \( l \) by \( \Delta l \), we obtain:

\[ (C + \Delta C) = \frac{Q}{V} = \frac{Q}{E (l - \Delta l)} \]

Therefore, the capacitance increases of \( \Delta C \), and the new resonant frequency is:

\[ \omega_{\text{TUNED}} = \frac{1}{\sqrt{L (C + \Delta C)}} < \omega_{\text{INITIAL}} = \frac{1}{\sqrt{LC}} \]
By lowering, with the tuners, the volume of the electrical stored energy, the resonant frequency decreases. Vice versa, incrementing the volume of the electrical stored energy, the resonant frequency rises up.

**Simulations of the tuners**

The effect of the tuners on the resonant frequencies has been simulated with HFSS. The variation of the resonant frequency with respect to the tuner penetration is quadratic, so the data have been fitted with a parabolic curve. Figure 2.35 shows the effect of one tuner of the reference cavity: it can increase the resonant frequency by nearly 7.3 MHz. Three tuners can therefore increase the resonant frequency by nearly 22 MHz.

![Figure 2.35: Simulation and fitting of the reference cavity tuner.](image)

Figure 2.36 shows the effect of one tuner of the position cavity: it can decrease the resonant frequency by nearly 5.9 MHz.

From measurements, it appears that the reference cavity tuners can increase the frequency by 20 MHz. The simulations of the reference cavity tuners are in good agreement with the measurements. In the position cavity we made only a fine tuning of 2.5 MHz.
2.4 Manufacturing

This section concerns the machining of the cavity BPM, made of copper [26]. All the cavities have been produced by the “Cinel Strumenti Scientifici” [7] manufacturing company. The RF cavities are obtained with three stacked parts: the reference cavity, the position cavity with the four waveguides acting as cap of the reference cavity, and the cap of the position cavity. On the reference cavity and on the position cavity cap, two protruding collars are machined to braze on them onto entrance and exit flange. As the resonant cavities have a rotational symmetry, the high precision features for RF and brazing are mostly done by turning; the parts are then milled to reduce their weight and to realize the auxiliary features: aperture for the antenna on the reference cavity, interfaces for the tuners, references for the machining of the rectangular waveguides on the position cavity part. This last part is particularly complex. In fact, the geometry resulting from the combination of waveguides and position cavity, does not allow using the wire-EDM (Electro Discharge Machining), because the four rectangular waveguides do not pass through the position cavity itself and have to stop at a certain distance from the beam axis (Figure 2.37). At the beginning, the rectangular waveguides were obtained by spark erosion on a classic sinker-EDM (Electro-Discharge
Such machine works with an electrode that ablates part of the copper following the electrode shape. The relevant depth of the waveguides combined with the relatively small cross section made an extensive preliminary milling almost impossible to rough machine the waveguides themselves. Therefore, long machining times were needed for the spark erosion phase.

In addition there was the need to use several electrodes in sequence to achieve the tight tolerances imposed on the rectangular aperture along the whole length and the almost sharp corners (several roughing and finishing electrodes in sequence). The strong advantage in terms of productivity and control of the geometrical tolerances given by the use of the wire-EDM, has driven the designers to study a solution allowing the use of this technique. The solution consists of the temporarily removal of the core of this part of the cavity BPM, to be inserted as a separate part after the machining of the waveguides by wire-EDM. Before giving final approval of this procedure, we tested the quality factor and the vacuum performances of the cavity BPM obtained with the new approach. Clearly, it was necessary to verify a perfect electrical contact and the absence of negative effects on the vacuum performance due to the large mating surfaces. Since both the verification tests were successfully passed, showing no appreciable differences compared to the
original configuration, this interesting innovation in the production process was implemented and optimized. The wire-EDM machining of the waveguides, as shown in Fig. 2.38, allows for fast machining step (compared to the sinker-EDM process), tight tolerances on the waveguide transversal dimensions, better surface roughness on the waveguide walls and perfect alignment between two opposite waveguides.

Figure 2.38: *Wire-EDM machining phase with central cylinder removed. This approach represents a huge global advantage in terms of precision, finishing and costs.*

The new configuration requires adding an insert that restores the temporarily removed material. Figure 2.38 shows the wire-EDM machining, with the central part removed by turning, so that the wire can pass through the whole copper block. The missing part is manufactured separately, constituting the “insert” that will complete the final structure. This insert is slightly larger than its seat to make a coupling by interference and optimize in this way the mechanical contact and, consequently, the electrical continuity: to make the insertion easier, before being put in place it is immersed in liquid nitrogen, to make it shrink. After that, it is suddenly placed into its seat on the cross piece, constituting the main body of the position cavity (Figure 2.39).

When the insert reaches the environmental temperature, it expands and remains locked inside the structure. The final result is shown in Figure 2.40.
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

Figure 2.39: Insertion of main body core. To guarantee a perfect mechanical and electrical contact, the tolerances are combined to get an interference of 0.02 mm. To ease the insertion, the insert itself is cooled down in liquid nitrogen just before insertion to make it shrink.

Note that to comply with the tight tolerances of the cavity itself, some material is left on the main body and on the insert. A further turning is performed on the position cavity (outside diameter and depth), and on the hole of insert together with the main structure beam pipe. The turning has also the effect of making the two parts touch perfectly at the interfaces since the turning makes a combination of cutting and plastic deformation of the material, making the two surfaces mechanically “melt” at the interface.

With this new technique, the manufacturing has been simplified and made more reliable. A series of 25 cavity BPMs has been produced at Cinel Strumenti Scientifici [7]; 10 cavities have been employed in the FEL 1 beamline. Figure 2.41 shows the building blocks of copper that have been assembled to realize the cavity BPMs.

The tolerances are $\pm 10 \mu m$ on all the relevant features such as diameters of passing holes and cavities (position and reference), height of cavities and transversal dimensions of waveguides, involving both the machining processes of turning and EDM. Such high precision in the manufacturing process strongly reduces the crosstalk between the orthogonal ports of the position cavity. For precaution only, the manufacturing procedure still foresees a resonant frequency measurement before final turning of cavities diameters, left
2.4. MANUFACTURING

Figure 2.40: The passing through hole and the position cavity (highlighted in green colour) are finished by turning after insertion to guarantee the tight tolerances (± 10 microns).

Figure 2.41: Building blocks of the cavity BPM, before brazing.

until this moment with 0.05 mm of material more on the radius. The required thickness to remove with the finishing phase has been calculated based on the measured resonant frequency, but this was always corresponding to the excess material with the uncertainty corresponding to the geometrical tolerance specified. The resonant frequencies of the cavities have been inspected before and after brazing, as reported in next section.
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

2.5 Workbench measurements

This section describes the results of the measurements performed with a vector network analyzer. Precisely, we report the resonant frequencies of both cavities and the crosstalk in the position cavity.

2.5.1 The resonant frequencies

The resonant frequencies of the cavities have been measured both before and after brazing, to ensure that the resonant frequency of the reference cavity is lower than that of the position cavity [30]. Only if this condition is satisfied, the two frequencies can be equalized with the tuners. If the reference cavity has resonant frequencies greater than those of the position cavity, the reference cavity must be rectified to decrease its resonant frequency. Four of the produced reference cavities showed this problem. Thus, they have been undergone to an additional turning of 30 \( \mu m \) on the diameter. The resonant frequencies of the reference cavity are reported in Table 2.9.

<table>
<thead>
<tr>
<th>Reference cavity ( f_{RES} ) [MHz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cavity BPM</td>
</tr>
<tr>
<td>Number</td>
</tr>
<tr>
<td>------------</td>
</tr>
<tr>
<td>1</td>
</tr>
<tr>
<td>2</td>
</tr>
<tr>
<td>3</td>
</tr>
<tr>
<td>4</td>
</tr>
<tr>
<td>5</td>
</tr>
<tr>
<td>6</td>
</tr>
<tr>
<td>7</td>
</tr>
<tr>
<td>8</td>
</tr>
<tr>
<td>9</td>
</tr>
<tr>
<td>10</td>
</tr>
<tr>
<td>11</td>
</tr>
<tr>
<td>12</td>
</tr>
<tr>
<td>13</td>
</tr>
</tbody>
</table>

Table 2.9: Resonant frequencies \( (f_{RES}) \) of the reference cavity. An asterisk indicates the rectified cavities.
The frequencies of the position cavities, measured before and after brazing, are reported in Table 2.10.

<table>
<thead>
<tr>
<th>Cavity BPM number</th>
<th>Before Brazing</th>
<th>After Brazing H. Polarization</th>
<th>After Brazing V. Polarization</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>6506.7</td>
<td>6502.8</td>
<td>6502.3</td>
</tr>
<tr>
<td>2</td>
<td>6509.8</td>
<td>6502.5</td>
<td>6502.6</td>
</tr>
<tr>
<td>3</td>
<td>6506.5</td>
<td>6503.7</td>
<td>6503.3</td>
</tr>
<tr>
<td>4</td>
<td>6508.2</td>
<td>6503.8</td>
<td>6504.2</td>
</tr>
<tr>
<td>5</td>
<td>6508.8</td>
<td>6504.5</td>
<td>6504.6</td>
</tr>
<tr>
<td>6</td>
<td>6505.6</td>
<td>6503.9</td>
<td>6504.4</td>
</tr>
<tr>
<td>7</td>
<td>6507.3</td>
<td>6504.0</td>
<td>6504.0</td>
</tr>
<tr>
<td>8</td>
<td>6507.3</td>
<td>6505.3</td>
<td>6505.0</td>
</tr>
<tr>
<td>9</td>
<td>6509.3</td>
<td>6503.1</td>
<td>6503.7</td>
</tr>
<tr>
<td>10</td>
<td>6507.9</td>
<td>6504.4</td>
<td>6505.0</td>
</tr>
<tr>
<td>11</td>
<td>6510.3</td>
<td>6505.8</td>
<td>6505.6</td>
</tr>
<tr>
<td>12</td>
<td>6501.1</td>
<td>6503.2</td>
<td>6503.1</td>
</tr>
<tr>
<td>13</td>
<td>6501.1</td>
<td>6501.4</td>
<td>6501.0</td>
</tr>
</tbody>
</table>

2.5.2 Crosstalk

The crosstalk between the orthogonal ports has also been evaluated, by measuring the transmission between the ports that should be ideally isolated. The ports of the position cavity are numbered in Figure 2.42, and the crosstalk values are listed in Table 2.11.

![Figure 2.42: Numbering of the ports.](image-url)
Table 2.11: Crosstalking ($Ct$)

<table>
<thead>
<tr>
<th>Cavity BPM number</th>
<th>$S_{14}$</th>
<th>$S_{21}$</th>
<th>$S_{32}$</th>
<th>$S_{43}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-41.2</td>
<td>-40.7</td>
<td>-41.5</td>
<td>-41.7</td>
</tr>
<tr>
<td>2</td>
<td>-50.0</td>
<td>-49.0</td>
<td>-50.0</td>
<td>-50.0</td>
</tr>
<tr>
<td>3</td>
<td>-49.0</td>
<td>-48.7</td>
<td>-48.6</td>
<td>-49.0</td>
</tr>
<tr>
<td>4</td>
<td>-74.0</td>
<td>-84.0</td>
<td>-77.0</td>
<td>-72.0</td>
</tr>
<tr>
<td>5</td>
<td>-47.0</td>
<td>-48.0</td>
<td>-49.0</td>
<td>-49.0</td>
</tr>
<tr>
<td>6</td>
<td>-51.0</td>
<td>-51.0</td>
<td>-48.4</td>
<td>-49.0</td>
</tr>
<tr>
<td>7</td>
<td>-50.0</td>
<td>-50.0</td>
<td>-50.5</td>
<td>-51.7</td>
</tr>
<tr>
<td>8</td>
<td>-50.0</td>
<td>-50.0</td>
<td>-51.0</td>
<td>-51.6</td>
</tr>
<tr>
<td>9</td>
<td>-48.6</td>
<td>-48.5</td>
<td>-50.5</td>
<td>-50.0</td>
</tr>
<tr>
<td>10</td>
<td>-42.0</td>
<td>-42.0</td>
<td>-42.0</td>
<td>-42.0</td>
</tr>
<tr>
<td>11</td>
<td>-70.0</td>
<td>-70.0</td>
<td>-65.0</td>
<td>-63.0</td>
</tr>
<tr>
<td>12</td>
<td>-52.5</td>
<td>-52.0</td>
<td>-50.0</td>
<td>-51.0</td>
</tr>
<tr>
<td>13</td>
<td>-75.0</td>
<td>-59.0</td>
<td>-60.0</td>
<td>-60.0</td>
</tr>
<tr>
<td>Average value</td>
<td>-54.3</td>
<td>-49.2</td>
<td>-49.9</td>
<td>-51.0</td>
</tr>
</tbody>
</table>

These values indicate that a good isolation exists between the orthogonal ports. Only two of the ports are used by the electronic measurement system. The other two ports are terminated on a 50 Ω load.

2.6 Installation

Ten cavity BPMs have been installed in the “FEL 1” beamline. When a vacuum of $10^{-9}$ mbar has been achieved, the cavities are ready for the operations described below. The order of installation is different from the manufacturing one, because the cavities with the best characteristics have been chosen for the launching region.

2.6.1 Tuning

The tuning was performed in the tunnel using a portable network analyzer. Table 2.12 reports the tuning of the ten cavity BPMs of FEL 1, in order of installation.
Table 2.12: Tuning of the position cavities; ‘H.’ is the Horizontal polarization of the position cavity; ‘V.’ is the Vertical polarization of the position cavity; ‘R.’ is the resonance frequency of the reference cavity.

<table>
<thead>
<tr>
<th>#</th>
<th>R. Cav.</th>
<th>pos H.</th>
<th>pos V.</th>
<th>H.-V.</th>
<th>H.-R.</th>
<th>V.-R.</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>6503.03</td>
<td>6503.42</td>
<td>6503.03</td>
<td>0.39</td>
<td>0.39</td>
<td>0.00</td>
</tr>
<tr>
<td>2</td>
<td>6507.35</td>
<td>6507.58</td>
<td>6507.68</td>
<td>-0.10</td>
<td>0.23</td>
<td>0.33</td>
</tr>
<tr>
<td>3</td>
<td>6507.19</td>
<td>6507.77</td>
<td>6507.39</td>
<td>0.38</td>
<td>0.58</td>
<td>0.20</td>
</tr>
<tr>
<td>4</td>
<td>6504.60</td>
<td>6504.66</td>
<td>6504.70</td>
<td>-0.04</td>
<td>0.06</td>
<td>0.10</td>
</tr>
<tr>
<td>5</td>
<td>6503.58</td>
<td>6503.69</td>
<td>6503.63</td>
<td>0.06</td>
<td>0.11</td>
<td>0.05</td>
</tr>
<tr>
<td>6</td>
<td>6504.27</td>
<td>6504.04</td>
<td>6504.11</td>
<td>-0.07</td>
<td>-0.23</td>
<td>-0.16</td>
</tr>
<tr>
<td>7</td>
<td>6504.78</td>
<td>6504.94</td>
<td>6505.04</td>
<td>-0.10</td>
<td>0.16</td>
<td>0.26</td>
</tr>
<tr>
<td>8</td>
<td>6503.98</td>
<td>6503.89</td>
<td>6504.11</td>
<td>-0.22</td>
<td>-0.09</td>
<td>0.13</td>
</tr>
<tr>
<td>9</td>
<td>6505.64</td>
<td>6505.71</td>
<td>6505.67</td>
<td>0.04</td>
<td>0.07</td>
<td>0.03</td>
</tr>
<tr>
<td>10</td>
<td>6502.52</td>
<td>6502.56</td>
<td>6502.53</td>
<td>0.03</td>
<td>0.04</td>
<td>0.01</td>
</tr>
</tbody>
</table>

With an accurate design for the positioning of the tuners, the resonant frequencies can be adjusted with errors lower than 400 kHz.

2.6.2 Thermal drifts

This section addresses the effects of the temperature variations on the resonant frequencies of both the reference and the position cavities. Such frequencies are given by the expressions:

\[
\begin{align*}
    f_{\text{REF}} &= \frac{c}{2\pi R_{\text{REF}}} j_{01} \\
    f_{\text{POS}} &= \frac{c}{2\pi R_{\text{POS}}} j_{11}
\end{align*}
\]  

(2.19)

where \( R_{\text{REF}} \) and \( R_{\text{POS}} \) are the radii of the two cavities.

Effect of variation of the radius on the resonant frequency

The radii \( R_{\text{REF}} \) and \( R_{\text{POS}} \) are designed in such a way that the two resonant frequencies \((f_{\text{REF}} \text{ and } f_{\text{POS}})\) are as close as possible to each other. Due to the mechanical tolerances, this is very difficult to achieve. To ensure that \( f_{\text{REF}} \approx f_{\text{POS}} \), we make use of tuners.

A perturbation of the cavity radius modifies the resonant frequency. Calculating the derivatives of the two resonant frequencies with respect to the
radius by (2.19), and approximating such derivatives with finite differences, yields:

\[
\frac{\Delta f_{\text{REF}}}{\Delta R_{\text{REF}}} \approx -\frac{c}{2\pi R_{\text{REF}}^2} \frac{j_{01}}{R_{\text{REF}}} \quad ; \quad \frac{\Delta f_{\text{POS}}}{\Delta R_{\text{POS}}} \approx -\frac{c}{2\pi R_{\text{POS}}^2} \frac{j_{11}}{R_{\text{POS}}}
\]  

(2.20)

which hold true for small variations of the radii.

**Effect of the temperature on the radius**

With a thermal drift, a copper bar of length \( l \) is subject to a length variation \( \Delta l \). As is well-known, the relative length variation and the temperature variation are related as follows:

\[
\frac{\Delta l}{l} \frac{1}{\Delta T} = 1.6 \cdot 10^{-5} \quad ^\circ\text{C}^{-1}
\]

(2.21)

where \( \Delta T \) is the temperature variation. Therefore, the radius variation with respect to the temperature variation is given by:

\[
\frac{\Delta r}{\Delta T} = 1.6 \cdot 10^{-5} \cdot r
\]

(2.22)

where \( r = R_{\text{REF}} \) for the reference cavity, and \( r = R_{\text{POS}} \) for the position cavity.

**Effect of the temperature on the resonant frequency**

Combining (2.20) and (2.22), yields the relations:

\[
\frac{\Delta f_{\text{REF}}}{\Delta T} = -1.6 \cdot 10^{-5} \cdot \frac{c}{2\pi R_{\text{REF}}} \frac{j_{01}}{R_{\text{REF}}} = -1.6 \cdot 10^{-5} \cdot f_{\text{REF}}
\]

\[
\frac{\Delta f_{\text{POS}}}{\Delta T} = -1.6 \cdot 10^{-5} \cdot \frac{c}{2\pi R_{\text{POS}}} \frac{j_{11}}{R_{\text{POS}}} = -1.6 \cdot 10^{-5} \cdot f_{\text{POS}}
\]

As already observed above, the two cavities are designed to obtain \( f_{\text{REF}} \approx f_{\text{POS}} \). Therefore, the frequency thermal drift is nearly the same for the two cavities, and is given by:
\[
\frac{\Delta f_{\text{REF}}}{\Delta T} = \frac{\Delta f_{\text{POS}}}{\Delta T} = -1.6 \cdot 10^{-5} \cdot f_{\text{RES}} \tag{2.23}
\]

where we assumed \( f_{\text{RES}} = f_{\text{REF}} = f_{\text{POS}} \).

Setting \( f_{\text{RES}} = 6.5 \text{ GHz} \), the thermal drift (for both cavities) is therefore given by:

\[
\frac{\Delta f_{\text{RES}}}{\Delta T} = -1.6 \cdot 10^{-5} \cdot f_{\text{RES}} \approx -100 \text{ kHz/}^{\circ}\text{C} \tag{2.24}
\]

This result can also be used where a further tuning is needed, with a thermostated control unit.

### 2.7 The electronic system

This section describes the new approach adopted for the electronics used in the FERMI@Elettra BPM cavities. Our measurement system is entirely original: it has never been thought before. This approach avoids the use of the IQ demodulation, whose circuitry is complex, and is based on non-linear electronic stages. The new approach of the electronic measurement system is based on a 180° hybrid device [25] [31]. It gives the sum(\( \Sigma \)) and the difference(\( \Delta \)) of the signals generated by the reference (\( V_{\text{REF}} \)) and by the position (\( V_{\text{POS}} \)) cavities. If the beam is crossing the cavities in the center, the position signal is equal to zero, and the hybrid output signals are equal. When the beam is crossing the cavities off axis, generating the signals of Fig. 2.17, the hybrid output signals are different. By analyzing the unbalance of the signal magnitudes, it is possible to determine the beam position. The magnitude of the signals is obtained by using AM detectors, placed after the hybrid. This new approach gives the maximum amount of signal when the electron beam is travelling through the cavity center. Moreover, the first stages of the electronics are made with simple components, and the RF-frontend is easier to be realized. We developed two versions of this circuit, one that works properly when the component of the bunch tilt is negligible with respect to the bunch offset. The second version measures the beam offset, rejecting the unwanted tilt component.
2.7.1 The first electronic system

The schematic of the first kind of circuit is depicted in Fig. 2.43 (the unused ports are closed on a matched-load).

![Figure 2.43: First type of electronic circuit.](image)

The signals of the reference and of the BPM cavity go to the 180° hybrid, which gives the sum (Σ) and the difference (Δ) of such signals. After each cavity, there is a 6.5 GHz band-pass filter used to eliminate the signal of other resonant modes. When the electron beam is crossing the cavity with small offsets, the position signal is very weak, but both output signals from the hybrid (Σ and Δ) have a strong magnitude. In particular, with zero offset, the sum and difference have the same level. This is the first advantage of this circuit, which allows to have a strong output signal even with small electron beam offsets.

As shown in Fig. 2.43, the circuitry follows with an active detector, and with an integrator that calculates the area of the pulse. The phase shifter is tuned in order to have the hybrid input signals ($V_{REF}$ and $V_{POS}$) in phase, when the beam has a positive offset.

From a theoretical point of view, the following calculations represent the electronic system. The hybrid input signals are:

$$V_{REF} = A \cos(\omega t) e^{-t/\tau_R}$$

$$V_{POS} = (B_o \cos(\omega t) + B_t \sin(\omega t)) e^{-t/\tau_p}$$  \hspace{1cm} (2.25)

where “$A$” is the amplitude of the reference signal, “$B_o$” is the offset signal of the position cavity, “$B_t$” is the tilt signal of the position cavity, “$\tau_R$” and
“τ_P” are the reference and the position cavity time constants, respectively. The sum and the difference output signals from the hybrid are, respectively:

\[ \Sigma = (Ae^{-t/\tau_R} + B_\sigma e^{-t/\tau_P})\cos(\omega t) + B_t e^{-t/\tau_P} \sin(\omega t) \]

\[ \Delta = (Ae^{-t/\tau_R} - B_\sigma e^{-t/\tau_P})\cos(\omega t) - B_t e^{-t/\tau_P} \sin(\omega t) \]

The detectors extract the amplitude of such signals, giving therefore:

\[ \Sigma_D = \sqrt{(Ae^{-t/\tau_R} + B_\sigma e^{-t/\tau_P})^2 + (B_t e^{-t/\tau_P})^2} \]

\[ \Delta_D = \sqrt{(Ae^{-t/\tau_R} - B_\sigma e^{-t/\tau_P})^2 + (B_t e^{-t/\tau_P})^2} \] (2.26)

The integrators numerically calculate the area of the pulses. Such values \(A_\Sigma\) and \(A_\Delta\) go to the FPGA-board [32]. The latter allows to estimate the electron beam offset. In this case the tilt signal \(B_t\) must be negligible with respect to the offset signal \(B_\sigma\), because in the offset measurement the tilt component is an unwanted signal. Eq. (2.11) gives the trajectory tilt signal, while Eq. (2.10) gives the desired position signal. Since, for our C-BPMs, it is \(k = \omega/c = 136\) rad/m, and \(l=10^{-2}m\), the ratio between the tilt and the offset signal \(V_1^{110}/V_1^{100}\) is \(1.18 \cdot 10^{-3}\). Thus, for 1 \(\mu m\) of resolution, the beam tilt component must be approximately less than 0.1 mrad. The effect of the head-tail angle, of Eq. (2.12), is always negligible because the bunch length term \(\sigma_z\) is very small in the undulator hall.

The beam position is estimated as follows:

\[ Pos[mm] = k \cdot \frac{A_\Sigma - A_\Delta}{A_\Sigma + A_\Delta} \] (2.27)

where \(k\) is a constant calibration factor from arbitrary units (areas) to \(mm\). It is determined during the cavity BPM calibration, performed with the movers.

### 2.7.2 Rejection of the tilt signal

By using the same kind of electronics it is possible to reject the tilt signal with the configuration of Fig. 2.44.

The “d” signal is therefore:

\[ d = \Sigma_D^2 - \Delta_D^2 = 4AB_\sigma e^{-t/\tau_R} e^{-t/\tau_B} \] (2.28)
Hence, the integrator gives:

$$A_d \propto 4AB_\sigma$$  \hspace{1cm} (2.29)

This result is tilt-free and is analogous to that obtained with the coherent demodulation. The signal $A_d$ is proportional to the beam offset. Since in the FERMI@Elettra project the electron beam does not have a tilt that exceeds $0.1 \mu\text{rad}$, the first approach has been used for the electronic system.

### 2.7.3 Calibration of the electronics

The above equations hold true only for ideal electronic components. In the real world, analog electronic components are affected by differences and tolerances on the gains of the two branches (sum and difference). We used an automatic calibration system that cancels the unbalances of the hybrid, detectors, ADC and of the integrators. The calibration is automatically performed in the time interval between two beam shots.

When the electron beam is travelling through the electrical center of the cavity, the sum must be equal to the difference ($A_\Sigma = A_\Delta$). Unfortunately this does not happen in the real world, and the electronic calibration corrects the tolerances of the two analog branches. It operates by simulating the signal of a perfectly centered beam. For calibrating the first kind of circuit of Fig. 2.43 we use the configuration of Fig. 2.45, where we inject a generated signal into $V_{\text{REF}}$, while $V_{\text{POS}}$ is set to zero.

In this way we measure $A_{\Sigma,\text{cal}}$ and $A_{\Delta,\text{cal}}$, which will be not equal. This inequality is used to balance the two analog branches, by calculating the following two correcting gains with Eq. (2.30).
2.8. ELECTRON BEAM MEASUREMENTS

Figure 2.45: Calibration signals

\[
\begin{align*}
G_{\Sigma} &= \frac{2 \cdot A_{\Delta \text{cal}}}{A_{\Sigma \text{cal}} + A_{\Delta \text{cal}}} \\
G_{\Delta} &= \frac{2 \cdot A_{\Sigma \text{cal}}}{A_{\Sigma \text{cal}} + A_{\Delta \text{cal}}}
\end{align*}
\]  

(2.30)

These two gains are used to compensate the two branches unbalances, and the beam position is estimated as follows [33]:

\[
\text{Pos[mm]} = k \cdot \frac{G_{\Sigma} \cdot A_{\Sigma} - G_{\Delta} \cdot A_{\Delta}}{G_{\Sigma} \cdot A_{\Sigma} + G_{\Delta} \cdot A_{\Delta}}
\]  

(2.31)

where \(k\) is a constant calibration factor from arbitrary units (areas) to \(\text{mm}\). It is determined during the cavity BPM calibration, performed with the movers.

2.8 Electron beam measurements

This section describes the measurements performed in the control room with the electron beam. The entire electronic and data acquisition system has been developed in-house [32] [33]. The “Real Time” feature allows the synchronous data acquisition of a set of shots stamped with the bunch number.

2.8.1 Measuring the output signal levels

This subsection presents the test of the cavity BPM prototype, in presence of the electron beam. The objective is to assess the level of the output signals. The cavity BPM is placed between two microstrip BPMs (“BPM A” and “BPM B”), as shown in Fig. 2.46. An oscilloscope has been connected directly to the reference and to the position cavity. The electric charge is 270 pC, while the duration of the pulse (\(\sigma_t\)) is nearly 10 ps.
CHAPTER 2. THE CAVITY BEAM POSITION MONITORS

The levels of the output signals are measured varying the offset of the electron beam and reading its position with two microstrip BPMs placed near the beam pipe. Five tests have been performed, by varying the offset of $\pm 1$ millimeter in the vertical and in the horizontal plane and by recording the oscilloscope signals levels. Table 2.13 summarizes the values of the output signals given by both cavities.

<table>
<thead>
<tr>
<th>Cavity BPM tunnel measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reference cavity</td>
</tr>
<tr>
<td>Position cavity, X offset</td>
</tr>
<tr>
<td>Position cavity, Y offset</td>
</tr>
</tbody>
</table>

The signals of the cavities are lower, but are in a good agreement with the estimated values. The harmonic content of the position cavity signal has been analyzed by calculating its FFT. The spectrum is reported in Figure 2.47.

The harmonic content of such output signal is satisfactory, as it indicates that the connectors do not generate parasitic resonances. Moreover, in the RF frontend we use a band-pass filter centered at the dipole mode frequency of the resonant cavity.

2.8.2 Calibration of the cavity BPM

The calibration of the cavity BPM is required for determining the conversion factor $k$ of Eq. 2.31. This parameter translates the arbitrary unit value measured by the electronics into a displacement of the electron beam (expressed in mm) with respect to the center, with a 10% of accuracy [30]. The calibration algorithm works assuming that the electron beam is not changing its position with respect to the cavity BPM center. It is based on the following operations:
2.8. ELECTRON BEAM MEASUREMENTS

Figure 2.47: Spectrum of the position cavity output signal. Peak 1: dipole mode of the position cavity, at 6.48 GHz; peak 2: rectangular waveguide resonance, at 7.71 GHz; peak 3: dipole mode of the reference cavity, at 8.48 GHz; peak 4: quadrupole mode of the position cavity, at 9.05 GHz.

1. Move the cavity BPM with the mover of a known step;
2. Read the variation of the arbitrary unit number given by the electronics;
3. Repeat the previous steps for a complete sweep of the mover’s range, and calculate the calibration coefficient.

The calibration factor has been calculated by averaging 160 acquisitions. Fig. 2.48 shows the result of the calibration: each read value is consistent with the corresponding mover displacement.

2.8.3 Resolution of the cavity BPM

The resolution of a single BPM is assessed among cavities with the same characteristics. This resolution method correlates the reading of the BPM of interest with the readings of the others BPMs [34]. The procedure assesses the resolution with ‘p’ pulses and ‘n’ BPMs. The BPM of interest is indexed with ‘m’, and is placed as shown in Fig. 2.49.

The reading of the electron beam offset measured with the BPM of interest (‘m’) is compared with the predicted value by all the other BPMs. In the presence of beam transversal instability, the data of many bunches are collected.
The standard deviation of the errors is related to the resolution of the BPM of interest as follows:

$$\sigma_m = \sqrt{\frac{1}{p} \sum_{i=1}^{p} (x_{i,\text{measured}} - x_{i,\text{predicted}})^2} \quad (2.32)$$

According to [22], the resolution value estimated from three BPMs is given by:

$$\text{BPM resolution} = \frac{\sigma_m}{\sqrt{1^2 + \left(\frac{z_{12}}{z_{13}}\right)^2 + \left(\frac{z_{23}}{z_{13}}\right)^2}} \quad (2.33)$$

where $z_{ij}$ is the distance between the $i_{th}$ and the $j_{th}$ installed BPM.

The first three cavity BPMs in the spreader section have been used for the resolution assessment (Figure 2.50).
The readings of these cavity BPMs are obtained shot by shot in “Real-Time” mode, in presence of beam jitter. Fig. 2.51 represents the measured-predicted value of the cavity BPM of interest. The measured resolution is nearly 1.8 \( \mu \text{m} \) with 500 \( \text{pC} \) of bunch charge.

A series of improvements in the electronics chain are already known to be beneficial for the increase of the measurement accuracy. Upgrade activities comprise:

1. Adding low-pass filters before the ADC.

2. Insertion of a programmable attenuator in the reference signal to adjust the “weight” of the position signal at different beam charges.

3. Digitalization of the entire sum and difference measurement system.
Chapter 3

The RF linear accelerators

3.1 The travelling wave linear accelerators

The RF travelling wave linear accelerators are travelling wave disk loaded structures in which an electromagnetic field is used to accelerate the electron bunches [35] [36] [37]. The main contribution on the acceleration is due to the electric field. The front-end injection systems of the FERMI@Elettra linac [1] [2] produce high brightness electron beams that contribute to the performance of the Free Electron Laser (FEL). As shown in chapter 1, the “Linac 0” photoinjector mainly consists of the radiofrequency (RF) gun and of two S-band RF accelerating structures which accelerate the beam, working at 2.998 GHz and operating with a 16 MeV/m accelerating electric field. The main challenge in the microwave design of such structures is that of obtaining a high gradient of acceleration without degrading the beam emittance, which is a constrained parameter in FELs [38]. However, the couplers of the accelerating structure have field asymmetries that are responsible for the deflection of the electron beam [39] [40] [41]. There are two kinds of undesired effects:

1. The beam transversal kick;

2. The beam head-tail deflection.

Fig. 3.1 shows the beam trajectory kick. If the trajectory deviation is not large, the trajectory kick can be compensated with the steering dipole at the linac entrance.

Fig. 3.2 shows the beam head-tail deflection (called also head-tail kick). This effect cannot be compensated, and causes the degradation of the beam...
quality, described by the so called emittance [9]. Thus, the beam head-tail deflection must be controlled in the linac design.

The field asymmetries are responsible for both the unwanted effects on the electron beam. As will be demonstrated later on, the beam trajectory kick is generated by the phase gradient of the electric field, while the beam head-tail deflection is caused by the amplitude gradient. The accelerating structures of Fig. 3.1 and 3.2 are of “Single feed” type. Such kind of couplers have a strong field asymmetry, referred to the dipolar component, as explained in the following section. A possible solution capable of reducing the dipolar field is that of translating the coupler along the $-x$ axis, as shown in Fig. 3.3.

This solution only compensates the dipolar field asymmetry in small regions close to the $z$ axis. Moreover, due to the coupler translation, the mechanical asymmetry introduces high order modes, such as the “sextupole”,

---

**Figure 3.1:** Beam trajectory kick.

**Figure 3.2:** Beam head-tail deflection.
The structures of the “Linac 0” of the FERMI@Elettra linac [1] are of “Single feed translated” type. They must be replaced by new accelerating structures having a more symmetric field, so as to avoid negative effects on the beam emittance.

A possible solution for reducing the dipolar field is that of using the dual feed coupler [9] [42]. The latter solution requires a power splitter and two feeding waveguides that must be calibrated both in phase and in attenuation. Fig. 3.4 shows such linac assembled with the power splitter and the double waveguide feeding system.

Figure 3.3: Single feed translated linac.

Figure 3.4: Dual feed linac with waveguide feeding system.
However, this approach reduces the dipolar component only if the waveguides have the same attenuation and the same phase delay. Moreover, the power splitter and the additional components increase the final cost of the device. In order to avoid the use of the power splitter, the dipolar field can be reduced also by realizing a single feed coupler endowed with an adjustable short-circuit on the opposite side [43]. The latter approach provides the same field magnitude distribution as the dual feed structure, with the advantage of avoiding the use of power splitters. Fig. 3.5 shows the difference between the dual feed structure and the single feed with movable short-circuit structure.

Figure 3.5: Dual feed (a) and Single feed with movable short-circuit (b).

The latter approach (Single feed with movable short-circuit) allows the reduction of the dipolar component of the electric field magnitude. The quadrupolar component is reduced by extending the racetrack concept [9] [41] [44] to the single feed with movable short-circuit structure. This new approach will be analyzed and discussed later.

3.2 Beam dynamics in the coupler: overview

Accelerating structures endowed with a single feed coupler (Fig. 3.6) cause deflection and degradation of the electron beam properties, due to the asymmetry of the electromagnetic field.

The longitudinal electric field can be expanded in multipoles, where the
monopole is the accelerating mode, while the other modes (such as the dipole and the quadrupole modes) are unwanted and cause degradation of the beam quality. In a first-order approximation, the longitudinal electric field of the coupler can be expressed as [39]:

\[ E_z = \left[ E_{z0} + \frac{\Delta E_d}{2a} x \right] \exp \left( j \frac{\Delta \varphi_d}{2a} x \right) \]  \hspace{1cm} (3.1)

where \( \Delta \varphi_d / 2a \) is the phase gradient of \( E_z \), and \( \Delta E_d / 2a \) is the amplitude gradient. The latter quantities are mean values along the z-axis of the coupler, which take into account the transit factor [39]. This is an expression of the dipole field, where the \( x \) and \( y \) axes are chosen in such a way that \( x \) is parallel with the dipole momentum \( p_x \). The Panofsky-Wenzel theorem [40] shows that the electron beam receives a transversal kick as well. Precisely, the transversal momentum \( (p_\perp) \) is given by:

\[ p_\perp = \frac{q}{\omega} \int_0^l \nabla_t E_z dz \]  \hspace{1cm} (3.2)

where \( q \) is the electron charge, \( \omega \) the angular frequency, and \( l \) the coupler length. The transversal dynamics of the particle can be evaluated by substituting (3.1) into (3.2), obtaining:

\[ p_\perp = \frac{q \cdot E_{z0} \cdot l}{\omega} \left( \frac{\Delta \varphi_d}{2a} \cos(\varphi_{RF}) + \frac{\Delta E_d}{2aE_{z0}} \sin(\varphi_{RF}) \right) \]  \hspace{1cm} (3.3)
where $\varphi_{RF}$ is the RF phase. The crest ($\varphi_{RF} = 0$) corresponds to the RF phase that gives the maximum acceleration. An electron bunch can be modeled as two particles, simulating the head (particle 1), and the tail (particle 2). After crossing the coupler, the beam quality degradation is evaluated by calculating the difference between the two particle momentums:

$$\Delta p_\perp = p_{\perp,1} - p_{\perp,2}$$

(3.4)

By substituting Eq. (3.3) into Eq. (3.4) an analytical estimate of the beam quality degradation can be obtained. The particles 1 and 2 have different RF phase values $\varphi_{RF}$. However if we consider short bunches accelerated in crest, on Eq. (3.3) we can set $\cos(\varphi_{RF}) \approx 1$, and $\sin(\varphi_{RF}) \approx \varphi_{RF}$, and therefore the phase field asymmetry contribution $\Delta \varphi_d$ to $\Delta p_\perp$ of Eq. (3.4) is negligible.

For short bunches, the main beam emittance degradation is given by the magnitude electric field asymmetry, while the phase electric field asymmetry gives only a transversal kick that can be compensated with the steering dipole at the linac entrance.

3.2.1 Evaluation of the field asymmetry

The multipole field contributions can be easily quantified from the simulation results. The field asymmetry can be evaluated by measuring the variation of the electric field along the coupler midplane. Assuming that the dipole field, which is oriented in the ‘x’ direction (Fig. 3.6) is predominant with respect to the higher modes, it varies as a cosinusoidal function of the azimuth angle $\phi$. The contribution of the dipole can be evaluated as follows:

$$\Delta E_d/E_z = \frac{|E_z(x = a)| - |E_z(x = -a)|}{|E_z(x = 0, y = 0)|}$$

(3.5)

$$\Delta \varphi_d = \angle E_z(x = a) - \angle E_z(x = -a)$$

where $a$ is the iris radius.

On the contrary, if the quadrupole field is predominant and varies as $\cos(2\phi)$, the quadrupole contribution can be evaluated as follows:

$$\Delta E_q/E_z = \frac{|E_z(x = a)| - |E_z(y = a)|}{|E_z(x = 0, y = 0)|}$$

(3.6)

$$\Delta \varphi_q = \angle E_z(x = a) - \angle E_z(y = a)$$
3.3. THEORY OF COUPLER MATCHING

The general multipole decomposition is performed by expanding the amplitude and the phase of the field component $E_z$ calculated in the midplane. This gives the following Fourier coefficients $C_{k}^{\text{mag}}$ (referred to the magnitude of $E_z$) and $C_{k}^{\text{arg}}$ (referred to the phase of $E_z$):

$$C_{k}^{\text{mag}} = \frac{1}{2\pi} \int_{0}^{2\pi} |E_z(\phi, r = a, z = \frac{l}{2})| e^{jk\phi} d\phi$$

$$C_{k}^{\text{arg}} = \frac{1}{2\pi} \int_{0}^{2\pi} \angle E_z(\phi, r = a, z = \frac{l}{2}) e^{jk\phi} d\phi$$

(3.7)

Eqs. (3.5) and (3.6) are related with (3.7) as follows:

$$\Delta E_d/E_z = 4 \cdot |C_{k}^{\text{mag}}| \cdot \cos(\angle C_{k}^{\text{mag}})$$

$$\Delta \phi_d = 4 \cdot |C_{k}^{\text{arg}}| \cdot \cos(\angle C_{k}^{\text{arg}})$$

$$\Delta E_q/E_z = 4 \cdot |C_{k}^{\text{mag}}| \cdot \cos(\angle C_{k}^{\text{mag}})$$

$$\Delta \phi_q = 4 \cdot |C_{k}^{\text{arg}}| \cdot \cos(\angle C_{k}^{\text{arg}})$$

(3.8)

Such equations will be used to evaluate the field asymmetries on the coupler midplane.

3.3 Theory of coupler matching

In this section we provide a necessary and sufficient condition for the coupler matching, expressed in terms of the S parameters. A linear accelerating structure is a device powered by a klystron, whose power is typically of some tens of MW. A linac works in a proper way if the input reflected power is negligible, precisely if the reflection coefficient at the input port has amplitude: $|\Gamma| < 0.02$. Moreover the correct phase advance per cell must be preserved.

3.3.1 Single feed

A single feed accelerating structure can be modeled with the two-ports network of Fig. 3.7. The first and the last two-ports blocks represent the input and the output couplers, respectively, while the inner blocks represent the linac cells, where $Z_C$ is the characteristic impedance of the accelerating hybrid mode, $L$ is longitudinal length of the the cell and $\beta_C$ is the wave number of the hybrid mode at the working frequency [35].
The terms $a_1$ and $b_1$ represent the forward and the backward wave intensity, respectively. The input reflection coefficient is therefore:

$$\Gamma_1 = \frac{b_1}{a_1} = S_{11} + S_{12}S_{21} \frac{S_{22}e^{-2n_2\beta_CL}}{1 - S_{22}^2e^{-2n_2\beta_CL}}$$

that can also be expressed as:

$$\Gamma_1 = S_{11} + S_{12}S_{21} \sum_{t=1}^{\infty} S_{22}^{t-1} e^{-2n_2\beta_CLt}$$

The coupler is matched when both parameters $S_{11}$ and $S_{22}$ converge to zero value during the coupler optimization process. This conditions assure that there is no reflection at the input port and no reflection inside the linac, therefore the cells work with the designed phase advance.

### 3.3.2 Dual feed

The previous calculations can be extended so as to design the dual feed coupler. It is important to consider the three-port circuit of Fig. 3.8 to analyze its matching in the dual-feeding case.

The dual feed is modeled setting the forward wave intensity of port 1 equal to the one of port 3. The input reflection coefficient at port 1, defined as: $\Gamma_1 = b_1/a_1$, must be evaluated by solving the following system:
3.3. THEORY OF COUPLER MATCHING

Figure 3.8: Three-ports circuit with dual feeding, closed on generic load.

\[
\begin{align*}
\begin{cases}
 b_1 &= S_{11}a_1 + S_{12}a_2 + S_{13}a_1 \\
 b_2 &= S_{21}a_1 + S_{22}a_2 + S_{23}a_1 \\
 b_3 &= S_{31}a_1 + S_{32}a_2 + S_{33}a_1 \\
 \Gamma_1 &= \frac{b_1}{a_1} \\
 \rho_2 &= \frac{b_2}{a_2}
\end{cases}
\end{align*}
\]

which gives:

\[
\Gamma_1 = (S_{11} + S_{13}) + S_{12}(S_{21} + S_{23}) \frac{\rho_2}{1 - S_{22}\rho_2} \quad (3.11)
\]

This three-port circuit is therefore matched when \((S_{11} + S_{13}) = 0\) and \(\rho_2 = 0\).

The full dual feed linac network is depicted in Fig. 3.9.

Applying Eq. (3.11) to the network of Fig. 3.9, the reflection coefficient at port 1 is given by:

\[
\Gamma_1 = (S_{11} + S_{13}) + S_{12}(S_{21} + S_{23}) \frac{S_{22}e^{-j2n\beta_CL}}{1 - S_{22}e^{-j2n\beta_CL}} \quad (3.12)
\]

which, by a series expansion of the term \(1/(1 - S_{22}e^{-j2n\beta_CL})\), yields:

\[
\Gamma_1 = (S_{11} + S_{13}) + S_{12}(S_{21} + S_{23}) \sum_{t=1}^{\infty} S_{22}^{2t-1}e^{-j2n\beta_CLt} \quad (3.13)
\]

Eq. (3.13) outlines the sequence of reflections inside the structure. The dual feed linac is therefore matched when \((S_{11} + S_{13}) = 0\) and \(S_{22} = 0\).
Symmetries of the dual feed coupler

The dual feed coupler has three ports and can be modeled with a 3x3 S-matrix. However, due to the geometrical symmetries, some S-parameters are equal. According to Fig. 3.8, the symmetries imply the following equalities:

\[
\begin{align*}
S_{11} &= S_{33} \\
S_{21} &= S_{23} \\
S_{12} &= S_{32} \\
S_{13} &= S_{31}
\end{align*}
\]  

(3.14)

3.3.3 Single feed with movable short-circuit

The structure involving the single feed with movable short-circuit consists of a coupler with the input port (port 1); the port connected to the accelerating cells (port 2); an additional port connected to a movable short-circuit, represented by \( \rho_3 \) (port 3). The input coupler is illustrated in Fig. 3.10. For the sake of simplifying the study, we suppose that the output coupler is already matched. In this way, the input coupler is connected, through port 2, to the characteristic impedance \( Z_C \).

Unlike the dual feed case, the coupler in Fig. 3.10 is not necessarily symmetric. The coupler is connected with the waveguides with two apertures, called “coupling windows”. In the latter case, the coupling windows of port 1 and port 3 are different. This introduces an additional degree of freedom in the design procedure as there are two width values of the coupling windows instead of only one. However, in order to ensure an easier design and man-
3.4. THE MICROWAVE DESIGN

This section describes our microwave design of the accelerating structure. The design of the basic cell is shown, and it is followed by the design of the new couplers.
3.4.1 The cell design

The basic cell has been designed with the HFSS [5] code. The geometry of Fig. 3.11 has been simulated using the eigenmode solver with the Master and Slave boundary condition, imposing the phase advance of the $2\pi/3$ mode, and the working frequency of 2.99801 GHz. The simulated building material is copper, while the cell medium is vacuum.

![Figure 3.11: Basic cell of the linac.](image)

The dimensions of the basic accelerating cell are listed in Table 3.1

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>11.000</td>
</tr>
<tr>
<td>b</td>
<td>39.366</td>
</tr>
<tr>
<td>L</td>
<td>33.333</td>
</tr>
<tr>
<td>t</td>
<td>5.840</td>
</tr>
</tbody>
</table>

The RF parameters are listed in Table 3.2. The sensitivities of the geometrical parameters are reported in Table 3.3: the external radius and the iris radius are the most important parameters for the mechanical tolerances.

3.4.2 The coupler design

In realizing the coupler-waveguide connection, we used a rectangular waveguide input with the “WR284” commercial standard. This allows to directly
3.4. THE MICROWAVE DESIGN

Table 3.2: RF parameters of the basic cell.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{RES}$</td>
<td>2998.01 MHz</td>
</tr>
<tr>
<td>$Q$</td>
<td>13300</td>
</tr>
<tr>
<td>$R_{SHUNT}$</td>
<td>60.3 $M\Omega/m$</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>0.1888 $Nm$</td>
</tr>
<tr>
<td>$v_g/c$</td>
<td>0.01251</td>
</tr>
<tr>
<td>$E_{z,0}$ @ 19 MW</td>
<td>20 MV/m</td>
</tr>
<tr>
<td>$</td>
<td>E</td>
</tr>
<tr>
<td>$BW$</td>
<td>41.52 MHz</td>
</tr>
<tr>
<td>$f_{MIN}$</td>
<td>2966.83 MHz</td>
</tr>
<tr>
<td>$f_{MAX}$</td>
<td>3008.35 MHz</td>
</tr>
</tbody>
</table>

Table 3.3: Sensitivities of the basic cell.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Sensitivity</th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>0.2 MHz/10 $\mu$m</td>
</tr>
<tr>
<td>b</td>
<td>-0.8 MHz/10 $\mu$m</td>
</tr>
<tr>
<td>t</td>
<td>0.05 MHz/10 $\mu$m</td>
</tr>
<tr>
<td>L</td>
<td>-0.03 MHz/10 $\mu$m</td>
</tr>
</tbody>
</table>

connect the coupler to the waveguide without using a taper. The end of the waveguide, before the coupling window, is rounded with the radius $R_G$ (Fig. 3.12). This simplifies the machining operation when the coupler is realized with the mill.

![Figure 3.12: Sizing of the coupler-waveguide connections, 3D view (a), XZ plane view (b).](image)

These dimensions are listed in Table 3.4.
Table 3.4: Sizing of the coupler-waveguide connections [mm].

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>$a_G$</td>
<td>72.136</td>
</tr>
<tr>
<td>$b_G$</td>
<td>34.036</td>
</tr>
<tr>
<td>$R_A$</td>
<td>1.5</td>
</tr>
<tr>
<td>$R_G$</td>
<td>8</td>
</tr>
<tr>
<td>$t$</td>
<td>5.84</td>
</tr>
<tr>
<td>$a$</td>
<td>11.000</td>
</tr>
</tbody>
</table>

The coupler design is described in detail in the following sections, where the other dimensions will be specified. We analyze the following different structures:

1. Single feed
2. Single feed translated
3. Dual feed
4. Dual feed with racetrack
5. Single feed with movable short-circuit
6. Single feed with movable short-circuit and racetrack

**Single feed**

The design of the single feed coupler has been performed by using Ansys HFSS [5]. The design aims to achieve the following results:

- reducing the magnitude of the reflection coefficient of the input waveguide in order to obtain $|\Gamma_1| < 0.02$;

- obtaining the correct phase advance per cell, for the $2\pi/3$ mode;

The coupler matching [45] is usually achieved by varying the coupler radius $R_c$ and the width $w$ of the coupling window. As already pointed out before, this design uses the feeding waveguide standard WR284, and such waveguide is directly connected to the coupler without the use of a taper. Therefore, the beampipe radius $R_{BP}$ is introduced as an additional degree of freedom (Fig.3.13).
3.4. THE MICROWAVE DESIGN

Figure 3.13: Single feed matching variables.

The matching is achieved with a trial-and-error procedure on the variables \( R_c, w \) and \( R_{bp} \). The coupler is considered as matched when \( |\Gamma_1| < 0.02 \) and it has the correct phase advance per cell [45]. The dimensions of the designed single feed coupler are listed in Table 3.5.

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_C )</td>
<td>38.290</td>
</tr>
<tr>
<td>( w )</td>
<td>25.460</td>
</tr>
<tr>
<td>( R_{BP} )</td>
<td>8.06</td>
</tr>
<tr>
<td>( t )</td>
<td>5.84</td>
</tr>
<tr>
<td>( R_A )</td>
<td>1.5</td>
</tr>
<tr>
<td>( R_G )</td>
<td>8</td>
</tr>
<tr>
<td>( a_G )</td>
<td>72.136</td>
</tr>
<tr>
<td>( b_G )</td>
<td>34.036</td>
</tr>
</tbody>
</table>

This matching strategy has been used also for the single feed translated, for the dual feed and for the dual feed with racetrack structures.

**Single feed translated**

The design of the single feed translated coupler has been performed with the same procedure of the previous single feed, introducing an additional
degree of freedom, that is the coupler translation, as shown in Fig. 3.3. The coupler translation allows to reduce the dipolar field asymmetry down to 0.1%. The dimensions of the designed single feed translated coupler are listed in Table 3.6, where $T$ is the translation along the $x$ axis, according to Fig. 3.3.

### Table 3.6: Dimensions of the single feed translated coupler [mm].

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_C$</td>
<td>38.300</td>
</tr>
<tr>
<td>$w$</td>
<td>25.457</td>
</tr>
<tr>
<td>$R_{BP}$</td>
<td>8.135</td>
</tr>
<tr>
<td>$T$</td>
<td>-1.303</td>
</tr>
<tr>
<td>$t$</td>
<td>5.84</td>
</tr>
<tr>
<td>$R_A$</td>
<td>1.5</td>
</tr>
<tr>
<td>$R_G$</td>
<td>8</td>
</tr>
<tr>
<td>$a_G$</td>
<td>72.136</td>
</tr>
<tr>
<td>$b_G$</td>
<td>34.036</td>
</tr>
</tbody>
</table>

**Single feed with movable short-circuit coupler**

The design of the single feed with movable short-circuit coupler has been performed by using Ansys HFSS [5]. The design aims to achieve the following results:

- reducing the magnitude of the reflection coefficient of the input waveguide in order to obtain $|\Gamma_1| < 0.02$;

- obtaining the correct phase advance per cell, for the $2\pi/3$ mode;

- reducing the dipolar field in the coupler, so as to satisfy the constrain $|\Delta E_d/E_z| \leq 0.01\%$.

The coupler matching is usually achieved by varying the coupler radius $R_c$ and the width $w$ of the coupling window [45]. As already pointed out before, this design uses the feeding waveguide standard WR284, and such waveguide is directly connected to the coupler without the use of a taper. Therefore, the beampipe radius $R_{bp}$ is introduced as an addition degree of freedom (Fig. 3.14).
3.4. THE MICROWAVE DESIGN

For any given position of the short-circuit specified by the length $SC$ of Fig. 3.14, the matching is achieved with a trial-and-error procedure on the variables $R_c$, $w$ and $R_{bp}$. The coupler is considered as matched when $|\Gamma_1| < 0.02$ and it has the correct phase advance per cell [45]. However, even if the coupler is matched, this does not ensure that the dipolar field is minimum. Its intensity depends on the short-circuit position. The reduction of the dipolar component is achieved by iteratively changing the short-circuit position, re-matching the coupler at each time. Fig. (3.15) shows how the dipole component varies with the short-circuit position, when the matching is achieved at each step.

We stopped the search of the optimal short-circuit position when the absolute value of the dipole component $|\Delta E_d/E_z|$ was nearly 0.01%, or less. The dimensions of the single feed with movable short-circuit coupler are listed in Table 3.7.

**Single feed with movable short-circuit and racetrack coupler**

In the previous design the quadrupolar component was: $\Delta E_q/E_z = 2.3\%$. Now we show that such component can be reduced. To this aim, we extend the racetrack geometry concept [9] [41] [44], showing that it can be applied to a single feed with movable short-circuit, obtaining the coupler structure of Fig. 3.16.
Figure 3.15: *Plot of the dipole component for a small variation of the dimension “SC”.*

Table 3.7: Dimensions of the single feed with movable short-circuit coupler [mm].

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_C$</td>
<td>37.710</td>
</tr>
<tr>
<td>$w$</td>
<td>25.520</td>
</tr>
<tr>
<td>$R_{BP}$</td>
<td>8.67</td>
</tr>
<tr>
<td>$SC$</td>
<td>99.5</td>
</tr>
<tr>
<td>$t$</td>
<td>5.84</td>
</tr>
<tr>
<td>$R_A$</td>
<td>1.5</td>
</tr>
<tr>
<td>$R_G$</td>
<td>8</td>
</tr>
<tr>
<td>$a_G$</td>
<td>72.136</td>
</tr>
<tr>
<td>$b_G$</td>
<td>34.036</td>
</tr>
</tbody>
</table>

The aim of our design is that of achieving the following results:

- reducing the reflection coefficient at the input waveguide ($|\Gamma_1| < 0.02$);
- obtaining the correct phase advance per cell, for the $2\pi/3$ mode;
- reducing the quadrupolar field intensity so as to satisfy the constrain $|\Delta E_q/E_z| \leq 0.01\%$. 
3.4. THE MICROWAVE DESIGN

Figure 3.16: Racetrack geometry applied to the single feed coupler with movable short-circuit. Front view (a), back view (b).

- reducing the dipolar field so as to satisfy the condition $|\Delta E_d/E_z| \leq 0.01\%$.

For given values of $RT$ and $SC$ (Fig. 3.16), the matching is achieved with a trial and error procedure on the variables $R_c$, $w$ and $R_{bp}$. The coupler is matched with $|\Gamma_1| < 0.02$ and with the correct phase advance per cell of the $2\pi/3$ mode [45]. The quadrupolar field component depends on the $RT$ dimension. Therefore, the reduction of the quadrupolar component is achieved by iteratively changing the racetrack dimension, matching the coupler at each step. In order to keep low the dipolar component, the short circuit $SC$ parameter can be adjusted. Fig. 3.17 shows how the quadrupolar component reduces with the racetrack geometry, in a matched single feed with movable short-circuit and racetrack geometry.

The optimum point has been chosen in order to achieve an electric field asymmetry lower or equal to 0.01% of both the dipolar and the quadrupolar field components. The dimensions of the single feed with movable short-circuit and racetrack geometry are listed in Table 3.8.

3.4.3 Comparison of the coupler asymmetries

The coupler field asymmetries of the six types of couplers listed in subsection 4.2 have been evaluated with Eqs. (3.5), (3.6), (3.7) and (3.8). Table 3.9 summarizes the simulated results, where the contributions of the dipolar,
quadrupolar, sextupolar and octupolar components are indicated with “d”, “q”, “s” and “o”, respectively.

Column 5 shows that the magnitude gradient of the dipole electric field is strongly reduced in the case of symmetric coupler with one waveguide input. Column 6 shows that the magnitude gradient of both the dipole electric and the quadrupole electric field, are strongly reduced in the case of single feed
3.5 Prototype manufacturing

This section shows how a linac prototype has been realized at Elettra - Sincrotrone Trieste S.C.p.A. [1]. All the prototype components are fabricated in aluminum. The machined accelerating structure has a single feed with movable short-circuit coupler and a normal single feed coupler. The reason of this choice is to allow a comparison between the simulated results obtained with the coupler with movable short-circuit, and the results obtained using a single feed coupler.

### 3.5.1 Machining of the cells

As a first step, we machined the cells, with the aim of performing some preliminary standing wave measurements and tuning. Five cells like that of Fig. 3.11 have been realized with a lathe, the dimensions are reported in Table 3.1. Such cells have been designed with the ‘C’ profile, as indicated in Fig. 3.18. A threaded hole, together with a screw that reduces the volume of the cell, is used to tune the resonant frequency.

<table>
<thead>
<tr>
<th></th>
<th>Single feed</th>
<th>Single feed translated</th>
<th>Dual feed</th>
<th>Dual feed racetrack</th>
<th>Single feed movable short</th>
<th>Single feed movable short and racetrack</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Delta E_d / E_z$</td>
<td>6%</td>
<td>-0.1%</td>
<td>0</td>
<td>0</td>
<td>0.01%</td>
<td>-0.01%</td>
</tr>
<tr>
<td>$\Delta \varphi_d$</td>
<td>0.6°</td>
<td>0.6°</td>
<td>0°</td>
<td>0°</td>
<td>0.6°</td>
<td>0.6%</td>
</tr>
<tr>
<td>$\Delta E_q / E_z$</td>
<td>1.2%</td>
<td>1.2%</td>
<td>1.6%</td>
<td>0.01%</td>
<td>2.3%</td>
<td>-0.01%</td>
</tr>
<tr>
<td>$\Delta \varphi_q$</td>
<td>0.1°</td>
<td>0.13°</td>
<td>0.12°</td>
<td>0.12°</td>
<td>0.13°</td>
<td>0.12°</td>
</tr>
<tr>
<td>$\Delta E_s / E_z$</td>
<td>0.24%</td>
<td>0.3%</td>
<td>0</td>
<td>0</td>
<td>8e-4%</td>
<td>1e-3%</td>
</tr>
<tr>
<td>$\Delta \varphi_s$</td>
<td>0.02°</td>
<td>0.025°</td>
<td>0°</td>
<td>0°</td>
<td>0.02°</td>
<td>0.03°</td>
</tr>
<tr>
<td>$\Delta E_o / E_z$</td>
<td>0.23%</td>
<td>0.24%</td>
<td>0.27%</td>
<td>0.27%</td>
<td>0.32%</td>
<td>0.33%</td>
</tr>
<tr>
<td>$\Delta \varphi_o$</td>
<td>7e-4°</td>
<td>1e-4°</td>
<td>6e-4°</td>
<td>1e-3°</td>
<td>7e-5°</td>
<td>2e-3°</td>
</tr>
</tbody>
</table>

input with movable short-circuit and racetrack geometry. Moreover, these two latter single feed solutions do not require neither the use of a power splitter nor the use of a dual waveguide feeding.
In order to perform the standing wave measurements, two auxiliary semi-cells have been machined, each one closed with two metallic plates acting as short-circuits, which allow the standing wave generation. Each auxiliary semi-cell has half volume of each of the five main cells. All these cells are close by three threaded rods, obtaining the structure of Fig. 3.19, that will be used for the standing wave measurements.

The components are:

1. Five main cells;
2. Inferior half cell;
3. Superior half cell;
4. Inferior plate;
5. Superior plate;
6. Threaded rod;
7. Antennas with SMA (Sub Miniature version A) connectors, aimed to realize the standing wave measurements.

The antennas will be used for the standing wave measurements.
3.5. PROTOTYPE MANUFACTURING

3.5.2 Machining of the coupler

After the cell machining, the linac is completed with the production of the input and output coupler. As explained above, one coupler is the single feed with movable short-circuit, while the other is the normal single feed. The single feed with movable short-circuit has the geometry of Fig. 3.14, and has been fabricated using a mill. It is closed with a lid, as shown in Fig. 3.20. The cavity radius $R_C$ and the coupling window $w$ (Fig. 3.14) have been machined with $10\mu m$ of precision. The reason is that they are very critical parameters.

The lid is the beginning of the beampipe, in which the evanescent mode of the accelerating RF power decays. In the other side, the cells are connected with the coupler. Since the cells are of ‘C’ type (Fig. 3.18), an indentation has been made in the coupler, as shown in Fig. 3.21.

The other coupler has been designed and machined in the same way,
with the only difference that it is a single feed coupler, so that it has only one coupling window. Fig. 3.22 shows a photo of the produced single feed coupler.

Since the single feed coupler of Fig. 3.22 has the same indentation of the other coupler (Fig. 3.21), an additional cell is needed to connect the ‘C’ cells with the two couplers. This is a ring, as shown in Fig. 3.23, and it is endowed with a tuner.
3.6 Microwave measurements

This section reports the microwave measurements performed on the realized prototype. Two types of measurements have been performed:

1. Standing wave measurements, made with the cells;
2. Travelling wave measurements, made with the bead-pull method on the linac.
3.6.1 Standing wave measurements

The standing wave measurements have been performed on the structure of Fig. 3.25 whose scheme is shown in Fig. 3.19.

The threaded rods are closed with the torque wrench, that allows to equally close the nuts, with a torque that can vary from 6 Nm to 8 Nm. A high closing torque is needed in order to achieve a good electric contact between the cells. The environmental temperature, during all our measurements was $26^\circ C$. The frequency for a S-band structure varies of nearly $-50kHz/^\circ C$.

Since the structure is in air, the resonant frequency is about $1MHz$ less than the simulated value in vacuum. The measurements have been carried out with a vector network analyzer, measuring the $S_{21}$ parameters between the two antennas, placed in the superior plate. The antennas are entirely contained in the hole made in the plate, in order to avoid modifications of the resonant frequencies. The typical $S_{21}$ parameter is -50dB, this means...
that the antennas are strongly undercoupled. At first, the measurements were performed only on the two semi-cells, in order to tune the resonant frequencies of the 0 and of the $\pi$ mode. After this, a cell is included between the two semi-cells, obtaining a structure with three coupled cavities, measuring therefore the three resonant frequencies of the mode 0, $\pi/2$ and $\pi$. The measurements went on adding a cell at each time, by measuring and tuning the resonant frequencies and the respective quality factors, until reaching the full structure of Fig. 3.25, composed by seven coupled cavities. The resonant frequencies are listed in Table 3.10.

Table 3.10: Measured resonant frequencies [MHz], with a closing torque of 8 Nm.

<table>
<thead>
<tr>
<th>Mode, #</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>AVG</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>2965.72</td>
<td>2965.52</td>
<td>2965.64</td>
<td>2965.57</td>
<td>2965.44</td>
<td>2965.36</td>
<td>2965.54</td>
</tr>
<tr>
<td>$\pi/6$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2968.37</td>
</tr>
<tr>
<td>$\pi/5$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2969.57</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$\pi/4$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2971.93</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$\pi/3$</td>
<td>-</td>
<td>-</td>
<td>2976.43</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2976.16</td>
</tr>
<tr>
<td>$2\pi/5$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2980.42</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$\pi/2$</td>
<td>-</td>
<td>2986.58</td>
<td>-</td>
<td>2986.96</td>
<td>-</td>
<td>-</td>
<td>2986.47</td>
</tr>
<tr>
<td>$3\pi/5$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2992.97</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$2\pi/3$</td>
<td>-</td>
<td>-</td>
<td>2996.58</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>2996.5</td>
</tr>
<tr>
<td>$3\pi/4$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3000.82</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$4\pi/5$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3002.91</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>$5\pi/6$</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>3004.12</td>
</tr>
<tr>
<td>$\pi$</td>
<td>3006.68</td>
<td>3006.96</td>
<td>3007.34</td>
<td>3007.38</td>
<td>3007.17</td>
<td>3007.07</td>
<td>3007.1</td>
</tr>
</tbody>
</table>
The measured resonant frequencies can be used to construct the dispersion diagram of the cell. The simulated and the measured dispersion diagram (after tuning), are reported in Fig. 3.26, where the points have been interpolated with a Lagrange polynomial.

![Simulated and measured dispersion diagrams of the cell.](image)

The two curves are in good agreement. After the above procedure, the cells have been tuned and can be assembled with the two couplers in order to build the whole accelerating structure.

### 3.6.2 Bead-pull measurements

The bead-pull measurements [46] have been performed on the whole linac. A picture of the assembled linac with the bead-pull measuring system is shown in Fig. 3.24

The four threaded rods are closed with the torque wrench, that allows to apply a uniform torque on the closing nuts. The typical closing torque is 6 Nm or 8 Nm. The measurements have been carried on with a vector network analyzer, measuring the $S_{11}$ parameter at one port, while the other is closed on a matched load. The bead-pull measuring system is based on a step by step motor, that moves a nylon rope on which a metallic needle is placed. The needle is a cylinder having radius of 0.5 mm and length of 3 mm. It perturbs the electromagnetic field in the $z$ direction. When the needle is outside the linac, the perturbation produced by the rope is negligible. When the motor
is enabled, the metallic needle enters in the linac, thus perturbing the field. Such perturbation is seen on the input port by watching a small detuning of the structure, and the $S_{11}$ parameter increases. Other perturbators can be used, such as a metallic or dielectric sphere.

**Longitudinal field measurements**

The first bead-pull measurement is made to verify the correct phase advance per cell, once the linac has been tuned. This measurement is made by moving with constant velocity the metallic needle inside the linac, starting from the input coupler and arriving to the output coupler. The vector network analyzer is set on the “Time domain” mode, with a constant frequency of 2996.37 MHz. The VNA records the amplitude and the phase variations when the bead is crossing the center section of the linac. Fig. 3.28 shows the magnitude $[\text{dB}]$ and the phase $[\text{o}]$ of the $S_{11}$ parameter as a function of the bead position.

These plots show a good tuning of the linac, in fact, we measured the correct phase advance of the $2\pi/3$ mode. The next step is that of measuring the coupler field.

**Transversal field measurements**

The purpose of this measurement is to evaluate the field asymmetry in the two realized couplers (Single feed with movable short-circuit and single feed coupler). The asymmetry measurement is performed by putting the bead in the center of the coupler, and by moving it transversally, simultaneously recording the $S_{11}$ parameter. According to [46], the electric field magnitude
Figure 3.28: *Longitudinal bead-pull measurements performed when the bead is moving through the linac center.*

is given by:

\[ |E_z(x)| = k \cdot \sqrt{|S_{11}(x)|} \]  

(3.18)

where \( x \) is the bead position and \( k \) is a constant. With this equation a relative field amplitude variation can be evaluated. The term \( \Delta E_d/E_z \) can be calculated by substituting Eq. (3.18) into (3.5), and the unknown term \( k \) is simplified.

Fig. 3.29 illustrates the normalized electric fields, measured in the single feed with movable short coupler midplane. The optimal short-circuit position resulted to be \( SC = 93.85 \) mm. The corresponding field asymmetry is \( |\Delta E_d/E_z| = 0.1\% \), according to the precision of our bead-pull measurement system. In the normal single feed coupler, the field asymmetry, measured in the coupler midplane, is illustrated in Fig. 3.30.

In this coupler the field asymmetry resulted to be 6\%, in agreement with the simulation results. These measurements show that the single feed with movable short-circuit coupler allows to adjust the field asymmetry as wanted within the measurement system precision.
3.7 Evaluation of the particle beam motion

This section describes a method for evaluating the beam dynamics in the designed couplers. As a first step, the case of a single particle is considered. Subsequently, the beam is approximated with two particles: the head particle and the tail particle. The space-charge effects are neglected. The method
exploits a particle tracking software developed in-house with MatLab [47]. The algorithm discretizes the relativistic Lorentz force [9] and numerically integrates it in the time domain.

### 3.7.1 Discretizing the Lorentz force

The relativistic Lorentz Force is given by:

$$
\vec{F} = \frac{d\vec{p}}{dt} = m_0 c \frac{d(\gamma \vec{\beta})}{dt} = e(\vec{E} + c \vec{\beta} \times \vec{B})
$$

(3.19)

where the electric field $\vec{E}$ and the magnetic induction $\vec{B}$ fields are given by the simulator. The following system of differential equations is solved in the three unknowns $\gamma \beta_x$, $\gamma \beta_y$ and $\gamma \beta_z$:

$$
\begin{align*}
\frac{d(\gamma \beta_x)}{dt} &= \frac{e}{m_0 c} [E_x + c(\beta_y B_z - \beta_z B_y)] \\
\frac{d(\gamma \beta_y)}{dt} &= \frac{e}{m_0 c} [E_y + c(\beta_z B_x - \beta_x B_z)] \\
\frac{d(\gamma \beta_z)}{dt} &= \frac{e}{m_0 c} [E_z + c(\beta_x B_y - \beta_y B_x)]
\end{align*}
$$

(3.20)

where $\gamma$ is the relativistic factor, given by:

$$
\gamma = \frac{1}{\sqrt{1 - \beta_z^2}}
$$

(3.21)

In the time domain, the variations of the unknowns $\gamma \beta_x$, $\gamma \beta_y$, $\gamma \beta_z$, in a small time interval $\Delta t$, are given by:

$$
\begin{align*}
\gamma \beta_x(t+1) - \gamma \beta_x(t) &= f_x(\vec{E}, \vec{B}, \beta_y, \beta_z)\Delta t \\
\gamma \beta_y(t+1) - \gamma \beta_y(t) &= f_y(\vec{E}, \vec{B}, \beta_x, \beta_z)\Delta t \\
\gamma \beta_z(t+1) - \gamma \beta_z(t) &= f_z(\vec{E}, \vec{B}, \beta_x, \beta_y)\Delta t
\end{align*}
$$

(3.22)

In Eq. (3.22), the values of $\beta_x$, $\beta_y$ and $\beta_z$ are updated at each step. Observing that, since $\gamma = \sqrt{1 + (\gamma \beta_z)^2}$, we can write the following relations:

$$
\begin{align*}
\beta_x &= \frac{\gamma \beta_x}{\sqrt{1 + (\gamma \beta_x)^2}} \\
\beta_y &= \frac{\gamma \beta_y}{\sqrt{1 + (\gamma \beta_x)^2}} \\
\beta_z &= \frac{\gamma \beta_z}{\sqrt{1 + (\gamma \beta_z)^2}}
\end{align*}
$$

(3.23)
3.7. EVALUATION OF THE PARTICLE BEAM MOTION

that can be substituted into Eq. (3.22). This approach allows to calculate, at each step \( t \), the values \( \gamma \beta_x, \gamma \beta_y, \gamma \beta_z, \beta_x, \beta_y, \beta_z \), and the particle spatial positions, given by:

\[
\begin{align*}
  x(t + 1) &= x(t) + c \beta_x \Delta t \\
  y(t + 1) &= y(t) + c \beta_y \Delta t \\
  z(t + 1) &= z(t) + c \beta_z \Delta t
\end{align*}
\] (3.24)

3.7.2 Multipole decomposition

The next step consists of decomposing in multipoles the transversal momentum:

\[
\gamma \vec{\beta}_\perp = \gamma \beta_x \hat{x} + \gamma \beta_y \hat{y}
\] (3.25)

where \( \hat{x} \) and \( \hat{y} \) are the unit-vectors of the axes \( x \) and \( y \), respectively, depicted in Fig. 3.6. To this aim, \( \gamma \vec{\beta}_\perp \) can be expanded as:

\[
\begin{align*}
  \gamma \beta_x &= A_0 \cdot x + D_x + Q \cdot x + S \cdot y + O_x(x, y) \\
  \gamma \beta_y &= A_0 \cdot y + D_y - Q \cdot y + S \cdot x + O_y(x, y)
\end{align*}
\] (3.26)

where: \( A_0 \) is the RF focusing due to the beampipe fringe fields and to the field distribution of the accelerating mode, \( D_x \) and \( D_y \) are the dipoles oriented in the \( x \) and \( y \) direction, \( Q \) is the quadrupole, \( S \) the skew quadrupole, \( O_x(x, y) \) and \( O_y(x, y) \) are higher-order infinitesimal terms. The calculation of these five values can be carried out using one of the following two methods, described below:

1. Fitting;

2. Fourier transform.

Fitting

Running the algorithm in correspondence of a number \( N \) (\( N = 121 \) in our case, obtained with a square grid composed of 11 values in \( x \), times 11 values in \( y \)) of different initial conditions \( x = x_0 n, y = y_0 n, n = 1, 2, ..., N \), Eq. (3.26) leads to the following system of \( 2N \) equations in the five unknowns \( A_0, D_x, D_y, Q \) and \( S \):
\[
\begin{align*}
\gamma \beta_x &= A_0 x'_0 + D_x + Q x'_0 + S y'_0 \\
\gamma \beta_y &= A_0 y'_0 + D_y - Q y'_0 + S x'_0
\end{align*}
\] (3.27)

with \(n=1,2,\ldots, N\)

where the higher-order infinitesimal terms are neglected, and \(x'_0, y'_0\) are the final positions of the particle. However, since the particle is considered as relativistic, using the thin-lens approximation, only variations of the momentum can be considered, \(x'_0 \approx x_0\) and \(y'_0 \approx y_0\). This system is solved with the Least Square Method. If the sextupole and and the ottupole modes are not neglectable, they can be added in Eq. (3.26). In the latter case, a possible alternative approach is the use of the Fourier transform, described in the next subsection.

**Fourier transform**

The term \(\gamma \beta_r(R, \phi)\), with \(R\) constant is expanded in Fourier series, and the Fourier coefficients allow to easily calculate the unknowns. We considered a number \(N\) of initial conditions

\[
\begin{align*}
x_{0n} &= R \cdot \cos(\phi_n) \\
y_{0n} &= R \cdot \sin(\phi_n)
\end{align*}
\] (3.28)

where \(R\) is assigned and \(\phi_n = \frac{2\pi}{N}(n - 1)\), with \(n = 1, \ldots, N\). Running the algorithm for each of these \(N\) initial conditions, provides the quantities \(\gamma \beta_x\) and \(\gamma \beta_y\), and therefore \(\gamma \beta_r = \gamma \beta_x \cdot \cos(\phi) + \gamma \beta_y \cdot \sin(\phi)\) which is the focusing component in our case. Since \(\gamma \beta_x = \gamma \beta_r \cdot \cos(\phi)\) and \(\gamma \beta_y = \gamma \beta_r \cdot \sin(\phi)\), introducing the sextupole and the ottupole terms in Eq. (3.26), and manipulating, yields:

\[
\gamma \beta_r = A_0 R'_n + D_x \cos(\phi'_n) + D_y \sin(\phi'_n) + Q R'_n \cos(2\phi'_n) + S R'_n \sin(2\phi'_n) + E R'_n^2 \cos(3\phi'_n) + S E R'_n^2 \sin(3\phi'_n)
\] (3.29)

where \(R'_n\) and \(\phi'_n\) are the final polar coordinates of the particle at the end of the coupler. As observed above, using the thin-lens approximation, we can write \(R'_n \approx R\), and \(\phi'_n \approx \phi_n\). The term \(\gamma \beta_r(\phi)\) can be expanded in Fourier series, that is:

\[
\gamma \beta_r(\phi) = a_0 + \sum_{m=1}^{+\infty} (a_m \cos(m \cdot \phi) + b_m \sin(m \cdot \phi))
\] (3.30)
3.7. EVALUATION OF THE PARTICLE BEAM MOTION

A comparison between Eq. (3.29) and Eq. (3.30) yields immediately:

\[ A_0 = a_0/R \]
\[ D_x = a_1 \]
\[ D_y = b_1 \]
\[ Q = a_2/R \]
\[ S = b_2/R \]
\[ E = a_3/R^2 \]
\[ S_E = b_3/R^2 \]

(3.31)

where \( a_0, a_1, b_1, a_2, b_2, a_3, b_3 \) are the Fourier coefficients, given by:

\[ a_0 = \frac{1}{2\pi} \int_0^{2\pi} \gamma \beta_r(\phi) d\phi \]
\[ a_m = \frac{1}{\pi} \int_0^{2\pi} \gamma \beta_r(\phi) \cdot \cos(m\phi) d\phi \]
\[ b_m = \frac{1}{\pi} \int_0^{2\pi} \gamma \beta_r(\phi) \cdot \sin(m\phi) d\phi \]

(3.32)

Note that, with the latter approach, it is possible to evaluate the effect of the sextupole \((E)\), of the skew sextupole \((S_E)\), and of the higher-order terms.

3.7.3 Tracking results

The above numerical approach has been applied in the case where the beam energy at the input coupler is 5 MeV, while the linac accelerating gradient is 20 MeV/m. After evaluating the beam dynamics, the multipole decomposition has been carried on using both the “Fitting” and the “Fourier transform” methods, obtaining the same results. For each of the six types of couplers itemized in subsection 4.2, the terms \( D_x, D_y, A_0, Q, S \) and \( E \) have been calculated varying the RF phase \( \varphi_{RF} \) (the particle beam is in crest when \( \varphi_{RF} = 0 \)).

The dipolar component has been analyzed on the “Single feed” type accelerating structures. According to the reference system of Fig. 3.6, the dipole is oriented along the \( x \) direction, so that \( D_y = 0 \). For the couplers “Single feed”, “Single feed translated”, “Single feed with movable short-circuit”, and “Single feed with movable short-circuit and racetrack” the \( D_x \) component is shown in Fig. 3.31.

Eq. (3.4) can be used to calculate the difference between the momentum of two particles that have different RF phase values. We consider a bunch 5°
long, which centroid is accelerated -2.5° off-crest. The dipole bunch head-tail angle can be evaluated by:

\[ \Delta x' = \frac{\Delta p_x}{p_x} = \frac{\Delta\gamma\beta_x}{\gamma} \]  \hfill (3.33)

By using the simulation results of Fig. 3.31, it is possible to evaluate the quantity \( \Delta(\gamma/\beta x) \), listed in Table 3.11.

Table 3.11: Comparison of the dipole effects for a 5 MeV bunch.

<table>
<thead>
<tr>
<th>Head tail dipole effects in the input coupler</th>
<th>( \Delta(\gamma/\beta x) )</th>
<th>HT angle (( \mu )rad)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Single feed</td>
<td>6.5e-3</td>
<td>602</td>
</tr>
<tr>
<td>Single feed translated</td>
<td>0.1e-3</td>
<td>9.5</td>
</tr>
<tr>
<td>Single feed s.c.</td>
<td>0.24e-3</td>
<td>23</td>
</tr>
<tr>
<td>Single feed s.c. and r.t.</td>
<td>0.25e-3</td>
<td>23</td>
</tr>
</tbody>
</table>

The skew quadrupole component (\( S \)) is not present in the simulated device, due to the symmetry of the structure. The effect of the quadrupole (\( Q \)), analyzed on the “Dual feed” and on the “Dual feed with racetrack” structures, is shown in Fig. 3.32.

The effects of the quadrupole (\( Q \)), analyzed on the structures “Single feed”, “Single feed translated”, “Single feed with movable short-circuit”, and
3.7. EVALUATION OF THE PARTICLE BEAM MOTION

Figure 3.32: Plot of the quadrupole kick for the dual feed structures.

“Single feed with movable short-circuit and racetrack” are shown in Fig. 3.33.

The quadrupole bunch head-tail effects can be evaluated as previously done for the dipoles. For the same bunch 5° long, accelerated -2.5° off-crest, with an initial energy of 5 MeV, the values of $\Delta(\gamma\beta_x)$ for the quadrupole are listed in Table 3.12 (In the table, $\Delta(\gamma\beta_y)$ is omitted, being equal to $-\Delta(\gamma\beta_x)$).
Table 3.12: Comparison of the quadrupole effects for a 5 MeV bunch.

<table>
<thead>
<tr>
<th></th>
<th>Δ(γβ_x)/m</th>
<th>HT angle (mrad/m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Single feed</td>
<td>230e-3</td>
<td>21</td>
</tr>
<tr>
<td>Single feed translated</td>
<td>240e-3</td>
<td>22</td>
</tr>
<tr>
<td>Dual feed</td>
<td>315e-3</td>
<td>29</td>
</tr>
<tr>
<td>Dual feed racetrack</td>
<td>7.4e-3</td>
<td>0.7</td>
</tr>
<tr>
<td>Single feed s.c.</td>
<td>450e-3</td>
<td>42</td>
</tr>
<tr>
<td>Single feed s.c. and r.t.</td>
<td>5.6e-3</td>
<td>0.5</td>
</tr>
</tbody>
</table>

From the tracking results, a strong correlation can be observed between the field asymmetries evaluated in Table 3.9 and the final kick received by the beam. The quadrupolar kick given by the “Single feed with movable short-circuit and racetrack” structure has the same order of magnitude as the one of the “Dual feed with racetrack” structure. This kick cannot be completely eliminated, due to the presence of a quadrupole phase gradient.

The sextupolar component has been analyzed on the “Single feed” type accelerating structures. According to the reference system of Fig. 3.5, the sextupole is oriented along the x direction, so that the skew sextupole is null. For the couplers “Single feed”, “Single feed translated”, “Single feed with movable short-circuit”, and “Single feed with movable short-circuit and racetrack” the E component is shown in Fig. 3.34. The sextupole bunch head-tail effects can be evaluated as previously done, with the same bunch 5° long, accelerated -2.5° off-crest, the values of Δ(γβ_x) for the quadrupole are listed in Table 3.13. The “Single feed with movable short-circuit and racetrack” gives the lowest sextupole kick between these four single feed structures. The biggest sextupole kick is given in the “Single feed trans-
3.7. EVALUATION OF THE PARTICLE BEAM MOTION

Figure 3.34: Plot of the sextupole kick calculated varying the RF phase.

lated” structure, because the coupler translation introduces an additional geometrical asymmetry that increases the high order terms.
Chapter 4

The high energy RF deflectors

4.1 The RF deflectors for FERMI@Elettra

The RF deflectors are travelling wave disk loaded structures in which an electromagnetic field can interact with ultra-relativistic electron bunches, providing a constant transversal force [48]. When the deflector is turned on, the electron bunch is stretched by a transversal deflecting voltage and is forced to collide with a detector screen, as illustrated in Fig. 4.1, thus recording the beam longitudinal profile. The beam trace on the screen is then converted into an optical signal to be processed. When the deflector is turned off, the detector screen is removed and the beam simply passes through the device without being deflected, and is used to produce a free electron laser (FEL) radiation.

![Figure 4.1: RF beam deflection.](image)

The hybrid deflecting mode is called $H E M_{11}$ [48]. It has a phase velocity
synchronous with the speed of the relativistic electron bunch, and allows an
effective energy exchange, as described by the Panofsky-Wenzel theorem [40].
Radio-frequency deflectors are used to perform bunch length measurements
and phase space diagnostics in the light source accelerators [49] [50]. These
measurements are used to assess the quality of the electron beam, before
entering in the undulator chains, at 1.2 GeV.

Two High Energy RF Deflectors (HERFD) have been realized for the
FERMI@Elettra project. They have the aim of stretching the electron bunch
vertically and horizontally, respectively. The two cavities are individually
powered by the same klystron and a switch system is used to choose the
deflection plane.

The specifications of the two HERFD are reported in Table 4.1.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency ($f$)</td>
<td>2998.01 MHz</td>
</tr>
<tr>
<td>Deflecting voltage ($V_t$)</td>
<td>20 MV</td>
</tr>
<tr>
<td>Filling time ($t_f$)</td>
<td>500 ns</td>
</tr>
<tr>
<td>Input power ($P_{in}$)</td>
<td>15 MW</td>
</tr>
<tr>
<td>Length</td>
<td>2.5 m</td>
</tr>
</tbody>
</table>

The following sections will show the HERFD measurements.

### 4.2 RF measurements

This section reports the RF measurements performed during the acceptance
test, carried out in the PMB-Alcen manufacturing company [8]. The high
energy deflector, is composed by 72 cells. Fig. 4.2 shows the cells endowed
with the tuners, while Fig. 4.3 shows the whole structure assembled for the
RF measurements.

For the correct operation of these two devices, the most important param-
eter verified in the acceptance test, is the phase advance $\Delta \varphi$, which is 120°
$\pm$ 1.5° (RMS), for the $2\pi/3$ mode. If the phase advance specification is not
respected, the deflecting voltage $V_t$ is lower than the designed one [48]. The
phase advance has been measured with the “short” method [51], by using a
plunger acting as a short circuit. Such plunger slides inside the linac and
allows to measure the phase advance of each cell, step by step. Figures 4.4
4.2. RF MEASUREMENTS

Figure 4.2: Cells of the high energy deflector.

Figure 4.3: High energy deflector composed by seventy two cells plus two couplers.

and 4.5 show the final phase advance of the vertical and horizontal deflector, after brazing.
CHAPTER 4. THE HIGH ENERGY RF DEFLECTORS

Figure 4.4: Phase advance of the vertical deflector, after brazing.

Figure 4.5: Phase advance of the horizontal deflector, after brazing.

Tables 4.2 and 4.3 show the RF parameters of the vertical and horizontal deflectors, respectively, before and after brazing.

The “SWR” parameter is also important: when it lies in the interval [1.0, 1.1], the structure is considered as matched. The resonant frequency $f$, reported in Tables 4.2 and 4.3 is measured in air and at the environmental temperature of 20 °C. When the deflector is installed and placed under vac-
4.2. RF MEASUREMENTS

Table 4.2: Acceptance test RF measurements for the vertical deflector.

<table>
<thead>
<tr>
<th></th>
<th>Specifications</th>
<th>Before brazing</th>
<th>After brazing</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency [MHz]</td>
<td>-</td>
<td>2997.94</td>
<td>2998.09</td>
</tr>
<tr>
<td>SWR</td>
<td>1.0 ÷ 1.1</td>
<td>1.005</td>
<td>1.02</td>
</tr>
<tr>
<td>$S_{21}$ [dB]</td>
<td>-2.97</td>
<td>-4.47</td>
<td>-3.27</td>
</tr>
<tr>
<td>Attenuation [Np/m]</td>
<td>0.137</td>
<td>0.208</td>
<td>0.145</td>
</tr>
<tr>
<td>Filling time [ns]</td>
<td>500</td>
<td>525</td>
<td>521</td>
</tr>
<tr>
<td>Quality factor</td>
<td>13200</td>
<td>9600</td>
<td>13500</td>
</tr>
<tr>
<td>$\Delta \phi$ [°]</td>
<td>120 ± 1.5</td>
<td>120.22 ± 0.46</td>
<td>119.85 ± 0.69</td>
</tr>
</tbody>
</table>

Table 4.3: Acceptance test RF measurements for the horizontal deflector.

<table>
<thead>
<tr>
<th></th>
<th>Specifications</th>
<th>Before brazing</th>
<th>After brazing</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency [MHz]</td>
<td>-</td>
<td>2997.56</td>
<td>2997.64</td>
</tr>
<tr>
<td>SWR</td>
<td>1.0 ÷ 1.1</td>
<td>1.007</td>
<td>1.016</td>
</tr>
<tr>
<td>$S_{21}$ [dB]</td>
<td>-2.97</td>
<td>-3.72</td>
<td>-3.39</td>
</tr>
<tr>
<td>Attenuation [Np/m]</td>
<td>0.137</td>
<td>0.173</td>
<td>0.158</td>
</tr>
<tr>
<td>Filling time [ns]</td>
<td>500</td>
<td>524</td>
<td>525</td>
</tr>
<tr>
<td>Quality factor</td>
<td>13200</td>
<td>11400</td>
<td>12700</td>
</tr>
<tr>
<td>$\Delta \phi$ [°]</td>
<td>120 ± 1.5</td>
<td>119.85 ± 0.26</td>
<td>119.82 ± 0.70</td>
</tr>
</tbody>
</table>

uum, its working frequency is nearly 1 MHz higher than that measured in air. The value of 2998.01 MHz is obtained with the temperature tuning of the device, knowing that, for an S-band structure, the frequency variation nearly -50 kHz/°C. The other RF parameters depend on the geometry of the structure. The quality factor increases after the brazing process, due to the soldering material that improved the electric contact between the cells.

The two deflectors have been accepted and installed at the end of the linac 4 of the FERMI@Elettra project, which layout is depicted in Fig. 4.6.

Fig. 1.3 of chapter 1 shows the vertical deflector, while Fig. 4.7 shows the horizontal deflector, both successfully installed.
4.3 RF conditioning

The RF conditioning is the first power operation of the device. Since in the waveguides components of dust, water and impurities are present, the startup RF cannot run with the designed full power. The RF heating of the waveguides and of the structure walls cause the degradation of the vacuum, which will not be sufficient to support such amount of power. The conditioning process is made by gradually increasing the RF power, starting with a RF power lower than 1 MW, and with an RF pulse width of 100 ns. When the power rises up for the first time, the RF heating of the waveguides and of the structure walls releases some impurities that cause peaks in the vac-
4.3. RF CONDITIONING

Uum, as shown in Fig. 4.8. With this phenomenon, called “Outgassing”, due to the RF power, the metallic surfaces are heated and gradually release the impurities. The ionic vacuum pumps, which are always turned on, remove the impurities so that the vacuum is restored. If a vacuum peak overcomes a threshold, of nearly $10^{-6}$ mbar, the RF power must be kept constant or even decreased, in order to restore the previous vacuum level. The conditioning runs until reaching the maximum level of 15 MW with 100 ns of pulse width. After that, the conditioning from 1 MW to 15 MW is performed again with larger RF pulses, increasing the pulse of nearly 200 ns at each step. The HERFD is considered as conditioned when it can be used with 15 MW of RF power, with 3000 ns of RF pulse width, and with the vacuum levels below the threshold of $5 \cdot 10^{-8}$ mbar. The conditioning has been performed with a pulse repetition rate of 10 Hz, reaching the power of 12 MW with the RF pulse of 700 ns in the first eight hours of operation of the structure. The maximum power of 15 MW with the pulse length of 3000 ns has been reached after 24 hours of conditioning.

During the conditioning process, RF discharges are also present. They are generated by the extraction of electrons from the metallic surface, which cause an arc, and are visible in the reflected RF power. Fig. 4.9 shows four oscilloscope signals, each proportional to the following power levels:

Figure 4.8: Vacuum levels, with outgassing peaks.
• Blue: Power generated by the Klystron.
• Yellow: Forward power, measured at the HERFD input coupler.
• Red: Backward power, measured at the HERFD input coupler.
• Green: Output power, measured at the HERFD output coupler.

Figure 4.9: RF discharge in the reflected power.

The oscilloscope is set with the “Trace persistence”, that records every sweep on the screen. With the persistence, the RF discharge on the reflected power is recorded, and is visualized as a peak, in which the trace deviates from its normal shape. The RF discharges are present even after the conditioning process, but with a lower frequency. The next section will analyze the frequency of the discharge events in the normal operation of the device.

4.4 Break down rate measurements

After 40 hours of operation, we started at fixed conditions (15 MW) to get breakdown rate (BDR), by observing the discharges on the backward RF power over about two hours of operation. This observation time has been chosen because it is the most common using time of the high energy deflector,
in order to make a slice emittance measurement. The BDR measurements are reported in Fig. 4.10, where the BDR is calculated as:

\[
BDR = \frac{\text{\# of RF discharges}}{\text{\# of total RF pulses}}
\] (4.1)

When the RF pulse width increases, the probability of RF discharge increases. Since the filling time of the HERFD is nearly 500 ns, the RF pulse has been chosen to be 900 ns. Such value ensures the complete filling of the structure, and the breakdown rate is kept low.

### 4.5 Electron beam measurements

The electron bunch length has been measured with the two deflectors, at first by using the vertical deflector, and then by the horizontal one. The bunch length values have been compared. Fig. 4.11 shows the beam vertical deflection, while Fig. 4.12 shows the beam horizontal deflection.

The measured bunch lengths RMS values are 0.44 ps by using the vertical deflector, and 0.43 ps by using the horizontal deflector. These measurements
are in good agreement. The deflectors are ready to be used in slice emittance and phase-space measurements [52].
Conclusions

In this thesis we showed three fundamental devices for the Free Electron Lasers (FEL): The cavity beam position monitor, the travelling wave linear accelerator and the high energy RF deflector.

The research on the cavity BPM focuses on a novel electromagnetic design of the structure and on a new signal processing, based on a simple system that avoids the use of IQ demodulation. The new electronic system is entirely original and has never been though before. The cavity BPM activity, described in this thesis, comprises electromagnetic design, simulation, optimization, and the design of the frequency tuners, making use of the softwares HFSS and CST. A first prototype of cavity BPM has been produced and measured, followed by the production of a series of 25 cavity BPMs for the FERMI@Elettra project. The cavity BPMs have been successfully installed and tuned in the undulator hall, and the data acquisition system has been implemented with the new type of electronic system. A friendly graphic user interface has been realized to perform the beam position measurements. The calibration and the evaluation of the cavity BPM performances have been showed. A series of improvements in the electronics chain are already known to be beneficial for the increase of the measurement accuracy. Upgrade activities comprise:

1. Adding low-pass filters before the ADC.
2. Insertion of a programmable attenuator in the reference signal to adjust the “weight” of the position signal at different beam charges.
3. Digitalization of the entire sum and difference measurement system.

The research on the linear accelerators has proposed a different approach to realize an input coupler that does not degrade the electron beam properties. Precisely, the alternative approach is based on a single feed coupler
Conclusions

with movable short-circuit on the opposite side. The coupler is geometrically symmetric, and the electric field can be symmetrized by suitably choosing the short circuit position. The racetrack concept has been extended to this kind of coupler, reducing the quadrupolar component as well. All the simulations have been performed with HFSS. The single feed with movable short-circuit coupler has been realized at the Elettra Synchrotron Trieste workshops. The thesis has shown the preliminary standing wave measurements performed on the cells. After that, the full linac has been assembled and the bead-pull measurements allowed to measure the longitudinal electric field which accelerates the electrons. With the same method, the transversal electric field magnitude has been measured on the single feed with movable short-circuit coupler and in the single feed coupler. The latter measurements showed that the electric field dipolar component, which is unwanted, is strongly reduced in the single feed with movable short-circuit coupler. A particle tracking software has been developed to show the transversal kick received by the electron beam, in six types of designed couplers. Future activities comprise the measurement of the quadrupolar component with the bead-pull method, the improvements of the linac tuning methods and the design of pulse compression system, which allows to have a bigger peak accelerating gradient, by using the same klystron.

The last activity reported in this thesis, describes the high energy deflectors. The RF parameters have been measured during the acceptance test of the two structures, the vertical one and the horizontal one, respectively. They have been successfully installed after the linac. During the RF conditioning the RF power and the RF pulse width were gradually incremented, until reaching the final designed value of RF power of 15 MW and the pulse length of 3000 ns. We reached such values without problems, in the first 24 hours of machine operation. The breakdown rate measurements have been reported. The two deflecting structures are ready to be used for the bunch length and the beam slice emittance measurements. Finally, the beam deflection has been illustrated, showing the deflection in the X and in the Y plane. Future activities are the study of an ‘X’ band deflector, which allows to have a bigger deflection with the same amount of input power.
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Publications on international journals


Conference proceedings


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