Design of the 15 GHz BPM test bench for the CLIC test facility to perform precise stretched-wire RF measurements

This content has been downloaded from IOPscience. Please scroll down to see the full text.

2015 Meas. Sci. Technol. 26 094005

(<http://iopscience.iop.org/0957-0233/26/9/094005>)

View [the table of contents for this issue](#), or go to the [journal homepage](#) for more

Download details:

IP Address: 137.138.254.22

This content was downloaded on 30/07/2015 at 15:05

Please note that [terms and conditions apply](#).

Design of the 15 GHz BPM test bench for the CLIC test facility to perform precise stretched-wire RF measurements

Silvia Zorzetti^{1,2}, Luca Fanucci², Natalia Galindo Muñoz¹ and Manfred Wendt¹

¹ CERN, 1217 Meyrin, Geneva, Switzerland

² Università di Pisa, 56126 Pisa PI, Italy

E-mail: silvia.zorzetti@cern.ch

Received 12 January 2015, revised 28 May 2015

Accepted for publication 10 June 2015

Published 29 July 2015

Abstract

The Compact Linear Collider (CLIC) requires a low emittance beam transport and preservation, thus a precise control of the beam orbit along up to 50 km of the accelerator components in the sub- μm regime is required. Within the PACMAN³ (Particle Accelerator Components Metrology and Alignment to the Nanometer Scale) PhD training action a study with the objective of pre-aligning the electrical centre of a 15 GHz cavity beam position monitor (BPM) to the magnetic centre of the main beam quadrupole is initiated. Of particular importance is the design of a specific test bench to study the stretched-wire setup for the CLIC Test Facility (CTF3) BPM, focusing on the aspects of microwave signal excitation, transmission and impedance-matching, as well as the mechanical setup and reproducibility of the measurement method.

Keywords: alignment, BPM, linear collider, CLIC, PACMAN, beam diagnostic, impedance transformer

(Some figures may appear in colour only in the online journal)

1. Introduction

A beam position monitor (BPM) is a diagnostic instrument used to measure the position of the beam with respect to the geometric centre of the beam pipe. From a beam dynamics point of view, the optimal beam trajectory with minimum emittance blow-up will be in the magnetic centre of each quadrupole, where non-linear field components are at a minimum. A series of BPMs distributed along the accelerator are needed to measure and control the beam orbit precisely along all the quadrupoles. One challenge lies in the pre-alignment of the BPM electrical centre to the quadrupole magnetic centre in the sub- μm regime, which is mandatory for a successful commissioning of a 50 km long accelerator with low emittance beams.

³ The PACMAN project is funded by the European Union's Seventh Framework Programme for research, technological development and demonstration under grant agreement no. 606839.

The strategy chosen for our alignment studies is based on two steps: the first step is to separately characterise the BPM and the quadrupole using stretched-wire measurement methods; with the second and final step we integrate both components into a dedicated standalone test bench using the same stretched wire.

Among the other components, the test stand includes: precision translation stages to mechanically scan the BPM cavity using the wire as a probe, seismic sensors and metrology equipment. The wire-based calibration of the BPM will define the electrical centre position associated with the electrical signal induced by the particle beam. Referencing the measurement of the magnetic centre of the quadrupole magnet to the BPM electrical centre, the quad-BPM system will be independent of external references. In this context particular attention has to be paid to the wire chosen, and to the impact that it may have on the BPM pickup.

At DESY for the TESLA Test Facility phase II (TTF2), similar measurements have been performed, achieving a

Figure 1. Schematic view of the stretched-wire alignment setup at DESY for TTF2.

Figure 2. CTF3 BPM test stand schematic view.

reproducible resolution of $10\ \mu\text{m}$ [1]. A schematic view of that stretched-wire test bench for the alignment is shown in figure 1, which can be used as reference for the future test bench in the frame of the PACMAN activities. However, the DESY BPM was a stripline BPM operating at 375 MHz, while the CTF3 BPM is a passive RF cavity operating at 15 GHz. Operating at microwave frequencies offers a higher resolution potential in the nm regime, and many aspects of the measurement technologies, RF electronics and precision mechanics are very challenging.

2. PACMAN BPM test stand

Within the frame of the PACMAN project and the BPM study, the first goal is a study of the methodology on a standalone RF test stand (figure 2), evaluating the cavity BPM with stretched-wire techniques to study signal excitations and measurements at microwave frequencies. With the help of metrology equipment available at CERN, i.e. a 3D Coordinate-Measuring Machine (Leitz Infinity CMM from Hexagon Metrology) with a minimum uncertainty of $300\ \text{nm}$ ⁴, we will reference the measured electrical centre to mechanical fiducials, and prove the reproducibility of the method.

⁴Leitz Infinity CMM Measuring uncertainty: $\pm(0.3\ \mu\text{m} + L/1000)$, where L is the dimension of the object under test.

The test bench, assembled on an active optical table (to damp vibrations) includes:

Hexapod to precisely move the BPM with respect to the wire in sub- μm resolution steps in 6 degrees of freedom (DOF). In fact, we will use only 4 DOF, the two axes perpendicular to the beam directions and the respective angles. The HXP100-MECA Hexapod by Newport allows us to move on the three linear axes with a resolution of $0.25\ \mu\text{m}$ on the vertical axis, and of $0.5\ \mu\text{m}$ on the horizontal ones. The error has been measured in the CMM, and it is on a single direction of about $0.5\ \mu\text{m}$, while the bi-directional repeatability is typically $4\ \mu\text{m}$.

Beam Pipes There are two types of beam pipes foreseen, two without bellows, directly connected to the flanges of the BPM, and another two, each including two bellows, to allow movement of the BPM setup with respect to the wire.

Adjustable Supports for BPM and beam pipes, fixed on the optical table.

Comparators and metrology instrumentation to measure the physical position of the BPM and other elements of the setup, required for the pre-alignment of the system.

Stretched-wire tools, various mechanics and electronics, such as stepper motors, are required to precisely stretch the wire in the centre of the cavity.

RF matching networks for the signal excitation of the wire and to reduce the reflections on the transmission line.

Figure 3. 15GHz BPM cavity resonator used for CLIC and CTF3 experiments.

Figure 4. Pillbox schematic view and dimensions.

3. The beam position monitor in CTF3

The BPM used for CLIC is a passive resonant cavity, operating at 14 GHz (figure 3), optimized for both, an excellent spatial resolution ($<50\text{nm}$), and a temporal resolution ($<50\text{ns}$) to resolve parts of the beam pulse [2]. For beam tests at the CLIC test facility (CTF3) the resonant frequency of this design was modified to 15 GHz, to comply with the beam structure.

The cavity BPM can be seen as a simple resonant pillbox, in which the following relation holds: $d < 2a$ ($d = 2\text{mm}$ is the length and $a = 11.24\text{mm}$ is the radius of the cavity, see figure 4). According to the geometric properties of the cavity, the fundamental modes excited are transverse modes (TM), two of them are of particular interest: the monopole mode TM_{010} at approximately 11 GHz, and the dipole mode TM_{110} at 15 GHz, which has two polarisations.

The beam-excited dipole mode is slot-coupled to rectangular waveguides, having a cut-off frequency between the TM_{010} and the TM_{110} resonant frequencies, to suppress the signal excited by the fundamental monopole mode. Thus the output signal, picked up by RF connectors attached to a coaxial-to-waveguide transition, depends only on the dipole mode, and it is proportional to the displacement of the beam to the cavity electrical centre.

In figures 5 and 6, plots of the electric field, with the TM_{010} and TM_{110} modes excited, are depicted. There is no coupling through the waveguides if the monopole mode (TM_{010}) is excited, while in presence of the dipole mode TM_{110} the signal is passed through the waveguides to the coaxial ports. The two different polarisations of the dipole mode (figure 6) are underlined, and, according to the polarisation, the respective orthogonal set of external waveguides is also excited.

For a resonant cavity BPM, based on beam excited dipole eigenmodes, the electrical centre is the physical

Figure 5. E-Field TM_{010} .

position where the electric field is zero. Because of manufacturing and material imperfections, the electrical centre may not match with the mechanical centre; for this reason the cavity needs to be calibrated with the dedicated BPM test bench.

4. Stretched-wire setup

We consider two methods to locate the electrical centre of the cavity BPM, both by means of a stretched, coaxial wire:

- **Signal excitation**

A 15 GHz RF signal is fed to the stretched wire and excites the TM_{110} dipole mode in a similar manner to the beam. With the wire in the electrical centre, no signal will be transferred between the wire and the relevant BPM waveguide ports, therefore a S21 measurement between those ports will be performed.

- **Perturbation analysis**

The dipole mode is excited using the attached waveguides, performing a S21 measurement between two opposite waveguide ports. Offsetting the stretched wire will perturb the modes, the perturbation is at a minimum when it is in the electrical centre, i.e. does not couple to the dipole mode.

Currently we consider the stretched-wire signal excitation as the primary method to detect the electrical centre of the BPM and to characterise it.

A stretched wire in a coaxial arrangement with the surrounding beam pipe yields in a transmission-line, and when excited with a RF signal it develops a TEM field pattern similar to that of a charged particle beam. Different from using bunched beams, we operate the test setup with sine waves (continuous wave—CW) signals, exactly at the TM_{110} mode frequency. Unfortunately, the characteristic impedance of the stretched-wire setup differs substantially from the standard impedance of RF and microwave measurement equipment $Z_0 = 50\Omega$. With the given cross-section dimensions of the coaxial wire setup, the characteristic impedance is in the order $Z_{\text{coax}} \approx 250\Omega$. It is necessary to match those impedances to minimise reflections, see the simplified block schematic in figure 7.

Figure 6. E-Field TM_{110} .

Figure 7. Stretched-wire simplified setup.

Figure 8. BPM transverse section. The wire is swept from the centre of the cavity (x-axis and y-axis intersection), along a 45-degree line.

BPM stretched-wire simulations A metallic wire was simulated inside the beam pipe and the cavity BPM, using the model in figure 8, where the parameter ‘p’ refers to the distance between the axis intersection and the position of the wire. The result of the simulation in *CST Studio* is depicted in figure 9 for a few samples, the cavity is excited from *port 1* and the signal is picked up from *port 3*. The graph in figure 10(a) shows the behaviour of the BPM output signal, at a fixed frequency $f_0 = 14.988$ GHz ≈ 15 GHz (dipole mode frequency), with respect to the radial position of the simulated wire (parameter ‘p’ in figure 8). The *quasi-linear* range is limited to $\sim \pm 0.3$ mm, as indicated in figure 10(b).

Figure 9. Result of the simulation in *CST Studio* of a metallic wire sweep in the BPM cavity for a limited number of samples.

The final setup, which will include the BPM and the quadrupole magnet rigidly mounted together, will have their electrical and magnetic centres measured with the same stretched wire, and, operating in the BPM linear region, the offset vector between those centres will be compensated as correction term in the BPM data post-processing.

BPM stretched-wire measurements Preliminary stretched-wire measurements have been realised on the BPM to study the working principle of the instrument under this operating conditions, and to compare the results with numerical simulations. The two traces in figure 11 compare the TM_{110} mode measurements, with and without the presence of the stretched wire. The dipole mode frequency is slightly shifted, but the wire does not alter it substantially, and some expected change in the Q-value was observed.

5. Impedance matching

To excite the stretched wire with a RF signal we use a standard signal source: a vector network analyser (VNA), which is based on a characteristic impedance of 50Ω. The

Figure 10. $|S_{31}|$ at dipole mode eigen-frequency in function of the radial position of a simulated metallic wire. (a) Sweep from 0 mm to 3 mm with reference to the centre of the cavity. (b) Zoom on the *quasi-linear region*, sweep from 0 mm to 0.5 mm with reference to the centre of the cavity.

stretched-wire setup uses a very thin ($d = 0.125$ mm) wire to limit perturbation the effects of the eigenmode, which results in a rather high characteristic impedance (250Ω) of the coaxial line setup. A direct connection between the VNA signal source and the wire setup without any matching network will cause unwanted signal reflections at the $50\text{--}250\Omega$ transition, and also at the end of the wire if no termination is foreseen. Following the transmission-line theory, this will result in standing waves along the lines, with difficulty in predicting and unstable signal conditions.

The reflection coefficient is a function between the load impedance (Z_L) and the characteristic impedance of the line (Z_0):

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0}. \quad (1)$$

It is linked to the Voltage Standing Wave Ratio (VSWR), which is defined as the ratio between the reflected wave V^- and the incident wave V^+ ; this parameter characterises the quality of the RF signal transmission. The reflections generated by impedance mismatches produce standing waves along the entire transmission line, in our case the wire setup.

$$\text{VSWR} = \frac{V^-}{V^+} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (2)$$

Figure 11. TM_{10} on one of the BPM pickups.

As a rule of thumb, reflections $\leq 10\%$, $|\Gamma| < 0.1$ are typically considered as acceptable in RF and microwave systems. This corresponds to a return loss of 20 dB, and results in $\text{VSWR} < 1.2$, based on equation (2).

The signal is launched through SMA connectors, with an external (reference) impedance of $Z_0 = 50\Omega$, to the coaxial line, with an equivalent impedance of $Z_L = 250\Omega$. Design

limitations are due to a reasonable minimum microstrip width and a minimum outer diameter for the coaxial line. The microstrip width is optimal for low impedance values, while the coaxial outer diameter better fits with high impedances. As matching from $Z_0 = 50\Omega$ to $Z_L = 250\Omega$ cannot be achieved using a single transformer technology, we choose an hybrid multi-section impedance transformer. An impedance value ($Z' = 125\Omega$), in between Z_0 and Z_L , was picked as technology transition, designing two transformers, passing from $Z_{01} = 50\Omega$ to $Z_{L1} = 125\Omega$ on the PCB, and from $Z_{02} = 125\Omega$ to $Z_{L2} = 250\Omega$ on the coaxial wire-line.

Out of a variety of impedance matching schemes such as resistive networks with absorbing material, exponential taper, *Klopfenstein* taper, quarter-wave transformers etc, we chose the latter for our tests, as best fit between reflection attenuation and design feasibility.

Popular designs are based on binomial (*Bessel*) or *Chebyshev* functions. Here a binomial transformer approach is preferred to achieve a flat response and a linear phase [3].

5.1. Quarter-wave transformer

Considering a section of lossless transmission-line of impedance Z_0 , the impedance seen at the input of the line computes to

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan \theta}{Z_0 + jZ_L \tan \theta}, \quad (3)$$

where Z_0 is the characteristic impedance of the line, Z_L is the load impedance, and $\theta = \beta l$ is the electrical length, with $\beta = 2\pi/\lambda$.

If $l = \lambda/4$, $\tan \theta = \tan \beta l = \tan \pi/2 \rightarrow \infty$, this quarter-wave long transmission-line section acts like an impedance transformer

$$Z_{in} = \frac{Z_0^2}{Z_L} \quad (4)$$

The impedance matching condition, zero reflections ($\Gamma = 0$), is given for a characteristic impedance of the quarter-wave line of:

$$Z_0 = \sqrt{Z_{in} Z_L} \quad (5)$$

Applying this technique will result in an optimal impedance matching at the desired frequency, in our case $f_0 = 15$ GHz, which is the eigen-frequency of the dipole mode TM_{110} . However, the bandwidth of this single section quarter-wave transformer is rather narrow, and is therefore sensitive to mechanical tolerances.

5.2. The multisection quarter-wave transformer

A multisection transformer consists of N quarter-wave lines, in which the characteristic impedance is gradually adapted to the load. This allows an impedance matching over a larger frequency band, and by varying the impedance of the steps, the matching response function can be optimised.

As a result of this N -section transformer the reflection coefficient is given by:

Figure 12. Reflection coefficient with and without approximations. Γ_T (equation (11)), Γ_{p1} (equation (13)) and Γ_{p2} (equation (14)).

$$\Gamma(\theta) = A(1 + e^{-2j\theta})^N = A \sum_{n=0}^N C_n^N e^{-2jn\theta}, \quad (6)$$

where

$$C_n^N = \frac{N!}{(N-n)!n!}, \quad (7)$$

is the binomial coefficient, and

$$A = 2^{-N} \frac{Z_L - Z_0}{Z_L + Z_0} \quad (8)$$

is a constant that can be calculated for $f \rightarrow 0 \Rightarrow \theta = \beta l \rightarrow 0$.

The magnitude of the equation (6) is given by:

$$|\Gamma(\theta)| = 2^N |A| |\cos \theta|^N. \quad (9)$$

For a specific line section n , the reflection coefficient is:

$$\Gamma_n = \frac{Z_{n+1} - Z_n}{Z_{n+1} + Z_n} \quad (10)$$

5.3. Impedance matching design

As the design choice it was decided to use two three-stage transformers ($N = 3$), the former on the PCB, and the latter on the coaxial wire-line.

The sum of the two filters does not represent a pure binomial filter, but it can be still studied using the theory of the small reflections.

$$\Gamma_T(\theta) = \Gamma_0 + \Gamma_1 e^{-2j\theta} + \Gamma_2 e^{-4j\theta} + \dots + \Gamma_N e^{-2jN\theta} \quad (11)$$

Where Γ_n is the reflection coefficient at the section n , as in equation (10).

To find a good approximation for the present transformer, in order to study some properties, such as the bandwidth, we first calculate separately the function of the two binomial transformers.

Figure 13. Ideal transformer block diagram.

Figure 14. Ideal transformer response.

For the first transformer, located on the PCB, we match $Z_{01} = 50\Omega$ to $Z_{L1} = 125\Omega$, for the second one, performed in coaxial technology, we continue with $Z_{02} = 125\Omega$, and go to $Z_{L2} = 250\Omega$.

$$\begin{aligned}\Gamma_1(\theta) &= A_1(1 + e^{-2j\theta})^N \\ \Gamma_2(\theta) &= A_2(1 + e^{-2j\theta})^N\end{aligned}\quad (12)$$

Where $N = 3$ and A_1 and A_2 are calculated as in equation (8).

It is possible to demonstrate that a good approximation of this filter is done by the following equation:

$$\Gamma_{p1}(\theta) = \Gamma_1(\theta) + e^{-j6\theta}\Gamma_2(\theta) \quad (13)$$

While it may be more simply approximated with the pure sum of the two three stages transformers only around the central frequency $f_0 = 15$ GHz.

$$\Gamma_{p2}(\theta) = \Gamma_1(\theta) + \Gamma_2(\theta) \quad (14)$$

$$|\Gamma_{p2}(\theta)| = |\Gamma_1 + \Gamma_2| = 2^N |A_1 + A_2| |\cos\theta|^N \quad (15)$$

The plot in figure 12 shows the comparison between the reflection coefficients of a pure six-stage binomial transformer (Γ_T , equation (11)) and the two approximations Γ_{p1} and Γ_{p2} , equations (13) and (14).

An important parameter to consider and to estimate the efficiency of the matching network for our purpose is the bandwidth. Using the approximation $\Gamma_{p2}(\theta)$ in equation (14), around f_0 , the bandwidth may be calculated as follows:

$$\frac{\Delta f}{f_0} = \frac{2(f_0 - f_m)}{f_0} = 2 - \frac{4}{\pi} \cos^{-1} \left[\frac{1}{2} \left(\frac{\Gamma_m}{|A_1 + A_2|} \right)^{\frac{1}{N}} \right] \quad (16)$$

where Γ_m is the acceptable reflection coefficient at the bandwidth limit, and f_m is the respective frequency.

For $\Gamma_m = 0.1$, the relative bandwidth is

$$\frac{\Delta f}{f_0} = 0.68 \quad (17)$$

The result found with this setup is $\Delta f \approx 10$ GHz, but it may be almost doubled if the reflection coefficient is fixed to $\Gamma_m = 0.2$; this result would not have been achieved with a pure 6th order binomial transformer.

This multi-section quarter-wave transformer approach ensures good matching conditions over a wide frequency range, or in other words, allows for fabrication tolerances without compromising good impedance matching conditions.

5.4. Calculation of the characteristic impedances and wavelengths

The following impedance values were calculated for the two matching networks.

PCB Termination Matching $Z_{01} = 50\Omega$, $Z_{L1} = 125\Omega$, in microstrip technology

$$\begin{aligned}\frac{Z_1}{Z_{01}} &= 1.113 \rightarrow Z_1 = 55.62\Omega \\ \frac{Z_2}{Z_{01}} &= 1.539 \rightarrow Z_2 = 76.96\Omega \\ \frac{Z_3}{Z_{01}} &= 2.128 \rightarrow Z_3 = 106.4\Omega.\end{aligned}\quad (18)$$

Coaxial Line Termination Matching $Z_{02} = 125\Omega$, $Z_{L2} = 250\Omega$

$$\begin{aligned}\frac{Z_4}{Z_{02}} &= 1.087 \rightarrow Z_4 = 135.88\Omega, \\ \frac{Z_5}{Z_{02}} &= 1.398 \rightarrow Z_5 = 174.75\Omega, \\ \frac{Z_6}{Z_{02}} &= 1.798 \rightarrow Z_6 = 224.75\Omega.\end{aligned}\quad (19)$$

Figure 15. Microstrip PCB and Coaxial-line transformer 3D Model.

Figure 16. Microstrip PCB and Coaxial-line transformer |S₁₁| dB Simulation.

Figure 17. CPWG-CPW-Coaxial transition.

Figure 14 shows the reflection coefficient versus frequency for our double three-stage transformer (block diagram in figure 13), using the impedance values calculated above.

5.5. Simulations and design

Applying well-known equations [4], the impedance values calculated in the previous section were translated into parameters for the microstrip PCB and the coaxial-line designs. The characteristic of the hybrid quarter-wave transformer was verified by 3D numerical electromagnetic simulations, evaluating the S-parameters.

The model in figure 15 includes the PCB and the coaxial line sections, the response is displayed in figure 16 and can be compared with the ideal response in figure 14.

The transition from the microstrip to the coaxial line is a critical point in our design, as it represents a source of radiation loss, causing additional reflections and uncontrolled

Figure 18. BPM with PCB Microstrip and Coaxial-line transformer, 3D Model.

impedances. Also, in the practical implementation the transition is challenging, as it is difficult to ensure the ground signal path along the discontinuity. Moreover, the design is complicated by the necessity of fixing the wire to the PCB.

Figure 19. BPM with PCB Microstrip and Coaxial-line transformer, $|S_{11}|$ dB simulation.

The final PCB is designed in CPWG (coplanar waveguide with ground) with a transition to CPW (coplanar waveguide) at the last stage to allow the wire soldering before the coaxial line [5].

The termination design, with the CPWG-CPW-Coaxial transition figure 17, was integrated with the entire BPM model figure 18. An aluminium plate was added to the PCB ground, with the double objective of making the PCB mechanically stiffer and improving the ground path.

The resonant dipole frequency f_0 for this BPM is around 15 GHz; the value simulated was 14.988 GHz while the measured one was 14.995 GHz, as can be evaluated, respectively, by figures 9 and 11. The final structure is not simply a pill box; lateral waveguides are added to cancel out the monopole frequency and to pick up the signal. These structures will cause additional reflections and the response will not be flat as for the simple transformer (simulation in figure 16). Nevertheless, the attenuation at f_0 (14.988 GHz for the simulated model) is less than -20 dB when the wire is off-set and couples the electrical field. In figure 19 the attenuation achieved can be appreciated for a wire offset of 0.3 mm with respect to the centre of the cavity.

6. Conclusions

Simulations, analytic calculations and preliminary measurements were performed on a 15 GHz cavity BPM to demonstrate the proof-of-principle of this instrument in a stretched-wire measurement setup.

The BPM test bench was fully designed, including the mechanical design, as well as the termination design. The first results and simulations give confidence that this method has good potential to perform very precise measurements.

References

- [1] Noelle D *et al* 2004 BPMs with precise alignment for TTF2 *AIP Conf. Proc.* **732** 166–173
- [2] Chritin N *et al* A high resolution cavity BPM for the CLIC test facility [arXiv:1207.4768](https://arxiv.org/abs/1207.4768)
- [3] Pozar D M 2004 *Microwave Engineering* (New York: Wiley)
- [4] Wadell B C 1991 *Transmission Line Design Handbook* (Norwood, MA: Artech House)
- [5] Holzman E L 2005 Microstrip transitions *Encyclopedia of RF and Microwave Engineering* (New York: Wiley)