

Impedance Optimization of In-Vacuum Undulator Components

Dissertation
submitted for the award of the title
“Doctor of Natural Sciences”
to the Faculty of Physics, Mathematics, and Computer Science
of the Johannes Gutenberg-University
in Mainz

Paul Volz
born in Berlin

Mainz, 20.02.2026





Impedance Optimization of In-Vacuum Undulator Components © 2026 by Paul Volz is licensed under CC BY 4.0. To view a copy of this license, visit <https://creativecommons.org/licenses/by/4.0/>

Date of doctoral viva: 24.07.2026

1. Reviewer: Prof. Dr. Atoosa Meseck
2. Reviewer: Prof. Dr. Kurt Aulenbacher

Contents

1	Introduction and Motivation	9
1.1	BESSY II	9
1.2	Undulators	10
1.3	In-Vacuum Undulators	12
1.4	CPMU17	13
1.5	Elliptical In-Vacuum Undulator IVUE32	15
2	Theory	19
2.1	Theory of Impedance in Particle Accelerators	19
2.1.1	Wakefields	20
2.1.2	Impedances	22
2.1.3	Loss and Kick Factor	23
2.1.4	Impedance of Parallel Plates	24
2.2	Theory of the Goubau Transmission Line	27
2.2.1	Characteristic Impedance	30
3	Simulations	33
3.1	CST Solver Settings	33
3.2	IVUE32 Taper Simulations	37
3.2.1	Design Philosophy	37
3.2.2	First Taper Design	38
3.2.3	Second Taper Design	41
3.2.4	Simplified Step	43
3.2.5	Double Finger Designs	46
3.2.6	Foil Tensioner Design	48
3.2.7	Compound Taper	49
3.3	Foil Slit Simulations	53
3.3.1	Modeling	53
3.3.2	Real BESSY Beam Parameters	54
3.3.3	Impedance of Sinusoidal Slit	55
3.3.4	Symmetric Slit Deviation	61
3.3.5	Power Loss	61
3.4	Goubau Line Test Stand	65
4	Measurement Setup	69
4.1	Goubau Line Test Stand	69
4.1.1	Network Analyzer	71

Contents

4.1.2	Wire Properties	73
4.1.3	Impedance Step Models	76
4.1.4	Foil Taper Models	77
4.1.5	Transmitter and Receiver	78
5	Measurement Results	81
5.1	Impedance Step Model	81
5.1.1	Time Gating and Reference Scaling	83
5.1.2	Field Extension Test	86
5.2	Foil Taper Models	89
6	conclusion	91
7	Appendix	93
7.1	Effect on TDR Measurement of Different Support Materials	93
7.1.1	Introduction	93
7.1.2	Measurement Devices	93
7.1.3	Calibration and Set Up	93
7.1.4	Measurement	94
7.1.5	Polymer Profile at different locations	98
7.1.6	Conclusion	101
7.2	Data Sheet for Test Stand Wire	102
7.3	Foil Slit Simulations - Additional Considerations	105
7.3.1	Modeling and Meshing	105
7.3.2	Straight Slit Simulations	107
7.4	IVUE32 Taper Simulations - Additional Considerations	111
7.4.1	Mesh Convergence Study	111
7.4.2	Gap and Cut-Off Frequency	115
7.5	Horn Improvement Studies	116
	Bibliography	119
8	Acknowledgments	123

Abstract

In this thesis the concept of an elliptical in-vacuum undulator following the APPLE-II design, IVUE32, is introduced together with the design challenges specific to the impedance contribution of the device. The theory of impedances in particle accelerators is discussed together with the concept of a Goubau transmission line test stand. Extensive wakefield simulations of IVUE32 components are presented, focusing on the taper section and the shielding foils. A Goubau transmission line test stand is presented together with measurement results of simplified models. In the scope of this thesis several different taper section designs were analyzed and improvements towards a reduced impedance contribution were made.

Zusammenfassung

In dieser Dissertation wird das Konzept eines elliptischen In-Vakuumundulators mit dem APPLE-II-Design, IVUE32, vorgestellt und die spezifischen Herausforderungen hinsichtlich des Impedanzbeitrags Undulators erläutert. Die Theorie der Impedanz in Teilchenbeschleunigern wird zusammen mit dem Konzept eines Goubau-Leitungsmessstandes diskutiert. Umfangreiche Wakefeldsimulationen von IVUE32-Komponenten werden präsentiert, wobei der Fokus auf der Tapersektion und den Magnetschutzfolien liegt. Ein Goubau-Leitungsmessstand wird vorgestellt, zusammen mit Messergebnissen vereinfachter Modelle. Im Rahmen dieser Arbeit wurden verschiedene Designs von Tapersektionen analysiert und Verbesserungen hinsichtlich des Impedanzbeitrags erzielt.

1 Introduction and Motivation

Synchrotron radiation facilities have evolved greatly over the past 60 years. The first experiments using this radiation were parasitic measurements at facilities built mostly for high-energy or nuclear physics in the 1960s and 1970s. The first purpose built synchrotron light sources were commissioned in the early 1980s with BESSY I being one example. With advances in solid state physics, the demand for synchrotron radiation grew dramatically. Synchrotron radiation became the most important tool for studying the structure of matter in various fields. Chemistry, biology, protein crystallography and various disciplines of physics drove the demand for this radiation. Today, increasingly advanced synchrotron radiation facilities are being built around the world to meet this demand from industry and academia. Existing facilities are upgraded, pushing ever higher photon fluxes and wider spectral ranges. Many of them introduce new in-vacuum undulators (IVUs) to achieve this goal. BESSY II is one of them.

In this chapter, the BESSY II light source and its in-vacuum devices are introduced and discussed, motivating the studies presented in this thesis. The second chapter discusses the theory of impedance in particle accelerators and the concept of a Goubau transmission line. In the third chapter wakefield simulations of various in-vacuum undulator components are presented together with simulations of the Goubau line test stand, which is described in detail in chapter 4. Chapter 5 presents measurement results from the test stand and compares them to simulations. Chapter 6 concludes the results of this thesis and gives an outlook, followed by the appendix.

1.1 BESSY II

BESSY II is a third generation synchrotron light source. It was commissioned in September of 1998 [1]. It consists of a booster synchrotron and a full energy storage ring with a circumference of 240 m. The storage ring circulates electrons at an energy of 1.7 GeV with a beam current of up to 300 mA. The double bend achromat lattice of BESSY II with its 32 dipole magnets, results in 16 straight sections. Of those straight sections, 13 are occupied by undulators. The remaining three straight sections are occupied by the 500 MHz accelerating cavities, the injection infrastructure, and a 7 T superconducting wavelength shifter [2].

Since October 2012, BESSY II runs in top up operation [3]. This means that electrons are constantly injected from the booster into the storage ring, keeping the beam current and therefore the intensity of the light output constant. This was achieved by introducing fast orbit feed back to BESSY II [4]. In 2018 CPMU17 [5]

1 Introduction and Motivation

was installed at BESSY II as its first in-vacuum Undulator (IVU). A second IVU is currently under development.

BESSY II offers a standard user operating mode [6] with the filling pattern shown in Fig. 1.1. Special operating modes like low- α for especially short light pulses are

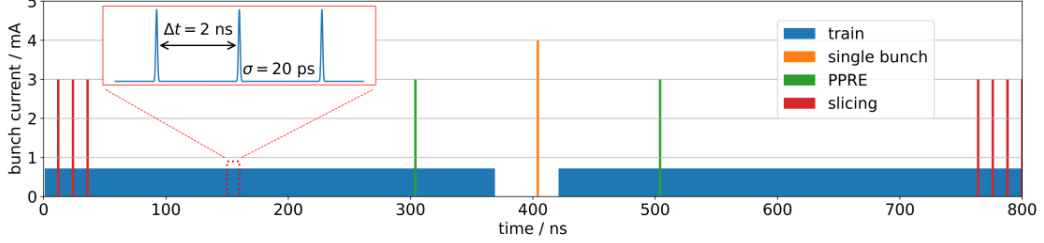


Figure 1.1: BESSY II standard user filling pattern taken from [6].

offered regularly. The beam parameters for the standard user filling pattern and the other specialized beams are shown in Tab. 1.1.

1.2 Undulators

As a third generation light source, BESSY II uses both dipole bending magnets and undulators as a source of radiation.

An undulator is a magnetic structure consisting of periodic dipole fields with alternating polarity. Relativistic electrons that pass through this magnetic field are forced onto an undulating trajectory, hence the name. This magnetic field constellation and the resulting radiation was first reported by H. Motz in May 1951 [7]. In July 1953 Motz *et al.* published their first experimental results for an undulator structure [8]. Their experiment used an electromagnet undulator with 25 dipole magnets. The period length of the field was 40 mm with a strength of 0.56 T. The experiment was driven by the 100 MeV Stanford Linear Accelerator.

As mentioned above, the periodic magnetic field of an undulator forces the relativistic electrons onto a sinusoidal or undulating trajectory. Each change in direction causes the emission of a cone of radiation similar to the case of a bending magnet. Depending on the viewing angle θ certain parts of the spectrum get amplified as the phase of each emission lines up to interfere constructively. The wavelength of the emitted light is described by [9]

$$\lambda = \frac{\lambda_u}{2\gamma^2} \left(1 + \frac{K^2}{2} + \gamma^2 \theta^2 \right) \quad (1.1)$$

where λ_u is the period length of the undulator, γ is the Lorentz factor of the accelerated electron, and K is the undulator strength parameter. The undulator strength parameter

$$K = \frac{eB_0\lambda_u}{2\pi mc} \quad (1.2)$$

Table 1.1: Overview of bunch parameter for different filling patterns taken from [6].

	Standard User (Top Up operation)	Low Alpha (decay operation)	
		Mode A	Mode B
Ring current (mA)	≤ 300	≤ 15	≤ 100
Multi-bunch-train	0.9 mA	45 μA	285 μA
bunch length (FWHM)*	45 ps	7 ps	16 ps
Single bunch	4.0 mA	≤ 43 μA	≤ 300 μA
Position	201	201	201
Single gap	56 ns	128 ns	128 ns
Purity in range +/-40 ns	> 1000	> 1000	> 1000
bunch length (FWHM)*	80 ps	7 ps	16 ps
Slicing bunches	7 x 3.0 mA	not offered	
Position	379, 385, 391, 397, 3, 9, 15		
bunch length (FWHM)*	80 ps		
PPRE bunch	3.0 mA	not offered	not offered
Position	151,251		
bunch length (FWHM)*	70 ps		
“low current” bunch	not available	≤ 30 μA	
Position		225	
bunch length (FWHM)*		7 ps	
Ring current Lifetime	> 5.0 h	> 20 h	> 10 h

* streak camera measurements

is determined by the magnetic flux B_0 at the trajectory of the electron, with e being the elemental charge, and m the mass of the electron.

The resulting photon wavelength λ can be understood as a Lorentz boosted and Doppler shifted undulator wavelength λ_u .

Compared to the radiation of bending magnets, undulator radiation has a very narrow bandwidth but much higher flux as shown in Fig. 1.2. Most experiments call for monochromatic radiation or very narrow bandwidth with the highest possible photon flux. Therefore specifically tuned undulators are much more desirable as radiation sources than bending magnets combined with monochromators.

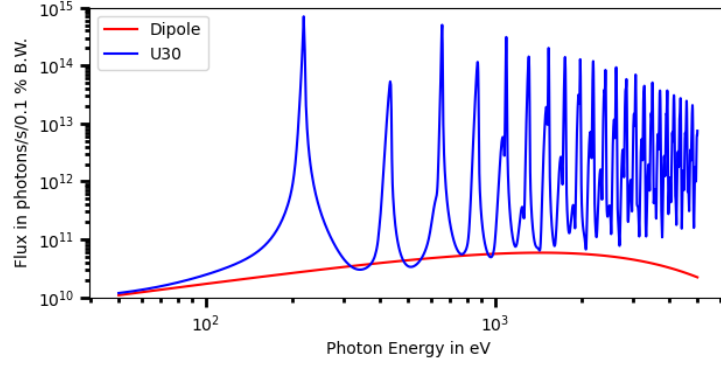


Figure 1.2: Spectrum comparison of undulator U30 in blue and a dipole in red for the BESSY II beam parameters, courtesy of Atoosa Meseck.

1.3 In-Vacuum Undulators

Most modern undulators utilize rare earth permanent magnets, as they can reach high flux densities of 1 to 1.5 T while maintaining very compact dimensions, allowing for smaller undulator periods. The magnetic flux density, however, falls off quickly with distance from the magnets. Accordingly, large undulator gaps result in reduced magnetic field strength at the electron trajectory. This is especially true for short undulator periods. Therefore, it is desirable to minimize the undulator gap as much as possible to achieve high K -values. The K -Value is also used for tuning the wavelength of the emitted undulator radiation. Looking at Eq. 1.1 most parameters are difficult to change during beam operation. The undulator period λ_u is defined during the design and construction of the insertion device. Changing γ would require changing the kinetic energy of the electrons in the storage ring, which would affect all other insertion devices and also the beam optics. The observation angle θ is defined by the beamline optics and can only be adjusted slightly, if at all. The K -value can be adjusted by changing the undulator gap, as that changes the on axis magnetic field. Changing the gap of an undulator can be done during beam operation and does not affect any other device in the storage ring, giving the user great flexibility for their experiments.

The larger the usable gap range of an undulator, the wider the spectral range of the produced radiation. Conventional undulators increase their tunability and spectral range by reducing the vertical dimension of the undulator vacuum chamber. This cannot be done indefinitely.

This limit gave rise to the idea of eliminating the vacuum chamber in the undulator gap and putting the whole magnetic structure into a much larger vacuum chamber. These so called in-vacuum undulators achieve gaps of only a few millimeters. This way, IVUs can provide photon energies and power output, unreachable for conventional in air undulators. Their large gap range allows the users to tune the IVUs over a very large spectral range.

Another approach to increase magnetic flux on the electron trajectory, is to use superconducting electromagnet undulators (SCU) [10]. SCUs tune their magnetic field by adjusting the electric current through the superconducting coils. These devices feature constant geometry vacuum chambers like conventional in air insertion devices and are therefore not the focus of this thesis.

Putting large complex structures that need to be movable with micrometer accuracy into a vacuum chamber, comes with some obvious mechanical challenges that require careful engineering and design.

There are however additional significant challenge that IVUs need to solve. For conventional undulators the vacuum chamber acts as a shield for the permanent magnets, protecting them from the electromagnetic fields of the relativistic electron bunches. IVUs rely on thin copper shielding foils that are stretched over the magnets. At high beam currents, the impedance of the structure can cause significant heating, concentrated at the conductive elements closest to the beam, like the shielding foils. This could lead to demagnetization of the undulator magnets or damage to the shielding foils in the worst case.

Furthermore, IVUs require complex taper sections to maintain continuity with the shielding foils at any gap setting. This makes the impedance considerations much more complex as an arbitrary number of vacuum chamber geometries need to be considered. Figure 1.3 shows the the taper section of CPMU17 as an example, which consists of two tensioned copper foils.

Investigating the impedance as cause for heat load on foils and electromagnetic resonances of IVUs is the subject of this thesis.

1.4 CPMU17

In 2018 the cryogenic permanent magnet in-vacuum undulator CPMU17 [5] with a period length of 17 mm was installed in the BESSY II storage ring. Figure 1.4 shows the magnet girders inside the vacuum chamber and the taper section at the entrance of the vacuum chamber, courtesy of the HZB undulator group photo archive [11].

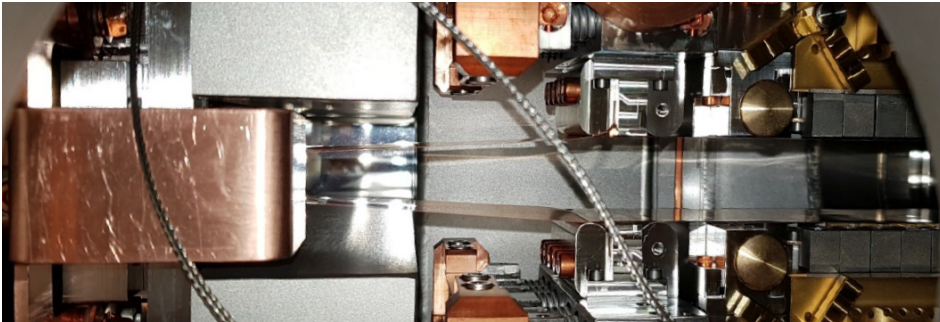


Figure 1.3: Photo of the foil taper section of CPMU17 [11].

1 Introduction and Motivation

During commissioning higher than expected heating of the taper sections was observed [5]. The observed heating is strongly dependent on beam current and bunch length. The downstream taper experiences a higher heat load than the upstream taper. Their geometry is not fully mirrored so geometric impedance seemed plausible as a source of the different heating.

It was later discovered that synchrotron radiation from an upstream dipole was the cause of the increased heat load of the downstream taper [12]. Additional cooling for the taper sections and a synchrotron radiation absorber were added in summer of 2019. This allowed CPMU17 to be commissioned to its design minimum gap of 7 mm for user operation.

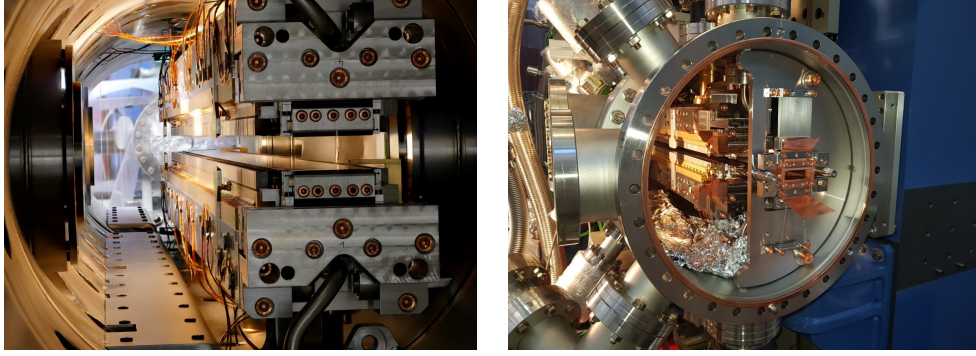


Figure 1.4: Photo of the magnet girders and shielding foil of CPMU17 on the left. The photo on the right shows the taper section installed and the vacuum chamber of CPMU17 [11].

The magnetic material used for CPMU17 was specifically developed in cooperation with Vacuumschmelze to operate at cryogenic temperatures [13]. Depending on the ferromagnetic material, reducing the temperature, increases the flux of the remanent magnetic field. Figure 1.5 shows the remanent magnetization and magnetic coercivity of the magnetic compound used in CPMU17 [13] as a function of temperature. The left graph shows that the flux of the remanent magnetization increases as the temperature decreases. The CPMU17 magnet sample shows a flux of 1.42 T at room temperature compared to 1.67 T when cooled with liquid nitrogen. This marks a significant increase in flux for a reasonable increase of cooling infrastructure complexity.

The graph on the right shows the coercivity of the material as a function of temperature. As the temperature decreases, the coercivity increases, meaning the material becomes more resistant to demagnetization.

Another upside of cryogenically cooling the permanent magnets is making the shielding foils more resilient to impedance induced heating. The thin copper foils are coated with nickel on one side, making them ferromagnetic. This keeps them in contact with the cooled magnets, which act as a heat sink.

For the next IVU currently under development at HZB, IVUE32, the complex mag-

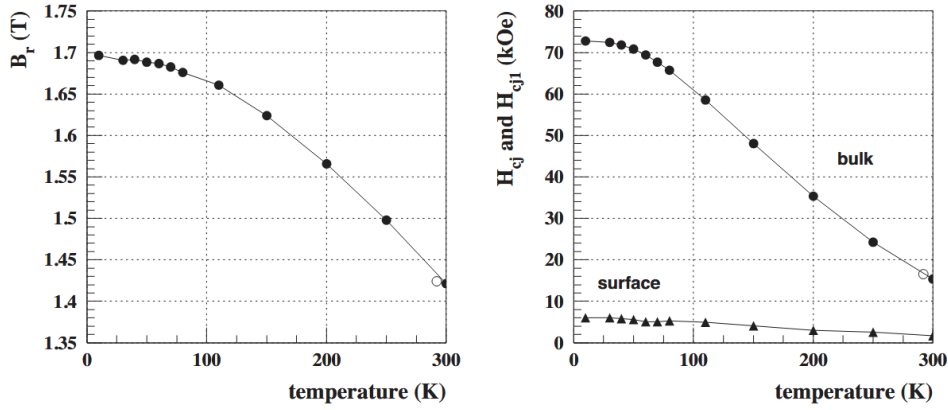


Figure 1.5: Remanence (left) and coercivity (right) of the magnetic material used in CPMU17 as a function of temperature. At cryogenic temperatures the magnetic flux increases significantly as well as the resistance to demagnetization. Taken from [13].

net keeper geometry does not leave a lot of room for cooling infrastructure. Controlling the impedance induced heating is therefore paramount for the longevity of the shielding foils and magnets.

1.5 Elliptical In-Vacuum Undulator IVUE32

IVUE32 is an in-vacuum elliptical undulator following the APPLE-II design with a period length of 32 mm. APPLE stands for **A**dvanced **P**lanar **P**olarized **L**ight **E**mitter. An APPLE undulator has four magnet girders, splitting the top and bottom girders in the middle along the axis of the beam. This allows for the magnet arrays to be moved relative to each other in the direction of the beam axis. With that phasing movement of the magnets, the polarization of the emitted synchrotron radiation can be changed. The polarization can be arbitrarily chosen anywhere between planar and circular. This added axis of movement requires a more complex support infrastructure for the magnet girders. More driving motors are required and more magnetic forces need to be compensated. Either by heavy and sturdy mounts or by complex magnetic force compensation schemes.

Bizen *et al.* presented an exotic concept of a revolver IVU that was installed in the storage ring at the Pohang Accelerator Laboratory in 2003 [14]. The device features revolving magnet girders that hold permanent magnets for four different undulators that can be selected by rotating the girders by 90° . This design only permits supports at either end of the magnet girders resulting in a very long support span of 1.12 m. This set up requires compensation of the attractive magnet forces between the girders as the long span would otherwise flex significantly. Secondary rows of permanent magnets are located at either side of the undulator function magnets. They are oriented in a Halbach array just like the function magnets but with their

1 Introduction and Motivation

polarity reversed. This is the first instance where permanent magnets are used as force compensation.

Since then other IVUs have been developed that feature magnetic force compensation schemes using permanent magnets. All of them however only consider one axis of the magnetic force. This is sufficient for planar undulators but the phasing movement of the APPLE-II design creates significant magnetic forces on all three axes that need to be compensated.

Heavy spring systems used for in-air APPLE undulators are very difficult to fit into the vacuum chamber of an IVU. Enlarging the support columns of the magnet girders to overcome the forces with structural rigidity is difficult because of the long levers involved. Thick support columns also act as thermal bridges which is undesirable for possible cryogenic IVU variants.

IVUE32 uses a complex force compensation scheme to compensate the magnetic force of the function magnets in all three directions and for all shift states [15].

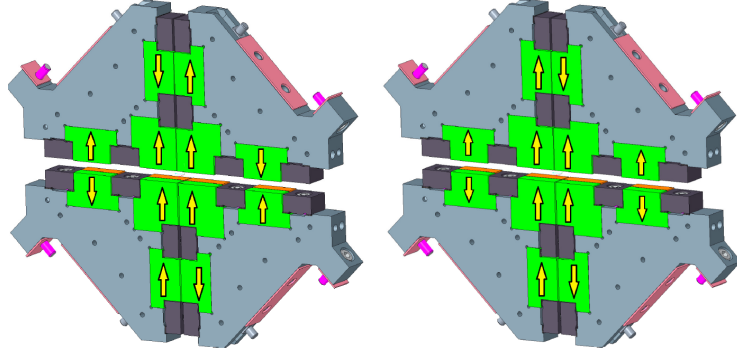


Figure 1.6: Magnet keepers of IVUE32 with undulator function magnets in the middle and smaller force compensation magnets on the perimeter. Taken from [15].

As Ryota Kinjo and Takashi Tanaka note in [16], controlling the three dimensional magnetic forces during the phasing operation is just the first of two major challenges that need to be overcome in realizing an in-vacuum APPLE-II undulator.

The second challenge, according to them, is the longitudinal movement itself, disqualifying established foil taper designs of other IVUs.

The initial shielding foil designs for IVUE32, presented at IPAC 2025 [17], feature a longitudinal slit along the magnet array to allow for the phasing movement. Shift states other than the neutral position would cause a transversal gap asymmetry in the taper sections. This will be mitigated by dividing the taper into two sections. The first section will compensate the gap movement with a solid copper foil taper similar to that of CPMU17. The second section will only compensate the longitudinal phasing movement. This is where the central slit in the shielding foils is introduced. A CAD model of the taper section is shown in Fig. 1.7 depicting the foil tensioning principle for a smooth transition. The shift compensation allows the magnet girder to move longitudinally for 16 mm while keeping the position of

the foil attachment points of the gap taper constant. This prevents vertical asymmetries in the gap taper gradient. Wakefield simulations for this taper model are presented in Chapter 3.

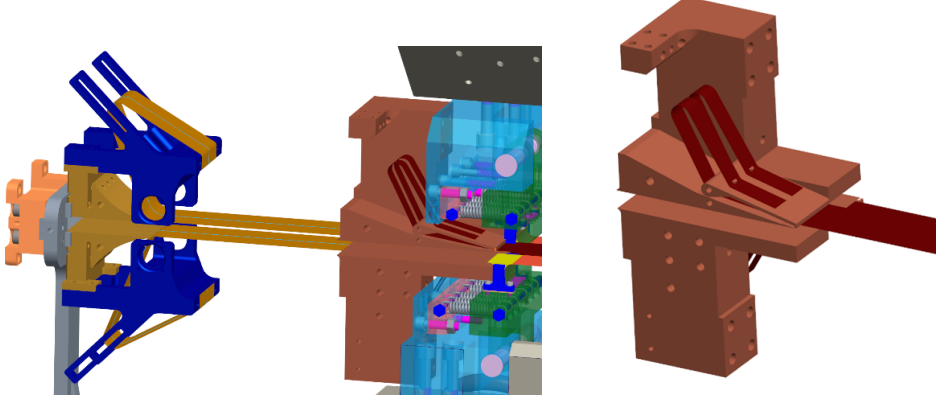


Figure 1.7: CAD model of the IVUE32 taper section. Full taper with gap and shift compensation on the left. On the right the foil tensioning principle of the shift taper is shown. Taken from [17]

Since then a possible constraint due to the β -function in the designated straight section has come up, limiting the minimum gap of the taper section. Introducing a compound gap and shift taper would break the transverse symmetry at shift positions other than neutral. Wakefield simulations of such a taper section are presented in Chapter 3.

The successor to BESSY II, BESSY III is in its conceptual design phase [18]. With a higher electron energy of 2.5 GeV and the goal to supply a broad spectrum of photon energies ranging from 100 eV to 10 keV, BESSY III will feature many IVUs of different designs [19]. Therefore, the impedance studies of IVUE32 presented in this thesis will be invaluable for future projects at HZB.

2 Theory

In this chapter we will discuss the interaction of a charged particle beam with its surrounding vacuum chamber components by introducing the concept of wake fields and impedance, followed by theoretical considerations of the electric fields surrounding a Goubau transmission line and its potential to simulate a charged particle beam.

2.1 Theory of Impedance in Particle Accelerators

When a charged particle travels through a conductive structure like a stainless steel vacuum chamber, its electric field induces mirror charges in the surrounding structure. When the cross section of the conductive structure changes, the distance between the charged particle and the induced mirror charges changes as well, affecting the electromagnetic (EM) fields between them. Figure 2.1 shows the electric field lines reacting to a change in cross section as a charged particle bunch travels through a tapered section of a beam pipe [20].

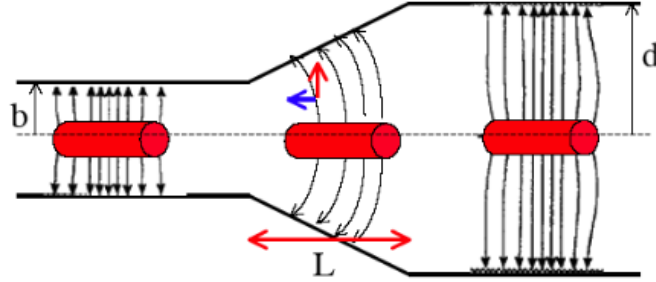


Figure 2.1: Wakefield lines of a charged particle bunch moving through vacuum pipe with a tapered section, taken from [20]

Depending on the geometry of the vacuum vessel there is a cut-off frequency above which EM fields can propagate and exit the section, below that frequency however, fields can excite resonant modes in the structure, effectively trapping the field energy. In a storage ring the time between two subsequent particle bunches passing a certain point of the vacuum vessel is in the order of a few nanoseconds, giving trapped modes very limited time to decay. The interaction of the induced EM fields with subsequent bunches described by the so called *coupling impedance*. It is usually separated into transverse and longitudinal components. The longitudinal impedance can cause electron bunches to lose energy, increasing the load on

2 Theory

the accelerating cavity. The power loss can also lead to heating of vacuum chamber components, stressing the cooling systems, increasing vacuum pressure or even damaging sensitive components close to the beam like permanent magnets of IVUs. The transverse impedance can introduce a vertical or horizontal kick. This can lead to beam steering, defocusing or in extreme cases beam loss. These effects are proportional to the bunch charge and might have a current threshold above which beam operations are significantly disturbed. Structures like IVUs are especially critical as their geometry changes with the gap settings during ID operation. The permanent magnets of IVUs are usually covered by thin copper shielding foils. Due to the nature of ultra-relativistic particle beams, the excited EM fields have high frequencies and therefore very low skin depths. This means that only the conductive surfaces are important when investigating the interaction with a charged particle beam.

One way to measure the impedance of IVUs is to use an accelerator's bunch-by-bunch feedback system and excite the structure using the beam. Such grow-damp and drive-damp measurements have been conducted at BESSY II by Huck *et al.* [12] for CPMU17. No changes in coupled bunch damping could be observed when changing the gap of the IVU. CPMU17 however still shows changes in temperature dependent on the gap setting. This hints at wakefield effects undiscovered by the previously described measurements.

EM fields can also be simulated to determine the impedance of a structure like IVUs. Their vacuum chambers are usually several meters long and filled with complex infrastructure to support operation. Accurate simulations at high frequencies however require a high mesh resolution of the structure, making the computation prohibitively resource intensive for such large structures. There are recent computational methods like domain decomposition that can split the problem into more manageable parts. This method was demonstrated by Quetscher *et al.* in 2023 for a planned IVU at the PETRA IV light source [21, 22]. As part of the collaborative research project WAKE [23] CPMU17 will be simulated and investigated to see if the initial measurements have missed potential trapped modes and if the gap dependent heating can be explained.

In this thesis we will only focus on the power loss aspect of impedance, possible trapped modes of structures are not investigated.

The Formulas and relations presented in the following sections regarding the interaction of charged particle beams with surrounding structures are based on the work of Zotter *et al.* [24].

2.1.1 Wakefields

The EM fields of a moving charge q_1 interact with the surrounding vacuum chamber and can influence subsequent charges. These fields interacting with a test charge q_2 a distance s behind the leading charge q_1 are called the wakefields of the leading charge, as the test charge is figuratively riding the wake of the leading charge. This relation is shown in Fig. 2.2 for a charged particle bunch traversing a structure

with a discontinuity. If the leading charge is a point charge the wakefields are

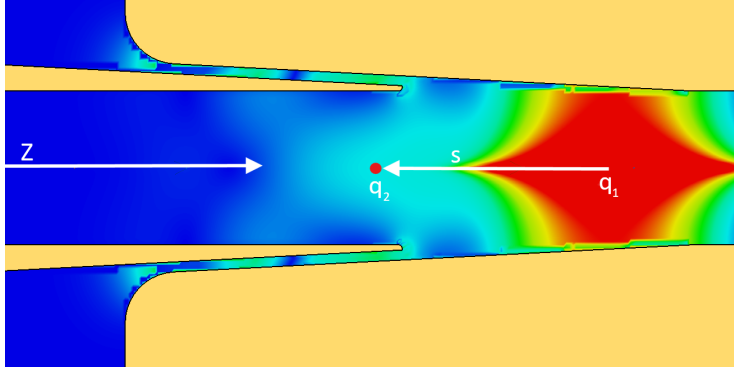


Figure 2.2: A charged particle bunch q_1 traversing a structure with a discontinuity, inducing EM fields that interact with a test charge q_2 a distance s behind q_1 .

referred to as a wake function. The integral of the EM forces acting on the test charge determines the wake function. The trajectories of both the exciting and test charges are assumed to be parallel to the axis of the vacuum chamber. The longitudinal wake function $G_{\parallel}$ only requires the integral of the longitudinal electric field component. The magnetic field information is not needed as magnetic forces only act perpendicular to the particle trajectory [24].

$$G_{\parallel}(\vec{r}_{q_1}, \vec{r}_{q_2}, s) = -\frac{1}{q_1} \int_{-\infty}^{\infty} dz E_z \left(\vec{r}_{q_1}, \vec{r}_{q_2}, z, t = \frac{z+s}{c} \right) \quad (2.1)$$

The vectors $\vec{r}_{q_1}$ and $\vec{r}_{q_2}$ denote the transverse offset of the exciting and test charge respectively. The wake function is normalized by the exciting point charge q_1 . The unit of the wake function is therefore V/C. The minus sign is a convention so that the wake function has positive values for decelerating field components. It is assumed that both the leading and the test charge travel at close to the speed of light so to maintain causality the electric field is evaluated at $t = (z+s)/c$. The practical definition of the longitudinal wake function is the change in energy ΔE experienced by the test charge q_2 trailing behind a point charge q_1 at a distance s

$$\Delta E = -q_1 q_2 G_{\parallel}(s). \quad (2.2)$$

Analogous to the longitudinal wake function, the transverse wake function is defined as the integral over the transverse components of the EM fields.

$$G_{\perp}(\vec{r}_{q_1}, \vec{r}_{q_2}, s) = -\frac{1}{q_1} \int_{-\infty}^{\infty} dz \left(\vec{E} + \vec{v}_{q_2} \times \vec{B} \left(\vec{r}_{q_1}, \vec{r}_{q_2}, z, t = \frac{z+s}{c} \right) \right)_{\perp} \quad (2.3)$$

Here the magnetic field needs to be evaluated. The transverse wake function is a vector with two components, most commonly defined as horizontal and vertical.

2 Theory

Instead of energy changes, the transverse wake function describes steering or focusing effects on a trailing particle bunch.

In the limit $v \approx c$ wake functions are only dependent on the geometry of the surrounding conductive structure. If the exciting charge is a point charge, the wake functions can be considered a Green's function for the geometry. In practice accelerators operate with charge bunches with a spatial distribution and not point charges. The distance $s = v_{q_2} \tau$ between the test charge and the leading charge is now considered to the center of the leading bunch charge. In the ultra-relativistic limit there are no wake fields ahead of the exciting charge and the wake function vanishes. To calculate the wake effects of these bunches, the wake function is convoluted with the charge distribution $\lambda(t)$ to yield the wake potential [24]

$$W_{\parallel}(\vec{r}_{q_1}, \vec{r}_{q_2}, s = v_{q_2} \tau) = \int_{-\infty}^{\tau} dt G_{\parallel}(\tau - t) \lambda(t) \quad (2.4)$$

for the longitudinal case and

$$W_{\perp}(\vec{r}_{q_1}, \vec{r}_{q_2}, s = v_{q_2} \tau) = \frac{1}{\xi} \int_{-\infty}^{\tau} dt G_{\perp}(\tau - t) \lambda(t) \quad (2.5)$$

for the transverse wake potential. Here ξ is the transverse offset of the exciting charge distribution. The upper limit τ of the time integral ensures causality.

2.1.2 Impedances

As wake potentials describe the EM forces in the time domain, impedance describes the EM forces in the frequency domain. Investigating the impedance of a structure is useful because several material properties can be frequency dependent.

Shorter bunches excite EM fields at higher frequencies. The frequency range of an electron bunch can be approximated by $\omega \approx 1/\sigma_t$ with σ_t being the standard deviation of a Gaussian bunch length in the time domain. The vacuum chamber acts as a waveguide and has a cut-off frequency dependent on its cross section. EM fields above that frequency can propagate, fields below that frequency can excite resonances called trapped modes that decay over time. In storage rings these trapped modes need to be investigated. Trapped modes that persist longer than the time between subsequent bunches can cause instabilities or even beam loss. The propagating high frequency components of wakefields are less concerning for beam stability unless they are reflected at other parts of the vacuum chamber and return at the correct phase to interact with subsequent bunches.

For the longitudinal case the impedance is simply the Fourier transform of the wake function [24]

$$Z_{\parallel}(\omega) = \int_{-\infty}^{\infty} d\tau G_{\parallel}(\tau) e^{-i\omega\tau}. \quad (2.6)$$

The transverse impedance is a complex function where the real (resistive) part

causes changes in amplitude of the betatron oscillation while the imaginary (reactive) part can cause changes in frequency of the betatron oscillation. For transverse motion both the magnetic and electric field components act on the beam and need to be considered. For vacuum chambers with transverse symmetry the transverse impedance disappears on axis. It is therefore usually defined for a transverse offset $\vec{r}_{q1}$ of the exciting charge [24]

$$Z_{\perp}(\vec{r}_{q1}, \omega) = \frac{i}{q_1 r_{q1}} \int_{-\infty}^{\infty} dz \left[\tilde{E}(\vec{r}_{q1}, z) + \vec{v}_{q2} \times \tilde{B}(\vec{r}_{q1}, z) \right]_{\perp} e^{\frac{i\omega z}{v}} \quad (2.7)$$

with $\sim$ denoting the Fourier transform of the field components. As with the wake function the transverse impedance has two components, usually divided into vertical

$$Z_y(\omega) = \frac{i}{y_{q1} \tilde{I}} \int_{-\infty}^{\infty} dz \left[\tilde{E}_y(z) + v_{q2,z} \tilde{B}_x(z) \right] \Big|_{x=0, y=y_{q1}} e^{ikz} \quad (2.8)$$

and horizontal

$$Z_x(\omega) = \frac{i}{x_{q1} \tilde{I}} \int_{-\infty}^{\infty} dz \left[\tilde{E}_x(z) - v_{q2,z} \tilde{B}_y(z) \right] \Big|_{x=x_{q1}, y=0} e^{ikz} \quad (2.9)$$

impedance.

2.1.3 Loss and Kick Factor

The loss factor is introduced to simplify the calculation of energy loss a charged particle bunch experiences when traveling through a vacuum chamber. The change in energy depends on the bunch charge q itself and the Voltage $V_b = -Z_{\parallel} I_b$ induced in the surrounding structure by said charge. The current $I_b = q\omega_0/2\pi$ is in turn proportional to the charge q with ω_0 denoting the revolution frequency for circular accelerators or repetition rate in case of linear accelerators. The resulting change in energy

$$\Delta E = -\kappa_{\parallel} q^2 \quad (2.10)$$

is proportional to the square of the charge with $\kappa_{\parallel}$ denoting the positive defined longitudinal loss factor [24]

$$\kappa_{\parallel}(\sigma) = \frac{1}{\pi} \int_0^{\infty} d\omega \Re \left\{ Z_{\parallel}(\omega) \right\} h(\omega, \sigma). \quad (2.11)$$

The loss factor can be calculated in the frequency domain as the convolution of the real part of the longitudinal impedance and the power density distribution $h(\omega, \sigma) = \tilde{\lambda}(\omega) \tilde{\lambda}^*(\omega)$, where $\tilde{\lambda}$ denotes the bunch spectrum meaning the charge distribution in the frequency domain. It can also be calculated from the wake potential in the time domain as

$$\kappa_{\parallel}(\sigma) = \int_{-\infty}^{\infty} d\tau W_{\parallel}(\tau) \lambda(\tau) \quad (2.12)$$

2 Theory

with λ being the line charge distribution.

In the transverse direction the physical action of the EM fields on the bunch is better described as a kick in trajectory rather than an energy change. The transverse loss factor is therefore often referred to as kick factor. It can be obtained in the frequency domain

$$\kappa_{\perp}(\sigma) = \frac{1}{\pi} \int_0^{\infty} d\omega \Im \{ Z_{\perp}(\omega) \} h(\omega, \sigma). \quad (2.13)$$

by convolution of the imaginary part of the transverse impedance with the power density distribution and in the time domain

$$\kappa_{\perp}(\sigma) = \int_{-\infty}^{\infty} d\tau W_{\perp}(\tau) \lambda(\tau) \quad (2.14)$$

from the transverse wake potential and the line charge distribution.

2.1.4 Impedance of Parallel Plates

The shielding foils stretched over the magnet assemblies of an IVU are very similar in structure to two parallel plates. The impedance of a charged particle bunch traveling between two parallel plates can be calculated analytically. Piwinski calculates the wakefields and Ohmic losses of flat vacuum chambers [25] analytically. The Ohmic losses for two parallel plates are

$$P_w = \frac{\Gamma\left(\frac{3}{4}\right) r_e E_0 N_p^2 c}{\pi h_1 \sigma_s^{3/2} \sqrt{2\mu_r Z_0 \sigma}} f_1(h, h_0, h_1) \quad (2.15)$$

$$f_1(h, h_0, h_1) = \frac{h_1}{h} \left(1 + \frac{\pi h_0}{2h} \tan\left(\frac{\pi h_0}{2h}\right) \right) \quad (2.16)$$

with $\Gamma\left(\frac{3}{4}\right) \approx 1.225$, r_e the classical electron radius, E_0 the particle rest mass, N_p the number of charged particles in the bunch, c the speed of light, σ_s the standard deviation of the Gaussian bunch length, μ_r and σ the relative permeability and conductivity of the plates, Z_0 the impedance of free space, h_0 the vertical offset of the beam, $\pm h$ the vertical position of the top and bottom plate, and $h_{1,2} = h \mp h_0$ the vertical distance of the beam to the top and bottom plate respectively. Piwinski also gives a formula for the wakefield force of a particle bunch traveling between two parallel plates

$$F_s = \frac{r_e E_0 N_p}{2h_1 \sqrt{2Z_0 \mu_r \sigma}} \left| \frac{u}{\sigma_s} \right|^{3/2} f_1(h, h_0, h_1) \left(I_{1/4} - I_{-3/4} \pm I_{3/4} \mp I_{-1/4} \right) e^{-u^2/4} \quad (2.17)$$

$$u = \frac{s - vt}{\sigma_s} \quad (2.18)$$

where I_{ν} are the modified Bessel Functions of the first kind with $u^2/4$ as the argument. For a particle bunch traveling in the middle of two parallel plates the wakefield

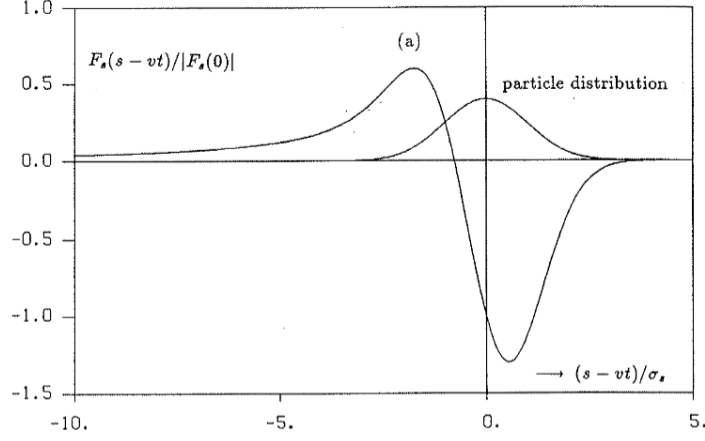


Figure 2.3: Longitudinal wakefield forces F_s induced by the depicted particle distribution. Taken from [25]

force yields a well known curve progression shown in Fig 2.3.

There are also analytical descriptions for the impedance of tapered structures of certain cross sections. The shielding foil geometry of IVUE32 and other IVUs in general can be compared to a tapered structure with rectangular cross section. An analytical description of the impedance of rectangular tapered transitions has been presented by Stupakov [26]. He presents an approach using perturbation theory to approximate the electric field needed to calculate the Green's function in Eq. 2.1. Equation 2.6 then yields the longitudinal impedance. The perturbation theory approach splits the longitudinal impedance into two components [26]

$$Z_{\parallel} = Z_{\parallel}^{(1)} + Z_{\parallel}^{(3)}. \quad (2.19)$$

$Z_{\parallel}^{(1)}$ is a frequency independent impedance base value that only depends on the cross sections at the entrance and exit of the tapered structure [26]

$$Z_{\parallel}^{(1)}(\vec{r}_{q1}, \vec{r}_{q2}) = \frac{1}{c} [\phi_0(s \rightarrow \infty) - \phi_0(s \rightarrow -\infty)]. \quad (2.20)$$

Here ϕ_0 is a Green's function for a rectangular cross section that satisfies [26]

$$\nabla_{\perp}^2 \phi_0 = -4\pi\delta(\vec{r}_{q2} - \vec{r}_{q1}). \quad (2.21)$$

If the entrance and exit cross sections of the tapered structure are identical, $Z_{\parallel}^{(1)}$ disappears. For mismatched cross sections, the longitudinal impedance does not vanish in the low frequency limit and will be offset by $Z_{\parallel}^{(1)}$ over the entire frequency range. This will be referenced in chapter 3 as some simulations only model the entrance of IVUE32. The final design of IVUE32 will have symmetric entrance and exit cross sections at any gap and shift setting.

2 Theory

Stupakov assumes small frequencies and therefore neglects terms of $\mathcal{O}(\omega^2)$. $Z_{\parallel}^{(3)}$ is included because it is the first term that does not vanish if entrance and exit cross sections are identical.

2.2 Theory of the Goubau Transmission Line

The Goubau line was developed by Georg Goubau in 1950 as a long range communication transmission line with lower attenuation than coaxial transmission lines [27]. The main difference to conventional wave guides is the single conductor. Surface waves are guided on the surface of a wire. This phenomenon was discussed by Sommerfeld in 1899 [28]. Goubau discovered that modifying the surface of the conductor greatly limits the field expansion around the conductor, making practical applications feasible. The main parts making up a Goubau line are a transmitter, a receiver and a dielectrically coated wire. Horn antennas are often used as transmitter and receiver as shown in Fig. 2.4. The transmitted waves are guided as

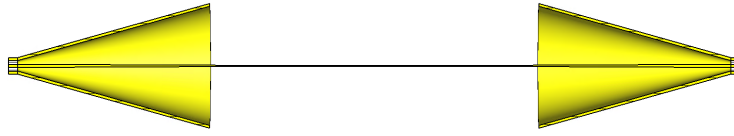


Figure 2.4: Schematic of a Goubau line consisting of two conical horn antennas and a dielectrically coated wire similar to the design at Argonne APS [29].

transverse magnetic modes on the coated wire. Figure 2.5 shows the orientation of electric and magnetic fields along the coated wire. The radial electric E_r , longitudinal electric E_z and magnetic H_ϕ fields are described by cylindrical functions. The corresponding formulas are derived in Goubau's original paper [27] and in a

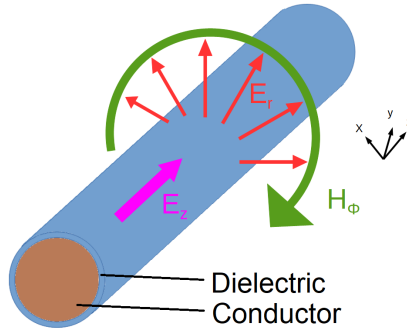


Figure 2.5: Schematic of field orientation of a Goubau line. The transverse electric field E_r mimics the field generated by a charged particle beam.

2 Theory

revision considering modern computational advances by B. Vaughn *et al.* [30].

$$E_r = iA_i \frac{h}{\gamma_i} Z_1(\gamma_i r) e^{i(\omega t - hz)} \quad (2.22)$$

$$E_z = A_i Z_0(\gamma_i r) e^{i(\omega t - hz)} \quad (2.23)$$

$$H_\phi = iA_i \frac{k_i^2}{\omega \mu \gamma_i} Z_1(\gamma_i r) e^{i(\omega t - hz)} \quad (2.24)$$

with

$$\gamma_i^2 = k_i^2 - h^2 \quad (2.25)$$

where A_i is a complex field amplitude, Z_n is the n -th order cylinder function, k_i is the total wavenumber, γ_i is the radial wavenumber, and h is the longitudinal propagation constant. The subscript $i = c, d, o$ denotes the region of the field. The three regions are the conductor (c), the dielectric coating (d), and the region outside of the wire (o) assumed to be free space. We assume a perfect conductor so any fields are immediately compensated by current flow

$$Z_{0,c} = Z_{1,c} = 0. \quad (2.26)$$

Inside of the dielectric the fields are described by Bessel functions of the first and second kind [27]

$$Z_{0,d} = J_0 + mY_0 \quad (2.27)$$

$$Z_{1,d} = J_1 + mY_1. \quad (2.28)$$

The parameter m is determined by the boundary condition between the conductor and the dielectric layer [27]

$$m = -\frac{J_0(\gamma_d r_c)}{Y_0(\gamma_d r_c)} \quad (2.29)$$

with r_c as the radius of the conductor. The field outside of the wire is described by Hankel functions

$$Z_{0,o} = H_0^{(1)} \quad (2.30)$$

$$Z_{1,o} = H_1^{(1)} \quad (2.31)$$

as they correctly describe the field fall off at infinity.

$$E_r(\omega, r, z) = iA \frac{h}{\gamma_o} H_1^{(1)}(\gamma_o r) e^{i(\omega t - hz)} \quad (2.32)$$

$$E_z(\omega, r, z) = AH_0^{(1)}(\gamma_o r) e^{i(\omega t - hz)} \quad (2.33)$$

$$H_\phi(\omega, r, z) = iA \frac{k_o^2}{\omega \mu \gamma_o} H_1^{(1)}(\gamma_o r) e^{i(\omega t - hz)} \quad (2.34)$$

2.2 Theory of the Goubau Transmission Line

The fields are continuous at the boundary between the dielectric and free space. The ratio E_z/H_ϕ yields the following relation at the outside radius r_d of the dielectric

$$\frac{\gamma_o}{\varepsilon_r} \left[\frac{J_0(\gamma_d r_d) Y_0(\gamma_d r_c) - J_0(\gamma_d r_c) Y_0(\gamma_d r_d)}{J_1(\gamma_d r_d) Y_0(\gamma_d r_c) - J_0(\gamma_d r_c) Y_1(\gamma_d r_d)} \right] = \gamma_o \frac{H_0^{(1)}(\gamma_o r_d)}{H_1^{(1)}(\gamma_o r_d)} \quad (2.35)$$

This relation can be used to numerically determine the propagation constant h for the Goubau line, taking into account equation 2.25.

The radial electric field in Eq. (2.32) is proportional to $1/r$ close to the wire where the zero argument approximation of the Hankel function holds.

Goubau focused his work on minimizing the field extension, as most of the power flows in that region. He wanted a robust long range transmission line and things in the environment like tree branches would interfere with transmission if they were to reach into that field region around the wire. We on the other hand want to extend the region in which the field follows a $1/r$ gradient as it is the region in which the radial electric field of the Goubau line resembles that of a charged particle beam. The extend of that region is determined by the radial wave number γ_o , which depends on the properties of the conductor, the dielectric coating, and the frequency of the field. Reducing γ_o extends the distance from the wire at which the Goubau line can emulate the fields of a charged particle beam. The needed frequency is determined by the bunch length of the particle beam to be emulated by the Goubau line.

As stated in Chapter 1, BESSY II features bunches as short as $\sigma_s = 5.1$ mm, which equates to a bunch spectrum of up to 20 GHz. For meaningful measurements the field needs to extend over the aperture of the device under test. In the case of IVUE32 the aperture at maximum gap is 22 mm, so a field extension of at least 11 mm at 20 GHz is required. Figure 2.6 shows the field amplitude as a function of distance at a frequency of 20 GHz for the case of a polyamide coated 12 AWG copper wire compared to a $1/r$ gradient. The field amplitude starts to deviate from the $1/r$ gradient less than 10 mm from the wire. This example does not deliver the required field extension for our measurements. The appropriate wire properties for our test stand will be discussed in chapter 4

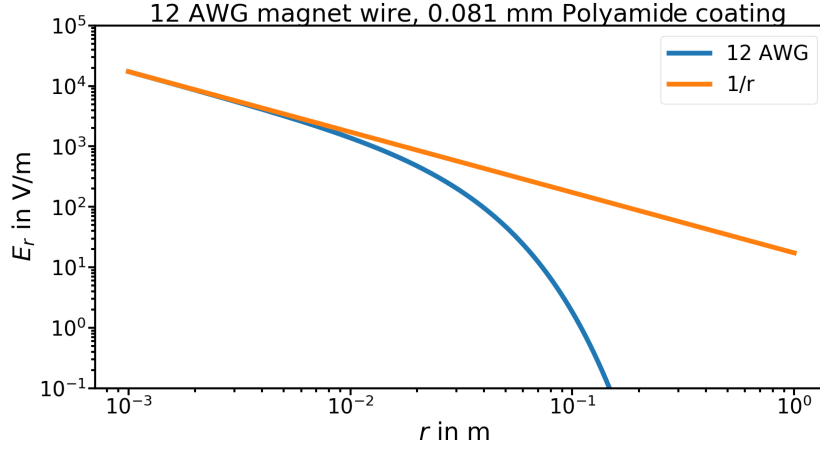


Figure 2.6: Amplitude of transverse electric field of a Goubau line using a polyamide coated 12 AWG magnet wire at 20 GHz compared to a $1/r$ gradient.

2.2.1 Characteristic Impedance

In order to efficiently impart the surface waves onto the Goubau line, the characteristic impedance Z_{GB} of the coated wire needs to be taken into account. The characteristic impedance of the Goubau line depends on the wire parameters and also the frequency of the guided waves. It can be calculated with an equivalent TEM expression as shown by B. Vaughn *et al.* [30].

$$Z_{GB} = \frac{Z_0 \ln\left(\frac{b}{r_c}\right)}{2\pi} \quad (2.36)$$

where $Z_0 = 377 \Omega$ is the intrinsic impedance of free space, r_c is the radius of the Goubau line conductor, and b is the radius of the outer conductor of an equivalent coaxial cable. This equivalent radius is obtained by equating the power flow through the Goubau line to an equivalent coaxial cable

$$P_{eq} = \frac{1}{2} \Re\{V_g I_z^*\} = \frac{4A^2 \omega \epsilon_0 \Re\{h\}}{\pi |\gamma_o|^4} \ln\left(\frac{b}{r_c}\right) \quad (2.37)$$

$$V_g \approx \frac{iAh}{\gamma_o^2} \left(H_0^{(1)}(\gamma_o r_c) - H_0^{(1)}(\gamma_o b) \right) \quad (2.38)$$

$$I_z = \frac{2\pi\sigma A r_c}{\gamma_o} H_1^{(1)}(\gamma_o r_c) \left(\frac{k_o}{k_c} \right)^2 \quad (2.39)$$

with

$$k_c = \omega \sqrt{\mu_0 \left(\frac{\sigma}{i\omega} + \epsilon_0 \right)}$$

where σ is the conductivity of the Goubau line wire. Equation (2.37) is solved numerically for b and the characteristic impedance is then obtained via Eq. (2.36). Figure 2.7 shows how the characteristic impedance of a Goubau line wire changes

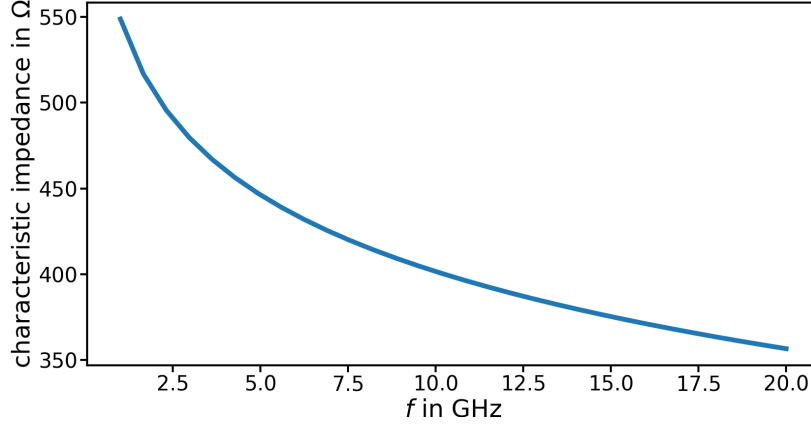


Figure 2.7: Characteristic impedance of a Goubau line using a 1 mm diameter copper wire coated with 500 nm of dielectric with $\epsilon_r = 4$ as a function of frequency.

with the frequency of the guided wave. In this example a 1 mm diameter copper wire coated with 500 nm of $\epsilon_r = 4$ dielectric is used. As the characteristic impedance varies with frequency, it is impossible to find a broad band taper that matches the impedance of the coaxial feed line. The test stand described in Chapter 4 uses conical horn antennas as impedance tapers. These horn antennas are not optimized and the measurements presented in Chapter 5 show the effects of an improperly matched impedance.

3 Simulations

In this chapter the results of numerical investigation of the wakefields and impedance of the relevant parts of IVUE32 are presented. As discussed previously, the investigation focuses on parts of the in-vacuum insertion devices that are specific to the APPLE II design. These parts are the slit shielding foil geometry and the taper section.

Simulations are done using the CST Studio Suite® [31] software by Dassault Systèmes. It is a powerful software package for calculating electromagnetic fields in three dimensions. It includes several electromagnetic field solvers in both the time and frequency domain as well as particle tracking algorithms. These solvers can be coupled for hybrid simulations, enabling the analysis of entire systems comprising multiple components.

In Section 3.1 the settings and parameters used in the simulations are discussed. Section 3.2 presents different design prototypes for the IVUE32 taper sections and the corresponding simulation studies. In Section 3.3 the study on slit shielding foil geometry is presented and different geometric errors and their effect on impedance are discussed. Followed by time domain simulations of the Goubau line test stand set up in Section 3.4.

3.1 CST Solver Settings

The simulations presented in this chapter have been carried out using the CST Studio Suite [31] software. For the taper simulations presented in Sec. 3.2 and the shielding foil slit simulations presented in Sec. 3.3 the wakefield solver is used. The simulation of the Goubau-line test stand presented in Sec. 3.4 utilize the time domain solver.

The wakefield solver specifies a particle beam, described in Sec. 3.1.0.3, which is passed through the modeled structure. The electromagnetic fields caused by the charged particle beams are evaluated with a time domain solver. From these electromagnetic fields, the wake potential is calculated as described in Chapter 2. Since this presets the background material to Perfect Electrical Conductor (PEC), a vacuum cylinder with a 40 cm diameter is added to the taper simulations presented in Sec. 3.2. This vacuum cylinder spans the length of the simulated structure and mimics the vacuum chamber, the final undulator will be installed into.

For the shielding foil slit simulations presented in Sec. 3.3 background is set to vacuum. The transverse boundary conditions are set to `open (add space)` instead of modeling a vacuum chamber. This is done to save computation time as the shielding foils are modeled for a much larger distance than the taper sections and

3 Simulations

adding a vacuum chamber to the full length would dramatically increase the number of mesh cells. The longitudinal boundary conditions are set to **electric** ($E_t = 0$). To further save computation time, the YZ-plane is defined as a magnetic symmetry plane for the shielding foil simulations. This assumes there are no magnetic field components with a direction in that plane. So the magnetic field vectors at points on the symmetry plane are completely orthogonal to the plane.

The time domain solver is essentially identical to the wakefield solver, it just differs in the method of field excitement. Instead of a charged particle beam, the time domain solver uses Gaussian, or otherwise defined, pulse at a defined waveguide port to excite the system.

3.1.0.1 Wakefield Solver Parameters

The fields are simulated for a time interval specified by the wakelength in the solver settings.

To investigate long term field excitation, a wakelength of 20 m is chosen for the taper section simulations, resulting in a simulated time interval of 68.6 ns. This provides a high resolution for the impedance calculations, which is necessary to detect possible resonances of the structure.

For the computationally more demanding shielding foil simulations the calculated wakelength is reduced to 10 m to keep the computation time manageable. This allows for a simulated time span of 36.8 ns.

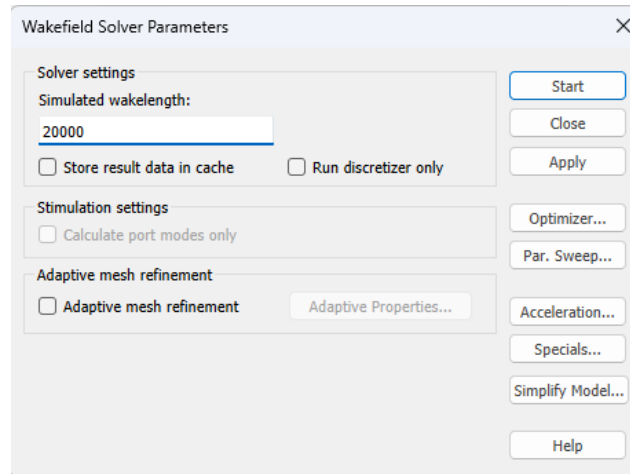


Figure 3.1: CST wakefield solver settings with the wakelength given in mm

3.1.0.2 Power Loss Monitor

The solver settings include an option for calculating the power dissipated by the lossy materials in the simulation. This can be done as a function of frequency or time. In our case we are interested in the loss as a function of time. In the power

loss monitor settings the option *Store per solid* can be activated. This shows which part is responsible for how much power dissipation. Figure 3.2 shows the settings menus to activate the power loss monitor.

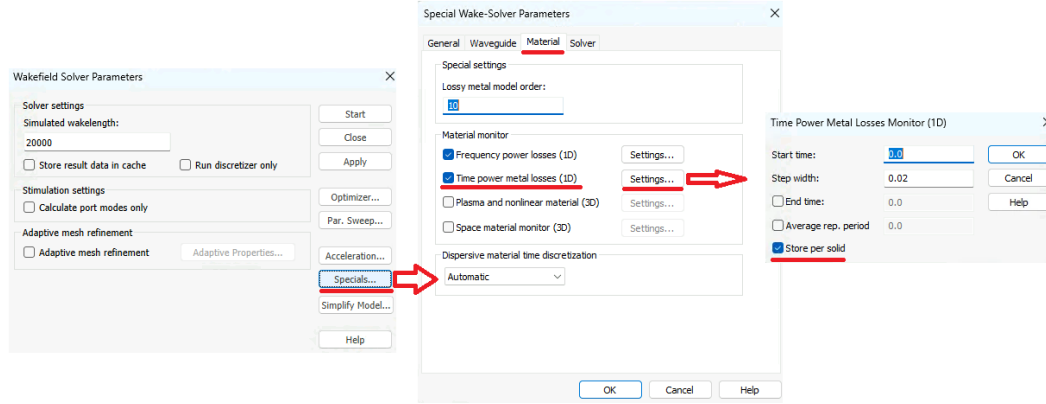


Figure 3.2: CST Wakefield solver settings to activate the power loss monitor

3.1.0.3 Beam Parameters

The beam is defined as ultra-relativistic with a longitudinal Gaussian profile. CST only models pencil beams so there is no transverse charge distribution considered. There are two possible options for the current injection scheme *Transmission line* and *Analytic*. The *Transmission line* scheme was chosen here. This choice together with the mesh settings is explained in section 7.4.1. A screenshot of the beam settings is shown in Fig. 3.3.

Both the taper and shielding foil simulations are run with a test beam of greatly exaggerated bunch charge 10.4 nC. The longitudinal RMS bunch length is set to $\sigma_z = 5.1$ mm, comparable to the standard user bunch length of BESSY II. The high beam current is chosen to highlight resonances in the impedance of the simulated structure and to exaggerate losses. The idea is to easier identify geometric problems in the taper design prototypes.

To verify the impedance results and to calculate the actual power loss in the structure, significant geometries are also simulated with beam parameters of BESSY II operating modes. The parameters of select BESSY II operating modes and the test beam are compared in Tab. 3.1.

Table 3.1: Comparison of bunch length and bunch charge of different simulated beams

bunch type	test beam	standard user	single bunch	low- α A	low- α B
charge [nC]	10.4	0.72	3.2	0.036	0.228
σ_z [mm]	5.1	5.73	10.18	0.89	2.04

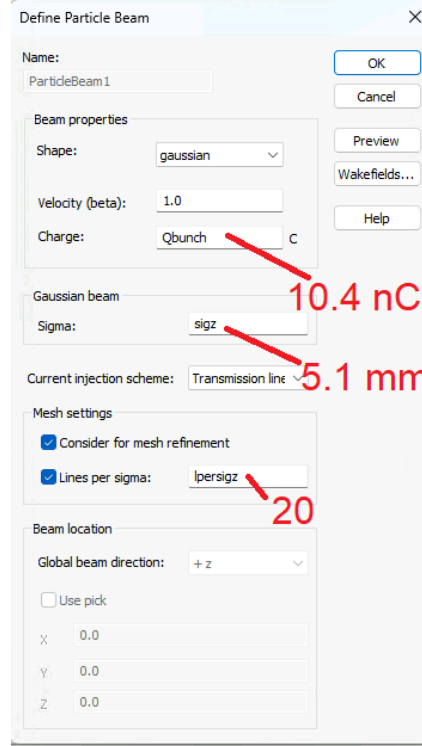


Figure 3.3: CST Wakefield solver electron beam settings

3.1.0.4 Mesh Settings

The global mesh settings are kept fairly coarse at 5 cells per wavelength. The wavelength refers to the highest frequency excited by the beam, 20.0767 GHz in this case. This is done to keep the mesh cell count low and to save simulation time. The simulation area is rather large due to the vacuum chamber cylinder but there is very little field propagation far away from the beam. So a coarse resolution in the outer areas does not affect the accuracy of the simulation. However, much higher local mesh resolutions are defined for several significant areas:

- The slit between the shift shielding foils is only 0.2 mm wide. To accurately simulate that structural feature a resolution of 4 mesh lines across the slit was defined. This fine resolution would be prohibitively expensive for the rest of the structure so it only covers the small area of the actual foil slit.
- The second significant structural feature is the cavity of the shift block presented in Sec. 3.2.2. Here a resolution of 200 mesh lines across horizontally and 100 across vertically is defined.
- The third region of greater interest is the beam channel. This area is already covered by the increased resolution of the shift block cavity and therefore no separate mesh resolution was defined.

These resolutions are defined by placing vacuum bricks in the areas of interest and defining local mesh properties for them. Intersecting vacuum components with metal structures is not a problem because any metal takes priority over vacuum as far as the solver is concerned. These mesh settings mainly cover the transverse mesh resolution. The longitudinal mesh resolution is determined by the beam injection scheme and discussed in section 7.4.1. This results in a horizontal and vertical transverse mesh cell size of around 0.35 mm with slight variations depending on which surfaces the mesh lines are snapped to.

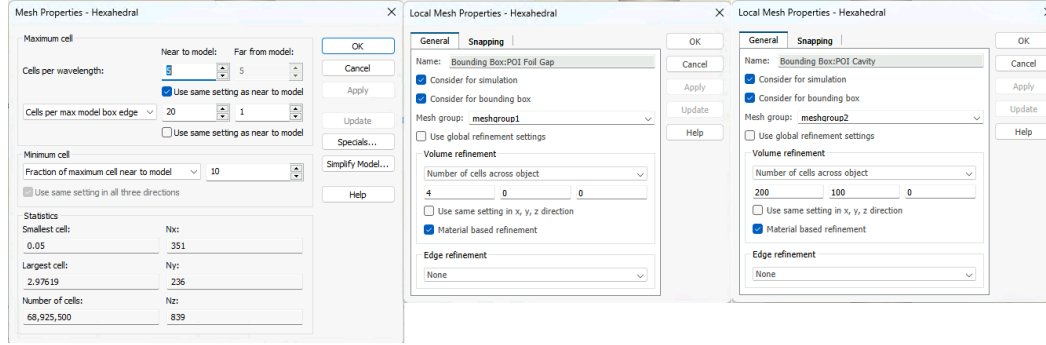


Figure 3.4: Mesh resolution settings, global settings on the left with the two specified local resolutions on the right.

3.2 IVUE32 Taper Simulations

3.2.1 Design Philosophy

The IVUE32 taper geometry described in Section 1.5 is the result of several rounds of feedback between design ideas and wakefield simulations.

The initial idea was a taper section similar to the proven design of CPMU17 [5, 32] but with a longitudinal separation matching the split of the magnet girders. This would result in two independent taper assemblies next to each other, where tensioned shielding foils form the transition from the vacuum chamber entrance to the magnet girders. Changing the shift state of the IVUE would then result in different slopes for the two tensioned foils right next to each other. To avoid this gradient mismatch, it was decided to separate the mechanical compensation of the gap and shift movement.

The first iteration, achieving this separation, is shown in Fig. 3.5. The shift movement compensation is done by slides that move in and out of a cavity. There is no physical contact between the slides and the cavity walls. This was done to preserve vacuum quality. Further details and wakefield simulations of this design are discussed in Sec. 3.2.2. This taper design would significantly impede beam operations, making it not feasible.

Subsequent design iterations follow the same principle of separating the movement

3 Simulations

compensation. The gap compensating section of the taper remains largely unchanged for the following design iterations.

Another simpler taper design idea keeps the physical discontinuity but eliminates the cavities in the shift block. A CAD model of this shift compensation taper is shown in Fig. 3.7. Wakefield simulations of this design are discussed in Sec. 3.2.3. This design still shows prohibitively high impedance, suggesting that the discontinuity is a more significant problem than the cavities.

This is confirmed by adding simple sliding contacts where the discontinuity gap would otherwise be, as discussed in Sec. 3.2.3.1. Removing the discontinuity drastically reduces the longitudinal impedance of the taper structure for frequencies below 12 GHz.

This taper design still has a vertical step in gap of about 1 mm where the slides contact the block. At shift states other than neutral, this step is not at the same longitudinal position for the left and right slide, causing a transverse asymmetry in geometry. Section 3.2.4 discusses a simplified step model to investigate the impedance implications of such a transverse asymmetry.

A later design iteration, discussed in Sec. 3.2.6, foregoes both issues of transverse asymmetry and and step in gap geometry with the help of a complex foil tensioning system. This tensioning mechanism also eliminates any difference in gap. This design was presented at IPAC 2025 [17] and is shown in Fig. 1.7.

3.2.2 First Taper Design

The first design proposal that separates the gap and shift movement compensation is shown in Fig. 3.5. This design does not have direct electrical continuity between

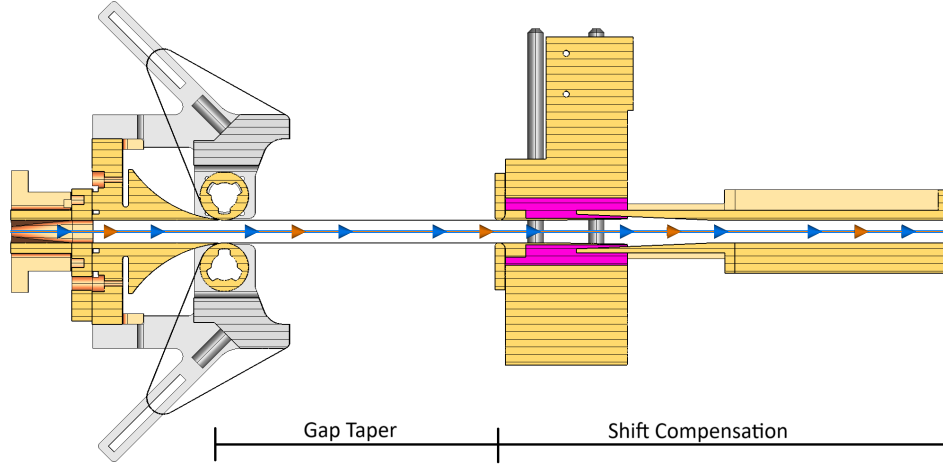


Figure 3.5: Cross section of the first IVUE32 taper design. Gap taper with foil tensioning mechanism on the left and shift movement compensation on the right. Cavity of the shift block highlighted in pink.

the slides and the block with the cavity, as they are not physically touching. This was done because of the relatively long ± 16 mm travel of the shift movement and because during operations, changes in shift state are expected frequently. Having touching metal surfaces frequently slide against each other in a high vacuum environment is not ideal, as cold welding and ejection of metal dust are a concern.

The shift movement compensation mechanism was simulated in different configurations. The main concern was exciting long lasting modes in the shift block cavity. The electric field is recorded with a 2D field monitor that records the Y-Z-plane and is located in the middle of the longitudinal slit.

Figure 3.6 shows the absolute value of the electric field as the beam passes the gap between the slides and the block at different time steps. Large field amplitudes can be seen in the gap between the two surfaces. This area shows high field amplitudes even long after the beam passes, with some oscillating fields persisting for more than 2 ns at which point the next bunch would arrive. This indicates structural resonances that are excited by the passing beam, which could cause instabilities for the following bunches. If resonances are hit, a lot of energy could be deposited into the shift block assembly causing overheating issues. This taper design would significantly impede beam operations, making it not feasible.

3 Simulations

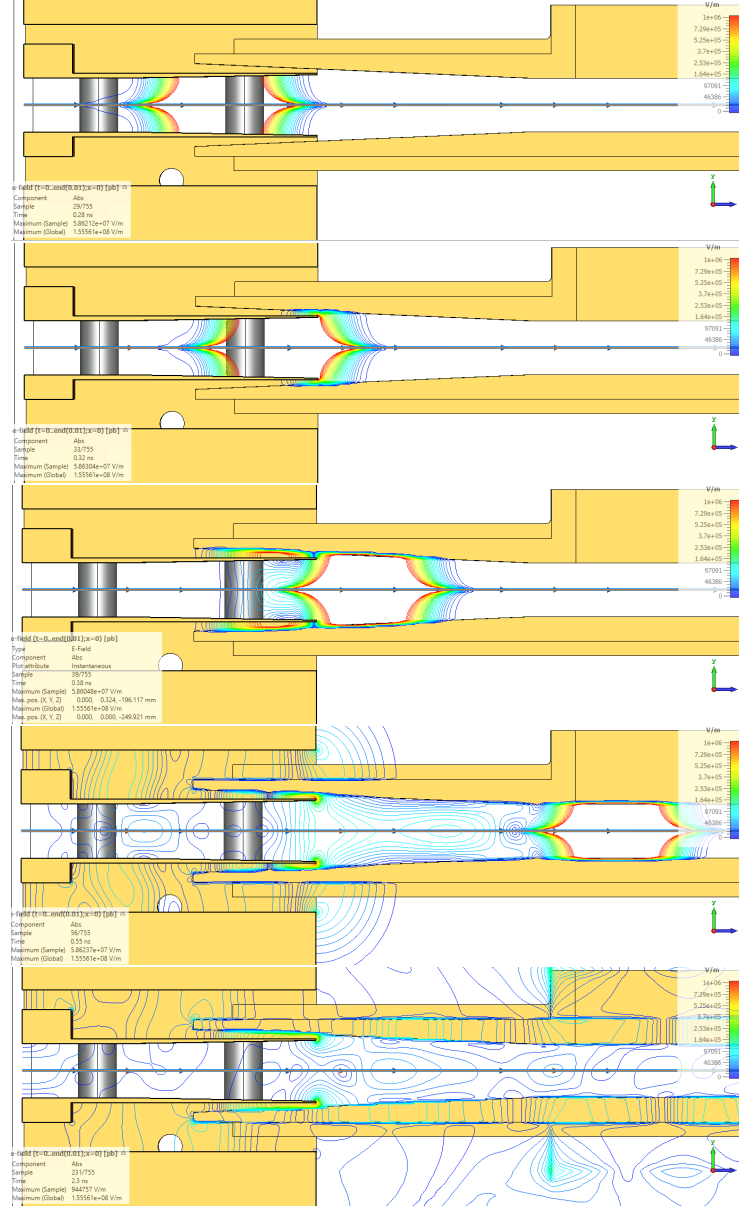


Figure 3.6: Absolute value of electric field shown as isolines for the shift movement compensation assembly of the entrance taper. Cross section of the Y-Z-plane. The top three pictures show the beam propagating over the gap in the shift block and exciting a field in the cavity. The fourth picture shows the field developing and radiating out into the larger vacuum chamber. The fifth picture shows the field state 2 ns after the first picture, when the next bunch would arrive at the gap.

3.2.3 Second Taper Design

The second taper has a somewhat simpler design. There is no cavity in the shift block, just a taper that contours to the shift slides. A cross section of the CAD model is shown in Fig. 3.7. This shift movement compensation assembly is simu-

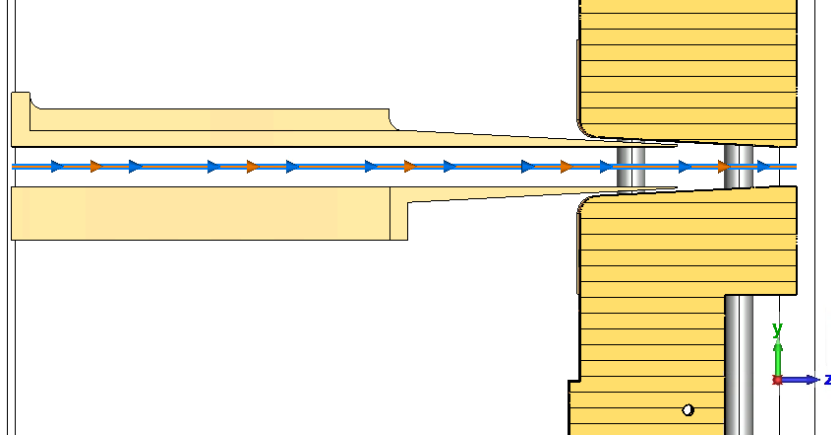


Figure 3.7: Cross section of the shift movement compensation assembly of the second taper design.

lated similar to the entrance taper. The mesh settings are kept the same and the electric field is recorded with a 2D field monitor on the Y-Z-plane.

Figure 3.8 shows the electric field propagation with a focus on the gap between the shift block and the shift slides. As with the previous design, the gap at the discontinuity shows high remaining field amplitudes long after the beam passes. Some oscillating fields are excited that persist for more than 2 ns at which point the next bunch would arrive.

3 Simulations

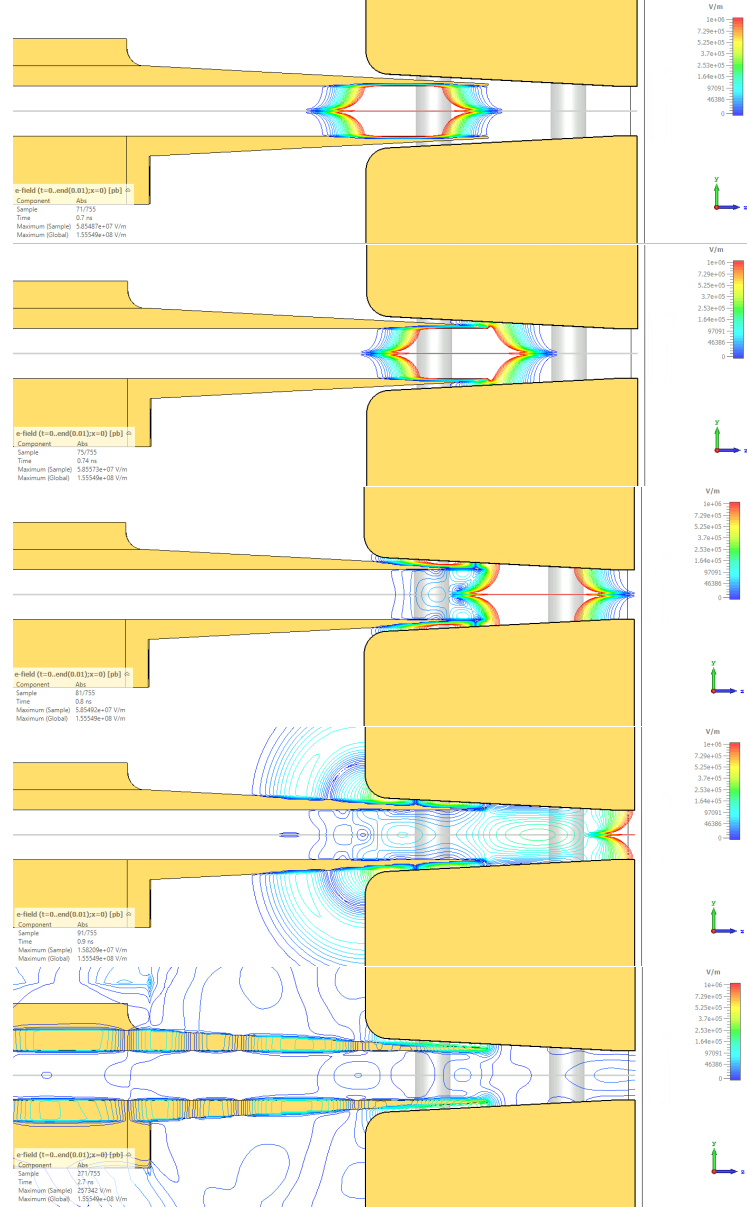


Figure 3.8: Absolute value of electric field shown as isolines for the shift movement compensation assembly of the second taper design. Cross section of the Y-Z-plane. The top three pictures show the beam propagating over the gap between the shift block and the shift sleds and exciting a field therein. The fourth picture shows the field developing and radiating out into the larger vacuum chamber. The fifth picture shows the field state 2 ns after the first picture, when the next bunch would arrive at the gap.

The initial idea for the first and second taper designs was that a step with an increase in vacuum chamber cross section would be less disruptive than a step with a decrease in cross section. In theory, the impedance of a structure is strictly dependent on the geometry and should be independent of the beam direction. The second taper design was simulated again with a reversed beam direction to test that theory. Figure 3.9 shows the longitudinal wake impedance of the second taper design for both beam directions. At frequencies below 13 GHz there is no significant difference between the beam direction. Above 13 GHz the reversed beam shows a smaller impedance. The simulations were done for a gap of 11 mm. As explained in Section 7.4.2, the cut-off frequency for this gap is close to 13 GHz, allowing excited waves to propagate above that frequency. The impedance values above the cut-off frequency are most likely just numerical noise.

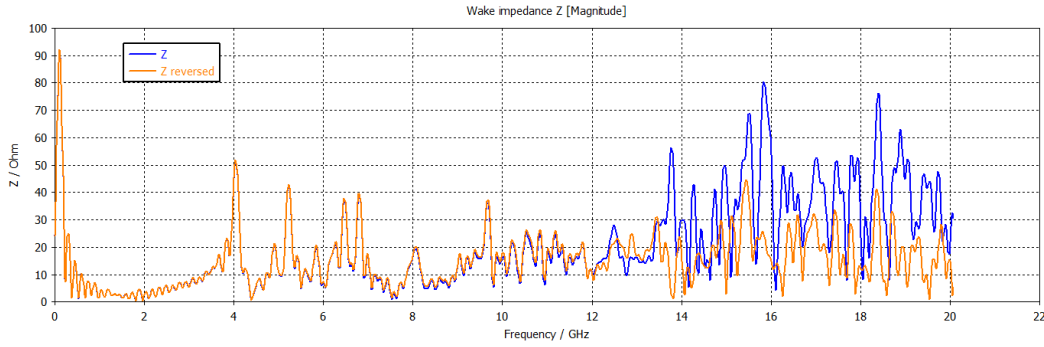


Figure 3.9: Longitudinal wake impedance of the second taper assembly. Comparison with a reversed beam direction. At frequencies below 13 GHz there is no significant difference between the beam direction. Above 13 GHz the reversed beam shows a smaller impedance.

3.2.3.1 Sliding Contacts

After examining the impedance, there is no feasible way of allowing a discontinuity so close to the beam path. Therefore both initial designs for the taper section are disqualified. Although there are concerns with vacuum quality some sort of sliding contact will have to be added to the shift compensating mechanism. This was done in CST to the second taper design as a proof of concept, shown in Fig. 3.10. The longitudinal impedance is greatly impacted by the sliding contacts. Especially in the lower frequency range as shown in Fig. 3.11. The large low frequency peak disappears as the discontinuity is removed.

3.2.4 Simplified Step

Design iterations featuring sliding contacts to ensure electrical continuity show a dramatic decrease in longitudinal impedance as shown in Fig. 3.11 compared to

3 Simulations

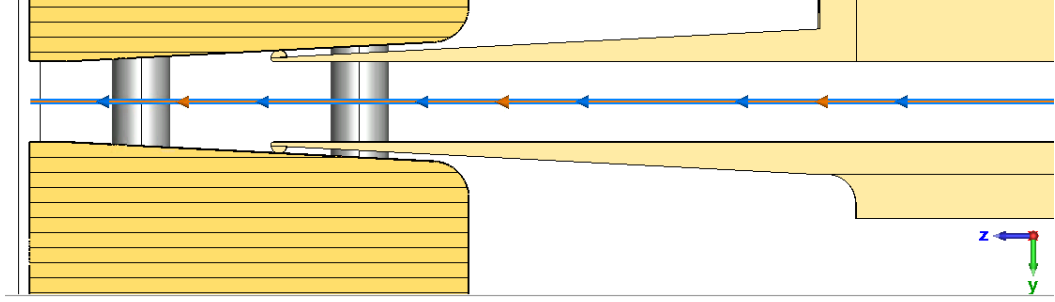


Figure 3.10: CAD model of sliding contacts added to the second taper design.

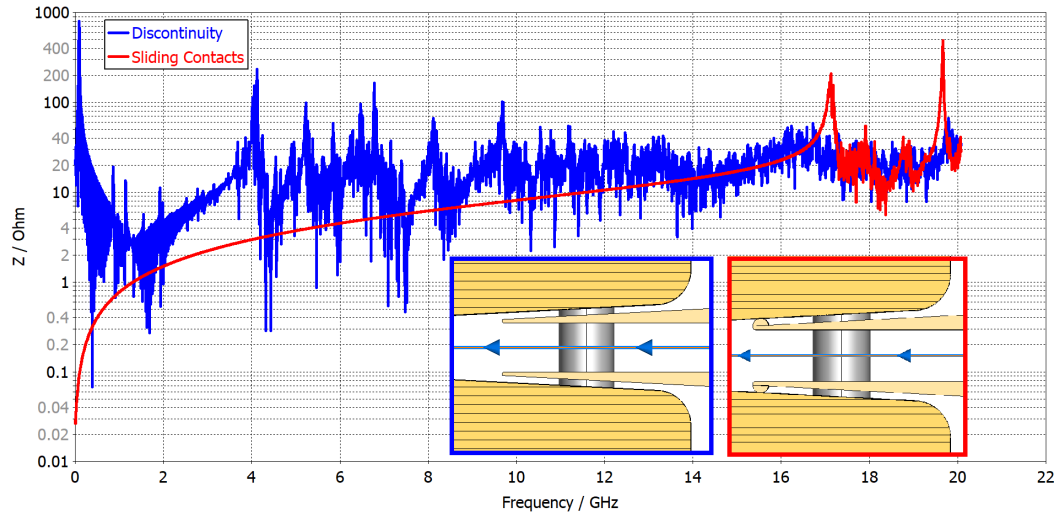


Figure 3.11: Longitudinal impedance of the second taper design of shift compensating assembly at a gap of 7 mm. Comparison of discontinuity and sliding contacts. The sliding contacts significantly reduce the impedance for lower frequencies.

designs with a discontinuity. Their geometry however brakes both vertical and horizontal transverse symmetry during any shift state other than the neutral position of planar polarization.

A simplified model was created to study the effects of a horizontal and vertical asymmetry created between the adjacent shift sliders at different shift states. This simplified design features tapered shift sliders that increase the gap before returning to the original gap in a single step. The step is kept at a fixed position regardless of the shift state of the sliders.

The CST model is shown in Fig. 3.12. Three positions are investigated, the neutral position without shift, a maximum symmetric shift of 16 mm where two diagonally opposing sliders are moved in the same direction and a maximum asymmetric shift where two diagonally opposing sliders are moved in the opposite direction.

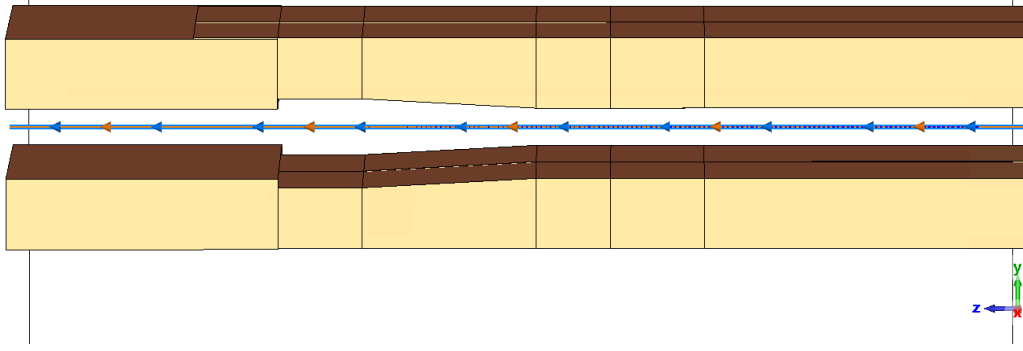


Figure 3.12: Simplified design to study transverse asymmetries created by tapered shift sliders during shift states. Neutral shift position shown.

Figure 3.13 shows the asymmetry of one such study together with the transverse horizontal impedance at different shift states. The magnitude of the transverse

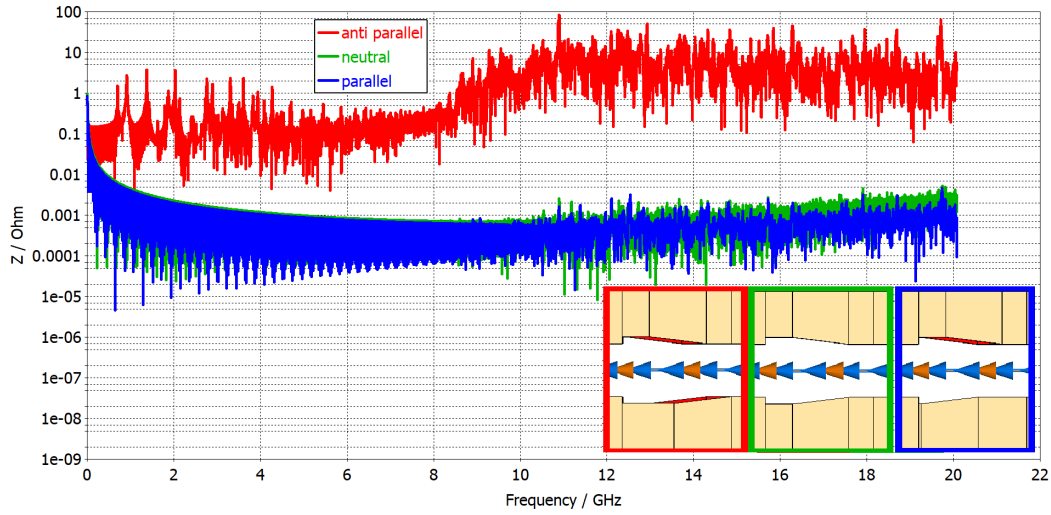


Figure 3.13: Horizontal impedance of tapered shift compensation at different shift states producing transverse geometric asymmetries highlighted in red

impedance for the anti parallel shift state is significant and could result in a kick factor, destabilizing the beam.

The longitudinal impedance for these three cases is shown in Fig. 3.14. The transverse asymmetry during shift states only has a marginal effect on the longitudinal impedance. There are some peaks in the lower frequency range for the anti parallel shift state, however with small amplitudes. The simulations were done with a gap of 15 mm so the signal is very noisy above the cut-off frequency of around 10 GHz.

In conclusion it can be said that a transverse geometric asymmetry has a more significant effect on the transverse impedance than on the longitudinal impedance.

3 Simulations

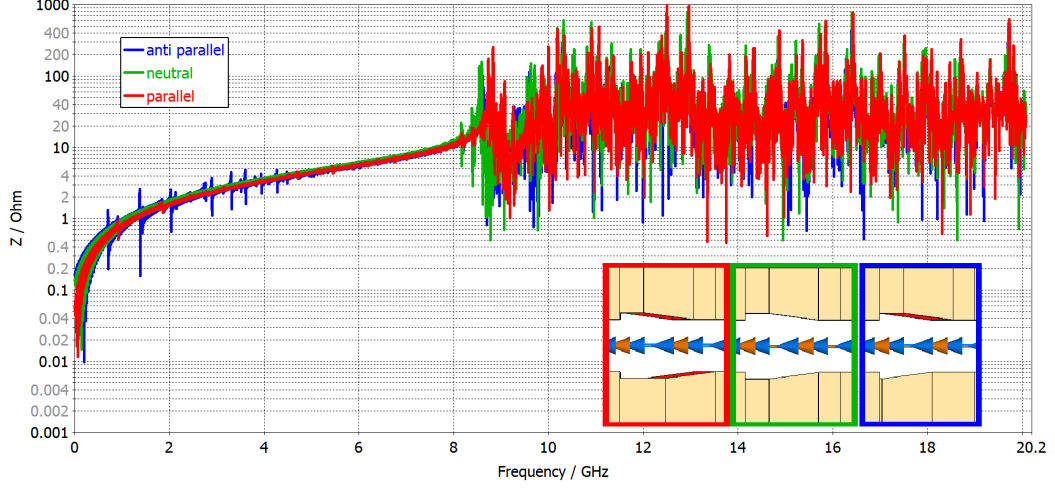


Figure 3.14: Longitudinal impedance of tapered shift compensation at different shift states producing transverse geometric asymmetries highlighted in red.

This means defocusing or steering effects on the beam pose are more concerning than power losses when it comes to transverse asymmetries. The acceptable magnitude of these transverse asymmetries depends on the available beam orbit correction mechanisms.

3.2.5 Double Finger Designs

A design idea with two opposing sliding fingers was simulated. A schematic of the design is shown in Fig. 3.15. The key feature of this design idea is to keep the position and height of the longitudinal step independent of the shift state and constant. This however means, that the gap decreases by 2 mm at the point of the step. This means at the minimum undulator gap of 7 mm, the gap at the shift block will only be 5 mm, which might be a problem with the β -function of the storage ring. The longitudinal impedance was compared to the sliding contact design and is shown in Fig. 3.16. The double finger design shows a low baseline impedance of around 8Ω over the lower frequency spectrum. At 10 GHz we see the noise expected for a 15 mm gap. Compared to the sliding contact design the impedance is remarkably flat up to the cut-off frequency. The power loss was also compared for the two designs. Figure 3.17 shows that the double finger design has a grater initial power loss as the beam passes but shows less ringing afterwards. This might indicate greater stability with repeated excitation by following bunches.

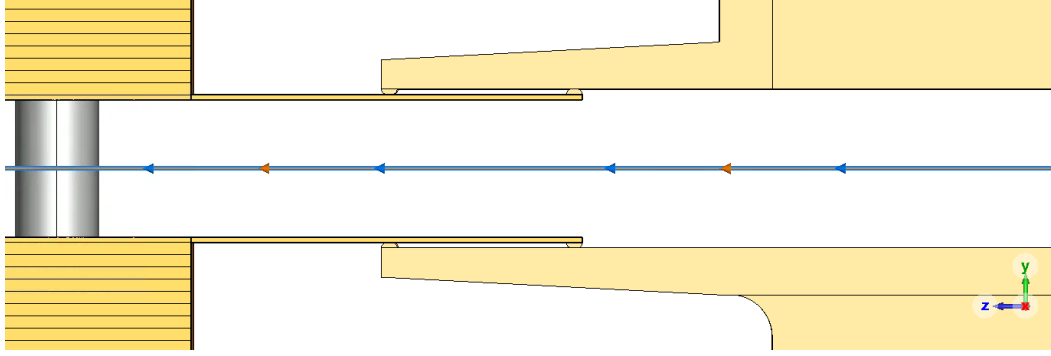


Figure 3.15: Schematic of double finger design. This design features a longitudinal step that is independent of shift state.

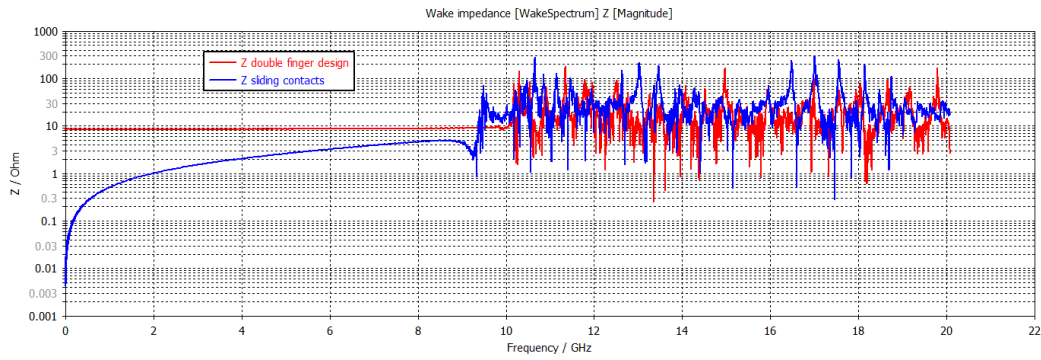


Figure 3.16: Longitudinal impedance of double finger design compared to the second taper design with sliding contacts. Both simulations are done at a gap of 16 mm.

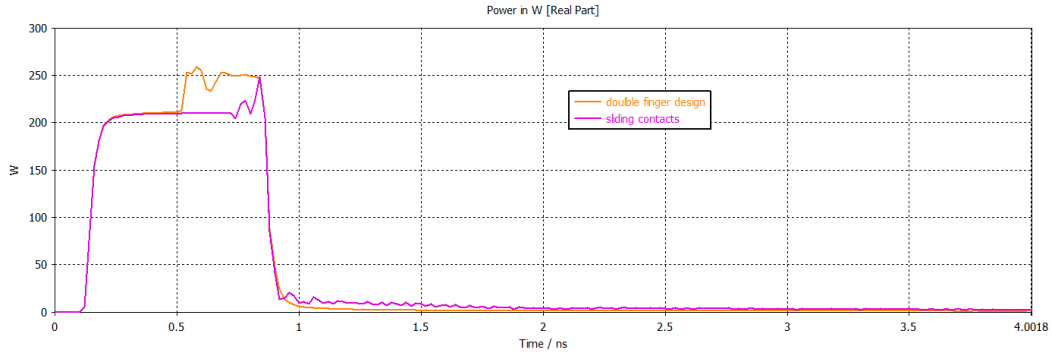


Figure 3.17: Power loss data for the first 4 ns. Double finger design compared to sliding contacts. The double finger design has greater initial power loss as the beam passes but has less ringing afterwards.

3.2.6 Foil Tensioner Design

A complex foil tensioning system is introduced to the shift compensating assembly, that avoids tapered sliders and does not break transverse symmetry regardless of shift state. A split shielding foil is introduced, which is reeled in or out by a tensioning mechanism as the magnet girder moves longitudinally during phasing operations. A CAD model of this mechanism together with the gap taper is shown in Fig. 3.18 There is only a minimal gap change of 0.1 mm where the movable foil is

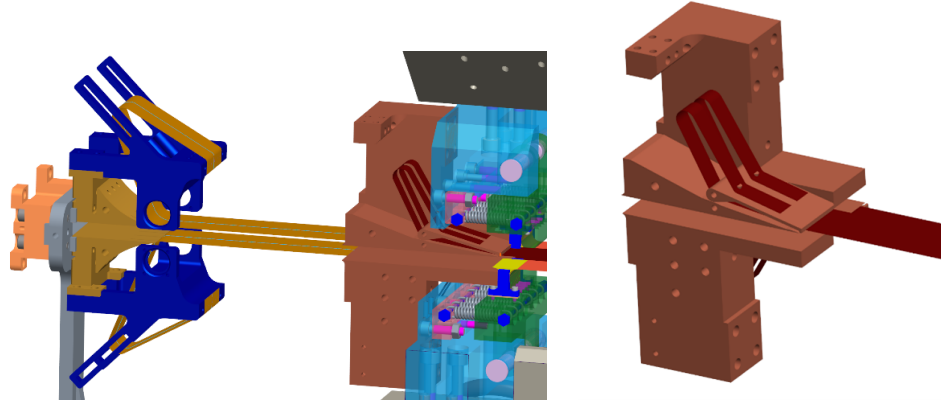


Figure 3.18: CAD model of the IVUE32 taper section. Full taper with gap and shift compensation on the left. On the right the foil tensioning principle of the shift taper is shown. Taken from [17]

let in or out by the tensioning mechanism. The position of which does not change during the phasing movement.

The longitudinal impedance of the foil tensioner design is compared to that of the second taper design with added sliding contacts for a gap of 7 mm in Fig. 3.19. The foil tensioner design has a lower longitudinal impedance accross the whole simulated frequency spectrum. The improvements are especially notable at frequencies above 16 GHz where the 1 mm step of the sliding contacts shows resonant behavior. The longitudinal impedance of the foil tensioner design, resembles that of two parallel plates. Which is not surprising, as the relevant geometry closely matches that of two parallel plates as well. From a beam dynamics standpoint this design is close to optimal. Now the challenge remains with the mechanical implementation of such a design.

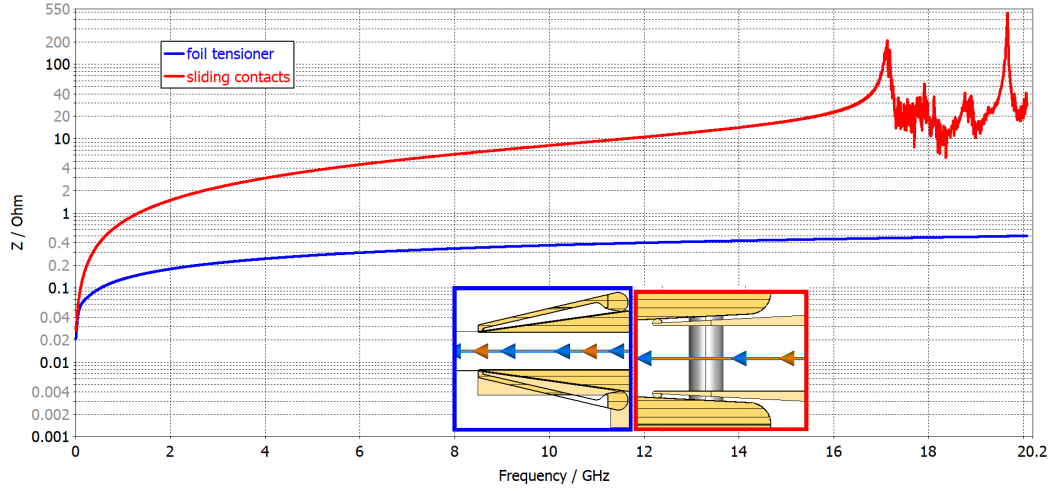


Figure 3.19: Longitudinal impedance of foil tensioner design compared to earlier design iteration with sliding contacts at a gap of 7 mm

3.2.7 Compound Taper

The current BESSY II beam optics have a minimum gap limitation for the straight section IVUE32 will be installed into. Due to the β -function and to maintain safety margins for the dynamic aperture, it could be necessary to immediately increase the gap outside of the magnet girders. This would prohibit the foil tensioner shift compensation mechanism, that maintains the same gap of the magnet girders over a distance of about 30 cm.

This study investigates a compound taper, that increases the gap from 7.4 mm to 9 mm over the length of the shift compensating mechanism. This introduces transverse asymmetries, similar to the case discussed in Sec. 3.2.4. The magnitude of the asymmetry is much smaller, due to the limited gap gradient.

The full taper section from the vacuum chamber entrance to the magnet girders is modeled as simplified foils. The gap taper section is unchanged and the compound taper replaces the shift compensating section. The CST model is shown in Fig. 3.20 for three different shift states. The horizontal axis is scaled up to make the asymmetries, highlighted in red, more visible.

Figure 3.21 shows the longitudinal impedance of the compound taper section simulated with the BESSY II standard user beam. There are a few peaks visible below the cut off frequency of about 14 GHz. The frequency and amplitude varies slightly for the different shift states. The amplitudes remain relatively small with the largest peak reaching $27\,\Omega$, which is still only $3\,\Omega$ higher than the $24\,\Omega$ base impedance due to the aperture mismatch explained in Ch. 2.

As with the model described in Sec. 3.2.4, the longitudinal impedance seems manageable despite the transverse asymmetries. This leaves the question of the transverse impedance. The horizontal and vertical impedance are shown in Fig. 3.22.

3 Simulations

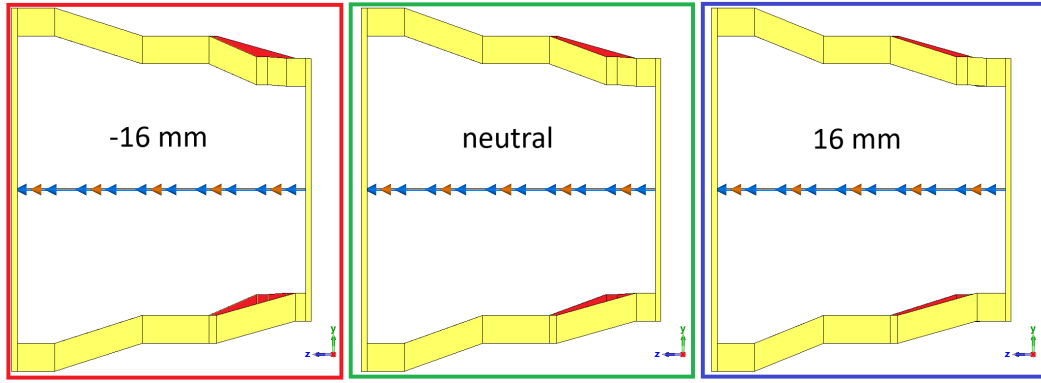


Figure 3.20: Model of compound taper at three different shift states. The vertical dimension is scaled up to emphasize the asymmetry, highlighted in red.

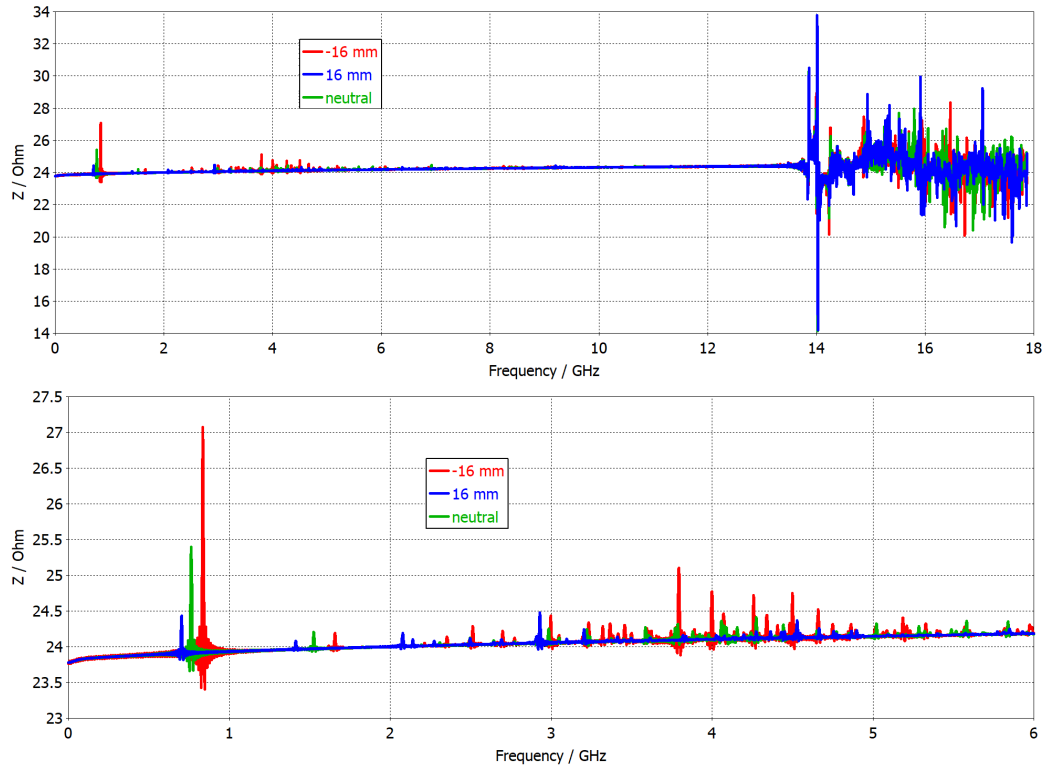


Figure 3.21: Longitudinal impedance of compound taper section simulated with the BESSY II standard user beam. Three shift states are compared. Full frequency range at the top, low frequency excerpt at the bottom.

The transverse impedance for all three shift states of the compound taper section shows a low magnitude without any discernible resonances.

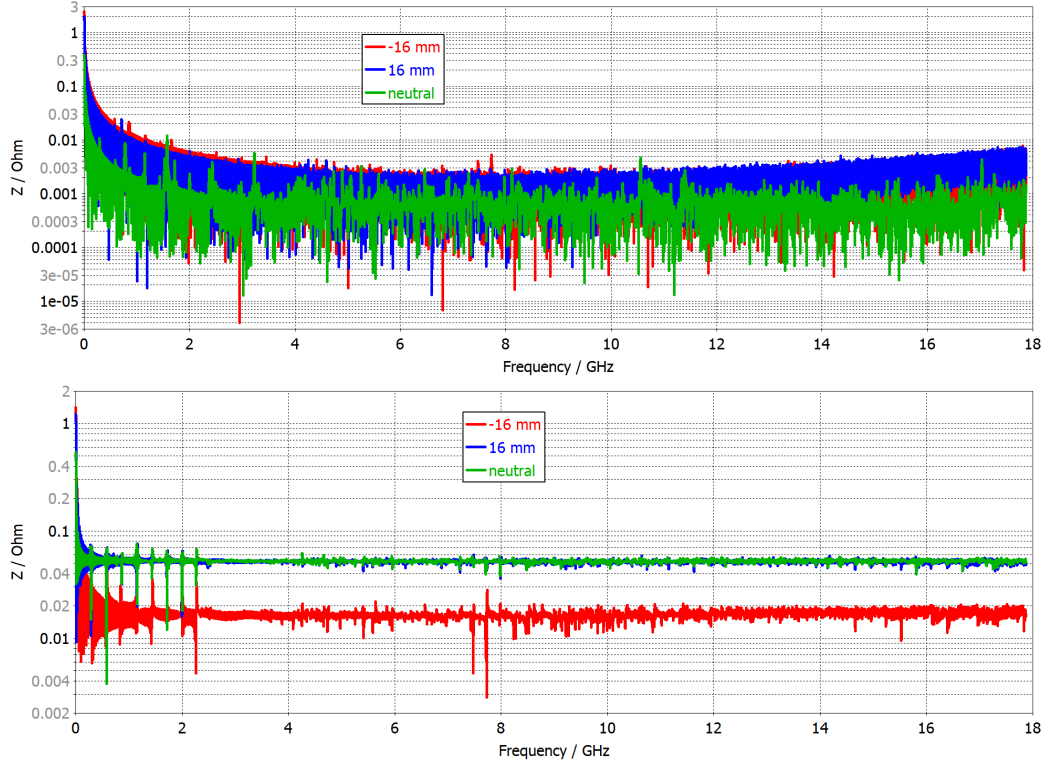


Figure 3.22: Transverse impedance of compound taper section simulated with the BESSY II standard user beam. Three shift states are compared. Horizontal impedance at the top, vertical impedance at the bottom.

This indicates that slight asymmetries due to compounding gap and shift changes in the taper section are manageable.

3.2.7.1 BESSY Beams

This compound taper section was simulated with the specified test beam and the most relevant beams found in regular BESSY II operations. Figure 3.23 compares the longitudinal impedance of the compound taper in neutral shift state for the test beam, and the BESSY II standard user and single bunch. The impedance matches closely for all three beams. This is expected, as the impedance of a structure is defined by its geometry. The bunch length defines the upper limit of the simulated frequency, therefore the longer single bunch is only shown for frequencies below 10 GHz. The much shorter low- α beams have been omitted because the compound taper shows a cut-off frequency of about 14 GHz. Simulating a short beam, that excites much higher frequencies would not yield any new information and would just waste substantial simulation time.

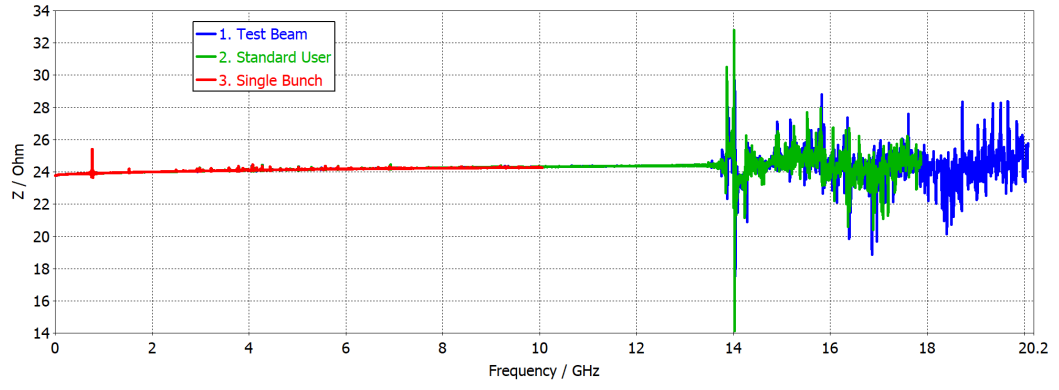


Figure 3.23: Longitudinal impedance of the compound taper in neutral shift state simulated with several different beam parameters. As a geometric property, the impedance matches closely for all beams.

3.3 Foil Slit Simulations

The APPLE-II design of IVUE32 requires a longitudinal slit in the shielding foils along the magnet rows to allow for the shift movement. The current design calls for the shielding foils to be folded into the gap between the magnet rows as shown in Fig. 3.24. This assures that the parts of the magnets closest to the beam trajectory

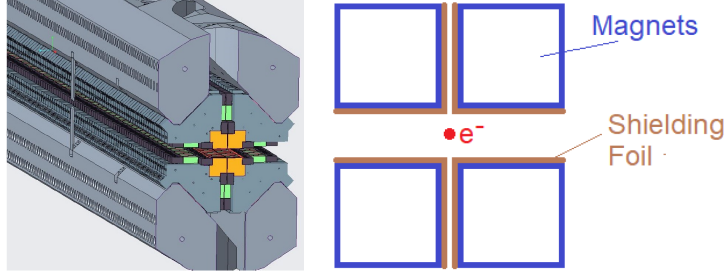


Figure 3.24: CAD model of IVUE32 magnet assembly on the left and schematic of how the shielding foils will be folded into the magnet gap on the right.

are fully covered by the shielding foil. Previous communications with potential foil manufacturers revealed certain possible deviations that can occur during the bending process [33]. Sinusoidal deviations can be expected at the folded edge. To assess the impact of these deviations a series of simulation studies have been carried out. Only a 1 m section of shielding foils is simulated to save simulation time. The length of the shielding foils of IVUE32 will be close to 2.5 m. Periodic geometric errors are investigated which repeat several times over the 1 m section. Impedance and resonances caused by these errors will therefore be detectable. Electric losses can be calculated per unit length and extrapolated to the final dimension of the IVUE structure.

The Foils include a longitudinal slit with a width of 1 mm. The slit starts 20 mm into the shielding foils and ends after 960 mm, leaving a 20 mm section at the beginning and the end of the foils without a slit. Several different sinusoidal deviations are compared. In one case the width of the slit is kept constant at 1 mm with different amplitudes and period lengths of the sinusoidal deviations. This type of variation will be referred to as *asymmetric*. Another case mirrors the sinusoidal deviations, resulting in a variation of slit width, which will be referred to as *symmetric*. A schematic of each of these variations is shown in Fig. 3.25.

These variations are then compared to a simulation with a straight slit of 1 mm width and also two parallel foils without any slit as a baseline.

3.3.1 Modeling

The first results of these simulation studies were presented at the 2023 IPAC conference [34]. As a first step, the parameter space relevant for IVUE32 was explored

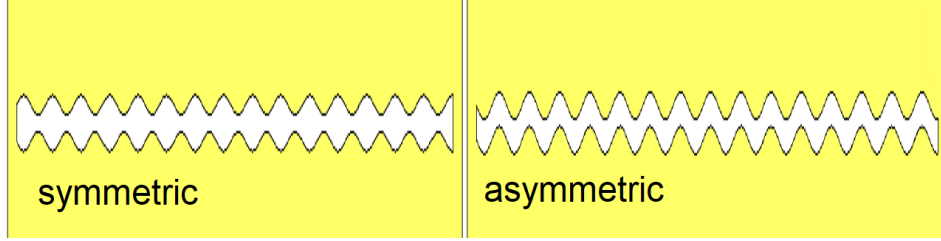


Figure 3.25: Schematic top view of slit geometry in shielding foils. Symmetric sinusoidal variations of the slit width on the left, asymmetric sinusoidal modulation of slit with constant width of 1 mm on the right.

in a qualitative manner. The focus was on fast simulation times and examining large numbers of variations. This meant fairly coarse meshing and limited wake-length in each individual simulation. This was done to search for obvious show stoppers.

In a second step, the simulation settings were optimized for more quantitative studies of a few likely variations, increasing the mesh resolution and simulated wake-length. The different optimization steps are discussed in detail in Sec. 7.3.1.

The shielding foils will be made of copper and have a thickness of about 40 μm with a nickel coating on one side. The Nickel coating is on the side opposite the electron beam and is used to attract the foil to the permanent magnets of the undulator.

Modeling metallic objects in CST with dimensions smaller than the mesh cell size can be problematic. Therefore the Foils are modeled as thick sheets with the slit cut out. The relevant frequencies for the simulation are in the GHz range, making the skin depth very small. The solver itself only considers metallic surfaces for this reason. Simulations with different foil thicknesses are carried out to optimize the simulation settings. This process is presented in Sec. 7.3.1.

3.3.2 Real BESSY Beam Parameters

Several different foil slit deviations are simulated with the actual beam parameters for the BESSY II operation modes described in Sec. 1.1.

The real BESSY II bunches feature a much lower charge than the 10.4 nC of the test beam used in the other simulations. Especially the very short bunches of the low- α operating mode have very little charge. The test beam parameters, especially the high current, are chosen to exaggerate resonances in the impedance. This makes geometric problems of taper and foil designs more obvious.

To make sure the test beam only effects the scaling of losses and impedance and not the quality of the results, the test beam is compared with the standard user beam of BESSY II. Figure 3.26 shows the longitudinal and horizontal impedance for the standard user beam and the test beam. The resonances are at exactly the same frequencies. The magnitude however is significantly lower for the standard user beam. Therefore the test beam is very well suited to find resonances and

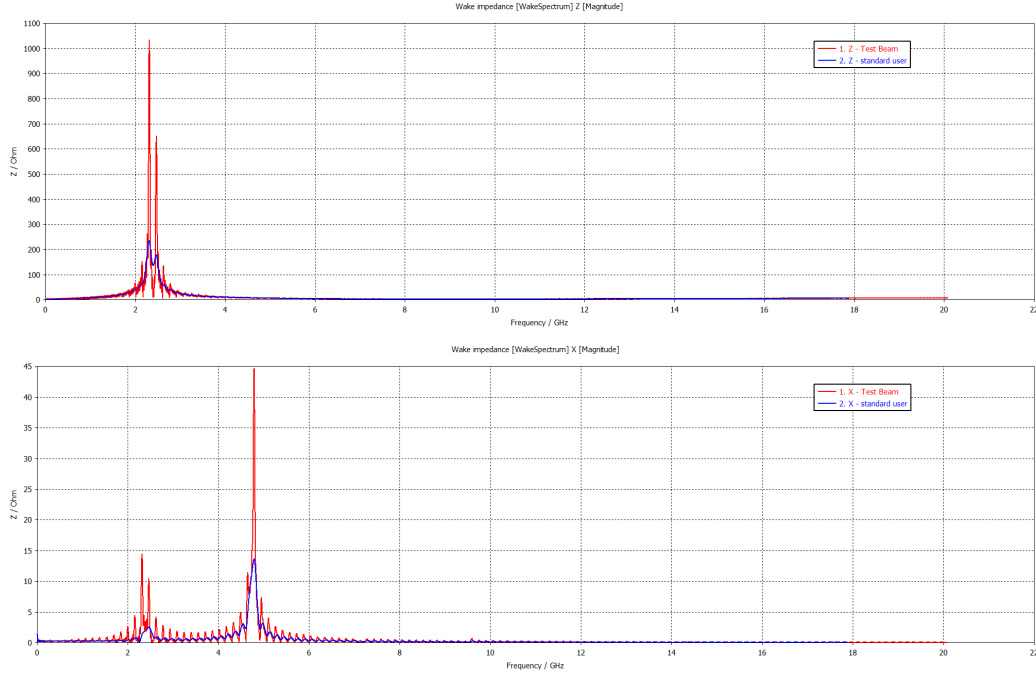


Figure 3.26: Impedance of sinusoidal deviation of the foil slit. Test beam compared to BESSY II Standard User Beam. Longitudinal impedance at the top and horizontal impedance at the bottom.

identify their sources in the geometry. However, for actual power loss calculations, the bunch charge of the real BESSY II beams is used.

3.3.3 Impedance of Sinusoidal Slit

As a starting point the sinusoidal slit discussed above is modeled with an amplitude of 1 mm, meaning the slit location varies over 2 mm in the horizontal direction. This is completely exaggerated as the slit between the two magnet rows will only have a width of about 1 mm.

The period length of the sinusoidal oscillation is $\lambda = 20\pi \text{ mm} \approx 62.83 \text{ mm}$. Figure 3.27 shows two longitudinal resonant peaks at Frequencies of 2.31 GHz and 2.47 GHz. One could reasonably assume the resonant frequencies have some connection to the period length λ of the sinusoidal deviations

$$f = \frac{c}{\lambda} \quad (3.1)$$

with c being the velocity of the charged particle bunch. Here the bunch is traveling at the speed of light.

For a sinusoidal period of $\lambda = 20\pi \text{ mm}$ we get a horizontal frequency of 4.77 GHz. Figure 3.27 shows the longitudinal and horizontal impedance of a foil with that

3 Simulations

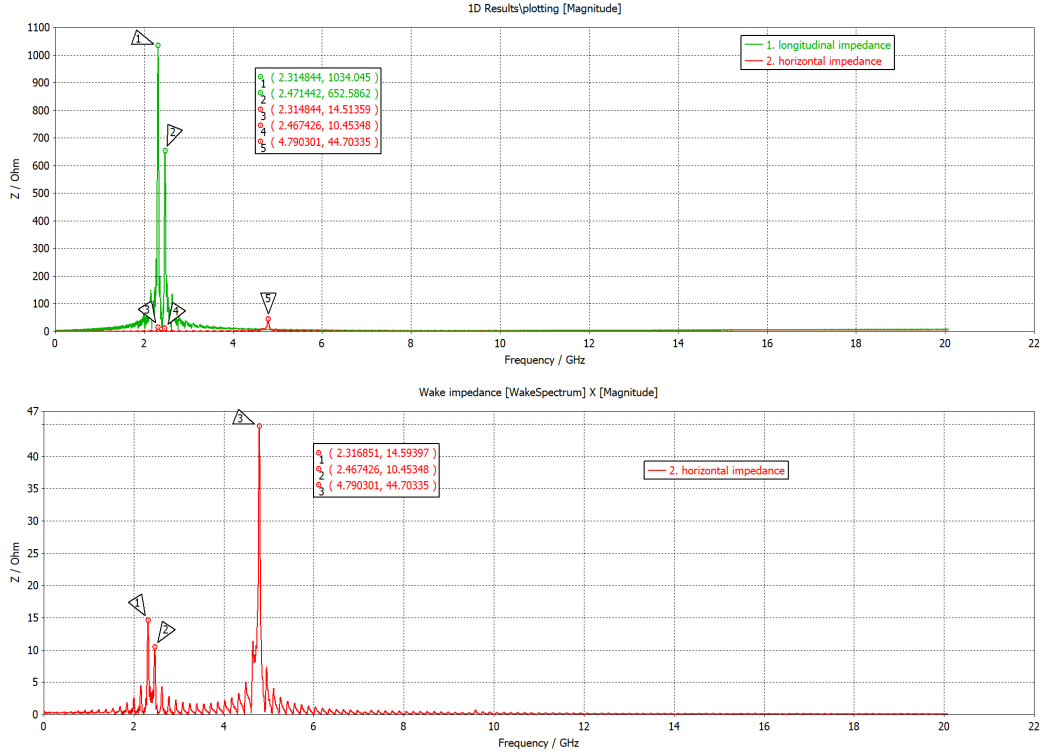


Figure 3.27: Impedance of shielding foils with sinusoidal slit deviation. Resonant frequencies are marked. The top compares the horizontal and longitudinal impedance, the bottom shows the horizontal impedance alone.

sinusoidal deviation in the slit. The longitudinal impedance does not show any resonance near the 4.77 GHz frequency. As noted above, there is only a double resonance peak at about half that frequency. The horizontal impedance also shows the double peak at about half the expected frequency. But there is also a higher resonant peak at 4.79 GHz which closely matches the expected frequency from the sinusoidal deviation.

Figure 3.28 shows the surface current for the sinusoidal slit. In the longitudinal direction the surface current changes polarity every full period length of the sine function. This means the current alternates with half the frequency of the sine function, which matches the resonant frequency peaks seen in the impedance.

The surface current in the horizontal direction is less obvious. There are more polarity changes and the regions of similar polarity have different sizes. The picture resembles a beat pattern of two sine functions where the second function has double the frequency of the first. Figure 3.29 shows such a beat pattern for $f(x) = 7.8 \sin(4.77x) + 5.2 \sin(2.36x)$ where the amplitudes and frequencies are taken from the resonance peaks in the horizontal impedance. This beat pattern

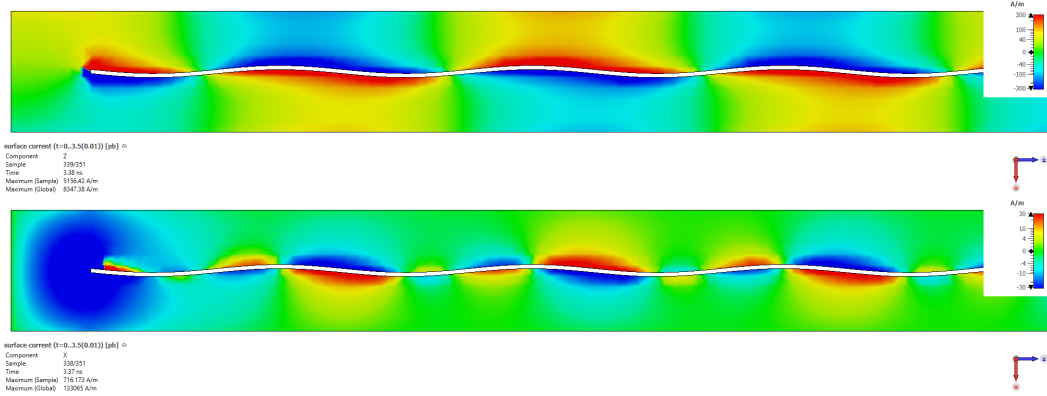


Figure 3.28: Surface current of shielding foils with sinusoidal slit deviation. Currents in longitudinal direction at the top and horizontal direction at the bottom.

matches the changes in polarity for the surface currents in X-direction.

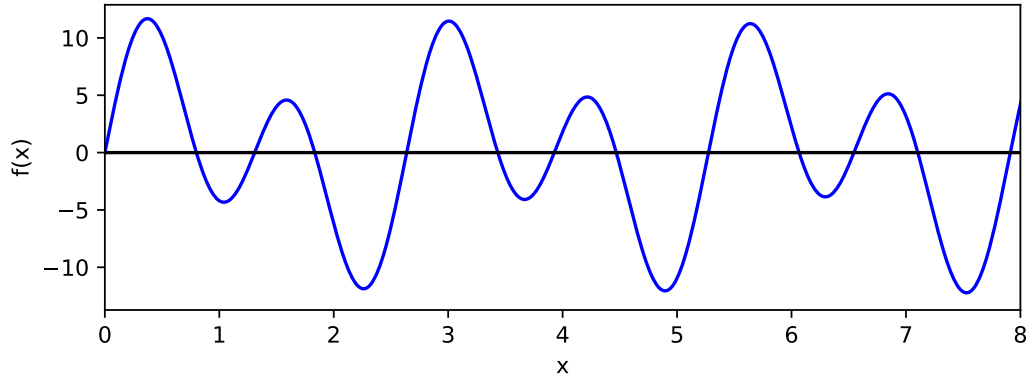


Figure 3.29: Sine function beat pattern. $f(x) = 7.8 \sin(4.77x) + 5.2 \sin(2.36x)$.

If we imagine the shielding foils as the boundaries of a rectangular waveguide where the sinusoidal deviation of the slit act as perturbations that define a cavity, we can use the dimensions to calculate the frequencies of the resonant modes. As a rough estimation we assume a rectangular cavity so that

$$f_{nml} = \frac{c}{2} \sqrt{\left(\frac{n}{x}\right)^2 + \left(\frac{m}{y}\right)^2 + \left(\frac{l}{z}\right)^2} \quad (3.2)$$

gives the resonant frequency for the nml -th mode. Here c is the speed of light and x, y, z are the respective dimensions of the cavity. If we set $z = 62.8$ mm, the period length of the sinusoidal deviation, we get $f_{001} = 2.39$ GHz. This matches the resonance in the longitudinal impedance very well.

3 Simulations

Setting $x = 30$ mm, the modeled width of the foil, yields $f_{100} = 5$ GHz, which is close to the 4.77 GHz resonance we see in the horizontal impedance.

3.3.3.1 Amplitude and Oscillation Period

To investigate the correlation with the longitudinal impedance, several different periods and amplitudes of the sinusoidal deviation are simulated. The horizontal sinusoidal deviation is described by

$$x(t) = A \sin(2\pi \frac{t}{P}) \quad (3.3)$$

with x being the horizontal deviation in mm from the center line, A is the Amplitude in mm and P the period length in mm. The width of the slit is kept constant at 1 mm.

First the period length is varied while keeping the amplitude at 1 mm. Figure 3.30 shows the longitudinal impedance of for different period lengths.

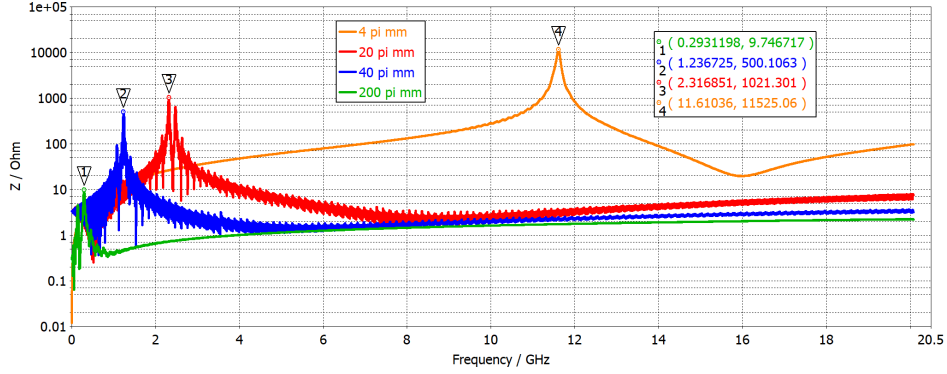


Figure 3.30: Longitudinal impedance of sinusoidal slit variations of varying period length.

The period length clearly affects the resonant frequency of the sinusoidal deviation. Shorter period length result in higher resonant frequencies. Table 3.2 compares the simulated resonant frequencies to the cavity model described by Eq. 3.2. The simulated resonant frequencies closely match the 001-mode of a rectangular cavity of the period length.

Table 3.2: Resonant frequencies in longitudinal impedance for sinusoidal slit deviations of different period lengths. Simulated results compared to cavity model of Eq. 3.2.

Period Length [mm]	4π	20π	40π	200π
resonant frequency [GHz]	11.61	2.32	1.24	0.29
cavity model [GHz]	11.93	2.39	1.19	0.24

The amplitude of the peak impedance also increases with frequency. The deviation with a period length of 4π mm shows a significantly higher longitudinal impedance across the whole frequency range. An explanation for the increased impedance could be the FWHM-length of 11.98 mm of the simulated test beam, which approaches $4\pi \approx 12.57$ mm.

To investigate the impedance effect of different amplitudes, the period length is kept at 20π mm. Figure 3.31 shows the longitudinal impedance for sinusoidal slit deviations of different amplitudes. The resonant peaks show the same frequency for all different amplitudes. This reinforces the direct correlation between the period length and the resonant frequency.

Amplitudes of 0.5 mm and above show a lot of noise and drastically higher longitudinal impedance. For amplitudes larger than 0.5 mm, the slit in the shielding foil crosses over the beam trajectory transversely. Please note, that this is just an academic study as the width of the overall foil slit will be limited to 1 mm by the magnet assembly of IVUE32, preventing large amplitudes of slit geometry deviations.

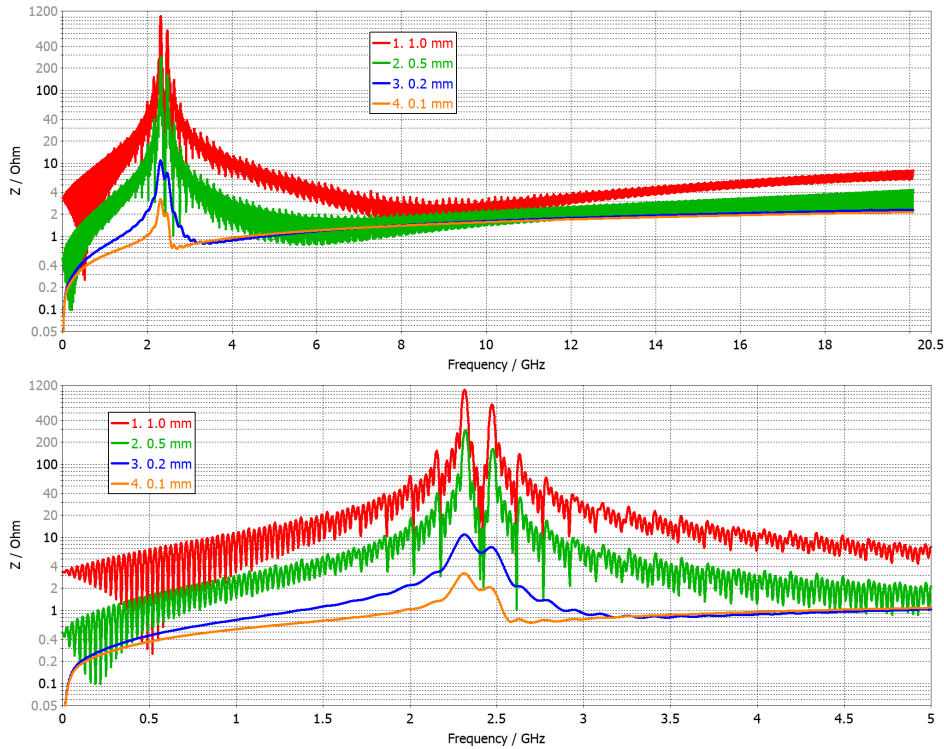


Figure 3.31: Longitudinal impedance of sinusoidal slit variations of varying amplitudes. Full frequency range at the top, low frequency excerpt at the bottom.

Figure 3.32 compares the longitudinal impedance of low amplitude slit deviations to that of a perfectly straight slit. An amplitude of 0.1 mm has a resonant peak

3 Simulations

impedance of only $3\ \Omega$, which would be manageable. The expected amplitude of any deviation is much smaller than 0.1 mm. The simulated deviations maintain perfect periodicity over the length of 960 mm, which is very unlikely for any real manufacturing tolerances. Therefore, the longitudinal slit in the shielding foil should not make any problematic contribution to the longitudinal impedance of IVUE32.

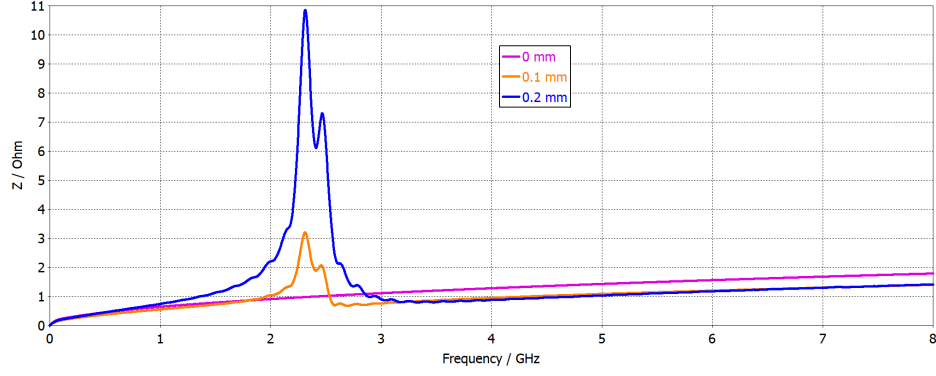


Figure 3.32: Longitudinal impedance of sinusoidal slit variations of varying amplitudes. Low amplitudes are compared to a perfect straight slit.

3.3.4 Symmetric Slit Deviation

To investigate the effects of a symmetric sinusoidal slit deviation, introduced in Sec. 3.3.1, several amplitudes of deviation are simulated. As with the asymmetric deviations, discussed in Sec. 3.3.3, the symmetric horizontal deviations are modeled after Eq. 3.3. However, the deviation is mirrored along the longitudinal plane, varying the width of the slit, as shown in Fig. 3.25.

Figure 3.33 shows the longitudinal impedance of symmetric deviations of different amplitudes compared with a perfectly straight and the impedance of two parallel plates.

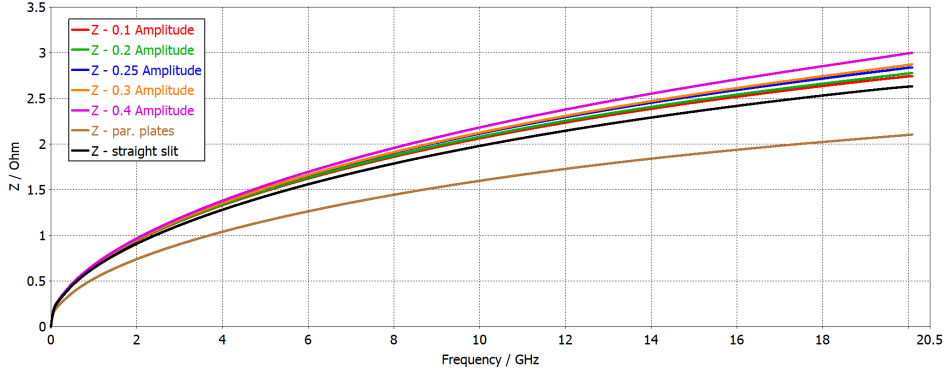


Figure 3.33: Longitudinal impedance of symmetric sinusoidal slit variations of varying amplitudes compared to a straight slit and parallel plates.

The longitudinal impedance does not show any resonances over the whole frequency spectrum. Symmetric slit deviations show a longitudinal impedance similar to that of a perfectly straight slot or two parallel plates. These deviations will not significantly increase the overall impedance of IVUE32.

Figure 3.34 compares the longitudinal impedance of a symmetric deviation with that of an asymmetric deviation, discussed in Sec. 3.3.3. The asymmetric deviation shows a clear resonance and much higher overall impedance.

The surface current visualizes the impedance resonance. The passing charged particle beam induces a mirror charge in the shielding foils, that travels alongside the bunch as a surface current. Figure 3.35 visualizes the surface current density for the two different types of deviations. The surface current can travel along the symmetric deviations unimpeded, while the asymmetric deviation capture a part of the beams energy in the form of an oscillating current behind the passing beam.

If it is possible to influence the type of deviation created during the bending process of the shielding foil, symmetric deviations are much less problematic.

3.3.5 Power Loss

The CST wakefield solver offers a surface current monitor which shows the current density on the surface of the simulated model. The resistance of the copper foils is

3 Simulations

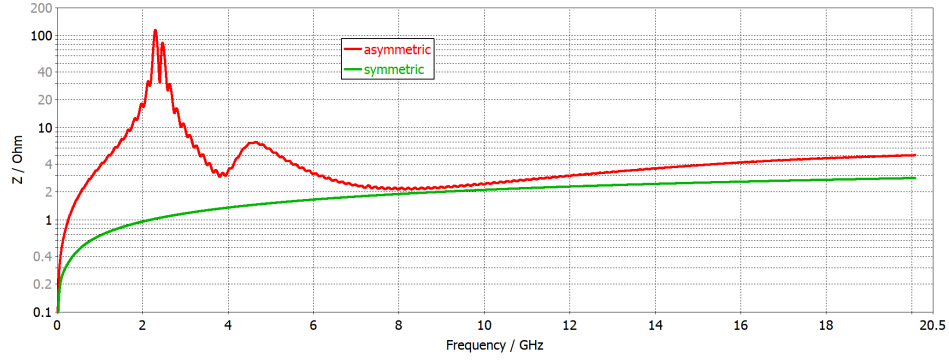


Figure 3.34: Longitudinal impedance of symmetric sinusoidal slit variation with an amplitude of 0.25 mm compared to asymmetric variation an amplitude of 1 mm.

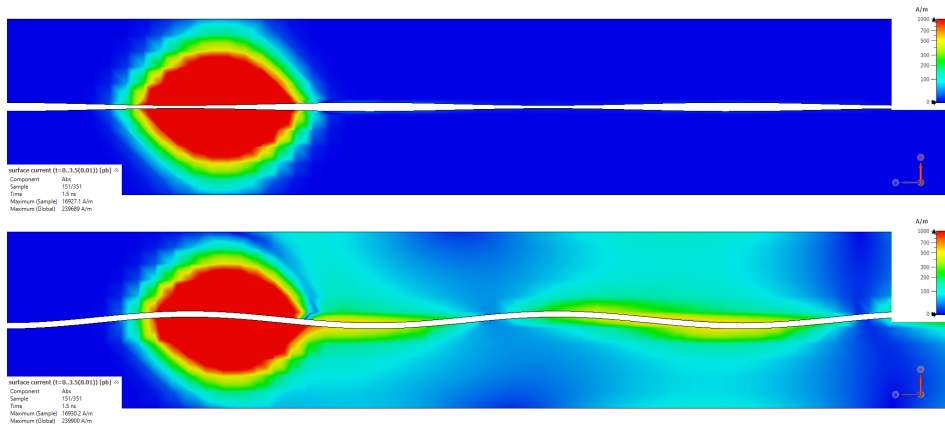


Figure 3.35: Surface current density for sinusoidal slit variations. Symmetric variation at the top, asymmetric variation at the bottom.

the only source of power loss. Therefore the surface current can be directly related to the power loss

$$P = RI^2 \quad (3.4)$$

with the help of Ohm's law. The power loss is proportional to the square of the current flowing through the shielding foil. The surface current shown in Fig. 3.36 shows where the power loss is concentrated on the shielding foil. The highest current density is close to the passing electron beam. This means the highest current density can be found right where the shielding foils are folded into the slit between the magnet girders. Figure 3.36 also shows that the surface current does not extend far into the slit.

The power loss can be directly translated into heat load on the shielding foils as there are no other means of power dissipation.

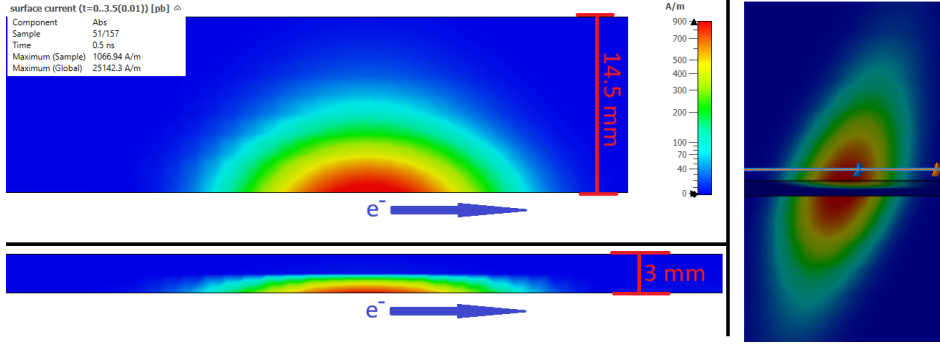


Figure 3.36: Surface current density as the beam passes the split shielding foils. The horizontal distribution is shown at the top and the vertical at the bottom. The right shows the 3D slit model with the surface current.

To quantify the total heat load on the IVUE32 shielding foils, a section with a perfect longitudinal slit was simulated for each of the different bunches found in the BESSY II fill pattern. The power losses for a 30 cm section of one folded shielding foil are shown in Fig. 3.37. Once the beam has fully entered the structure, the power loss is constant for the duration, the beam passes the structure. Higher current bunches, such as the single or PPRE bunch, induce stronger surface currents and therefore greater power loss. The slight offset on the time axis is due to the different bunch length, which affects the starting time of the simulation. The time needed to pass the simulated section of foil, is the same for every bunch.

To calculate the total heat load on the IVUE32 shielding foils, the power loss of each bunch is integrated over time to gain the total energy E_b each bunch deposits into the structure. It is then multiplied by the amount of this type of bunch in the BESSY II fill pattern n_b and the revolution frequency f_{rev} . Finally, the full length of IVUE32, 2.5 m, compared to the length of the simulated section is taken into account to yield

$$P_{tot} = 4E_b \cdot n_b \cdot f_{rev} \frac{l_{IVUE32}}{l_{sec}} \quad (3.5)$$

the total heat load on the shielding foils. The factor 4 takes the number of shielding foils into account. For the BESSY II standard user fill pattern shown in Fig. 1.1 the shielding foils will experience a heat load of 16.1 W.

In the context of an enclosed vacuum environment 16.1 W of heating are not negligible. The only possible method to dissipate this heat is to transmit it through the undulator structure. In this case the permanent magnets and their keepers act as the heat sink for the shielding foils. Compared to CPMU17 with ample cryogenic cooling in the magnet support structure, IVUE32 is planned without active cooling in the magnet keepers. Heat load simulations of the magnet support structure will answer the question, whether additional cooling is required.

3 Simulations

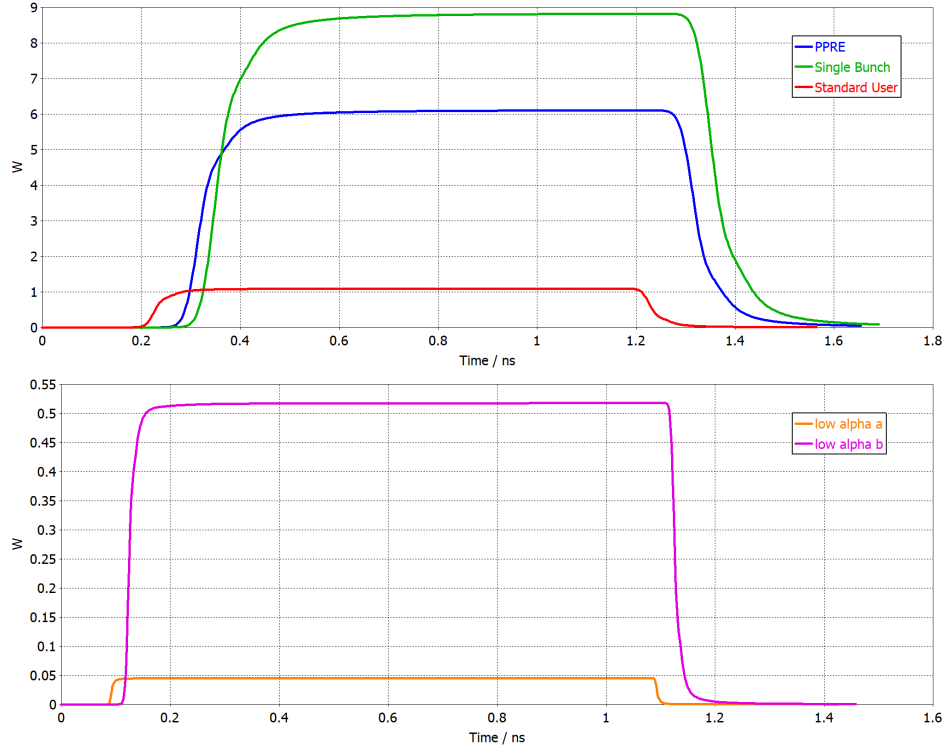


Figure 3.37: Power loss of a 30 cm section of one folded shielding foil for different BESSY II beam parameters. Standard user fill pattern at the top, low- α beams at the bottom.

Another concern is the concentration of the heat load on the folded edge of the shielding foil. This is not only the furthest part of the foil from any possible cooling infrastructure but also the most likely part to lose contact with the magnets due to the sharp bending radius. Direct contact with the magnets as a heat sink is vital for the copper shielding foil with a thickness in the range of a few tens of μm . Correctly seating the folded foil on the magnets and testing the heat load can be done in a vacuum chamber test stand with a model of the magnet assembly.

3.4 Goubau Line Test Stand

The Goubau-line test stand was modeled and simulated using CST studio [31]. The mounting plate, feed in, cone antenna, and wire were modeled to the exact parameters of the real test stand. Figure 3.38 shows the modeled test stand with parallel slit foils as a DUT. The width of the slit is 0.5 mm and the gap between the foils is 10 mm. The CST Time Domain Solver was used with a frequency range from 0 GHz to 20 GHz. The second metal plate at the end of the Goubau-line was omitted in the simulations and replaced by an open boundary condition. Magnetic symmetries in the XZ and YZ plane were used, where permitted by the geometry of the DUT, to reduce simulation time. The global mesh was kept fairly coarse at 5 cells per wavelength with higher mesh resolutions around key geometric features like the wire, the antenna feed in, and the foil slit. The simulation results were compared to the actual measurements. Figure 3.39 shows the agreement between the simulated TDR signal and the NWA measurement of an empty test stand without any DUT mounted. The graph in Fig. 3.39 shows a decaying oscillation

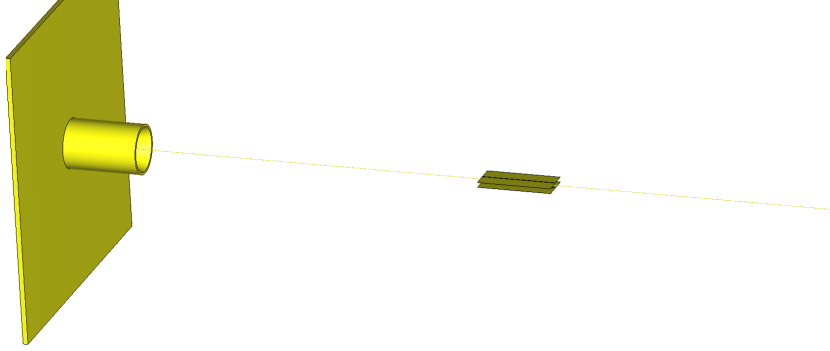


Figure 3.38: Goubau-line test stand modeled in CST studio with parallel slit foils as a DUT. The width of the slit is 0.5 mm and the gap between the foils is 10 mm

from about 2 ns onward with a period length of about 3 ns. Field simulations show the internal reflections of the horn antenna to be the source of these oscillations. A part of the incoming pulse gets reflected at the opening of the horn antenna and travels back and fourth in the horn. Figure 3.40 shows the TDR signal of the empty test stand together with the port signals. The output port signal shows periodic peaks after 3 ns that correlate with the impedance changes. The impedance for TDR is defined as

$$Z(t) = Z_0 \frac{\int_0^\infty i(\tau) d\tau + \int_0^t o(\tau) d\tau}{\int_0^\infty i(\tau) d\tau - \int_0^t o(\tau) d\tau} \quad (3.6)$$

with $i(\tau)$ the input and $o(\tau)$ output port signal, and $Z_0 = 50 \Omega$ the reference impedance of the port.

The Goubau-line is simulated without any DUT so there are no sources of reflections

3 Simulations

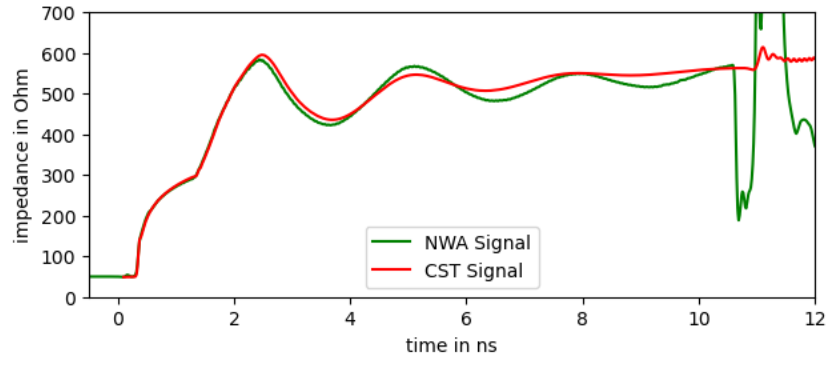


Figure 3.39: TDR signal of the Goubau-line test stand without DUT. Simulated signal in red compared to the NWA measurement in green.

past the end of the horn antenna. Therefore, multiple internal reflections inside the horn antenna are the only possible cause of the changing output port signal.

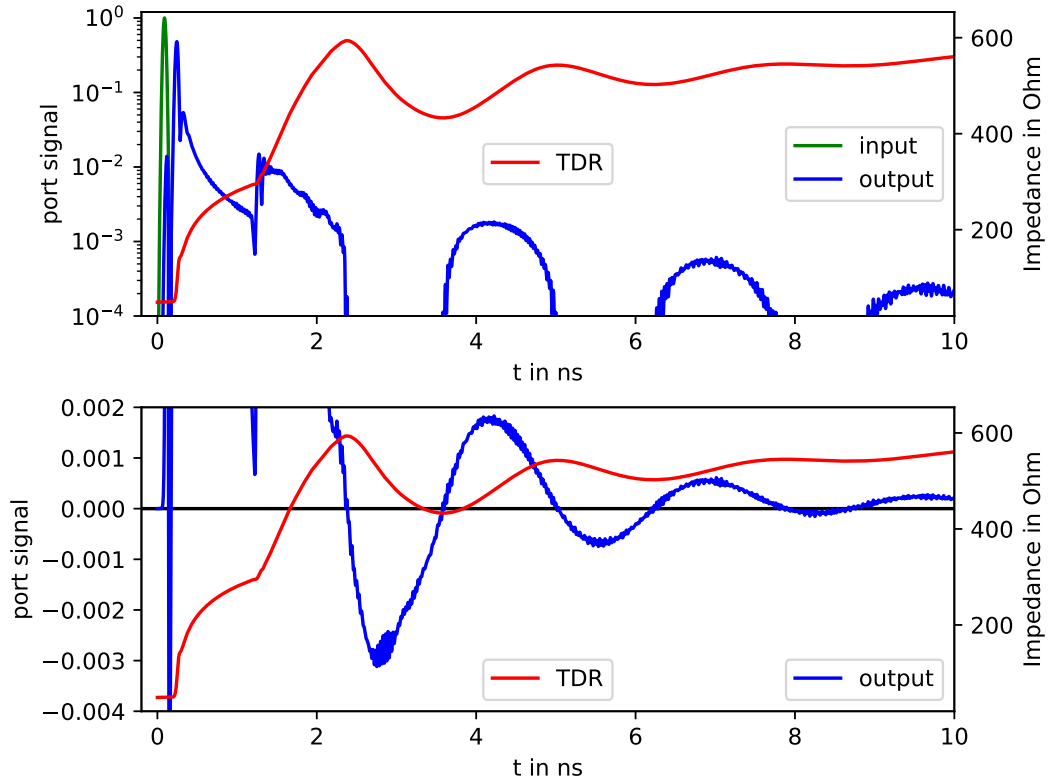


Figure 3.40: TDR signal of the Goubau-line test stand without DUT with the in- and output port signals shown. Logarithmic scaling of the port signals on top and zoomed view of output signal at the bottom.

This is a concern for time gating, as the time coordinate of the multiple reflections inside the horn overlap with the response of the DUT. Future improvements to the horn antenna aim to minimize this effect. Some simulations in that regard are presented in Sec. 7.5.

An alternative to time gating would be to subtract the signal of the empty test stand as a reference from the signal with a DUT mounted. This would allow for a qualitative comparison of impedance of different structures. Section 5.1.1 explores this in greater detail.

4 Measurement Setup

The structure and materials of the IVUE32 components are very complex and therefore cannot easily be simulated in full detail. This makes it necessary to measure the impedance of individual components. The Goubau line test stand presented in this chapter aims to determine which aspects of the complex assemblies need to be included in the simulations and which can be omitted, to reduce the computational cost of simulating the entire structure for different operation modes. Thus, the measurements are needed to validate the undulator model used simulations.

4.1 Goubau Line Test Stand

A Goubau line is a single wire transmission line developed in 1950 for long range communications [27]. As explained in chapter 2 the transverse electric field around the transmission line resembles that of a charged particle beam out to a certain distance. For that reason, the Goubau line was tested as a way to measure the impedance of accelerator components in recent years [29, 35]. At the Argonne Advanced Photon Source a Goubau line test stand was used to measure the coupling impedance of different vacuum chamber components, including a beam position monitor assembly. The test stand was able to detect an insufficient contact in one of their rf-gasket designs [29]. Bergoz Instrumentation in France also investigated the feasibility of a Goubau line test stand to measure the impedance of their products [35].

Figure 4.1 shows a photo of the Goubau line test stand developed in the context of the presented work. The frame of the test stand is made up of wood polypropylene

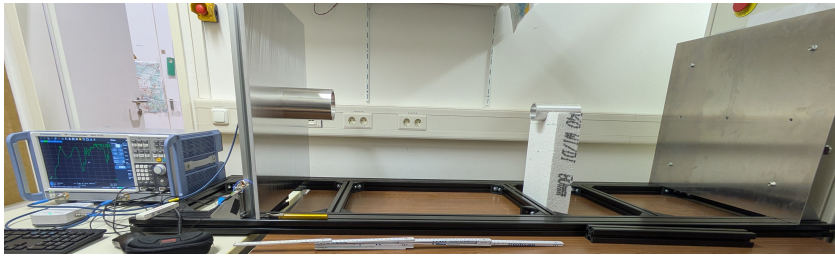


Figure 4.1: Photograph of Goubau line test stand with an impedance step structure set on expanded polystyrene (EPS) support next to the wire as device under test (DUT).

composite profiles. This material was chosen to limit conductive components in

4 Measurement Setup

proximity of the wire. Compared to aluminum profiles of the same geometry, these composite profiles affect the measurements significantly less. A study of different support materials and their effect on the measurement is included in the Appendix 7.1. The horn antenna uses a simple conical geometry and is turned from aluminum. A detailed description of the horn antenna can be found in section 4.1.5. The horn is mounted on a 10 mm thick aluminum plate as shown in Fig. 4.2. The plate is a square with 60 cm long sides. An enameled copper wire of 0.75 mm diameter



Figure 4.2: Horn antenna mounted on aluminum base plate.

is soldered to the inner conductor of the panel mount connector as shown in Fig. 4.3. Further parameters of the wire are discussed in Section 4.1.2. The wire is

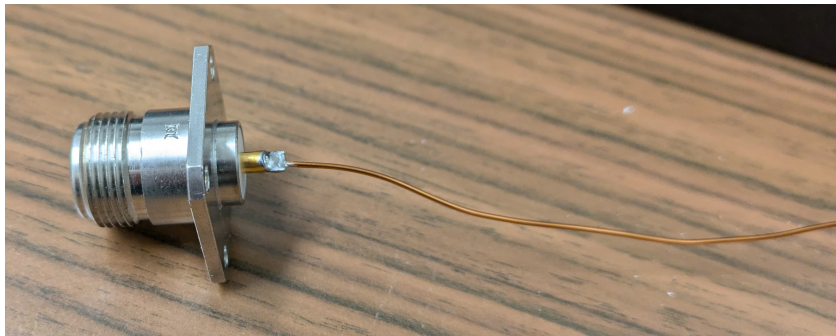


Figure 4.3: N-Type panel mount connector with 0.75 mm diameter enameled wire soldered to the inner conductor.

stretched for 1.55 m and goes through a hole in the middle of a second square metal plate. The wire is mounted on the back side of the second plate with a 3D printed tensioner as shown in Fig. 4.4. The second plate constitutes a significant change in impedance so the wire is just coiled around the tensioner without any further termination.

The measurement can be time gated so that impedance changes at the end of the test stand can be ignored and don't affect the measurement of the device under test. The measurement device for the test stand is a *Rohde und Schwarz* ZNB

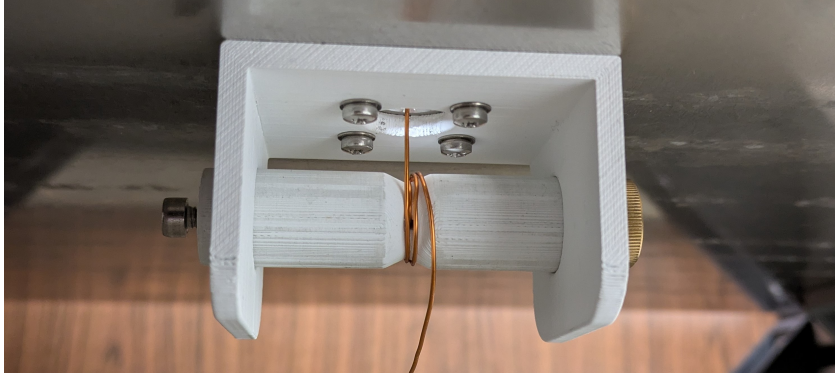


Figure 4.4: 3D printed wire tensioner mounted on the back plate of the Goubau line test stand.

vector network analyzer with software packages for time domain analysis. It is connected to the Goubau line via a 1 m long PC 3.5 coaxial cable. This coaxial cable is connected to the N-Type feed in via an adapter.

4.1.1 Network Analyzer

The set up uses a *Rohde und Schwarz* ZNB vector network analyzer (VNA) with a frequency range from 100 kHz to 43.5 GHz. It has two ports with male 2.92 mm connectors. In order to connect the PC 3.5 coaxial cables, which also have male ends at each side, adapters are used at the ports of the VNA as shown in Fig. 4.5. An automatic calibration unit ZN-Z54 is used to calibrate the connections to the test stand. The calibration unit is characterized for the full frequency range of the VNA. The VNA comes with software packages to enable time domain analysis, specifically time domain reflectometry (TDR). Even though the VNA is more capable, the top end of the frequency range is set to 20 GHz for the measurements. The shortest bunches in the BESSY II operating modes only have a bandwidth up to 20 GHz. Analyzing higher frequencies at the test stand would be counter productive as it might show problems that wouldn't exist in the accelerator.

The VNA is set up as shown in Fig. 4.6 with a 100001 point frequency sweep for maximum resolution. This results in a sweep time of 21.7 s. The frequency range is set from 199.998 kHz to 20 GHz resulting in a 46.25 ps, 10 % to 90 % rise time for the time domain measurements. After setting the resolution and frequency range, the VNA is calibrated using the ZN-Z54 automatic calibration unit. The measurement is set to TDR, obtaining the change of impedance from the S_{11} parameter. The S_{11} parameter measures the reflected signal from the DUT as a function of frequency. The impedance Z in the frequency domain is given by

$$Z = Z_0 \frac{1 + S_{11}}{1 - S_{11}} \quad (4.1)$$

4 Measurement Setup



Figure 4.5: Photo of the *Rohde und Schwarz* ZNB vector network analyzer used for the Goubau line test stand.

with a reference impedance $Z_0 = 50 \Omega$. The ZNB-K2 software package installed on our ZNB VNA is used to Fourier transform the result into an impedance change over time.

The TDR measurement is set to use a low pass step pulse and the impulse response is transformed to the time domain using the normal profile. The TDR curve is displayed from -1 ns to 12 ns with respect to the reference plane set by the calibration process. The reference plane is located at the rf-feed in of the test stand. The data past 12 ns is not usable as it would originate from behind the back plate of the test stand.

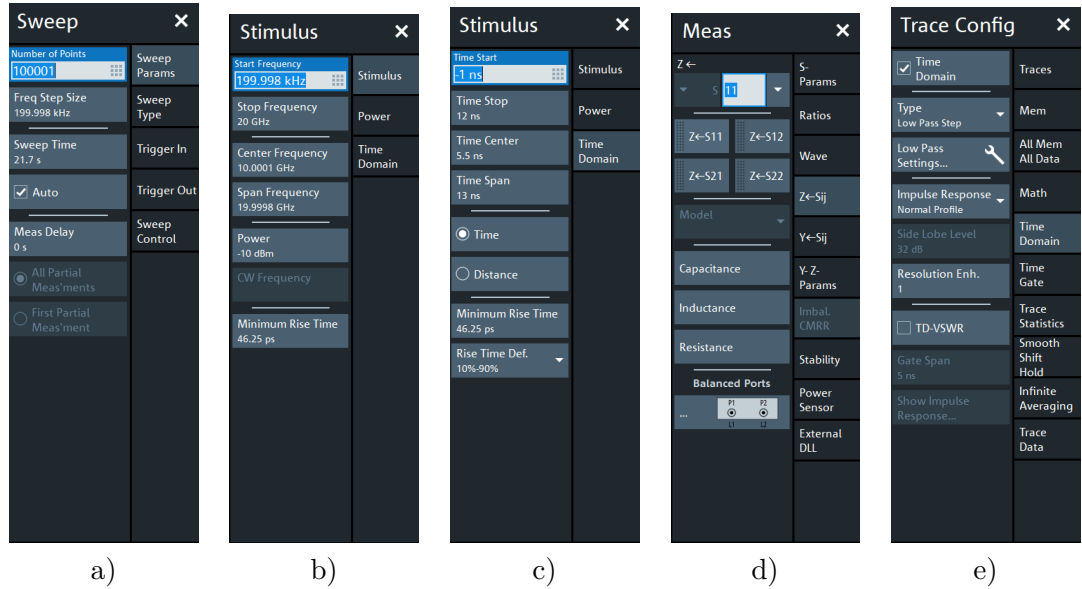


Figure 4.6: Screenshots of ZNB settings

4.1.2 Wire Properties

In order to cover the maximum aperture of the IVUE32 taper structure with a sufficient margin, a field extension of 30 mm around the Goubau line wire is required. The field needs to reach that extension for frequencies up to 20 GHz.

A python script is used to numerically solve equation 2.35 to determine the propagation constant h for the electric field. Subsequently equation 2.32 can be evaluated to yield the field extension for any frequency.

There are several variables that can be adjusted to increase the field extension. The diameter of the conductor is one of them. A larger diameter will also extend the field further. For a practical reasons a maximum 2 mm was chosen to maintain some flexibility of the wire. Readily available magnet wire was chosen as a starting point.

The test stand prototype uses a 0.75 mm diameter enameled copper wire. Figure 4.7 shows the spool of wire used for the test stand. This wire was chosen because it was available and still flexible enough to be pulled taut by the tensioner. Thicker wires were tested but the tension required to straighten out all the small bends was too high. The center conductor of the panel mount connector was pulled out before the wire could be properly tensioned.



Figure 4.7: Photo of enameled copper wire used in the Goubau line test stand.

According to the manufacturer the wire features a *Grade 1* insulation and complies with the IEC 60317-20 standard. In the data sheet, included in the Appendix 7.2, the insulating material is stated as modified polyurethane. According to the IEC 60317-20 standard the minimum increase in diameter due to the insulation is 0.030 mm while the maximum final outside diameter of a 0.75 mm conductor wire is set at 0.805 mm. This means the insulation layer has a thickness between 0.015 mm and 0.0275 mm. Unfortunately the manufacturer does not provide any further detail on the type of modified polyurethane insulation nor any information on the

4 Measurement Setup

electric permittivity of the insulating material.

Beverte investigates the permittivity of polyurethane foams [36] and gives a value of $\epsilon_r = 3.35$ for monolithic polyurethane. Figure 4.8 shows the field extension of a Goubau line using a 0.75 mm diameter wire with the specified insulation at a frequency of 20 GHz. The red plot shows the field extension for the minimum insulation thickness according to the IEC standard, while the blue graph represents the upper limit of insulation thickness. Both are compared to a $1/r$ gradient, shown in black. The dotted lines represent the deviation of the field amplitude from the gradient as a percentage value. The vertical dotted line marks the distance at which that deviation reaches 5 %. This 5 % threshold is reached at 4.7 mm from the wire for the upper limit of insulation and at 6.6 mm for the lower limit. Both cases fall significantly short of the 30 mm target, making this wire unsuitable for high frequency measurements.

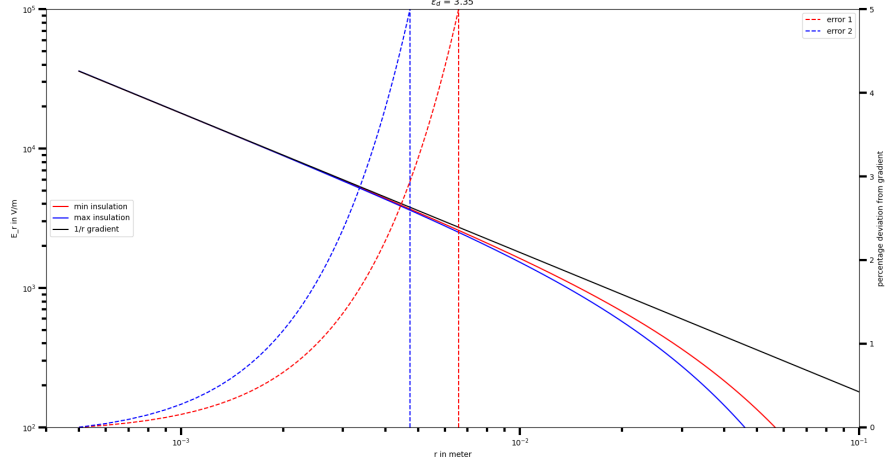


Figure 4.8: Transverse extension of electric field at 20 GHz around a 0.75 mm diameter Goubau line wire coated with polyurethane. The electric permittivity is assumed to be $\epsilon_r = 3.35$ for monolithic polyurethane.

Qian *et al.* investigate the electric properties of polyurethane elastomers containing modified silica [37]. They show the frequency dependent dielectric constant for several different polyurethane composites for frequencies up to 1 MHz. The values for most composites seem to converge on $\epsilon_r = 5$ at the high frequency limit. Figure 4.9 shows the field extension for a 0.75 mm diameter wire insulated with a material of $\epsilon_r = 5$ electrical permittivity at a frequency of 20 GHz. Here the 5 % deviation threshold is reached at 6.1 mm for the minimum insulation thickness and at 4.4 mm for the upper limit of insulation thickness.

A higher electric permittivity further reduces the field extension. In order to extend the electric field, a more suitable dielectric coating needs to be found. There are

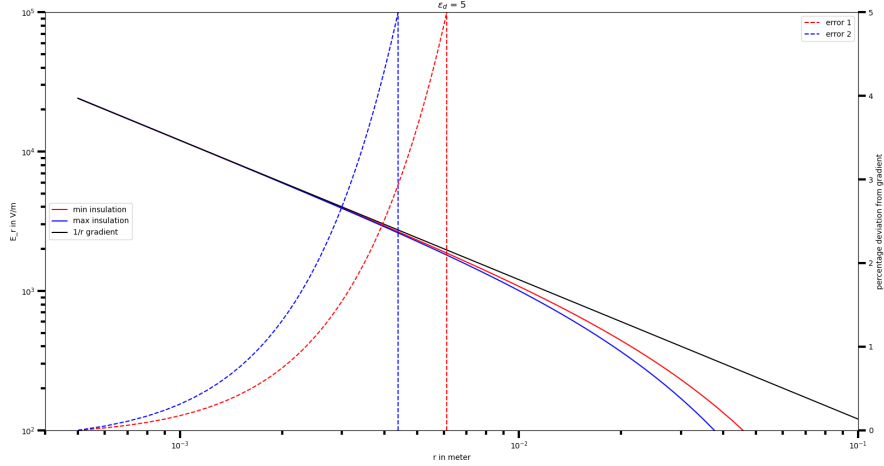


Figure 4.9: Transverse extension of electric field at 20 GHz around a 0.75 mm diameter Goubau line wire coated with polyurethane. The electric permittivity is assumed to be $\epsilon_r = 5$ for modified polyurethane.

a wide range of dielectric coatings available on the market. A study is conducted with a 1 mm diameter copper wire with a coating of varying thickness and a dielectric constant of $\epsilon_r = 4$. Figure 4.10 shows that for a coating thickness of less than

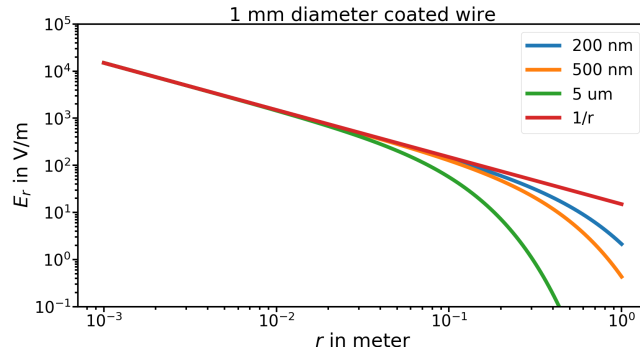


Figure 4.10: Amplitude of transverse electric field of a Goubau line using a 1 mm diameter copper wire with several thicknesses of $\epsilon_r = 4$ dielectric coating at 20 GHz compared to a $1/r$ gradient.

500 nm the electric field amplitude follows a $1/r$ gradient out to at least 30 mm. Such a wire would fulfill our requirements.

4.1.3 Impedance Step Models

To test the capability of the Goubau line test stand a cylindrical impedance step was created. It is a 15 cm long aluminum cylinder with different diameters bored out. It has three sections with a length of 5 cm each. The sections at either end have an inner diameter of 50 mm. The middle section has a smaller diameter of only 20 mm. The change in diameter happens in a single step. A schematic of the cross section is shown in Fig. 4.11.

When placed on the Goubau line test stand it forms a coaxial wave guide together with the Goubau line wire. The characteristic impedance of coaxial wave guides can be calculated analytically,

$$Z_0 = \frac{60}{\sqrt{\epsilon_r}} \ln \frac{d_2}{d_1} \quad (4.2)$$

depending only on the diameter d_1 of the inner conductor, d_2 of the outer conductor and the permittivity ϵ_r of the dielectric material between them.

With an inner conductor diameter of 0.75 mm, the impedance at either end of the structure is 252Ω , stepping down to 197Ω in the center section. This can be used to benchmark the TDR measurement of the Goubau line test stand.

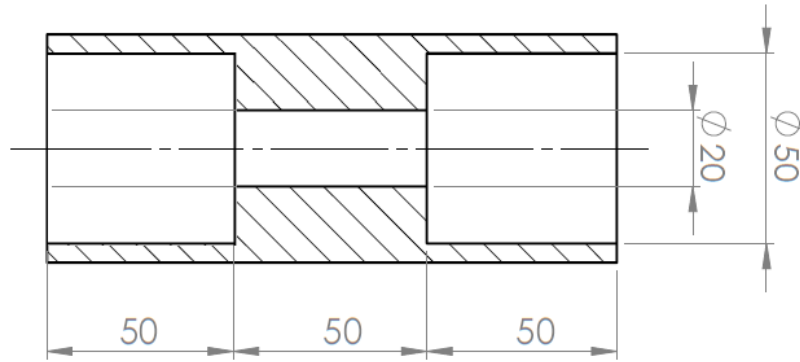


Figure 4.11: Schematic of the cross section of an impedance step test structure.

The results of the TDR measurements on the Goubau line test stand are shown in chapter 5.

4.1.4 Foil Taper Models

The shielding foils of IVUE32 will have to allow not only for gap movements but also for longitudinal shifts of the magnet rows. This calls for a design with a longitudinal slit in the shielding foils. Simple taper models were made to investigate the impact of a longitudinal slit in the shielding foils. The models are made up of 15 cm long, 6 cm wide copper foils with a constant gap of 10 mm or a tapered gap going from 10 mm to 20 mm, both with and without a longitudinal slit of 0.5 mm width. These foils are glued to a support structure made from expanded polystyrene (EPS) foam using double sided tape. Several materials were tested for their impact on the TDR signal, EPS was found to have minimal effect on the signal while still providing sufficient structural rigidity. A summary of these test measurements is included in the appendix 7.1. Figure 4.12 shows one of these taper models on the Goubau line test stand.

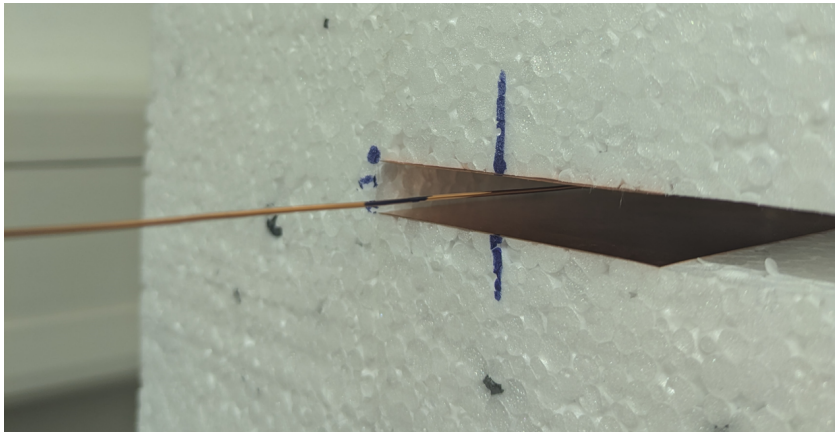


Figure 4.12: Photograph of a foil taper model placed on the Goubau line wire test stand.

The foil taper models are only supported on one side, leaving the other side open. This allows for the Foam models to be changed and repositioned without having to untension and reconnect the Goubau line wire. This is a key quality for reproducibility of measurements and also saves the wire. Repeated winding around the wire tensioner can cause metal fatigue in the copper wire, leading to breaks. The measurements of the foil taper models are shown in chapter 5.

Due to the positional accuracy limitations of the EPS material these taper models are very simplistic. More detailed taper models that closer reflect the actual taper design of IVUE32 will be 3D printed.

4.1.5 Transmitter and Receiver

The TEM mode feeding in from the coaxial cable needs to be translated to the TE mode surface wave, traveling along the Goubau line wire. The test stand prototype uses a conical horn antenna for that task. Figure 4.13 shows a schematic of the cross section. The inner diameter starts out at 7.5 mm and then tapers uniformly to 90 mm over a distance of 136.5 mm. The smaller inner diameter is bored out to accept an N-Type panel mount connector as the rf-feed in. The N-Type connector with the wire soldered to it, as shown in Fig. 4.3, is attached to the back of the horn antenna. The outer conductor of the panel mount connects to the body of the horn antenna.

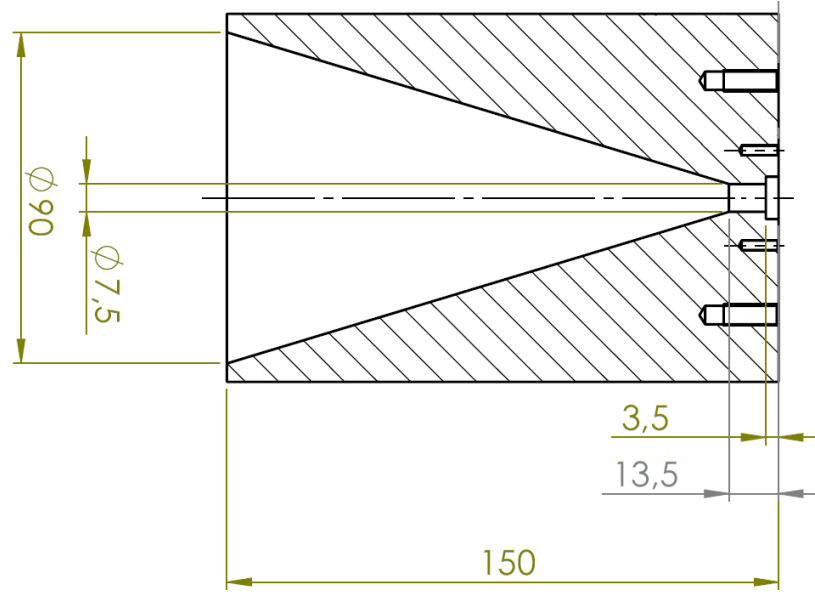


Figure 4.13: Schematic of the cross section of the horn antenna.

This assembly represents a coaxial line with increasing outer conductor diameter, also known as a transmission line taper. The goal is to smoothly match the characteristic impedance of the Goubau line wire to the $50\ \Omega$ impedance of the coaxial feed in. As stated in chapter 2, the characteristic impedance of a Goubau line is frequency dependent. Therefore it is only possible to properly match the impedance of a Goubau line to the feed in for a single frequency. Since the TDR measurement uses a sharp step pulse in the time domain or a broad frequency sweep, parts of the spectrum will always be mismatched. An impedance mismatch on a transmission line is a source of reflections. Ideally one would only record the reflections from the DUT for a TDR measurement. Unfortunately there is no practical way to establish a reference plane on the Goubau line after the horn antenna. There is however a causality argument for distinguishing the origin of reflections by examining the time at which they are recorded. Reflections from the feed in can be disregarded by

time gating the measurement. The only variable that can be used for time gating is the time at which the signal is recorded by the VNA. Reflections from different origins that are recorded at the same time cannot be distinguished that way. This means that the impedance mismatch of the feed in is only problematic if it causes multiple internal reflections that overlap the response of the DUT.

The TDR measurements in chapter 5 show a significant variation of impedance over time for the empty test stand. The copper wire and its insulation are manufactured to a standard that does not allow for variations of this magnitude. The impedance should be constant along the Goubau line wire. Simulations of the test stand presented in chapter 3 confirm, that the cause of these variations are multiple internal reflections at the horn antenna.

This means that the simple geometry of the horn antenna of the test stand prototype is not well suited for TDR measurements. Several other geometries have been simulated to find an improved design for future iterations of the test stand. This includes exponential horn profiles as well as corrugated designs, commonly used in satellite communication equipment [38]. These simulation studies are presented in Sec. 7.5.

5 Measurement Results

5.1 Impedance Step Model

The impedance step, described in Sec. 4.1.3 was placed on the Goubau line test stand 815 mm from the reference plane. Figure 5.1 shows the TDR measurement of the DUT compared to the empty test stand as a reference. The reflection signal of the DUT is registered at 5.7 ns, marking the beginning of the DUT. At that point a sharp decrease in impedance from $570\ \Omega$ to $530\ \Omega$ is visible. At 6 ns another sharp decrease down to $340\ \Omega$ is visible. The impedance oscillates until it reaches $450\ \Omega$ at 6.4 ns followed by a sharp increase up to a peak of $640\ \Omega$. After that the impedance oscillates with decreasing amplitude around the reference value of the empty test stand.

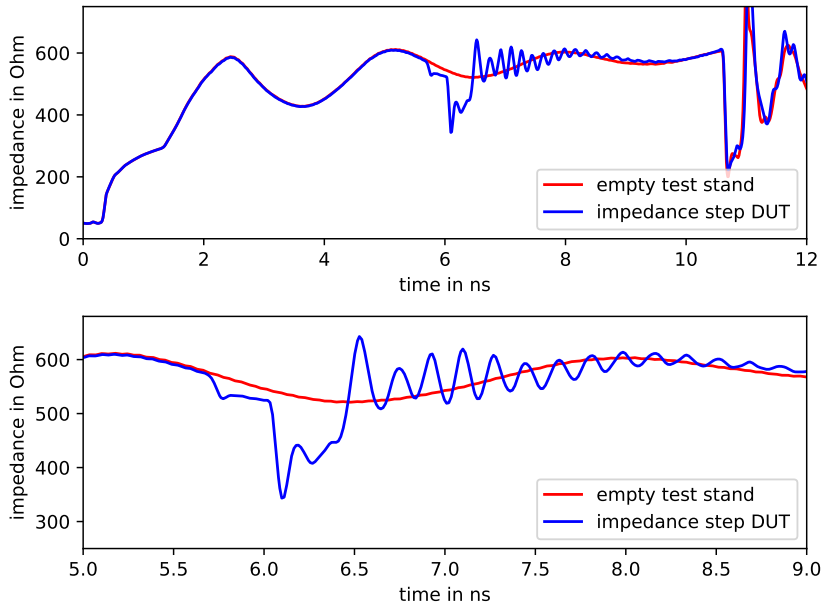


Figure 5.1: TDR measurements of impedance step DUT on the Goubau line test stand compared to the empty test stand as a reference. Full time span at the top and excerpt of DUT at the bottom.

The DUT has three 5 cm sections with constant impedance in each section and a step without any taper as a transition. Each impedance step is a source of reflection. The first section of the DUT is clearly visible in the TDR signal as a plateau of

5 Measurement Results

constant impedance. After each subsequent impedance step, the TDR signal shows increased ringing. This is due to multiple reflections between the impedance steps of the DUT. For example, the signal response of the second impedance step, i.e. the reflection at that step, needs to pass the first step again, before it can be recorded by the network analyzer. While passing the first step on the return, part of the signal is reflected again, altering the recorded impedance value for the second impedance step. This is unavoidable and causes ringing in TDR measurement when a DUT has multiple sources of reflection.

The impedance of the different sections of the DUT was calculated according to Eq. 4.2 as described in Sec. 4.1.3. The larger diameter sections should measure $252\,\Omega$ and the middle section $197\,\Omega$. The results shown in Fig. 5.1 do not match the calculated impedance but show significantly higher values for the entire DUT. This indicates that further calibrations for the Goubau line test stand are required. Comparing the measurement to the simulated test stand shows discrepancies in impedance values, especially after 4 ns as shown in Fig. 5.2. The general trend and

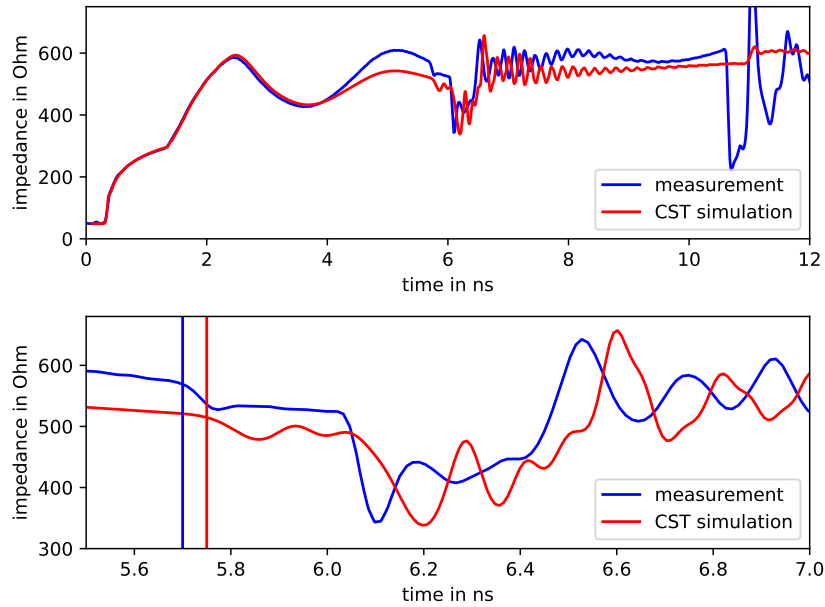


Figure 5.2: TDR signal of impedance step DUT comparing Goubau line measurements with CST simulations. Full time span at the top and excerpt of DUT at the bottom.

response of the DUT, however is comparable. There is a slight temporal mismatch of the reflection at the DUT entrance, highlighted by the blue and red vertical lines. This is most likely due to inaccuracies when placing the DUT at the test stand. The temporal difference is about $0.05\,\text{ns}$ which corresponds to a distance of $7.5\,\text{mm}$, taking into account that the $0.05\,\text{ns}$ are a round trip time difference. An uncertainty of $7.5\,\text{mm}$ is plausible for the experimental set up. In general the

CST simulations agree well with the measurements of the Goubau line test stand. Adjusting simulation parameters and mesh settings could be explored to achieve an even closer match.

5.1.1 Time Gating and Reference Scaling

Time gating is a technique in TDR where the causality argument is used to isolate reflected signals from specific parts of the DUT. This is done by only examining the time interval of the signal that corresponds to the part of the DUT under investigation. Figure 5.3 compares the full sample range and time gated signal from 0 to 9.5 ns of the TDR measurement of the empty test stand. This way the large reflection signal of the aluminum back plate can be omitted.

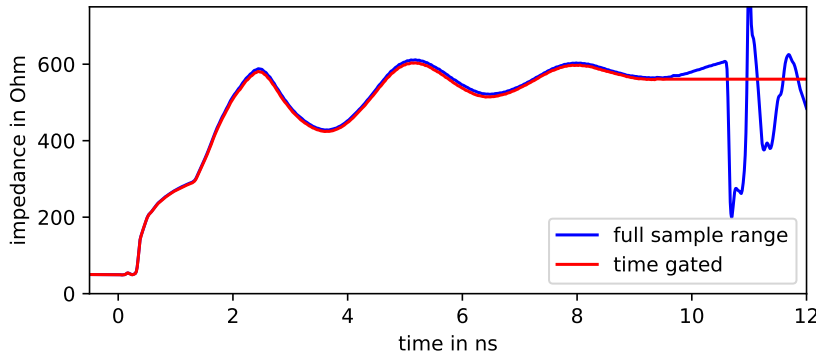


Figure 5.3: TDR of the empty Goubau line test stand, comparing the full sample range and time gated signal from 0 to 9.5 ns.

Time gating only the end of a TDR measurement might seem rather trivial as there is no difference to just ignoring the data after a certain point in time. There are however implications for spectral parameters like S_{11} . After selecting the data range in the time domain, the modified S parameters can be calculated via Fourier transformation. Figure 5.4 shows the S_{11} parameter for the full sample range and the time gated signal from 0 to 9.5 ns, which excludes the large reflection at the end of the test stand. The time gated S_{11} parameter shows a lot less noise.

This technique enables the comparison of S_{11} parameter between the measurement and the simulation, which differ greatly in how the end of the test stand is realized. The initial S_{11} parameter of the measurement and simulation are compared in Figure 5.5. There is clearly a mismatch for both cases but the time gated S_{11} parameter show a distinct matching feature at low frequencies, which is hidden by noise in the initial data.

Isolating the TDR response of the DUT is also possible, as shown in Fig. 5.6. Calculating the TDR signal requires a reference impedance Z_0 both in the time domain as shown by Eq. 3.6 or the frequency domain represented by Eq. 4.1. It is not always possible to determine the impedance of the structure at the point of

5 Measurement Results

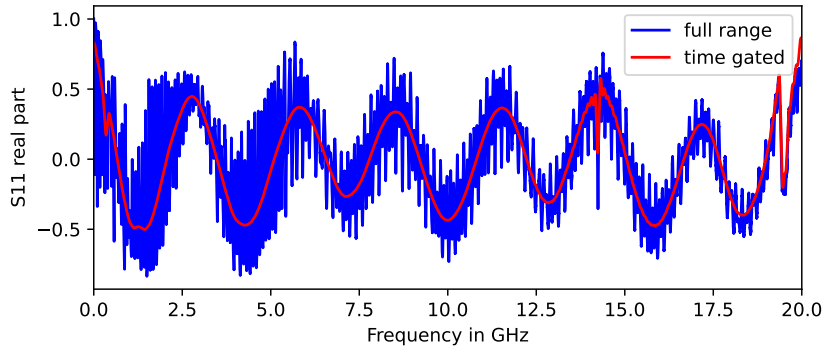


Figure 5.4: Real part of S_{11} parameter for the empty Goubau line test stand, comparing the full sample range and the time gated value from 0 to 9.5 ns

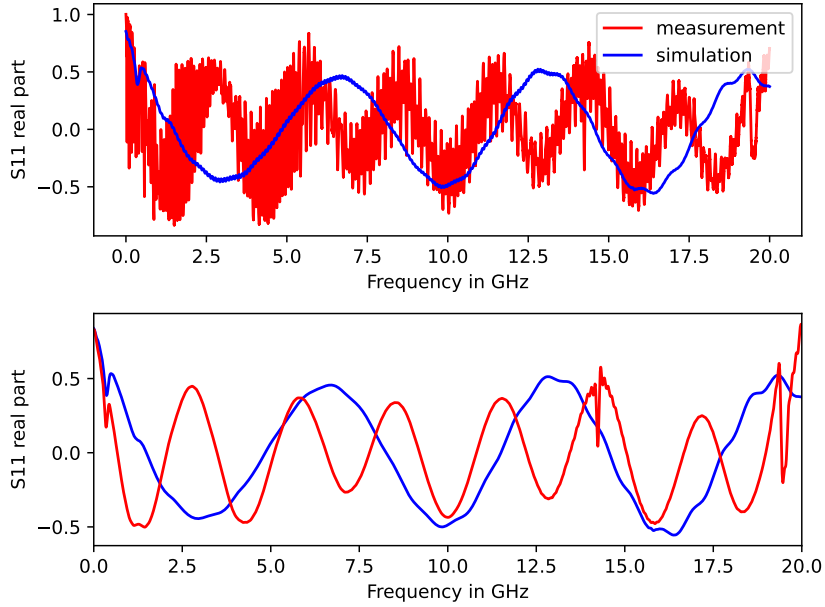


Figure 5.5: Real part of S_{11} parameter for the empty Goubau line test stand, comparing measurement and simulation. The top shows the full sample range and the bottom shows the time gated value from 0 to 9.5 ns

the time gate, relative impedance changes can however be visualized by choosing an arbitrary value for Z_0 .

With the isolated TDR signal it is possible to calculate the S_{11} parameter for the DUT itself. Figure 5.7 compares the S_{11} parameter derived from the time gated TDR signal of the DUT. This is done for both the measurement and the simulated data for a time gate from 5.3 to 10.5 ns. The measured results agree very well

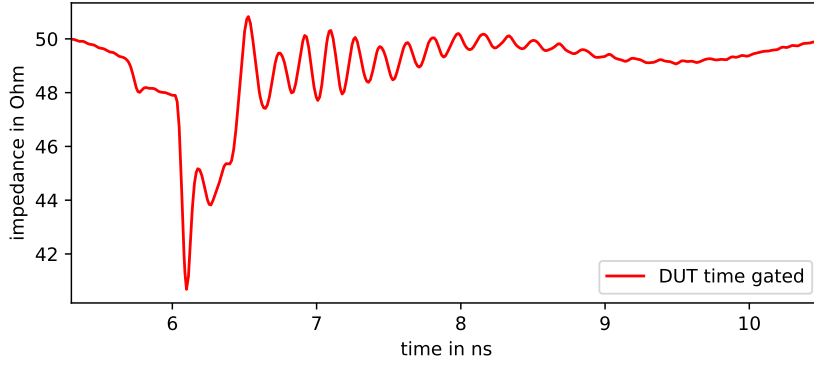


Figure 5.6: TDR signal of impedance step DUT time gated from 5.3 to 10.5 ns.

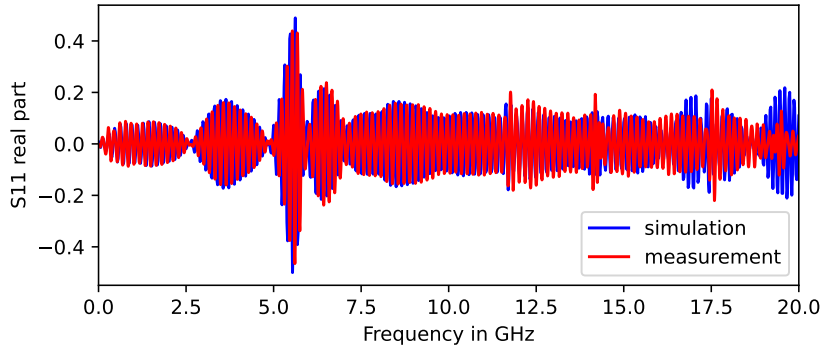


Figure 5.7: Real part of S_{11} parameter for the Goubau line test stand with impedance step DUT, time gated from 5.3 to 10.5 ns. Comparing measurement and simulation.

with the simulation, especially for frequencies below 12 GHz. This is despite the discrepancies shown in Fig. 5.5.

Another way to isolate the TDR signal of the DUT is to subtract the TDR signal of the empty test stand as a reference. This process is shown in Fig. 5.8 In order to compare this method, the time gated signal was scaled and offset. Figure 5.9 shows that both methods arrive at similar results. Sources of reflection are identifiable with both methods but time gating allows for further analysis like calculating the S_{11} parameter.

5 Measurement Results

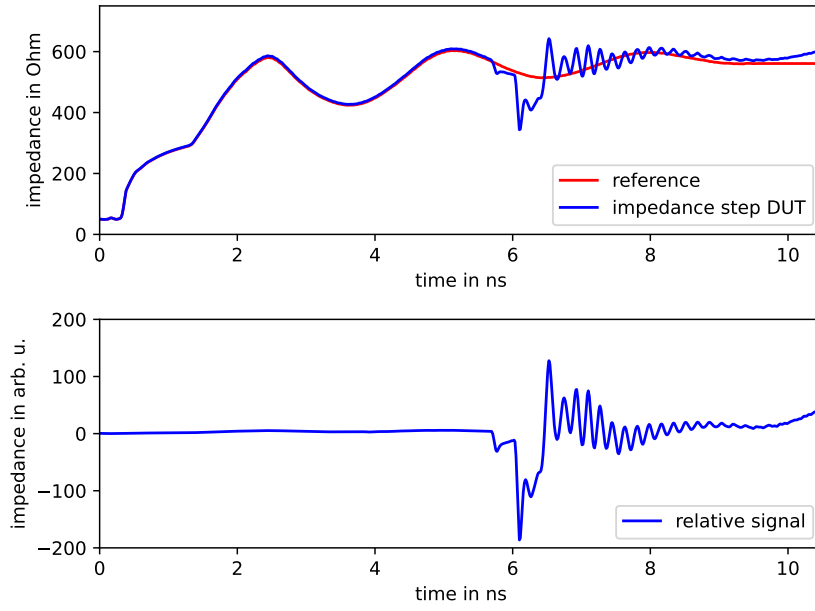


Figure 5.8: TDR signal of the Goubau line test stand with the impedance step DUT and empty as a reference at the top and the relative signal at the bottom.

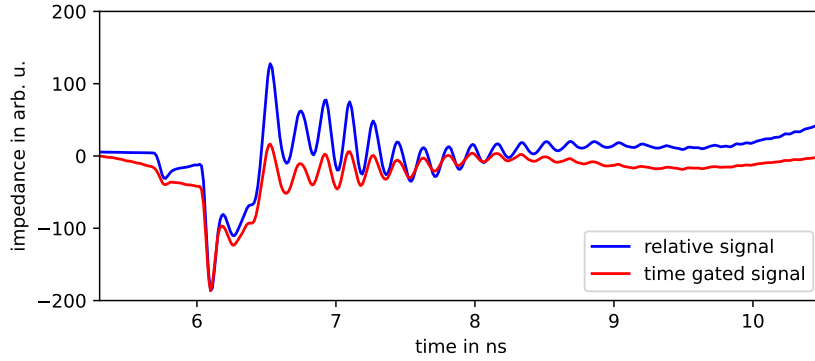


Figure 5.9: TDR signal of impedance step DUT comparing the time gated and relative impedance.

5.1.2 Field Extension Test

A set of measurements were conducted to determine the field extension around the wire of the Goubau line test stand. The impedance step described in Sec. 4.1.3 was set up right next to the wire of the test stand, as shown in Fig. 5.10. This measurement was then repeated with an increasing distance between the wire and impedance step. The distance was increased in 10 mm increments up to 50 mm.

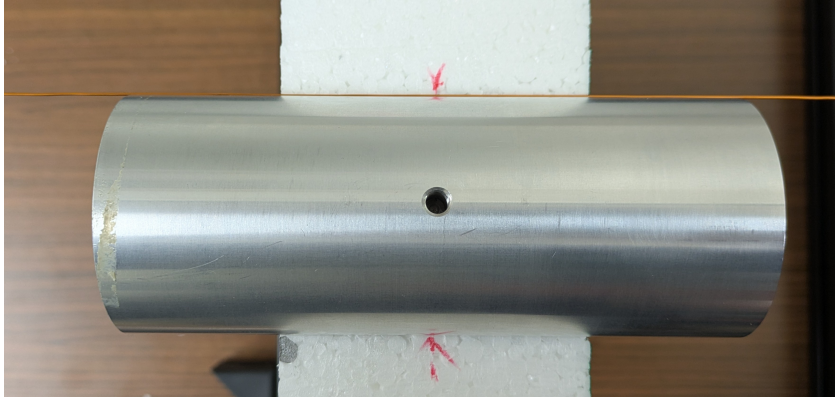


Figure 5.10: Impedance step DUT set up right next to the Goubau line wire to test the field extension.

Figure 5.11 shows the DUT 10 mm from the Goubau line wire.

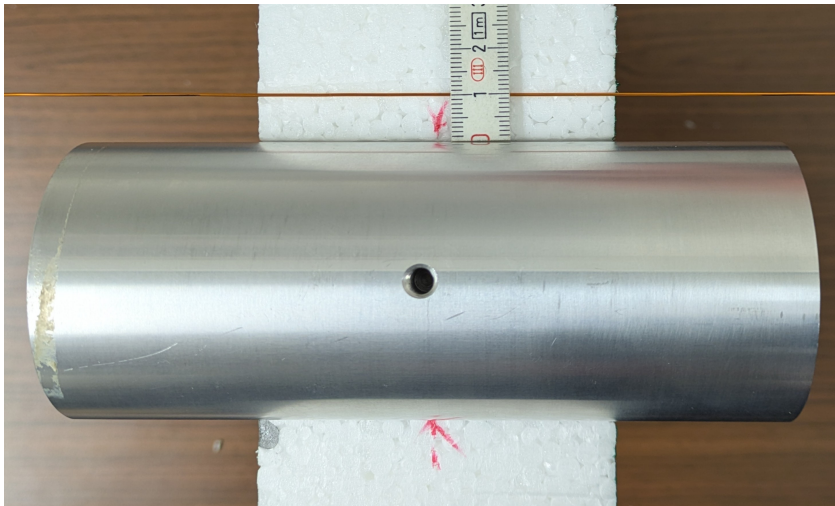


Figure 5.11: Impedance step DUT set up 10 mm from the Goubau line wire to test the field extension.

The results are compared and shown in Fig. 5.12. The case where the DUT is touching the wire shows the most dramatic decrease of impedance in the TDR measurement. It also shows the most temporal detail. The plots for a gap of 10 mm to 50 mm between the DUT and wire have a very similar effect on the impedance. Merely the amplitude of the change decreases as the gap is increased. Placing the DUT 50 mm from the wire still clearly affects the TDR measurement. This means the fields around the wire must extend further than that distance.

The measurement was repeated for distances between 0 mm and 10 mm in 1 mm. Figure 5.13 shows the TDR signal for select distances of the DUT from the wire.

5 Measurement Results

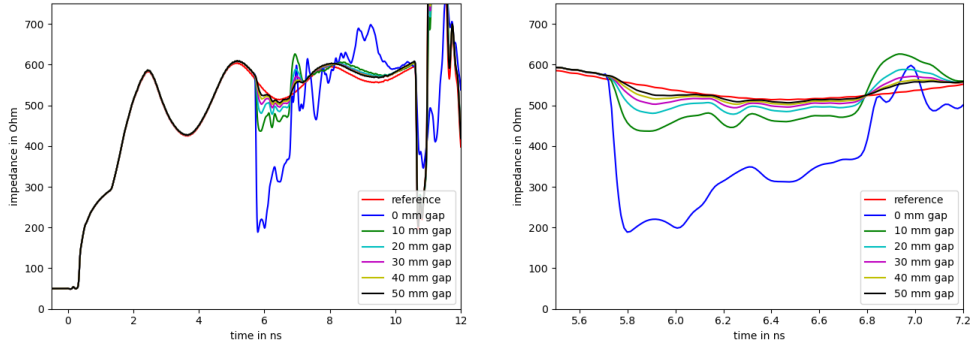


Figure 5.12: TDR measurement of impedance step DUT set up at different distances from the Goubau line wire to test the field extension. Full TDR plot on the left, zoomed view of the DUT response on the right.

The magnitude of the TDR response quickly decreases with distance.

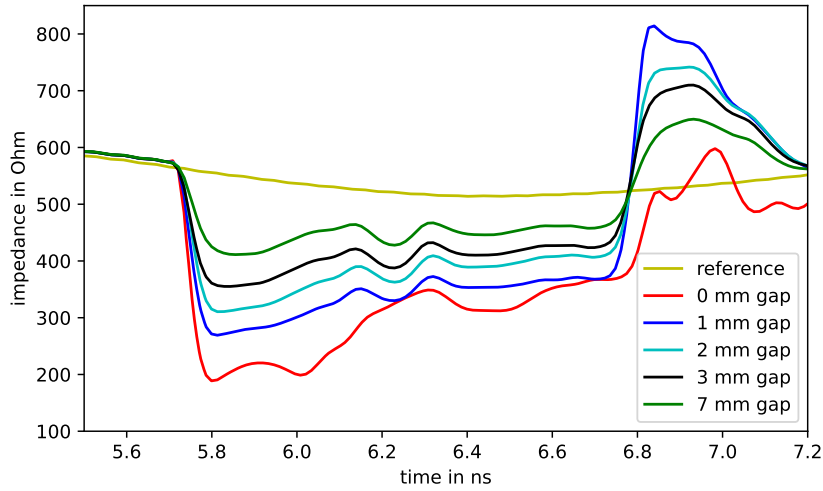


Figure 5.13: TDR measurement of impedance step DUT set up at different distances from the Goubau line wire to test the field extension, zoomed view of the DUT response.

Looking closely at the first reflection of the DUT, different gradients of impedance become apparent in Fig. 5.14. Not only does the magnitude of the response decrease with distance, the gradient does as well. This is an indicator of decreased temporal resolution of the TDR measurement. As discussed in Sec. 4.1.2, the wire of the used for the Goubau line test stand limits the field extension at high frequencies. As the DUT moves further from the wire, it moves out of range of the high frequency field components, resulting in decreased temporal resolution.

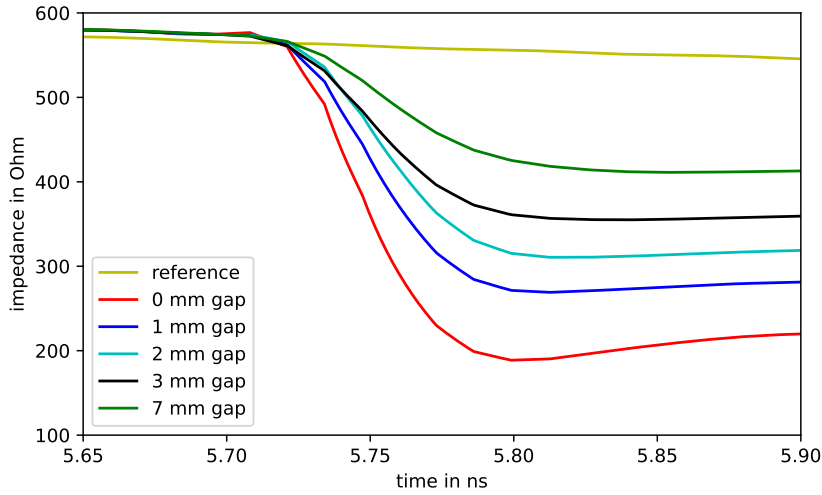


Figure 5.14: TDR measurement of impedance step DUT set up at different distances from the Goubau line wire to test the field extension, zoomed view of the first reflection from the DUT.

5.2 Foil Taper Models

The measurements of foil models with a constant 10 mm gap are shown in Fig. 5.15. The abrupt change in conductive material causes an impedance spike in the TDR measurement. This does not represent the reality of the undulator inside the storage ring, as electrical continuity will be maintained with the rest of the vacuum chamber. It is however necessary in the test stand but can be separated from the actual structure under test. The 0.5 mm wide longitudinal slit seems to have very little impact on the TDR measurement. To further investigate, two cases were simulated. One with a width of 0.5 mm and another with a width of 8 mm. Figure 5.16 shows the absolute value of transverse electric field around the Goubau-line wire as the pulse interacts with the shielding foils. A case without any slit is shown for reference. The narrow slit constrains the electric field very well and shows a similar field picture to the case without a slit. The field picture of the foils with an 8 mm slit shows significant field extension through the slit. The shortest wavelength in the simulated spectrum is 15 mm and therefore significantly larger than the slit of 0.5 mm width. The 8 mm slit on the other hand is large enough to be resolved by half a wavelength. This is probably the reason for the field behavior.

5 Measurement Results

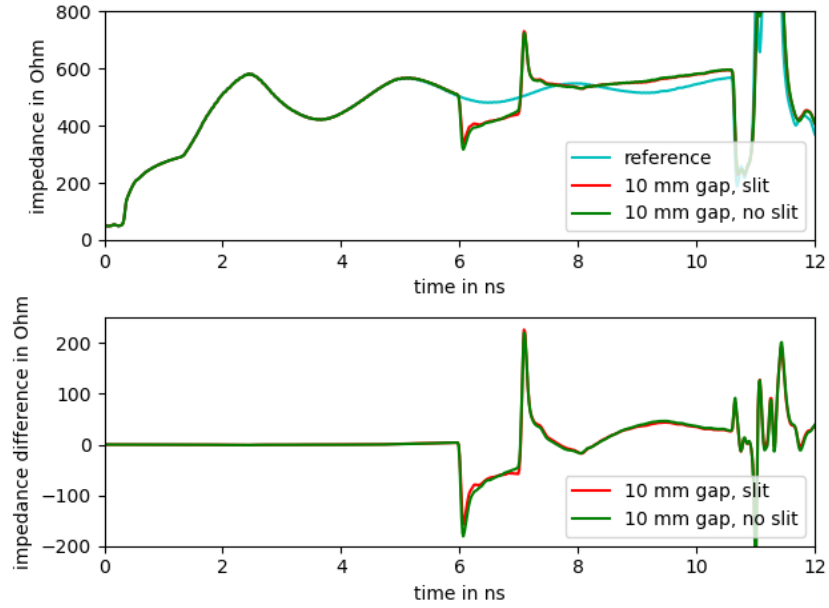


Figure 5.15: TDR signal of the Goubau-line test stand with a constant 10 mm gap foil model as DUT. Foil with longitudinal slit in red, without slit in green. The teal signal is a reference measurement of the test stand without DUT. The top graph shows the actual measurement, the bottom shows the signals with the reference subtracted.

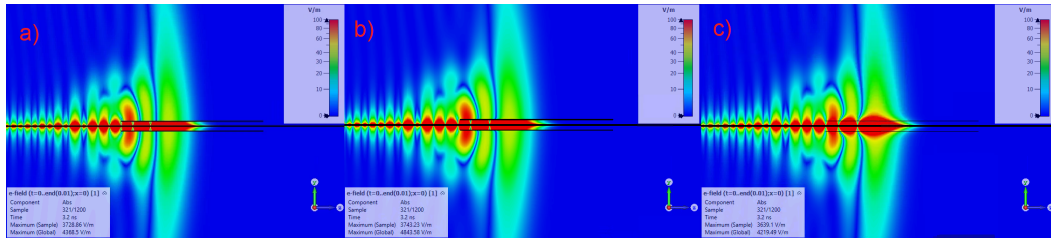


Figure 5.16: Absolute value of transverse electric field around the Goubau-line wire. Interaction of the pulse with shielding foils at a 10 mm gap a) without a slit, b) with a longitudinal slit of width 0.5 mm, and c) 8 mm. The narrow slit contains the electric field very well and shows a similar field picture to the case without a slit.

6 conclusion

In the scope of this thesis numerical simulations were conducted to study and calculate wakefields and impedances of in-vacuum undulator components, in particular IVUE32. The goal was to optimize the design towards minimizing the impedance contribution. In addition to the simulation studies, a Goubau line test stand prototype was developed and built. This test stand allows for the measurement of impedance of in-vacuum undulator components via time domain reflectometry. The test stand was found to produce measurements that agree well with simulated results. Since the measurement duration is independent of the geometric detail of the investigated components, the test stand can be used to compare detailed components to simplified models, determining which details are critical and what can be omitted in subsequent simulations. This reduces computational cost which heavily depends on the level of geometric detail that needs to be resolved. This is necessary to enable simulations of the whole undulator structure.

Feedback from wakefield simulations of the IVUE32 taper section lead to optimized designs in close cooperation with the engineers of the undulator group. Figure 6.1 shows a CAD model of the optimized taper design that separates the shift and gap movement compensation. These designs with limited impedance contributions are

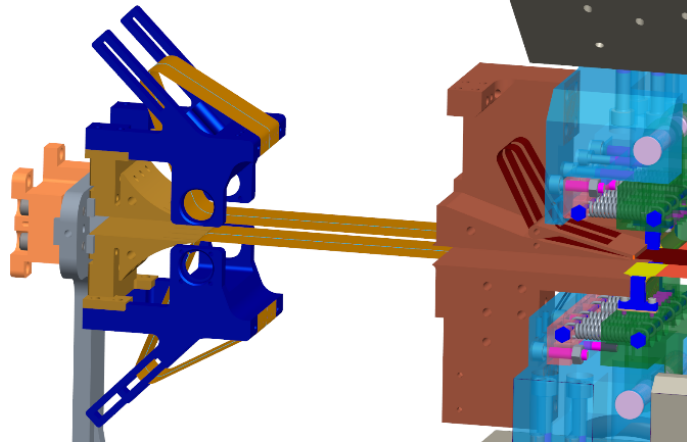


Figure 6.1: CAD model of the IVUE32 taper section with double foil tensioning mechanisms. Taken from [17]

currently being manufactured and will be examined on the Goubau test stand later in 2026. Further measurements will be conducted and possible constraints due to the machine optics need to be taken into account before the taper design can be

6 conclusion

finalized.

The overall design of IVUE32, including the vacuum chamber and magnet girder support structure is derived from CPMU17, an in-vacuum undulator that has been successfully commissioned and operated for several years. Beam driven impedance measurements of CPMU17 did not show any resonances [12]. Nevertheless the impedance of IVUE32 must be well understood and examined for possible resonances before it can be installed into the storage ring. In order to simulate large structures like IVUE32 with the necessary detail required for spectral resolutions up to 20 GHz, novel numeric methods need to be utilized. Code bases that can combine simulation results of individual sections to investigate the whole structure are being developed in the scope of the WAKE [23] and IVUS research projects. These simulations will be conducted in future studies.

7 Appendix

7.1 Effect on TDR Measurement of Different Support Materials

7.1.1 Introduction

This experiment is meant to determine the impact of possible support material for future DUTs on the TDR signal. Several materials are tested by being placed close to the Goubau line 70 cm from the end of the test stand.

7.1.2 Measurement Devices

7.1.2.1 NWA: ZNL20

- Rohde und Schwarz
- Bandwidth: 5 kHz to 20 GHz

7.1.3 Calibration and Set Up

The ZNL20 has an automated calibration kit that connects to both ports and to the device via USB. The calibration is automated and sets the reference plane for the TDR measurement at the point where port 1 connects to the calibration device. The record length is set to 100001 points on the frequency sweep. This results in a range from 199.998 kHz to 20 GHz. After that the automatic calibration routine using the USB calibration kit is executed.

For the time domain conversion the Hamming window is chosen that suppresses the first side lobe. This was chosen as a good compromise for temporal resolution without pronounced ringing.



Figure 7.1: Photo of wire test stand with hard foam block.

7.1.4 Measurement

Wood, two types of foam, extruded aluminum profile, and a wood polymer profile where tested as support material. The materials are shown in Fig. 7.2.

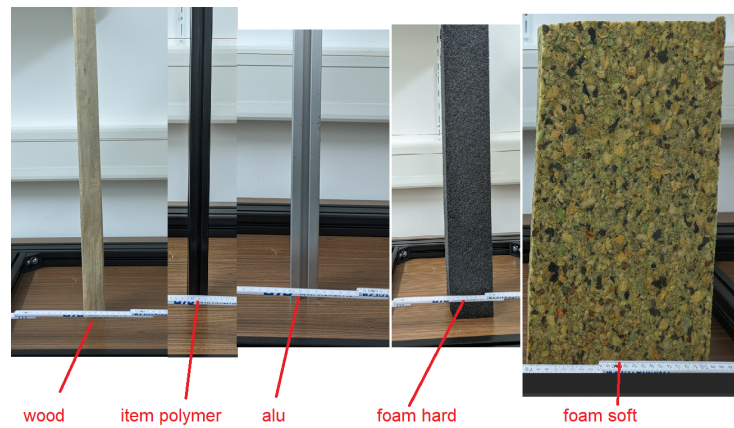


Figure 7.2: Photos of tested support material

The unimpeded wire test stand was taken as a reference signal to be compared with the signal from the different materials as shown in Fig. 7.3.

The item polymer profile and the extruded aluminum have a very similar geometry. Taking a closer look at these two cases reveals a similar response (Fig. 7.4). As expected the metal has a much larger impact on the impedance. The geometry of the extrusions is shown in Fig. 7.5 and has characteristics of a cavity. This could explain the ringing of the impedance signal for the metal extrusion.

The polymer profile has a wood fiber content of about 70 % according to the manufacturer. The impedance of a similar sized solid piece of wood is shown in Fig. 7.6 as a comparison. The magnitude of the impedance is comparable. The least impact on impedance was observed with the foam blocks. Two types of foam were tested. A fairly hard, black packaging foam and a soft mattress type foam. The impedance is shown in Fig. 7.7 Two polystyrene foam types were tested. An

7.1 Effect on TDR Measurement of Different Support Materials

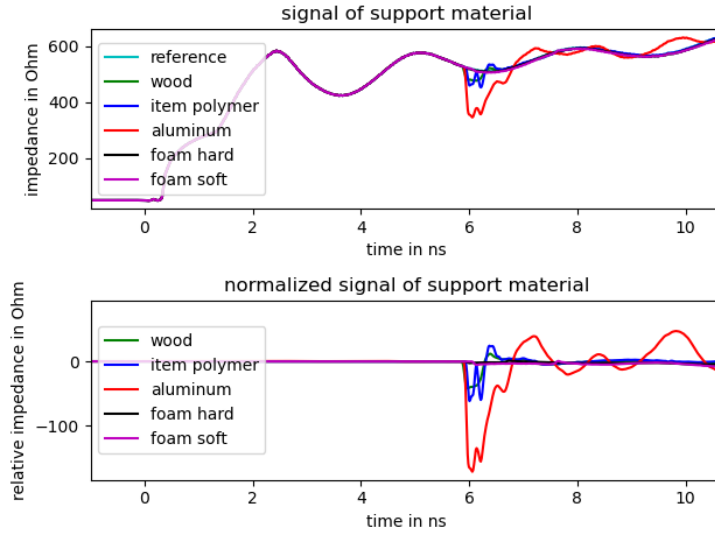


Figure 7.3: Impedance of different support materials compared to the unimpeded test stand as a reference. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

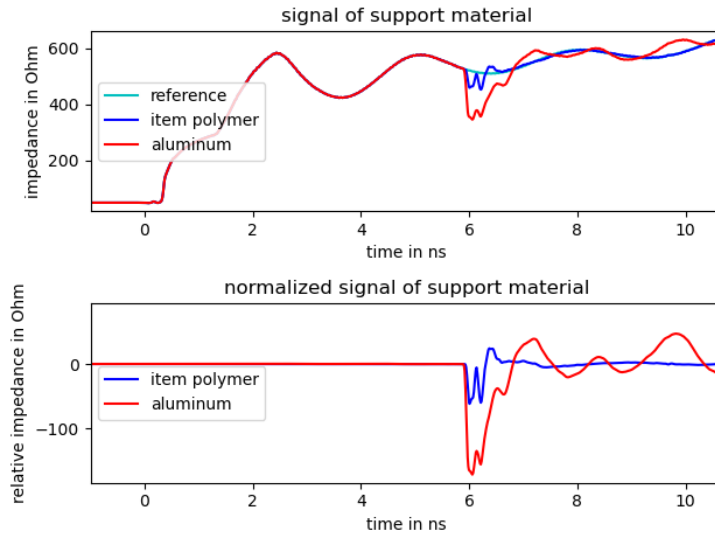


Figure 7.4: Impedance of different extruded profiles. One made of non-conducting wood polymer and one made of aluminum. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

expanded polystyrene foam (EPS) and extruded polystyrene (XPS). The material should be chemically identical. The difference is in air content and cell structure. XPS is more dense but also more sturdy and stiff. A 6 by 8 cm block was placed

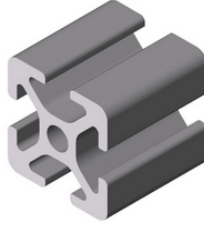


Figure 7.5: Geometry of extruded profiles. The T-slot has a similar geometry to a cavity.

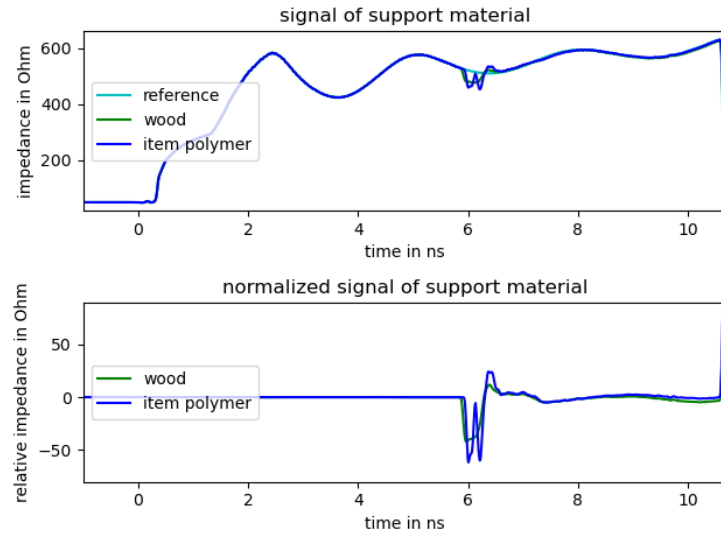


Figure 7.6: Impedance of wood polymer profile and similar sized solid wood plank. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

at 70 cm from the end of the test stand. Fig. 7.8 shows the impedance. XPS has a stronger impact on impedance than EPS, probably due to its higher density. Both seem to be less impactful than the other foam types tested. The other foam blocks however have different dimensions. For a definitive comparison, foam blocks of similar dimensions have to be compared.

7.1 Effect on TDR Measurement of Different Support Materials

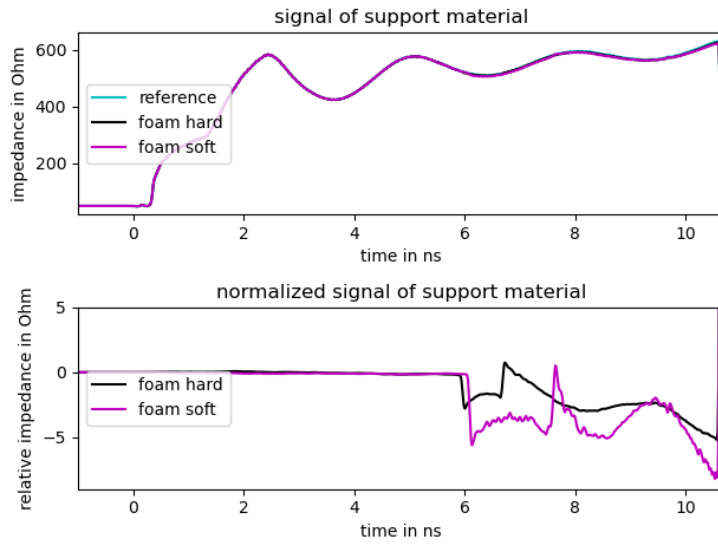


Figure 7.7: Impedance of two different types of foam. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

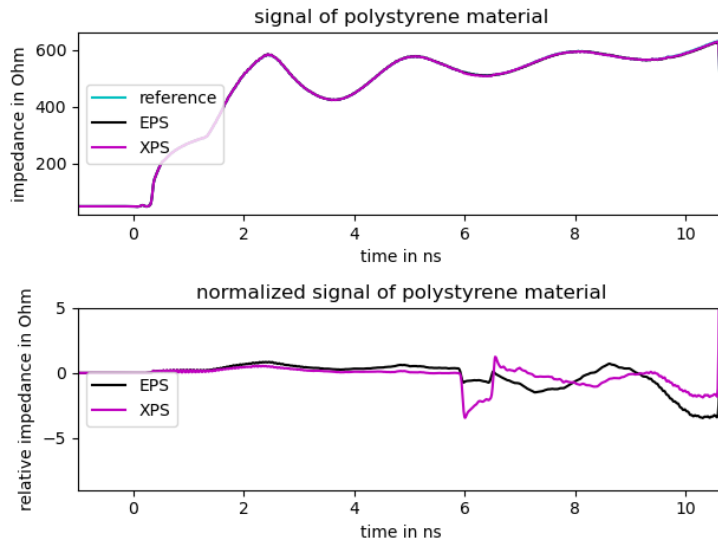


Figure 7.8: Impedance of two different types of polystyrene foam. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

7.1.5 Polymer Profile at different locations

This series of measurements puts the wood polymer profile at different distances from the cone antenna. The profile was placed at 4, 46, 75, 96, 121 cm. Fig. 7.9 shows the results.

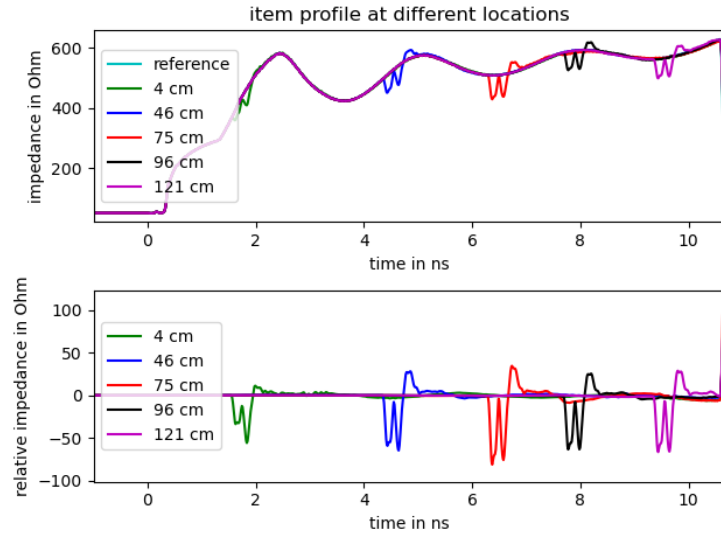


Figure 7.9: Impedance of wood polymer profile at different distances from cone antenna. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

Another measurement was done with two wood polymer profiles at 46 and 96 cm from the cone antenna. Fig. 7.10 shows the impedance of the two profiles and Fig. 7.11 compares the impedance to a single profile at the same distances.

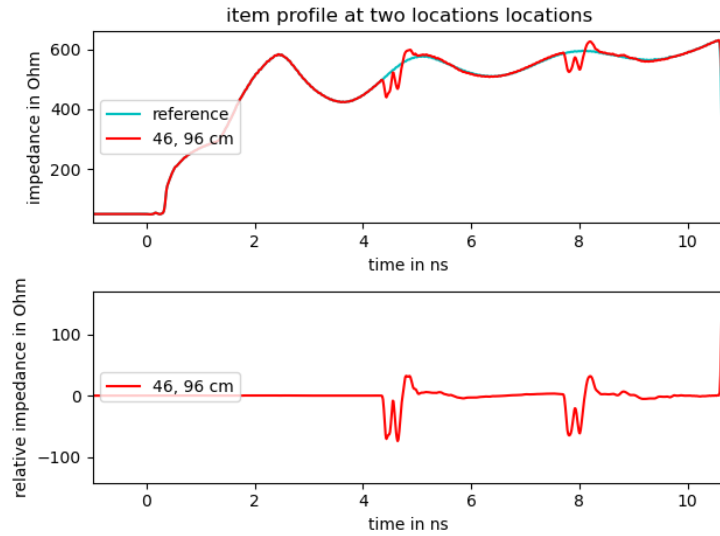


Figure 7.10: Impedance of two wood polymer profiles at different distances from cone antenna. Raw impedance at the top, relative impedance with reference subtracted at the bottom.

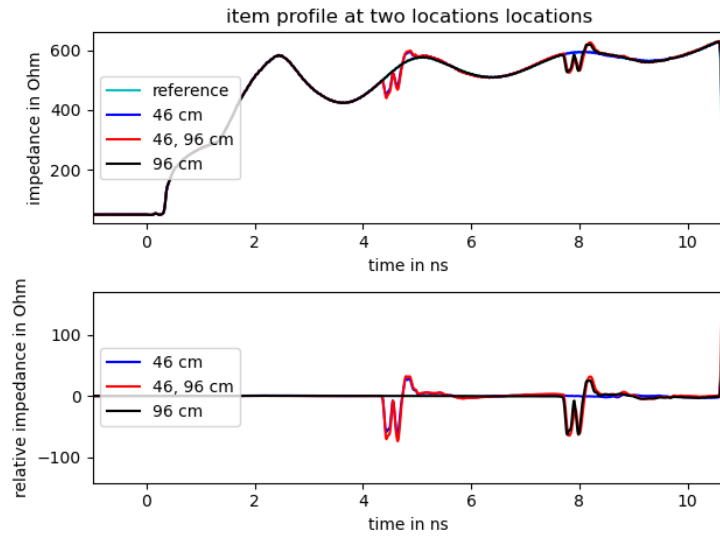


Figure 7.11: Impedance of two wood polymer profiles at different distances from cone antenna compared to a measurement of a single one at the same distances. Raw impedance at the top, relative impedance with reference subtracted at the bottom.



Figure 7.12: Photo of test stand with two wood polymer profiles at different distances from cone antenna.

7.1.6 Conclusion

A clear differentiation can be made in the impedance measurement between different types and geometries of materials placed close to the wire. The impedance impact of the wood polymer profile was much greater than anticipated. Foam as a support material for future DUTs might be a better choice. Although the positional accuracy for Foam supports might be inferior to the extruded polymer profiles. Another measurement should be conducted with Styrofoam as it might offer a better compromise than the tested foam types.

Considering the new polystyrene measurements, XPS seems most suited as support material. It is noticeably more rigid than the other foam types and still has a low impact on impedance.

7.2 Data Sheet for Test Stand Wire

Kupferlackdraht **CUL 100/0,75**



Abbildung zeigt CUL 100/0,22

Vorteile

Isolierlackbasis min. Polyurethan, mod.

Grad 1, TJ min. 155 °C

Lötbar

Anwendungen

Kupferlackdraht zur Herstellung von Spulen und Wicklungen sowie für den allgemeinen Laborbedarf. Bitte die einschlägigen Normen beachten.

Zulassungen



Kupferlackdraht CUL 100/0,75

Elektrische Daten

Typ	CUL 100/0,75
-----	--------------

Mechanische Daten

Typ	CUL 100/0,75
Betriebsdaten	
Strombelastung	1,55 A
Umwelt	
Dauerwärmeständigkeit	155 °C (CLF)
Isoliertackbasis	min. Polyurethan, mod.
Verzinnbarkeit	>350 °C
Maße und Gewichte	
Drahtdurchmesser	0,75 mm
Länge m (min.)	20
Gewicht	0,10 kg

Änderungen vorbehalten.

7.3 Foil Slit Simulations - Additional Considerations

7.3.1 Modeling and Meshing

The shielding foils were modeled as sheet geometries in CST to avoid structures with one dimension being significantly smaller than the other two dimensions. The sheet modeling together with a curved geometry results in many metal to non-metal interfaces within the mesh cells. This causes CST to represent these mesh cells in staircase mode and to consider these cells filled with perfect electric conductor (PEC). Depending on the size of the affected mesh cells, this can severely alter the simulated electric geometry from the original model. This effect is shown in Fig. 7.13 with the mesh cells filled with PEC drawn in teal. This significantly alters the sinusoidal shape of the foil slit. This problem can be reduced by locally increasing

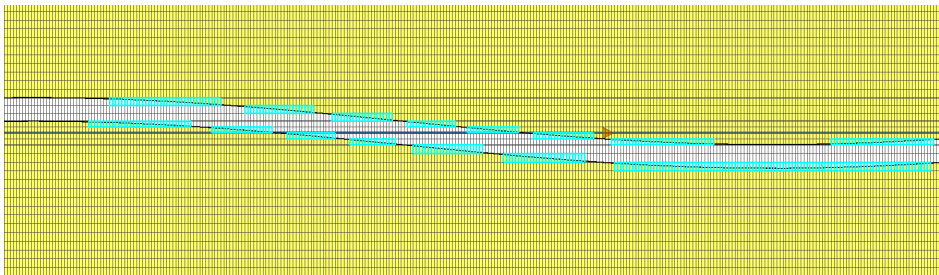


Figure 7.13: Mesh view of the CST shielding foil simulations. Mesh cells filled with PEC are shown in teal.

the mesh resolution, reducing the size of the affected mesh cells. Curved metal to non-metal interfaces can never be perfectly represented by a hexahedral mesh, but CST uses algorithms to correctly model these interfaces at sub mesh cell accuracy. The staircase approximation only gets used when the solver cannot correctly resolve the geometry otherwise. Dramatically increasing the mesh resolution didn't prevent the staircase approximation even for surfaces at very small angle to the mesh lines. Other simulations of the IVUE32 taper featured much larger angles of interfaces to mesh cells without resulting in staircase approximation. This leads to the assumption that two metal to non-metal interfaces in a single mesh cell cannot be resolved. This would make the sheet representation of the foils problematic. Modeling the foils with a thickness of 0.1 mm results in almost no cells being filled with PEC. Modeling the Foils as solid copper blocks with the slit cut out, results in no cells being filled with PEC.

The electric fields in these simulations have frequencies in the GHz range which means a skin depth of only a few or even sub μm . Therefore only the surface geometry is significant and changing the foil thickness should not change the simulation results. Figure 7.14 shows the three different modeling approaches for the shielding foils.

To compare the three different models the longitudinal impedance is examined.



Figure 7.14: Different model representations of shielding foils. Sheet representation on the left, modeled foil thickness of 0.1 mm in the middle, and a 3 mm solid block with carved slit on the right.

Figure 7.15 shows two distinct peaks close together for all three cases. Zooming in on the double peak reveals a slight difference in frequency and amplitude for the three different models. The peaks in impedance are at frequencies of about 2.3 GHz and 2.5 GHz.

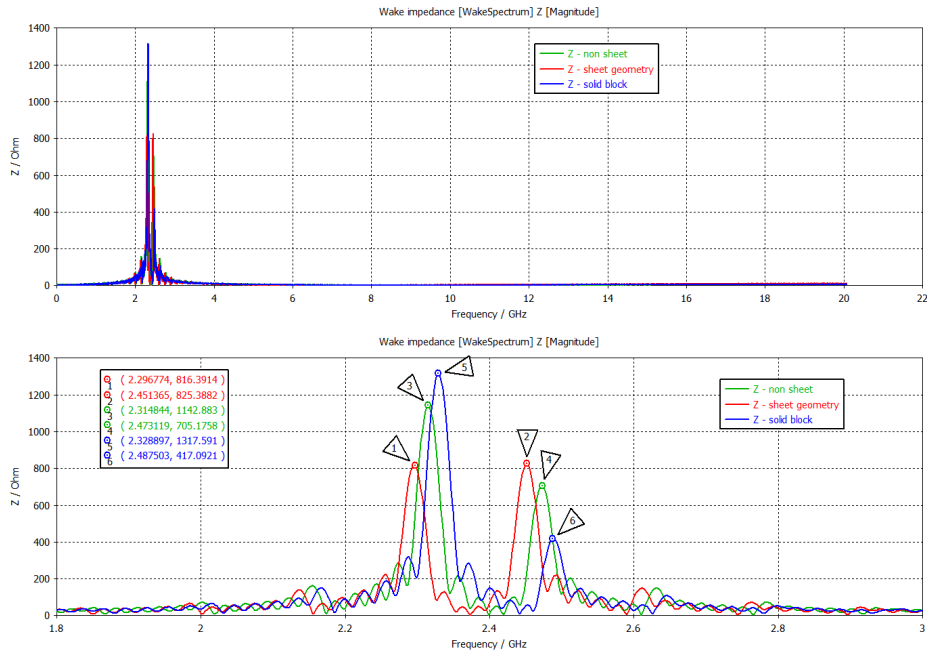


Figure 7.15: Longitudinal impedance of the three different sinusoidal slit models compared. Full frequency range at the top and zoomed view of the resonant peaks on the bottom.

Since the solver settings, beam parameters, and the meshing are the same for all three simulations the difference in resonant frequencies must be caused by differences in the model. The difference between the sheet model (red graph in Fig. 7.15) and the thin foil model (green graph in Fig. 7.15) lies in the PEC filled mesh cells. The electrically relevant surfaces are otherwise modeled the same way (except for a tiny difference at the start and end of the slit similar to what will be explained next for the solid model).

The solid block model (blue graph in Fig. 7.15) is different from the other two models at the beginning and end of the slit. As shown in Fig. 7.14 there is a vertical wall modeled, that does not exist in the other two models. These walls can reflect the electric fields and change the frequencies of resonant impedance. To test the effect of these vertical walls, the walls have been removed as shown in Fig. 7.16 and the simulation was rerun.

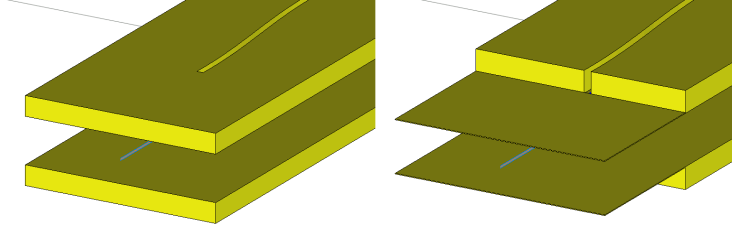


Figure 7.16: Block representation of shielding foils on the left and the same model with the vertical walls at either end of the slit removed on the right.

The longitudinal impedance is compared with the thin foil model in Fig. 7.17. There are no significant changes over the broad spectrum, so only the resonant peaks are shown. The resonant peaks for the solid block model with the vertical walls removed (orange graph in Fig. 7.17) now match the frequency of the resonances of the thin foil model (green graph in Fig. 7.17). There is only a small difference in amplitude.

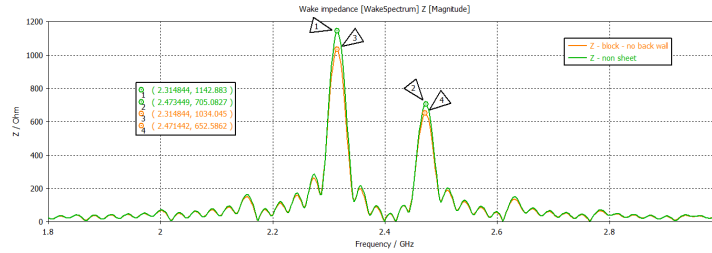


Figure 7.17: Resonant peaks in the longitudinal impedance for the thin foil model in green and the solid block model with the vertical walls removed in orange.

The design of IVUE32 as presented at IPAC 2025 [17] is depicted in Fig. 7.18. Depending on how far the folded section of the shielding foils will extend past the undulator magnets, either foil model will be closer to the actual design.

7.3.2 Straight Slit Simulations

The ideal case for the folded shielding foils would be a straight slit without deviation. This case was also simulated with three different foil models. As with the sinusoidal slit, the foils are modeled with a thickness of 0.1 mm in one case and as 3 mm thick

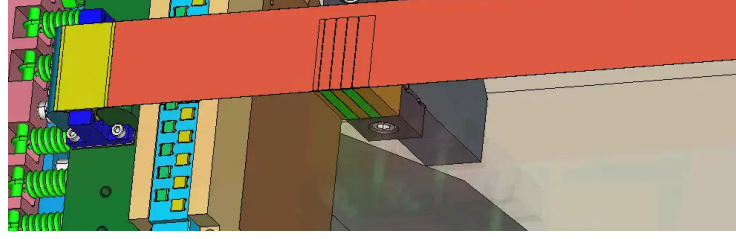


Figure 7.18: CAD model of IVUE32 showing a shielding foil at the end of the magnet assembly.

slabs with the slit cut out in the other cases. There are two different slab models, one with vertical walls at either end of the Slit and one with the vertical walls removed. Figure 7.19 shows the three different shielding foil models.

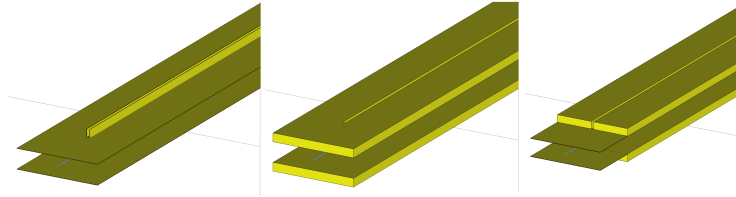


Figure 7.19: Different model representations of shielding foils with a straight longitudinal slit. Modeled foil thickness of 0.1 mm on the left, a 3 mm solid block with carved slit in the middle, and solid block model with vertical walls removed on the right.

Looking at the longitudinal impedance in Fig. 7.20, there are no resonant peaks. The impedance curve is smooth over the whole bandwidth and very similar to the impedance of two parallel plates. There are slight differences in impedance for the different Foil models, especially for the higher frequencies.

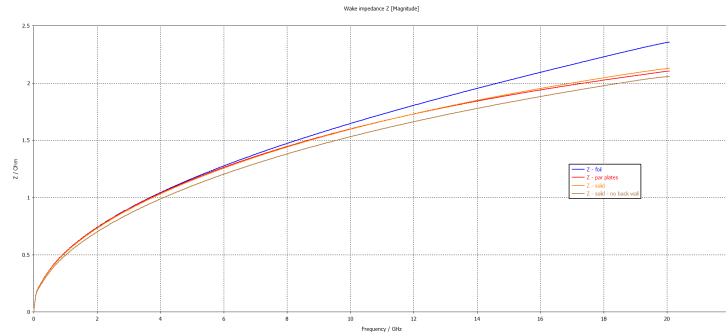


Figure 7.20: Longitudinal impedance of different straight slit models compared with two parallel plates.

These differences are less than half an Ohm at most. This means that shielding foils

with a perfect longitudinal slit behave very similar to parallel plates or shielding foils without a slit in terms of beam dynamics.

Now we need to determine a limit for the acceptable uncertainty when folding the shielding foils.

7.3.2.1 Localising Power Loss

To find out where in the shielding foils the power loss is concentrated, a surface current monitor is defined. Another method is to divide the shielding foil model into sections and look at the power loss for each component. The foil Surface towards the electron beam was divided into 0.2 mm wide strips with a thickness of 0.1 mm. The total power loss for the sectioned simulation was significantly different from the simulation where the foil is modeled as a single component, as shown in Fig. 7.21. This should not be the case, as the physical Model is not changed.

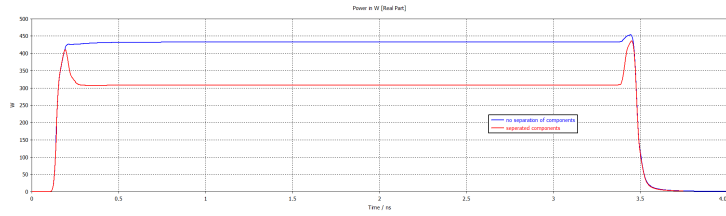


Figure 7.21: Power loss of straight slit foil simulation. Comparison between modeling the foil as a single component and dividing the foil into thin Strips. There should be no difference in the total loss.

Looking at the power loss for each strip of foil shows more inconsistencies. The expectation is a decrease in power loss as the strips move further from the beam. Figure 7.22 shows the power loss for each strip. As the index increases, the strip is located further from the beam. The expectation holds for the central region where there is a high resolution local mesh defined. Slice 8 and 10 show pretty much no losses while slice 7 and 9 show higher losses than expected.

This points towards an insufficient mesh resolution for the foil sections. Figure 7.23 shows the meshing for the segmented foil model. Close to the beam there are two mesh cells for every foil segment, compared to two foil segments per mesh cell outside of the high resolution area. To accurately calculate the losses for the segmented foil model, the mesh resolution needs to be increased.

With sufficient mesh resolution the power loss distribution behaves as expected with foil sections further from the slit experiencing less power loss than ones closer to the slit.

Unfortunately there is no field monitor available to show power loss in metals. The surface current however, can be visualized. The resistance of the copper foils is the only source of power loss. Therefore the surface current can be directly related to

7 Appendix

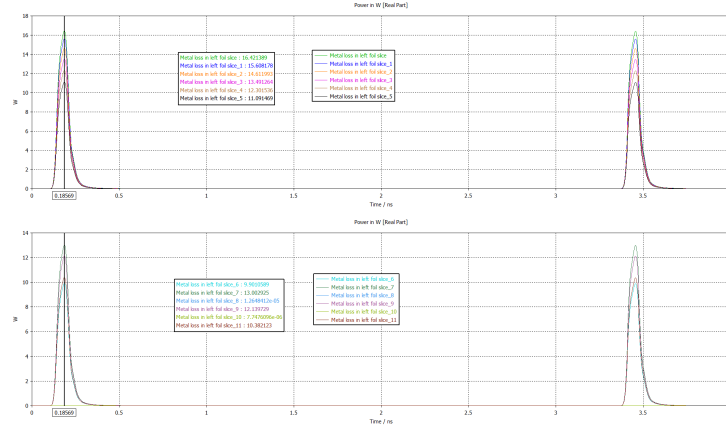


Figure 7.22: Power loss of straight slit foil simulation. Comparison between foil sections. The index denotes the increasing distance to the center of the foil. The top shows the region closest to the beam with a finer mesh resolution. The bottom shows the outer region with coarse mesh resolution. In the outer region some slices show higher losses than expected while others show almost no losses.

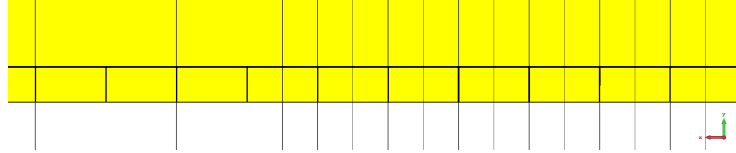


Figure 7.23: Changing mesh resolution for segmented foil model. On the right there are two mesh cells for every foil segment, compared to two foil segments per mesh cell on the left.

the power loss

$$P = RI^2 \quad (7.1)$$

with the help of Ohm's law.

7.3.2.2 Breaking Vertical Symmetry

To test the validity of the defined vertical magnetic field symmetry, the sinusoidal deviation is simulated without it. The structure still has a vertically symmetric geometry. The vertical impedance is shown in Fig. 7.24. With the defined symmetry plane, the vertical impedance is a constant 0 for all frequencies, as expected. The simulation without the defined symmetry is not constant over the frequency range. The amplitude however is very small, so that the symmetry approximation is valid. It is notable that the double resonance peak at around 2.3 GHz is visible in the vertical impedance.

7.4 IVUE32 Taper Simulations - Additional Considerations

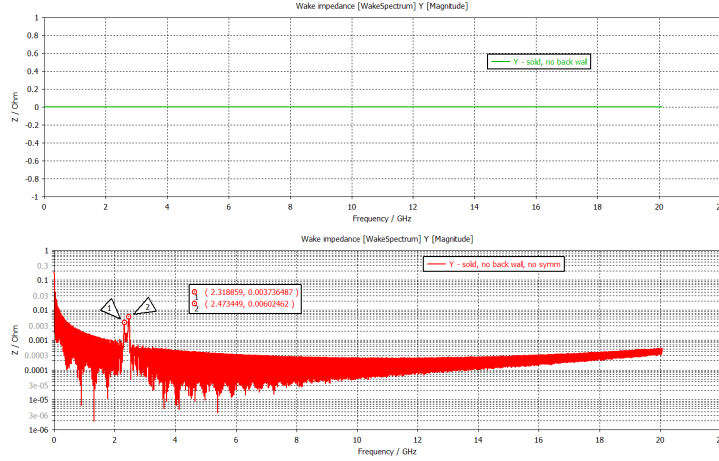


Figure 7.24: Vertical impedance of shielding foil with sinusoidal slit. At the top with defined magnetic symmetry plane, at the bottom without defined symmetry.

7.4 IVUE32 Taper Simulations - Additional Considerations

7.4.1 Mesh Convergence Study

The first simulations with lower longitudinal mesh resolutions showed questionable reflections in the electric field propagation. This prompted a mesh convergence study and a comparison between the two available beam injection schemes. The *Transmission line* and *Analytic* injection schemes are described in the CST user manual with the relevant excerpt shown in Fig. 7.25. Both schemes can produce

Scheme	Description
Analytic	This scheme integrates the Gaussian charge distribution very accurate at each time-step and computes the resulting current. Its advantage is that the current created by the travelling bunch is modeled very accurately. The disadvantage is that non-physical longitudinal electric fields appear due to numerical dispersion effects. This might affect the wake computation if the grid is too coarse.
Transmission line	This algorithm creates a current injection which is equal to a current traveling on a PEC wire. The advantage of this approach is that non-physical longitudinal electric fields are not created which improves the quality of the wake potential. Its disadvantage is, that current "reflections" appear if the grid is not homogeneous in longitudinal direction.

Figure 7.25: Current injection schemes as described by the CST documentation.

field errors if the longitudinal mesh resolution is not adequate. The longitudinal mesh resolution can be controlled with the *Lines per sigma* parameter in the beam settings (see Fig. 3.3). To find an adequate resolution a parameter scan was done for both injection schemes. Only the shift movement compensation assembly was used for the simulations and the *Lines per sigma* parameter was varied from 5 to 50 in steps of 5 for both injection schemes. The longitudinal wake potential and impedance was used to compare each resolution. Figure 7.26 shows the convergence

7 Appendix

of the longitudinal impedance for different resolutions using the analytic injection scheme. The convergence using the transmission line injection scheme is shown in Fig. 7.27

7.4.1.1 Analytic Injection Scheme

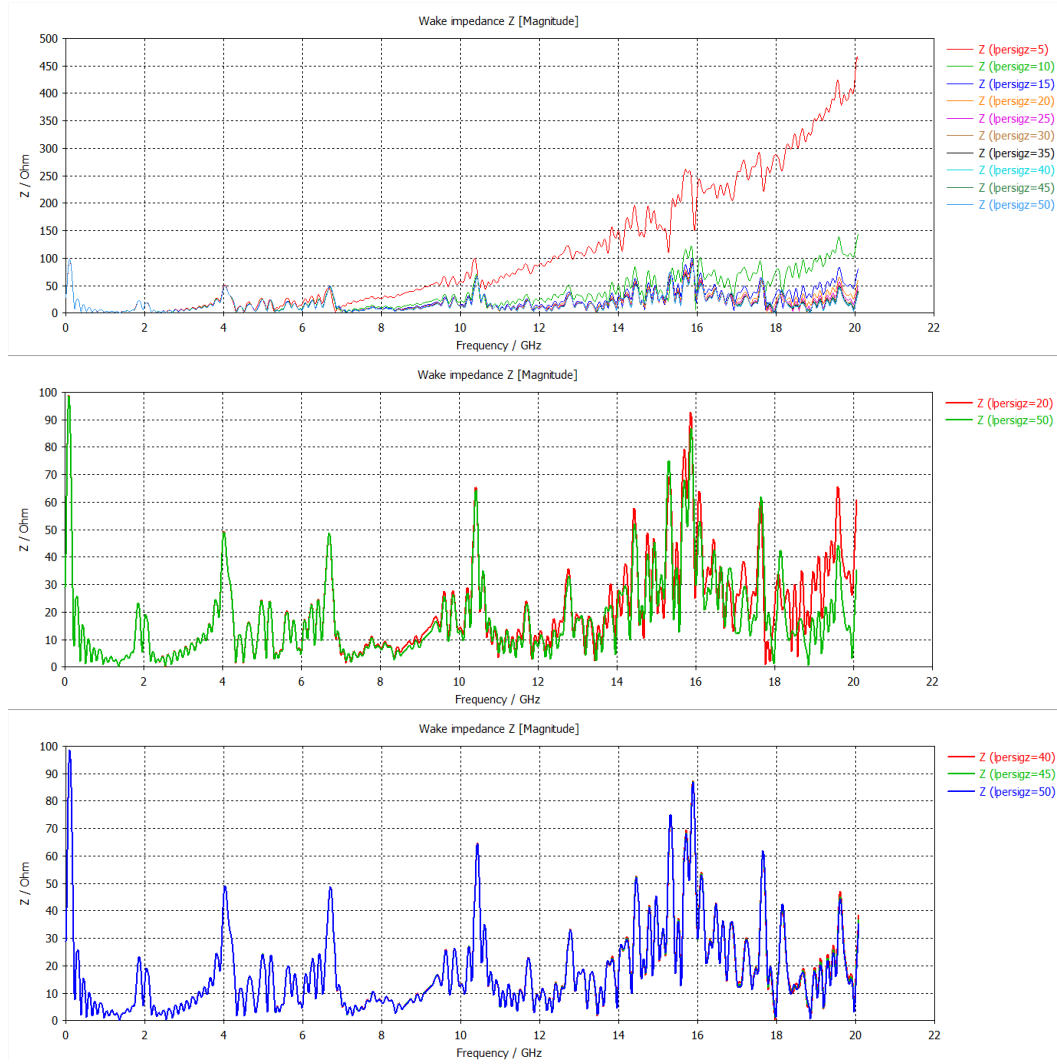


Figure 7.26: Analytic charge injection scheme: Longitudinal wake impedance compared for different longitudinal mesh resolutions. The parameter $l_{persigz}$ determines the amount of longitudinal mesh lines per beam size $\sigma_z = 5.1$ mm. The top plot shows the whole parameter scan with large discrepancies in the higher frequency range for the different resolutions. The middle plot compares the highest resolution with a medium resolution. There are significant discrepancies at higher frequencies. The bottom plot shows the highest three resolutions from the parameter scan. There are still small discrepancies visible above 16 GHz indicating that the longitudinal resolution is still inadequate.

7.4.1.2 Transmission Line Injection Scheme

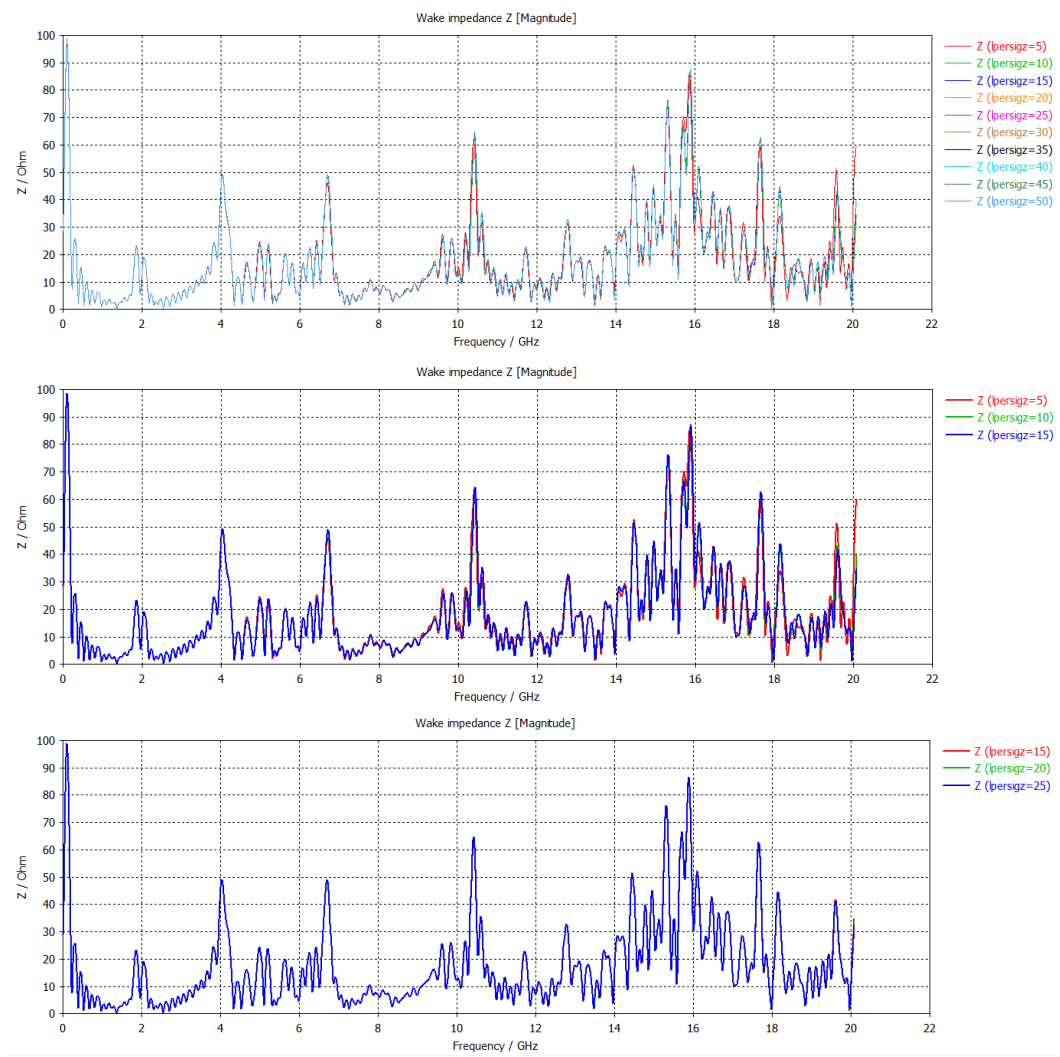


Figure 7.27: Transmission line charge injection scheme: Longitudinal wake impedance compared for different longitudinal mesh resolutions. The parameter $l_{persigz}$ determines the amount of longitudinal mesh lines per beam size $\sigma_z = 5.1$ mm. The top plot shows the whole parameter scan with significant discrepancies in the higher frequency range for the lower resolutions. The middle plot shows the lowest three resolutions. There are visible discrepancies above 10 GHz. The bottom plot shows the next three higher resolutions. The plots are practically identical, indicating that the results converge at a resolution of 10 lines per σ_z .

7.4.1.3 Conclusion

Figure 7.26 and 7.27 show the mesh convergence for the analytic and transmission line injection scheme respectively. The transmission line injection scheme converges much faster at a longitudinal resolution of 15 lines per σ_z . Figure 7.28 shows the highest resolution analytical injection with the converged transmission line injection. There are still visible discrepancies above 12 GHz, indicating that the resolution for the analytical scheme has not converged at 50 lines per σ_z . Further increasing the longitudinal resolution would become prohibitively expensive in terms of simulation time. The transmission line scheme, together with a longitu-

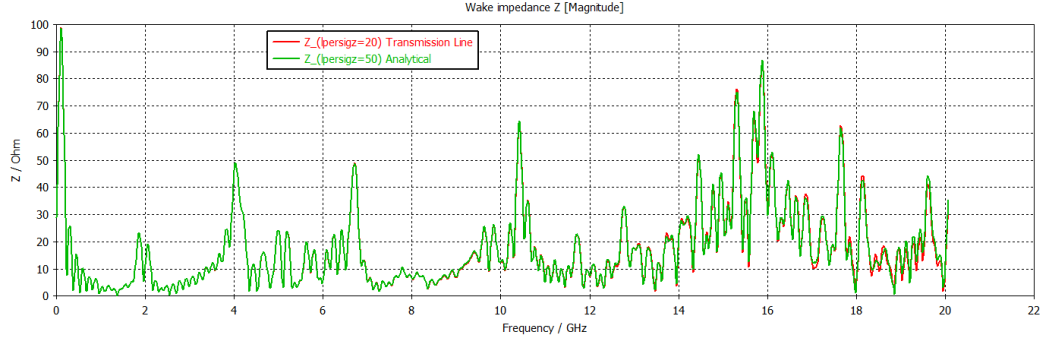


Figure 7.28: Comparing the highest resolution analytical injection with the converged transmission line injection. There are still visible discrepancies above 12 GHz, indicating that the resolution for the analytical scheme has not converged yet.

dinal mesh resolution of 20 lines per σ_z was chosen for the simulations presented in this thesis. This allows for some buffer in the resolution without increasing the simulation time too much.

7.4.2 Gap and Cut-Off Frequency

Simulations of the same geometry at different gap settings show a noise in the impedance signal which starts at certain distinguished frequencies. Simulations were done for gaps of 7 mm, 11 mm, and 15 mm. Since the geometry is roughly rectangular it can be approximated by a rectangular waveguide. The cut-off frequency of a rectangular waveguide is given by

$$f_c = \frac{c}{2} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2} \quad (7.2)$$

where c is the speed of light, m, n denote the order of the $TE_{m,n}$ mode, a and b are the dimensions of the waveguide. We set $a = 70.2$ mm as the width of the taper and b as the gap of the taper, the significant dimension for the $TE_{0,1}$ mode

7 Appendix

cut-off frequency. For gaps of 7 mm, 11 mm, and 15 mm we get the following cut-off frequencies:

$$f_{7\text{ mm}} = 21.4\text{ GHz}$$

$$f_{11\text{ mm}} = 13.6\text{ GHz}$$

$$f_{15\text{ mm}} = 10.0\text{ GHz}$$

With that reference we can look at Fig. 7.29 which compares these three gap settings for the exit taper with added sliding contacts. The first peaks in the longitudinal impedance signal are marked for each gap setting. The first peaks for each gap are roughly at the cut-off frequencies for the corresponding rectangular waveguide. Up to the first peak however, the signal between each gap is very similar. The frequency of the peak increases together with the magnitude, as the gap gets smaller, which is what we would expect. After the first peak there is a plateau of low signal before the impedance gets noisy with distinct peaks. The frequency at which the signal becomes noisy coincides very well with the rectangular waveguide cut-off frequencies for each gap.

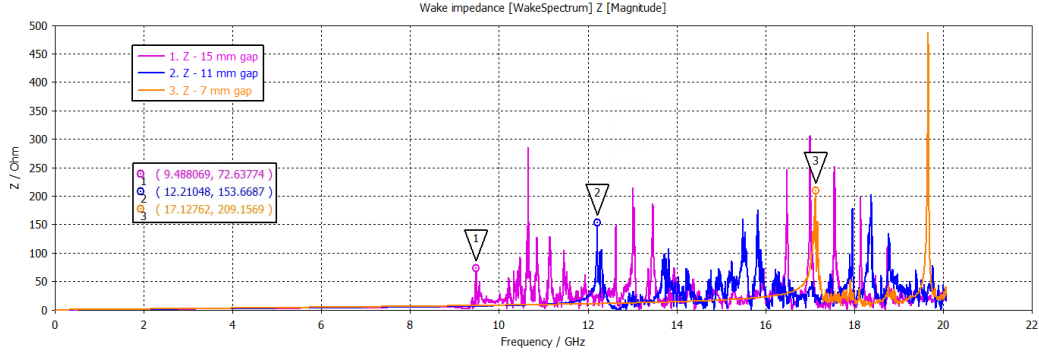


Figure 7.29: Longitudinal wake impedance of the exit taper assembly with sliding contacts added. Comparison between gaps of 7 mm, 11 mm, and 15 mm.

7.5 Horn Improvement Studies

Different horn antenna designs were studied in order to find a geometry that reduces internal reflections, improving the signal quality of TDR measurements on the Goubau line test stand. Figure 7.30 shows the cross section of selected horn antenna geometries that were simulated. All of them increase the opening diameter of the transmitter to 300 mm from the 90 mm diameter of the transmitter in the Goubau line test stand. Profile A) does so by increasing the opening angle and maintaining the same longitudinal dimension of the test stand transmitter at 136.5 mm. Profile B) maintains the same opening angle of the test stand transmitter and increases the length to 484 mm. Both profile A) and B) feature a simple

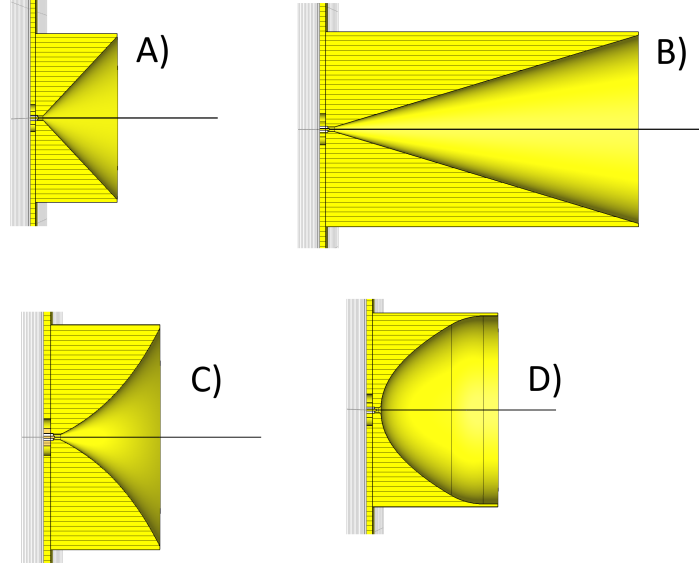


Figure 7.30: Cross section of different horn antenna geometries.

conical geometry. Profile C) features an exponential geometry and maintains the taper length of 136.5 mm. Profile C) features a parabolic geometry and a taper length of 196.5 mm. The simulated TDR signal of an empty Goubau line test stand with these transmitter geometries is shown in Fig. 7.31. The goal is to provide a

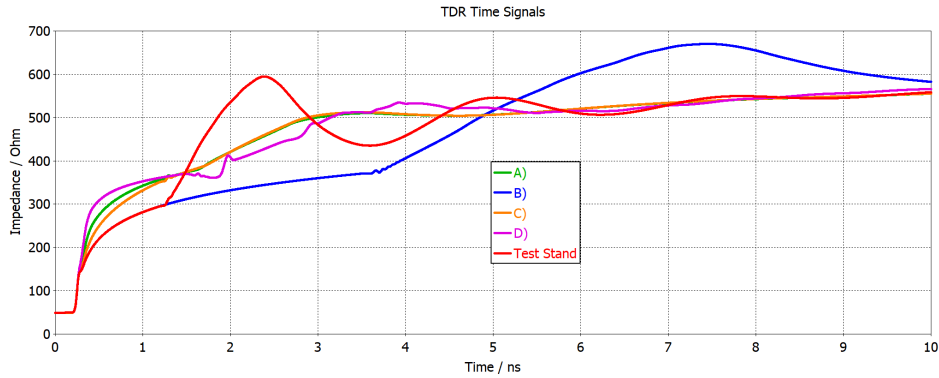


Figure 7.31: TDR signal of an empty Goubau line test stand for various transmitter geometries compared to the Goubau line test stand prototype.

constant TDR signal. Profiles A), C), and D) are the most promising candidates as the measured impedance after 3 ns shows very little variation. This means there are minimal internal reflections inside the transmitter. This allows for high fidelity time gating of possible DUTs. Future iterations of the Goubau line test stand will adopt one of these transmitter geometries.

Another geometry used in broad band antennas for satellite communications is the

7 Appendix

corrugated horn [38]. Corrugations are supposed to separate the guided waves from the walls of the horn, limiting reflections at the mouth of the horn. Two types of corrugations, as shown in Fig.7.32, were simulated. The TDR signals are shown in

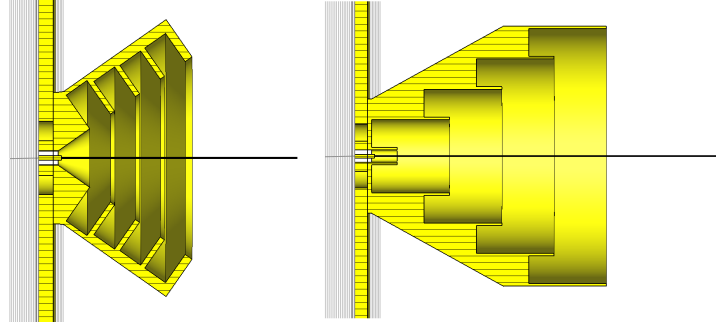


Figure 7.32: Cross section of horn antenna geometries with different types of corrugations.

Fig. 7.33 None of the simulated corrugated transmitters delivered satisfying results.

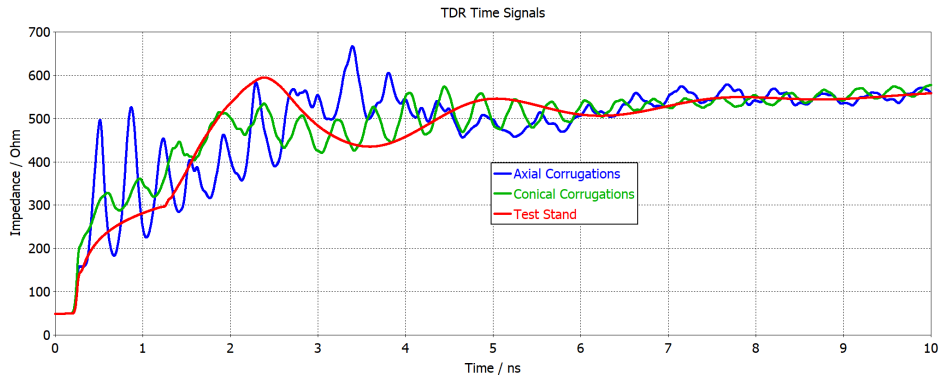


Figure 7.33: TDR signal of an empty Goubau line test stand for different corrugated transmitter geometries compared to the Goubau line test stand prototype.

The Goubau line differs from the intended use case of these corrugated antennas. They are designed for hollow waveguide feed ins and meant to transmit unguided, directional waves, where these designs are well proven. These geometries might not be suitable as transmission line tapers.

Bibliography

- [1] BESSY II facts and figures. https://www.helmholtz-berlin.de/forschung/quellen/bessy/bessy-in-zahlen_en.html. Accessed: 2026-01-08.
- [2] V.M. Borovikov, V.K. Djurba, M.G. Fedurin, V.V. Repkov, G.V. Karpov, G.N. Kulipanov, M.V. Kuzin, N.A. Mezentsev, V.A. Shkaruba, D. Kraemer, and D. Richter. Superconducting 7t wave length shifter for bessy-ii. *Nuclear Instruments and Methods in Physics Research Section A: Accelerators, Spectrometers, Detectors and Associated Equipment*, 467-468:181–184, 2001. 7th Int.Conf. on Synchrotron Radiation Instrumentation.
- [3] P Kuske, M Abo-Bakr, W Anders, T Birke, K B Buerkmann-Gehrlein, M Dirsat, O Dressler, V Dürr, F Falkenstern, W Gericke, R Görgen, F Hoffmann, T Kamps, J Kuszynski, D Lipka, I Müller, R Müller, K Ott, J Rahn, T Schneegans, D Schueler, T Westphal, and G Wüstefeld. Preparations of BESSY for Top Up Operation. 2008.
- [4] R. Müller, R. Goergen, R. Lange, I. Mueller, and J. Rahn. Introducing Fast Orbit Feedback at BESSY. In *Proc. ICALEPCS'09*, pages 773–775. JACoW Publishing, Geneva, Switzerland.
- [5] J. Bahrtdt et al. Characterization and Implementation of the Cryogenic Permanent Magnet Undulator CPMU17 at Bessy II. In *Proc. 10th International Particle Accelerator Conference (IPAC'19), Melbourne, Australia, 19-24 May 2019*, number 10 in International Particle Accelerator Conference, pages 1415–1418, Geneva, Switzerland, Jun. 2019. JACoW Publishing. <https://doi.org/10.18429/JACoW-IPAC2019-TUPGW014>.
- [6] Operating modi bessy ii. https://www.helmholtz-berlin.de/forschung/oe/be/operation-accelerator/betriebsmodi_en.html. Accessed: 2026-01-08.
- [7] H. Motz. Applications of the radiation from fast electron beams. *Journal of Applied Physics*, 22(5):527–535, 05 1951.
- [8] H. Motz, W. Thon, and R. N. Whitehurst. Experiments on radiation by fast electron beams. *Journal of Applied Physics*, 24(7):826–833, 07 1953.
- [9] Alexander Wu Chao and Maury Tigner. *Handbook of Accelerator Physics and Engineering*. World Scientific, P O Box 128, Fairer Road, Singapore 912805, 1999.

- [10] Kai Zhang and Marco Calvi. Review and prospects of world-wide superconducting undulator development for synchrotrons and fels. *Superconductor Science and Technology*, 35(9):093001, jul 2022.
- [11] Photo archive of hzb undulator group. https://www.helmholtz-berlin.de/forschung/oe/wi/undulators/index_en.html.
- [12] M. Huck, J. Bahrtdt, A. Meseck, G. Rehm, M. Ries, and A. Schällicke. Experimental Studies of the In-Vacuum-Cryogenic Undulator Effect on Beam Instabilities at BESSY II. In *Proc. IPAC'21*, number 12 in International Particle Accelerator Conference, pages 2929–2932. JACoW Publishing, Geneva, Switzerland, 08 2021. <https://doi.org/10.18429/JACoW-IPAC2021-WEPAB134>.
- [13] Johannes Bahrtdt and Carsten Kuhn. Cryogenic permanent magnet undulator development at hzb/bessy ii. *Synchrotron Radiation News*, 28(3):9–14, 2015.
- [14] T. Bizen, T. Hara, X. Marechal, T. Seike, T. Tanaka, and H. Kitamura. Development of in-vacuum revolver undulator. *AIP Conference Proceedings*, 705(1):175–178, 05 2004.
- [15] J. Bahrtdt and S. Grimmer. In-vacuum apple ii undulator with force compensation. *AIP Conference Proceedings*, 2054(1):030031, 01 2019.
- [16] Ryota Kinjo and Takashi Tanaka. Undulator configuration for helicity switching in in-vacuum undulators. *Phys. Rev. Accel. Beams*, 23:020705, Feb 2020.
- [17] A. Meseck P. Volz and S. Grimmer. Wakefield studies of the taper section of the elliptical in-vacuum undulator - ivue32. In *Proc. IPAC'25*, number 16 in IPAC'25 - 16th International Particle Accelerator Conference, pages 1591–1594. JACoW Publishing, Geneva, Switzerland, 06 2025.
- [18] P. Goslawski K. Holldack Z. Hüsches et al. Bessy ii the materials discovery facility. Technical Report DOI: 10.5442/r0004, Hahn-Meitner-Platz 1 14109 Berlin, Germany, 2022.
- [19] E. Rial et al. Undulators for bessy iii. In *Proc. IPAC'23*, number 14 in IPAC'23 - 14th International Particle Accelerator Conference, pages 1219–1222. JACoW Publishing, Geneva, Switzerland, 05 2023.
- [20] M. Ferrario, M. Migliorati, and L. Palumbo. Wakefields and Instabilities in Linear Accelerators. pages 357–375, 2014. 19 pages, contribution to the CAS - CERN Accelerator School: Advanced Accelerator Physics Course, Trondheim, Norway, 18-29 Aug 2013.
- [21] H. De Gersem F. Quetscher and E. Gjonaj. Shunt impedance calculations for an in-vacuum undulator at petra iv. In *Proc. IPAC'23*, number 14 in IPAC'23 - 14th International Particle Accelerator Conference, pages 1064–1067. JACoW Publishing, Geneva, Switzerland, 05 2023.

- [22] E. Gjonaj et al. Impedance computation for large accelerator structures using a domain decomposition method. In *Proc. IPAC'24*, number 15 in IPAC'24 - 15th International Particle Accelerator Conference, pages 3159–3162. JACoW Publishing, Geneva, Switzerland, 05 2024.
- [23] Wake: Wakefield and beam impedance studies for undulator structures at bessy ii and the eufel. <https://www.iae.uni-rostock.de/forschung/theoretische-elektrotechnik/elektromagnetische-felder-in-der-beschleunigerphysik/wake/>. Accessed: 2026-02-18.
- [24] Bruno W. Zotter and Semyon A. Kheifets. *Impedances and Wakes in High-Energy Particle Accelerators*. World Scientific, P O Box 128, Fairer Road, Singapore 912805, 1998.
- [25] A. Piwinski. Wake Fields and Ohmic Losses in Round Vacuum Chambers. 5 1992.
- [26] G. Stupakov. Low frequency impedance of tapered transitions with arbitrary cross sections. *Phys. Rev. ST Accel. Beams*, 10:094401, Sep 2007.
- [27] Georg Goubau. Surface waves and their application to transmission lines. *Journal of Applied Physics*, 21(11):1119–1128, 11 1950.
- [28] A Sommerfeld. Fortpflanzung elektrodynamischer wellen an einem zylindrischen leiter. *Annalen der Physik und Chemie*, 67:233–290, 1899.
- [29] Medani P. Sangroula, Ryan R. Lindberg, Robert M. Lill, and Randall Zabel. Measuring vacuum component impedance for the argonne advanced photon source upgrade. *Phys. Rev. Accel. Beams*, 23:082803, Aug 2020.
- [30] B. Vaughn and D. Peroulis. An updated applied formulation for the goubau transmission line. *Journal of Applied Physics*, 126(19):194902, 11 2019.
- [31] Dassault Systèmes. Dassault systèmes simulia cst studio suite.
- [32] M. Huck, J. Bahrtdt, H. Huck, A. Meseck, and M. Ries. Transverse Broad-band Impedance Studies of the New In-vacuum Cryogenic Undulator at Bessy II Storage Ring. In *Proc. IBIC'20*, number 9 in International Beam Instrumentation Conference, pages 263–267. JACoW Publishing, Geneva, Switzerland, 10 2020. <https://doi.org/10.18429/JACoW-IBIC2020-THPP26>.
- [33] Johannes Bahrtdt. personal communication.
- [34] P. Volz and A. Meseck. Loss simulations on shielding foil slit errors. In *Proc. IPAC'23*, number 14 in IPAC'23 - 14th International Particle Accelerator Conference, pages 3494–3497. JACoW Publishing, Geneva, Switzerland, 05 2023.

Bibliography

- [35] S.Y. Kim, F. Stulle, C.K. Sung, K.H. Yoo, J. Seok, K.J. Moon, C.U. Choi, Y. Chung, G. Kim, H.J. Woo, J. Kwon, I.G. Lee, E.M. Choi, and M. Chung. Characterization of the goubau line for testing beam diagnostic instruments. *Journal of Instrumentation*, 12(12):P12016, dec 2017.
- [36] I. Beverte. Investigation of the partial permittivity of rigid polyurethane foams by a circular one-side-access capacitive sensor. *Polymers*, 17(5):602, February 2025.
- [37] Miaomiao Qian, Xinru Wang, LiYang Yao, and Yanchao Zhu. Enhanced mechanical and dielectric properties of polyurethane elastomers containing modified sio₂. *ACS Omega*, 9(47):47315–47323, 2024.
- [38] Majid Abbas-Azimi, Farhad Mazlumi, and Fereidoon Behnia. Design of broadband constant-beamwidth conical corrugated-horn antennas [antenna designer’s notebook]. *Antennas and Propagation Magazine, IEEE*, 51:109 – 114, 11 2009.

8 Acknowledgments

I would like to thank Prof. Dr. Atoosa Meseck for her invaluable guidance throughout my PhD program.

I would like to thank Prof. Dr. Kurt Aulenbacher for reviewing my dissertation.

I would like to thank Prof. Dr. Adolfo Velez Saiz and Dr. Günther Rehm for their advice and guidance for measurements and getting my experiment started.

I would like to thank Prof. Dr. Ursula van Rienen, Dr. Erion Gjonaj, Alistair Muir and Frederik Quetscher together with the WAKE collaboration for providing valuable insights into the wakefield simulations using the CST Studio Suite.

I would like to thank Stefan Grimmer and the undulator group at HZB for their collaboration in the design process of the IVUE32 taper section.

Last but not least I would like to thank my wife Anika for putting up with me during these last weeks and motivating me in difficult times and our little daughter Frida for cheering me on.

Eigenständigkeitserklärung

Hiermit erkläre ich,

dass ich die vorliegende Arbeit selbstständig verfasst und keine anderen als die angegebenen Quellen und Hilfsmittel (dazu zählen auch KI-basierte Anwendungen oder Werkzeuge*) benutzt habe. Sämtliche wörtlichen oder sinngemäßen Übernahmen und Zitate sind kenntlich gemacht und nachgewiesen.

Ich habe keine KI-Tools verwendet.

Mit Abgabe der vorliegenden Leistung übernehme ich die Verantwortung für das eingereichte Gesamtprodukt. Ich verantworte damit auch jegliche KI-generierten Inhalte, die ich in meine Arbeit übernommen habe. Die Richtigkeit übernommener (KI-generierter) Aussagen und Inhalte habe ich nach bestem Wissen und Gewissen geprüft.

Mir ist bekannt, dass ein Verstoß gegen die genannten Punkte prüfungsrechtliche Konsequenzen hat.

Berlin, 20.02.2026

Ort, Datum und Unterschrift

